

Special Issue Reprint

Recent Advances in Power Quality Analysis and Robust Control of Renewable Energy Sources in Power Grids

Edited by
Dinko Vukadinović

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Guest Editor

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Recent Advances in Power Quality Analysis and Robust Control of Renewable Energy Sources in Power Grids

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In modern power grids with a large share of distributed power production, achieving high-power quality is a challenging task. The integration of renewable energy sources (RES) into power grids has introduced various disturbances, such as harmonic distortions, voltage sags, frequency variations, voltage unbalance, etc., leading to power quality degradation. In particular, the variable nature of RESs may cause rapid voltage changes, and consequently, the protection system can cause the disconnection of power system parts.

To enable the flexible operation of RESs, voltage source converters, and associated control systems are necessary. Very often, several RESs with storage devices, loads, and voltage source converters are connected in so-called microgrids. Microgrids are an emerging concept, closely connected with power quality, novel control methods, the Internet of Things, cyber security, etc. They are designed to function either in grid-connected or standalone modes, depending upon their energy potential and grid code requirements.

Keeping in mind that about 30% of the world's electricity generation comes from RESs (including hydropower), problems and challenges of RESs' integration into power grids and the design of associated control systems are the focus of many researchers.

This Special Issue includes eight research papers in the field of power factor correction, supercapacitor parameters identification, pulse width modulation, disturbance analysis in renewable energy systems, the vibration analysis of a line-start permanent magnet synchronous motor, the backstepping control of neutral point clamped converters, a distributed level phasor measurement unit, and the optimization of PI controller parameters using the Gray Wolf algorithm.

In [1], the analysis of currents in the three-phase system from the viewpoint of the displacement power factor and the total harmonic distortion is performed. Different loads are selected for both the industry and the home sector, as follows: induction motors, inductive loads, compact fluorescent lamps, and capacitive loads. This analysis is carried out depending on the number of steps used for the power factor controller, the values of the capacitor banks, the values of AC reactors connected in series with all loads, and the LC shunt filters. For unbalanced electrical networks with slightly deformed loads, it is recommended to use high-performance power factor controllers with a large number of capacitor banks to perform the regulation of the reactive power independently on each phase. As opposed to this, for unbalanced electrical networks with high deformation loads, it is recommended to use a small number of power factor controllers with LC filters to control the reactive power independently in each phase.

The authors in [2] parameterize the Zubieta model for supercapacitors, which involves identifying seven parameters using a hybrid metaheuristic gradient-based optimization approach. The effectiveness of this method is compared to the particle swarm optimization and to the two new methods developed in this paper, both based on particle swarm optimization. Rigorous experimentation considering various types of input excitation provides results indicating that the two new methods show improvements of 51% and 94%, respectively.

Grgić et al. [3] proposed two novel decoupled space-vector pulse-width modulation schemes for three-phase quasi-Z-source inverters. These schemes integrate dead time, and the shoot-through states are synchronized with the beginning of the zero-switching state. These novel schemes are designed to reduce switching losses and improve qZSI controllability. The obtained results indicate that the proposed pulse-width modulation schemes offer up to 6.8% greater efficiency and up to 7.5% reduced voltage stress compared to the closest competing pulse-width modulation schemes from the literature.

In [4], a condition monitoring strategy for the detection and identification of power quality disturbances in renewable energy generation systems is proposed. This strategy is based on the continuous wavelet transform and convolutional neural network. The proposed methodology is validated under a synthetic set of signals and a real set of signals acquired from a 30 MW photovoltaic system. The novelty of this work lies in the image processing that allows us to highlight the discriminant patterns through spectrograms into 2D images precisely.

Perez-Anaya et al. [5] deal with the vibration analysis of a line-start permanent magnet synchronous motor supplied with the voltage containing negative-sequence subharmonics. The results presented in this paper show the presence of unacceptable vibration for both no and full loads. For the full-load condition with voltage subharmonic values significantly less than those observed in real power systems, the maximal vibration velocity exceeds the threshold value of evaluation Zone D defined by the standard ISO 20816-1 Mechanical vibration—Measurement and Evaluation of Machine Vibration—Part 1: General Guidelines.

The modified backstepping methods to design controllers for neutral point-clamped converters interfacing with a DC/AC microgrid are proposed in [6]. These controllers are derived from the converter model and are designed to control the DC voltage and balance the two DC capacitor voltages in the neutral point-clamped converter. The proposed backstepping controller is free from high-frequency harmonics and ensures a very low total harmonic distortion ($\text{THD} < 1.7\%$) of AC currents and very small steady-state DC errors ($< 0.5\%$). It also balances the two DC capacitor voltages, controls the power factor, and shows bidirectional power transfer capability.

A hierarchical multi-stage approach to restore system voltage when it exceeds the limits recommended by the IEEE 1547-2018 standard is proposed in [7]. This approach is based on a distributed level phasor measurement unit at local controllers and a phasor data concentrator at the central control unit. The proposed algorithm reduces unnecessary curtailment, leading to a more efficient voltage control system. The simulation results obtained using the IEEE 13-bus radial distribution benchmark system confirm the efficacy of the proposed control system for achieving coordinated voltage control and ensuring the accuracy of the communication interface. In addition, to demonstrate its scalability, the proposed coordinated algorithm was successfully applied to the IEEE 33-bus distribution network, showcasing its effectiveness across a larger system.

A control method to reduce the speed oscillations in a five-phase asynchronous motor drive at low speed is proposed in [8]. This method is based on the Gray Wolf optimization algorithm with the goal of the PI controllers' parameter adjustment. The proposed approach exploits the rapid optimization process of the Gray Wolf optimization algorithm under real-time working conditions to determine the optimal stator voltage levels that ensure the desired tracking speed at high torque with less ripple.

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Article

Optimization of PI Controller Parameters by GWO Algorithm for Five-Phase Asynchronous Motor

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Abstract: Operation at low speed and high torque can lead to the generation of strong ripples in the speed, which can deteriorate the system. To reduce the speed oscillations when operating a five-phase asynchronous motor at low speed, in this article, we propose a control method based on Gray Wolf optimization (GWO) algorithms to adjust the parameters of proportional–integral (PI) controllers. Proportional–integral controllers are commonly used in control systems to regulate the speed and current of a motor. The controller parameters, such as the integral gain and proportional gain, can be adjusted to improve the control performance. Specifically, reducing the integral gain can help reduce the oscillations at low speeds. The proportional–integral controller is insensitive to parametric variations; however, when we employ a GWO optimization strategy based on PI controller parameters, and when we choose gains wisely, the system becomes more reliable. The obtained results show that the hybrid control of the five-phase induction motor (IM) offers high performance in the permanent and transient states. In addition, with this proposed strategy controller, disturbances do not affect motor performance.

Keywords: PI-controller; five-phase asynchronous motor; Gray Wolf optimization; speed oscillations

1. Introduction

The study and control of the asynchronous machine are currently one of the major concerns of researchers in the field compared to the direct current machine; this machine has many advantages in terms of dynamic performance, longevity, and its ability to withstand overloads, which justifies its use in various industrial environments, including electric traction [1]. However, the reduced number of phases can be a handicap in particular areas of application, such as rail traction, naval propulsion, automotive, and aerospace [2]. In contrast, the properties of polyphase machines are very appealing, as a rise in the number of phases results in a decrease in the ripples of the electromagnetic torque [3,4]. The fractionation of the power, which results in a reduction in the sizing constraints on the power electronic components supplying each phase in addition to this increase, makes it possible to operate one or more faulty phases [3]. This issue is fundamental for applications that must guarantee an excellent continuity of service.

Many works have studied the control of polyphase machines, such as the flux orientation control [3], the fuzzy logic control [5], the direct torque control (DTC) [4], the predictive control (PC) [6], and the sliding mode control (MGC) [7]. Research has aimed to develop a control system for the polyphase asynchronous machine that operates at low speed and

high torque to reduce the occurrence of periodic torque ripples. These ripples can cause speed oscillations and potentially harm the machine [8].

The fields of control and regulation theory have been considerably developed over the past decades; however, the importance of the proportional–integral–derivative (PID) controller in the industry remains unaffected. Because of its straightforward design, it is simple to comprehend and operate, as well as resilient and stable across a wide range [5].

Unfortunately, proportional–integral (PI) controllers are widely used in control systems to regulate a process variable, such as the speed or position of a motor. The PI controller has two parameters, the proportional gain (K_p) and integral gain (K_i), that determine the response of the controller. The PI controller parameters can be tuned using several methods, including metaheuristic algorithms [9]. Metaheuristic algorithms are optimization algorithms that can be used to find optimal solutions for complex problems. Here are some examples of metaheuristic algorithms that can be used to tune PI controllers: Inspired by the social behavior of bird flocks, Particle Swarm Optimization (PSO) is an optimization algorithm based on population. In PSO, a population of particles is used to search for the optimal solution by adjusting their positions based on the best solution [10]. Genetic Algorithm (GA) is an optimization algorithm that imitates the process of natural selection. It uses a group of individuals to explore the optimal solution by applying genetic operators, such as crossover and mutation, to produce new solutions [11–14]. The optimization algorithm known as Ant Colony Optimization (ACO) draws inspiration from the foraging behavior of ants. In ACO, a population of ants is used to search for the optimal solution by depositing pheromones on the paths that lead to better solutions [15,16]. Inspired by the foraging behavior of honeybees, Artificial Bee Colony (ABC) is an optimization algorithm. The Artificial Bee Colony (ABC) algorithm imitates the behavior of real bees, where a population of artificial bees is employed to explore the optimal solution [17]. Grey Wolf Optimization (GWO) is an optimization algorithm inspired by the hunting behavior of grey wolves [18]. In Grey Wolf Optimization (GWO), a population of wolves is employed to search for the optimal solution by mimicking the hunting behavior of a wolf pack. The Grey Wolf Optimizer and Particle Swarm Optimization are two popular metaheuristic optimization algorithms that are used to solve complex optimization problems. Both algorithms are based on the principles of swarm intelligence and are inspired by the behavior of animals in nature. However, there are some key differences between GWO and PSO: Optimization approach: GWO is a population-based optimization algorithm, where a group of grey wolves search for the optimal solution to a problem. PSO, on the other hand, is based on the behavior of a swarm of particles that explore the solution space. Solution representation: In GWO, the solution to a problem is represented as a position vector of the grey wolves, while, in PSO, it is represented as a position vector of the particles. Exploration vs. exploitation: GWO emphasizes the exploration of the search space in the early stages of the optimization process and gradually shifts towards exploitation in the later stages. PSO, on the other hand, maintains a balance between exploration and exploitation throughout the optimization process. Convergence rate: GWO has been shown to have a faster convergence rate than PSO for some optimization problems. However, PSO is known to be more effective at handling high-dimensional problems. Initialization: GWO requires an initial population of grey wolves to begin the optimization process, while PSO only requires an initial set of particles. In summary, both GWO and PSO are effective optimization algorithms that can be used to solve a wide range of problems. The choice between them depends on the specific problem being addressed and the performance requirements of the solution [17–20]. To achieve precise monitoring of the rotational speed with minimal ripple in the low-speed and high-torque operations of the five-phase asynchronous machine, the algorithm precisely tunes the parameters of the proportional–integral controller (PI) [19–24]. Although the advantages mentioned earlier are noteworthy, conventional PID controllers need a linear model, which cannot guarantee robustness against uncertainties because of the significant nonlinear dynamics of a five-phase induction motor (IM). In this paper, we propose a recently developed optimization method based on the Grey Wolf algorithm

(GWO). Under low speed and high torque conditions, this algorithm optimizes the gains of the PI controller to determine the optimal path of input commands that achieves the desired speed while minimizing ripple in the five-phase asynchronous machine [18].

This document is structured as follows: the first section covers the system description, and the second section addresses the modeling of the five-phase IM. The design methodology of the PI controller based on GWO is presented in the third section. The fourth section illustrates the results of the simulation and discusses the efficacy of the suggested control approach, and the final section provides a summary of the contributions of this work.

2. System Description

Achieving optimal control performance through the PID regulator requires a methodical approach to determine the appropriate values for the controller's proportional, integral, and derivative gains. We propose a meta-heuristic optimization technique based on the hunting style of grey wolves as illustrated in Figure 1. The five-phase IM control structure employs three control loops: the motor angular speed control loop Ω that uses a PI controller, and two PI controllers for the stator flux ψ and the electromagnetic torque C_{em} . The primary electromagnetic torque is obtained directly from the external speed loop. The two regulators output the reference stator voltage (v_{qsref}, v_{dsref}) along the q and d axes. After the inverse Park transformation, we obtain the real stator voltage, which produces the real speed of the motor. The inputs of the estimator are the real currents along the (d,q) axis obtained after the Park transformation of the real currents. The estimation block determines the instantaneous amplitude of the stator flux ψ_s , the electromagnetic torque C_{em} , and the stator electric angles θ_s .

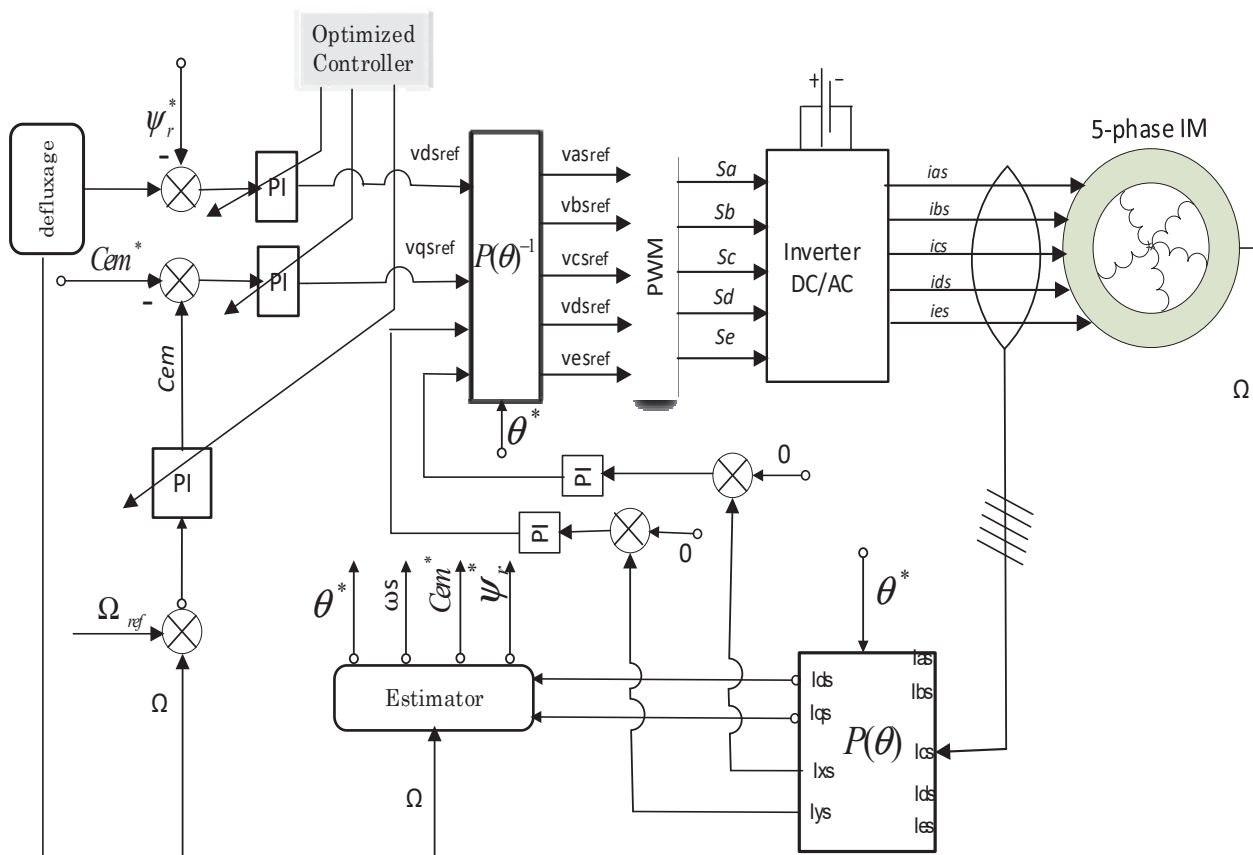


Figure 1. Schematic diagram of Optimized Controller applied on five-phase IM.

3. Modeling of Five Phase Induction Machine

The voltage model of the five phase induction motor is obtained as [5]:

$$\begin{cases} [V_s]_{1 \times 5} = [R_s]_{5 \times 5} [i_s]_{1 \times 5} + \frac{d}{dt} [\psi_s]_{1 \times 5} \\ [V_r]_{1 \times 5} = [R_r]_{5 \times 5} [i_r]_{1 \times 5} + \frac{d}{dt} [\psi_r]_{1 \times 5} \\ [\psi_s]_{1 \times 5} = [L_s]_{5 \times 5} [i_s]_{1 \times 5} + [M_{sr}]_{5 \times 5} [i_r]_{1 \times 5} \\ [\psi_r]_{1 \times 5} = [L_r]_{5 \times 5} [i_r]_{1 \times 5} + [M_{rs}]_{5 \times 5} [i_s]_{1 \times 5} \end{cases} \quad (1)$$

where $[R_s]_{5 \times 5}$ and $[R_r]_{5 \times 5}$ are the resistance matrices, $[L_s]_{5 \times 5}$ and $[L_r]_{5 \times 5}$ are the matrices of inductances, $[M_{sr}]_{5 \times 5}$ and $[M_{rs}]_{5 \times 5}$ are mutual inductance between stator and rotor matrices, $[V_s]_{1 \times 5}$ and $[V_r]_{1 \times 5}$ are the voltage matrices, $[i_s]_{1 \times 5}$ and $[i_r]_{1 \times 5}$ are the current matrices, $[\psi_s]_{1 \times 5}$ and $[\psi_r]_{1 \times 5}$ are flux linkages matrices, and the subscripts s and r refer to the stator and rotor windings, respectively. The mutual inductance matrices are given below.

$$[M_{sr}]_{5 \times 5} = [M_{rs}]_{5 \times 5}^t \quad (2)$$

All resistance matrices are symmetric as shown below, which is a direct result of the perfect symmetry of the machine:

$$\begin{aligned} [R_s]_{5 \times 5} &= R_s * [I]_{5 \times 5} \\ [R_r]_{5 \times 5} &= R_r * [I]_{5 \times 5} \end{aligned} \quad (3)$$

where R_s and R_r are stator resistance and stator inductance, respectively, and I is the matrix consisting of ones on the main diagonal and zeros elsewhere; this is referred to as the identity matrix.

The stator inductance matrix is obtained using:

$$[L_s]_{5 \times 5} = \begin{bmatrix} L_m + L_{fs} & L_m \cos\left(\frac{2\pi}{5}\right) & L_m \cos\left(\frac{4\pi}{5}\right) & L_m \cos\left(\frac{6\pi}{5}\right) & L_m \cos\left(\frac{8\pi}{5}\right) \\ L_m \cos\left(\frac{8\pi}{5}\right) & L_m + L_{fs} & L_m \cos\left(\frac{2\pi}{5}\right) & L_m \cos\left(\frac{4\pi}{5}\right) & L_m \cos\left(\frac{6\pi}{5}\right) \\ L_m \cos\left(\frac{6\pi}{5}\right) & L_m \cos\left(\frac{8\pi}{5}\right) & L_m + L_{fs} & L_m \cos\left(\frac{2\pi}{5}\right) & L_m \cos\left(\frac{4\pi}{5}\right) \\ L_m \cos\left(\frac{4\pi}{5}\right) & L_m \cos\left(\frac{6\pi}{5}\right) & L_m \cos\left(\frac{8\pi}{5}\right) & L_m + L_{fs} & L_m \cos\left(\frac{2\pi}{5}\right) \\ L_m \cos\left(\frac{2\pi}{5}\right) & L_m \cos\left(\frac{4\pi}{5}\right) & L_m \cos\left(\frac{6\pi}{5}\right) & L_m \cos\left(\frac{8\pi}{5}\right) & L_m + L_{fs} \end{bmatrix} \quad (4)$$

It is possible to express the mutual matrix inductance in the following manner:

$$[M_{sr}]_{5 \times 5} = L_m \begin{bmatrix} \cos(\theta) & \cos\left(\theta - \frac{2\pi}{5}\right) & \cos\left(\theta - \frac{4\pi}{5}\right) & \cos\left(\theta - \frac{6\pi}{5}\right) & \cos\left(\theta - \frac{8\pi}{5}\right) \\ \cos\left(\theta - \frac{8\pi}{5}\right) & \cos(\theta) & \cos\left(\theta - \frac{2\pi}{5}\right) & \cos\left(\theta - \frac{4\pi}{5}\right) & \cos\left(\theta - \frac{6\pi}{5}\right) \\ \cos\left(\theta - \frac{6\pi}{5}\right) & \cos\left(\theta - \frac{8\pi}{5}\right) & \cos(\theta) & \cos\left(\theta - \frac{2\pi}{5}\right) & \cos\left(\theta - \frac{4\pi}{5}\right) \\ \cos\left(\theta - \frac{4\pi}{5}\right) & \cos\left(\theta - \frac{6\pi}{5}\right) & \cos\left(\theta - \frac{8\pi}{5}\right) & \cos(\theta) & \cos\left(\theta - \frac{2\pi}{5}\right) \\ \cos\left(\theta - \frac{2\pi}{5}\right) & \cos\left(\theta - \frac{4\pi}{5}\right) & \cos\left(\theta - \frac{6\pi}{5}\right) & \cos\left(\theta - \frac{8\pi}{5}\right) & \cos(\theta) \end{bmatrix} \quad (5)$$

where L_s , L_{fs} , and L_m are stator, stator leakage, and magnetizing inductances, respectively.

The transformation matrix $[P_n]$ for the five-phase induction motor is defined as follows [7]:

$$[P_n] = L_m \begin{bmatrix} \cos(\theta) & \sin(\theta) & 0 & 0 & 0 \\ -\sin(\theta) & \cos(\theta) & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 & 0 \\ 0 & 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 0 & 1 \end{bmatrix} \quad (6)$$

After applying Park's transformation matrix to the stator and rotor equations, the five-phase system can be decomposed into the dq coordinate system as well as into an additional xy coordinate system. The stator voltages of the system are written in the $d-q-x-y$ axis as follows [5–7]

$$\begin{cases} V_{ds} = r_s i_{ds} - \omega_s \psi_{qs} + \frac{d}{dt} \psi_{ds} \\ V_{qs} = r_s i_{qs} + \omega_s \psi_{ds} + \frac{d}{dt} \psi_{qs} \\ V_{xs} = r_s i_{xs} + \frac{d}{dt} \psi_{xs} \\ V_{ys} = r_s i_{ys} + \frac{d}{dt} \psi_{ys} \end{cases} \quad (7)$$

The flux equations, which become:

$$\begin{cases} \psi_{ds} = (L_m + L_{fs}) i_{ds} + L_m i_{dr} \\ \psi_{qs} = (L_m + L_{fs}) i_{qs} + L_m i_{qr} \\ \psi_{xs} = L_{fs} i_{xs} \\ \psi_{ys} = L_{fs} i_{ys} \end{cases} \quad (8)$$

The expression for the electromagnetic torque C_{em} has been utilized to finalize the model of the five-phase asynchronous machine.

$$C_{em} = \frac{P}{2} ([i_s]^t \frac{\partial [M_{sr}]}{\partial \theta} [i_r] + [i_r]^t \frac{\partial [M_{sr}]}{\partial \theta} [i_s]) \quad (9)$$

where P is the number of pole pairs.

The mechanical model is obtained using:

$$J \frac{d}{dt} \Omega_r = C_{em} - C_r - f \Omega_r \quad (10)$$

where C_{em} is the electromagnetics torques, C_r is the resistance torque, and J is the inertia.

4. Grey Wolf Optimizer

Grey Wolf Optimization (GWO) is a population-based metaheuristic optimization algorithm inspired by the hunting behavior of grey wolves in nature [12]. The algorithm was proposed by Seyedali Mirjalili, Shima Saremi, and Seyed Mohammad Mirjalili in 2014. This algorithm is applied to the machine as stated by Ali Djeriou in [18]. GWO mimics the social hierarchy and hunting behavior of grey wolves. A group of wolves, comprising of alpha, beta, delta, and omega wolves, operates in this algorithm. The alpha wolf is the pack leader, and the remaining wolves follow its lead. The beta, delta, and omega wolves have a lower pack rank and serve as subordinates to the alpha wolf. The GWO algorithm starts by randomly initializing the positions of the alpha, beta, delta, and omega wolves in the search space. Then, the fitness function is evaluated for each wolf, and the position of the alpha wolf is updated based on the best fitness value obtained so far. The positions of the other wolves are updated using a formula that depends on the positions of the alpha (α), beta (β), delta (δ), and omega (ω) wolves. The GWO algorithm has been applied successfully to a variety of optimization problems, including function optimization, feature selection, and classification. One of the advantages of GWO is its simplicity and ease of implementation. However, as with other metaheuristic algorithms, GWO cannot be guaranteed to find the global optimum for all optimization problems [12].

4.1. Objective Function

A function that is considered as a potential objective function is referred to as a candidate objective function. The objective function is a mathematical function that takes one or more input variables and produces a single output value. The optimization process involves searching for the input values that result in the highest or lowest possible output value, depending on the specific problem being addressed. It is a fundamental tool for

decision making and problem solving, and it enables researchers and practitioners to find the best solutions to complex problems.

To minimize the disparity between the estimated torque and flux and their references, traditional PI regulators are utilized; alternatively, the GWO algorithm is employed to obtain optimal gains. The GWO algorithm boasts several benefits, including fast convergence, robustness, and minimal adjustment of gains. The following candidate objective functions can be optimized using GWO:

$$H[n] = \begin{bmatrix} K_1 & K_2 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & K_3 & K_4 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & K_5 & K_6 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & K_7 & K_8 \end{bmatrix} T[n] \quad (11)$$

where n is the current iteration, $K_1 \dots 8$ are the constant gains, and $H[n]$ is the objective function vector with two objective functions along the axis (d, q) .

The first objective function optimization can be applied to optimize the flux of a machine. The objective function in this case would be a function that quantifies the performance or efficiency of the machine in terms of the flux ψ_d with only one control input u_d .

The second objective function optimization can be used to optimize the speed $\Omega[k]$ and torque C_{em} of a machine by determining the input variables that result in the highest possible speed with only one control input u_q . The objective function in this case would be a function that quantifies the performance or efficiency of the machine in terms of its speed. The objective function could be the maximum torque (or speed) that the machine can achieve while maintaining a certain level of speed (or torque). The extent to which the function should be minimized is closely associated with the amplitude of the speed ripple:

$$T[k] = [T_d[k], T_q[k]]^T \quad (12)$$

with

$$T_d[k] = [\psi^{ref}[k] - \psi^*[k], \psi^{ref}[k-1] - \psi^*[k-1]] \quad (13)$$

$$T_q[k] = \begin{bmatrix} \Omega^{ref}[k] - \Omega[k], \Omega^{ref}[k-1] - \Omega[k-1], \\ C_{em}^{ref}[k] - C_{em}^*[k], C_{em}^{ref}[k-1] - C_{em}^*[k-1] \end{bmatrix} \quad (14)$$

4.2. Store the Best Particle

The GWO algorithm is based on a set of mathematical equations that are used to simulate the hunting behavior of grey wolves. Here is the general equation used in GWO [12]:

$$\begin{cases} H[k+1] = H_p[k] - P * D \\ D = |C * H_p[k] - H[k]| \end{cases} \quad (15)$$

here, k denotes the current iteration, H_p represents the position vector of the optimal solution, and H represents an objective function and also the position vector of a search agent. D indicates the distance between the i th wolf and the alpha wolf. P is a random number ranging from 0 to 2, which reflects the encircling behavior of the wolves. The values of P and C are calculated as follows:

$$\begin{cases} P = 2p * r_1 - p \\ C = 2p * r_2 \end{cases} \quad (16)$$

where r_1 and r_2 are random numbers between $[0, 1]$. The GWO algorithm also includes equations for updating the positions of the beta and delta wolves, as well as equations for adjusting the search space and controlling the exploration–exploitation trade-off. However, the general equation shown above is the key equation used in the encircling behavior of

the wolves in GWO. The hunt is guided by the leader (alpha), which is the best solution candidate; the beta and delta have better knowledge about the potential location of the prey.

$$\begin{cases} D_{x\alpha} = C_{x1}v_{x\alpha}(k) - H_x(k) \\ D_{x\beta} = C_{x2}v_{x\beta}(k) - H_x(k) \\ D_{x\delta} = C_{x3}v_{x\delta}(k) - H_x(k) \end{cases} \quad (17)$$

We use the optimal distance vector $[D_{x\alpha}, D_{x\beta}, D_{x\delta}]$ to formulate an intermediate input control variable $(v_{x\alpha}, v_{x\beta}, v_{x\delta})$ defined by the following equation:

$$\begin{cases} v_{x1}(k+1) = v_{x\alpha}(k) - P_{x1} * D_{x\alpha}(k) \\ v_{x2}(k+1) = v_{x\beta}(k) - P_{x2} * D_{x\beta}(k) \\ v_{x3}(k+1) = v_{x\delta}(k) - P_{x3} * D_{x\delta}(k) \end{cases} \quad (18)$$

where $C_{xi} = 2p_{xi}r_{xi}$, $P_{xi} = 2p_{xi}r_{xi} - p_{xi}$, $i \in (1, 2, 3)$, and $x = (d, q)$.

The following equation is used to compute the candidate dq -input control, which corresponds to the inverter voltages:

$$v_{xref}(k+1) = \sum_{i=1}^{i=n} \frac{v_{xi}(k+1)}{3} \quad (19)$$

Following the previous step, the search for the specific position for an ideal hunt is based on the control of the particular distance of each wolf. For this, the voltage reference is chosen as follows:

$$\begin{cases} u_{xref} = v_{xref}(k+1) \text{ if } (|P_{xi}|) < 1 \\ u_{xref} = v_{xref}(k) \text{ if } (|P_{xi}|) > 1 \end{cases} \quad (20)$$

where v_{xref} is the actual input control, $v_{xref}(k+1)$ is the future input control, and u_{xref} is the input control.

4.3. The Flowchart of the Algorithm

The Grey Wolf Optimization (GWO) algorithm is a metaheuristic optimization algorithm inspired by the social hunting behavior of grey wolves. The flowchart of the GWO algorithm is as follows:

- Initialize the grey wolves' population, which is a set of candidate solutions.
- Calculate the objective function of each grey wolf in the population.
- Assign the best, second best, and third best grey wolves as alpha, beta, and delta, respectively.
- Update the position of each grey wolf using Equation (14).
- Update the objective function of each grey wolf in the new population.
- Check if any of the updated grey wolves have a better fitness than alpha, beta, or delta. If so, update the position of alpha, beta, or delta accordingly.
- Repeat steps 4 to 6 until the stopping criteria are met, such as a maximum number of iterations or a desired level of convergence.

In order to compare the performance of the Grey Wolf Optimizer (GWO) with another optimization algorithm, we need to select a suitable algorithm for comparison. We propose Particle Swarm Optimization (PSO) as a potential candidate for this purpose. The flowchart for the basic PSO algorithm and GWO algorithm is shown in Figure 2.

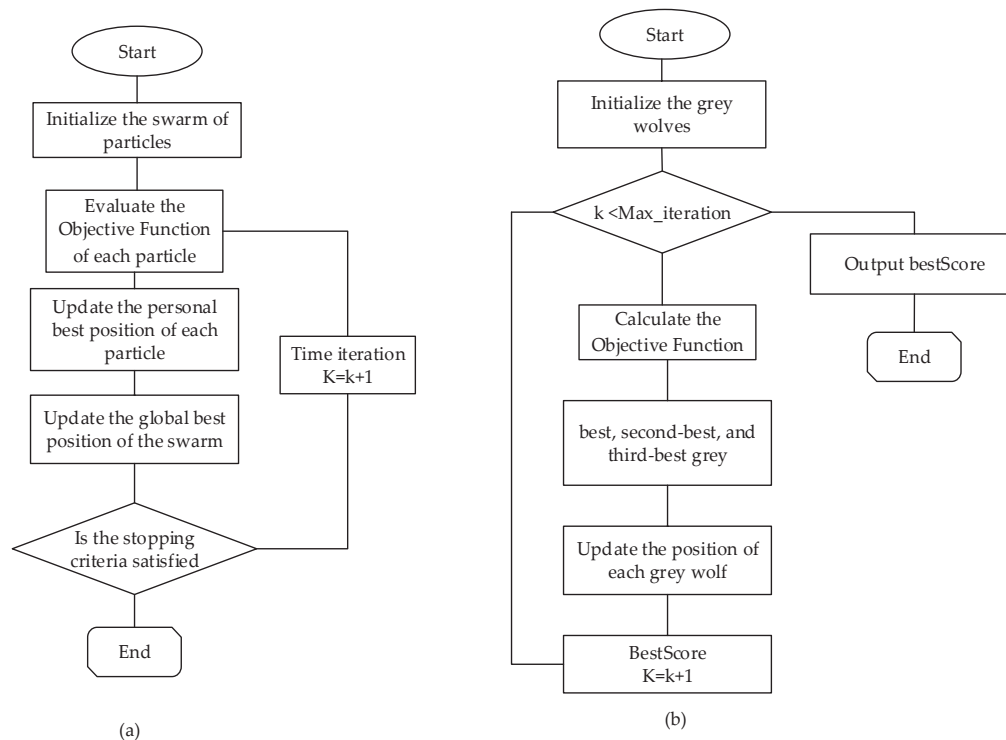


Figure 2. Flowchart for (a) basic PSO algorithm and (b) GWO algorithm.

5. PI Controller Parameters Using the GWO Algorithm

The following steps can be followed to optimize the PI controller parameters for a five-phase asynchronous motor using the GWO algorithm:

1. Define the problem: First, you need to clearly define the problem and identify the variables that need to be controlled. In this case, the objective is to optimize the PI controller parameters for controlling the speed of a five-phase asynchronous motor.
2. Choose the PI controller: A proportional–integral (PI) controller is commonly used in control systems to regulate a process variable. It is a type of feedback controller that uses the error between the desired setpoint and the measured process variable to adjust the control signal.
3. Determine the controller parameters: The PI controller has two parameters that need to be tuned: the proportional gain (K_p) and the integral gain (K_i). These parameters determine the response of the controller and can be adjusted to optimize the performance of the five-phase asynchronous motor.
4. Use the GWO algorithm for optimization: Grey Wolf Optimization (GWO) is a population-based optimization algorithm that is inspired by the hunting behavior of grey wolves. It is a metaheuristic optimization algorithm that can be used to find optimal solutions for complex problems. You can use GWO to optimize the PI controller parameters for the five-phase asynchronous motor.
5. Define the objective function: The fitness function is used to evaluate the performance of the PI controller. In this case, the fitness function should measure how well the controller is able to regulate the speed of the five-phase asynchronous motor.
6. Set up the GWO algorithm: The GWO algorithm requires several parameters to be defined, including the population size, maximum number of iterations, and search range for each parameter. The GWO algorithm can be set up to optimize the PI controller parameters for the five-phase asynchronous motor.
7. Run the optimization: Once the GWO algorithm is set up, it can be used to find the optimal values of K_p and K_i for the PI controller. The GWO algorithm will iterate through the population and adjust the values of K_p and K_i until the fitness function is optimized.

8. Evaluate the results.

Finally, the same can be performed to optimize the PI controller parameters for a five-phase asynchronous motor using the PSO algorithm.

6. Simulation

To validate the performance of the proposed scheme depicted in Figure 1, we present the simulations for a five-phase IM. The PI controller parameters are optimized using the GWO and PSO algorithm, whose rated values and nominal electrical and mechanical parameters are shown in Table 1. We obtained the values of PI controller gains using:

Table 1. Parameters of five-phase IM.

Parameters	Values	Parameters	Values
R_s	4.85Ω	L_s	0.024 H
R_r	3.805Ω	L_r	0.024 H
p	2	L_m	0.258 H
k	0.0005	j	0.085

Basic PSO algorithm: The controllers' gains are: $K_{pi} = 21.7200$, $K_{ii} = 47,700$, $K_{pi} = 11.7200$, $K_{ii} = 25,700$, $K_{p\omega} = 0.025375$ and $K_{i\omega} = 3.0290$.

GWO algorithm: The controllers' gains are: $K_{pi} = 22.1200$, $K_{ii} = 43,400$, $K_{pi} = 21.9200$, $K_{ii} = 42,300$, $K_{p\omega} = 0.032365$ and $K_{i\omega} = 2.0230$.

The results shown in Figure 3a–d demonstrate the favorable dynamic behavior of speed and torque achieved through the optimization of the PI controller parameters using the GWO algorithm. When the rotor speed and torque are in steady state, the stator resistance is changed from R_s to $R_s/4$ (stator faults). The PI control is a commonly used control strategy for induction motors, as it provides a simple and effective way to regulate the motor speed and torque. In this control strategy, the motor speed and torque are controlled by adjusting the magnitude and phase of the stator current. The speed reaches 314 V, given as Ω_{ref} at 0.05 s.

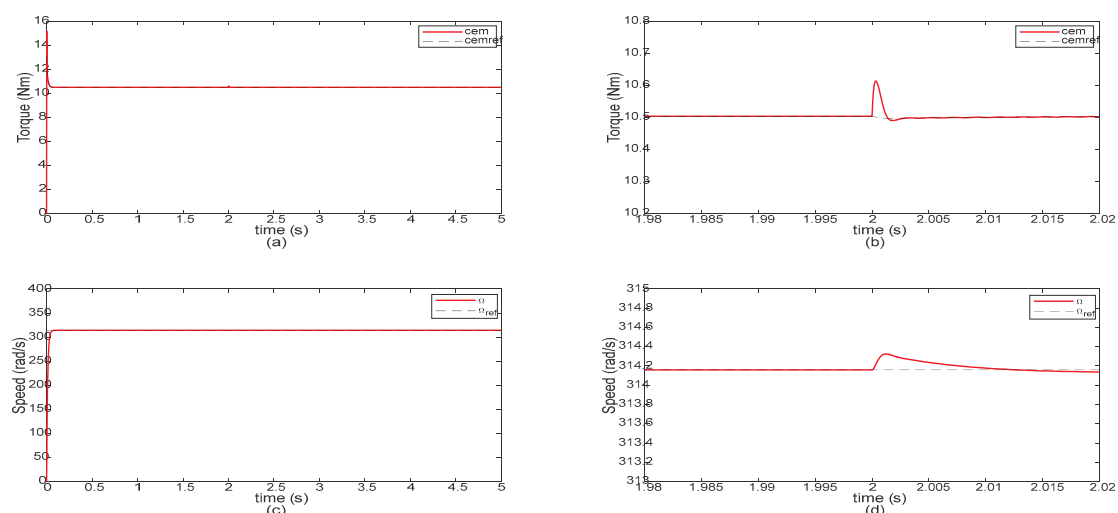


Figure 3. Torque (a,b) and speed (c,d) responses in the presence of stator faults with optimization of PI controller parameters using GWO algorithm.

In Figure 4a–d shows the current (i_{qs}) and relative tracking errors in the presence of stator faults in the optimization of the PI controller parameters using the GWO algorithm. The stator currents (i_{qs}) are shown in Figure 4a with the starting transients of the stator fault as it develops the relative tracking error, as displayed in Figure 4d. When the relative

tracking error is in steady state, the stator resistance is changed from R_s to $R_s/4$ to show the current harmonics.

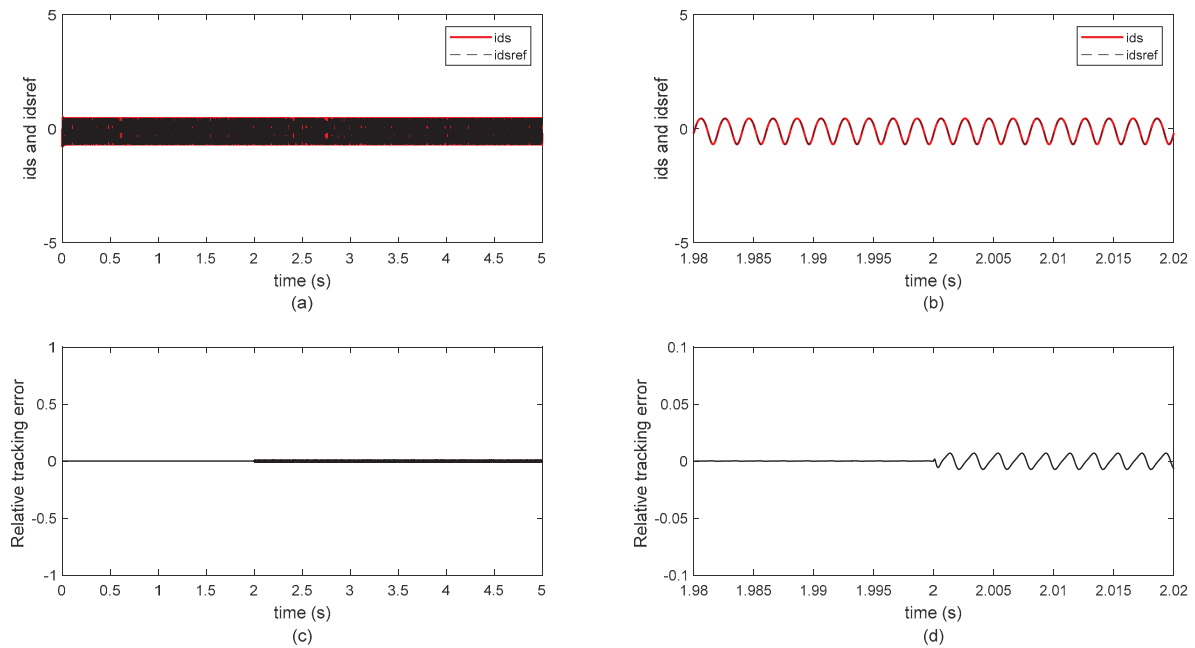


Figure 4. Current (i_{ds}) (a,b) and relative tracking error (c,d) in the presence of stator faults with the optimization of PI controller parameters using GWO algorithm.

Figures 5 and 6 show the current i_{qs} component and current i_{qp} component in the presence of stator faults with the optimization of PI controller parameters using the GWO algorithm, respectively. The currents (i_{qs}) are shown in Figure 4a with the starting transients of the stator fault as it develops the relative tracking error, as displayed in Figure 5d. When the relative tracking error is in steady state, the stator resistance is changed to show the current harmonics. This is also true for i_{qp} .

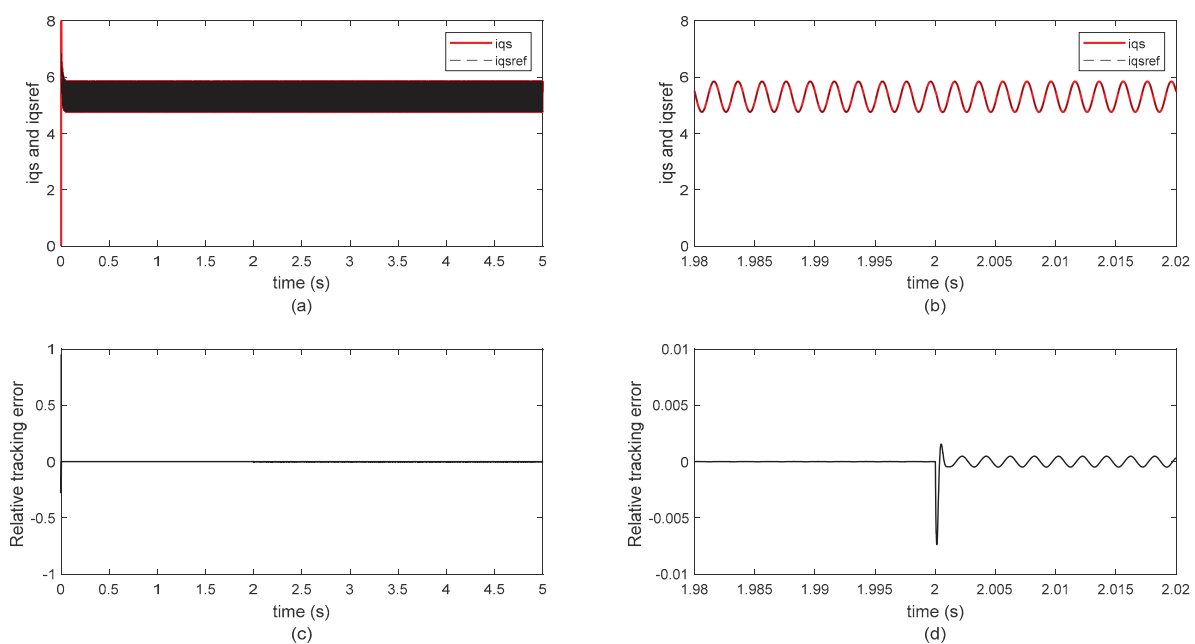


Figure 5. Current (i_{qs}) (a,b) and relative tracking error (c,d) in the presence of stator faults with the optimization of PI controller parameters using GWO algorithm.

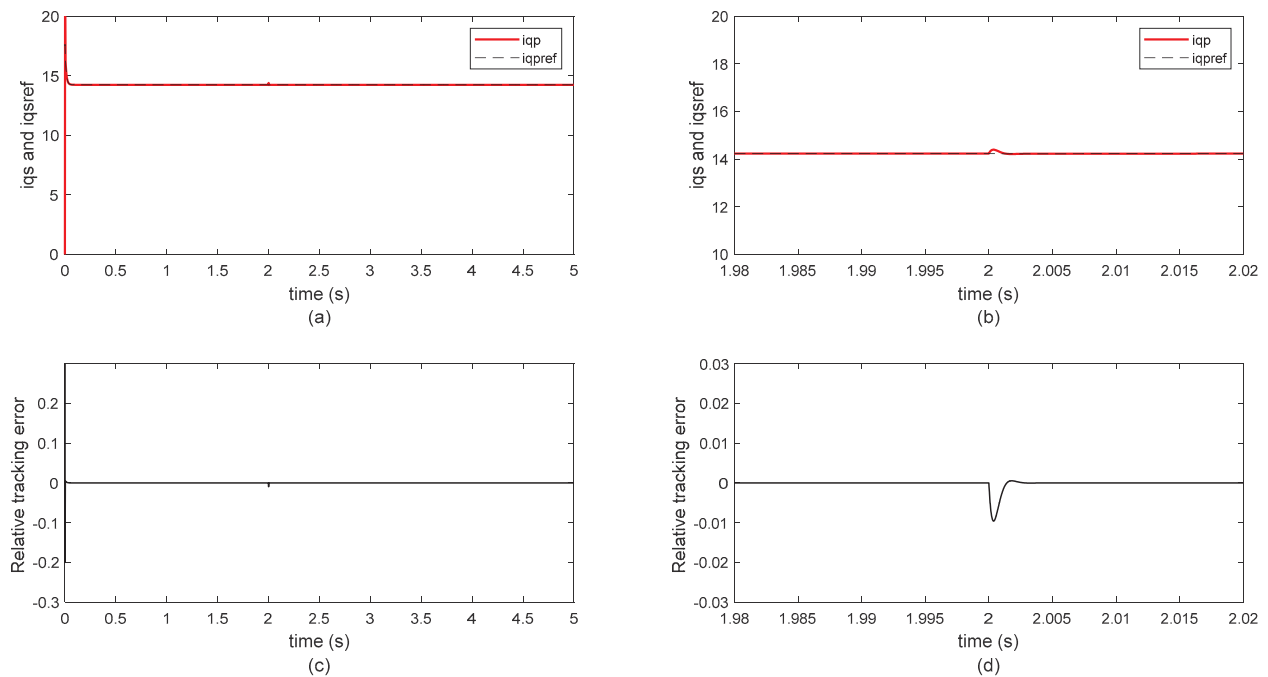


Figure 6. Current (i_{qp}) (a,b) and relative tracking error (c,d) in the presence of stator faults with the optimization of PI controller parameters using GWO algorithm.

Figure 7a–d show the electromagnetic torques' components (c_{emp} and c_{ems}) with the optimization of PI controller parameters using the GWO algorithm while subjecting the stator's resistance to a step changed from its nominal value to one-quarter of the nominal value (the speed reference is maintained at a constant level). The electromagnetic torque components for a five-phase machine can be expressed in the P–S reference frame. The total electromagnetic torque is the sum of the two torque components.

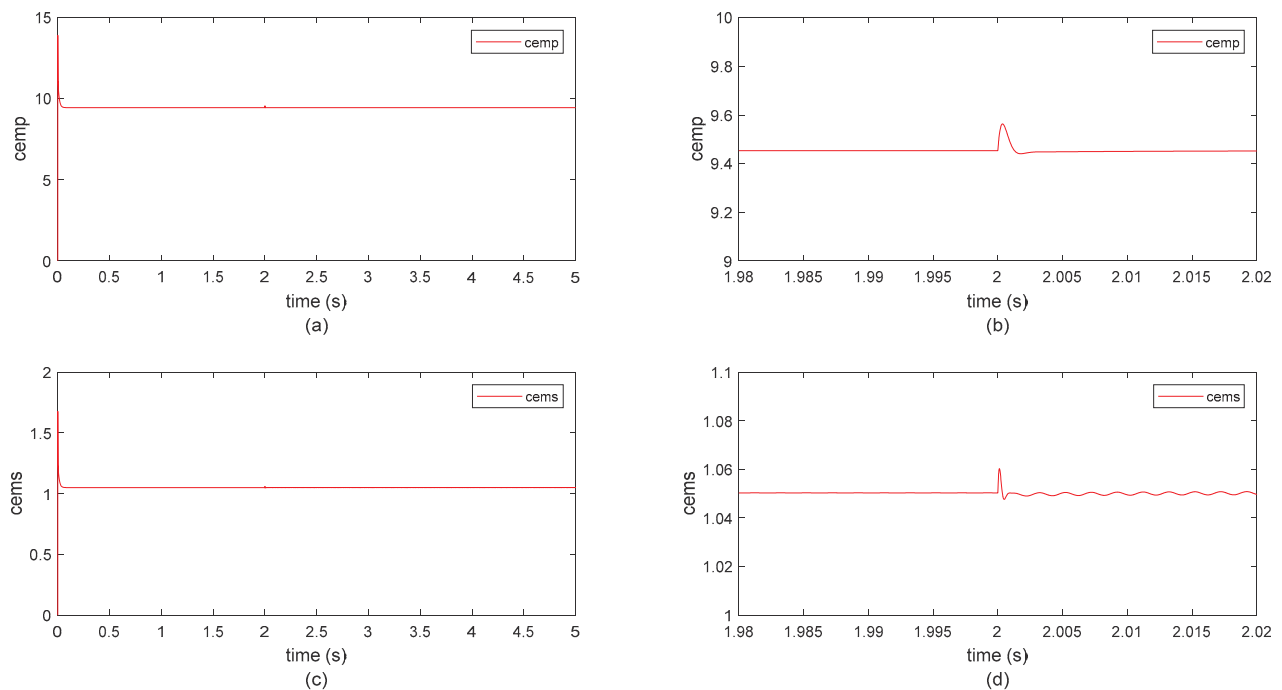


Figure 7. The electromagnetic torques' components (c_{emp} (a,b) and c_{ems} (c,d)) in the presence of stator faults with the optimization of PI controller parameters using the GWO algorithm.

The proposed control using GWO has a turnaround time of 35.3 μ s in Matlab, with 21.5 μ s used for implementing the control. In comparison, using PSO, the turnaround time is 56.3 μ s, with 39.5 μ s used for implementing the control. Additionally, the proposed method has minimal impact on the treatment time and central processing unit (CPU) utilization.

Figure 8 shows that the rotor speed remains in a steady state. However, at 2 s, the stator resistance changes from R_s to $R_s/2$ due to stator faults. The implementation of a PI controller with GWO Control is more robust and results in a higher performance than a PI controller utilizing PSO.

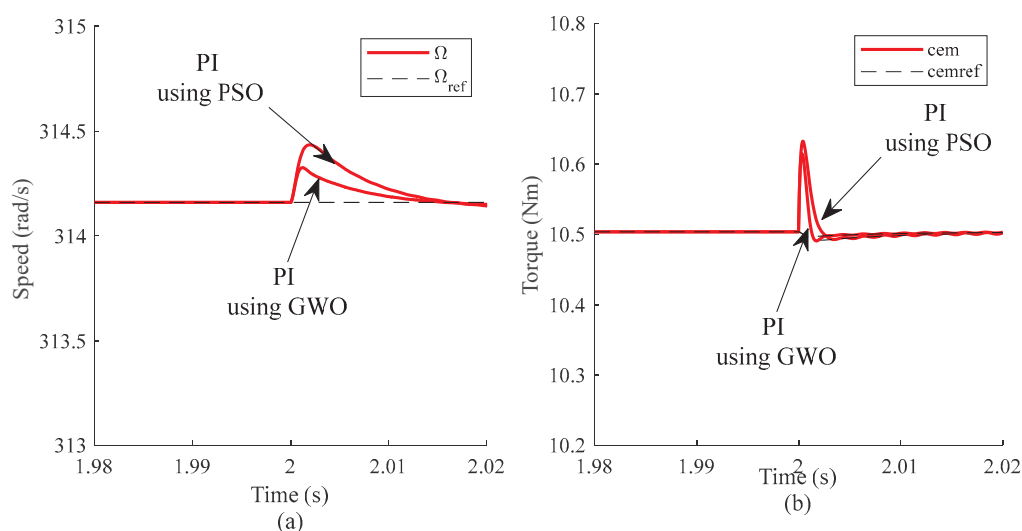


Figure 8. (a,b) Comparative study with a PI using GWO and PI using PSO.

From Table 2, the rotor speed is in a steady state, and the stator resistance is changed from R_s to $R_s/2$, $R_s/4$, and $R_s/10$ (stator faults) at 2 s. It can be asserted that utilizing the PI with GWO Control yields a high-performance control system that is more resilient than a conventional PI controller and PI using PSO (Setting time (T_{Se}) and Maximum Peak Overshoot (MPO)). To summarize the improvement in low-speed conditions, the rotor speed and electromagnetic torque of a five-phase induction motor with PI control depends on various factors such as the operating conditions, motor parameters, and control strategy used. When combined, the five-phase induction motor with PI control and GWO optimization can provide precise speed control and efficient operation of the motor, making it suitable for a wide range of applications.

Table 2. Comparative study with a classical PI controller, PI using GWO, and PI using PSO.

Stator Resistance	$R_s/2$			$R_s/4$			$R_s/10$		
Method	PI	PI using GWO	PI using PSO	PI	PI using GWO	PI using PSO	PI	PI using GWO	PI using PSO
Speed									
Setting time	0.2	0.04	0.09	0.24	0.07	0.101	0.32	0.078	0.103
Maximum Peak Overshoot	1%	0.02%	0.056%	2.5%	0.05%	0.076%	3.5%	0.066%	0.098%
Relatives' tracking error	6×10^{-3}	1×10^{-5}	4×10^{-4}	13×10^{-2}	2×10^{-5}	6.7×10^{-4}	5	2.5×10^{-5}	6.7×10^{-4}

7. Conclusions

In this paper, we evaluated the performance of the optimized PI controller by analyzing the output data and comparing it to the desired setpoint. If the speed of the five-phase asynchronous motor is regulated within an acceptable range and the ripples are minimized, then the PI controller parameters can be considered optimized for this application. If not,

the objective function may need to be adjusted. The proposed approach exploits the rapid optimization process of the GW technique under real-time working conditions to determine the optimal voltage vector that ensures the desired tracking speed at high torque and with less ripple. The mathematical principle of the proposed GWO controller is detailed, and the simulation results show that the GWO algorithm offers the best convergence towards optimal solutions with the best quality. Moreover, GWO shows a good balance between exploration and exploitation, which makes it a better algorithm to solve complex engineering optimization problems. To gain further insights into the effectiveness of the PI controller using GWO under real-time working conditions, it would be beneficial to conduct additional comparisons with advanced control methods such as intelligence-based algorithms. Additionally, the GWO algorithm could be extended to solve speed or torque-tracking tasks in other five-phase IM drives where low-order harmonics are undesired and traditional control algorithms are ineffective.

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Article

A Hierarchical Multi-Stage Coordination of Inverters for Voltage Control in Active Distribution Networks with a μ PMU-PDC

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Abstract: This paper proposes a hierarchical multi-stage approach based on a distributed level phasor measurement unit (μ PMU) at local controllers and a phasor data concentrator (PDC) at the central control unit to restore system voltage when it exceeds the limits recommended by the IEEE 1547-2018 standard. The proposed algorithm does not need an accurate system model or employ optimization solutions. Therefore, it has less implementation complexity and would be popular among distribution network service providers (DNSPs) and distribution network operators (DNOs) as it does not suffer from cost and computational complexity limitations. A PMU-PDC-based communication platform has been developed as a more efficient alternative to the supervisory control and data acquisition (SCADA) system, and provides superior characteristics, including a higher sample rate, higher data resolutions, and faster communication. The proposed coordinated algorithm aims to postpone power generation curtailment in distributed energy resources (DERs) with overvoltage problems (local DERs) by incorporating all the DERs that are not subjected to voltage violation (remote DERs) by absorbing their maximum reactive power capacity. If the system voltage has not recovered after absorbing all of the reactive power capacity of all the DERs, a reduced active power curtailment proposed by the algorithm is then applied to the system to control the voltage. The proposed strategy has been simulated in MATLAB and applied to IEEE 13-bus and IEEE 33-bus radial distribution benchmark systems to validate the performance of the system, in terms of its ability to coordinate voltage control and the accuracy of the PMU-PDC-based communication interface.

Keywords: voltage regulation; active distribution networks; phasor measurement unit (PMU); phasor data concentrator (PDC)

1. Introduction

Electrical distribution networks are undergoing a paradigm shift with the rapid deployment of distribution energy resources (DERs), enabling end-users to participate as prosumers in active distribution networks (ADNs). However, with the large integration of DERs and the consequent reverse power flow, voltage regulation has emerged as a major concern for distribution network operators (DNOs), due to voltage violations occurring outside the permitted range (e.g., $0.95 \text{ pu} < v_{bus} < 1.05 \text{ pu}$, recommended by ANSIC84.1-2016 [1] and IEEE 1547-2018 [2] standards). In response, various functions have been integrated into the inverters of DERs, enabling them to function as smart inverters (SIs) and ensure acceptable power quality and reliability of supply [3]. However, these functions typically operate based on local information/measurements and respond individually. To achieve more efficient voltage regulation in ADNs, coordination among the inverters is necessary to provide enhanced visibility and better control over the voltage profile across the distribution network.

However, to fully realize the benefits of coordinated voltage regulation, two technical challenges must be taken into consideration. The first is the design and implementation of an effective control algorithm that can coordinate the operation of smart inverters. The second is the development of a robust communication system that can reliably transmit information between the local controllers and a central control unit.

In the area of control algorithms, there have been several research studies on the coordinated operation of DERs with other voltage regulation devices, such as on-load tap changers (OLTC) and capacitor banks, to enhance the flexibility of a voltage regulation system [4]. One such approach is a two-stage online algorithm proposed in [5], which employs DERs as a complementary solution to control the voltage violation problem. In the first stage, OLTCs are scheduled using a multi-time-step optimization problem, and in the second stage, DERs are incorporated to control the voltage violation problem. The authors of [6] proposed an optimal coordinated approach for OLTC and DERs supplied by battery energy storage systems (BESSs) to minimize system voltage violations, with a focus on minimizing OLTC tap switching and reducing BESS usage. In [7], a model predictive control (MPC)-based method was suggested to optimally coordinate the generation in DERs, power of the BESS, and OLTC tap position to maintain the voltage within permissible ranges. The approach proposed in [8] is a coordinated strategy based on the alternating direction method of multipliers (ADMM) and optimal control to regulate the system voltage, which is shown to achieve a 20% improvement in voltage regulation while reducing the number of OLTC taps. The authors of [9] introduced a control approach based on the micro/recursive genetic algorithm optimization method to coordinate the performance of OLTC, the capacitor bank, and DER in the ADN, aiming to achieve optimum tap changing and minimum power losses. In [10], a coordinated control method based on a multi-agent system was proposed to minimize either the voltage violation or stress on passive voltage regulator devices.

In addition, various coordination approaches have been investigated in the literature to improve voltage regulation performance in ADNs. To minimize voltage deviations or power losses, a multi-objective hierarchical Volt/Var control (VVC) strategy was introduced in [11], which optimizes reactive power set points and p-droop characteristics in both local and central control units. Similarly, in [12], a coordinated voltage regulation algorithm was proposed that utilizes optimal reactive power from photovoltaic (PV)-based DERs and optimal charging/discharging power from battery energy storage systems (BESSs) for reactive and active power control. To allocate the optimal reactive power set point in DERs, a feedback-based leader–follower consensus algorithm was used to formulate the voltage regulation problem as an optimization problem in [13]. The authors of [14] proposed a globally optimum solution to control the system through coordinated active and reactive power exchanges between the DERs and distribution networks. Additionally, to minimize the incorporation of the main grid capacity and to specify the optimum operating point in DERs, a model-free coordinated strategy based on an extremum-seeking control approach was provided in [15].

Overall, these studies highlight the importance of coordinated voltage regulation in ADNs, which can lead to a more efficient control performance and better utilization of DERs. However, through a comprehensive review of the existing literature, the authors of this study have noticed that most current approaches to control and optimize power distribution systems require accurate models and complex optimization techniques. While these methods are theoretically sound, they can be cost-prohibitive and computationally burdensome for distribution network supply providers and distribution network operators (DNSPs and DNOs). To address these limitations, this paper proposes a novel algorithm that eliminates the need for complex optimization solutions and accurate system models. By doing so, the algorithm significantly reduces both the mathematical burden and overall cost, making it a more practical solution for DNSPs and DNOs.

In addition to control algorithms, communication systems play a critical role in voltage regulation in ADNs. Most utilities have traditionally used the supervisory control

and data acquisition system (SCADA) in the distribution networks for monitoring and control purposes [16]. The SCADA system captures unsynchronized voltage and current magnitudes, as well as active and reactive power readings from conventional measurement devices [17,18]. However, the low resolution of data and the inability to record the dynamic performance of the network are considered to be significant limitations of the SCADA-based communication system [19]. These limitations can lead to inaccurate system analysis and control, which can result in power quality issues, voltage fluctuations, and other operational problems.

On the other hand, the deployment of phasor measurement units (PMUs) that report to phasor data concentrators (PDCs), with a high time-stamp resolution, can provide system visibility enhancement much greater than that provided by SCADA. A distributed-level phasor measurement unit (D-PMU or μ PMU) has been developed as a more effective alternative to the SCADA system in the distribution network and provides superior characteristics, including a higher sample rate (maximum 100 and 120 frames per second for a 50 Hz and 60 Hz nominal system frequency, respectively, recommended by IEEE C37.118.1 standard [20]), higher data resolutions, fast communication, accurate synchronization achieved by utilizing global positioning system (GPS) antenna, and the capability to support a large number of devices [21,22]. Several research studies, such as [23–26], have reported on μ PMU application in distribution networks, in terms of monitoring, diagnostics, and control systems. In [27] and [25], μ PMU was represented as a system that could be incorporated into the planning and modeling of a system with a high penetration of DERs, as well as reconfiguration, power restoration, and island operation.

To ensure the effective coordination among the DERs in ADNs, a communication system based on μ PMUs and phasor data concentrators (PDCs) has emerged as a promising solution. The μ PMUs provide high-resolution, time-synchronized measurements of the system's voltage and current phasors, while the PDCs act as centralized data aggregators that collect and process the μ PMU data. The communication protocol used between the μ PMUs and PDCs is specified in the IEEE Std. C37.118.2 [28]. By leveraging this robust communication interface, the coordinated performance of DERs can be improved to enhance the overall stability and efficiency of the ADN.

Although there have been some papers published on the coordinated performance of DERs in ADN using μ PMU and PDC communication, the existing literature remains limited. This underscores the need to further develop and refine coordinated strategies using a μ PMU/PDC communication platform to fully realize the potential benefits of DERs in ADN.

This paper, therefore, presents a hierarchical multi-stage algorithm for voltage regulation in distribution networks with DERs. The algorithm aims to restore the system voltage within the limits recommended by IEEE 1547-2018 while alleviating power curtailments in case of over voltages. The proposed algorithm comprises four stages, where DERs are coordinated to regulate the system voltage. In the first stage, all DERs act independently to restore the voltage at the connection point through reactive power exchange with the network. In the next three stages, the central control unit and the local controllers communicate with each other to coordinate DER functions and to make decisions to minimize power generation curtailment in the system. To ensure the feasibility of the proposed algorithm, a communication interface based on phasor measurement units (PMUs) in local controllers and a phasor data concentrator (PDC) in the central control unit is proposed and tested. It is worth noting that while the use of μ PMUs is proposed in this paper to facilitate the coordination of DERs for voltage regulation, these devices can also be deployed to improve the visibility of the distribution network, which can benefit a range of other applications beyond voltage regulation.

The main benefits of the proposed algorithm are:

- The proposed algorithm does not require an optimization solution or an accurate system model, which reduces the mathematical burden and the overall cost considerably.

- The algorithm effectively postpones the PV power generation curtailment in DERs by incorporating all the DERs in absorbing reactive power using a coordinated strategy predicted in Stage II of the algorithm.
- The algorithm manages power generation curtailment in each DER by monitoring the overall curtailment index (CI) as well as the curtailment distribution (CD) in the entire system and making decisions to improve these indices in Stage III of the algorithm.
- The algorithm ensures the maximum capacity utilization of all the DERs for voltage restoration and avoids other solutions by considering the least power generation curtailment in Stage IV of the algorithm.

Overall, these features make the proposed algorithm a cost-effective and efficient solution for DNSPs and DNOs, offering improved voltage control and reduced power generation curtailment in DERs.

In addition, a mathematical assessment of the impact of remote distributed energy resources (DERs) on the voltage control of local DERs, particularly when subjected to voltage violations, is presented in the paper. This assessment is critical to ensure the overall stability and reliability of the proposed algorithm. To validate the effectiveness of this approach, rigorous simulation case studies are conducted using the IEEE 13-bus radial distribution benchmark feeder [29]. The results demonstrate the robustness and scalability of the algorithm in managing voltage violations and optimizing the utilization of DERs.

The rest of the paper is organized as follows. Section 2 provides a brief overview of the proposed algorithm for the coordination of voltage regulation devices. Section 3 presents the details of the WSN-based communication system used for monitoring and control. Section 4 presents the simulation results to demonstrate the effectiveness of the proposed algorithm and communication system. Finally, conclusions are drawn in Section 5.

2. The Proposed Voltage Regulation Strategy

This section outlines the proposed strategy for regulating the bus voltages, which is divided into two subsections. The first subsection introduces the general framework of the proposed algorithm, while the second subsection provides a detailed explanation of the proposed coordination algorithm using a conceptual flowchart.

2.1. The General Framework of the Proposed Algorithm

The framework of the proposed algorithm consists of two main parts: a local control and a central control (PDC), as shown in Figure 1. The local control assigns the reference (set) values of the active and reactive power (p_{ref}^i and q_{ref}^i) to the i th DER's PQ controller using the related p-droop and q-droop functions, respectively. These set values are updated as needed using the approach explained in the next subsection.

The central control modifies the parameters for the p- and q-droop functions in the PDC by analyzing the data received from all the PMUs located in the covered area of the distribution network. This is accomplished by executing the proposed coordinated strategy described in the next subsection. The proposed algorithm involves adjusting the dead-band voltages ($v_D^{p,i}$, $v_D^{q,i}$) and inflection points ($v_{inf}^{p,i}$, $v_{inf}^{q,i}$) in both the p- and q-droop functions related to the i th DER, as denoted in Figure 2. These control variables are used to fine-tune the setting points for the local controllers. The maximum exchangeable reactive power in the i th DER (Q_{max}^i), as a saturated level denoted in Figure 2, is also assigned according to the permitted ranges recommended by the IEEE 1547-2018 standard, as depicted in Figure 3, by considering the available active power generation (P_{ava}^i) in each DER.

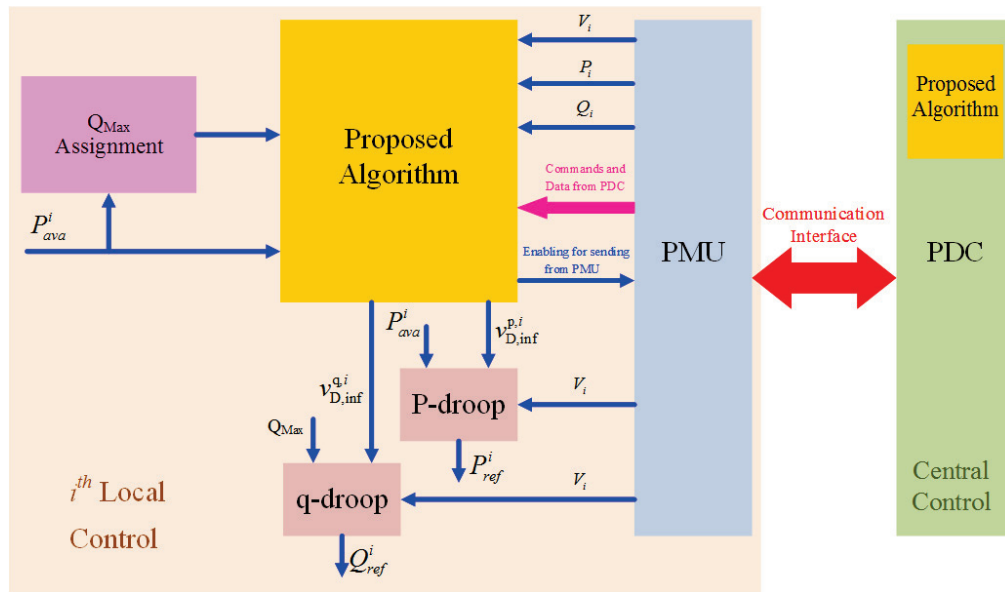


Figure 1. The general framework of the proposed control strategy.

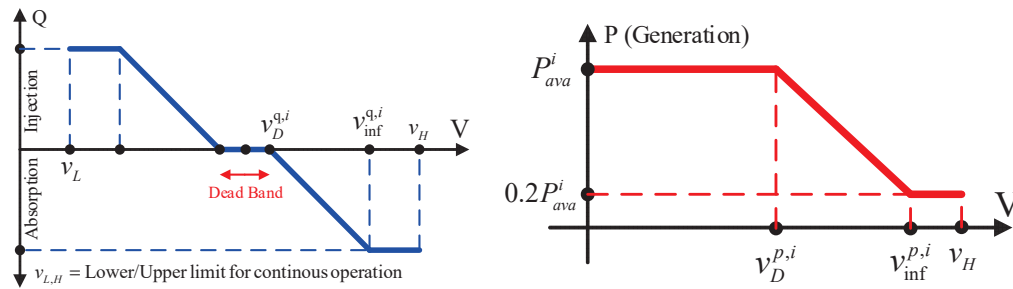


Figure 2. p-droop and q-droop functions associated with the i th DER.

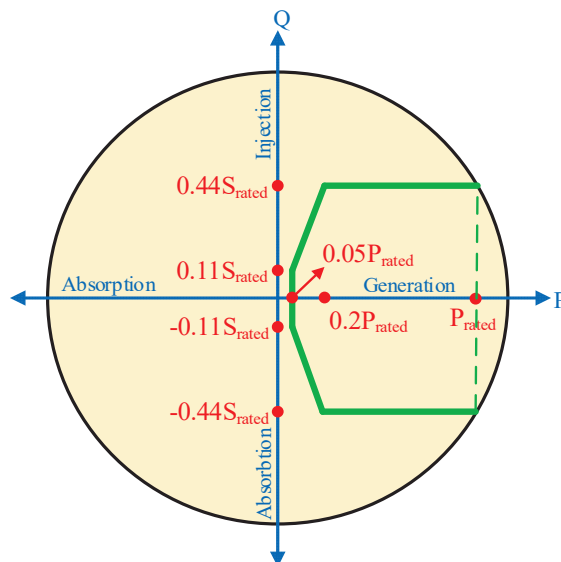


Figure 3. Permitted ranges for exchangeable reactive power recommended by the IEEE 1547-2018 standard (the green solid lines represent the boundaries for the reactive power limit).

While the measurements communicated to the PDC from the DERs include the magnitude (rms) values and phase angles of both the voltage and current at the buses, only the rms value of the voltage is used. The modified droop parameters calculated in the

local or central controllers are assigned to the droop functions, as elaborated in the following subsection. The communication network for the proposed algorithm is discussed in Section 3.

2.2. DERs Coordination Algorithm

To maintain the system voltage within acceptable limits and enhance power generation from the DERs, a four-stage coordination algorithm is proposed. This algorithm enables local and central control units to make measurements and decisions based on voltage violations, with the ultimate goal of regulating voltage while enhancing DER power generation. The algorithm is illustrated in a conceptual flowchart shown in Figure 4, where the measurements and decisions are performed by both local and central control units denoted by L and C, respectively.

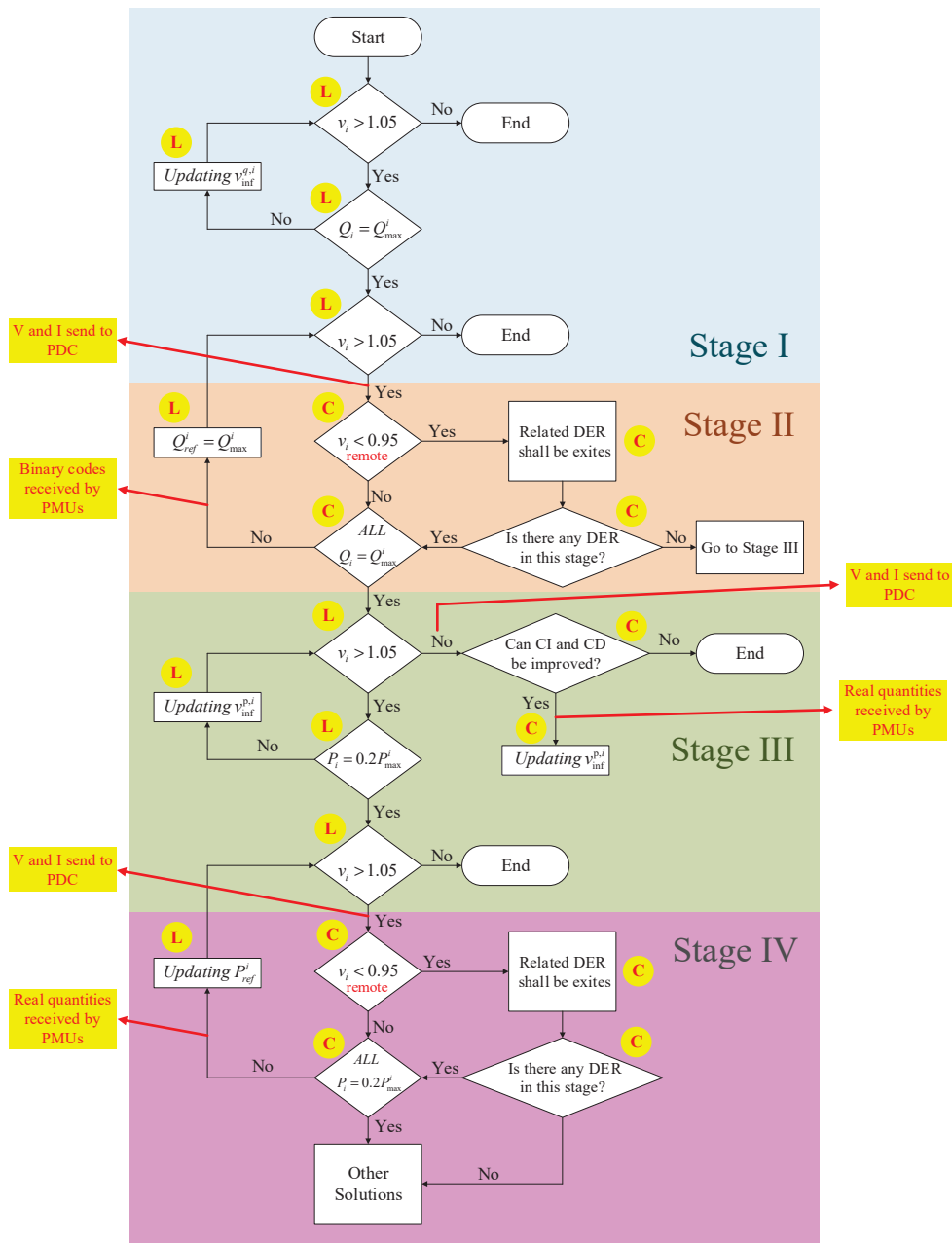


Figure 4. A conceptual flowchart for the proposed control algorithm.

In Stage I, DERs that experience exceeded voltage limits independently regulate the voltage at the point of connection (POC) by absorbing reactive power. The dead-band voltage ($v_D^{q,i}$) is maintained constant at 1.05 pu, which is the maximum permitted value recommended by the ANSIC84.1-2016 and IEEE 1547-2018 standards. The inflection points in the q-droop functions ($v_{inf}^{q,i}$) are updated through interpolation, with a starting value of 1.1 pu as an upper permitted voltage in a continuous range. If local controllers cannot regulate the system voltage in Stage I (when the reactive power of the DER reaches Q_{max}^i), the DERs that are not experiencing voltage violations are designated as remote DERs, and their specifications are sent to the PDC.

In Stage II, if the local controllers cannot regulate the system voltage through the reactive power exchange in Stage I, remote DERs are enlisted in the PDC to assist in voltage recovery. All the q-droops of the remote DERs are set to absorb the maximum capacity of reactive power (Q_{max}^i). However, as the voltage at the remote DERs starts to decrease due to reactive power absorption, it is necessary to ensure that their voltage does not drop below 0.95 pu. Therefore, a complementary loop is added to Stage II to check the lower limit of the voltage after checking the upper limit, as shown in Figure 4. If the voltage drops below the lower limit, the relevant DER is excluded from the central controller and returned to the local control unit to regulate the voltage. This ensures that only DERs with voltages above the lower limit are allowed to participate in voltage recovery. One of the advantages of this stage is that it postpones power curtailment by using the maximum capacity of the reactive power of all the DERs for regulating the system voltage.

In the event that the voltage remains unregulated at the end of Stage II, the algorithm proceeds to Stage III, in which any DER experiencing high voltage levels enters the p-droop mode. This stage involves the local controls of the DERs for curtailing power generation at the POC to regulate the voltage in a timely and effective manner. To achieve this, the active power reference values are updated continuously by adjusting the inflection points ($v_{inf}^{p,i}$) in the p-droop function. Starting with a value of 1.1 pu, the value of $v_{inf}^{p,i}$ is reduced in steps, while maintaining the dead-band voltage ($v_D^{p,i}$) constant at 1.05 pu. Figure 5 provides a visual representation of the p-droop function for various values of $v_{inf}^{p,i}$.

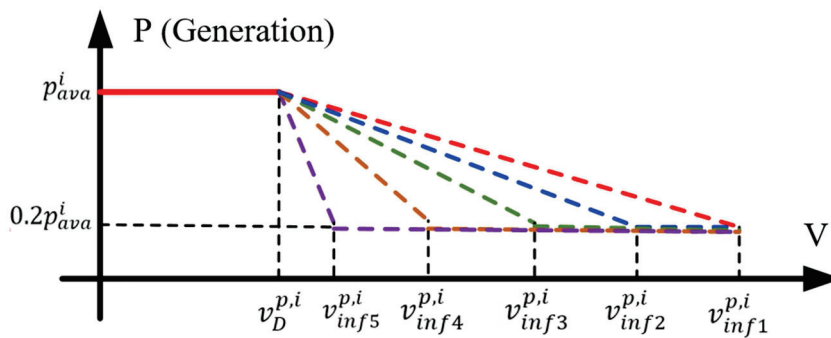


Figure 5. The p-droop function for different values of $v_{inf}^{p,i}$.

To enhance the efficiency of the system in terms of active power curtailment, Stage III includes an evaluation of the curtailment index (CI) and curtailment dispatch (CD) provided by (1) and (2), respectively. Note that lower values for CI and CD indicate better system conditions. To this end, an additional condition has been added to the flowchart after voltage recovery, as shown in Figure 4. In this section, the central controller evaluates other possibilities for curtailment to identify the most efficient option in terms of CI and CD. If a more efficient alternative is detected, a modified $v_{inf}^{p,i}$ value is assigned and sent to the local controllers before ending the flowchart in Stage III. By performing this evaluation,

the system can achieve a lower active power curtailment, leading to improved system performance and stability.

$$CI = \frac{\text{curtailed power}}{\text{available power}} = \frac{\sum_i P_{cur}^i}{\sum_i P_{ava}^i} \quad (1)$$

$$CD = \frac{CI}{N} \quad (2)$$

where n is the number of DERs in the covered area of the distribution network of the feeder.

In the fourth stage, if the voltage is still not regulated, the DERs that are not affected by the voltage violation are designated as remote DERs for power curtailment to lower the voltage in the local DERs that are still experiencing high voltage levels. To limit the amount of curtailment required, it is advisable to gradually decrease the power generation of the remote DERs by adjusting the active power reference in multiple stages. This involves using several reference active power values (p_{ref}^i) instead of starting at the minimum value ($0.2p_{ava}^i$, as per the standard). In this way, the curtailment is reduced, and the system experiences less power interruptions. The number of steps for remote curtailment per DER should be limited (up to a maximum of four steps in this study).

2.3. Assessment of the Impact of Remote DERs on Voltage Control in Local DERs

Since the proposed algorithm is based on the coordinated performance of DERs in system voltage control, it would be much more beneficial if the impact of the remote DERs on the voltage regulation at the local bus was mathematically formulized to increase the technical support of the proposed control strategy. To this end, the IEEE 13-bus radial distribution feeder (IEEE-13 feeder) was simplified, as illustrated in Figure 6. In this system, all the left side loads at B1 are modeled as P_{L1} and Q_{L1} . In addition, the left side loads at B2 are represented as PV in the active and reactive injection modes to model a voltage increase at B2. Three DERs are assumed to be installed at B2 as a local bus, and at B3 and B5 as remote buses. It is assumed that the operating condition is in Stage II of the algorithm and the remote DERs are incorporated to regulate the voltage at the local bus through only reactive power absorption.

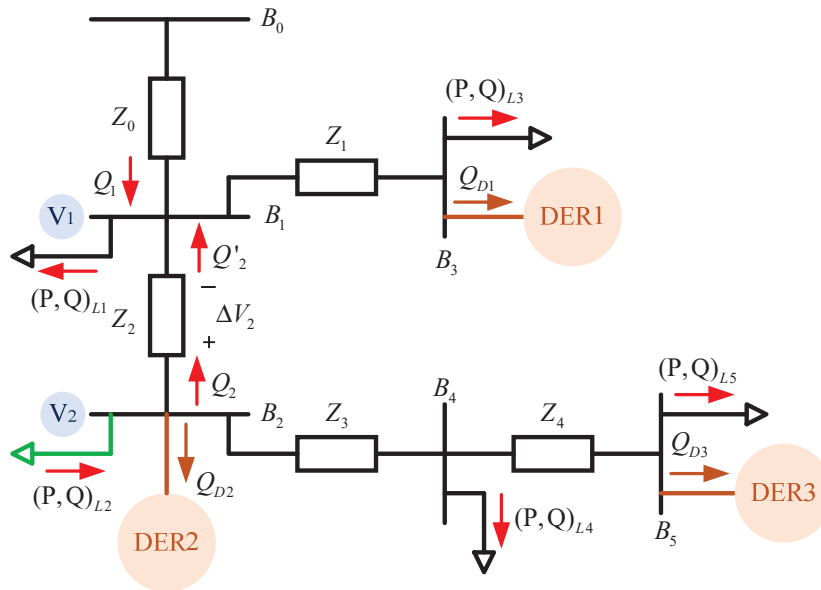


Figure 6. A simplified model of the IEEE 13-bus radial distribution feeder (IEEE-13 feeder).

By considering ΔV_2 as a voltage drop across Z_2 and ignoring the impact of the phase difference between V_1 and V_2 on voltage regulation, the relationship between the voltage at B1 and B2 can be explained as (3).

$$V_2 = V_1 + \Delta V_2 \quad , \quad \Delta V_2 = \sqrt{\frac{Z_2^2}{X_2} (Q_2 - Q'_2)} \quad (3)$$

where Q_2 and Q'_2 are the sending reactive power from B_2 and the receiving reactive power at bus B_1 , respectively, which can be expressed by (4) and (5), respectively.

$$Q_2 = Q_{L2} - Q_{L4} - Q_{L5} - Q_{D2} - Q_{D3} \quad (4)$$

$$Q'_2 = Q_{L1} + Q_{L3} - Q_1 + Q_{D1} \quad (5)$$

Q_{D1} , Q_{D2} , and Q_{D3} represent the reactive power absorbed by DER1, DER2, and DER3, respectively. The reactive power exchanged through all the loads can be assumed to be constant values during voltage regulation intervals. Moreover, since the reactive power absorbed by the local DER (Q_{D2} in this case) is set at the maximum value ($Q_{D2}^{\max}$) in Stage II, it can be assumed to be a constant value as well. Therefore, constants k_1 and k_2 can be defined as $k_1 = Q_{L2} - Q_{L4} - Q_{L5} - Q_{D2}$ and $k_2 = Q_{L1} + Q_{L3}$ and (3) is rewritten as (6).

$$V_2 = V_1 + \sqrt{\frac{Z_3^2}{X_3} (k_1 - k_2 + Q_1 - Q_{D1} - Q_{D3})} \quad (6)$$

By considering $k_3 = k_1 - k_2$, and assuming 1.05 pu as the maximum permitted voltage in an overvoltage area as recommended by IEEE1547, (6) is expressed as an inequity provided by (7).

$$V_2 = V_1 + \sqrt{\frac{Z_3^2}{X_3} (k_3 + Q_1 - Q_{D1} - Q_{D3})} < 1.05 \quad (7)$$

Figure 7 presents a 3D plot of (7) for five different values of V_1 . The surface plots show the relationship between the absorbed reactive power by DER1 and DER3, as well as the reactive power available at bus B_1 (Q_1). The plots were generated in response to a voltage rise of 1.05 pu at bus B_2 (the local bus) to identify the layers of the operating points. The operating points above the surface indicate a voltage level greater than 1.05 pu, while the operating points below the surface represent a voltage level less than 1.05 pu at B_2 .

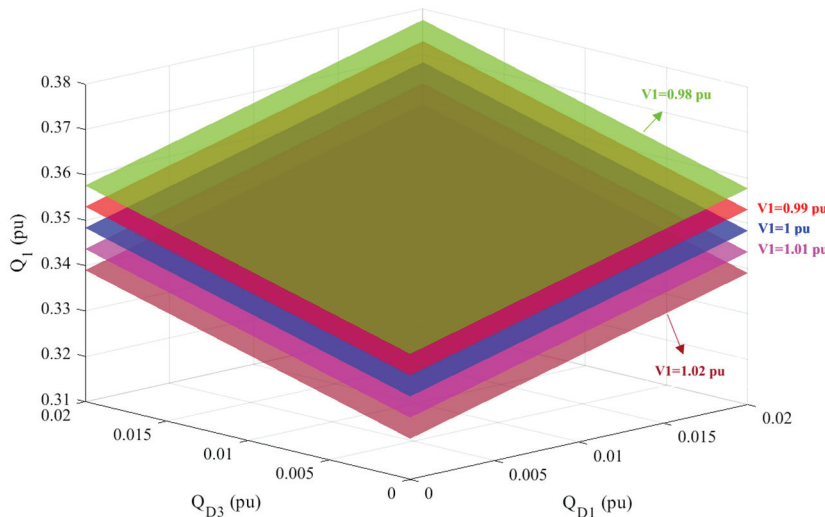


Figure 7. A 3D plot of (7) for five different values of V_1 .

To validate the analysis, the system shown in Figure 6 was simulated using MATLAB (R2022a), and the voltage at bus B₂ was measured for five different scenarios, including no compensation, compensation using only DER2, and compensation by deploying remote DERs with operating points determined by the values across/over/under the surface, as depicted in Figure 7. As seen in Figure 8, the voltage at B₂ was found to increase to 1.067 pu in the overvoltage case and decrease to 1.054 pu in Stage I. In Stage II, the remote DERs (DER1 and DER3) were integrated to regulate the voltage at the local bus (B₂). When the absorbed reactive power by the remote DERs was set at values from across the surface provided in Figure 7, such as $Q_{D1} = Q_{D3} = 0.02$ pu for $V_1 = 1$ pu, the voltage at B₂ was almost equal to 1.05 pu, as expected from Equation (7). A slight deviation was observed due to a steady-state error in the injected active and reactive power by the DER at B₂. Figure 8 indicates that if the absorbed reactive power by the remote DERs is assigned over or under the surface, the voltage at the local bus will be lower or higher than 1.05 pu, respectively, which validates the accuracy of the analysis provided in this section.

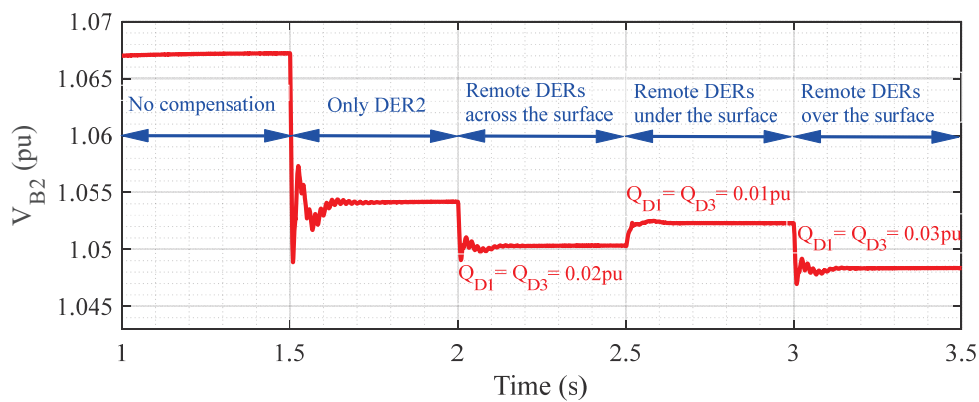


Figure 8. Voltage at B₂ for five different conditions.

3. Description of Proposed Communication System

The proposed communication system topology is based on a distributed level phasor measurement unit (D-PMU or μ PMU) and phasor data concentrator (PDC), as shown in Figure 1. This system leverages the electrical quantities measured by PMUs, which are either used in local controllers or transmitted to the PDC by a central control unit. The updated values of the parameters and/or control commands generated in the PDC are then sent to the local controllers of the DERs. The PMU reports four fundamental measurements in three phases, resulting in a total of 12 measurement channels. These measurements include voltage magnitude, voltage phase angle, current magnitude, and current phase angle per phase, with a maximum sampling rate of 100 and 120 frames per second for a 50 Hz and 60 Hz nominal system frequency, respectively, as recommended by the IEEE C37.118.1 standard. The μ PMU also utilizes a GPS clock to ensure precise time synchronization, which is achieved once the GPS antenna establishes a connection with satellites. This synchronization is achieved by the same-time sampling of voltage and current phasors using timing signals from the GPS receiver. With an accuracy angle of 0.01, a total vector error allowance of 0.05% (precision), an angle resolution of 0.002, and a magnitude resolution of 0.0002%, the μ PMU ensures reliable and accurate measurement. Its accuracy and precision are typically 0.01% and 0.05%, respectively, as per the standard [16].

The PDC is a device designed to receive measurement data packets from the PMUs using a specific protocol, such as TCP/IP packets with the C37.118.2 payload data format. The maximum wait time for a PDC to record phasor measurements from all the PMUs is 14 s. If the phasor data is received within this maximum wait time, the output data is created immediately; otherwise, the PDC waits until the time limit expires to generate the output data. The PDCs aggregate data and send it to downstream consumers in a format defined by IEEE Std. C37.118.2. Communication technologies in PMU-PDC-based systems can be broadly classified into two groups: wired and wireless. Wired communication

technologies offer high reliability, large bandwidth, and protection against interference, while wireless technologies provide advantages such as rapid deployment, low installation and maintenance costs, and access to remote locations [30].

Communication in the proposed algorithm uses wireless technology and can be visualized as a loop that includes PMUs, PDCs, and data transmission, as illustrated in Figure 9. The communication data and channels between the PMUs and the PDC(s) vary based on the stage of operation in the coordination algorithm, which are described below.

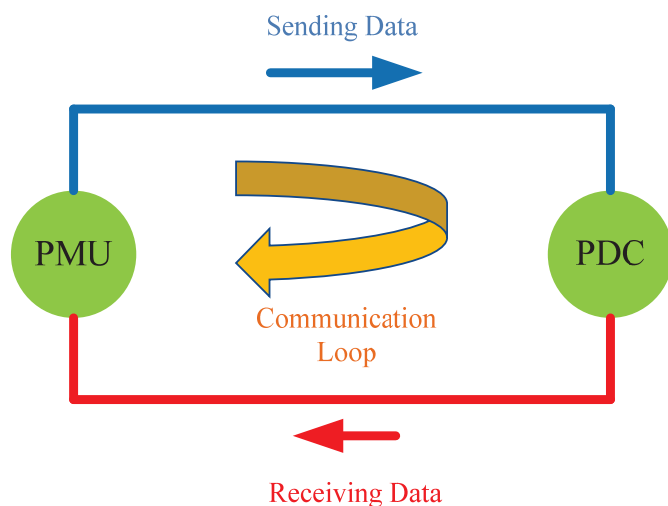


Figure 9. A general view of the proposed communication system.

In Stage I, the PMUs measure and transmit the usual voltage magnitude, voltage phase angle, current magnitude, and current phase angle per phase. In Stage II, the PDC uses the measured voltage magnitude to generate channels with the logical commands 0 (local) or 1 (remote), which includes two channels per PMU. Typically, a maximum of $n-1$ PMUs are involved in this stage, in a system with n PMUs. Stage III involves 12 channels for measuring the voltage magnitude, voltage phase angle, current magnitude, and current phase angle per phase of all the DERs, and the communication data in this stage include real quantities sent from the PDC and received by all the DERs for updating control parameters. Similarly, in Stage IV, there are 12 channels for voltage and current phasors, as well as an additional channel per DER for the real quantities sent by the PDC and received by the specified local controllers. Table 1 lists the specifications for these stages based on the test system used in this study, which includes three DERs.

Table 1. The specifications for each stage involved in communication.

Stage Number	Sending Data		Receiving Data		Number of Communication Loops
	Maximum Number of PMUs	Data Type	Maximum Number of PMUs	Data Type	
II	2	Voltage and current (Mag. and phase)	2	One-bit binary code	1
III	3	Voltage and current (Mag. and phase)	3	Real quantities	1
IV	2	Voltage and current (Mag. and phase)	2	Real quantities	Maximum 4

4. Evaluating the Results and Discussion

In order to demonstrate the effectiveness of the proposed approach for the coordinated performance of all DERs in voltage regulation, the benchmark IEEE 13-bus radial

distribution feeder (Figure 10) has been modeled in MATLAB. In this system, buses 634, 671, and 675 are assumed as area 1, area 2, and area 3, and selected for connecting DER1, DER2, and DER3, respectively, as illustrated in Figure 10. The performance of the system in terms of the ability of the proposed algorithm to coordinate voltage regulation as well as the accuracy of the communications interface are represented in two separate subsections.

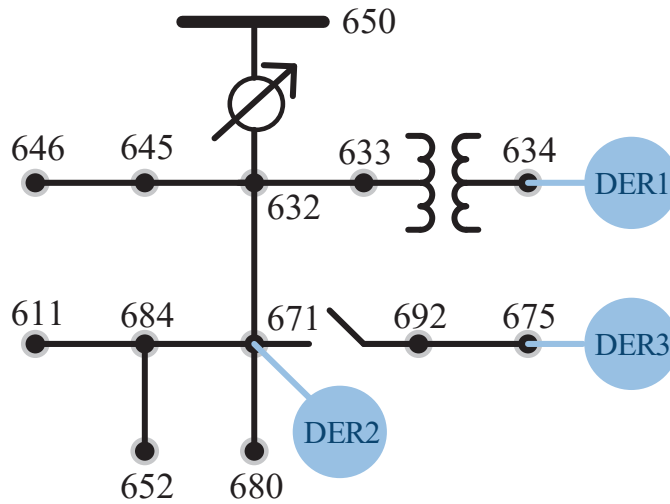


Figure 10. The studied system.

4.1. Ability of the Proposed Algorithm to Coordinate Voltage Control

Three different scenarios have been defined in this subsection to verify the performance and merits of the proposed algorithm.

4.1.1. Scenario 1

In this simulation scenario, an initial voltage increase was applied to the system at $t = 0.4$ s, resulting in voltage level increases in areas 1, 2, and 3 to 1.0655 pu, 1.065 pu, and 1.06 pu, respectively. Figure 11 presents the results of employing the proposed voltage regulation strategy in all three areas. In all the figures in this section, the maximum capability of the DERs in active power and reactive power exchanges is considered as the base value for the generated active power and absorbed reactive power of the DERs, respectively.

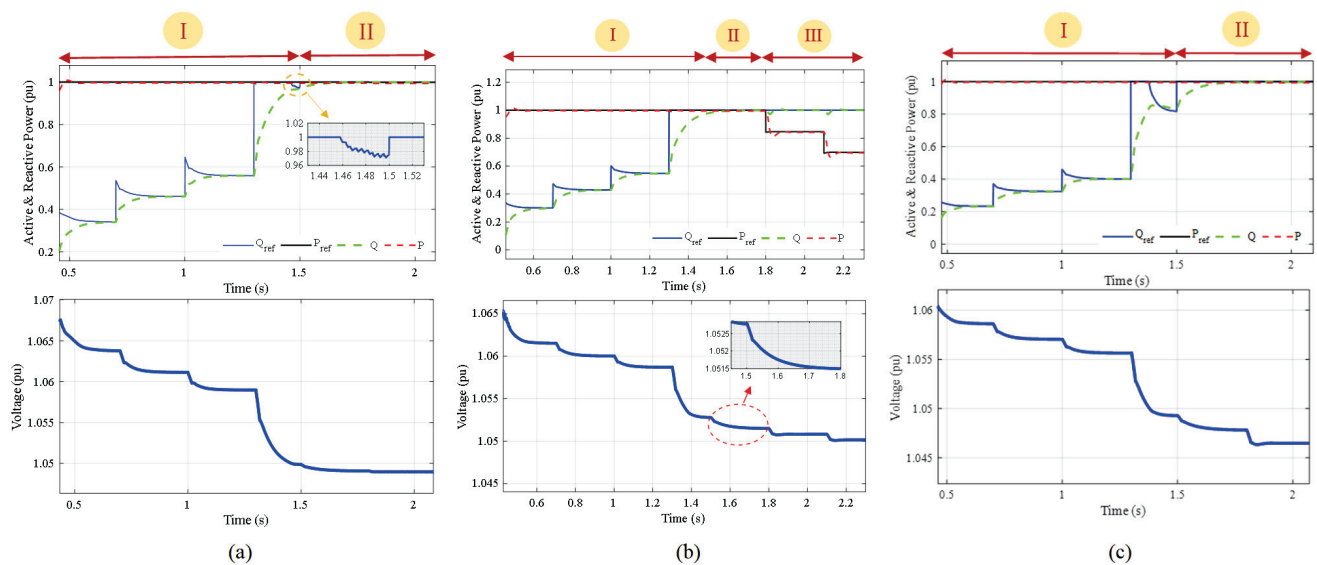


Figure 11. Voltage regulation in all three areas by employing the proposed strategy for scenario 1 in (a) area 1, (b) area 2, and (c) area 3.

Figure 11a,c show that DER1 and DER3 were able to regulate the voltage at the POC in the first stage of the algorithm, as depicted by the flowchart in Figure 4. However, Figure 11b demonstrates that DER2 was not able to control the voltage by locally exchanging reactive power. As a result, the algorithm proceeded to Stage II, and the free capacity of the reactive power of DER1 and DER3 (as depicted by the zoomed-in figures), was utilized. Under this condition, both DER1 and DER3 were listed as remote DERs and set at the maximum value as planned by the central controller. The voltage reduction resulting from the maximum reactive power absorption by DER1 and DER3 is clearly observable in Figure 11a,b. It is notable that all of the capacity of the reactive power resources was utilized before the curtailment occurred in DER2.

Despite the reduction in voltage in DER2 due to the incorporation of other DERs in Stage II, the voltage was not sufficiently reduced to the permitted value of 1.05 pu. Therefore, the algorithm proceeded to Stage III, where the voltage is controlled through local curtailment in area 2 without impacting the other remote DERs in terms of active power curtailment. As seen in Figure 11b, the voltage in area 2 approached 1.05 pu after a 0.3 pu active power curtailment was imposed on DER2 in Stage III. The figure demonstrates that if the voltage in area 2 had not been reduced in Stage II, an additional amount of power curtailment would have been required (without using the proposed algorithm) to restore the voltage to its prescribed level of 1.05 pu.

To assess the effectiveness of the proposed strategy and demonstrate its practical benefits over non-coordinated methods, the same test was repeated without implementing the proposed algorithm. Figure 12 shows the active and reactive power in all the DERs, as well as the voltage at the connection points. According to Figure 12a,c, the voltage in areas 1 and 3 were restored locally by absorbing reactive power at DER1 and DER3 within 1.1 s and less than 1 s, respectively. The results for area 2, depicted in Figure 12b, indicate that even with the maximum reactive power absorption and a minimum generation level of 0.2 pu (maximum curtailment) in DER2, the voltage could not be recovered. By comparing Figures 11 and 12, it becomes evident how the proposed coordinated approach helps to regulate the voltage in area 2. By incorporating DER1 and DER3 in absorbing reactive power at $t = 1.5$ s in Stage II of the algorithm, the voltage in area 2 was restored after only a 30% curtailment of DER2. The responses of the voltages in the three areas during this scenario are given in Figure 13, as compared to the case when only a local control is used (without coordination), which further illustrates the impact of the remote DERs in the proposed coordinated algorithm to regulate the voltage.

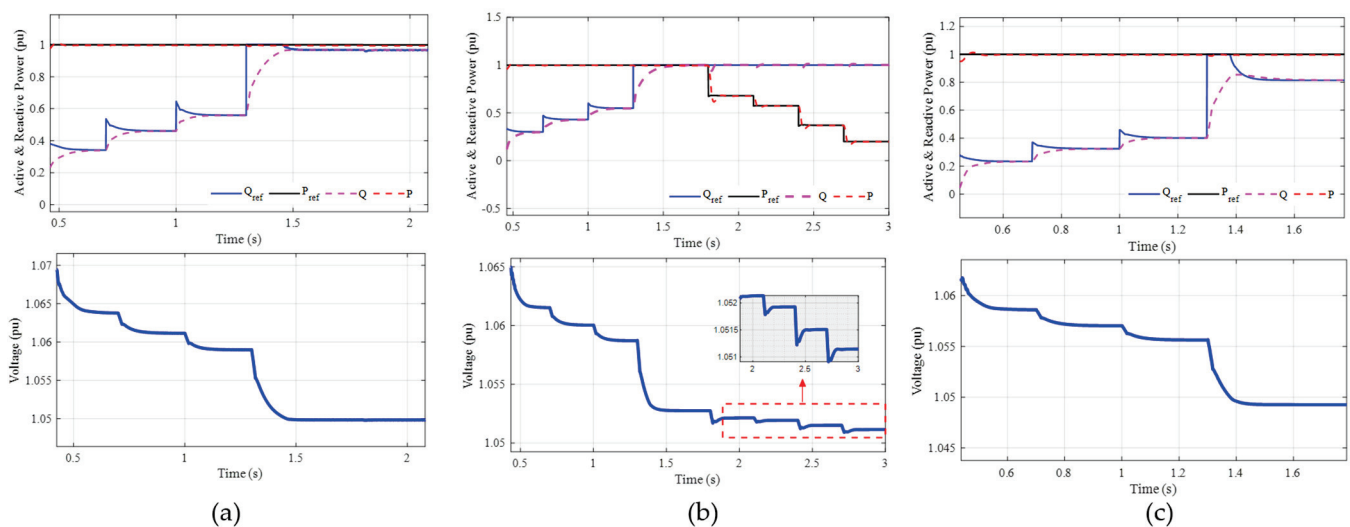


Figure 12. Voltage regulation in all three areas for scenario 1 with no coordination strategy in (a) area 1, (b) area 2, and (c) area 3.

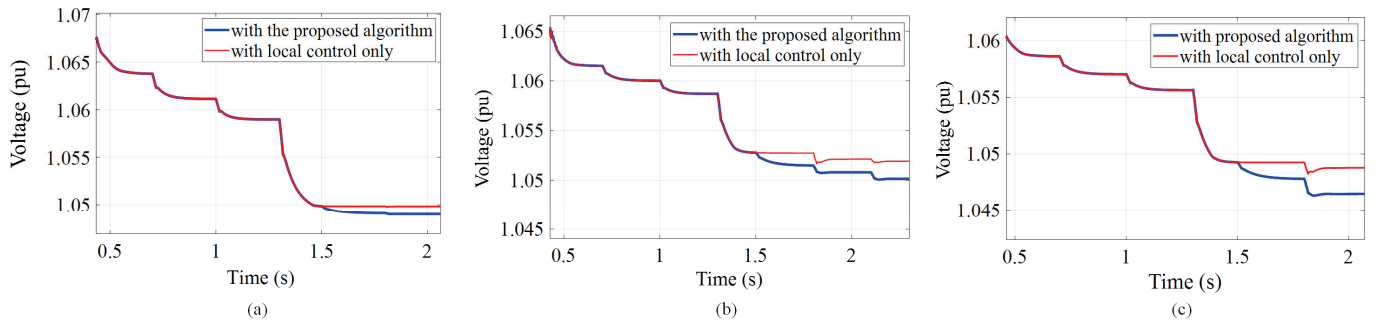


Figure 13. Comparison of the voltage response of the proposed coordinated approach with a local control only in (a) area 1, (b) area 2, and (c) area 3.

4.1.2. Scenario 2

This scenario aims to investigate the effectiveness of the supplementary condition implemented in Stage III of the proposed algorithm for improving active power curtailment in the system, as shown in Figure 4 and explained in Section 2.2. An increased voltage occurs at $t = 0.4$ s in this scenario. Figure 14 shows the active and reactive power in all three areas, as well as the voltage at the connection points when the supplementary loop is not employed in Stage III. As shown in Figure 14a, the voltage in area 1 was restored within 1.6 s after passing through three stages and imposing a 0.3 pu curtailment of DER1. In Figure 14b, it can be seen that the voltage in area 2 was controlled in Stage III within 2.2 s with a 60% curtailment of DER2. The voltage in area 3 was recovered in the first stage after less than 1 s, as depicted in Figure 14c. The impact of the central control unit in voltage regulation is highlighted in the zoomed-in Figure 14b. The curtailment index (CI) and the curtailment dispatch (CD) were calculated to be 0.3 and 0.1, respectively, based on equations (1) and (2). The system voltage in all areas was restored after 2.2 s.

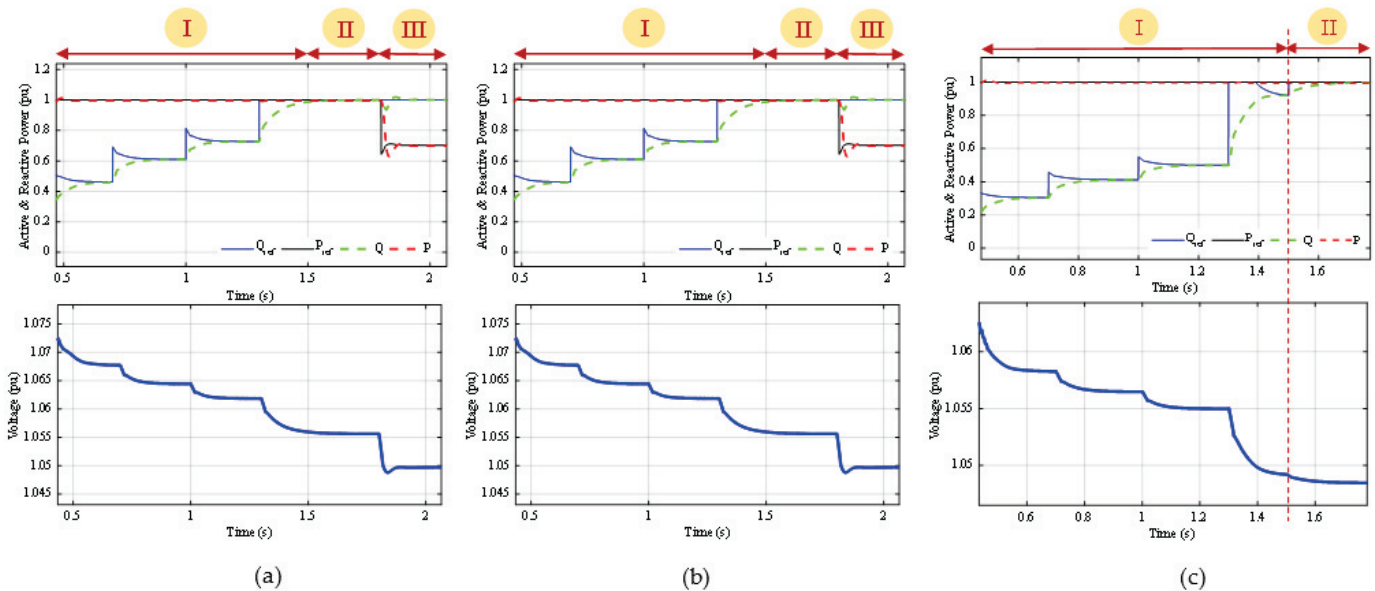


Figure 14. Voltage regulation in all three areas for scenario 2 with no supplementary loop in the Stage III strategy in (a) area 1, (b) area 2, and (c) area 3. The red vertical line highlights some significant transition times from a stage to another.

Figure 15 demonstrates the impact of incorporating the supplementary loop in Stage III. As shown in Figure 15b,c, the voltage in area 2 was restored within 1.6 s by making 0.258 pu curtailments in DER3 at $t = 1.8$ s, while suffering only a 0.3 pu curtailment of DER2. This approach results in a decrease in the curtailment index (CI) and the curtailment dispatch (CD) to 0.186 and 0.062, respectively, and a reduction in the recovery time to 1.6 s.

By rescheduling in the central controller before the algorithm finishes in Stage 3, not only is the recovery time decreased, but curtailment is also dispatched more equitably across the system with a lesser amount of CI.

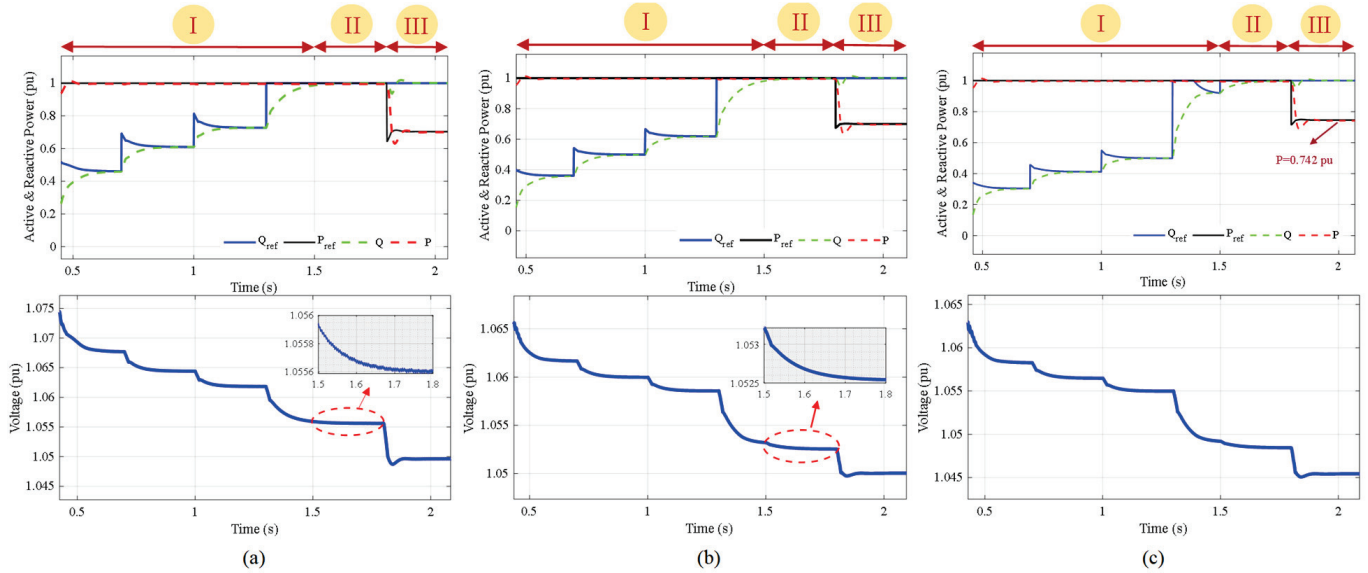


Figure 15. Voltage regulation in all three areas when the supplementary-loop strategy in Stage III is employed in (a) area 1, (b) area 2, and (c) area 3.

4.1.3. Scenario 3

The previous tests evaluated the system under conditions where all three areas experienced similar voltage increases, and all the DERs initially regulated the voltage at the POC through local controllers in Stage I. The remaining reactive power exchange capacity was utilized in Stage II to reduce the voltage rise before any curtailment occurred. This scenario, however, examines a situation where one area does not experience a voltage increase, but absorbs the maximum reactive power in Stage II to regulate the system voltage in the other areas. Figure 16 presents the results of this scenario, including the active and reactive power in all three areas, as well as the voltage at the connection points. As shown in the figure, only areas 2 and 3 experienced a voltage increase, and the voltage in area 1 remained within the permissible range. DER2 was able to regulate the voltage in area 2 at $t = 1.3$ s after passing through Stage I, as depicted in Figure 16b. In Figure 16c, it can be observed that the voltage in area 3 was not restored in Stage I, so the algorithm proceeded to Stage II, where DER1 absorbed its maximum exchangeable capacity, as shown in Figure 16a. As a result of absorbing reactive power in area 1, the voltage in area 3 decreased, as demonstrated in the zoomed-in figure. The system voltage was ultimately restored in Stage III, following a 0.5 pu curtailment of DER3 at $t = 2.4$ s. It is worth noting that the voltage in area 1 did not drop below the lower permissible limit (0.95 pu). However, in other systems with different power flows and specifications, the voltage may fall below the limit, which can be problematic. To address this issue, supplementary loops are included in Stages II and IV to exclude any potential effect on the DERs, as described in Section 2.

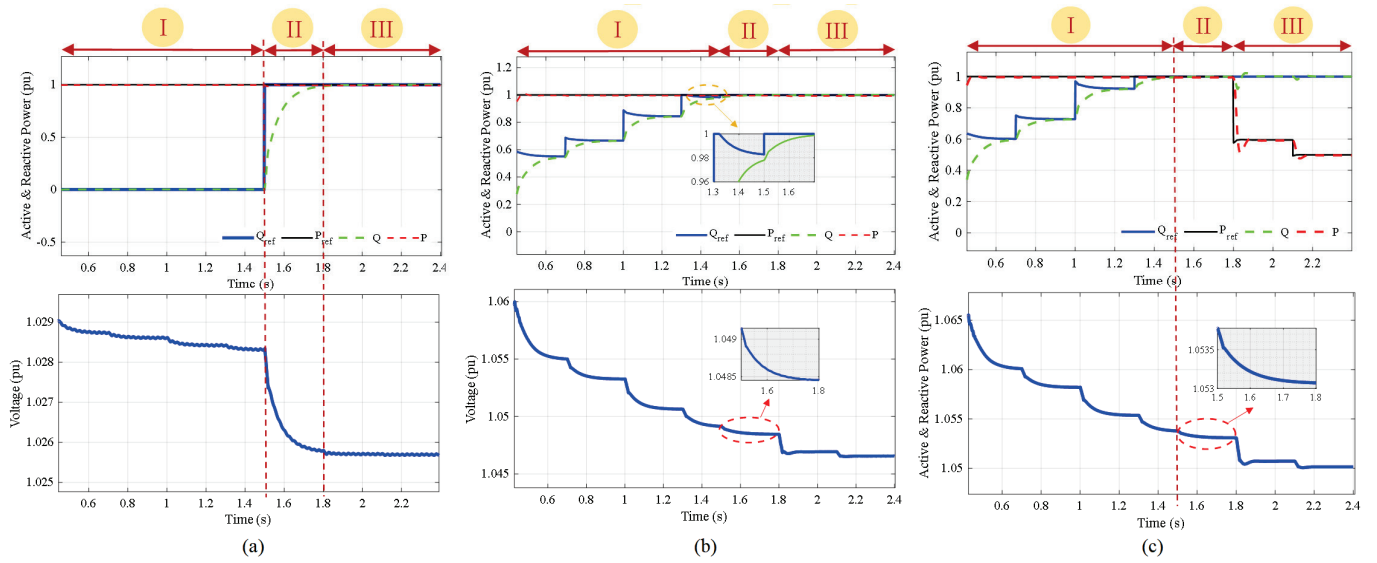


Figure 16. Voltage regulation in all three areas for scenario 3 when two DERs suffer from the increased voltage strategy in (a) area1, (b) area 2, (c) area 3. The red vertical line highlights some significant transition times from a stage to another.

4.2. Application of the Algorithm on a Larger Test System

In the previous section, a comprehensive examination of the proposed coordinated control strategy was conducted on a small-scale test system to establish its efficacy through a detailed analysis. To further evaluate its scalability, the performance of the coordinated control strategy was assessed on the IEEE 33-bus Benchmark system [31], serving as the second studied system. A single diagram of the network is presented in Figure 17.

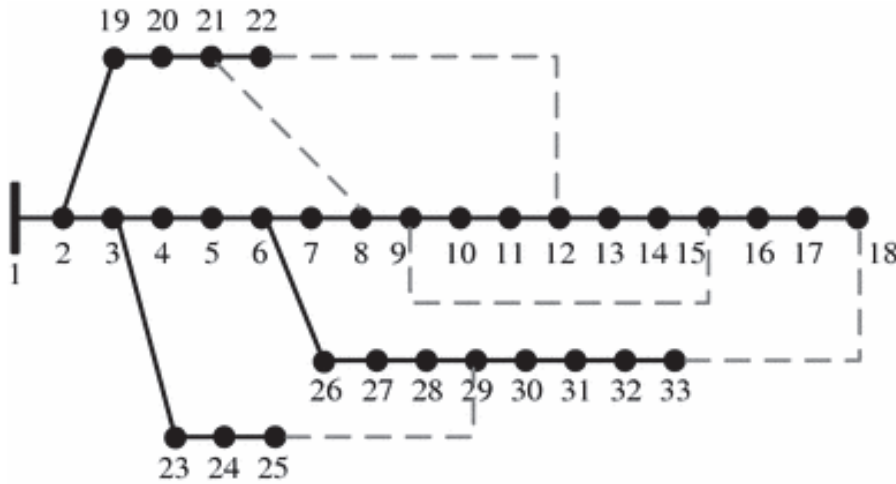


Figure 17. A single line diagram of the IEEE 33-bus system.

According to Figure 17, buses 6, 18, 21, 25, and 33 are assigned as the points of connection (POC) for five DERs. The voltage at B6, B18, B21, B25, and B33 are assumed to be increased to 1.0582, 1.584, 1.0649, 1.0537, and 1.0575, respectively. The DER connected to B21 is considered local, and the others are assumed to be remote buses. Figure 18 shows how reactive power absorption at the local DER has an impact on the remote ones.

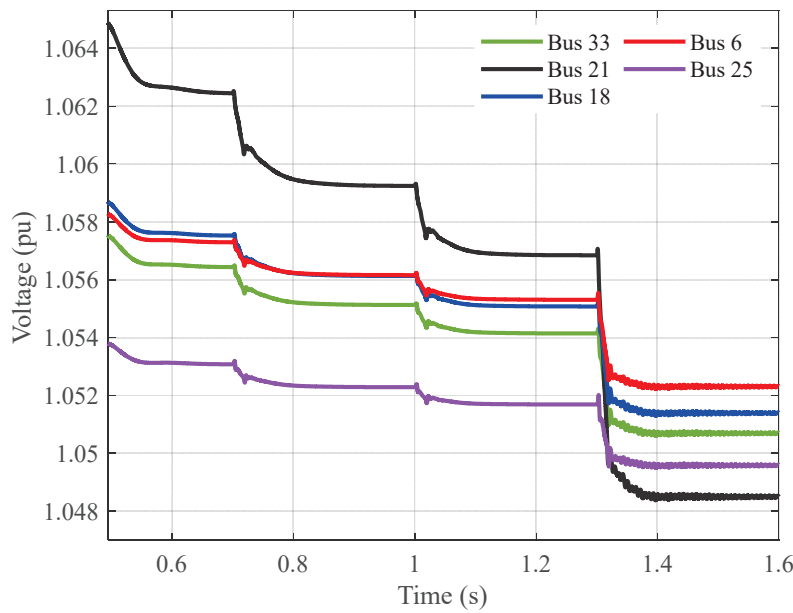


Figure 18. Voltage response of the selected DERs at the IEEE 33-bus system.

According to the figure, the local controller not only reduces the voltage at B21 to below 1.05 as the target value, but also is able to reduce the voltage at B6, B18, B25, and B33 to 1.0522, 1.513, 1.0495 (lower than 1.05), and 1.0507, respectively. The reactive power absorbed by the local DER and the reference value are illustrated in Figure 19. It is visible from the figure that the DER is able to follow the reference value accurately with no transient distortion or steady-state error. This test demonstrates that the proposed control system is able to be employed in systems with more buses and different configurations.

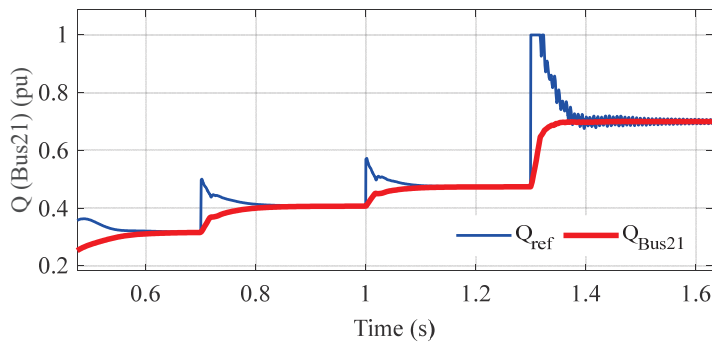


Figure 19. The reactive power absorbed by the local DER and the reference value.

It should be noted that, while there is no strict limit on the number of DERs that can be employed in the proposed algorithm, in practice, the limit depends on the capacity of the PDC to handle multiple PMUs. In general, the number of PMUs that can be connected to one PDC depends on its processing power, memory, and communication bandwidth. A PDC needs to efficiently manage the incoming data streams from multiple PMUs, process and analyze the phasor measurements, and store or transmit the data to the appropriate applications or control centers. Therefore, the capability of the PDC to handle a large number of PMUs may be limited by its technical specifications and processing capabilities.

4.3. Accuracy of the Communication System

The performance of the communication between the PMU and PDC is evaluated using the specifications for LTE (4G) technology. Each resource block is 0.5 ms long and a path loss exponent of three is considered. The path loss is modeled as $(d/d_0)^3$, where d is the

actual distance between the PMU and PDC and d_0 is the reference distance. The channel gain is inversely proportional to this path loss. The communication model considers a 4-QAM modulation. The accuracy of the communication system in the proposed algorithm is discussed in the following subsection in terms of the influence of the noise and the impact of the communication delay.

4.3.1. Influence of the Noise

The accuracy of the communication system is affected by the signal-to-noise ratio (SNR). When the SNR is low, there are an increasing number of samples which are incorrectly received, leading to potential errors in the received measurements. In this study, the impact of SNR on the accuracy of the communication system is investigated using monitoring data transmission from the PMU to PDC. Figure 20a–d show the voltage and current magnitude, as well as the voltage and current phase angles, respectively, for a certain instance of monitoring data transmission from the PMU to PDC when the SNR is 10 dB. On the other hand, Figure 21a–d show the voltage and current magnitude, and the voltage and current phase angles, respectively, for 0 dB SNR. The comparison between these two scenarios provides valuable insights into the effects of low SNR on the received measurements. It can be clearly observed that when the SNR is low, there are an increasing number of samples which are incorrectly received, with an error level of up to 30% in the voltage magnitude and 200% in the voltage phase angle. Since this paper is concerned with voltage regulations, errors in the voltage magnitudes could potentially impact the performance of the algorithm employed in this study.

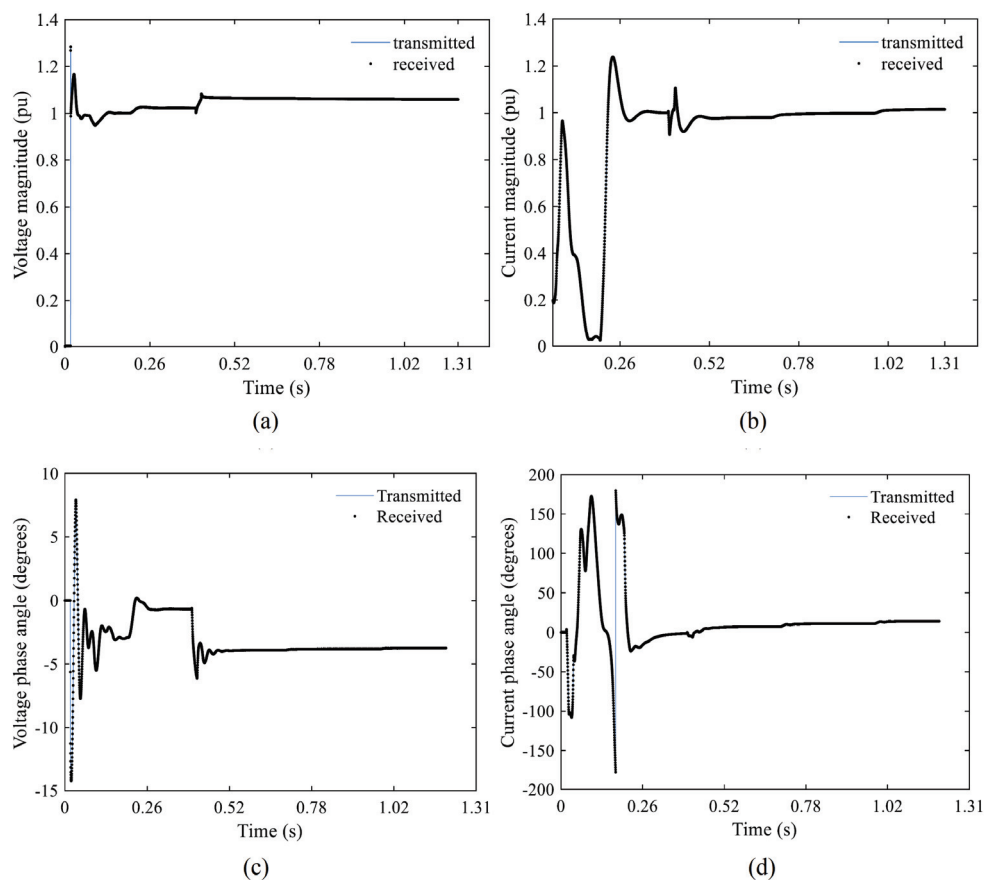


Figure 20. Transmitted and received voltage and current magnitudes (a,b) and phase angles (c,d) through the communication channel between the PMU and PDC with SNR = 10 dB.

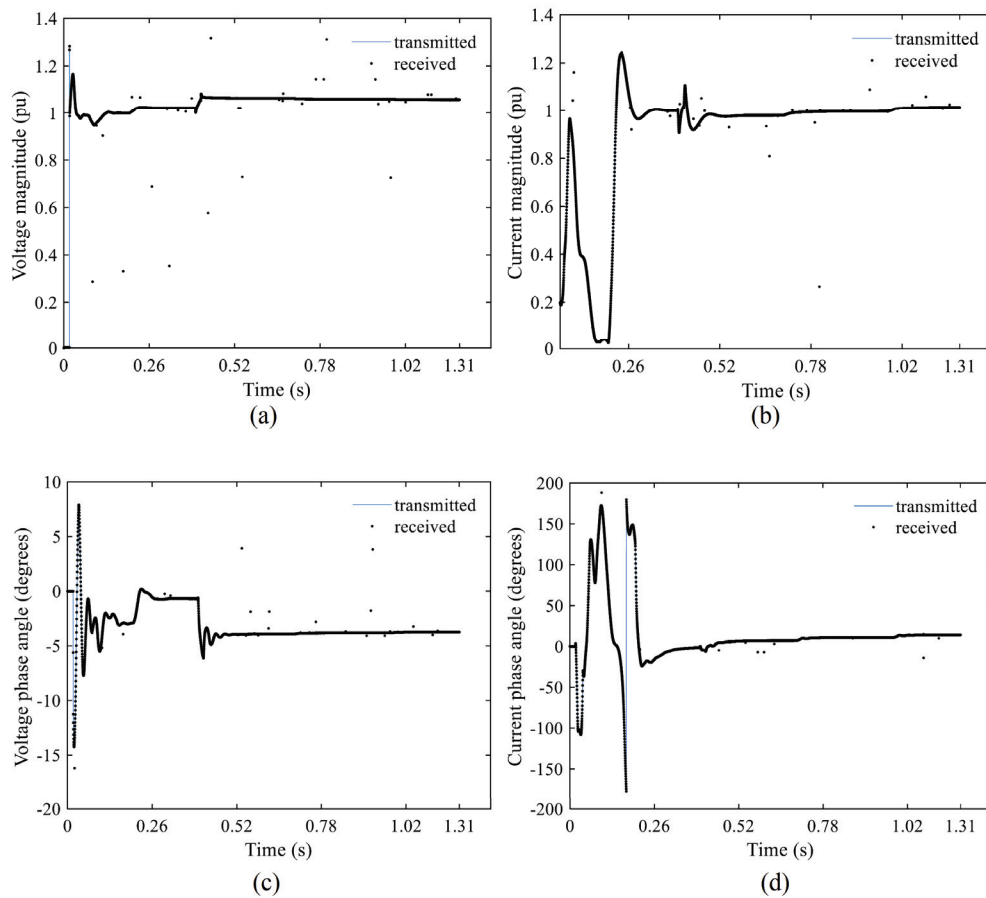


Figure 21. Transmitted and received voltage and current magnitudes (a,b) and phase angles (c,d) through the communication channel between the PMU and PDC with SNR = 0 dB.

In particular, the impact of voltage magnitude errors primarily affects Stage II of the algorithm. In this stage, the algorithm relies on accurate voltage measurements to make decisions regarding the inclusion or exclusion of DERs. If a DER's reported voltage is mistakenly received as a very low voltage level, it may be erroneously excluded from the voltage regulation process. Conversely, if a DER's reported voltage is mistakenly received as a high voltage level, it may be mistakenly included as a remote DER. However, it is important to note that if the actual voltage level at those DERs decrease due to reactive power absorption, the local control of those DERs will initiate voltage regulation to mitigate any potential issues.

4.3.2. Impact of Communication Delay

Figure 22 shows the communication delay between the PMU and PDC for (a) different channel gains (10 dB SNR) and (b) different noise power levels (channel gain 20) between the PMU and PDC. It can be observed that the communication delay varies between 0.5 ms and 2 ms (given that the LTE resource blocks have a discrete length). When the channel gain is high, most of the samples are transmitted within a 1.5 ms communication delay, which is not true when the channel gain is low. Similar observations can be made for different noise power levels. With an increasing SNR, more samples can be transmitted with a smaller delay.

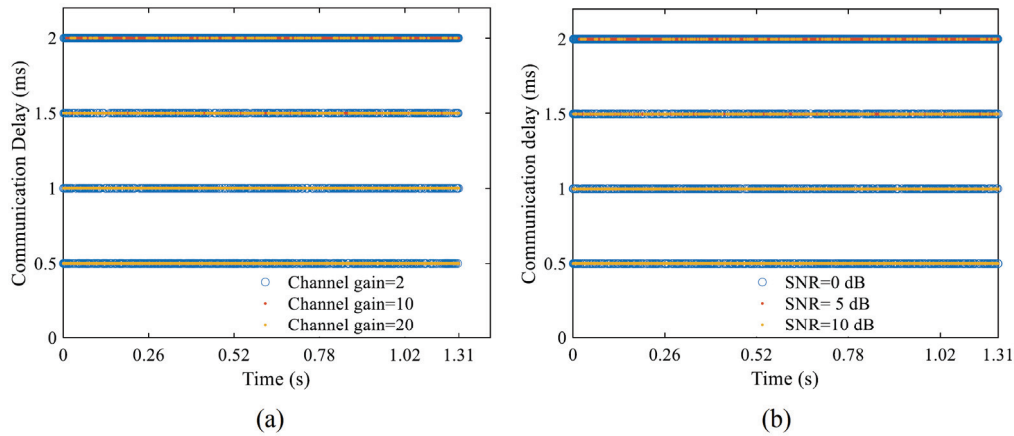


Figure 22. Communication delay between the PMU and PDC for (a) different channel gains and (b) different noise power levels.

Figure 23 shows the communication error rates for transmission between the PMU and PDC for (a) different channel gains (10 dB SNR) and (b) different noise power levels (channel gain 20). It can be observed that for a lower channel gain, the probability of receiving incorrect samples becomes higher. With a 10-fold increase in channel gain, this probability can be decreased by more than 10-fold. Similarly, for an increasing SNR, a smaller error probability can be observed.

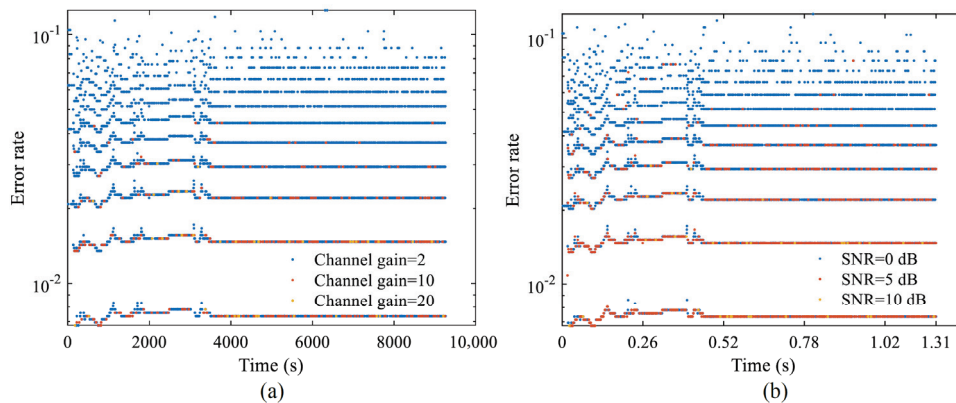


Figure 23. Error rate for communication between the PMU and PDC for (a) different channel gains and (b) different noise power levels.

It is pertinent to highlight that the failure of the communication system will not impact the operation of the DERs, as they will regulate the voltage locally in the traditional way. The coordination algorithm is added as an additional layer which ensures the efficient restoration of the voltage with minimum power curtailment.

5. Conclusions

The hierarchical multi-stage algorithm proposed in this study has demonstrated its effectiveness in achieving coordinated voltage control in distribution systems. The algorithm, which consists of four stages, employs a combination of reactive power absorption and power generation curtailment to restore the system voltage, with communication based on the PMU at the local controllers and the PDC at the central control unit facilitating data exchange between the DERs and the central controller. By prioritizing reactive power absorption over curtailment, the proposed algorithm reduces unnecessary curtailment, leading to a more efficient voltage control system based on PMU-PDC communication technology. The simulation results obtained using the IEEE 13-bus radial distribution benchmark system confirmed the efficacy of the proposed control system for achieving

coordinated voltage control and ensuring the accuracy of the PMU-PDC communication interface. In addition, to demonstrate its scalability, the proposed coordinated algorithm was successfully applied to the IEEE 33-bus distribution network, showcasing its effectiveness across a larger system. Overall, the proposed hierarchical control strategy based on PMU-PDC provides an effective means for improving voltage control in distribution systems and represents a promising area for future research.

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Article

Backstepping Control of NPC Multilevel Converter Interfacing AC and DC Microgrids

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Abstract: This work introduces modified backstepping methods to design controllers for neutral point clamped (NPC) converters interfacing a DC/AC microgrid. The modified backstepping controllers are derived from a proper converter model, represented in dq coordinates, and are designed to regulate the DC voltage and to balance the two NPC converter DC capacitor voltages through a DC offset in the sinusoidal pulse width modulation (SPWM) carriers. The averaged and separated dynamics backstepping controllers also enforce nearly sinusoidal AC currents at a given power factor. The two proposed NPC converter controllers are evaluated through MATLAB/Simulink simulations and experimental implementation using a laboratory prototype. Simulations and experimental results show that the two modified backstepping controllers regulate the microgrid DC voltage in steady state and in transient operation, even with load disturbances or DC voltage reference changes, while enforcing nearly AC sinusoidal currents at a given power factor or injected reactive power. The modified backstepping-controlled NPC converter is bidirectional, converting energy from DC renewable energy sources or storage systems to AC or charging storage systems from AC. The results also highlight the effective balancing of the NPC DC capacitor voltages.

Keywords: backstepping control; microgrids; multilevel converter; separate dynamics; capacitor voltage balance; pulse width modulation

1. Introduction

The adoption of microgrids is increasing globally, driven by their capacity to integrate decentralized renewable energy generation and operate autonomously from the conventional AC grid. Microgrids have the capability to incorporate distributed renewable generation sources like solar panels or wind turbines, along with community energy storage devices typically in the form of battery banks, enabling the microgrid disconnection from the national grid. Moreover, microgrids are designed to autonomously address local power issues, thereby ensuring reliable power supply [1–3].

Electricity consumption is experiencing ongoing global growth, and it is projected to persist in the coming years due to industrial and human activities, the increasing adoption of electric vehicles (EVs), and the drive toward decarbonizing the energy sector. Consequently, there has been a significant push to adopt distributed energy sources, with a particular emphasis on those derived from renewable energies [4,5].

Microgrids can operate only in AC or in DC or they may contain both AC and DC subsystems (hybrid). Power converters play a crucial role in connecting AC sources to DC microgrids or DC sources to DC microgrids. In the case of hybrid microgrids, converters may also be employed to facilitate the interconnection of storage systems, which may include batteries or fuel cells.

With the inclusion of DC power sources like PV panels, fuel cells, and EV batteries capable of supplying power to the grid through Vehicle-to-Grid (V2G) technology, as well as the existence of DC loads, the concept of hybrid microgrids combining AC and DC components has gained significant attention. As AC power networks have matured considerably, hybrid AC/DC microgrids that accommodate both AC and DC loads and sources are now seen as the most promising systems for efficient distribution and transportation of electrical energy [6].

The growing adoption of EVs in recent years suggests the potential utilization of their storage batteries for power/frequency control in microgrids. EV batteries, equipped with bi-directional power converters and controllers, can be viewed as mobile storage units capable of fast-charging other EVs with immediate high-power energy requirements. Additionally, they can enhance the performance of specific micro-regions within the microgrid [7,8].

Electronic power conversion within microgrids often encompasses power in the order of multiple megawatts, with voltages reaching kilovolt levels. In this context, multilevel converters emerge as a suitable solution for microgrids.

Besides energy conversion, the main role of converters integrated with microgrids is to uphold the quality of electrical power and to regulate the DC voltage, aiding photovoltaic (PV) panels to follow the maximum power point tracking (MPPT) voltage. When operating as a source on the AC side, converters work toward minimizing harmonic distortion. Conversely, when connected to a load on the DC side, they are responsible for regulating the level of DC voltage. In most unbalanced systems, voltage compensation is typically achieved through the utilization of active power filters [9–11].

The regulation of power/frequency and voltage/frequency in microgrids with distributed inverter-based generators is commonly achieved through the implementation of the droop control method. This method generally performs effectively in reactive power regulation [12–15] and can be of benefit if converter level controllers are fast and stable, ensuring low distortion AC waveforms and DC tracking capability.

Multiple topologies of multilevel converters exist, including the neutral point clamped (NPC) [16], flying-capacitors [17], and cascaded multicell with isolated DC sources [18]. Several modulation and linear control strategies have been devised for multilevel converters, such as sinusoidal pulse width modulation (SPWM) and space vector modulation (SVM), coupled to linear proportional integral (PI) controllers.

Coping with the nonlinearity and unpredictable disturbances in multilevel converters presents notable challenges in their control. To address these challenges effectively, the implementation of nonlinear controllers capable of handling unknown disturbances such as sliding mode and predictive control [19–26] emerged as first solutions, despite sliding mode chattering and power quality requirements needing high processing speeds for predictive control [26].

Multilevel converters enable operation at voltages and power levels of demanding microgrids and are also capable of power quality enhancement in microgrids [26–30] despite needing fast digital processors. To mitigate this issue, the backstepping technique stands out as a powerful tool for developing controllers that are both robust and stable [30–35]. By employing virtual control quantities and designing controllers for intermediate variables, backstepping enables a systematic approach, where the focus initially lies on the output and then progressively moves recursively towards the control input.

Backstepping controllers have several advantages due to their direct relationship with converter nonlinear dynamics and their ability to identify the control laws and the essential quantities to be measured or estimated. Furthermore, the backstepping approach is based on positive definite functions that have been chosen to ensure system stability and robust-

ness [35–43]. Recently, the authors of [44] presented a method to combine backstepping and integral sliding mode to control grid-tied parallel connected inverters. In comparison to predictive controllers, backstepping requires less computational power [36,44] and maintains constant switching frequencies, all while preserving the response speed and robustness exhibited by predictive controllers [21,26,45,46]. Furthermore, predictive controllers for NPC converters necessitate extremely fast processors, whereas traditionally designed backstepping controllers [36] may necessitate extremely high switching rates, overloading PWM modulators (more than one commutation per switching period, leading to increased non-constant switching frequencies).

To address the aforementioned issues, this paper proposes two modified backstepping control algorithms to give the bidirectional NPC converter bridging the DC and AC parts of a microgrid the capabilities of regulating the DC voltage, the AC currents, and the reactive power, while balancing the NPC internal bipolar voltages. The obtained modified averaged backstepping controller and the AC and DC separated dynamics backstepping controller both regulate the converter DC voltage tightly, control the AC currents ensuring very low total harmonic distortion (THD), and maintain the bipolar voltage balance of the NPC internal capacitors, using an industry-proven fixed frequency carrier-based SPWM technique.

This paper presents in Section 2 the dq coordinate model of the multilevel converter interfacing the AC grid to the DC microgrid. Two modified backstepping controllers are derived from the NPC converter dynamics model to control the DC voltage, AC current, and balance the capacitor voltages by regulating the SPWM carrier offset. The first modified backstepping controller assumes the control action takes the average value in steady state for the next backstep. The second modification adds the assumption of decoupled (independent) AC and DC dynamics. Section 3 outlines results obtained from simulation using MATLAB/Simulink and from a reduced-scale laboratory prototype. Results cover steady-state, transient performance, THD, DC voltage tracking errors, and the comparison to standard linear control based on SPWM coupled to PI controllers. Section 4 summarizes the main findings and conclusions.

2. Microgrid NPC Converter Model and Controllers

The multilevel NPC converter is connected to a DC grid, with loads and DC renewable sources, while interfacing the AC microgrid, as depicted in Figure 1.

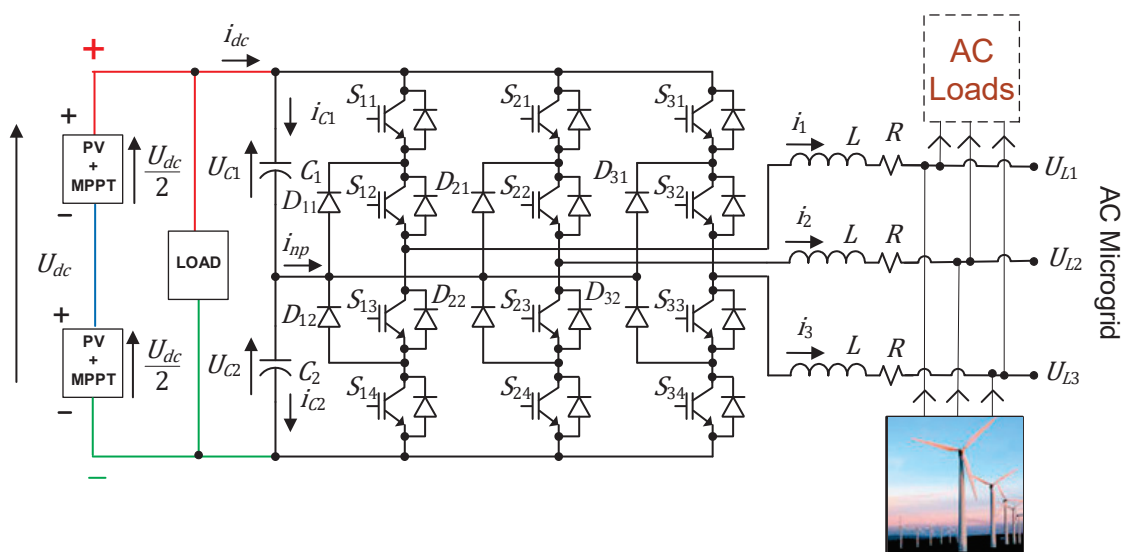


Figure 1. Multilevel NPC converter circuit connecting DC loads and distributed power sources to the AC microgrid.

The NPC converter currents i_1 , i_2 , and i_3 are AC three-phase currents, flowing into the AC microgrid, with voltages U_{L1} , U_{L2} , and U_{L3} , through coupling inductors L , with loss resistors, R . The NPC converter DC part has two capacitors, C_1 and C_2 , with voltages U_{C1} and U_{C2} , respectively, which should be kept approximately constant at $U_{dc}/2$. The DC grid voltage is U_{dc} and the NPC DC current i_{dc} flows into the DC grid.

2.1. Microgrid-Connected NPC Converter Model

Supposing ideal semiconductors, the dynamics of the NPC converter can be written considering that each NPC converter k leg ($k \in \{1, 2, 3\}$) has four switches. Assuming that when switch S_{k1} is closed, its state is written as $S_{k1} = 1$, otherwise when switch S_{k1} is open $S_{k1} = 0$, while $S_{k3} = 1 - S_{k1}$ and $S_{k4} = 1 - S_{k2}$. Each k leg outputs three voltages (states), defined by a variable γ_k , dependent on the leg switches states.

$$\gamma_k = \begin{cases} 1 & S_{k1} = 1 \text{ and } S_{k2} = 1 \\ 0 & S_{k1} = 0 \text{ and } S_{k2} = 1 \\ -1 & S_{k1} = 0 \text{ and } S_{k2} = 0 \end{cases} \quad (1)$$

The averaged switching model of the NPC converter can be obtained, assuming $C_1 = C_2 = C$ and applying Kirchhoff laws, Clark–Concordia, and Park transforms to Figure 1 circuit equations, while assuming the microgrid has nearly balanced DC voltages $U_{C1} \approx U_{C2}$ and DC currents, together with nil zero sequence AC currents. State variables chosen for the NPC dynamic model in Equation (2) are the balanced AC currents in dq coordinates, i_d and i_q , while the DC voltages are represented by U_{dc} , as the sum of the two U_{C1} , U_{C2} capacitor voltages ($U_{dc} = U_{C1} + U_{C2}$), and by E_{UC} as the difference or unbalance between the two capacitor voltages $E_{UC} = U_{C1} - U_{C2}$:

$$\frac{d}{dt} \begin{bmatrix} i_d \\ i_q \\ U_{dc} \\ E_{UC} \end{bmatrix} = \begin{bmatrix} \frac{-R}{L} & \omega & \frac{\gamma_d}{2L} & 0 \\ -\omega & \frac{-R}{L} & \frac{\gamma_q}{2L} & 0 \\ \frac{-\gamma_d}{C} & \frac{-\gamma_q}{C} & 0 & 0 \\ -\frac{2}{C}\gamma_d\Delta_\gamma + \Delta_d & -\frac{2}{C}\gamma_q\Delta_\gamma + \Delta_q & 0 & 0 \end{bmatrix} \begin{bmatrix} i_d \\ i_q \\ U_{dc} \\ E_{UC} \end{bmatrix} + \begin{bmatrix} -\frac{1}{L} & 0 & 0 \\ 0 & -\frac{1}{L} & 0 \\ 0 & 0 & \frac{2}{C} \\ 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} U_{Ld} \\ U_{Lq} \\ U_{dc} \\ i_{dc} \end{bmatrix} \quad (2)$$

where i_d and i_q are, respectively, the direct and quadrature components of the Park transform of the converter AC currents i_1 , i_2 , and i_3 , ω is the angular frequency of the microgrid. The Park transform components of the averaged values Γ_1 , Γ_2 , and Γ_3 , during one switching period of the semiconductor states, γ_1 , γ_2 , and γ_3 are γ_d , γ_q , and γ_0 , while U_{Ld} , U_{Lq} are the direct and quadrature voltages of balanced AC voltages of the microgrid. The quantities Δ_d , Δ_q (given in Equations (3) and (4), respectively) are received (Appendix A) starting from the switching period averaged values Γ_1 , Γ_2 , and Γ_3 , and considering their small signal responses to a Δ_γ , offset added to the SPWM carrier, as $\Gamma_1 + \Delta_\gamma$, $\Gamma_2 + \Delta_\gamma$, and $\Gamma_3 + \Delta_\gamma$, while applying the Clark–Concordia and Park transforms. The Δ_γ offset is the control input to balance the two capacitor voltages (to track $E_{UCRef} = 0$).

$$\Delta_d = \frac{\gamma_d^2 \cos \omega t}{C} \left(\frac{1+\sqrt{12}}{\sqrt{18}} \sin^2 \omega t - \frac{\cos^2 \omega t}{\sqrt{6}} \right) + \frac{\gamma_q^2 \cos \omega t}{C} \left(\frac{\sqrt{2}}{6} \cos^2 \omega t - \frac{\sin^2 \omega t}{\sqrt{6}} \right) + \frac{\gamma_d \gamma_q \sin \omega t}{C} \left(\frac{4\sqrt{6}+2\sqrt{2}}{6} \cos^2 \omega t - \sqrt{\frac{2}{3}} \sin^2 \omega t \right) - \frac{\gamma_d \gamma_0}{C} \frac{2}{\sqrt{3}} \quad (3)$$

$$\Delta_q = \frac{\gamma_d^2 \sin \omega t}{C} \left(\frac{3}{\sqrt{6}} \cos^2 \omega t - \frac{\sqrt{2}}{6} \sin^2 \omega t \right) + \frac{\gamma_q^2 \sin \omega t}{C} \left(\frac{\sin^2 \omega t}{\sqrt{6}} - \frac{\sqrt{2}+\sqrt{24}}{6} \cos^2 \omega t \right) + \frac{\gamma_d \gamma_q \cos \omega t}{C} \left(\sqrt{\frac{2}{3}} \cos^2 \omega t - \frac{4\sqrt{6}+2\sqrt{2}}{6} \sin^2 \omega t \right) - \frac{\gamma_q \gamma_0}{C} \frac{2}{\sqrt{3}} \quad (4)$$

As seen in Equation (2), the dynamics of i_d and i_q are characterized by cross-coupling and non-linearity. However, they can be controlled individually by γ_d and γ_q , respectively. As usually i_q is small compared to i_d , the DC voltage U_{dc} depends mainly on i_d . Therefore,

i_d must be used to control U_{dc} , while i_q may control the power factor (or the reactive power). The capacitor voltage balance error E_{UC} control input Δ_γ in Equation (2) has negative gains, $-\gamma_d/C$ and $-\gamma_q/C$ regarding the i_d and i_q currents, respectively.

2.2. Block Diagram of the Microgrid Backstepping Controllers

Due to battery storage or renewable resources, power transfer must be bidirectional between DC and AC microgrid sides. Therefore, backstepping DC voltage controllers that ensure AC sinusoidal line currents at a given power factor (Figure 2) are designed. They generate the modulation waveforms of a 3-phase 3-level SPWM modulator using two triangular-shaped carriers with adjustable offset Δ_γ .

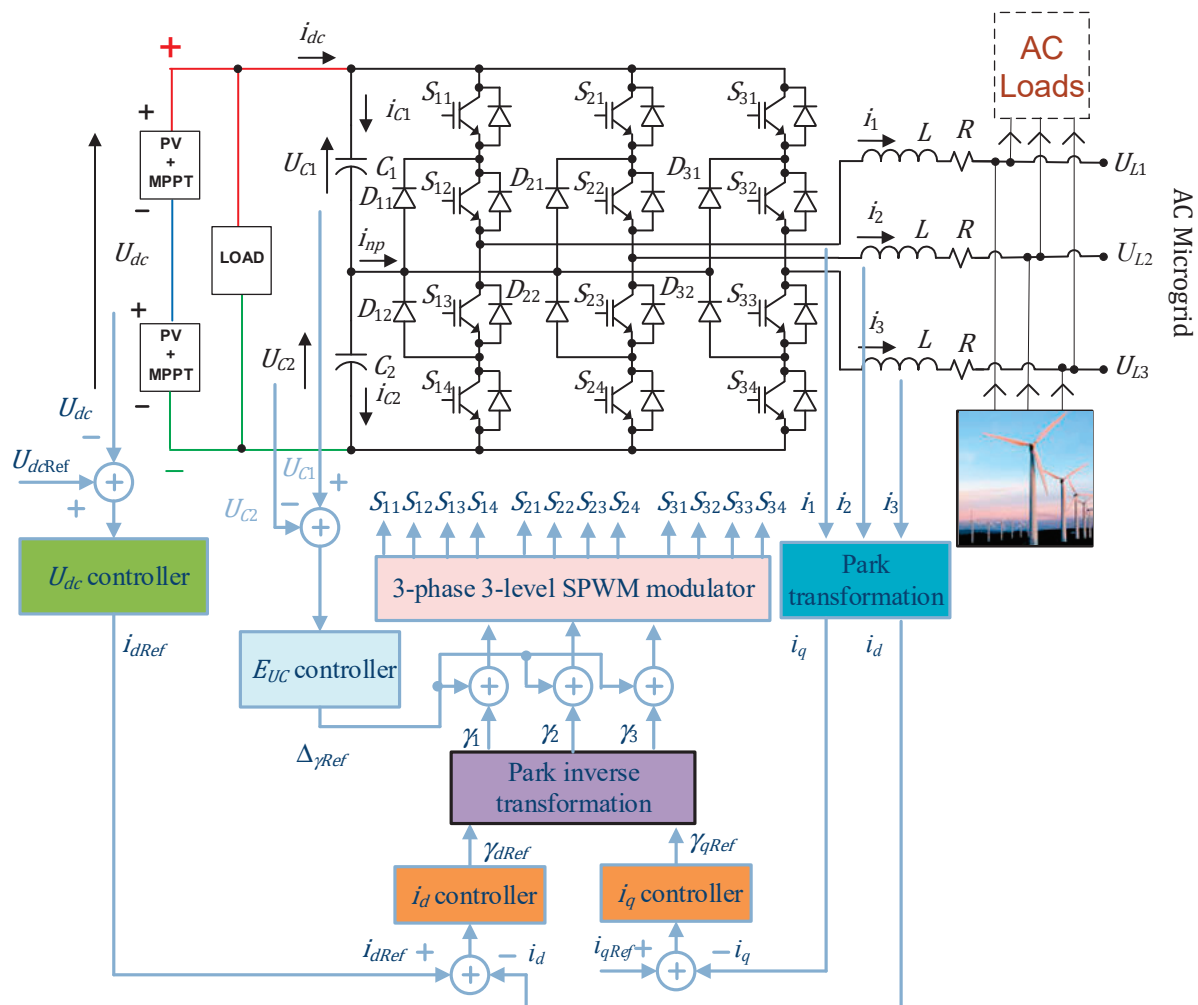


Figure 2. Block diagram of the microgrid backstepping controllers.

In the block diagram of the microgrid converter, the DC voltage is measured and compared to the reference value U_{dcRef} . The application of the backstepping method to the U_{dc} voltage dynamics in Equation (2) results in the U_{dc} controller, which calculates the virtual value of the current, i_{dRef} . The AC currents are measured and the Park transform is applied. These currents in dq , i_d , and i_q are compared with their magnitudes, i_{dRef} and $i_{qRef} = -Q_{Ref}/U_{Ld}$, where Q_{Ref} is the reactive power value. The application of the backstepping control method results in the command quantities of the converter legs, γ_{dRef} and γ_{qRef} , which are then converted into quantities in 123, γ_1 , γ_2 , and γ_3 . The backstepping control method is also applied to the voltage error of the capacitors to calculate the offset $\Delta_{\gamma Ref}$, which is added to the command quantities, γ_1 , γ_2 , and γ_3 , to balance the voltage of the capacitors.

2.3. Design of Microgrid Backstepping Controllers

2.3.1. Modified Backstepping Control of the Microgrid DC Voltage

To control the bidirectional DC/AC microgrid NPC converter interface, the backstepping approach starts from the DC side total voltage $U_{dc} = U_{C1} + U_{C2}$. Define the U_{dc} voltage tracking error, e_{Udc} , relative to its reference value U_{dcRef} , as:

$$e_{Udc} = U_{dcRef} - U_{dc}, \quad (5)$$

A feedback controlled variable x is asymptotically stable if there is a Lyapunov positive definite candidate function $V(x)$ obeying (1) $V(x=0)=0$; (2) $V(x \neq 0) > 0$; (3) $V(|x| \rightarrow \infty) \rightarrow \infty$; (4) $\dot{V}(x \neq 0) < 0$. Hence, the system stability condition is $V(x \neq 0) \dot{V}(x \neq 0) < 0$ [33]. Therefore, consider the e_{Udc} variable and select a positive definite function, V_1 , as follows:

$$V_1 = \frac{e_{Udc}^2}{2}, \quad (6)$$

where the V_1 function obeys the previous conditions. For stability, the time derivative of V_1 must be negative. This can be accomplished if dV_1/dt is enforced to equal $-K_{eUdc}e_{Udc}^2$, where K_{eUdc} is a positive definite constant:

$$\frac{dV_1}{dt} = e_{Udc} \frac{de_{Udc}}{dt} = -K_{eUdc}e_{Udc}^2. \quad (7)$$

Replace e_{Udc} from Equation (5) in the term de_{Udc}/dt of Equation (7) and use dU_{dc}/dt from Equation (2), to receive the control action that enforces the dynamics in Equation (7), called virtual control action i_{dV} in Equation (8). The direct current i_d is supposed to track the virtual control action i_{dV} value to enforce stability.

$$i_{dV} = \frac{C}{\gamma_d} \left(-K_{eUdc}e_{Udc} - \frac{dU_{dcRef}}{dt} + \frac{2}{C}i_{dc} \right) \quad (8)$$

The virtual current i_{dV} depends on the available control input γ_d ($\gamma_d \neq 0$). To enforce the current i_d to track the virtual control value i_{dV} , backstep recursively by defining the tracking error e_{id} of the i_d current as:

$$e_{id} = i_{dV} - i_d. \quad (9)$$

Consider a composite positive definite function, V_2 , as:

$$V_2 = \frac{e_{Udc}^2}{2} + \frac{e_{id}^2}{2}, \quad (10)$$

where the time derivative of V_2 must be negative to ensure stability, a condition that can be written as:

$$\frac{dV_2}{dt} = e_{Udc} \frac{de_{Udc}}{dt} + e_{id} \frac{de_{id}}{dt} = -K_{eUdc}e_{Udc}^2 - K_{eid}e_{id}^2 \quad (11)$$

where K_{eUdc} , K_{eid} are positive definite constants. From Equation (8), the term de_{id}/dt of Equation (11) would contain the time derivative of the control input γ_d as $di_{dV}/dt = -(i_{dV}/\gamma_d)(d\gamma_d/dt)$. This means the control input γ_d would be found integrating its time derivative [36]. This time derivative contains high frequency harmonics which strongly affect the THD of the 3-phase currents [37]. Therefore, knowing that in three-phase systems, dq variables are nearly constant in each operating regime, this paper proposes a modified averaged backstepping method. It supposes a conservative converter ($R \approx 0$) and neglecting the inductor voltage drop at the fundamental frequency $\omega L i_q \ll (\gamma_d U_{dc}/2 - U_{Ld})$, from di_d/dt in Equation (2), the γ_d average value ($\bar{\gamma}_d$) defined in steady state ($di_d/dt = 0$), is obtained by solving the resulting identity, $0 = \bar{\gamma}_d U_{dc}/L - U_{Ld}$, for $\bar{\gamma}_d$, giving $\bar{\gamma}_d = 2U_{Ld}/U_{dc}$. This

value replaces γ_d in Equation (8) to give a virtual control current i_{dv} not disturbed by high frequency harmonics:

$$i_{dv} = \frac{U_{dc}C}{2U_{Ld}} \left(-K_{eUdc}e_{Udc} - \frac{dU_{dcRef}}{dt} + \frac{2}{C}i_{dc} \right) \quad (12)$$

Replacing in Equation (11) e_{Udc} and e_{id} from Equations (5) and (9), respectively, and using the i_d current dynamics di_d/dt and the U_{dc} voltage dynamics dU_{dc}/dt , both from Equation (2), after some simplifications, the control component, γ_{dRef} , for the averaged backstepping is obtained (Appendix B):

$$\gamma_{dRef} = \frac{2L}{U_{dc}} \left(K_{eid}e_{id} - \frac{\gamma_d}{C}e_{Udc} + \frac{di_{dv}}{dt} + \frac{R}{L}i_d - \omega i_q + \frac{U_{Ld}}{L} \right) \quad (13)$$

The term $\gamma_d e_{Udc}/C$ is often negligible. Nevertheless, it can be estimated using the previous value of γ_{dv} and e_{Udc} . The control law γ_{dRef} will be applied to a SPWM modulator through the inverse Park transform (Figure 2).

2.3.2. Control of the Reactive Power Injected in the AC Microgrid

The control of the reactive power can be performed indirectly by regulating the quadrature current component i_q . Define the tracking error e_{iq} :

$$e_{iq} = i_{qRef} - i_q, \quad (14)$$

where i_{qRef} is the i_q current reference value related to the reactive power Q_{Ref} ($i_{qRef} = -Q_{Ref}/U_{Ld}$, $i_{qRef} = 0$ if unity power factor is sought), choosing a positive definite function, V_3 , given in Equation (15).

$$V_3 = \frac{e_{iq}^2}{2} \quad (15)$$

Computing the time derivative of V_3 , and forcing it to be negative, using a positive definite constant K_{eiq} , the stability criterion is guaranteed:

$$\frac{dV_3}{dt} = e_{iq} \frac{de_{iq}}{dt} = -K_{eiq}e_{iq}^2 \quad (16)$$

Replacing e_{iq} in Equation (16) using Equation (14) and the i_q current dynamics di_q/dt from Equation (2), the control component, γ_{qRef} , is obtained:

$$\gamma_{qRef} = \frac{2L}{U_{dc}} \left(K_{eiq}e_{iq} + \frac{di_{qV}}{dt} + \frac{R}{L}i_q + \omega i_d + \frac{U_{Lq}}{L} \right) \quad (17)$$

where the control action γ_{qRef} will be fed to the aforementioned inverse Park transform and applied to the SPWM modulator (Figure 2).

2.3.3. Balancing the DC Capacitor U_{C1} , U_{C2} Voltages

The multilevel converter DC side capacitor voltages U_{C1} , U_{C2} may drift from the midpoint value due to converter asymmetries in the semiconductors, voltage vector selection, or usage of the DC microgrid in a bipolar connection. To zero the unbalance value $E_{UC} = U_{C1} - U_{C2}$, a control input was created by adding the offset Δ_γ , in the triangular SPWM modulator (Figure 2). The resulting dE_{UC}/dt dynamics (2) contains negative gains, regarding the i_d and i_q currents.

Select a positive definite function, V_4 , for the unbalance E_{UC} :

$$V_4 = \frac{E_{UC}^2}{2} \quad (18)$$

To ensure stability, enforce the time derivative of V_4 to be negative, using a positive definite constant K_{EUC} :

$$\frac{dV_4}{dt} = E_{UC} \frac{dE_{UC}}{dt} = -K_{EUC} E_{UC}^2 \quad (19)$$

Then, replace in Equation (19) the dE_{UC}/dt term using Equation (2), where Δ_γ equals its reference value $\Delta_{\gamma Ref}$, to receive the control law (Appendix C) for this offset (Figure 2):

$$\Delta_{\gamma Ref} = \frac{C}{4i_{dc}} (K_{EUC} E_{UC} + \Delta_d i_d) \quad (20)$$

2.4. DC Microgrid Voltage Backstepping Control Assuming Separated AC and DC Dynamics

The DC microgrid voltage U_{dc} should be as constant as possible, while the AC current components i_d and i_q represent nearly sinusoidal AC currents with 50 Hz frequency. Therefore, inductive and capacitive components should be designed to allow the AC current with much faster dynamics regarding DC voltage dynamics. Consequently, in the AC current control, the much slower DC voltage dynamics can be neglected. This assumption enables simplifying the control law γ_{dRef} . Therefore, the separated dynamics backstepping process is similar to Section 2.3.1, except that the positive definite function, V_2 (Equation (10)), used for the current i_d to track the virtual control value i_{dV} , is replaced by V_{2s} that considers only the tracking error e_{id} :

$$V_{2s} = \frac{e_{id}^2}{2} \quad (21)$$

For stability, the time derivative of V_{2s} is made negative using a positive definite constant K_{eids} :

$$\frac{dV_{2s}}{dt} = e_{id} \frac{de_{id}}{dt} = -K_{eids} e_{id}^2 \quad (22)$$

Replacing in Equation (22) the e_{id} by Equation (9) and the i_d dynamics from di_d/dt in Equation (2), the separated dynamics control law is γ_{dsRef} :

$$\gamma_{dsRef} = \frac{2L}{U_{dc}} \left(K_{eids} e_{id} + \frac{di_{dV}}{dt} + \frac{R}{L} i_d - \omega i_q + \frac{U_{Ld}}{L} \right), \quad (23)$$

where comparing γ_{dsRef} in Equation (23) to γ_{dRef} in Equation (13), the only difference is the absence in Equation (23) of the DC voltage error, e_{Udc} , because the slow DC dynamics do not interfere with the faster AC current dynamics.

3. Microgrid Simulation and Experimental Results

This section presents the simulation and experimental results of the AC DC microgrid interface. The simulations were conducted implementing the microgrid model in the Power System Toolbox of MATLAB/Simulink, together with the controllers derived through the averaged and the separated dynamics backstepping methods. A low-power laboratory prototype of the multilevel NPC converter (Figure 1) was built to evaluate the performance of a scaled-down microgrid controlled by backstepping controllers. Digital processors are used to implement the averaged dynamics (i_{dV} from Equation (12), γ_{dRef} from Equation (13)) and the separated dynamics (i_{dV} from Equation (12), γ_{dsRef} from Equation (23)) backstepping controllers, together with the offset triangular carrier modulators at a sampling time t_s of 28 μ s. This sampling time is suitable for 50 to 60 Hz AC grid frequency f and NPC converters switching around 4 kHz compatible with MW AC microgrids.

Simulations and experiments were conducted in both steady-state and transient conditions, with distributed loads or power sources on the DC side. A comparison between the simulation and experimental performance of the modified backstepping controllers is performed to assess the performance of the averaged and the separated dynamics controllers.

The parameters and corresponding values used in the simulations and experimental results are shown in Table 1.

Table 1. Microgrid parameters and their descriptions and values.

Parameter	Value	Parameter	Value
U_{dc}	110 V $\pm$ 10%	$K_{eU_{dc}}$	126.6
$Z_{U_{dc}}$	100 Ω	K_{eid}	20/ t_s
C	4.4 mF	K_{eiq}	20/ t_s
L	15.1 mH	K_{EUC}	2/ C
R	0.1 Ω	K_{eids}	20/ t_s
U_L	24 V	f	50 Hz
ω	$2\pi f = 100\pi$	t_s	28 μ s

3.1. DC/AC Microgrid Steady-State Simulation and Experimental Results

Steady-state microgrid results are shown under constant DC voltage and constant load. Figure 3a shows the DC voltage simulation value (120 V) of the microgrid using either the averaged or the separated dynamics backstepping control methods. The simulation results of both backstepping DC voltage controllers are virtually identical, showing that, in steady-state conditions, both controllers effectively regulate the DC voltage with measured steady-state stationary U_{dc} error around 0.5 V (<0.5%). This supports the validity of the separated dynamics assumption. In the laboratory prototype, the results of DC voltage regulation in Figure 4a show the same value (120 V) as the simulation results, although some high frequency noise is apparent.

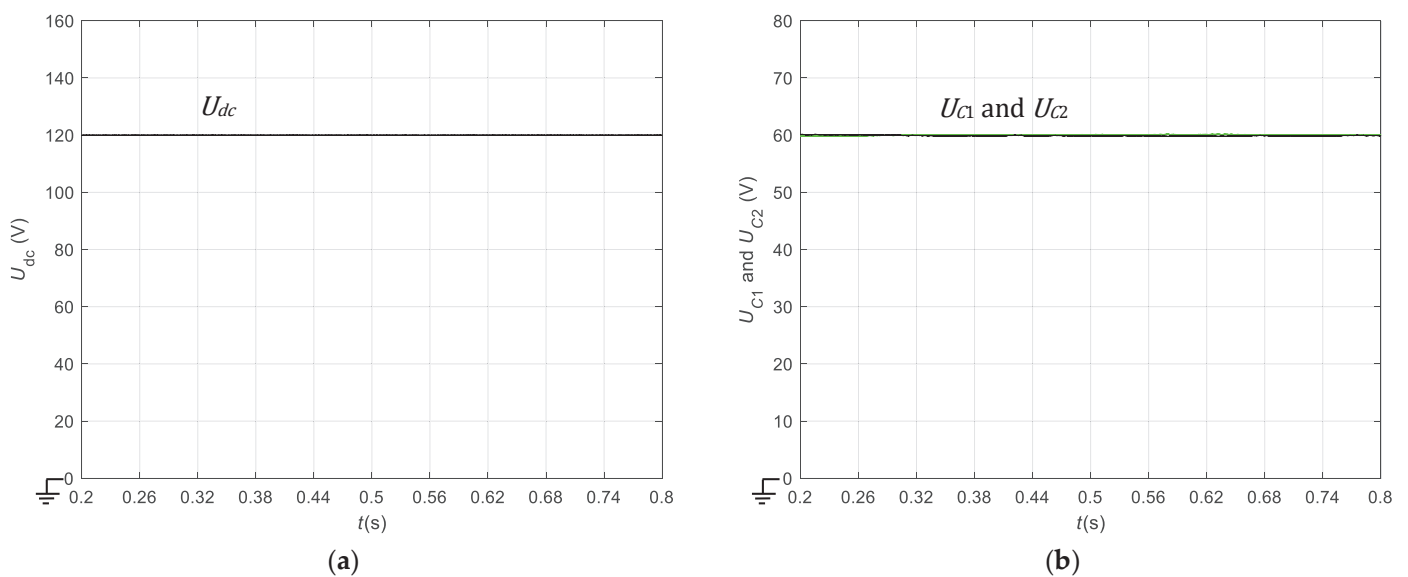


Figure 3. DC voltage simulation results in steady state using the averaged controller: (a) U_{dc} voltage; (b) U_{C1} and U_{C2} voltages.

Experimental results of balancing the DC capacitor U_{C1} , U_{C2} voltages are shown in Figure 4b, indicating that the backstepping algorithm that controls the SPWM carrier offset effectively balances the capacitor voltages (DC error below 0.5%). Similar simulation results were obtained (Figure 3b) using both the averaged and the separated backstepping methods.

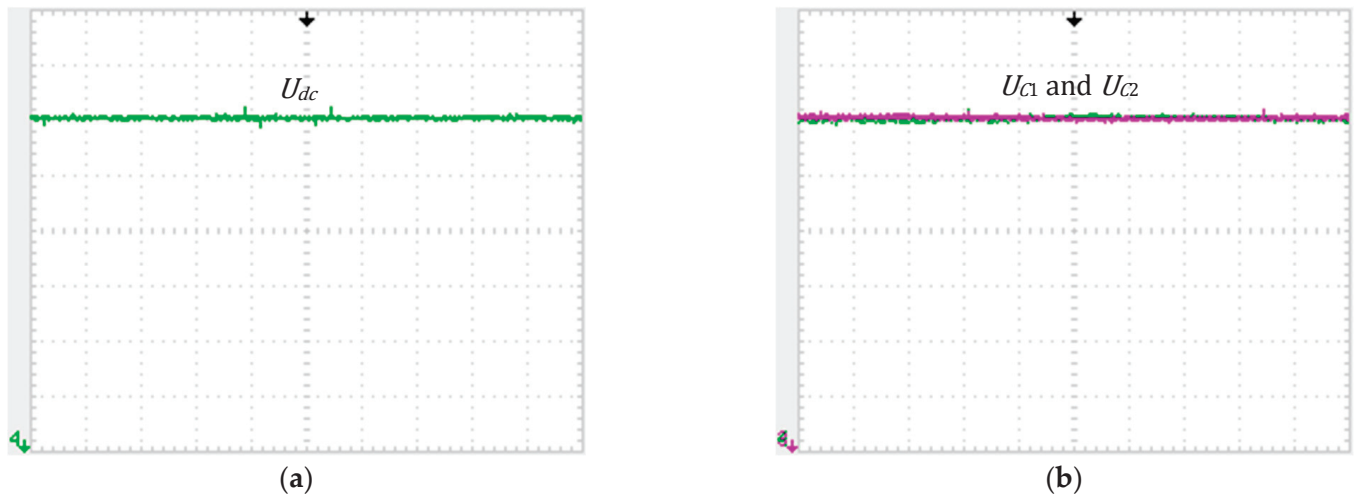


Figure 4. Experimental results in steady state using the separated dynamics backstepping controller: (a) DC voltage, vertical: 20 V/div; (b) Capacitor voltages, U_{C1} and U_{C2} , vertical: 10 V/div. Horizontal: 50 ms/div.

Simulation results in Figure 5 show the AC currents obtained using either the averaged dynamics (i_{dV} from Equation (12), γ_{dRef} from Equation (13)) or the separated dynamics (i_{dV} from Equation (12), γ_{dsRef} from Equation (23)) backstepping controllers. Offsets were added to each current to clearly show every single AC current. Currents are sinusoidal, balanced, with very low THD (simulations 1.6% (MATLAB/Simulink), and experimental 1.7% (Fluke 435)). These results further confirm the effectiveness of the averaged dynamics and the validity of the separated dynamics assumption. Experimental tests confirm the values in the simulations (Figure 5b). The experimental AC current, i_1 , THD was obtained from the class A power quality analyzer Fluke 435 to a csv file, whose harmonics are represented in Figure 6b.

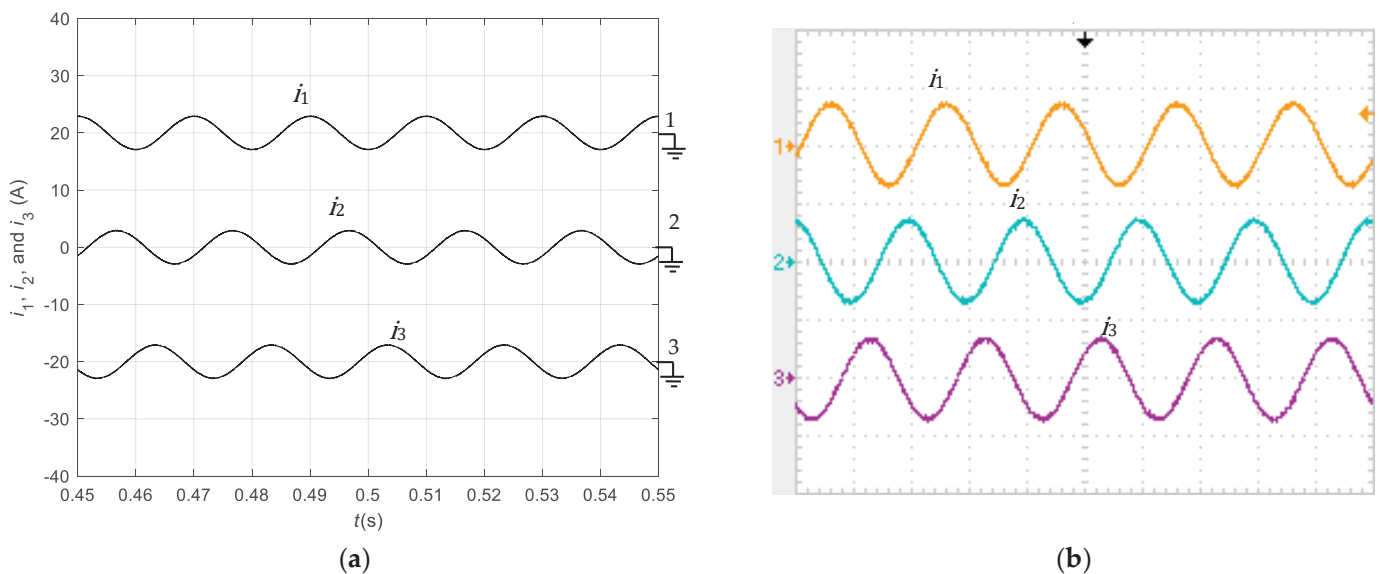


Figure 5. Results of AC current, i_1 , i_2 , and i_3 , in steady state with the separated dynamics backstepping controller: (a) Simulation; (b) Experimental results. Vertical: 4 A/div. Horizontal: 10 ms/div.

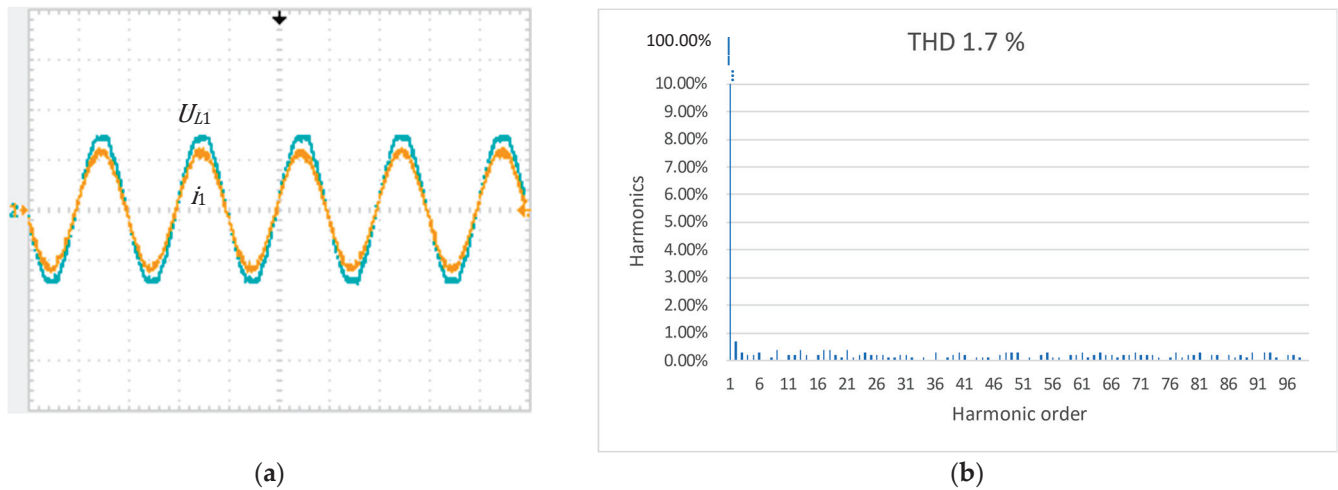


Figure 6. (a) Experimental results of AC voltage and AC current, U_{L1} and i_1 , in steady state with the averaged backstepping controller. Converter connected to a distributed power source. Vertical: 2 A/div and 20 V/div. Horizontal: 10 ms/div; (b) Experimental result of AC current, i_1 , THD.

Figure 6a shows experimental results of AC current i_1 and AC voltage U_{L1} , in steady state, with DC passive loads and renewable energy injection, with near unity power factor ($Q_{Ref} = 0$). Voltage U_{L1} and current i_1 are in phase meaning DC power is being injected in the AC microgrid (Figure 1) at unity power factor.

These steady-state results indicate that the averaged backstepping control is able to control the DC voltage, to balance the capacitor voltages U_{C1} , U_{C2} , to enforce almost sinusoidal AC currents with THD < 1.7%. The modified backstepping-controlled converter is bidirectional, as now, power is flowing into the AC grid (see Figure 1). Simulation results are similar to the experimental results. The separated dynamics backstepping controller presents also similar results, indicating AC and DC dynamics separation.

3.2. DC/AC Microgrid Simulation and Experimental Results in Transient Operation

The microgrid DC voltage U_{dc} applied to loads or distributed power sources may require regulation to suit the load needs, the renewable distributed power sources, or microgrid stability requirements. Results of transient microgrid operation are displayed for DC voltage varying set-points and for DC current i_{dc} variations.

When the DC reference voltage U_{dcRef} steps up from the lower operating value (100 V) to the upper operating value (120 V) at time $t = 0.5$ s, the simulated DC voltage U_{dc} transient behavior, using the averaged backstepping controller, is displayed in Figure 7a. The response time is found to be nearly 15 ms. Using the separate dynamics controller, the results are nearly the same, with the absence in Equation (23) of the DC voltage error, $e_{U_{dc}}$, having very little effect. For performance comparison purposes, Figure 7b shows the results of standard linear control based on SPWM, coupled to PI controllers, which has a response time of approximately 45 ms and some overshoot.

Figure 8a shows the simulated AC current behavior regarding the above U_{dcRef} step. The AC currents increase while the phase is conserved (rectifier operation) and present a rise time in the 1 ms range, in response to the increased power in the DC load. In the transient operation, the results do not show visible differences either using the averaged or the separated dynamics backstepping methods. This shows that the backstepping controllers can be designed using a fast inner loop current controller and a slow external loop DC voltage controller, which provides the values of the reference currents for the inner loop, as it is usually conducted when using linear PI controllers. The results for standard linear current control based on SPWM, coupled to PI controllers, as shown in Figure 8b, show that at the time (0.5 s) of the reference voltage variation U_{dcRef} , the controller response is slower regarding the averaged backstepping controller, leading to the U_{dc} voltage slower response.

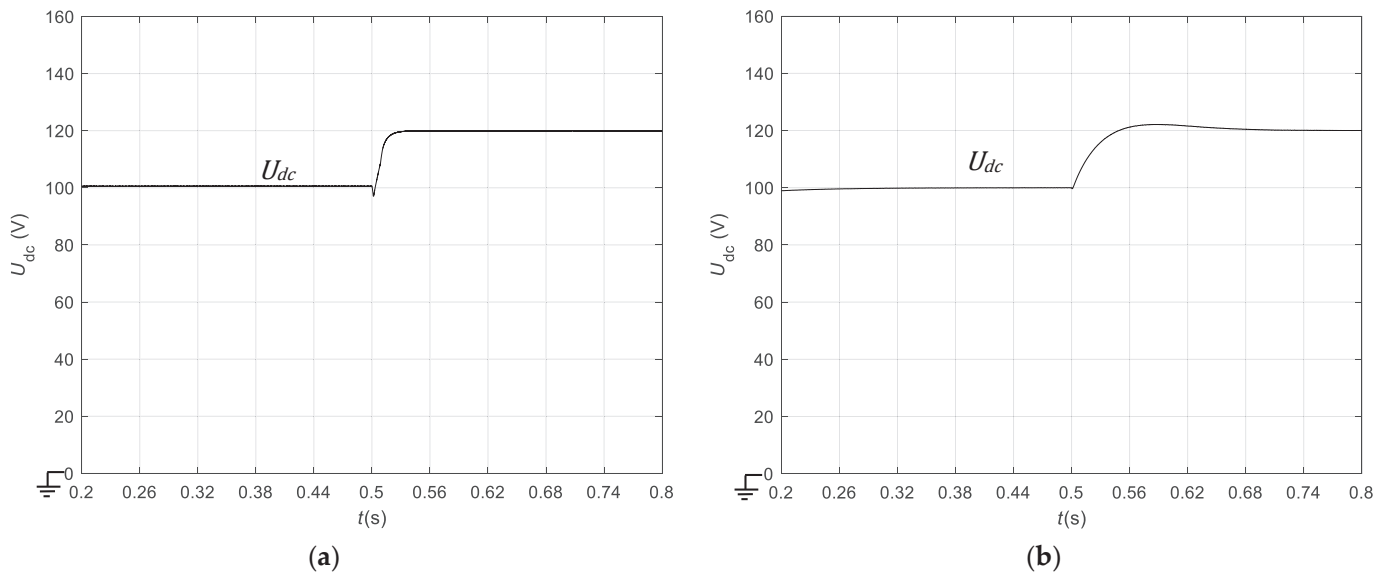


Figure 7. DC voltage simulation results in transient mode, when the reference changes from 100 V to 120 V: (a) With the averaged backstepping controller; (b) With PI controller.

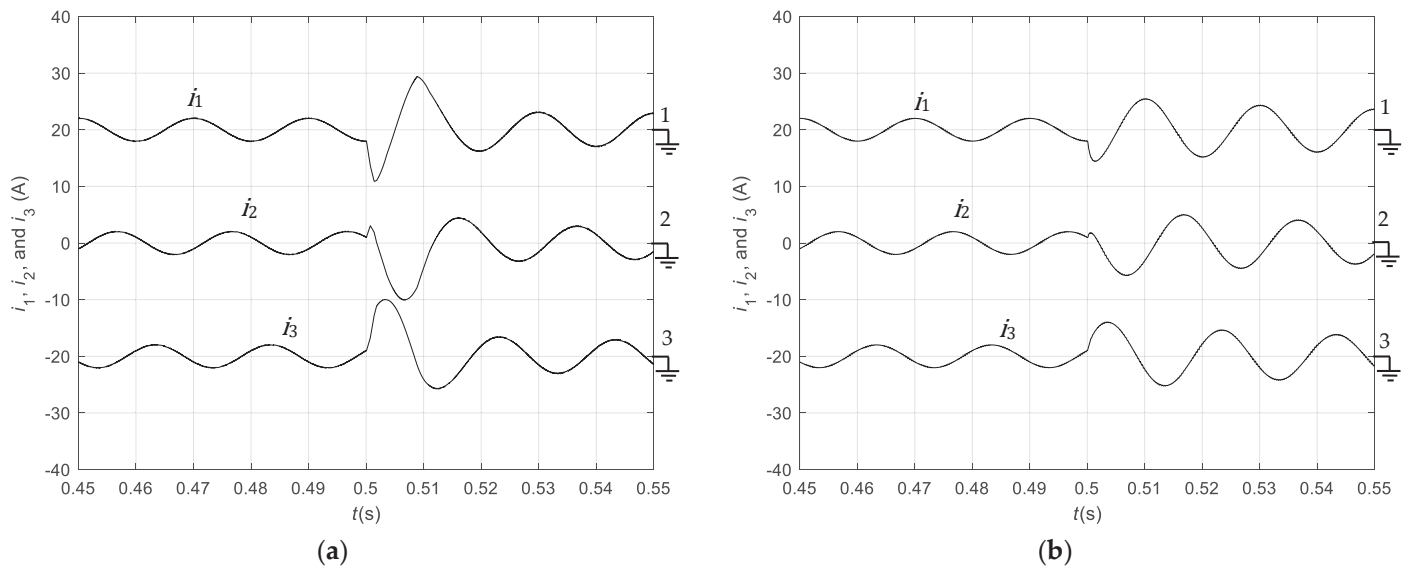


Figure 8. Simulation results of the AC currents, i_1 , i_2 , and i_3 , in transient mode, when the reference changes from 100 V to 120 V: (a) With the averaged backstepping controller; (b) With PI controller.

Simulation and experimental tests were conducted to evaluate the performance of the modified backstepping controllers to balance the DC capacitor U_{C1} , U_{C2} voltages in transient operation. Figure 9b shows that the capacitor voltages are balanced, when the reference DC voltage was changed from 120 V to 100 V. The DC voltage U_{dc} transition time is approximately 75 ms. The fall time is longer than the rise time (Figures 7 and 9), as the AC currents in Figure 9a show a fast phase reversal, transferring the DC capacitor's excess energy to the AC microgrid (inverter operation), correlated to the DC voltage decreasing ratio. As the U_{dc} voltage nears the new reference value, the currents rise again and change phase to transfer power to the DC grid (rectifier operation).

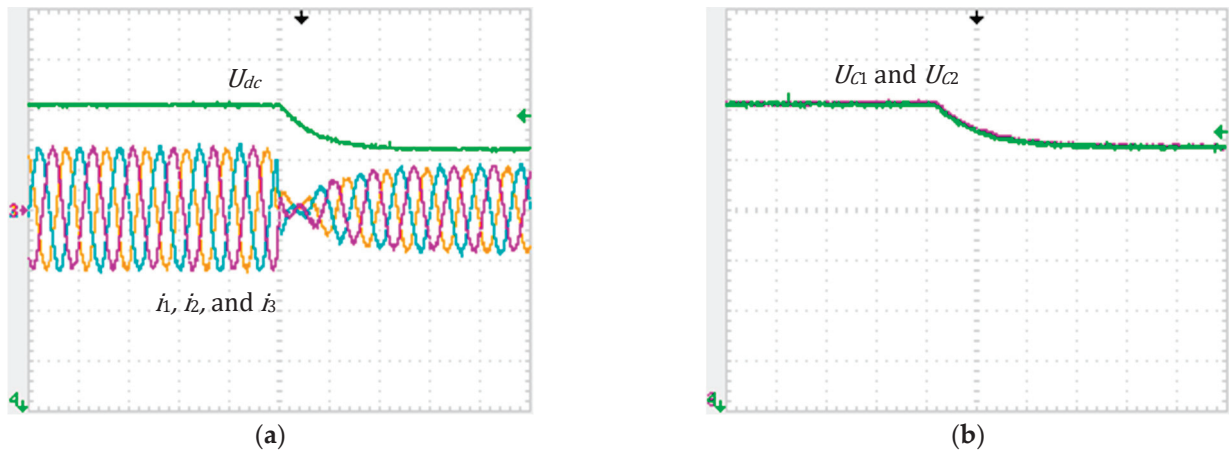


Figure 9. Experimental results in transient conditions when the reference changed from 120 V to 100 V, with the averaged backstepping controller: (a) DC voltage and AC currents, i_1 , i_2 , and i_3 , vertical: 2 A/div and 20 V/div; (b) Capacitor voltages, U_{C1} and U_{C2} , vertical: 10 V/div. Horizontal: 25 ms/div.

The experimental results of Figure 10 illustrate the response of separated dynamics backstepping controllers when DC loads change, leading to disturbances in the i_{dc} current. The U_{dc} voltage and the AC currents, i_1 , i_2 , and i_3 , are shown subjected to a step change in the i_{dc} current from 1.2 A to 2.4 A. The steep DC current increase causes a slight decrease (sag) of the U_{dc} voltage (Figure 10a) followed by a slow increase in U_{dc} , as the separated dynamics backstepping controller tries to track the input reference. Figure 10b illustrates the NPC converter behavior when subjected to a negative step disturbance in the i_{dc} current. The negative step current change increases momentarily the U_{dc} voltage (small swell), while the separated dynamics backstepping controller slowly tracks its reference. Results obtained using the averaged backstepping controller did not show improved performance, despite the presence of the DC voltage error, $e_{U_{dc}}$, in Equation (13).

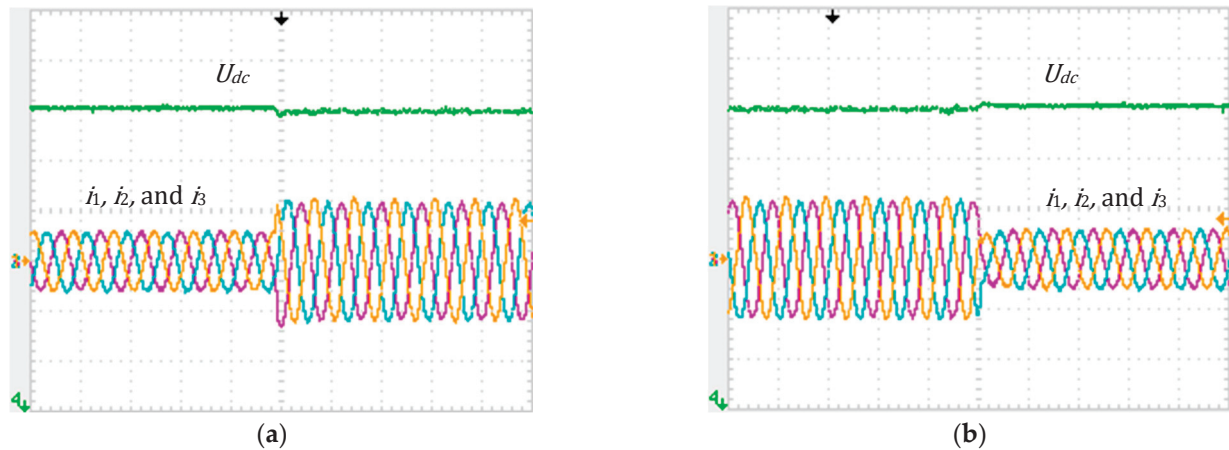


Figure 10. Experimental results of DC voltage and AC currents, i_1 , i_2 , and i_3 , with the separated dynamics backstepping controller: (a) Subjected to a step change in the i_{dc} current from 1.2 A to 2.4 A; (b) Subjected to a step change in the i_{dc} current from 2.4 A to 1.2 A. Vertical: 2 A/div and 20 V/div. Horizontal: 25 ms/div.

The averaged backstepping controller and the separated dynamics backstepping controller show an adequate regulation of DC voltage, capacitor voltage balance, AC current's THD, and reactive power control at $Q_{Ref} = 0$.

The microgrid DC voltage regulation can aid the power transfer of microgrid renewable energy sources, such as PV panels, energy storage batteries, and fuel cells, as both modified backstepping controllers are able to track the DC microgrid best reference voltage.

4. Conclusions

This paper presented a new method for obtaining backstepping controllers for multi-level NPC converters interconnecting AC to DC microgrids. The averaged backstepping controller is free from high-frequency harmonics, outputs very low total harmonic distortion (THD < 1.7%) AC currents, and very small steady-state DC errors (<0.5%). It also balances the two DC capacitor voltages, regulates the power factor, and shows bidirectional power transfer capability. The separated dynamics backstepping controller has virtually the same performance while needing less system information, as the DC voltage tracking error is not needed to generate the SPWM modulating waveform. The separated dynamics backstepping controller performance shows that, for NPC converters, backstepping controllers can be designed using a fast inner loop current controller and a slow external loop DC voltage controller.

To serve as the basis for the design of the backstepping controllers, a suitable NPC dynamic model was devised considering, as state variables, the AC currents in dq coordinates, the DC total voltage, and the unbalance voltage between the two capacitors.

The simulation and experimental results confirm that the AC currents exhibit sinusoidal behavior and track their reference values; the power factor is controllable using the q current component, while the total DC capacitor voltage is controlled using the d current component and tracks the reference value. To balance the DC capacitor voltages, an offset was added to the SPWM carriers. The converter is stable and bidirectional, able to regulate the DC voltage and convert energy from storage systems or DC renewable energy sources. The modified backstepping control algorithms are stable and cope with non-modeled disturbances, such as the DC current and typical component tolerances, which confirms the validity of using the control quantity steady-state averaged value for the backstepping design. The AC current's fast response and the DC voltage response times also indicate the validity of the assumption of using separated AC and DC dynamics.

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Appendix A

Derivation process for Equations (3) and (4). Analyzing the multilevel NPC converter circuit connecting DC loads and distributed power sources to the AC microgrid (Figure 1), the dynamic equation of the difference or unbalance between the two capacitor voltages can be written as:

$$\frac{dE_{UC}}{dt} = -\frac{1}{C}(\gamma_1^2 i_1 + \gamma_2^2 i_2 + \gamma_3^2 i_3). \quad (A1)$$

Considering their small signal responses to a Δ_γ offset added to the SPWM carrier, which is equivalent to changing variables, $\gamma_1 \rightarrow \gamma_1 + \Delta_\gamma$, $\gamma_2 \rightarrow \gamma_2 + \Delta_\gamma$ and $\gamma_3 \rightarrow \gamma_3 + \Delta_\gamma$ Equation (A1), as a function of Δ_γ , SPWM offset becomes:

$$\frac{dE_{UC}}{dt} = -\frac{1}{C}(\gamma_1 + \Delta_\gamma)^2 i_1 - \frac{1}{C}(\gamma_2 + \Delta_\gamma)^2 i_2 - \frac{1}{C}(\gamma_3 + \Delta_\gamma)^2 i_3. \quad (A2)$$

Applying the Clark–Concordia transform (A3) to γ_1 , γ_2 , and γ_3 and to i_1 , i_2 , and i_3 as follows:

$$\begin{bmatrix} x_1 \\ x_2 \\ x_3 \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & 0 & \frac{1}{\sqrt{2}} \\ -\frac{1}{2} & \frac{\sqrt{3}}{2} & \frac{1}{\sqrt{2}} \\ -\frac{1}{2} & -\frac{\sqrt{3}}{2} & \frac{1}{\sqrt{2}} \end{bmatrix} \begin{bmatrix} x_\alpha \\ x_\beta \\ x_0 \end{bmatrix}, \quad (\text{A3})$$

results in the dynamic equation of the difference or unbalance between the two capacitor voltages as a function of semiconductor states, γ_α , γ_β , and γ_0 and AC currents i_α and i_β :

$$\frac{dE_{UC}}{dt} = \left(-\frac{2}{C} \gamma_\alpha \Delta_\gamma - \frac{1}{\sqrt{6}} \frac{\gamma_\alpha^2}{C} + \frac{\sqrt{2}}{6} \frac{\gamma_\beta^2}{C} - \frac{2}{\sqrt{3}} \frac{\gamma_\alpha \gamma_\beta}{C} \right) i_\alpha + \left(-\frac{2}{C} \gamma_\beta \Delta_\gamma + \sqrt{\frac{2}{3}} \frac{\gamma_\alpha \gamma_\beta}{C} - \frac{2}{\sqrt{3}} \frac{\gamma_\beta \gamma_0}{C} \right) i_\beta. \quad (\text{A4})$$

The dq dynamic equation of the two capacitor voltages difference in Equation (A6) is obtained by applying the Park transform in Equation (A5) to Equation (A4):

$$\begin{bmatrix} x_\alpha \\ x_\beta \\ x_0 \end{bmatrix} = \begin{bmatrix} \cos\omega t & -\sin\omega t & 0 \\ \sin\omega t & \cos\omega t & 0 \\ 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} x_d \\ x_q \\ x_0 \end{bmatrix} \quad (\text{A5})$$

$$\begin{aligned} \frac{dE_{UC}}{dt} = & \left(-\frac{2}{C} \gamma_d \Delta_\gamma + \frac{\gamma_d^2 \cos\omega t}{C} \left(\frac{1+\sqrt{12}}{\sqrt{18}} \sin^2\omega t - \frac{\cos^2\omega t}{\sqrt{6}} \right) \right. \\ & + \frac{\gamma_q^2 \cos\omega t}{C} \left(\frac{\sqrt{2}}{6} \cos^2\omega t - \frac{\sin^2\omega t}{\sqrt{6}} \right) \\ & + \frac{\gamma_d \gamma_q \sin\omega t}{C} \left(\frac{4\sqrt{6}+2\sqrt{2}}{6} \cos^2\omega t - \sqrt{\frac{2}{3}} \sin^2\omega t \right) - \frac{\gamma_d \gamma_0}{C} \frac{2}{\sqrt{3}} \Big) i_d \\ & + \left(-\frac{2}{C} \gamma_q \Delta_\gamma + \frac{\gamma_d^2 \sin\omega t}{C} \left(\frac{3}{\sqrt{6}} \cos^2\omega t - \frac{\sqrt{2}}{6} \sin^2\omega t \right) \right. \\ & + \frac{\gamma_q^2 \sin\omega t}{C} \left(\frac{\sin^2\omega t}{\sqrt{6}} - \frac{\sqrt{2}+\sqrt{24}}{6} \cos^2\omega t \right) \\ & \left. + \frac{\gamma_d \gamma_q \cos\omega t}{C} \left(\sqrt{\frac{2}{3}} \cos^2\omega t - \frac{4\sqrt{6}+2\sqrt{2}}{6} \sin^2\omega t \right) - \frac{\gamma_q \gamma_0}{C} \frac{2}{\sqrt{3}} \right) i_q. \end{aligned} \quad (\text{A6})$$

Defining the quantities Δ_d in Equation (A7) (or Equation (3)) and Δ_q in Equation (A8) (or Equation (4)), we obtain Equation (A6) in compressed format in Equation (A9) to insert in the NPC dynamic model Equation (2):

$$\begin{aligned} \Delta_d = & \frac{\gamma_d^2 \cos\omega t}{C} \left(\frac{1+\sqrt{12}}{\sqrt{18}} \sin^2\omega t - \frac{\cos^2\omega t}{\sqrt{6}} \right) \\ & + \frac{\gamma_q^2 \cos\omega t}{C} \left(\frac{\sqrt{2}}{6} \cos^2\omega t - \frac{\sin^2\omega t}{\sqrt{6}} \right) + \\ & \frac{\gamma_d \gamma_q \sin\omega t}{C} \left(\frac{4\sqrt{6}+2\sqrt{2}}{6} \cos^2\omega t - \sqrt{\frac{2}{3}} \sin^2\omega t \right) - \frac{\gamma_d \gamma_0}{C} \frac{2}{\sqrt{3}} \end{aligned} \quad (\text{A7})$$

$$\begin{aligned} \Delta_q = & \frac{\gamma_d^2 \sin\omega t}{C} \left(\frac{3}{\sqrt{6}} \cos^2\omega t - \frac{\sqrt{2}}{6} \sin^2\omega t \right) \\ & + \frac{\gamma_q^2 \sin\omega t}{C} \left(\frac{\sin^2\omega t}{\sqrt{6}} - \frac{\sqrt{2}+\sqrt{24}}{6} \cos^2\omega t \right) + \\ & \frac{\gamma_d \gamma_q \cos\omega t}{C} \left(\sqrt{\frac{2}{3}} \cos^2\omega t - \frac{4\sqrt{6}+2\sqrt{2}}{6} \sin^2\omega t \right) - \frac{\gamma_d \gamma_0}{C} \frac{2}{\sqrt{3}} \end{aligned} \quad (\text{A8})$$

$$\frac{dE_{UC}}{dt} = \left(-\frac{2}{C} \gamma_d \Delta_\gamma + \Delta_d \right) i_d + \left(-\frac{2}{C} \gamma_q \Delta_\gamma + \Delta_q \right) i_q \quad (\text{A9})$$

Appendix B

The derivation process of Equation (13) begins with the first term simplification of Equation (11), using Equation (5):

$$e_{Udc} \frac{de_{Udc}}{dt} = e_{Udc} \frac{dU_{dcRef}}{dt} - e_{Udc} \frac{dU_{dc}}{dt}. \quad (\text{A10})$$

Using the DC voltage dynamics equation U_{dc} (2) and Equation (9), to represent the current i_d as a function of the virtual control value i_{dV} and error e_{id} , the Equation (A11) is obtained:

$$e_{Udc} \frac{de_{Udc}}{dt} = e_{Udc} \frac{dU_{dcRef}}{dt} - e_{Udc} \left(-\frac{\gamma_d}{C} i_{dV} + \frac{\gamma_d}{C} e_{id} - \frac{\gamma_q}{C} i_q + \frac{2}{C} i_{dc} \right) \quad (A11)$$

By replacing the virtual control action i_{dV} (8) in Equation (A11) and doing some simplifications, the following equation,

$$e_{Udc} \frac{de_{Udc}}{dt} = -K_{eUdc} e_{Udc}^2 - e_{Udc} \frac{\gamma_d}{C} e_{id}, \quad (A12)$$

is obtained. Using this term in Equation (11) gives,

$$\frac{de_{id}}{dt} = -K_{eid} e_{id} + \frac{\gamma_d}{C} e_{Udc}. \quad (A13)$$

Rewriting Equation (A13) as a function of current i_d and the virtual control current i_{dV} , considering Equation (9), and using the i_d current dynamics di_d/dt , from Equation (2), the control component, γ_{dRef} , for the averaged backstepping is obtained (Equation (A14)) (or Equation (13)):

$$\gamma_{dRef} = \frac{2L}{U_{dc}} \left(K_{eid} e_{id} - \frac{\gamma_d}{C} e_{Udc} + \frac{di_{dV}}{dt} + \frac{R}{L} i_d - \omega i_q + \frac{U_{Ld}}{L} \right) \quad (A14)$$

Appendix C

Derivation process for Equation (20). By writing Equation (19) as the difference or unbalance between the two capacitor voltages E_{UC} ,

$$\frac{dE_{UC}}{dt} = -K_{EUC} E_{UC} \quad (A15)$$

and using the dynamics, dE_{UC}/dt , from Equation (2), as well as considering a unity power factor ($i_q \approx 0$), the value of the offset Δ_γ equals its reference value $\Delta_{\gamma Ref}$, and is obtained in the triangular SPWM modulator,

$$\Delta_{\gamma Ref} = \frac{C}{2i_d \gamma_d} (K_{EUC} E_{UC} + \Delta_d i_d). \quad (A16)$$

To avoid division by $i_d \gamma_d$, the steady-state voltage dynamics equation is used ($dU_{dc}/dt \approx 0$) and unity power factor, obtaining:

$$i_d \gamma_d \approx 2i_{dc} \quad (A17)$$

where replacing in Equation (A16) gives the SPWM modulator offset Equation (A18) (or Equation (20)) to balance the two capacitor voltages:

$$\Delta_{\gamma Ref} = \frac{C}{4i_{dc}} (K_{EUC} E_{UC} + \Delta_d i_d) \quad (A18)$$

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Article

Line-Start Permanent Magnet Synchronous Motor Supplied with Voltage Containing Negative-Sequence Subharmonics

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Abstract: In some power systems, the voltage waveform contains frequency components less than fundamental, called subharmonics or subsynchronous interharmonics. Voltage subharmonics can be both positive- and negative-sequence, independent of their frequency (order). Subharmonics exert harmful effects on sundry electrical equipment, especially on rotating machinery; they cause various noxious phenomena, such as a local saturation of the magnetic circuit, increases in power losses and windings temperature, and torque pulsations leading to vibration of unacceptable severity. Notably, previous works reported excessive vibration of rotating machinery only under no-load, while under full load, rather moderate vibration occurred. This study deals with vibration analysis of a line-start permanent magnet synchronous motor (LSPMSM) supplied with the voltage containing negative-sequence subharmonics. Experimental investigations were conducted for a 3 kW, four-pole production LSPMSM for subharmonics of various values and frequencies. Voltage subharmonics of values significantly less than reported in real power systems were found to cause unacceptable vibration, especially under full load.

Keywords: AC motors; interharmonics; permanent magnet motors; power quality; subharmonics; synchronous motors; vibration; voltage distortions

1. Introduction

A prevalent power quality disturbance is from voltage waveform distortions. In practice, the voltage waveform is not a perfect sine but contains various additional frequency components. The undesirable components are usually harmonics, but in some systems [1–3], subharmonics (subsynchronous interharmonics) and interharmonics (SaIs) also occur—that is, components of a frequency less than fundamental or those that are not integer multiples of the fundamental frequency. Of note, subharmonics are often understood as a kind of interharmonics, but in some works (e.g., [4–6]), they are considered separately because of their especially negative effect on electrical equipment [7].

The presence of SaIs in a power system usually originates from the work of various power electronic appliances, renewable sources of energy, and loads of a fluctuating nature, including double-frequency conversion systems (inverters and high-voltage DC links), wind power stations, photovoltaic plants, arc furnaces, laser printers, AC motors driving piston compressors and other time-varying loads [2,8–22]. SaIs can also occur in an inverter output voltage because of voltage fluctuations in the DC link [6,13,23]. In fact, cyclic voltage fluctuations [7,24–26] are a superposition of the fundamental voltage component and SaIs [7,27,28].

Various works have reported voltage subharmonics of values up to about 1% in real power systems. For example, in [1], significant SaIs were observed in a building located near steelworks, containing numerous non-linear receivers. The maximal value of the

squared subharmonics subgroups of frequencies 5, 10, . . . , and 45 Hz was 0.99% for the 10-minute aggregation time. Authors in [3] analysed a subharmonic resonance of a wind farm. During the resonance, a voltage subharmonic of 8.1 Hz with a value of 1–2% occurred (based on correspondence with the authors of [3]). Furthermore, [2] reported a subharmonic value of 0.9% and frequency of 45 Hz, accompanied by interharmonics values of 1.17%, 0.89%, and 0.931% and frequencies of 135, 225, and 405 Hz, respectively. SaIs contamination was found in a system supplied with diesel generators, caused by high-power top-drive and mud-pump variable frequency drives.

As is well known, the sequence of voltage harmonics is linked to their order (in practice, frequency). Contrastingly, voltage SaIs can be both positive- and negative-sequence, independent of their frequency [21]. In practice, the sequence of SaIs is interconnected with their origination [21].

Subharmonics and, interconnected with them, voltage fluctuations cause various noxious phenomena in transformers, light sources, control systems, and power electronic and measurement equipment [4,7,16,24,27,29,30]. Especially sensitive to subharmonics is rotating machinery (based on [5,14,23,27,31–39]). Subharmonics have been reported to cause excessive transverse and torsional vibrations, increases in power losses and windings temperatures, local saturations of magnetic circuits, and speed fluctuations. Particularly high levels of vibration, even leading to drivetrain destruction [12,18], occur under a resonance condition [5,12,18,24,27,33,34]. The most exposed to harmful torsional vibration are drivetrains containing large synchronous generators and motors (based on [10,14]). As a result, allowable levels of voltage SaIs can significantly differ for various installations (e.g., low- vs. medium-voltage generation installations).

Power quality standards generally do not include permissible levels of voltage SaIs. For instance, Standard EN 50160: Voltage characteristics of electricity supplied by public distribution systems [38] states, “The level of interharmonics is increasing due to development of the application of frequency converters and similar control equipment. Levels are under consideration, pending more experience”. Of note, the comment concerns both interharmonics and subharmonics, understood as subsynchronous interharmonics. IEEE-519:2022 Standard for harmonic control in electric power systems [39] discusses the necessity of SaIs limitation, presenting two alternative proposals of limiting curves for non-generation installations. One curve generally limits subharmonics to 0.5%, and the other limits them to 0.3%.

Rotating machinery supplied with voltage containing SaIs is the subject of numerous research works. They usually concern either multi-megawatt powertrains [10,14] or induction motors [5,23,28,31–35]. Furthermore, the effect of SaIs on the line-start permanent magnet synchronous motor (LSPMSM) [40,41] was only preliminarily analysed in the authors’ works [36,37]. Of note, the investigations [36,37] were restricted to currents and vibration under positive-sequence subharmonics. However, determining the permissible levels of voltage SaIs requires comprehensive investigations of their effect on various elements of a power system, including the increasingly popular LSPMSM—a possible successor to the induction motor.

In addition, previous research [5,35–38] indicated high transverse vibration of the rotating machinery for no-load, while in other load conditions, the observed vibration was generally moderate. At the same time, at least temporarily, many motors carry a load much less than rated [42,43], but this is not typical. In practice, most motors usually work with significantly greater loads [43].

In light of the above considerations, the objectives of this study are formulated as follows: (1) to demonstrate that voltage subharmonics can cause high vibration of rotating machinery for both no-load and full load, and (2) to examine the vibration of the LSPMSM under negative-sequence voltage subharmonics. The results of this research can be useful for power quality standard modification.

2. Origin of Voltage Subharmonics and Interharmonics

The presence of voltage SaIs in the power system usually originates from the work of power electronic equipment, renewable energy sources, and time-varying loads.

The most important SaIs source is considered to be a double-frequency conversion system (DCS) [13], which is a power electronic appliance connecting two AC systems of various frequencies [16]. The DCS comprises a rectifier and an inverter interconnected by a DC link [13,16,21]—in practice, a capacitor or an inductor. During the DCS operation, voltage or current ripples occur in the DC link, which may result in SaIs presence in both input and output sides of the DCS [13,16,21,23]. The frequencies of subharmonics (f_{sh}) and interharmonics (f_{ih}) occurring in the input side of the DSC are (based on [16,21])

$$f_{sh/ih} = f_1 \mp f_r, \quad (1)$$

$$f_{sh/ih} = f_h \mp f_r, \quad (2)$$

where f_1 is the fundamental frequency, f_r is the frequency of ripples occurring in the DC link, and f_h is the harmonic frequency characteristic for the given kind of converter, for example, 250 Hz or 350 Hz (in a 50-Hz system). In practice, the DCS containing a p_1 -pulse rectifier and a p_2 -pulse inverter produces SaIs of frequencies equal to [16,21]

$$f_{sh/ih} = (p_1 m \mp 1) f_1 \mp p_2 n f_o \quad (3)$$

where f_o is the output frequency of the DCS, $m = 0, 1, 2, \dots$, $n = 1, 2, 3, \dots$

The above dependency (3) can be likewise applied to cycloconverters [16]. Of note, the PWM inverter produces additional interharmonics, whose frequency depends on the modulation frequency, strategy, and depth [15,16].

SaIs generated by DCSs can be of both positive and negative sequences. In other words, if the frequency described by (1) is greater than zero, the resulting subharmonic or interharmonic component is of positive sequence. Similarly, if the frequency calculated with (2) is positive, the SaI sequence is the same as the same sequence of the harmonic frequency f_h [21]. On the contrary, if the frequency determined with (1) or (2) is negative, the SaI sequence is the opposite of the sequence of the fundamental or harmonic component of frequency f_h [21], respectively.

Furthermore, ripples occurring in the DC link can be considerably amplified by resonance phenomena, resulting in high SaI contamination. For example, ripple amplification was reported [13] for inverter output voltage fluctuations corresponding to the natural torsional frequency of a power train. This phenomenon can be especially significant for the low inertia moment of a driven appliance [13].

Another reason for SaI occurrence is the work of renewable energy sources, such as photovoltaic plants [18,22] and wind power stations [3,7,12,20]. In the case of wind turbines, the presence of voltage fluctuations and subharmonics originates from such sources as tower resonance, blades passing in the front of the tower, small differences in the blades, wind turbulences, and rotor electrical or mechanical asymmetry [7,12]. Another SaI source is the DCS used in some wind turbines. In addition, in plants equipped with double-fed induction generators, the harmonics of the rotor circuit converter can be transmitted to the stator as SaIs [12].

Especially high SaI contamination [3] may occur during the sub-synchronous interaction of a power plant [3,20]. Of note, the interaction is not characteristic only for wind turbines but may also appear in thermal power stations [20]. The interaction may be categorised as subharmonic resonance and device-dependent sub-synchronous oscillations [20]. It can be either a purely electrical phenomenon or an event involving the torsional resonance of the generator rotating mass [20]. The sub-synchronous interaction can be initiated by a resonance of an LC circuit consisting of various system elements, including compensation capacitors, transformers, and generators, especially after the failure of a transmission

line [20]. Additionally, subharmonic resonance can be triggered by the interaction of the current loop controller of the converter of a double-fed induction generator or the interaction between a generator and various power electronic appliances, such as high-voltage DC links, voltage source converters, and power system stabilisers [20].

A prominent source of SaIs includes time-varying loads, which can be divided into regularly fluctuating loads, e.g., welder machines, laser printers, and devices with integral cycle control, and irregularly fluctuating loads, such as arc furnaces [16]. The work of time-varying loads causes current modulation [7,16], resulting in voltage fluctuations [7,25,26,35]. At the same time, sinusoidal voltage fluctuations can be analysed as a superposition of the fundamental voltage component, a subharmonic component, and an interharmonic component of the following frequencies [7,27,31,35]:

$$f_{sh} = f_1 - f_m \quad (4)$$

$$f_{ih} = f_1 + f_m \quad (5)$$

where f_{sh} is the frequency of the subharmonic component, f_{ih} is the frequency of the interharmonic component, and f_m is the frequency of voltage fluctuations.

If the frequency described with (4) or (5) is greater than zero, the corresponding subharmonic/interharmonic component is of positive sequence; otherwise, it is of negative sequence (based on [21]). Additionally, periodic voltage fluctuations of any shape can be considered as the simultaneous occurrence of sinusoidal modulations of various frequencies [31].

A specific fluctuating load is the arc furnace, whose feature is the variation of the generated SaIs because of the chaotic arc behaviour [16]. Of note, for the AC arc furnace, the most significant SaIs occur around the fundamental frequency [16,19], while for the DC arc furnace, they occur near the harmonics produced by the rectifier [16]. The arc furnace generates both positive and negative sequence SaIs (based on [16,19,21]).

Other specific fluctuating loads include synchronous and asynchronous motors driving appliances of the pulsating anti-torque, e.g., the piston compressor [8]. The motors may inject especially significant SaI currents when the frequency of the torque pulsations corresponds to the natural torsional frequency of the rigid-body mode [8]. As an example, a 12 MW synchronous motor under an anti-torque pulsation of 50% of the rated value produced an SaI current of frequencies $f_{sh} = 43$ Hz, $f_{sh} = 57$ Hz, and values of about 10% of the rated current [8].

Also noteworthy are some receivers and components of the power system that cannot produce SaIs for themselves but generate them when the supply voltage contains a subharmonic or an interharmonic (based on [5,28]). For instance, the induction motor supplied with voltage containing a positive-sequence interharmonic injects a subharmonic of a frequency equal to $2f_1 - f_{ih}$ into the power system [5,28]. Furthermore, the induction motor under a positive-sequence subharmonic generates an interharmonic of a frequency equal to $f_1 + f_{sh}$ [5,28]. In one study, a 3 kW induction motor supplied with the voltage containing the subharmonic value of 1% and frequency $f_{sh} = 26$ Hz injected a current interharmonic of frequency $f_{sh} = 76$ Hz and value of 23% of the rated current into the power system [5]. This phenomenon is especially significant in the conditions of rigid-body torsional resonance caused by SaI presence in the supply voltage [5].

In summary, both positive- and negative-sequence SaIs can be generated by various commonly used electrical appliances.

3. Measurement Setup

The experimental equipment comprised the LSPMSM, a multi-machine system for SaIs generation, power quality analysers and a Bruel&Kjaer (B&K) system for vibration measurement. The system included a computer with the installed B&K Connect software (version 26.1.0.251), a standalone four-channel data acquisition module B&K type 3676-B-040, an accelerometer calibrator B&K type 4294, and a three-axis accelerometer B&K

type 4529-B (frequency range of 0.3–12,800 Hz, a sensitivity of 10 mV/ms^{-2} , maximum shock level peak of 5100 g, and weight of 14.5 g). Before each measurement session, the accelerometer was calibrated, and the measurements were taken and analysed with the chief provisions of the standard ISO 20816-1 Mechanical vibration—Measurement and evaluation of machine vibration—Part 1: General guidelines [44]. The accelerometer indications were recorded and recalculated into the broad-band vibration velocity [44] using the B&K Connect software (version 26.1.0.251).

The LSPMSM tested was type WQuattro L100L-04 with 3 kW rated power. It was loaded with a DC machine (working as a generator), supplying a resistor bank. Table 1 lists the nameplate parameters of the LSPMSM, and Figure 1 shows a photograph with the mounted accelerometer. Because of some modifications, the results of vibration measurements presented in this study cannot be directly compared with previous research [36,37].

Table 1. Nameplate Data of the Investigated Motor WQuattro L100L-04.

Parameter	Value
Rated power (kW)	3
Rated frequency (Hz)	50
Rated voltage (V)	400
Rated current (A)	5.84
Rated power factor (—)	0.82
Rated rotational speed (rpm)	1500
Manufacturer	WEG Industries

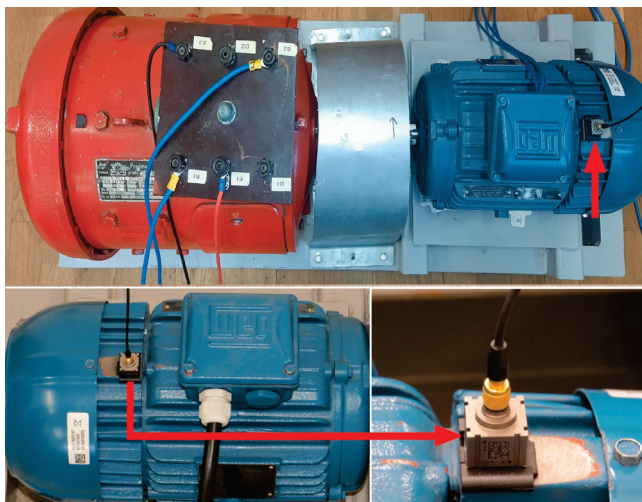


Figure 1. Investigated motor and accelerometer.

The multi-machine system for subharmonic generation (based on [45]) comprised two synchronous generators coupled via a transformer with an experimentally selected number of turns. One generator (type Leroy Somer LSA 42.2M6 6/4, rated power $S_{rat} = 23 \text{ kVA}$, rated current $I_{rat} = 33 \text{ A}$, direct-axis synchronous reactance $x_d = 190\%$, quadrature-axis synchronous reactance $x_q = 100\%$) was driven with a rotational speed corresponding to the fundamental frequency, and the other (type GTNSa132s2/08, $S_{rat} = 6.5 \text{ kVA}$, $I_{rat} = 9.9 \text{ A}$) was driven at the subharmonic frequency [5]. Voltage and current subharmonics were measured with a computer-based power quality analyser (with software elaborated by the authors, employing fast Fourier transform) and an estimator–analyser of power quality [46] developed at Gdynia Maritime University for commercial purposes and certified by the Polish Register of Shipping.

Figures 2 and 3 present a simplified diagram of measurement setup [36] and the flowchart of the vibration measurement (based on [35]), respectively.

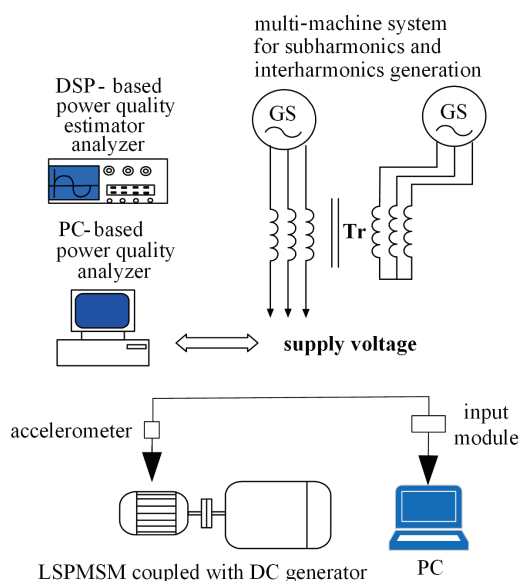


Figure 2. Simplified diagram of the measurement stand.

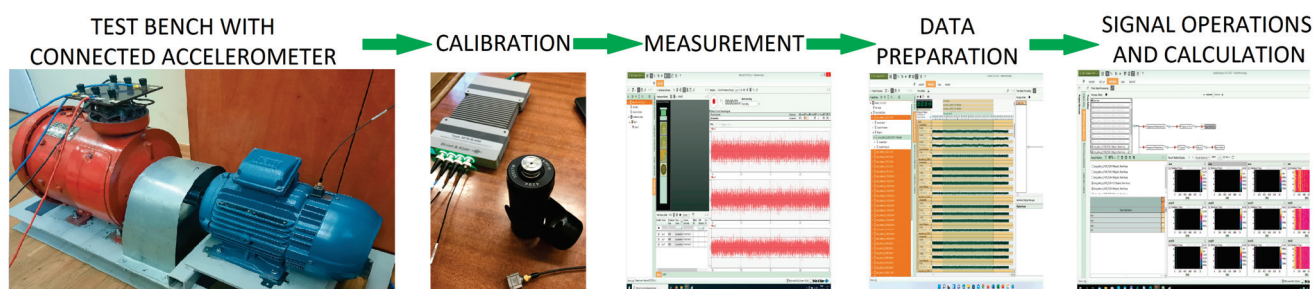


Figure 3. Flowchart of the vibration measurement.

4. Results

4.1. Assessment of Vibration Severity

Recommendations concerning the assessment of vibration severity are provided in the standard ISO 20816-1 Mechanical vibration—Measurement and evaluation of machine vibration—Part 1: General guidelines [44]. The standard [44] specifies four evaluation zones: Zone A, Zone B, Zone C, and Zone D. Vibrations corresponding to Zone A are generally characteristic of newly commissioned machines. Vibrations falling into Zone B are acceptable for long-term unrestricted operation, while those within Zone C can be accepted for a limited period, for example, until machine repair or replacement. Finally, vibrations within Zone D “are normally considered to be of sufficient severity to cause damage to the machine” [44].

As the threshold values of each evaluation zone are not explicitly specified in the current standard [44], for this study, they were assumed based on its previous edition—ISO 10816-1 Mechanical vibration -- Evaluation of machine vibration by measurements on non-rotating parts — Part 1: General guidelines [47]. For low-power electric motors, vibrations of broad-band velocity [44,47] up to 0.71 mm/s correspond to Zone A, between 0.71 and 1.8 mm/s correspond to Zone B, between 1.8 and 4.5 mm/s correspond to Zone C, and exceeding 4.5 mm/s correspond to Zone D.

4.2. Vibration under NoLoad

This subsection presents the results of investigations for the LSPMSM coupled with an unloaded DC machine. Figures 4 and 5 show the broad-band vibration velocity for subharmonics of values $u_{sh} = 0.5\%$ and $u_{sh} = 1\%$, respectively. As mentioned in Section 1, the subharmonics of values circa 1% were reported in real power systems [1–3]. As shown in Figures 4 and 5, the highest vibration occurred for $f_{sh} = 15$ Hz. The maximal vibration velocity reached 2.70 and 3.755 mm/s for $u_{sh} = 0.5\%$ (Figure 4) and $u_{sh} = 1\%$ (Figure 5), respectively, falling into Zone C. For frequencies f_{sh} below ~ 10 Hz and greater than 20 Hz, the vibration velocity was much lower, generally within the limits of Zone B. Furthermore, the vibration velocities for the supply voltage without subharmonics (reference test) were 0.739 mm/s, 0.512 mm/s and 0.634 mm/s in the vertical, horizontal, and axial directions, respectively.

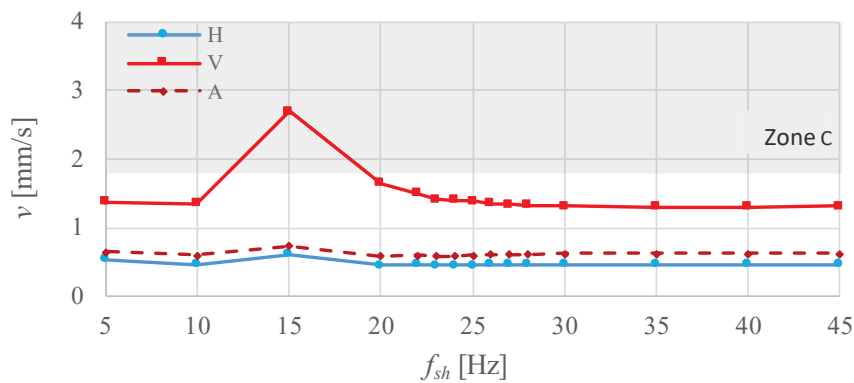


Figure 4. Measured broad-band vibration velocity in the horizontal (H), vertical (V), and axial (A) directions versus the voltage subharmonic frequency for subharmonic value $u_{sh} = 0.5\%$.

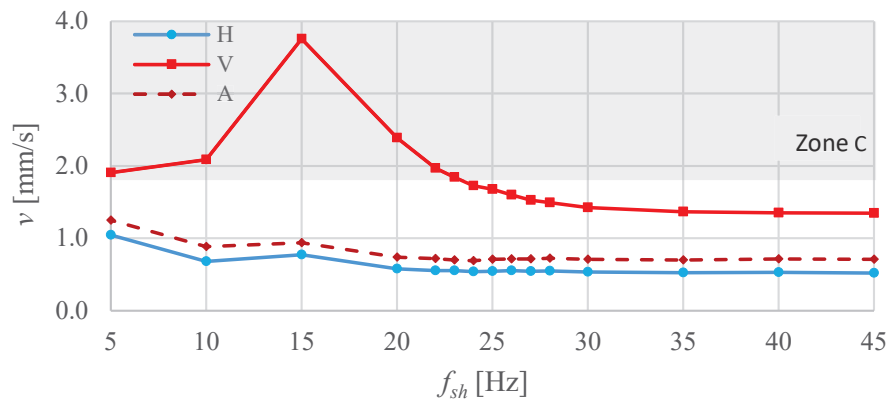


Figure 5. Measured broad-band vibration velocity in the horizontal (H), vertical (V), and axial (A) directions versus the voltage subharmonic frequency for subharmonic value $u_{sh} = 1\%$.

The primary source of excessive vibration under power quality disturbances is tangential electromagnetic forces (based on [5,48]), interconnected with the torque pulsations. In the case of rotating machinery supplied with voltage containing SaIs, torque pulsations are chiefly caused by the flow of current SaIs through the windings (based on [35]). In practice, SaI currents produce their own rotating magnetic field, interacting with the field because of the fundamental current harmonic [28].

An exemplary waveform of the supply motor current is shown in Figure 6 for the voltage subharmonic frequency $f_{sh} = 15$ Hz and value $u_{sh} = 1\%$, with the spectrum of the current waveform presented in Figure 7. The current waveform contained a subharmonic of value $I_{sh} = 16.72\%$ of the rated current (I_{rat}) and an interharmonic of frequency 115 Hz and value $I_{ih} = 2.86\%$ of I_{rat} and some minor components of frequency 10 Hz, 20 Hz, 85 Hz,

115 Hz and value 0.7%, 1.23%, 0.74%, 3.71% of I_{rat} , respectively. Next, Figure 8 presents the main current subharmonic component versus its frequency f_{sh} . The characteristic generally decreased as the frequency f_{sh} rose, and did not contain current peaks, characteristic of rigid-body resonance of the rotating mass [5,35–37]. For example, for the same power train (LSPMSM coupled with the unloaded DC machine) and positive-sequence subharmonics, the resonant peak occurred at the frequency $f_{sh} = 10$ Hz [36]. In practice, for this f_{sh} , the current subharmonic was higher by ~50% than those for $f_{sh} = 5$ and 15 Hz [36].

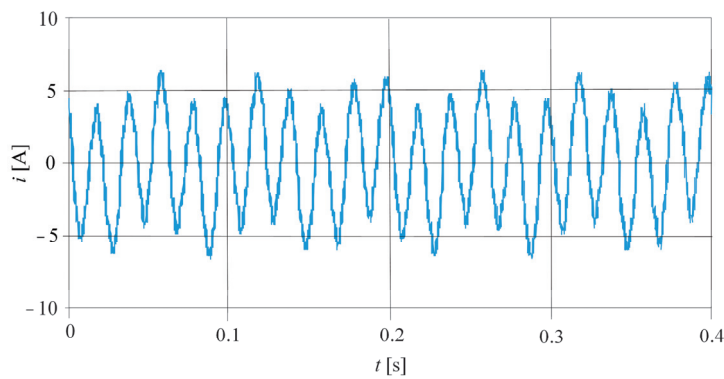


Figure 6. Measured current waveform for the subharmonic of frequency $f_{sh} = 15$ Hz and value $u_{sh} = 1\%$.

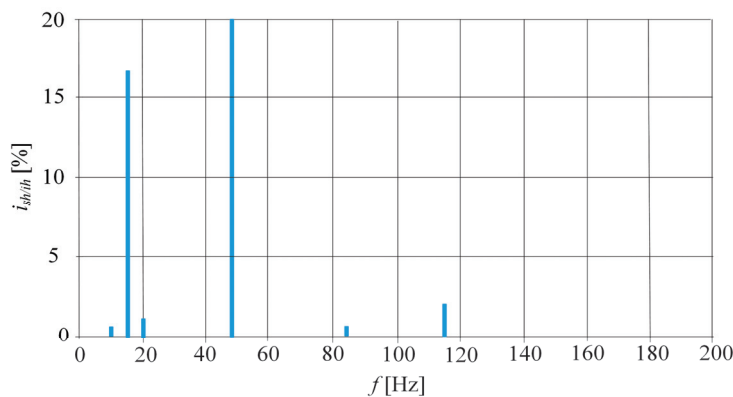


Figure 7. Spectrum of the current waveform presented in Figure 6. The frequency components are related to the rated current.

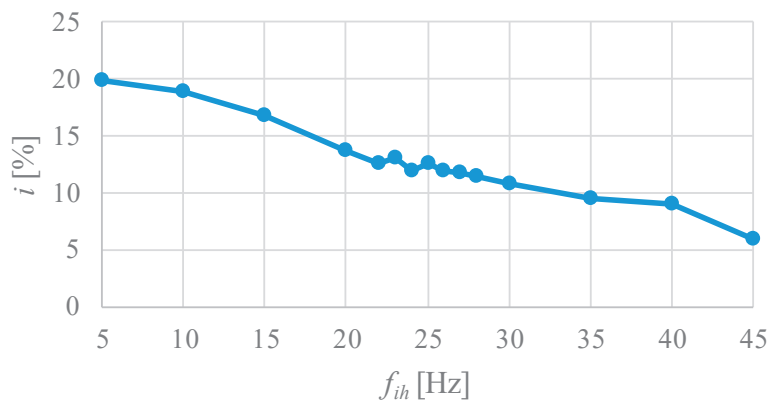


Figure 8. Measured current subharmonics versus their frequency for subharmonic value $u_{sh} = 1\%$, related to the rated current.

Although the observed vibration (Figures 4 and 5) was indirectly induced by the flow of the distorted current, the collocation of Figures 4, 5 and 8 shows that the vibration velocity was not directly interconnected with the value of current subharmonics. In practice, the vibration severity “directly depends on the mechanical behavior of the motor structure and the possibility of a resonance condition... on the structure of an entire unit or on the motor components” [48].

In summary, for the investigated motor, the negative-sequence voltage subharmonics resulted in high vibration around the resonant frequency of the mechanical structure, while for other frequencies, the vibration was comparatively low. Rigid-body resonance of the rotating mass was not observed.

4.3. Vibration under Rated Load

This subsection presents the results of the investigation on the vibration velocity under full load. Similarly to the previous subsection, the appropriate experimental investigations were carried out for the subharmonic of values $u_{sh} = 0.5\%$ and 1% . Additionally, the reference test was performed for the supply voltage without subharmonics; the measured vibration velocities were 1.136 mm/s, 0.825 mm/s, and 1.049 mm/s for the vertical, horizontal, and axial directions, respectively.

Figures 9 and 10 show the characteristics of the broad-band vibration velocity versus the subharmonic frequency f_{sh} . The maximal vibration velocity reached 5.058 mm/s for $u_{sh} = 0.5\%$ (Figure 9) and 7.344 mm/s for $u_{sh} = 1\%$ (Figure 10). Notably, these values were considerably greater than under no load (see Section 4.2). In contrast to the no-load condition, the vibration velocity took high values for the wide frequency range. For both the subharmonic values, almost all measurement points (except the vibration velocity for the frequency $f_{sh} = 45$ Hz) corresponded to Zones C and D. Of note, in Figures 9 and 10, the maximal vibration velocity occurred for the frequency $f_{sh} = 25$ Hz, while for no load, the highest vibration velocity was observed for the frequency $f_{sh} = 15$ Hz. According to the authors’ experience, the vibration under no load and full load significantly diverges (e.g., [5,35–37]). The differences may result from the non-linear coupling properties, among others. In particular, the coupling stiffness depends on the twist angle, which, in practice, is the carried torque. In addition, the anti-torque produced by the DC generator causes a change in the stress distribution of the structural components. At the same time, the vibration is extremely sensitive to the variation in any agents influencing the mechanical structure; a seemingly negligible change may significantly affect the vibration.

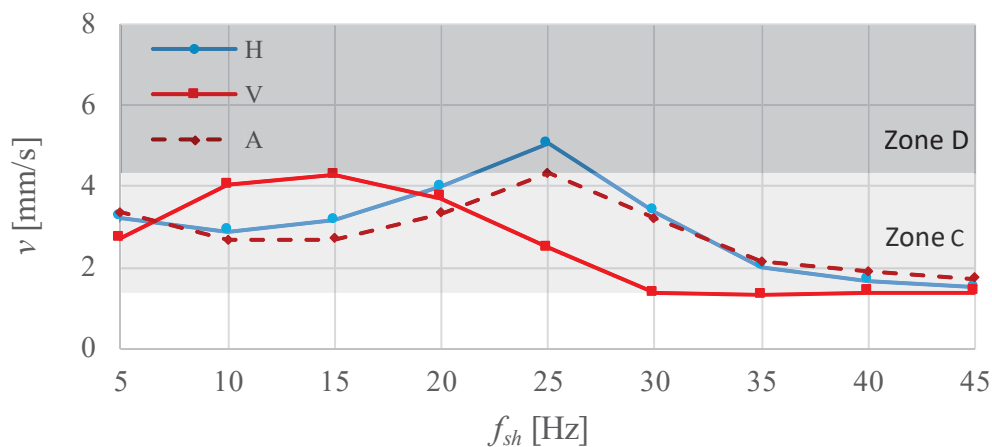


Figure 9. Measured broad-band vibration velocity in the horizontal (H), vertical (V), and axial (A) directions versus the voltage subharmonic frequency for subharmonic value $u_{sh} = 0.5\%$.

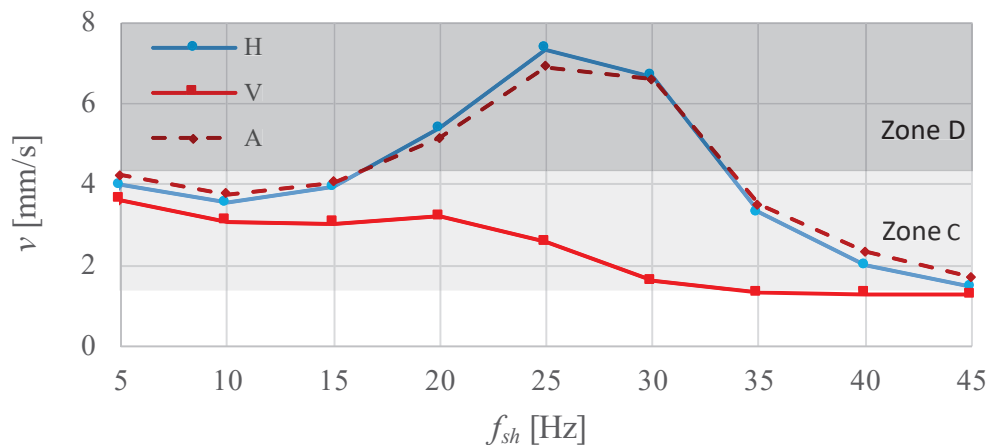


Figure 10. Measured broad-band vibration velocity in the horizontal (H), vertical (V), and axial (A) directions versus the voltage subharmonic frequency for subharmonic value $u_{sh} = 1\%$.

In summary, at full load, the measured vibration velocity was significantly higher than at no load. Additionally, for the investigated motor, the maximal vibration occurred for other frequencies at full and no load. A detailed analysis of the impact of load variation on the vibration will be a subject of future research.

5. Discussion

Voltage subharmonics exert a harmful impact on various electrical equipment, especially on rotating machinery. Among others, voltage subharmonics cause local saturations of the magnetic circuits, increased power losses and windings temperatures, torque pulsations, and unacceptable vibrations. Previous works on low-voltage machinery under subharmonics [5,28,31–37] focused on the negative phenomena appearing in the cage induction motor. The effects of voltage subharmonics on the LSPMSM were investigated in the authors' studies [36,37], but they were limited to positive-sequence subharmonics. In fact, both positive- and negative-sequence voltage subharmonics may occur in the power system, independent of their frequency (order) [21].

The previous studies on the rotating machinery under SaIs reported excessive vibration for no-load conditions, while in the load conditions, the vibration was generally moderate [5,35–37]. On the contrary, the results of the investigations presented in this work show that unacceptable vibration may occur for both no and full load. For the full-load condition with subharmonic values reported in real power systems ($\sim 1\%$ [1–3]), the measured vibration reaches an extremely high level—the maximal vibration velocity exceeds the boundaries of the evaluation Zone D by 60%. In addition, for the subharmonics value $u_{sh} = 0.5\%$, the vibration velocity falls into Zone D, the region where machine damage can occur [46]. Furthermore, for no load, the measured vibration velocity was appreciably lower than under full load. For both the considered subharmonic values— $u_{sh} = 0.5\%$ and 1% —the maximal vibration velocity corresponded to Zone C. In practice, the vibration was undue for long-time machine operation.

To protect energy receivers against malfunctions, strict limits on voltage subharmonics should be introduced into power quality standards. The determination of the appropriate permissible limits requires vibration investigations on rotating machinery. The research should include LSPMSMs of various manufacturers, rated powers, and numbers of poles under both positive- and negative-sequence voltage subharmonics.

6. Conclusions

Previous works [5,35–37] reported excessive vibration of the rotating machinery under subharmonics in the no-load conditions, while the vibration observed under full load was generally comparatively low. In contrast, for the investigated LSPMSM under negative-sequence subharmonics, the vibration at full load was much more significant than at no

load. For the full-load condition with voltage subharmonics values significantly less than observed in real power systems, the maximal vibration velocity exceeds the threshold value of the evaluation Zone D [46]. Additionally, for the no-load condition, the vibration velocity was unacceptable, corresponding to Zone C. To protect energy consumers from malfunctions caused by subharmonics, their limits should be introduced into power quality standards and rules. Determination of the permissible subharmonic levels requires comprehensive investigations of various electrical equipment, including LSPMSMs.

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Abbreviations

A	axial
AC	alternating current
B&K	Brüel&Kjær
DC	direct current
DCS	double-frequency conversion system
H	horizontal
LSPMSM	line-start permanent magnet synchronous motor
SalS	subharmonics and interharmonics
V	vertical

List of Symbols

f_1	fundamental frequency
f_h	harmonic frequency
f_{ih}	interharmonic frequency
f_m	voltage fluctuations frequency
f_o	DCS output frequency
f_r	DC link ripples frequency
f_{sh}	subharmonic frequency
g	gravitational acceleration
I_{ih}	current interharmonic
I_{rat}	rated current
I_{sh}	current subharmonic
m	zero or natural number
n	natural number
p_1	pulse number of the rectifier
p_2	pulse number of the inverter
S_{rat}	rated apparent power
u_{sh}	voltage subharmonic (related to the fundamental component)
x_d	direct-axis synchronous reactance
x_q	quadrature-axis synchronous reactance

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Article

Methodology for the Detection and Classification of Power Quality Disturbances Using CWT and CNN

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Abstract: Energy generation through renewable processes has represented a suitable option for power supply; nevertheless, wind generators and photovoltaic systems can suddenly operate under undesired conditions, leading to power quality problems. In this regard, the development of condition-monitoring strategies applied to the detection of power quality disturbances becomes mandatory to ensure proper working conditions of electrical machinery. Therefore, in this work we propose a diagnosis methodology for detecting power quality disturbances by means of the continuous wavelet transform (CWT) and convolutional neural network (CNN). The novelty of this work lies in the image processing that allows us to precisely highlight the discriminant patterns through spectrograms into 2D images; the images are cropped and reduced to a standard size of 128x128 pixels to retain the most relevant information. Subsequently, the identification of six power quality disturbances is automatically performed by a convolutional neural network. The effectiveness of the proposed method is validated under a set of synthetic data as well as a real data set; the obtained results make the proposal suitable for being incorporated into the monitoring of power quality disturbances in renewable energy systems.

Keywords: power quality disturbance; condition monitoring; continuous wavelet transform; convolutional neural network

1. Introduction

Nowadays, electric power generation has taken a turn towards the implementation of renewable energies (RE), a situation that supposes not only a change but a challenge. Therefore, the power system is suffering a transition that is being carried out in a gradual and controlled manner. The International Renewable Energy Agency (IRENA) estimates that by 2050, around 90% of the power demand around the world can and should come from renewable energies [1]. Currently, power generation systems are based on the burning of fossil and mineral fuels, but this way of production is facing environmental and economic problems, becoming less and less profitable every day. Thus, many global policies are encouraging the implementation of distributed generation systems that include renewable sources. This way, we can achieve greater sustainability because the use of RE reduces the amount of green-house gas emissions, which are related to air pollution and climate change. However, RE are susceptible to problems related to power quality (PQ), which is derived from the erratic behavior of renewable generation systems [2]. In this sense, some conditions like the shadows generated by clouds and buildings [3,4], the accumulation of dust, snow, or dirt generated by wildlife [5,6], and the temperature [7], all impact the performance of the photovoltaic system (PVS) by obstructing the solar radiation that reaches

the system, decreasing its performance by up to 20%, as shown in [8]. All these variations downgrade the amount of generated power, but they also degrade the PQ [9]. This is not a minor situation, because when the PVS is interconnected to the grid, the disturbances related to the use of RE are also incorporated into the conventional grid, affecting the end-users. Therefore, PQ has become a topic of interest because it has been shown to have an impact on the performance of both electrical and electronic equipment. The term PQ derives from a standard given by IEEE std 1159 [10], which specifies the ideal characteristics that the sinusoidal signal should have both in frequency and magnitude to ensure optimum operation of the equipment. Within this same standard, certain anomalous behaviors are already defined and called power quality disturbances (PQD). Some of the most common can be classified as sag, swell, fluctuations, harmonics, and transient impulses. Additionally, each of these disturbances has its own well-defined characteristics that serve as a basis for the detection, classification, and mitigation of these types of disturbances when they appear in an electrical signal.

Due to the importance of PQ, a number of processing tools and techniques have been developed to address the presence of disturbances and anomalies in power systems. The events related to PQ may be classified as stationary and non-stationary depending on the nature of the phenomenon [11], and they have different behaviors in time, frequency, and amplitude. Therefore, the developed processing techniques must be able to properly identify variations and deviations in all these magnitudes. Several signal processing techniques have been used to address this problem; some of them use frequency analysis by means of the Fourier transform (FT) [12]. Although this technique provides good results, it is limited because it does not provide temporal information and is focused only on the analysis of stationary signals. Additionally, some other more specialized techniques have been used to perform not only a frequency track, but also a time-frequency analysis, such as the wavelet transform (WT) [13], empirical mode decomposition (EMD) [14], the short time Fourier transform (STFT), the S-Transform (ST), continuous wavelet transform (CWT) [15], and the Hilbert Huang transform (HHT) [16], among others. Due to their ability to perform an adaptive time and frequency trace, these methods are suitable for analyzing transient, nonlinear and non-stationary signals, and they have shown good results in the analysis of PQD. Each of these techniques has characteristics that improve or decrease its performance depending on the analyzed signal. The CWT, for example, is one of the most used techniques due to its windowing characteristic that allows precise monitoring in frequency, time, and magnitude resolution, providing versatility and adaptability to treat a vast number of signals within different frequency ranges. Recently, these time–frequency techniques have been considered from an image processing point of view because the resulting spectrogram is in fact an image, and spectral component information may be extracted from it. For instance, in [17], the STFT and the S-Transform are used to obtain different spectrograms that help to characterize different PQD, such as voltage sag, voltage swell, momentary interruption, harmonic content, and DC offset. With the methodology proposed in that work, it is possible to visualize and classify these disturbances in an easy way, and it is demonstrated that the S-transform helps to improve the results delivered by the STFT. In a similar approach, the work in [18] uses the Chirplet transform (CT) and different time-frequency-scale transform (TFST) to show how it is possible to detect and classify different PQD through a spectrogram, facilitating the interpretation of the results. Even though these techniques have been shown to be useful in the detection of PQD, they still depend on the interpretation of the information by an expert user, whether performing a direct parameter comparison based on statistical indicators or by visual comparison. This implies a drawback, because there is the possibility of misinterpretation, resulting in false positive or false negative situations that may cause permanent damage to the equipment connected to the grid. Additionally, it is important to mention that the amount of data obtained from the time-frequency domain techniques and the spectrogram images is huge, complicating the correct analysis of the results even more.

In order to improve the results delivered by the space-transform-based methodologies and to avoid the misinterpretations resulting from human error, some other methodologies have been developed to receive the large amount of data delivered from the analysis of the power signals and to perform automatic detection of PQD. In this sense, the methodologies that make use of deep learning (DL) and machine learning (ML) have gained popularity in recent years. These approaches are provided with the ability to learn; thus, they can be adapted to perform continuous monitoring that allows the automatic classification of different disturbances related to PQ. On the machine learning side, the work developed in [19] proposes an approach based on dimensionality reduction through a linear discriminant analysis (LDA); moreover, statistical features are used to characterize the behavior of different PQD, and finally, using a Neural Network as a classifier, it is possible to perform an automatic classification of the disturbance in the signal. Another example of this approach is presented in the work of [20]. Here, four main steps are defined: generation of PQD classes, feature extraction, selection of highly discriminating features, and then the classification of PQDs using selected features. Both approaches demonstrated high classification rates related to PQDs. On the other hand, the work presented in [21] analyzes different machine learning approaches that have been carried out in recent years. In that work, the authors show different techniques, like the support vector machine (SVM), artificial neural network (ANN), k-nearest neighbors (kNN), and fuzzy logic (FL). They test the effectiveness of all of them in the detection and classification of different PQD, demonstrating that machine learning is a suitable tool for this task. On the deep learning side, several works have been developed. For instance, in [22,23] the authors use an auto-encoder neural network to find the most representative features of PQ-related events in an automated way. Also, in [24] the large number of images and information required for the correct training of the network takes at least 87 min; thus, it requires a high computational effort and hinders its application in continuous monitoring systems. Moreover, they do not explore if image pre-conditioning (like resizing and zones cutting) may reduce the computational burden, preserving the accuracy of the methodology; finally, they only perform tests in synthetic signals, and it is not shown how the method responds to real signals. Even with the difficulties that may arise, there are works that have demonstrated that CNN is a powerful tool for the identification of deviations in electric signals. For instance, in [25,26] the authors applied the CNN for analyzing electrical motor failures using spectrogram images that show the benefits of working under a deep-learning scheme. Given the results obtained in different fields, it is possible that similar results can be obtained in the PQ field. One of the limitations of the discussed works is that most of them only work with ideal synthetic signals without the effects of noise contamination, achieving classification accuracies up to 97%, which do not reflect their performance under real conditions.

This work proposes a methodology for the automatic detection and classification of PQD using CWT and CNN as main tools. This approach is based on image processing techniques and pattern recognition. Therefore, a 2D image bank is generated from the spectrograms of different disturbances obtained by means of the CWT. To ensure that only the most relevant information is retained in the images, they are cropped, reducing their resolution by resizing the images to a standard 128×128 pixels. This way, the amount of data is highly reduced, and an improvement is achieved in the classification process that turns to the next task. Such classification is performed by a CNN that receives the resized spectrogram images as inputs and delivers the PQD that exists in the signal as outputs. The proposed methodology is validated first using synthetic signals that simulate six different PQ conditions: healthy signal, sag, swell, harmonic distortion, voltage fluctuation, and impulsive transient. Additionally, with the aim to improve the methodology strength by giving a synthetic signal with a level of realistic behavior, each one of the synthetic disturbances considered adding four levels of white Gaussian noise (0, 10, 20 and 30 dB). These signals are used to train the deep learning model and determine whether the methodology is suitable for performing the automatic classification of the PQD. Next, the proposed approach is tested with real signals coming from a 30 MW photovoltaic

park located in Spain. Since in this type of system it is necessary to guarantee that the voltage signal remains within very specific ranges in amplitude and phase, the proposed methodology considers the voltage signals from the PV system as the inputs to be analyzed. The novelty of this work consists of the use of real signals from a PVS and a preprocessing stage consisting of the reduction of the data set by applying a resize to the spectrogram images and using signals of short duration (maximum one minute). The obtained results show that the developed method can detect and classify a wide variety of PQD, making it a useful tool to assess the impact that a PVS may cause regarding PQ.

2. Theoretical Background

This section describes the theoretical aspects that are considered for performing the proposed diagnosis approach, relevant definitions of power quality disturbances (PQD) are given, and the general description of a CNN structure is mentioned.

2.1. Power Quality Disturbances

PQ is a general term that refers to a phenomenon that can affect electrical equipment when the power supply presents instabilities or disturbances associated with amplitude, continuity, waveform, and frequency. Therefore, the PQ is often measured through different international standards where the most important are the IEC 61000, IEEE 1159, and the EN 50160 [27,28]. Although different standards can be used to describe PQD, the IEEE 1159 is the most recognized one, and it is recommended and implemented for monitoring PQ; thus, five main disturbances such as sags, swells, transients, fluctuations, and harmonics are defined by the IEEE 1159 [10].

Therefore, three of these five disturbances are directly related to the amplitude of the voltage signal, and the other two are associated with changes in the frequency. Certainly, the sag is represented as a phenomenon that shows a loss of amplitude. The occurrence of sag can also be measured in terms of the RMS value of the voltage signal within a range from 0.1 to 0.9 pu (per unit) with a duration greater than 5 cycles in 60 s. On the other hand, the swell occurs when the voltage signal is subjected to amplitude increases of around 1.1 pu in terms of the RMS value and with a duration greater than 30 to 60 s. Although the sag and swell modify the amplitude of the voltage signal, the transient impulse appears as a sudden change in steady-state frequency conditions that can produce changes on both signals, voltage, or current, and is unidirectional in polarity; hence, this kind of disturbance is measured by quantifying the amplitude of the disturbance, the onset time, and the duration.

Regarding the disturbances linked to changes in the frequency, the harmonic distortion is known as the result from the sum of multiple frequencies of the fundamental component at which the system is intended to operate; consequently, the occurrence of harmonics leads to a waveform distortion, and it can be evaluated by estimating the total harmonic distortion (THD). According to the standards, the THD level must be lower than 8% in power grids that operate under voltages lower than 1.0 kV. Finally, voltage fluctuations are conditions where random changes in the voltage signal appear. Such changes can also be defined as systematic variations of the signal envelope within ranges from 0.95 to 1.05 pu.

2.2. Continuous Wavelet Transform (CWT)

Condition monitoring strategies are commonly supported by signal processing techniques that allow highlighting those fault-related patterns; hence, the time domain, frequency domain, and time–frequency domain remain the most preferred processing techniques, and each one of them can be selected according to specific aims. Certainly, time–domain processing techniques represent a practical solution, since trends and changes of the signals can be modeled through the estimation of statistical features with an excellent trade-off between simplicity and performance [29]. By their part, frequency domain techniques represent some of the most reliable approaches since the existence of specific undesired frequency components on the signal usually indicates malfunctioning problems [30]. On

the other hand, time-frequency techniques are also incorporated for the processing of signals due to their capability of providing temporal and spectral information.

In this sense, the continuous wavelet transform (CWT) is considered a time–frequency technique that allows the analysis of multiscale non-stationary signals using a windowing approach to extract information from segments of a signal [31]. These signal segments are then evaluated by a window called mother wavelet [32], this process allows the CWT to have the versatility of lengthening or shortening the window size depending on the selected scale of the mother wavelet. Consequently, a smaller scale is associated with a greater sensitivity in frequency variations; Equation (1) depicts the mathematical expression:

$$C(a, \tau) = \int \frac{1}{\sqrt{a}} \psi\left(\frac{t - \tau}{a}\right) x(t) dt \quad (1)$$

where C is the transform, a is the scale factor, τ is the overlap time, ψ is the mother wavelet function, and $x(t)$ is the signal input as a function of time.

Hence, one of the main advantages of the CWT is its capability of contraction and dilation, which can lead to high spectral resolution depending on the needs of the signal to be analyzed. Likewise, an additional property of the wavelet function is that it can obey a certain limit, since the energy represented must be finite and must meet the admissibility condition. These parameters are described by Equations (2) and (3), respectively.

$$E = \int_{-\infty}^{\infty} |\psi(t)|^2 dt \quad (2)$$

$$E = \int_{-\infty}^{\infty} \psi(t) dt = 0 \quad (3)$$

where E is the energy of the wavelet and ψ is the wavelet mother function.

The graphical representation of the correlation between the waveform scale and time is known as a Scalogram; subsequently, the CWT automatically adjusts the time and frequency based on the resolution, depending on the contraction or dilation of the window that best fits the signal to be analyzed [33]. Within this same aspect, there is another representation offered by the CWT, which shows a time-frequency-scale distribution mapping the distribution of the energy contained within the signal expressing this energy per unit frequency. This representation is known as the spectrogram [34]. This last feature makes the CWT suitable for the analysis of non-stationary signals since this representation considers the three main characteristics that make up a signal, showing them graphically, and facilitating their interpretation.

2.3. Convolutional Neural Network

Convolutional neural networks (CNN) have been widely used in several methods based on image classification under large sets of information; certainly, a CNN is a feed-forward network that uses partial variance to learn behavioral patterns, and it is specifically designed for image processing [34]. As in any network structure, the input layer of the CNN evaluates the set of original images to be processed, obtaining an output block that contains the estimated features derived from the original image. Accordingly, a series of filters are applied for further processing. These filters consist of four main processes, where the first one consists of a convolutional layer, polishing, vectorization, and classification. These are described in more detail below.

Convolutional layer: A convolutional layer is the main component in a CNN containing a series of filters (Kernels), which map the image by generating a new space through convolutions [35]. This means that from the original image, the most representative features are extracted as the filtering process progresses. Equation (4) shows the process carried out by a convolutional network.

$$y_j^k = \sum W_{ij}^k * x_j^k + b_j^k \quad (4)$$

where y_j^k is the mapping feature output, x_j^k is the plane feature input, W_{ij}^k is the 2D filter set, and b_j^k is the bias training parameter.

Polishing layer: The polishing layer is a non-linear process that reduces the sample size. This process is performed to reduce the number of patterns that will be fed to the training process without losing a large amount of information in the process. This feature extraction process when performing max-pooling is given by Equation (5).

$$h_r = \max_1^{M \times M}(X_{pj}) \quad (5)$$

where h_r is the output of the features when mapping, X_{pj} is the element of the clustered region, and M is the size of the clustered mapping dimension.

Flattened layer: This layer is responsible for applying a vectorization process of the mapping performed by the previous layers, forming a vector that contains the information already processed by the whole stages of convolution and polishing, generating a concentrated vector of features that can be more easily processed by the classification layer.

Classification layer: The classification layer oversees processing all the information delivered by the vectorization layer when performing a class classification process employing activation functions. One of the most used for the solution of class classification problems is the “Softmax” [36]. Image processing through CNN is a highly recommended and widely used procedure to tackle different classification and pattern detection problems, some related works can be found in references [35,37].

3. Proposed Methodology

The proposed methodology for detecting and classifying PQ disturbances occurring in photovoltaic cogeneration systems is presented through the general block diagram of Figure 1. As can be noticed, three main stages encompass the methodology to know: (i) Signals bank, (ii) Image processing, and (iii) Pattern recognition and classification.

First, a description of the stage “Signals banks” is presented. Here, two different sets of signal banks (voltage signals) are used for processing by the proposed methodology; on one hand, there is the bank of synthetic signals, and on the other hand, there is the bank of real signals. The set of synthetic voltage signals are generated having a fundamental frequency of 60 Hz with a sampling rate of 8 kHz during 0.3 s. Therefore, these signals presented in Figure 2. comprise the healthy state (ideal sine wave behavior) and five different types of power quality disturbances, such as sag (SAG), swell (SWL), flicker (FLC), harmonic content (HAR), and transient impulse (IMP). Additionally, with the purpose of being considered as power disturbances, their generation considered the normativity established by the standard IEEE-1159 in relation to parameters like duration, amplitude, and frequency [10]. Now, assuming that a synthetic signal behaves somewhat ideally and cannot provide enough variability, some noise is added to test the strength of the methodology. As a consequence, white Gaussian noise is added to the synthetic signal, and this is repeated in four different noise levels (0 SNR dB, 10 SNR dB, 20 SNR dB, and 30 SNR dB), generating 250 images per level, leaving a total of 1000 images per synthetic signal. By its part, the bank of real signals comprises a signal from a photovoltaic cogeneration system in Spain, located in the south-central area of the country, which is a 30 MW photovoltaic power plant. The acquisition of the signal was performed on a 100 kW three-phase inverter at 230 RMS volts with a fundamental frequency of 50 Hz with sag disturbances. In Figure 3, an example of the acquired signal under the occurrence of SAG is shown. It must be emphasized that signal acquisition should always be performed as close as possible to the point of interest for the analysis since this ensures measurements without environmental or structural interferences that may affect the data collection. In this specific case, the subjects of interest were the PQ signals generated by the cogeneration system, so the acquisition was performed as close as possible to it in one of the phases of the inverter. The amplitudes of all the signals are normalized, and the units remain as *per-units* (pu).

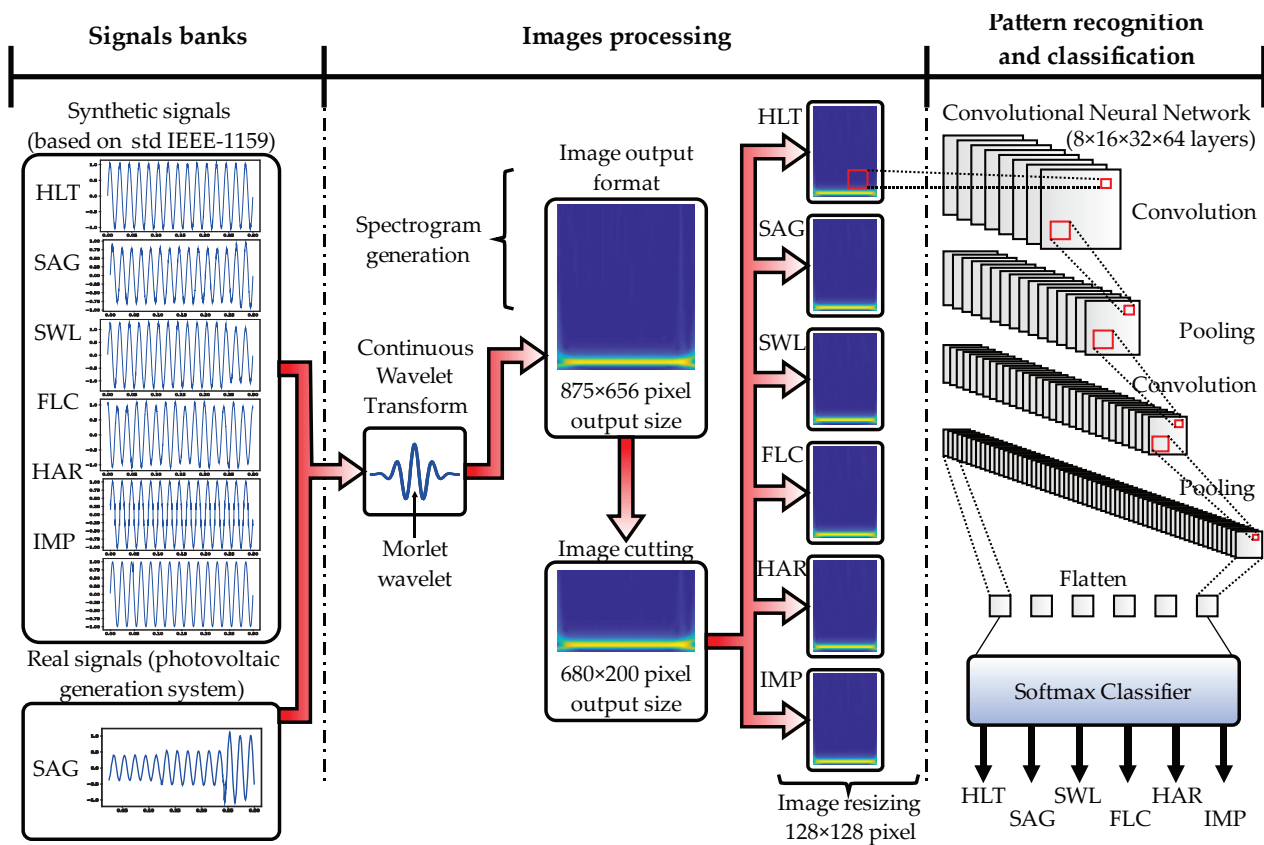


Figure 1. General block diagram of the proposed methodology based on CWT and CNN for detecting and classifying standardized PQ disturbances.

In the second stage of the proposed methodology “Images processing”, the synthetic signals coming from the first bank are processed through the continuous wavelet transform (CWT). To perform the CWT, the analytic Morlet-Gabor is selected as the mother wavelet, and an extended signal configuration is used to mitigate the boundary effects. This way, the normalized voltage signals are transformed into 2D images, having the information about whether a power disturbance appears or not. Thus, through the CWT, a spectrogram image is generated from the original signal as output with an original format of 875×656 pixels. However, since not all the zones of the image contain relevant information about the power disturbance, it is necessary to cut the upper area of the spectrogram, leaving only the bottom area where the magnitude components are observed, yielding a clipped frame of 600×200 pixels size. Additionally, with the aim of reducing the computational effort required to process the data, a resizing is performed to this new frame, yielding a final image with a format of 128×128 pixels. This procedure will be repeated for each one of the six signals (normal signal and the five signals with power disturbances), considering the free noise (250 samples) signal and the three noise levels per disturbance condition (750 spectrograms); then, a total of 6000 resized spectrograms are finally generated.

The third stage, named “Pattern recognition and classification”, consists of applying a classifier structure based on deep learning, which in this case is a convolutional neural network (CNN) ideal for processing data in image format. This way, the final images of 128×128 pixels with the different noise levels are fed into the network with a binary structure that considers four hidden convolutional layers of 8, 16, 32, and 64 neurons, respectively. The vectorized output from the CNN (flatten layer) is finally processed by a SoftMax multiclass classifier with the purpose of labeling the type of power disturbance contained in the spectrogram. The training process of the CNN considers Stochastic Gradient Descent Momentum (SGDM), and the convolution layers have a Rectified Linear Unit (ReLU) activation function. With this, the network training speed is increased, and

the convergence of the gradient descent algorithm is ensured. What exactly happens is that the CNN learns patterns about the power disturbances contained in the spectrogram image, and then SoftMax provides a final label of the specific disturbance found. It is worth mentioning that the images are first fed to the CNN for training purposes, but some images are reserved for validation, and additionally, some images from the real data bank are tested to probe the methodology effectiveness. The detailed description of the network structure, and the percentage of data considered for training and validation, are also specified in the next section.

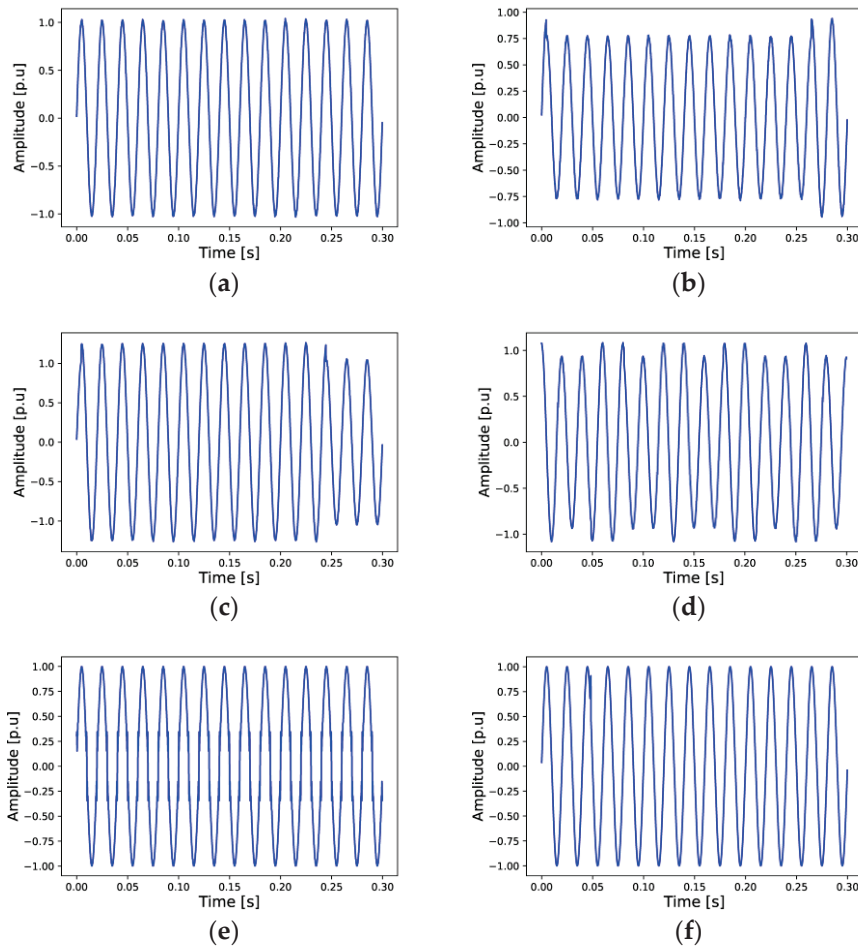


Figure 2. Synthetic signals considered for the evaluation of the proposed methodology; each one of the signals represents: (a) healthy condition (HLT), (b) voltage sag (SAG), (c) voltage swell (SWL), (d) flicker (FLC), (e) harmonic content (HAR), and (f) transient impulse (IMP).

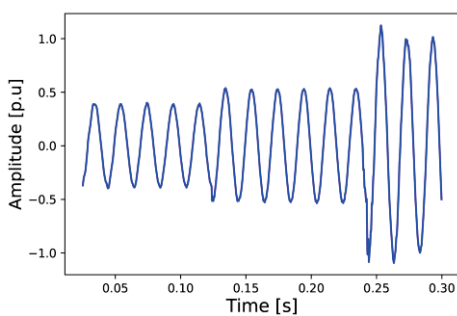


Figure 3. Real signal acquired from a photovoltaic cogeneration system with sag disturbance.

4. Results

This section is divided into two main parts. The first one presents the results that are achieved when the proposed diagnosis methodology is applied to a bank of synthetic signals, and the second one shows that the results that are carried out when real signals from a 30 MW PVS located in Spain are evaluated through the proposed approach.

4.1. Competency of the Proposed Methodology in Front of Synthetic Signals

The proposed methodology is first evaluated under a bank of synthetic signals in order to validate its applicability to real systems. Thus, as mentioned previously, a bank of synthetic signals is generated involving a total of six different conditions: healthy signal (HLT), voltage sag (SAG), voltage swell (SWL), harmonic distortion (HAR), voltage fluctuations (FLC), and impulsive transient (IMP). Such synthetic signals are generated by considering the mathematical formulation presented in [38], thereby, for each one of the considered conditions, a total of 1000 signals with a duration of 0.3 s are synthetically generated. Indeed, it should be mentioned that these signals were randomly generated, but the duration, location, and severity are within the parameters established by the standard IEEE 1159. Moreover, white Gaussian noise has been added to the generated signals to ensure their similarity to those that appear in real conditions; hence, the noise was added by considering random SNR values from 0 dB to 40 dB. Accordingly, the synthetic signals were generated as shown in Figure 2, where it is possible to appreciate the waveform of the conditions under study, such as for the healthy condition, voltage sag, voltage swell, flicker, harmonics, and transients, respectively.

In addition, it can be observed from the synthetic signals of Figure 2 that the amplitude of each one of them is expressed as per unit (p.u.) as is defined in the standard IEEE 1159. Certainly, the use of p.u. is a mandatory suggestion that must be followed during the implementation of monitoring strategies since this particular scale defines the amplitude parameters of any PQD. Consequently, the implementation of the proposed methodology to real signals may consider a normalization process in order to ensure the proper assessment and identification of disturbances. Additionally, it is observed that the disturbances are present in the signals, but its identification usually requires expertise; on the other hand, it is common to make mistakes during the identification of PQD when the assessment is carried out by visual inspection. This assertion is due to how there are some disturbances that are difficult to detect even by experts. For instance, the IMP disturbance in Figure 2f presents a low amplitude and a short duration, making it difficult to find. Therefore, the development of an efficient strategy to identify and classify the PQD remains necessary.

Thus, following the proposed methodology, the synthetic signals are then subjected to a processing stage by means of the CWT; as a result, a spectrogram represented by a 2D image is obtained for each one of the signals. Consequently, a set of 1000 images is generated for each condition under study; afterward, the images are preprocessed with the aim of reducing the amount of data that must be treated without compromising the information provided by the image. Such preprocessing is mainly performed for resizing the original images as shown in Figure 4; thereby, all the images are scaled to a new size by preserving the most important information.

The 2D spectrogram presented in Figure 4 belongs to one of the generated signals that simulate a voltage sag. The spectrogram is obtained by applying the DWT to the signal under evaluation, and then it is saved as an image with an original size of 875×656 pixels. It is observed that the original image provides relevant information regarding the disturbance behavior; for instance, a frequency component linked to the fundamental frequency appears horizontally in yellow at the bottom of the image, but at the beginning and at the end of the 2D spectrogram, it can be noticed that there are changes that affect the amplitude of the fundamental component. These amplitude changes are due to the occurrence of the voltage sag, and this is explained because the signal being analyzed is healthy at the beginning, but around 0.02 s, the voltage sag starts; then, the voltage sag remains until 0.25 s, when the healthy condition returns. It should also be highlighted that

during the transition from sag to healthy condition, a faint line appears that is relevant for the identification of amplitude variations. In this regard, a first size reduction is carried out by making a crop to the original image with the aim of removing most of the redundant information, but the most relevant information is preserved. The region of high frequencies is completely removed from the original 2D spectrogram by applying the cropping, thus, although it could be thought that relevant information is being eliminated, it should be mentioned that high frequencies do not provide significant information from the viewpoint of power quality analysis. Finally, the image is resized to a standard 128×128 pixel size; hence, despite the size reduction to 128×128 pixels, it is still possible to observe the amplitude changes at the beginning and at the end of the signal. Truthfully, the faint line that depicts the transition from the disturbance to the healthy state remains in the final image, proving that the relevant information is preserved after the resizing. Although the processing that is applied to the images seems trivial, it has a significant impact, since it allows us to significantly reduce the amount of data processed by the CNN; consequently, the computational burden is also reduced.

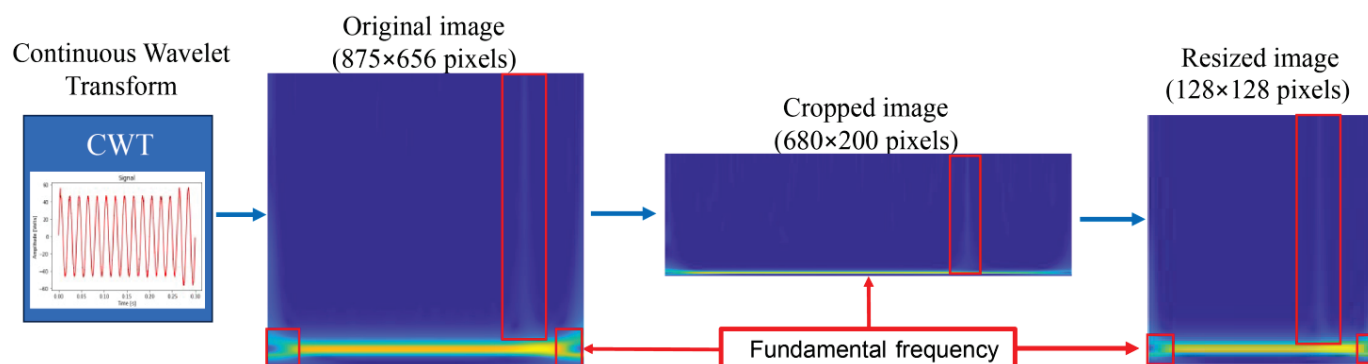


Figure 4. Resizing process to reduce the amount of data without losing information.

Hence, the resizing procedure is applied to all images of all considered conditions; then, the new images are used as input in the CNN structure in order to carry out the classification task. Due to there being a total of 6000 spectrograms (1000 per disturbance) that compose the dataset under assessment, 60% of the 2D spectrograms is used during the training process, 20% is used for testing purposes, and the remaining 20% is used for validation. Afterward, the CNN structure is subjected to the training procedure for 550 iterations, leading to achieving an accuracy of around 99.41% during the test process and a maximum value of around 99.48% during the validation. In Figure 5, the qualitative representation of the accuracy that was reached during the training of the CNN is shown. The achieved accuracy is promising, and it can be considered as a metric that validates the effectiveness of the proposed methodology and its feasibility for being applied in the detection and classification of PQD. On the other hand, Figure 6 shows the qualitative representation of loss achieved by the CNN, which at the end of the process tends to zero. From the results presented in Figure 6, it can be inferred that the hyperparameters of the CNN are well-adjusted, and therefore, it is expected that the classification errors remain low. Regarding the individual classification for each one of the assessed conditions, the confusion matrix obtained during the individual classification is presented in Table 1. As observed, for the HLT, FLC, HAR, and IMP conditions, 100% of the individual classification is achieved, whereas the SAG and SWL conditions reached an individual classification of about 99.0% and 97.5%, respectively. Only two and five misclassification errors were made between the SAG, SWL, and the HLT conditions. Although the proposed classifier sometimes produces classification errors and confuses the SAG and SWL conditions with the HLT one, this is an expected situation, because all the generated signals have at least a part of the HLT condition. Nevertheless, the high-performance of the accuracy indicates that the proposed methodology has the capability to detect and classify different disturbances, and it can be used as a diagnostic tool for assessing the quality of power grids.

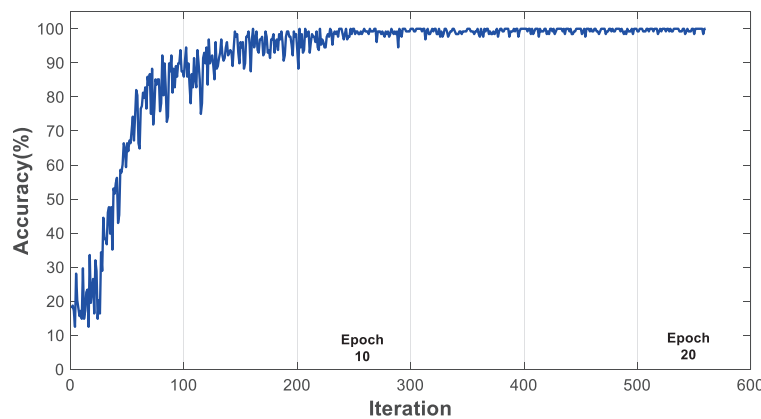


Figure 5. Qualitative representation of the accuracy achieved by the CNN structure during the training procedure.

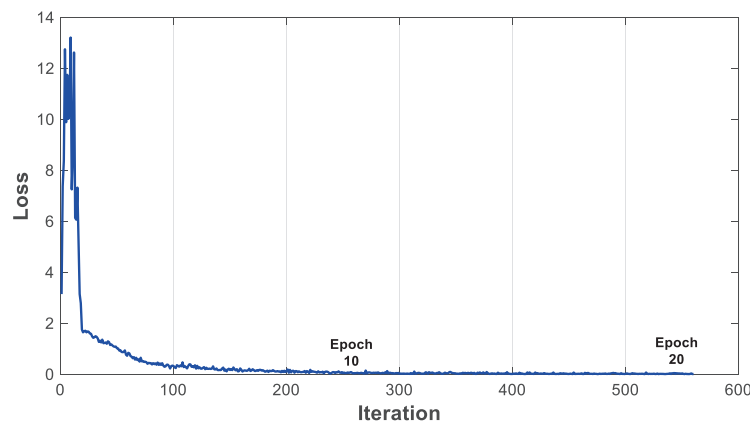


Figure 6. Qualitative representation of the loss achieved by the CNN structure during the training.

Table 1. Confusion matrix achieved by the CNN structure during the validation for all assessed conditions.

		True Class					
		HLT	SAG	SWL	FLC	HAR	IMP
Assigned Class	HLT	200	2	5	0	0	0
	SAG	0	198	0	0	0	0
	SWL	0	0	195	0	0	0
	FLC	0	0	0	200	0	0
	HAR	0	0	0	0	200	0
	IMP	0	0	0	0	0	200

Finally, for practical implementation, it should be mentioned that the proposed methodology has been implemented in a personal computer with an intel core i7-12700H 12th gen (4.7 GHz), and the image processing is configured to be performed by a dedicated GPU NVIDIA GeForce RTX 3070 (Manufactured by Micro-Star International-MSI Co., Ltd., New Taipei City, Taiwan). In this sense, the whole process is carried out in an approximated time of 120 s, which is a relatively short time of response considering that the processing is purely carried out on images.

4.2. Study Case: 30 MW Photovoltaic System

Once the proposed method is successfully applied to identify and classify different PQD through the analysis of synthetic signals, the analysis of real signals acquired from a real electric system is then carried out. Certainly, the experimentation is performed with

the voltage signals acquired from a PV generation plant located in south-central Spain; such PVS can generate a total of 30 MW, and it is interconnected to the local grid. The measurement of electric signals is carried out in a 100 kW three-phase PV inverter with a nominal voltage of 230 Vrms at 50 Hz. The acquired signals are measured in this place because it has been demonstrated that the PQD that appear in a power grid are produced due to the action of the loads that are connected to grid; therefore, the measurement is performed in the AC side of the PV inverter, which is the device in charge of feeding the loads connected to the grid. Hence, it is important to mention that current signals in a PV system can present many natural variations throughout the day. These variations occur because the current delivered by the PV system is determined by the amount of solar radiation that reaches the PV panels. In this sense, the current signals are not suitable to be analyzed, but the voltage signals must always remain at an amplitude of 230 Vrms with a 50 Hz frequency; therefore, any deviation from these values may be dangerous for the equipment connected to the grid. Accordingly, the analysis of PQD through the analysis over voltage signals is adequate.

Consequently, the signals are acquired and stored using a proprietary data acquisition system (DAS) based on field programmable gate array (FPGA) technology; the DAS is able to acquire the three-phase voltages and currents at the same time at a sampling frequency of 8000 Hz. Although the PVS is designed to keep the failures and disturbances to a minimum, it eventually presents some undesired phenomena. As mentioned before, the real signals are subjected to a normalization procedure in order to transform the original amplitudes into the p.u. scale. Afterward, the proposed methodology is applied as described in Section 3 in order to carry out the assessment and identification of PQD. In this sense, Figure 7a shows a voltage sag that was identified by the proposed methodology; as observed, the occurrence of two different severities of voltage sag produces effects in the voltage signal during 3 s. Thus, two main different amplitudes are identified. The first one is around 0.4 p.u., and the second one is about 0.6 p.u. Finally, at around the 0.25 s, the signal returns to a normal amplitude that falls within the permitted range established by the standard IEEE 1159. Additionally, the corresponding 2D spectrogram obtained by the CWT is shown in Figure 7b, where it can be seen that there are two faint lines that represent the two changes in the amplitude of the signal. However, the fundamental component of frequency cannot be clearly appreciated in Figure 7b since it is hindered by the background noise. In this sense, the image resizing procedure is then applied as a filter that allows highlighting the fundamental frequency component due to non-useful frequencies (high-frequencies) being removed from the original spectrogram (see Figure 7c)). Also, it is observed that the two lines that depict the amplitude changes of the signal are more visible in the resized image than in the original one, marking the importance of the resizing process in this methodology being able to achieve accurate results while reducing the computational burden.

The proposed methodology also has the capability to detect other different PQD; hence, an impulsive transient is also identified in the 30 MW PVS as presented in Figure 8a. In fact, it is necessary to zoom in over the affected signal zone in order to clearly identify the appearance of the impulsive transient; hence, it must be highlighted that the identification of this kind of disturbance is difficult through a visual inspection. Then, the CWT is applied over the affected signal to obtain its corresponding 2D spectrogram (see Figure 8b), and a line appears extended over the entire range of frequencies. This line is located around 0.1 s, approximately. Afterward, the resizing procedure is carried out, highlighting the details of the 2D spectrogram (see Figure 8c). The final 2D spectrogram may facilitate the identification and classification of the disturbance performed by the CNN.

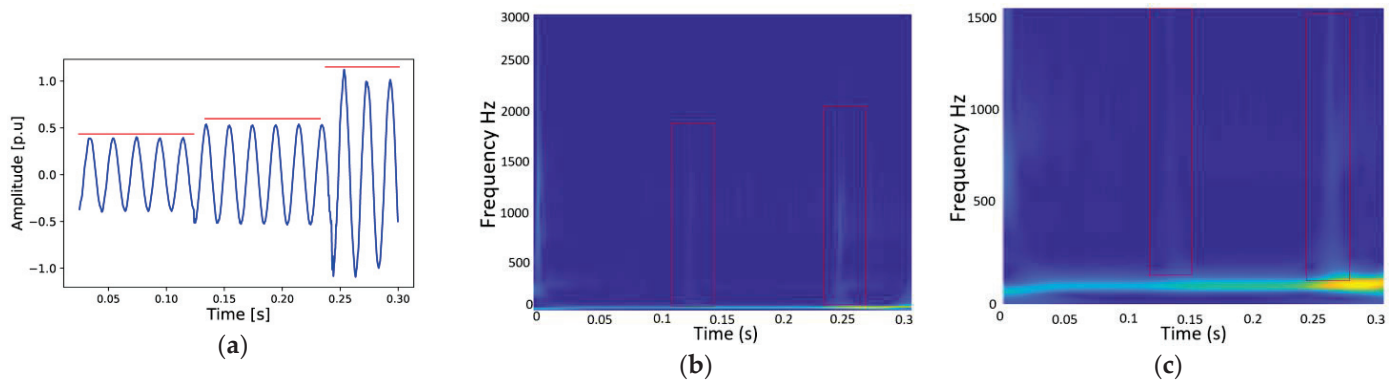


Figure 7. Real signal acquired from the 30 MW photovoltaic system, (a) voltage signal under the occurrence of SAG (horizontal red lines represent different SAG severities), (b) its original 2D spectrogram obtained by means of the CWT and, (c) the resized 2D spectrogram obtained after the resizing procedure.

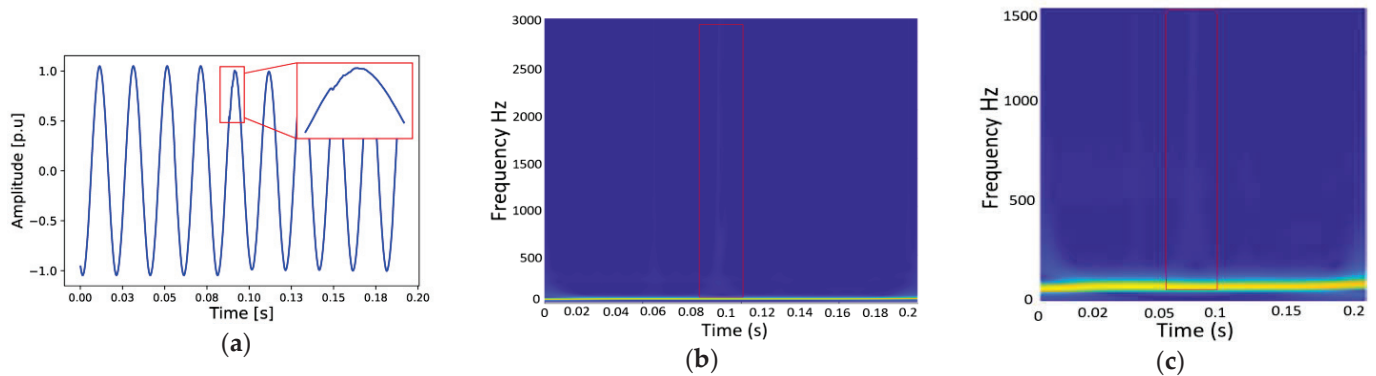


Figure 8. (a) real signal from the PVS with an impulsive transient (red rectangle), (b) original spectrogram image obtained by means of the CWT over the signal with the impulsive transient (red rectangle indicates the occurrence of impulsive transient), and (c) resized spectrogram of the signal with impulsive transient.

It is worth mentioning that there are some other methodologies that do not require the use of images. Such methodologies use a waveform approach by analyzing the amplitude and phase angle of the electric signals. However, the waveform-based methodologies commonly perform a track of patterns and behaviors in the signals by performing a feature extraction. These features used to be statistical, entropy-based features, energy features, among others. The selection of these features is not a trivial task, and it has a very deep impact on the accuracy of the results. Therefore, to properly select the relevant features and to discard redundant information, the waveform-based methodologies implement complementary processing for the optimized feature selection. Additionally, the use of several features results in a high number of parameters to describe a single PQD, and to obtain an accurate classification, it is necessary to implement a complex classification scheme that requires of the tuning of a high number of parameters; many times, the optimum performance is not achieved even after multiple trials. Thus, in this work, a 2D-CNN is used that performs the feature extraction and the classification in a single step, without requiring of the tuning of many parameters. The tradeoff of using this technique is that the input of the 2D-CNN is an image, and the only preprocessing that is performed is a resizing, which is a simple task. In this sense, the use of image processing is helpful for reducing the complexity of the optimal feature selection and the high number of parameters to be tuned in the classification task by using a single technique (2D-CNN) to perform all these steps.

4.3. Comparative Analysis

Observing the results shown above, it is clear that the proposed methodology has been able to adequately classify the different phenomena related to PQDs with a global classification ratio up to 99.48%, for all the six power disturbances. However, with the purpose of demonstrating the effectiveness of the proposed approach, a comparative analysis versus some previously mentioned works in the introduction section is carried out. In this sense, Table 2 summarizes the results obtained by the previous literature and also shows the conditions evaluated, as well as the used classification technique.

Table 2. Comparison of different techniques and performances.

Ref	PQD	ND Technique	Noise	Total Performance (%)
[19]	Normal, sag, swell, harmonic, fluctuation	Feature extraction/ NN	N/A	99.7% ($\approx$ 79% for sag)
[20]	Normal, harmonics, interruption, notch, sag, swell, transient impulse	Cubic SVM	N/A	99.7% ($\approx$ 79% for sag)
[22]	Normal, swell, swell with harmonics, harmonics with oscillatory transient	OC-SVM	Combination of PQDs	94.2%
[23]	Normal, interruption, harmonics	Auto encoder	N/A	80.0%
[24]	Normal, impulsive, interruption, oscillatory, sag, swell	2D-CNN	30 SNRdB	96.67%
Proposed	Normal, sag, swell, flicker, harmonic, transient impulse	2D-CNN with image resizing	60 SNRdB	99.4%

Results reported in [19,20] present general values with high accuracy (99.7%); however, a classification percentage around 79% has been reported when the sag disturbance appears. This situation may be explained by the fact that the behavior of a signal with a sag is very similar to one that presents a healthy signal. Another important aspect is that the signals analyzed in [19,20] are free of noise, which is an important issue that significantly affects the performance of classification when real signals are under evaluation. For the remaining works, such as in [22–24], the overall percentages show variations within the range of 80% to 96.67%, which are lower than the accuracy reached by the proposed approach. Therefore, it seems that conventional methods provide high percentages of classification accuracy, but such results are reached only because the analyzed signals are ideally perfect without noise. In contrast, methodologies where noise is considered show accuracy percentages below 97% (lower than the proposed methodology). This is because the phenomena related to PQDs are very susceptible to noise conditions, reducing the classification accuracy. Therefore, it can be emphasized that the proposed method, in comparison with approaches of previous works, shows better results of up to 99.48%, even in high noise conditions.

At this point, it is also noticeable that the proposed methodology aims to fill some of the existent gaps in the field of power quality monitoring and in the analysis of distributed generation grids with renewable sources. Beyond the high classification accuracy achieved by the proposed methodology, this approach is not affected by the window size, since the CWT performs an automatic windowing process that allows us to enhance the performance of the overall technique. In this regard, most of the proposed methodologies so far are highly dependent on the window size, and they fail in classification if the window size is not properly tuned. Also, most of the reported works so far only deal with synthetic signals where the classification of disturbances is easier than in real environments. Moreover, the analysis of distributed generation grids that contain renewable generation sources like the photovoltaic one has been scarcely reported. In this sense, the proposed work demonstrates that is able to work with signals from real environments and to perform an analysis in a power grid with PV generation, making it a tool to be used in the reliable assessment of renewable generation sources. Finally, it must be mentioned that some of the reported works so far treat the quasi-stationary disturbances and the transient disturbances

separately; the methodology proposed here is able to deal with both types of disturbances (quasi-stationary and transient) at the same time. Table 3 summarizes the strengths of the proposed approach compared with some previously reported works.

Table 3. Summary of the bullet points and strengths of the proposed methodology compared with some previously reported works.

Ref	Window Size Sensitive	Transient Disturbance Identification	Quasi-Stationary Disturbance Identification	Use Real Signals	Consider Renewable Generation Power Grids
This work	No	Yes	Yes	Yes	Yes
[19]	Yes	No	Yes	No	No
[20]	Yes	Yes	Yes	No	No
[22]	Yes	Yes	Yes	No	No
[23]	Yes	No	Yes	No	No
[24]	No	Yes	Yes	No	No

5. Conclusions

In this work, we propose a condition monitoring strategy for the detection and identification of power quality disturbances that can occur in energy generation systems that are based on renewable processes. The capability of continuous wavelet transform to highlight those representative fault-related features that are linked to the occurrence of PQD should be emphasized; hence, the obtained results demonstrate that the appearance of a PQD produces changes that can be identified through a visual analysis over the obtained spectrograms; that is, the occurrence of any of the studied disturbances may produce a specific pattern. Additionally, the processing of the spectrograms as 2D images also influences the resulting diagnosis outcome that is achieved by the proposed CNN; specifically, the selection of the optimal image size plays an important role in the training of the CNN. The proposed methodology is first validated under a synthetic set of signals where six different PQD are considered, and then, validation under a real set of signals acquired from a 30 MW photovoltaic system is also performed. According to the obtained results, the method is suitable to be implemented since it can be exported to any software containing the necessary tools for CNN development, for example PYTHON, which would facilitate its implementation in maintenance programs to detect the occurrence of disturbances in renewable-energy-based generation systems.

Also, it is worth mentioning that certain conditions contribute to generating more than one power quality disturbance at the same time, acting in a combined way, and this situation could be reflected in the collected data. For instance, the switching of several resistance load types together with the startup of some induction motors in the system can induce phenomena of sags, swells, harmonic content, and transient spikes in the power lines at the same time or combinations of them. Thus, the proposed methodology could handle these situations because of the considered deep feature-learning approach. For instance, once the signals with combined power disturbances are preprocessed in stage two, spectrogram images are created containing information of these combined disturbances. Therefore, they will be used for training the CNN structure for detection, and the SoftMax for classification. Of course, it is assumed that the signals will have different behavior in their spectrograms when an isolated disturbance is present than when combined disturbances occur. Thus, this training will provide the networks with the capability to recognize the patterns of combined disturbances to give them a category. However, these cases are detected as an area of opportunity in this work, and they will be addressed as future work. Also, hardware implementation is considered future work in order to develop a diagnostic tool for online implementations. Another point of interest is the realization of a method that allows the optimization of the area to be cut out of the image to improve the process by only containing the most relevant information of the phenomenon, depending on the sampling

frequency. This can vary according to the needs of each user, allowing the automation of the entire process.

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Article

Novel Space-Vector PWM Schemes for Enhancing Efficiency and Decoupled Control in Quasi-Z-Source Inverters

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Abstract: This paper investigates the development of pulse width modulation (PWM) schemes for three-phase quasi-Z-source inverters (qZSIs). These inverters are notable for their voltage boost capability, built-in short-circuit protection, and continuous input current, making them suitable for low-voltage-fed applications like photovoltaic or fuel cell-based systems. Despite their advantages, qZSIs confront challenges such as increased control complexity and a larger number of passive components compared to traditional voltage source inverters (VSIs). In addition, most existing PWM schemes for qZSIs lack the capability for independent control of the amplitude modulation index and duty cycle, which is essential in closed-loop applications. This study introduces innovative space-vector PWM (SVPWM) schemes, addressing issues of independent control, synchronization, and unintentional short-circuiting in qZSIs. It evaluates several established continuous and discontinuous PWM schemes, and proposes two novel decoupled SVPWM-based schemes that integrate dead time and in which the shoot-through occurrence is synchronized with the beginning of the zero switching state. These novel schemes are designed to reduce switching losses and improve qZSI controllability. Experimental validation is conducted using a custom-developed electronic circuit board that enables the implementation of a range of PWM schemes, including the newly proposed ones. The obtained results indicate that the proposed PWM schemes can offer up to 6.8% greater efficiency and up to 7.5% reduced voltage stress compared to the closest competing PWM scheme from the literature. In addition, they contribute to reducing the electromagnetic interference and thermal stress of the related semiconductor switches.

Keywords: quasi-Z-source inverter; inverter control; performance optimization; system efficiency; decoupled control; pulse-width modulation

1. Introduction

Three-phase quasi-Z-source inverters (qZSIs) belong to the category of single-stage inverters with voltage boost capability [1]. Over the years, they have gained significant attention due to their distinctive features and capabilities, such as the inherent short-circuit protection, continuous input current, easy integration of batteries, and enhanced reliability. These qualities make qZSIs particularly well-suited for low-voltage-fed applications, such as photovoltaic (PV) [2–4] or fuel cell-based systems [5,6]. The voltage boost capability of qZSIs is achieved by utilizing an additional, shoot-through (ST) switching state and the input impedance network comprising two inductors, two capacitors, and a diode. The ST state is activated by short-circuiting one or all the inverter legs, typically during the traditional zero switching states of the utilized pulse width modulation (PWM) scheme so that the output voltage waveform remains unaffected. This approach allows the input dc voltage to be boosted effectively. During the ST state, the inverter bridge transistors are utilized in a similar way that the transistor is utilized in traditional dc–dc boost converters. Consequently, the qZSI topology does not introduce any additional active switches

when compared to the conventional voltage source inverter (VSI) topology. In contrast, traditional VSIs are typically combined with various dc–dc boost converter topologies in low-voltage-fed power systems to achieve the desired output ac voltage, resulting in increased complexity.

The qZSI is derived from the original ZSI topology proposed in [7], whereas it offers several comparative advantages, including continuous input current and a lower voltage rating for one of the impedance network capacitors. These advantages are achieved through a different arrangement of the impedance network, setting it apart from its predecessor. Despite the advantages offered by the qZSI, it is important to acknowledge and address its limitations. These include the increased complexity in control algorithms compared to traditional VSIs, a higher passive component count compared to conventional two-stage architectures, and the requirement of an additional ST switching state that is not typically present in conventional PWM schemes.

Among the commonly used sinusoidal PWM (SPWM) schemes for ZSI-based topologies are the simple boost control [8], maximum boost control [9], and maximum constant boost control [10]. However, these schemes do not allow for decoupled control of the amplitude modulation index (M_a) and the duty cycle (D_0), which is essential for closed-loop applications. To overcome these challenges and further enhance the efficiency and controllability of qZSI, it is crucial to explore innovative PWM schemes. Space-vector PWM (SVPWM) schemes offer higher maximum output ac voltage for a given input dc voltage than the SPWM schemes. According to the existing literature, SVPWM-based schemes may be divided into continuous [11,12] and discontinuous [13,14]. The continuous schemes are continuously pulse-width modulated (i.e., the triangular carrier signal and the reference signals intersect within each switching period), whereas the discontinuous involve clamping of a phase leg to the positive or negative dc rail (i.e., the triangular carrier signal and the reference signals do not intersect within each switching period). A substantial number of continuous schemes use two vertically displaced ac reference signals to control the two transistors in the same inverter phase leg [11]. These schemes differ by the number of ST states per switching period (T_{sw}), ranging from two to six. The most prominent scheme of this type named ZSVM6—originally proposed in [12]—has six ST states per T_{sw} , resulting in the lowest inductor current ripple compared to other schemes from this group. Other continuous schemes inject the ST states by utilizing two additional dc reference signals [15]. The primary objective of employing discontinuous schemes is to minimize switching losses. The discontinuous scheme proposed in [13] focuses on reducing the switching losses but with the ST states occurring in one phase leg at a time, whereas in [14,16], the ST states occur in all three legs simultaneously, so the ST-induced current is evenly distributed between the phase legs. The latter feature ensures lower current stress of the switches in the bridge, as well as reduced power losses and thermal stress. In some applications, the PWM scheme has to enable independent variation of M_a and D_0 within feasible limits [10], as, for example, in PV systems with the qZSI [3,16–18]. In [3,17], this was achieved by means of the ZSVM6 scheme, whereas in [16], the SPWM scheme with additional dc signals was utilized for the same purpose.

In the existing literature on qZSIs and their PWM schemes, it has been observed that the synchronization between the beginning of the ST state and the zero switching state is not typically achieved. This lack of synchronization, as demonstrated in [19], leads to the occurrence of additional transitions to zero switching states that would not exist otherwise. Consequently, this results in increased switching losses. To address this issue, a method called “zero-sync” was proposed in [19], which effectively eliminates these additional switching transitions by synchronizing the switching states in question. This synchronization proves to be crucial in enhancing the efficiency of qZSI. Additionally, it was found in [19] that although the application of dead time is not necessary in ZSI-based topologies due to their inherent short-circuit protection, omitting the dead time leads to unintended short-circuiting in the inverter phase legs. This unintended short-circuiting occurs outside of the ST intervals due to the non-ideal switching of the involved transistors,

resulting in an uncontrollable increase in voltage boost and stress. The severity of this problem becomes more pronounced as the voltage across the bridge increases, primarily due to the prolonged turn-off time of the transistors. To mitigate this issue, the minimal necessary dead time was implemented in [19] to prevent unintended short-circuiting and maintain control over the voltage. It is important to note that excessive dead time leads to reduced voltage boost and a decrease in the quality of the output voltage waveforms.

In this study, several continuous and discontinuous PWM schemes proposed in [12–15] were considered, including the decoupled variants of the SVPWM-based schemes from [13,14], which have not yet been considered. Further modifications to the SVPWM-based schemes from [13,14] are proposed to overcome the observed drawbacks and limitations, namely the introduction of dead time and zero-sync operation. This resulted in two novel decoupled SVPWM-based schemes with reduced switching losses and enhanced controllability, and, in some cases, reduced number of reference signals. To evaluate the effectiveness of these novel PWM schemes, experimental validation was conducted. An electronic circuit board was developed specifically for this purpose, enabling the implementation of a wide range of known PWM schemes, as well as the two newly proposed schemes. Through rigorous experimentation and analysis, the performance and advantages of the novel SVPWM-based schemes were assessed, providing valuable insights into their potential for enhancing efficiency and control in qZSI applications. The obtained results highlight the benefits of the proposed schemes over the existing ones with regard to several performance indicators including the system efficiency, voltage/current/thermal stress, inductor current ripple, and the output current's harmonic content.

The remaining sections of the paper are organized as follows. Section 2 provides a comprehensive review of the existing SVPWM schemes that are commonly used for controlling the three-phase qZSI. In Section 3, two proposed decoupled zero-sync SVPWM schemes with dead time are introduced and explained. Section 4 presents the laboratory setup of the qZSI that was utilized for experimental evaluation of the considered PWM schemes, with a particular emphasis on the circuitry employed for generating switching pulses. The experimental results obtained from the evaluation are presented and thoroughly discussed in Section 5. Finally, Section 6 concludes the paper by summarizing the findings and drawing overall conclusions based on the research conducted.

2. Review of SVPWM Schemes for Three-Phase qZSIs

The three-phase qZSI consists of the impedance network and the three-phase inverter bridge. It is in Figure 1 depicted as part of a standalone (i.e., off-grid) system supplying an autonomous three-phase load. The impedance network comprises two inductors ($L_1 = L_2 = L$), two capacitors ($C_1 = C_2 = C$), and a diode (D). The three-phase inverter bridge is composed of six insulated-gate bipolar transistors (IGBTs) featuring integrated freewheeling diodes. In the considered standalone system, an LCL filter is additionally inserted between the inverter output and the load. The filter includes inductors (L_{f1}, L_{f2}), capacitors (C_f), and damping resistors (R_d).

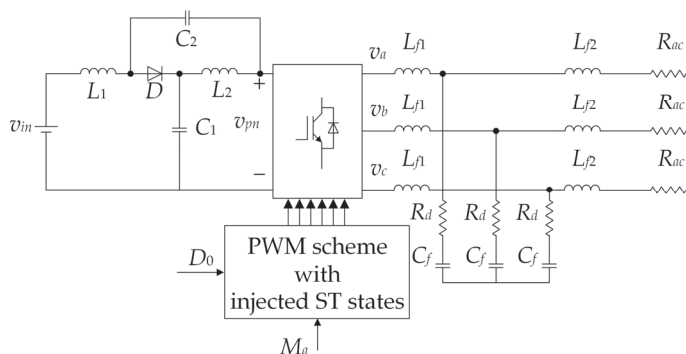


Figure 1. Standalone configuration of the qZSI.

The analyzed qZSI enables boosting of the input dc voltage (v_{in}) through the utilization of the impedance network and the implementation of ST states. These states involve short-circuiting one or all the inverter phase legs, which is strictly prohibited in conventional VSIs. The equivalent circuits of the qZSI in the non-ST state and ST state are shown in Figure 2a and Figure 2b, respectively. In the non-ST state, the impedance network is coupled to the load allowing power to flow from the dc source to the load (except during the zero switching states). This state results in a forward biased diode D (Figure 2a), charging of the capacitors C_1 and C_2 , and a decrease in the currents through the inductors L_1 and L_2 . During the ST state (typically occurring during the zero switching state), the inverter bridge is short-circuited, which causes reverse biasing of the diode D (Figure 2b), discharging of the capacitors C_1 and C_2 , and an increase in i_{L1} and i_{L2} .

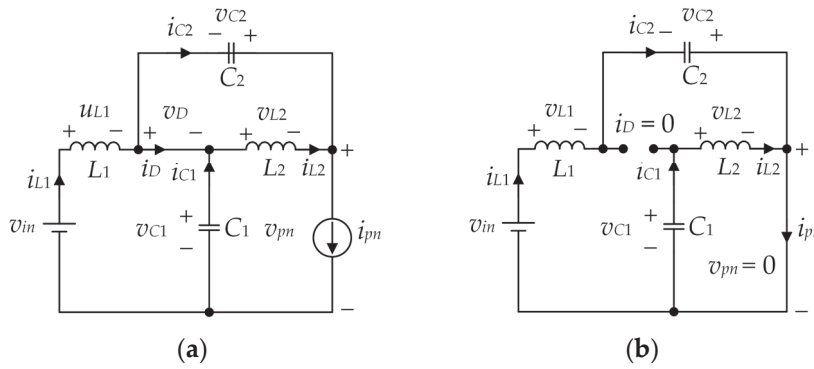


Figure 2. Equivalent circuits of the qZSI: non-ST state (a) and ST state (b).

The input voltage boost (B) and the mean values of the voltages across the capacitors C_1 and C_2 , denoted as V_{C1} and V_{C2} , respectively, are determined as follows:

$$\begin{aligned} B &= \frac{1}{1-2D_0} \\ V_{C1} &= \frac{1-D_0}{1-2D_0} \\ V_{C2} &= \frac{D_0}{1-2D_0} \end{aligned} \quad (1)$$

PWM schemes recently proposed for the qZSI are based on the conventional SPWM or SVPWM schemes but adapted to enable injection of the ST states. However, most are based on the conventional SVPWM scheme due to the lower total harmonic distortion (THD) and $2/\sqrt{3}$ times higher achievable output ac voltage for the same M_a value compared to the conventional SPWM scheme. In fact, similar can be achieved by injecting a third-harmonic component ($1/6$ amplitude) into the fundamental sinusoidal signals of the conventional SPWM scheme. This is because the conventional SVPWM is equivalent to an SPWM with injected triangular component of the triple fundamental frequency [20].

2.1. Continuous SVPWM Schemes

The most straightforward approach to inject ST states of constant duration involves introducing two additional dc signals—one positive and one negative—commonly referred to as the simple boost method [15]. In Figure 3a, the waveforms of reference and carrier signals are illustrated for the case of simple-boost space-vector modulation (SBSVM). In this figure, the M_a value is set to 0.7, while the frequency modulation index (M_f) is set to 10. The conventional switching pulses are obtained through the comparison of ac reference signals (v_{mA} , v_{mB} , and v_{mC}) with the triangular carrier signal (v_{trian}). Simultaneously, the ST state pulses are generated by comparing dc reference signals (v_P and v_N) with v_{trian} .

$$\begin{aligned} v_{mA/B/C} &= v_{smA/B/C} - 0.5[\max(v_{smA}, v_{smB}, v_{smC}) + \min(v_{smA}, v_{smB}, v_{smC})] \\ v_{smA} &= \frac{2}{\sqrt{3}} M_a \sin(\omega t); v_{smB} = \frac{2}{\sqrt{3}} M_a \sin(\omega t - \frac{2\pi}{3}); v_{smC} = \frac{2}{\sqrt{3}} M_a \sin(\omega t + \frac{2\pi}{3}) \end{aligned} \quad (2)$$

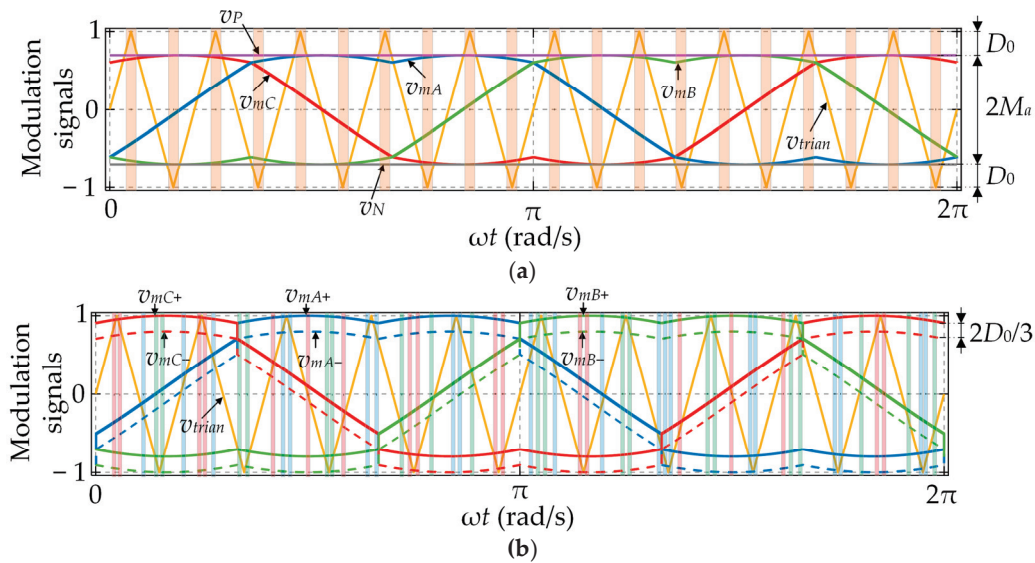


Figure 3. Waveforms of the SBSVM (a) and the ZSVM6 (b), with $M_a = 0.7$, $M_f = 10$ (ST states are denoted by vertical stripes, with different colors denoting different phase legs).

The ST state is activated whenever v_{trian} is higher than v_P or lower than v_N . The switching pulses for the transistors are obtained as the output of the logic OR operation between the ST state pulses and the standard switching pulses. This results in the short circuiting of all the inverter phase legs during the ST state, which occurs twice per switching period.

However, the implementation of the SBSVM scheme is more complicated than the standard VSI SVPWM scheme because it requires two additional reference signals (v_P and v_N) to generate the gate signals.

Another method of injecting ST states involves controlling the upper and lower transistors in the same inverter phase leg using two vertically displaced ac reference signals. The SVPWM scheme depicted in Figure 3b was proposed in [12] and named ZSVM6 in [11]. It is the most widely used SVPWM scheme employing the reference signal displacement and it is notable for having the least amount of inductor current ripple. A small vertical displacement is introduced between the reference signals for the upper and lower transistors, as shown in Figure 3b. It is important to note that this displacement may be adjusted within feasible limits, which allows for decoupled control of M_a and D_0 . The upper transistors are turned on when the corresponding reference signals (v_{mA+} , v_{mB+} , and v_{mC+}) are larger than the triangular carrier signal (v_{trian}), whereas the lower transistors are turned on when the corresponding reference signals (v_{mA-} , v_{mB-} , and v_{mC-}) are smaller than the carrier signal. Consequently, due to the displacement, an overlap is generated between each pair of switches in each phase leg. This results in the ST state every time the normal switching transition is bound to occur, which is analogous to implementing negative dead time. The ST state occurs in one phase leg at a time, whereas in each phase leg it occurs twice per switching period. Therefore, there are six ST states in total per switching period, each lasting one-sixth the total ST time. However, they are not evenly distributed within a switching period due to the variable duration of individual active switching states (Figure 3b). This leads to a variable D_0 value and induces a low-frequency (LF) ripple in both the currents of the inductors and the voltages across the capacitors in the impedance network. Due to the same reason, the effective switching frequency within the impedance network is, in the worst case, reduced to only twice the bridge switching frequency ($2f_{sw}$). This reduction contributes to an increase in the high-frequency (HF) ripple.

The fact that the ZSVM6 scheme requires six reference signals for three-phase qZSIs makes its implementation more complicated compared to the standard VSI SVPWM scheme, which requires half as many reference signals.

2.2. Discontinuous SVPWM Schemes

The previously discussed, SBSVM and ZSVM6 schemes fall into the group of continuous PWM schemes, in which all the phase legs are continuously pulse-width modulated. This is not the case with discontinuous PWM schemes because they involve clamping of a phase leg to the positive or negative dc rail for one-third of each fundamental period, while the remaining two phase legs remain pulse-width modulated [13,14]. As a result, the number of switch commutations is reduced compared to the continuous PWM schemes.

The discontinuous SVPWM scheme proposed in [14] for the three-phase qZSI is illustrated in Figure 4a. It is known as the simple-boost discontinuous space-vector (SBDSV) modulation and requires an additional negative dc reference signal (v_N). The ST states are injected in all three phase legs simultaneously when the carrier signal (v_{trian}) is above the largest ac reference signal ($\max(v_{mA}, v_{mB}, v_{mC})$) or below the dc reference signal. If these conditions are not met, the qZSI is modulated as the conventional VSI. The simultaneous injection of the ST states in all three phase legs results in lower current stress of the bridge transistors compared to ZSVM6. However, since the ST state occurs twice per switching period (as with the SBSVM scheme), the HF ripple in the inductor currents is higher compared to ZSVM6. Still, the need for an additional dc reference signal complicates the implementation of the SBDSV scheme.

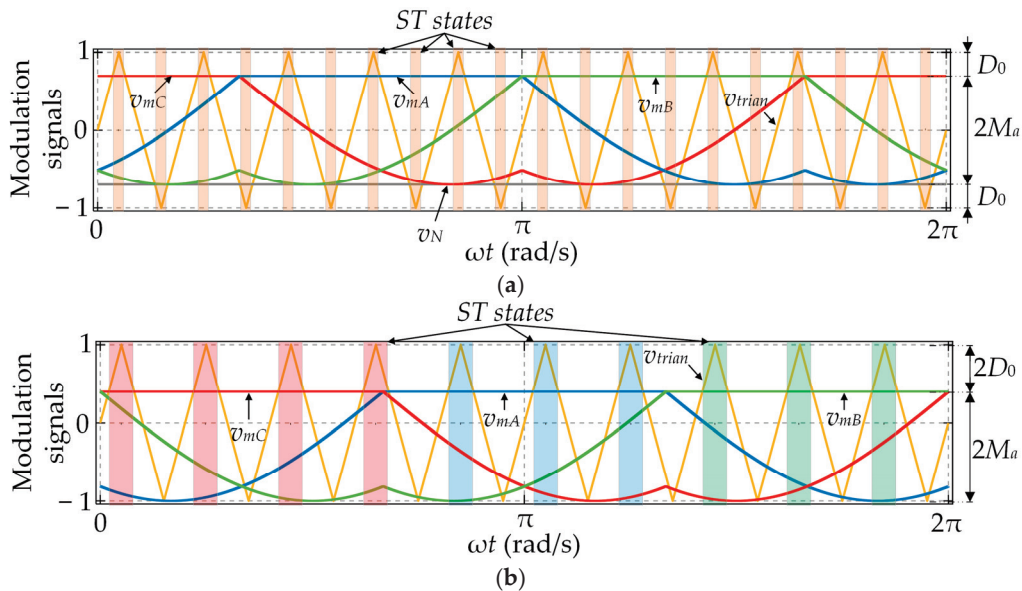


Figure 4. Waveforms of the SBDSV (a) and the SBMSV (b), with $M_a = 0.7$, $M_f = 10$ (ST states are denoted by vertical stripes, with different colors denoting different phase legs).

The corresponding reference signals are given as follows:

$$\begin{aligned} v_{mA/B/C} &= v_{mA/B/C}^{SV} - \max(v_{mA}^{SV}, v_{mB}^{SV}, v_{mC}^{SV}) - v_N \\ v_N &= -M_a \end{aligned} \quad (3)$$

where $v_{mA/B/C}^{SV}$ denotes $v_{mA/B/C}$ obtained from (2).

Figure 4b shows the waveforms of the discontinuous SVPWM scheme proposed in [13]. It is known as the simple-boost modified space-vector (SBMSV) modulation. In this case, the three-phase qZSI is modulated by comparing the reference signals (v_{mA} , v_{mB} , and v_{mC}) with the carrier signal (v_{trian}), as is the case in standard PWM schemes. In other words, there is no requirement for any additional ac or dc reference signals. Furthermore, each of the upper transistors is turned on when the corresponding reference signal is larger than or equal to the carrier signal or the other two reference signals, whereas the lower transistors are turned on when the corresponding reference signal is smaller than the carrier signal. This inevitably leads to overlapping between the two transistors in the same phase leg. The

ST state occurs in one phase leg at a time, as is the case with ZSVM6. This means that the transistors must withstand higher peak currents compared to SBSVM or SBDSV, where the ST-induced current is evenly distributed in all three phase legs. Lastly, the ST state occurs only once per switching period, which increases the HF ripple in the inductor currents compared to the other previously discussed schemes and particularly ZSVM6.

The corresponding reference signals are given as follows:

$$v_{mA/B/C} = -v_{mA/B/C}^{SV} + \min(v_{mA}^{SV}, v_{mB}^{SV}, v_{mC}^{SV}) + 2M_a - 1 \quad (4)$$

where $v_{mA/B/C}^{SV}$ again denotes $v_{mA/B/C}$ obtained from (2).

It is important to note that in all the SVPWM schemes discussed in this section, except for the ZSVM6 scheme, the ST duty cycle does not vary over time. By this, the LF ripple in the inductor currents and the capacitor voltages is effectively eliminated. In addition, in these schemes, dead time is not implemented to prevent short-circuiting in the inverter phase legs that may occur outside of the intended ST intervals and lead to higher voltage boost and stress than expected [19]. Moreover, in the case of the SBSVM and SBDSV schemes, the beginning of the ST state is not always synchronized with the beginning of the zero switching state, which may lead to additional switching transitions and, hence, higher switching losses [19]. Finally, again except for ZSVM6, the decoupled control of D_0 and M_a has not been implemented for the PWM schemes discussed in this section, although such a possibility was mentioned in [14] for the SBDSV scheme and in [13] for the SBMSV scheme.

3. Proposed Decupled Zero-Sync SVPWM Schemes with Dead Time

This section introduces two novel discontinuous SVPWM schemes for the three-phase qZSI. Both involve dead time injection, as proposed in [19], to prevent uncontrollable increase in the voltage boost and stress due to unintended ST states. The ST states are injected according to the zero-sync method, also proposed in [19], so as to reduce the switching losses. Another significant feature of the two proposed SVPWM schemes is decoupled control of D_0 and M_a , provided that $D_0 < D_{0max}$ ($D_{0max} = 1 - M_a$), which opens the possibility of application in closed-loop control systems. The first proposed scheme, named DSV2ST, utilizes two ST states per switching period and is based on the SBDSV scheme from [14], whereas the second proposed scheme, named DSV1ST, utilizes a single ST state per switching period and is based on the SBMSV scheme from [13].

3.1. Decoupled SVPWM with Two ST States per Switching Period (DSV2ST)

Starting from the SBDSV scheme described in Section 2.2, the decoupling of D_0 and M_a from Figure 4a can be achieved by shifting the ac reference signals (v_{mA} , v_{mB} , and v_{mC}) upwards and the dc reference signal (v_N) downwards by the same amount, as illustrated in Figure 5a. The resulting reference signals are given as

$$\begin{aligned} v_{mA/B/C} &= v_{mA/B/C}^{SV} - \max(v_{mA}^{SV}, v_{mB}^{SV}, v_{mC}^{SV}) - v_N \\ v_N &= D_0 - 1 \end{aligned} \quad (5)$$

Note the similarity between (5) and (3), where the only difference is in the v_N value, which, in turn, affects the vertical shift of the ac reference signals. In this case, v_N is defined by D_0 and not by M_a , with the latter parameter contained in $v_{mA/B/C}^{SV}$. This shows that D_0 and M_a may be independently set within feasible limits. Therefore, the PWM scheme presented in Figure 5a may be referred to as the decoupled SBDSV scheme. Still, apart from the decoupled control, the scheme in Figure 5a shares the same demerits of the SBDSV scheme, discussed in Section 2.2.

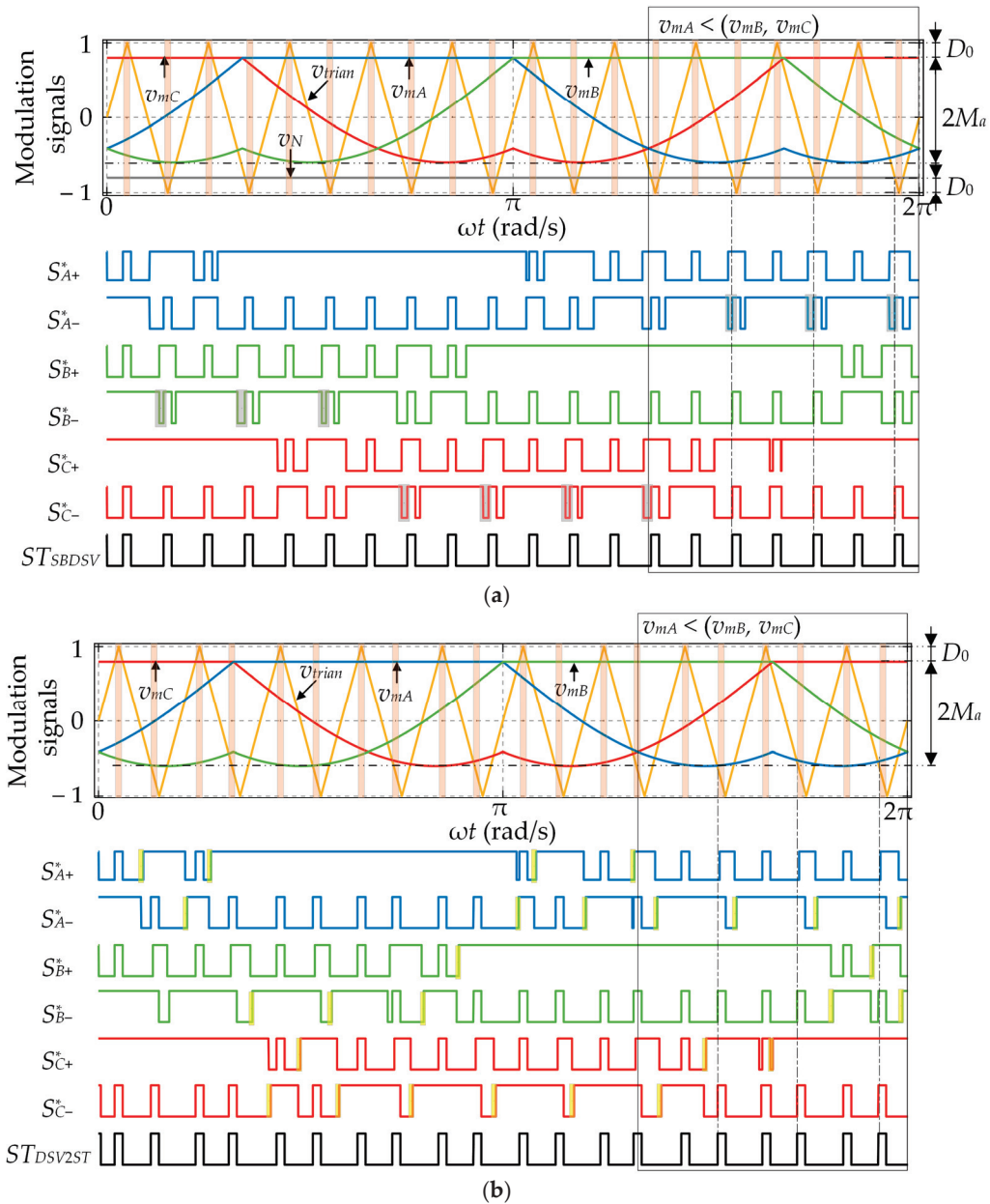


Figure 5. Waveforms of the decoupled SBDSV (a) and the proposed DSV2ST (b), with $M_a = 0.7$, $M_f = 10$, $D_0 = 0.2$ (ST states are denoted by vertical stripes).

In the proposed DSV2ST scheme with dead time, illustrated in Figure 5b, a dc reference signal is removed, whereas a timer is utilized instead to ensure adjustable but constant ST state duration, as shown in Figure 6. The ST state occurs twice per switching period in all three phase legs when the carrier signal is higher than the upper envelope of the ac reference signals or lower than the dc reference signal. Moreover, the ST states are synchronized with the zero switching states, thanks to which all the switching transitions marked with gray color in the lower part of Figure 5a are effectively eliminated. It can be seen that only the lower transistors in the bridge are affected by the proposed synchronization, so the number of eliminated switching transitions is in this case equal to $2M_f$, which is twice lower than the number achieved in [19]. Dead-time intervals are marked with yellow color in the lower part of Figure 5b. The corresponding reference signals are given as follows:

$$v_{mA/B/C} = v_{mA/B/C}^{SV} - \max(v_{mA}^{SV}, v_{mB}^{SV}, v_{mC}^{SV}) - D_0 + 1 \quad (6)$$

In (5) and (6), $v_{mA/B/C}^{SV}$ denotes $v_{mA/B/C}$ obtained from (2).

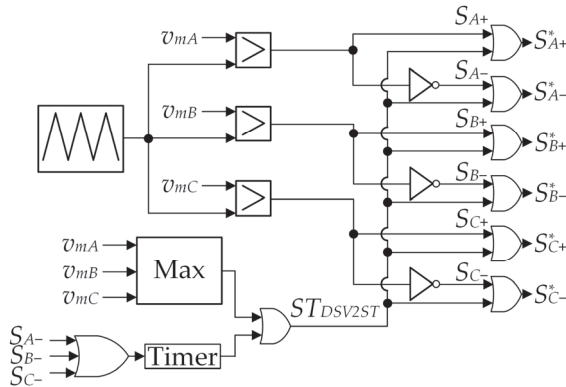


Figure 6. PWM pulses generation for the proposed DSV2ST scheme.

3.2. Decoupled SVPWM with a Single ST State per Switching Period (DSV1ST)

Starting from the SBMSV scheme described in Section 2.2 (Figure 4b), the respective reference signals can be shifted vertically so that the minimum of the lower envelope is not necessarily aligned with -1 . Such modification is illustrated in Figure 7a, whereas the corresponding reference signals are defined as

$$v_{mA/B/C} = -v_{mA/B/C}^{SV} + \min(v_{mA}^{SV}, v_{mB}^{SV}, v_{mC}^{SV}) + 1 - 2D_0 \quad (7)$$

This PWM scheme may be referred to as the decoupled SBMSV scheme. However, it suffers from the same other issues discussed in Section 2.2 for the SBMSV scheme.

Given the above, a novel DSV1ST scheme with dead time is proposed in which the upper envelope of the reference signals is aligned with 1, as illustrated in Figure 7b. The ST state occurs once per switching period when the carrier signal is below the lower envelope of the reference signals, but, as opposed to SBMSV, it occurs in all three phase legs. This results in lower current stress of the bridge transistors. On the downside, the number of switching transitions of the lower transistors in the bridge is higher compared to SBMSV. Again, a timer is utilized to ensure constant and adjustable ST state duration, as shown in Figure 8, while the ST states are synchronized with the zero switching states. Dead-time intervals are marked with yellow color in the lower part of Figure 7b. The corresponding reference signals are defined as

$$v_{mA/B/C} = -v_{mA/B/C}^{SV} + \min(v_{mA}^{SV}, v_{mB}^{SV}, v_{mC}^{SV}) + 1 \quad (8)$$

Once again, $v_{mA/B/C}^{SV}$ in (7) and (8) corresponds to $v_{mA/B/C}$ obtained from (2). Also note that (8) does not contain D_0 because its value is not defined by the reference signals, but is instead controlled by utilizing the timer.

3.3. Proposed vs. Existing PWM Schemes—A Comparative Study

The equations describing the HF ripple in the inductor currents and the capacitor voltages within the impedance network can be derived based on the voltage-current conditions in the equivalent circuits in Figure 2 and are given as follows:

$$\Delta I_{L,hf} = \frac{V_{C1} T_{ST}}{n_{ST} L} \quad (9)$$

$$\Delta V_{C,hf} = \frac{I_{L1} T_{ST}}{n_{ST} C} \quad (10)$$

where T_{ST} represents the total duration of ST states per switching period and n_{ST} denotes the number of ST states per switching period (see Table 1).

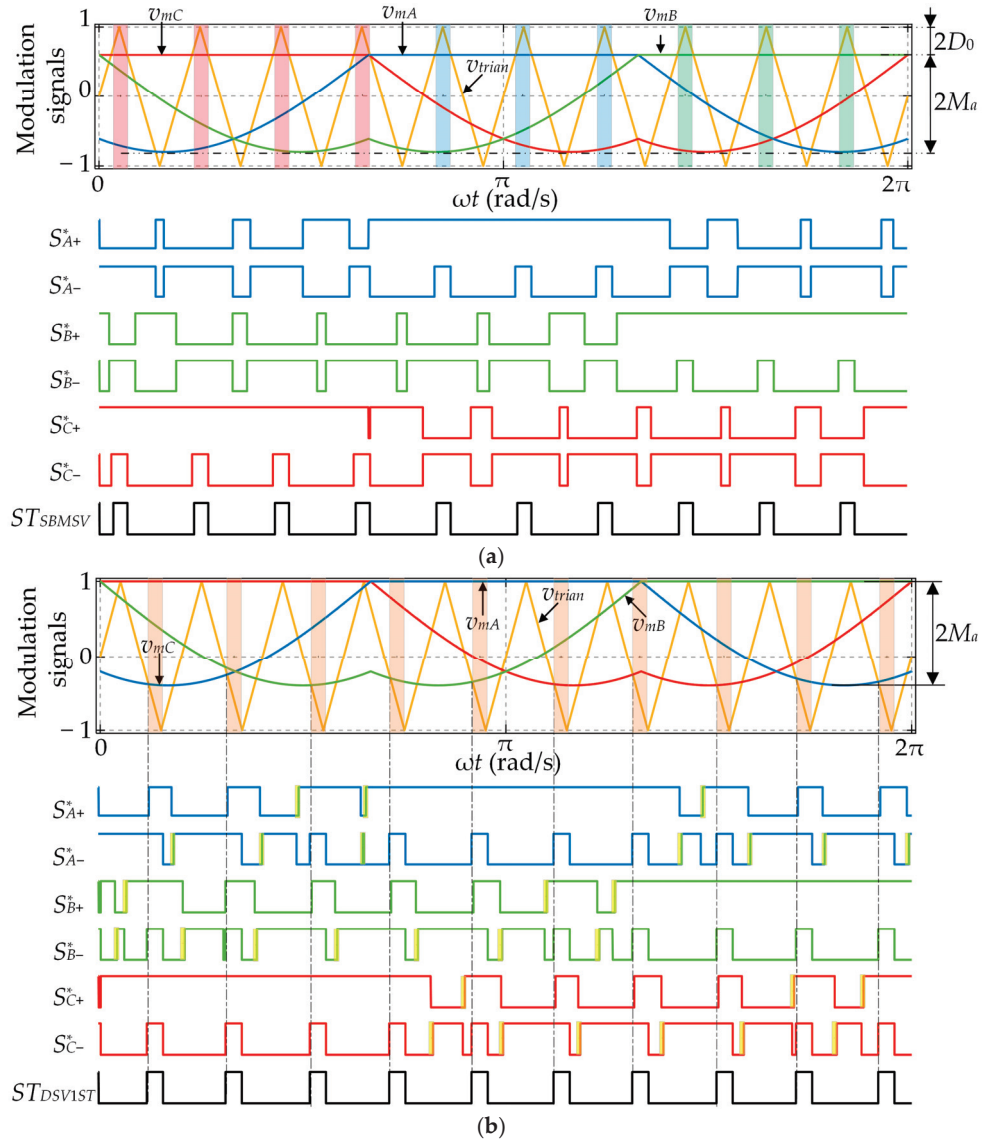


Figure 7. Waveforms of the decoupled SBMSV (a) and the proposed DSV1ST (b), with $M_a = 0.7$, $M_f = 10$, $D_0 = 0.2$ (ST states are denoted by vertical stripes, with different colors denoting different phase legs).

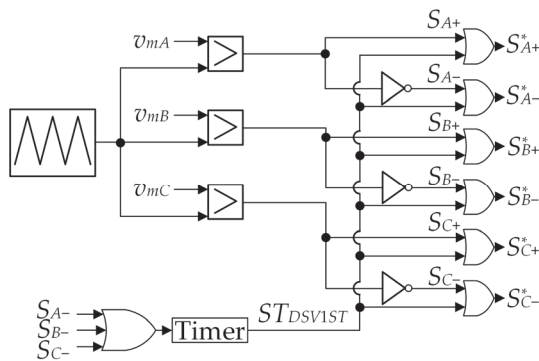


Figure 8. PWM pulses generation for the proposed DSV1ST scheme.

Table 1. Comparison of the PWM schemes' features.

PWM Scheme	SBSVM [15]	SBDSV [14]/ dec. SBDSV	SBMSV [13]/ dec. SBMSV	ZSVM6 [12]	Proposed DSV2ST	Proposed DSV1ST
Modulation signals	Figure 3a	Figure 4a/ Figure 5a	Figure 4b/ Figure 6a	Figure 3b	Figure 5b	Figure 6b
Number of modulation signals	5	4	3	6	3	3
Number of ST states per T_{sw}	2	2	1	6	2	1
Number of phase legs used for ST injection	3	3	1	1	3	3
ST state duration	constant	constant	constant	constant	constant	constant
Duty cycle variation	constant	constant	constant	Variable ⁽¹⁾	variable	variable
Decoupled M_a and D_0	No	No/Yes	No/Yes	Yes	Yes	Yes
Additional zero switching states	Yes	Yes	Yes	No	No	No
Unintentional ST states	Yes	Yes	Yes	No	No	No
Number of switchings in the bridge per T_{sw}	24	20	10	12	18	12
Upper switches effective switching frequency	$2f_{sw}$	$4f_{sw}/3$	$2f_{sw}/3$	f_{sw}	$4f_{sw}/3$	$2f_{sw}/3$
Lower switches effective switching frequency	$2f_{sw}$	$2f_{sw}$	f_{sw}	f_{sw}	$5f_{sw}/3$	$4f_{sw}/3$
Input dc side seen switching frequency	$2f_{sw}$	$2f_{sw}$	f_{sw}	$2f_{sw}$ ⁽²⁾	$2f_{sw}$	f_{sw}
Output ac side seen switching frequency	f_{sw}					
Peak current in each switch ⁽³⁾	$I_{ph}^{(4)}/2 + (2I_L^{(5)} + \Delta I_L^{(6)})/3$	$I_{ph}/2 + (2I_L + \Delta I_L)/3$	$2I_L + \Delta I_L$	$2I_L + \Delta I_L$	$I_{ph}/2 + (2I_L + \Delta I_L)/3$	$I_{ph}/2 + (2I_L + \Delta I_L)/3$

⁽¹⁾ In [12] is incorrectly stated that the duty cycle variation of the ZSVM6 is constant. ⁽²⁾ This is considered the worst case (it varies from $2f_{sw}$ to $6f_{sw}$ due to the varying duration of active switching states). ⁽³⁾ In [12], the peak current in each switch is incorrectly defined for the ZSVM6 and SBMSV. ⁽⁴⁾ Peak value of the inverter phase current. ⁽⁵⁾ Mean value of the inductor current. ⁽⁶⁾ Inductor current ripple (peak-to-peak).

In the PWM schemes where the ST states are injected based on the comparison of the triangular carrier signal and the reference signals with constant upper and lower envelopes (SBDSV and SBMSV), the LF ripple does not exist due to the constant D_0 value. However, in PWM schemes like ZSVM6, DSV1ST, and DSV2ST, the LF ripple component is present despite the constant duration of the ST state. This is due to the varying D_0 value caused by uneven duration of the non-ST states (e.g., unequal spacing between the ST states in Figure 3b). Nevertheless, it is worth noting that the LF ripple is relatively small for the considered PWM schemes, whereas its mathematical derivation is beyond the scope of this study.

Table 1 compares some of the key features of the SVPWM schemes under consideration so as to clarify and highlight the advantages of the proposed schemes. The proposed

schemes enable decoupled control of M_a and D_0 while also eliminating additional zero switching states and unintended ST states. Among the other schemes listed in Table 1, only ZSVM6 meets all these criteria, but it does so at the expense of needing twice as many reference signals and experiencing notably higher peak currents in the bridge switches, resulting in higher semiconductor losses. As for the current stress imposed by the proposed methods, it is among the lowest recorded, as demonstrated by the experimental waveforms presented later in Section 5.

4. Experimental Setup

4.1. Laboratory Setup of the Quasi-Z-Source Inverter

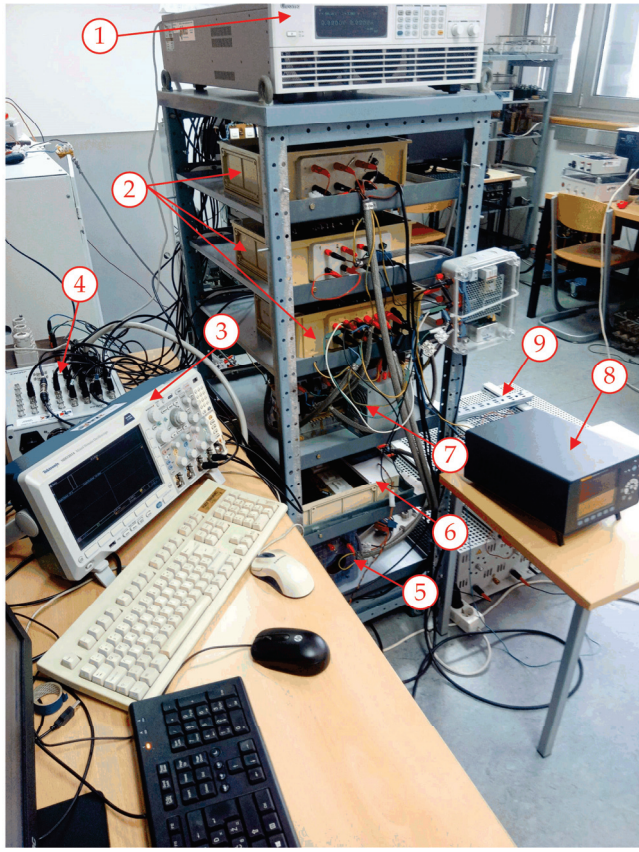
The experimental setup of the standalone qZSI system from Figure 1, utilized for the evaluation of the proposed PWM schemes, is shown in Figure 9a. The meaning of the numerical markings shown in Figure 9a is as follows:

1. Dc power supply 62050H 600S (Chroma, Foothill Ranch, CA, USA); maximum power of 5000 W with voltages up to 600 V and currents up to 8.5 A.
2. Hall effect transducers LA 55-P/SP52 (for the inductor current) (LEM International SA, Plan-les-Ouates, Switzerland), CV 3–500 (for the qZSI input voltage) (LEM International SA, Meyrin, Switzerland), and DVL 500 (for the voltages across C_1 and C_2) (LEM International SA, Meyrin, Switzerland).
3. Oscilloscope MDO 3014 (Tektronix, Beaverton, OR, USA).
4. MicroLabBox controller board (dSpace, Paderborn, Germany) for the qZSI control.
5. qZSI impedance network inductors, built with powder cores T520-26 (Micrometals, Anaheim, CA, USA), with $L_1 = L_2 = 20$ mH (unsaturated), $R_L = 0.5 \Omega$ (at 25 °C), and polypropylene capacitors MKSPI35-50U/1000 (Miflex, Kutno, Poland) with $C_1 = C_2 = 50 \mu\text{F}$ and $\text{ESR} = 7.8 \text{ m}\Omega$.
6. qZSI three-phase inverter bridge (IXBX75N170 IGBTs (Littelfuse, Chicago, IL, USA) with SKHI 22B(R) (Semikron-Danfoss, Nuremberg, Germany) gate drivers) and the custom-developed electronic circuit board for the PWM pulses generation (Figure 9b).
7. Self-made LCL filter at the qZSI output stage ($L_{f1} = 8.64 \text{ mH}$, $L_{f2} = 4.32 \text{ mH}$, $R_{f1} = 0.1036 \Omega$, $R_{f2} = 0.0518 \Omega$, $C_f = 4 \mu\text{F}$, $R_d = 10 \Omega$).
8. Power analyzer Norma 4000 (Fluke, Everett, WA, USA) utilized for the measurement of the load power.
9. Adjustable three-phase resistive load.

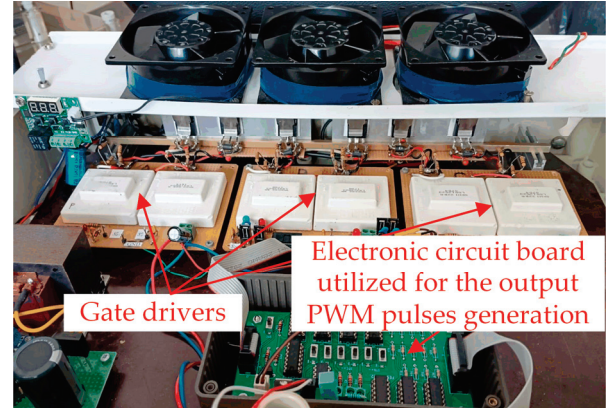
4.2. Hardware Circuitry for Gate Pulses Generation

In this study, an electronic circuit board was specially designed and built to facilitate the implementation of a range of existing PWM schemes for the qZSI, as well as the two proposed schemes. The logic diagram for generating the PWM pulses is depicted in Figure 10. Six PWM channels of the EMC Multi Channel PWM block are utilized to generate the input PWM pulses, namely S_{A+} , S_{A-} , S_{B+} , S_{B-} , S_{C+} , and S_{C-} . This allows for individual control of each transistor in the inverter bridge, enabling the implementation of any PWM scheme that utilizes two vertically displaced ac reference signals in the same inverter leg. The remaining two PWM channels are used to generate the S_P and S_N pulses, which correspond to the ST signals in the case of PWM schemes that utilize the dc reference signals for ST injection (e.g., SBSVM). It is important to note that the input variables for the EMC Multi Channel PWM block in Figure 10 are the duty cycles d_A , d_B , and d_C , calculated in the MATLAB-Simulink (version 8.6 (R2015b)) environment, which can be obtained from the corresponding ac reference signals v_{mA} , v_{mB} , and v_{mC} as

$$d = 0.5(1 + v_m) \quad (11)$$



(a)



(b)

Figure 9. Laboratory setup of the considered qZSI system (a) and the qZSI bridge with the corresponding gate drivers and electronic circuit board (b).

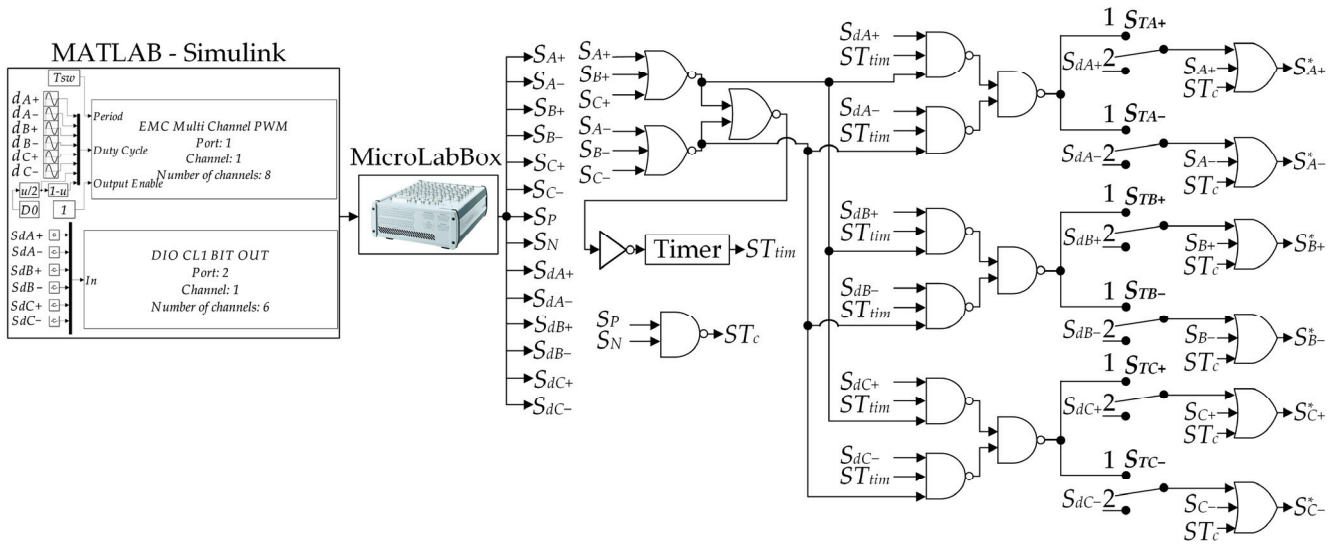


Figure 10. Logic diagram of the PWM pulses generation.

Digital outputs of the MicroLabBox controller board (i.e., *DIO CL1 BIT OUT* block in Figure 10) are utilized to generate six additional digital signals, namely S_{dA+} , S_{dA-} , S_{dB+} , S_{dB-} , S_{dC+} , and S_{dC-} . These signals are, for example, required for the implementation of the decoupled variants of the SBMSV and SBDSV schemes.

The logic diagram includes NOR gates in the upper left part to detect the occurrence of zero switching states. The timer is initiated by the logic signal 1 at the beginning of these states. From that moment, the timer holds the logic signal 1 at its output for a precisely set time, so it can be used to control the duration of ST states. This facilitates the injection of zero-synced ST states of adjustable duration, represented by the output signal ST_{tim} . As required by some PWM schemes, this signal can be additionally combined with the digital signals S_{dA+} , S_{dA-} , S_{dB+} , S_{dB-} , S_{dC+} , and S_{dC-} . If the timer is used for the ST injection, the switches S_{TA+} , S_{TA-} , S_{TB+} , S_{TB-} , S_{TC+} , and S_{TC-} are set to position 1, whereas they are set to position 2 if the ST states are injected by other means. In the output stage, OR gates are used to generate the output PWM pulses S^*_{A+} , S^*_{A-} , S^*_{B+} , S^*_{B-} , S^*_{C+} , and S^*_{C-} .

Figure 11 shows a photo of the developed electronic circuit board for PWM pulses generation. Apart from the utilized integrated circuits, connectors, and switches, the photo shows the resistor R_t and the capacitor C_t . These two components determine the time constant $T_t = R_t \cdot C_t$, by which the desired achievable range of D_0 can be easily implemented. The reader is referred to [19] for more details.

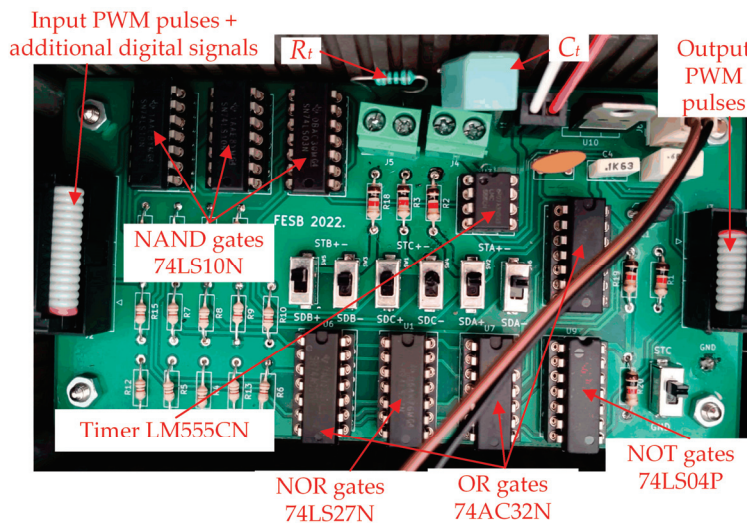


Figure 11. Photo of the electronic circuit board utilized for PWM pulses generation.

Six PWM schemes were considered for the experimental analysis in this study. These are as follows:

1. ZSPWM—continuous zero-sync SPWM scheme with injected third harmonic and dead time, proposed in [19]; it has two ST states per switching period, each occurring in all three phase legs simultaneously (similar to the SBSVM scheme from Figure 3a, but with implemented zero-sync and dead time).
2. ZSVM6—continuous decoupled SVPWM scheme reported in [12], and illustrated in Figure 3b; it incorporates six ST states per switching period, each occurring in one phase leg at a time.
3. dec. SBDSV—decoupled variant of the discontinuous SVPWM scheme from [14], presented in Figure 5a, with two ST states per switching period, each occurring in all three phase legs simultaneously.
4. dec. SBMSV—decoupled variant of the discontinuous SVPWM scheme from [13], illustrated in Figure 5b, with a single ST state per switching period, occurring in a single phase leg.
5. DSV2ST—proposed decoupled and discontinuous zero-sync SVPWM scheme with dead time and two ST states per switching period, each occurring in all three phase legs simultaneously (Figure 5b).

6. DSV1ST—proposed decoupled and discontinuous zero-sync SVPWM scheme with dead time and a single ST state per switching period, occurring in all three phase legs simultaneously (Figure 7b).

Table 2 provides the corresponding switches position, additional digital signal values, and the values of the S_P and S_N signals. Note that d_{trian} in Table 2 is obtained from v_{trian} , as per (11). Three out of the six considered PWM schemes include dead time implementation, namely, ZSPWM, DSV2ST, and DSV1ST. The corresponding value was in this study set to $0.7 \mu\text{s}$ to prevent unintended short-circuiting while avoiding previously mentioned negative effects of excessive dead time. The same three PWM schemes utilize the timer for injecting the zero-synced ST states. It is important to note that the developed electronic circuit board, shown in Figure 11, facilitates the implementation of various other PWM schemes that were not considered in this study.

Table 2. Switches positions and additional digital signal values for considered PWM schemes.

PWM Scheme	ZSPWM	dec. SBDSV	dec. SBMSV	ZSVM6	DSV2ST	DSV1ST
S_{TA+}	1	2	2	2	2	2
S_{TA-}	1	2	2	2	1	1
S_{TB+}	1	2	2	2	2	2
S_{TB-}	1	2	2	2	1	1
S_{TC+}	1	2	2	2	2	2
S_{TC-}	1	2	2	2	1	1
S_{dA+}	1	0	1 if $d_A > (d_B, d_C)$; else 0	0	0	0
S_{dA-}	1	0	0	0	1	1
S_{dB+}	1	0	1 if $d_B > (d_A, d_C)$; else 0	0	0	0
S_{dB-}	1	0	0	0	1	1
S_{dC+}	1	0	1 if $d_C > (d_A, d_B)$; else 0	0	0	0
S_{dC-}	1	0	0	0	1	1
S_P	1	0 if $d_{trian} > \max(d_A, d_B, d_C)$; else 1	1	1	0 if $d_{trian} > \max(d_A, d_B, d_C)$; else 1	1
S_N	1	0 if $D_0/2 > d_{trian}$; else 1	1	1	1	1

5. Experimental Results and Discussion

The experimental results presented in this section are obtained for three distinct switching frequencies, namely 5 kHz, 10 kHz, and 15 kHz, and three distinct load powers, namely 1000 W, 2000 W, and 3000 W. For all the considered operating points, the input dc voltage (v_{in}) remained constant at 500 V, and the amplitude modulation index (M_a) was maintained at 0.71. At the same time, the duty cycle (D_0) was varied in the range 0.1–0.25, which was enabled by the decoupled control feature. To prevent excessive power losses and to avoid overheating the semiconductor switches, D_0 was limited to 0.25, which is 0.04 less than the theoretical maximum value.

5.1. System Efficiency and Inverter Voltage Stress

Figure 12 illustrates the system efficiency, which is calculated as the ratio between the load power and the dc source power, thereby encompassing the filter losses as well. However, it is important to note that the filter losses are generally stable for specific load

power and f_{sw} settings, regardless of the PWM scheme used. Consequently, the variations in the characteristics shown in Figure 12 accurately reflect the differences in inverter efficiency for the various PWM schemes under consideration. It is evident for all six PWM schemes that the corresponding efficiency reduces with increasing duty cycle. This reduction in efficiency is attributed to the greater voltage stress (as shown in Figure 13), leading to increased switching losses.

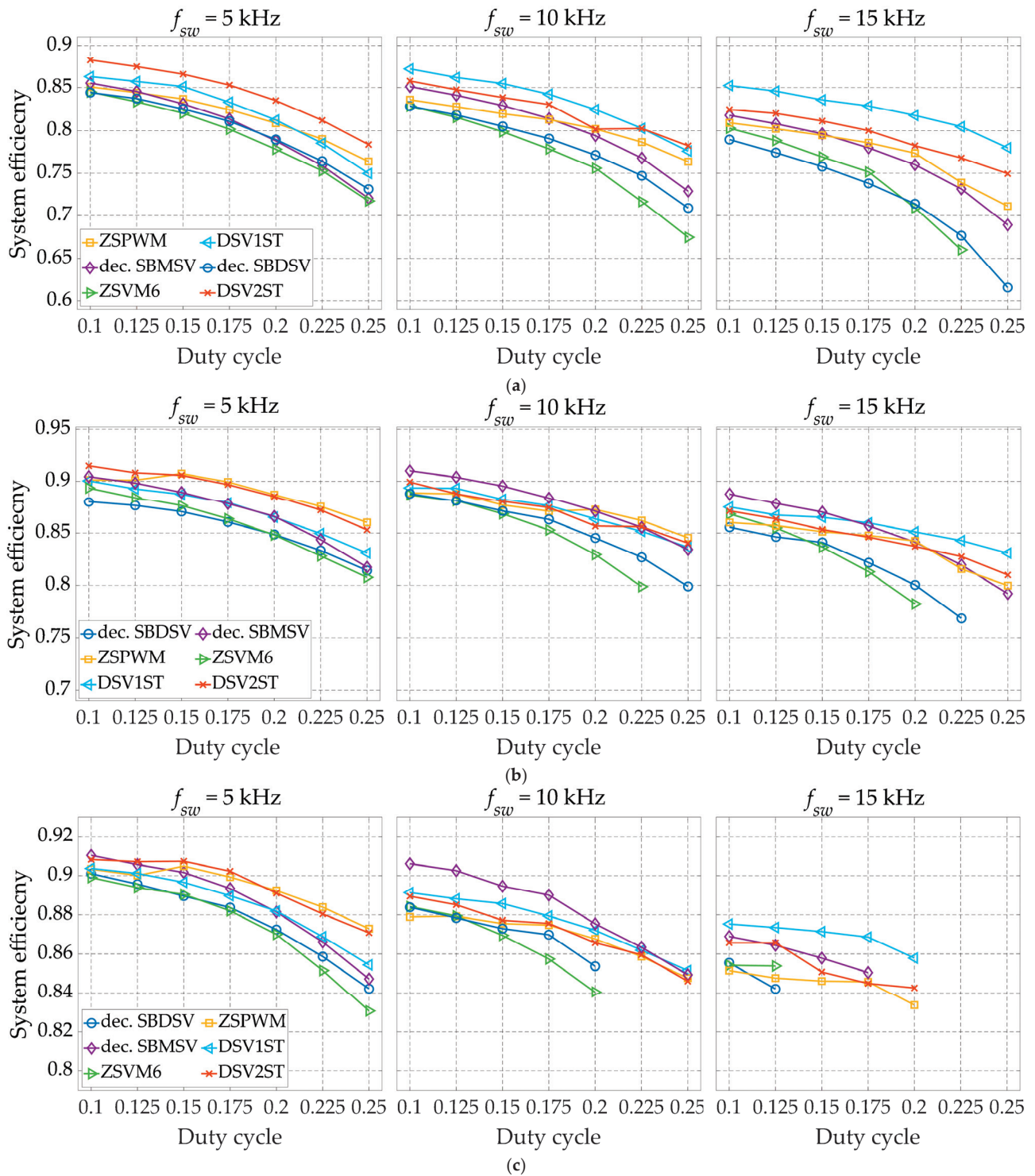


Figure 12. System efficiency as a function of the duty cycle for the different switching frequencies and inverter power of 1000 W (a), 2000 W (b), and 3000 W (c).

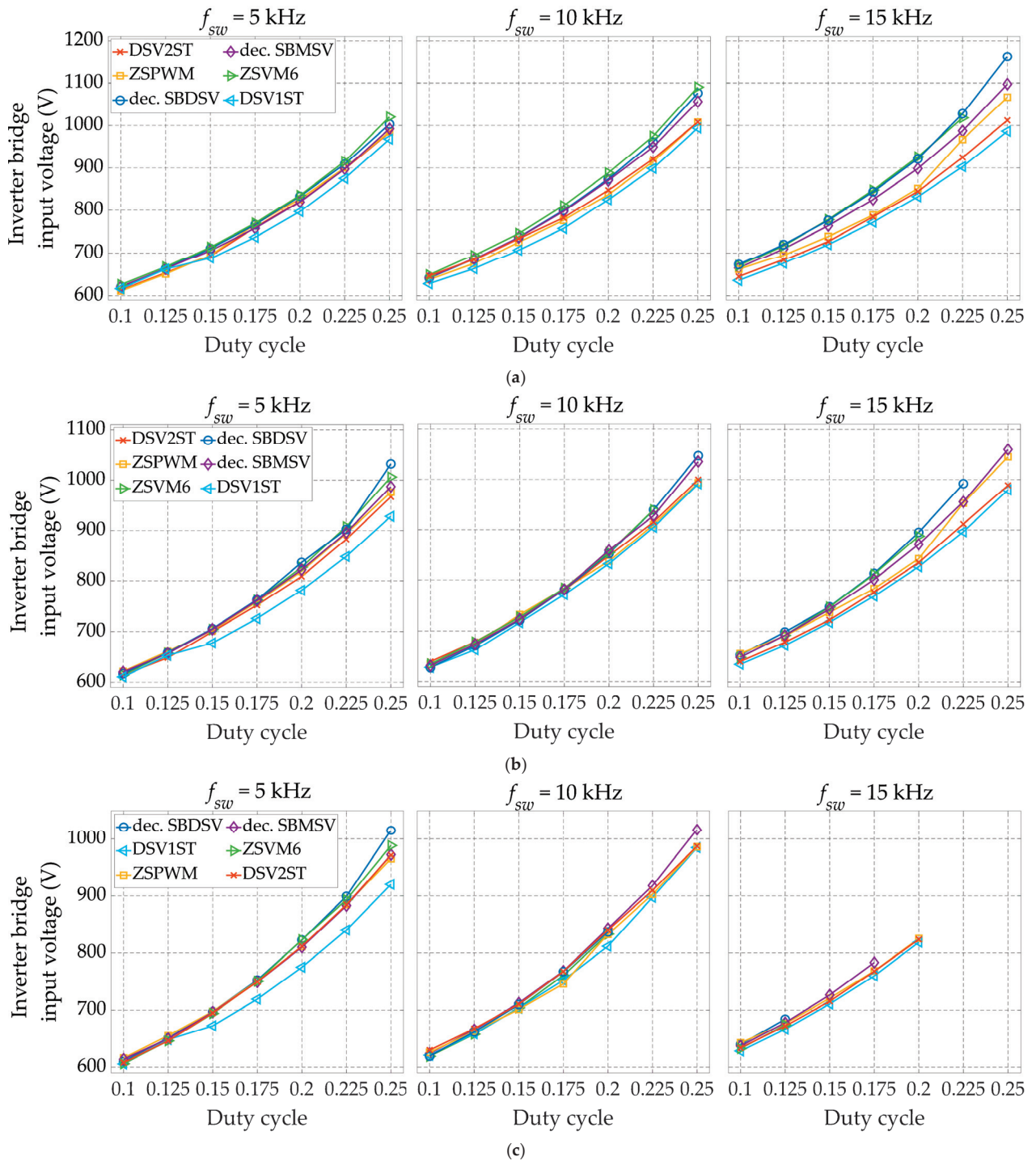


Figure 13. Inverter bridge input voltage as a function of the duty cycle for the different switching frequencies and inverter power of 1000 W (a), 2000 W (b), and 3000 W (c).

The ZSVM6 scheme exhibits the lowest efficiency, which is due to the six ST states per switching period (i.e., two ST states per phase leg), each occurring in one phase leg at a time. This results in increased current stress, switching losses, and conduction losses of the involved transistors, especially at higher D_0 values (i.e., longer ST states). As the duty cycle and switching frequency increase, the efficiency difference between ZSVM6 and the other five PWM schemes grows. Additionally, the ZSVM6 scheme yields the highest voltage stress overall.

The decoupled SBDSV scheme shows only marginal efficiency improvement, despite having three times less ST states per switching period than ZSVM6. On the other hand, the number of switch commutations per switching period is almost doubled with this PWM scheme compared to ZSVM6. In addition, it is the only considered scheme that does not include zero-sync or dead time (ZSVM6 operates with “negative” dead time and does not depend on zero switching states for ST state injection). Both SBDSV and ZSVM6 suffer the most significant efficiency drop with increasing switching frequency. In fact, at the highest considered switching frequency ($f_{sw} = 15$ kHz), certain data points at higher D_0 values could not even be recorded for these two schemes due to the risk of exceeding the maximum allowable temperature of the IGBT-diode pairs in the bridge.

Having only one ST state per switching period proved to be an advantageous feature of the decoupled SBMSV scheme in the context of switching losses, particularly at higher switching frequencies. Moreover, the ST states are for this scheme inherently synchronized with the zero switching states, leading to comparatively higher efficiency than ZSVM6 and SBDSV. However, unlike the remaining three schemes, the decoupled SBMSV scheme does not incorporate dead time, which explains its lower efficiency in most cases.

The PWM schemes with implemented (positive) dead time, namely ZSPWM, DSV2ST, and DSV1ST, show reduced voltage stress, with DSV1ST particularly excelling in this area. These schemes generally achieve higher efficiency, with DSV1ST performing best at higher switching frequencies, whereas DSV2ST is the most efficient at lower switching frequencies. The latter scheme ensures the lowest voltage stress, whereas the DSV1ST scheme is second best with regard to that criterion. The ZSPWM scheme, while reducing voltage stress compared to the considered existing PWM schemes, still has higher voltage stress than the DSV1ST scheme and similar to that of the DSV2ST scheme. Its efficiency is medium to high at low switching frequencies and load powers, but declines notably with increasing switching frequency, particularly at heavier loads.

5.2. Inductor Current Ripple and Output Current THD

Figure 14 demonstrates that the ZSVM6 scheme excels in terms of inductor current ripple, outperforming other schemes based on this metric. The superior performance of ZSVM6 in this aspect was expected and is attributed to its high number of ST states per switching period, which, on average, results in the highest effective switching frequency of the impedance network. On the other hand, the worst PWM schemes by the same metric are those with only one ST state per switching period, namely the proposed DSV1ST scheme and the decoupled SBMSV scheme. However, it is exactly due to the high number of ST states per switching period, occurring in individual phase legs and leading to relatively high switching and conduction losses, that the ZSVM6 scheme is characterized by the narrowest achievable range of operation. For the remaining three PWM schemes, the inductor current ripple values are close to those recorded for ZSVM6, but they manage to maintain a broader operating range, which is only slightly compromised at the highest considered switching frequency and load power. Overall, as the switching frequency increases, the discrepancies in ripple values among the different PWM schemes become less pronounced.

The system under consideration has a nominal power of 3 kW. If, for example, a tolerance for the inductor current ripple at the nominal power is set at 20%, the results in Figure 14 indicate that only the ZSVM6 scheme can achieve this requirement across most of the considered D_0 range at $f_{sw} = 5$ kHz. However, when the switching frequency is increased to 10 kHz, the number of PWM schemes capable of meeting this ripple requirement expands to four, including the proposed DSV2ST scheme. Nonetheless, this increase in switching frequency also limits the highest attainable D_0 value to 0.2 for ZSVM6 and SBDSV. It is important to note that for DSV1ST and SBMSV, the inductance in the impedance network would still need to be doubled to reach the target ripple value. Further increasing the switching frequency to 15 kHz allows all considered PWM schemes to achieve the desired

inductor current ripple. However, at this frequency, none of them can exceed $D_0 = 0.2$. Specifically, for ZSVM6 and SBDSV, the maximum D_0 value is further restricted to 0.125.

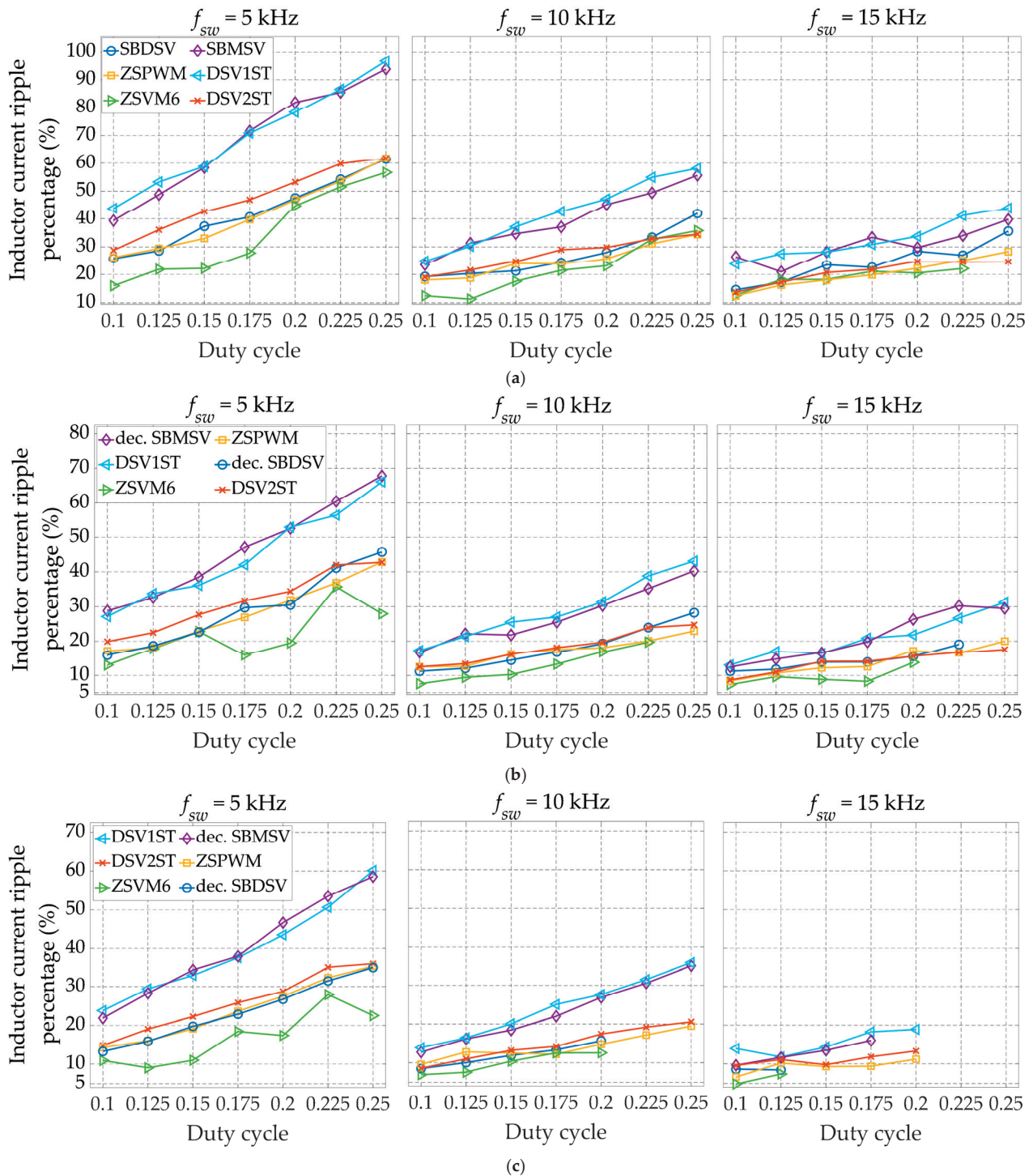


Figure 14. Inductor current ripple percentage as a function of the duty cycle for the different switching frequencies and inverter power of 1000 W (a), 2000 W (b), and 3000 W (c).

Figure 15 shows the recorded THD values of the load phase currents. For all PWM schemes, in most observed cases, there is a trend of increasing THD with the duty cycle. The proposed DSV1ST scheme ranks as the least effective in terms of THD. However, even in the worst case considered (i.e., 5 kHz and 1000 W), the corresponding THD value remained below 4%. Interestingly, the ZSVM6 scheme is the only one exhibiting a noticeable

improvement in THD as the D_0 increases. Despite ZSVM6 being one of the better schemes in this regard, it is important to remind that this comes at the expense of a reduced D_0 range at higher switching frequencies. The ZSPWM and SBMSV schemes also stand out as being among the top three in terms of THD performance, whereas the other proposed scheme, DSV2ST, demonstrated moderate performance with respect to THD.

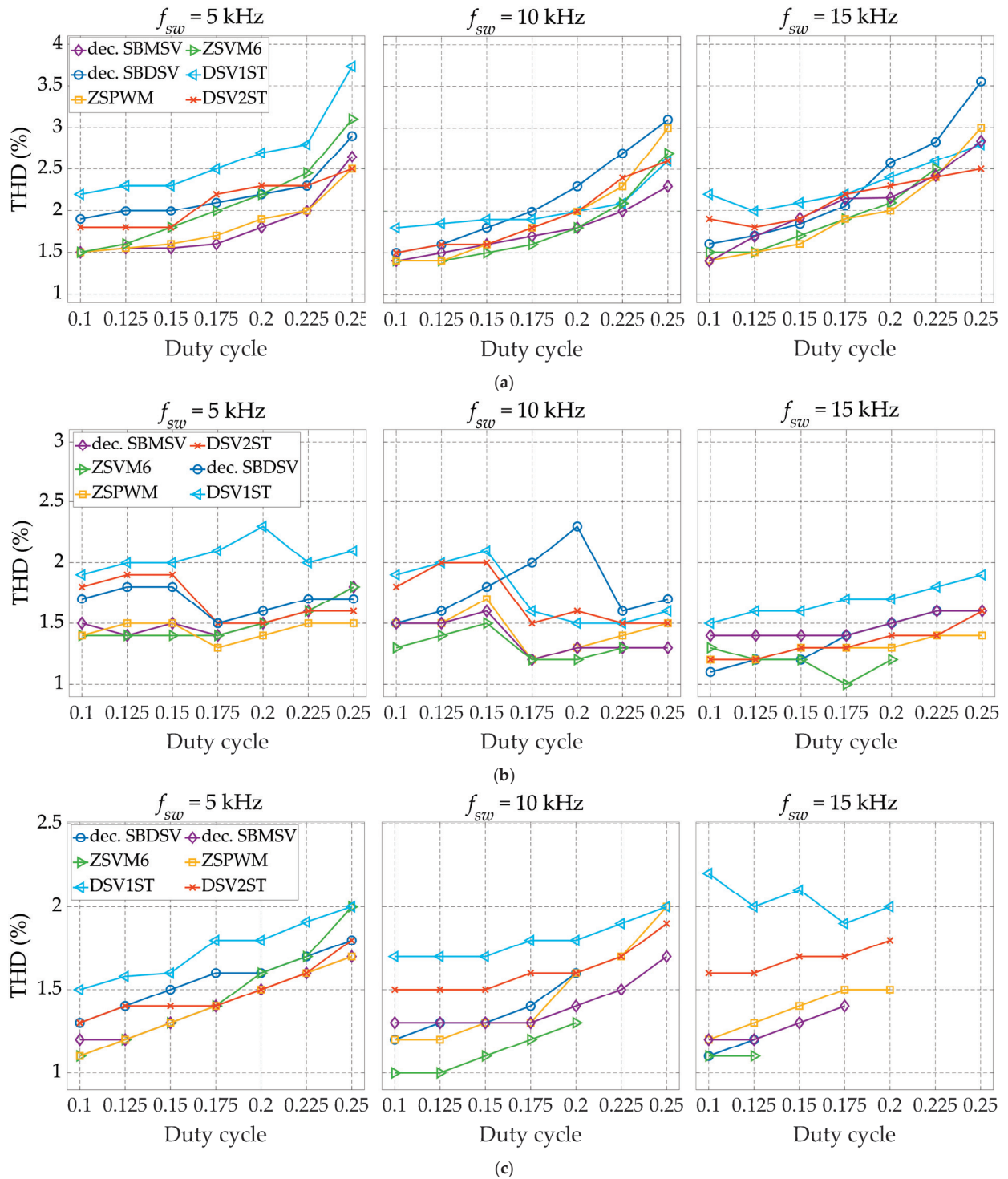


Figure 15. Output current THD as a function of the duty cycle for the different switching frequencies and inverter power of 1000 W (a), 2000 W (b), and 3000 W (c).

Merits (+) and demerits (−) of the proposed PWM schemes are summarized below. Some of them are common to both methods as follows:

- + Higher output voltage for the same M_a value compared to the conventional SPWM scheme.
- + Decoupled control of D_0 and M_a .
- + Reduced number of switch commutations compared to continuous PWM schemes.
- + Lower current stress of the bridge transistors compared to PWM schemes where ST is injected in one leg at a time.
- + No additional reference signals compared to the conventional SPWM or SVPWM schemes.
- − Uneven number of switch commutations of the upper and lower transistors in the bridge.
- − Timer required for D_0 control.

The following features are specific to the DSV1ST scheme:

- + Highest efficiency at higher switching frequencies (otherwise moderate efficiency).
- + Lowest voltage stress for all considered combinations of switching frequency and load power.
- − Highest inductor current ripple, along with the SBMSV scheme, due to a single ST state per switching period.
- − Highest load current THD, particularly at lower switching frequencies and/or heavier loads.

Lastly, the following features are specific to the DSV2ST scheme:

- + Highest efficiency at lower switching frequencies (otherwise moderate efficiency).
- + Second lowest voltage stress for all considered combinations of switching frequency and load power.
- + Similar inductor current ripple as with other PWM schemes having two ST states per switching period (ZSPWM and SBDSV).
- − Moderate load current THD, which gets slightly higher at heavier loads in combination with medium to high switching frequencies.

5.3. Voltage and Current Waveforms

In addition to the previously presented steady-state characteristics, experimental waveforms were recorded for the six considered PWM schemes. These waveforms, presented in Figures 16–18, include the current through L_1 (cyan), the load phase current (green), and the voltages across C_1 (magenta) and C_2 (yellow). The currents were measured by means of the current clamps TT-CC 770 (Testec, Frankfurt, Germany), whereas the voltages were measured by means of the differential probes TT-SI 9101 (Testec, Frankfurt, Germany). All the waveforms were recorded for the following parameters: $f_{sw} = 10$ kHz, $v_{in} = 500$ V, $M_a = 0.71$, $D_0 = 0.225$, and the load power of 1000 W. Note that the time axes of the first three channels are all aligned and positioned at the bottom of the screen, whereas the time axis of the fourth channel is centered vertically.

The decrease of the inductor current ripple with the increase of the number of ST states per switching period is clearly visible. Consequently, the smallest ripple is observed for the ZSVM6 scheme (Figure 16c,d), whereas the largest ripple is observed for the two PWM schemes in Figure 17. At the same time, the ZSVM6 scheme, having the highest number of ST states per period, induces the highest electromagnetic interference (EMI) in the recorded waveforms, whereas on the other side of the spectrum is the DSV1ST scheme with the lowest number of switch commutations and, hence, the lowest induced EMI noise (Figure 17c,d). Both the proposed PWM schemes induce lower levels of EMI noise in comparison with the discontinuous PWM schemes that have the same number of ST states per switching period, which is evident from the lack of HF oscillations in the corresponding zoomed waveforms of the inductor current in Figure 17b,d and Figure 18b,d, respectively.

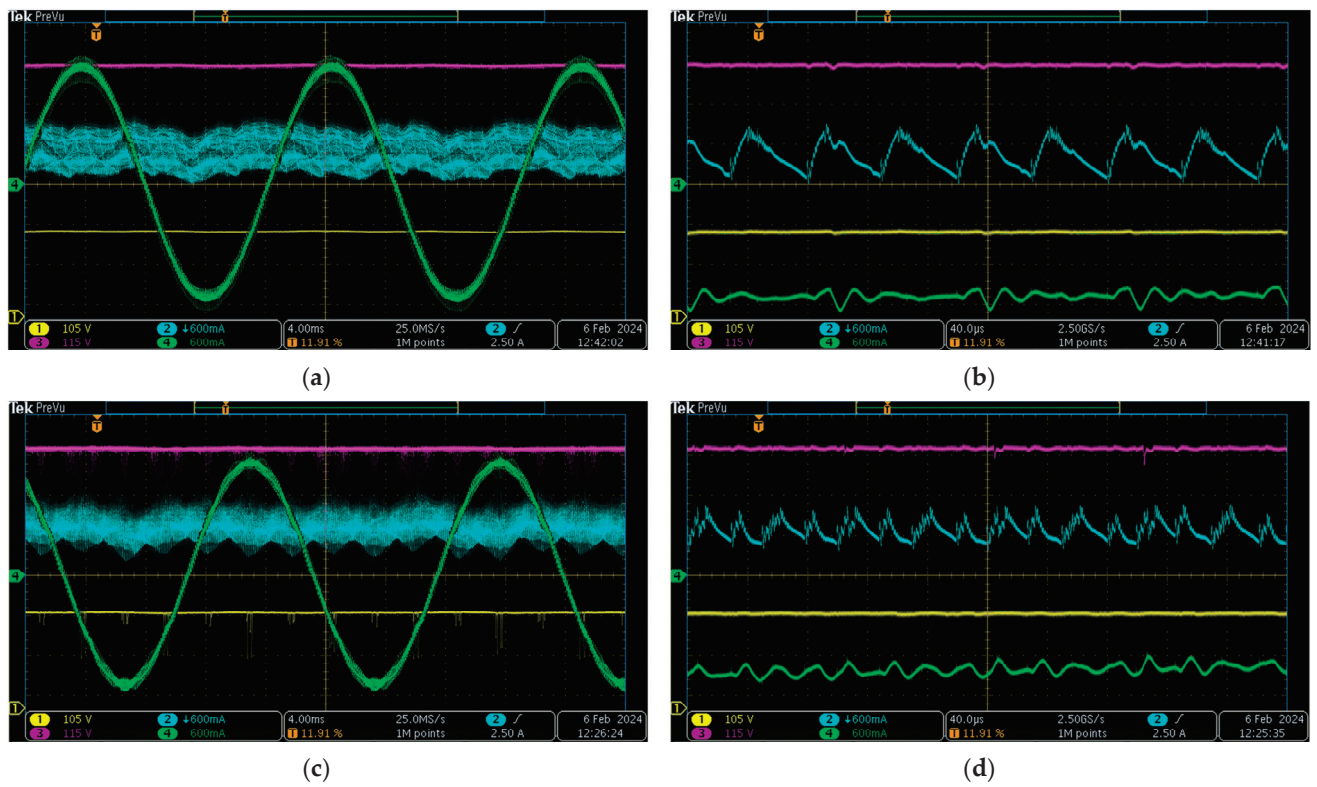


Figure 16. Experimental waveforms of the considered continuous PWM schemes: ZSPWM (a), zoomed ZSPWM (b), ZSVM6 (c), and zoomed ZSVM6 (d).

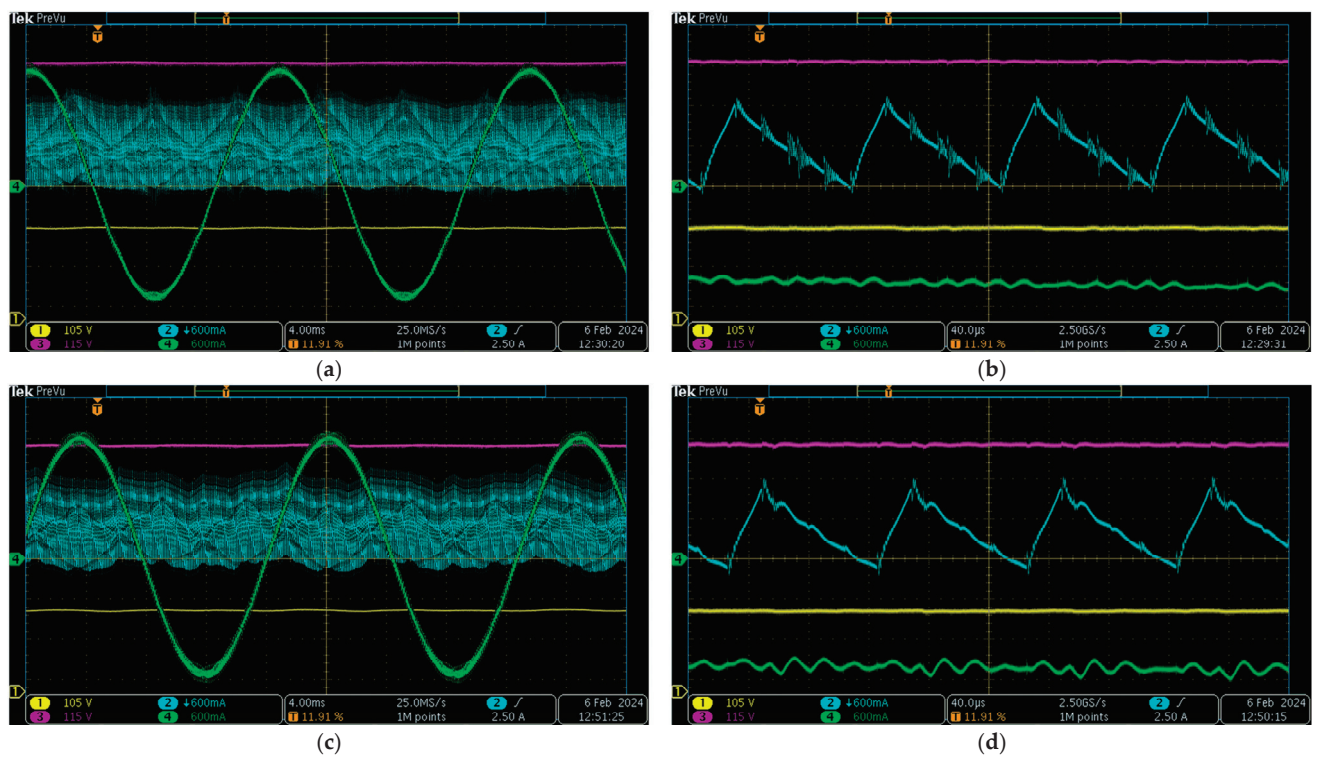


Figure 17. Experimental waveforms of the considered discontinuous PWM schemes with one ST state per switching period: dec. SBMSV (a), zoomed dec. SBMSV (b), DSV1ST (c), and zoomed DSV1ST (d).

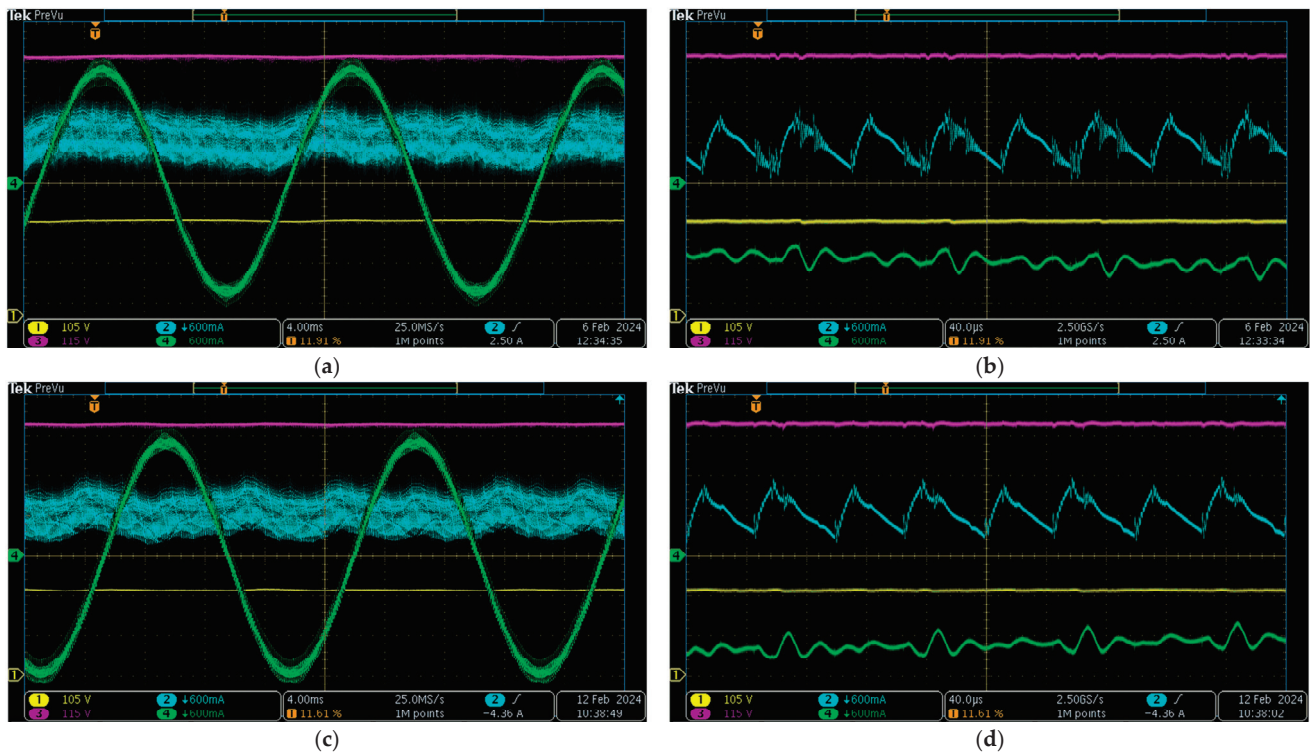


Figure 18. Experimental waveforms of the considered discontinuous PWM schemes with two ST states per switching period: dec. SBDSV (a), zoomed dec. SBDSV (b), DSV2ST (c), and zoomed DSV2ST (d).

5.4. Current Stress

Based on prior theoretical analysis, different considered PWM schemes induce varying levels of current stress on the bridge switches. Firstly, it is observed that the current stress of the upper and lower switches is balanced only in the case of continuous PWM schemes, namely, ZSPWM and ZSVM6. Furthermore, the peak current experienced by the bridge switches tends to be higher in PWM schemes where ST states are injected sequentially, (i.e., phase by phase), namely, ZSVM6 and SBMSV (Table 1). Figure 19 shows the experimentally obtained instantaneous values of the current through L_1 (cyan), the phase current at the bridge output (green), and the current through the upper transistor of the same phase (magenta), which was measured by means of the Rogowski coil probe CWT UM/015/B/1/80 (PEM, San Carlos, CA, USA). The upper transistor is here chosen because the respective maximum current is higher or equal to that of the lower transistor in the same phase leg for all the considered PWM schemes. The depicted time frame spans two switching periods and coincides with the phase current reaching a positive peak. The ST states are distinguished since the inductor current reaches its maximum value at the end of these states. The phase current and the inductor current combined define the maximum current through the upper transistor, as per Table 1, whereas the recorded waveforms confirm that the largest corresponding value is inherent to ZSVM6 and SBMSV schemes (Figures 19b and 19d, respectively). Note that for the ZSPWM scheme (Figure 19a), which is not included in Table 1, the maximum switch current is the same as for other schemes in which ST states are simultaneously injected in all three phase legs.

5.5. Thermal Stress

Thermal stress is known to affect the reliability of semiconductor switches and power converter efficiency. To assess the thermal stress induced by the considered PWM schemes, the case temperature of the upper and lower IGBT-diode pair from the same phase leg was recorded by means of the thermal camera Testo 865 (Testo SE Titisee-Neustadt, Germany).

Figure 20 shows the results obtained for the same operating parameters as those presented in Figures 16–18, where “H” represents the hotspot.

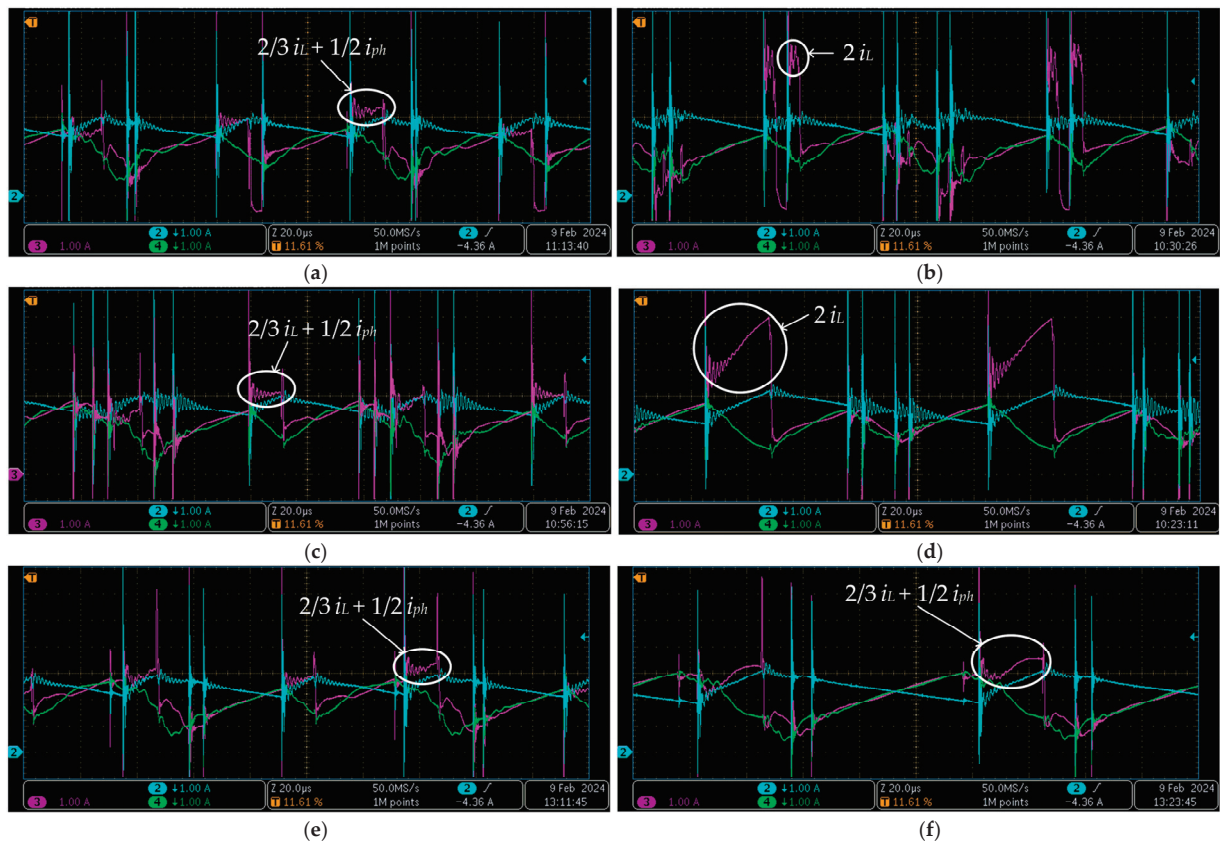


Figure 19. Current stress of the considered PWM schemes: ZSPWM (a), ZSVM6 (b), dec. SBDSV (c), dec. SBMSV (d), DSV2ST (e), and DSV1ST (f).

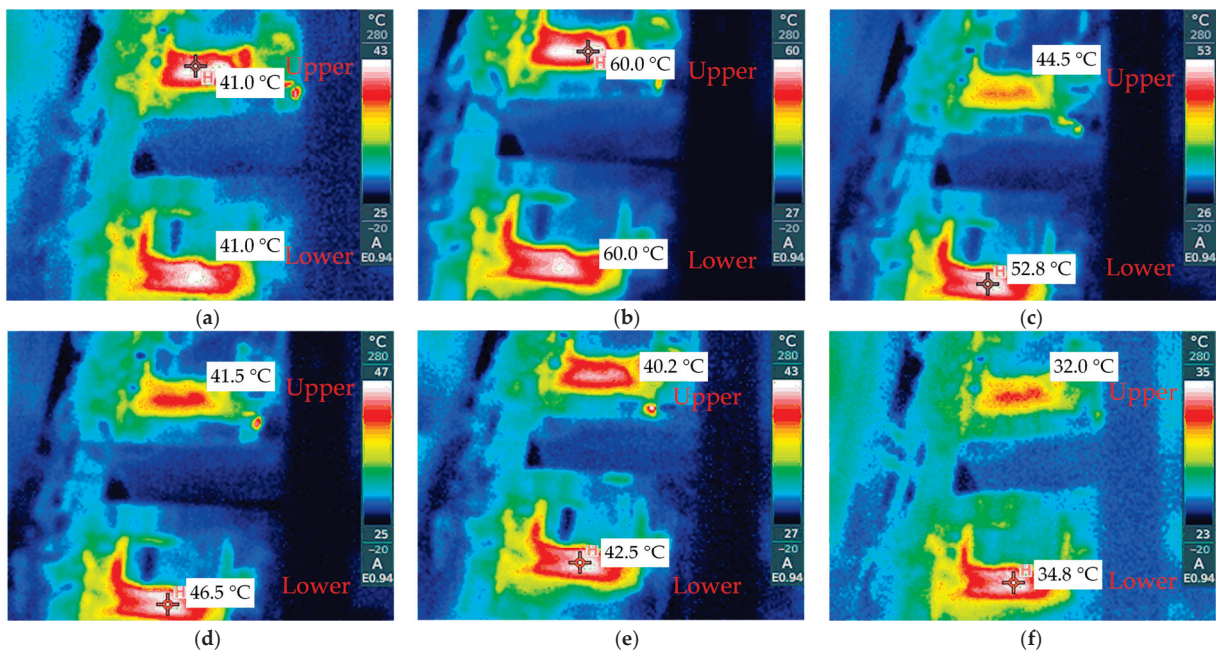


Figure 20. Thermal stress of the considered PWM schemes: ZSPWM (a), ZSVM6 (b), dec. SBDSV (c), dec. SBMSV (d), DSV2ST (e), and DSV1ST (f).

Only the continuous ZSPWM and ZSVM6 schemes result in symmetric thermal stress of the upper and lower switches, whereas the ZSVM6 scheme yields the highest overall temperature. The remaining discontinuous PWM schemes lead to higher thermal stress of the lower IGBT-diode pair due to the uneven number of switch commutations of the upper and lower switches. The lowest thermal stress was observed for the proposed DSV1ST scheme, with almost twice lower temperatures compared to the ZSVM6 scheme.

6. Conclusions

This paper has successfully addressed the development and implementation of advanced PWM schemes for three-phase qZSIs, highlighting their significant potential in enhancing the efficiency and reliability of low-voltage-fed applications. Two innovative SVPWM schemes were introduced that overcame the limitations of existing PWM schemes by providing independent control of the amplitude modulation index and duty cycle, crucial for optimizing performance in closed-loop applications. The experimental testing showed that the proposed decoupled SVPWM schemes not only ensure improved synchronization of the zero switching states and the ST states and prevent unintentional short-circuiting outside of the ST states, but also demonstrates up to 6.8% higher efficiency and up to 7.5% reduced voltage stress. The enhanced efficiency and reduced voltage stress contribute directly to lower operational costs and improved sustainability of systems employing qZSIs. This is particularly valuable in applications where maximizing energy conversion efficiency is of great importance, such as renewable energy systems (e.g., PV panels and wind turbines), electric vehicles (longer ranges and shorter charging times), and uninterruptible power supplies. Moreover, reduced EMI noise means that qZSI systems employing the proposed SVPWM schemes can meet stricter regulatory standards and exhibit less interference with nearby electronic devices, whereas lower thermal stress extends the lifespan of semiconductor components, thereby enhancing system reliability and reducing maintenance costs. The findings of this research underscore the importance of continuous innovation in PWM techniques to meet the evolving demands of power electronics systems. They have the potential to significantly improve the performance and controllability of qZSIs, thereby promoting the widespread adoption of this power converter topology in various applications and facilitating the development of more efficient and reliable power conversion systems.

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Article

A Novel Supercapacitor Model Parameters Identification Method Using Metaheuristic Gradient-Based Optimization Algorithms

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Abstract: This paper addresses the critical role of supercapacitors as energy storage systems with a specific focus on their modeling and identification. The lack of a standardized and efficient method for identifying supercapacitor parameters has a definite effect on widespread adoption of supercapacitors, especially in high-power density applications like electric vehicle regenerative braking. The study focuses on parameterizing the Zubieta model for supercapacitors, which involves identifying seven parameters using a hybrid metaheuristic gradient-based optimization (MGBO) approach. The effectiveness of the MGBO method is compared to the existing particle swarm optimization (PSO) and to the following algorithms proposed and developed in this work: ‘modified MGBO’ (M-MGBO) and two PSO variations—one combining PSO and M-MGBO and the other incorporating a local escaping operator (LCEO) with PSO. Metaheuristic- and gradient-based algorithms are both affected by problems associated with locally optimal results and with issues related to enforcing constraints/boundaries on solution values. This work develops the above-mentioned innovations to the MGBO and PSO algorithms for addressing such issues. Rigorous experimentation considering various types of input excitation provides results indicating that hybrid PSO-MGBO and PSO-LCEO outperform traditional PSO, showing improvements of 51% and 94%, respectively, while remaining comparable to M-MGBO. These hybrid approaches effectively estimate Zubieta model parameters. The findings highlight the potential of hybrid optimization strategies in enhancing precision and effectiveness in supercapacitor model parameterization.

Keywords: supercapacitor; Zubieta model; energy storage; particle swarm optimization; gradient-based optimization; metaheuristic algorithms; local escaping operator; parameter identification

1. Introduction

Recently, there has been a noteworthy global upswing in the adoption of renewable energy sources. This increased interest is a proactive response to economic, political, and social factors, all directed toward diminishing dependence on conventional fossil fuels [1,2]. The efficiency of power systems that encompass a substantial share of renewable energy sources is shaped by various factors. Challenges arise from the inherent unpredictability associated with renewable energy sources. Consequently, the use of energy storage systems plays a pivotal role in effectively bringing renewable energy systems into commercial viability [3].

In recent years, there has been a marked uptick in interest surrounding supercapacitors (SCs). They are being viewed as a viable supplementary power source owing to their outstanding high power density and relatively high energy density [4]. The successful

incorporation of supercapacitors into energy storage systems has been observed across diverse industrial applications, such as electric vehicles and solar energy systems [5,6]. The integration of SCs provides significant advantages in maintaining stability within electrical power systems. This is achieved by augmenting the energy supply derived from batteries and the unpredictable nature of renewable resources.

To optimize the integration of supercapacitors into energy storage systems, it is crucial to establish a precise dynamic representation that effectively represents the static and dynamic properties of supercapacitors. The precise definition and modeling of the system's characteristics is a vital step in enhancing and managing any energy storage system. SC dynamic modeling is used to identify and characterize electrical and thermal performances, condition diagnostics-monitoring and estimation of state of charge (SOC), state of power (SOP), state of health (SOH), and control mechanism design [7–9]. Supercapacitors are modeled based on the characteristics to be monitored, with five key models: equivalent circuit models, electrochemical models, thermal models, fractional order models, and intelligent models [7,10].

Different models have been developed to explain the behavior of supercapacitors; however, the electrical equivalent circuit models are found to be a convenient and common way to simulate the electrical behavior of SCs. To characterize and simulate the electrical behavior of SC, equivalent circuit models utilize parameterized RC networks defined by ODEs. Utilizing electrical models enables the assessment of supercapacitor capacitance, considering its variations with bias voltage, voltage drop, as well as power loss attributed to internal resistance, self-discharge, and leakage current effects. Additionally, these models account for the electric dynamic behavior influenced by ion diffusion [11,12].

Modeling and characterization of supercapacitors have been extensively investigated using electrochemical impedance spectroscopy (EIS) approaches [13–16] or time response experiments and simulations [17–20]. Both techniques aim to provide a variety of equivalent circuit representations that are required for characterizing the state of the SC cell in the course of operation. Hence, when analyzing the time domain system, various circuit models have been proposed. Classical models, multi-stage ladder models, and dynamic models are the three categories of equivalent circuit models. In addition, a wide range of nonlinear models have been formulated for supercapacitors, which include fractional order models, as well as methodologies for determining parameters, thereby demonstrating the effectiveness of these models [21,22]. Nevertheless, the procedure of determining parameters for these models sometimes entails multiple stages of linearization and extensive experimental investigation.

Methods such as adaptive filter algorithms, metaheuristic optimization methods, and intelligent artificial intelligence are being deployed in the literature to find the optimum parameters for the different SC equivalent models. The recursive least squares (RLS) method was used to identify the parameters of the classical equivalent circuit model in [23–25]. In [25], the authors used a voltage-dependent capacitor in the first branch instead of the constant capacitance capacitor, making it simpler to identify the time-variant parameters. Similarly, in [26], a nonlinear least square method was used to identify the parameters of a supercapacitor model. The authors provided a framework for the branch number selection criteria and order reduction. The total least-squares method was used in [27] to identify the three parameters (C_0 , K_V , and C_2) of the reduced two-branch Zubieta model [28]. The first branch resistor, R_0 , was assumed to be given by the manufacturer, and the second branch resistor, R_2 , and the leakage resistor, R_{lea} , were found using the circuit analysis method. The voltage response, input current, and their second- and first-order derivatives were used to define the estimation problem. The derivatives of the electric variables were obtained using filter differentiation.

Moreover, Kalman filters were used to identify supercapacitor dynamic model parameters online [29]. The authors used the extended Kalman filter to automate the estimation of the error bound. The researchers proved that the proposed estimator model can be used to accurately depict the voltage behavior of the SC under various scenarios.

Optimization is still a challenging computational task. Consequently, numerous algorithms have been proposed to address this challenge. Two questions must be answered to ensure the best possible solution to this problem: how to identify global and local optimization and how to preserve such optimization until the end of the search. Swarm intelligence, which includes particle swarm optimization (PSO) [30], the grey wolf optimizer (GWO) [31], whale optimizer [32], ant-lion optimizer (ALO) [33], genetic algorithm (GA) [34], artificial ecosystem-based optimizer (AEO) [35], and many others, evolved as a result of these questions over the last two decades. These metaheuristic algorithms are widely used to solve and optimize various real-world problems. However, few researchers in the field of energy storage have used metaheuristic optimization algorithms as an appealing tool for identifying electrical circuit model parameters.

In 2022, a study by researchers [36] employed a straightforward and reliable methodology, using the bald eagle search (BES) optimization algorithm to identify the parameters of the Zubieta model's supercapacitor equivalent circuit [20]. The envisaged bald eagle search algorithm mimics bald eagle hunting behavior to demonstrate the consequences of each stage of hunting. The BES approach's robustness was assessed in comparison to other metaheuristic algorithms for two supercapacitors (SCs) with modules of 470 F and 1500 F. The authors attained an MSSE of 1.32×10^{-8} for the 1500 F SC when employing the BES. In contrast, the PSO, the most closely related optimizer based on the obtained results, yielded a higher MSSE of 6.69×10^{-6} .

In [37], a similar approach was used with the interior search optimizer (ISA), a notable metaheuristic optimization algorithm proposed by Gandomi [38]. The mean square error (MSE) for parameterization of a 470 F capacitor using the ISA was 0.004487% compared to 0.0310895 for GWO, 0.0045% for GA, and 0.03109% for WA. These findings suggest that the ISA method can be used to optimize the parameters of the Zubieta model, and they are comparable to the genetic algorithm approach.

In [39], a real-time modeling approach based on the weighting bat algorithm (WBA) was proposed. The model was used for parameter identification in the reduced Zubieta model, enabling real-time power management in embedded systems powered by supercapacitors, as well as predicting supercapacitor behavior. The WBA convergence of the fitness function in 50 iterations is comparable to the genetic algorithm with lower computational time, 0.4328% and 0.43815%, respectively.

Furthermore, the study in [40] introduced an electrical model that is based on the optimization of parameters for a passive electrical circuit model. There exist multiple methodologies for the parameter identification of supercapacitor models. These include the analytical method [20], Segmentation Optimization [41], the binary quadratic equation fitting method [42], Universal Adaptive Stabilization and Optimization [11], Particle Swarm Algorithm (PSO) [43], the recursive least-squares method [44], and other alternative techniques.

Over the past few years, there has been a notable influx of research focusing on the development and application of advanced optimizers for parameter identification, particularly in the field of energy storage systems. Metaheuristic algorithms have continued to dominate the landscape of parameter estimation, offering innovative solutions to address challenges associated with accuracy and convergence rates.

One prominent optimizer in recent literature is the Sine Cosine Algorithm (SCA), introduced by Mirjalili et al. [45]. SCA has shown promise in optimizing complex and nonlinear functions, making it suitable for parameter identification tasks. Studies have applied SCA to various energy storage system models, demonstrating its effectiveness in achieving accurate parameter estimates within a reduced computational time.

Another development, as mentioned earlier, is the grey wolf optimizer (GWO), proposed by Mirjalili et al. [31]. GWO draws inspiration from the social hierarchy of grey wolves and has been successfully applied to parameter identification tasks in renewable energy systems. Its ability to strike a balance between exploitation and exploration makes it well-suited for handling complex and high-dimensional optimization problems.

Other noteworthy contributions are the Dandelion Optimization Algorithm [46] used for the parameter estimation of proton exchange membrane fuel cells (PEMFCs) and the Metaheuristic Mountain Gazelle Optimizer for parameter estimation of single- and double-diode photovoltaic cell models [47]. These new optimization strategies demonstrated improved performance compared with traditional algorithms, showcasing the potential of metaheuristic optimization paradigms to address the complexities associated with renewable energy system models.

Nevertheless, the existing literature points out limitations in the application of metaheuristic optimizers to identify variables in the SCs equivalent electrical circuit model. Additionally, the currently employed algorithms have their own constraints and highlight notable gaps that warrant further investigation. Existing studies often lack a standardized approach, leading to inconsistencies in parameter estimation. Challenges include the absence of a universally accepted model and the need for improved convergence rates and accuracy in optimization algorithms. Moreover, there is a limited exploration of hybrid optimization strategies that could potentially enhance the efficiency of parameter identification for supercapacitors. Addressing these gaps is crucial to advancing the reliability and performance of supercapacitors in energy storage systems, promoting a more standardized and effective methodology for parameter identification. Consequently, to build an accurate model for SCs, it is vital to use an efficient optimization technique that effectively addresses the limitations of current optimizers.

In this study, the authors propose a more straightforward and reliable approach by employing the metaheuristic gradient-based optimization (MGBO) algorithm [48,49] to ascertain the parameters of the SCs electrical circuit model, specifically the Zubieta model. Having a precise model is crucial for accurately characterizing the behavior of SCs, thereby facilitating further research in the field. Moreover, to enhance the effectiveness of PSO and address the underlying constraints, numerous variations of PSO have been developed. Multiple implementations of PSO algorithm variants have been proposed in the literature. These variants include adaptive inertia weight PSO, constriction PSO, and hybrid PSO [50,51]. These changes encompass the incorporation of additional strategies, such as the dynamic manipulation of the inertia weight, the self-adjustment of acceleration constants, and integrating it with other optimization techniques [52].

Hence, this research develops a modified metaheuristic gradient-based optimization (M-MGBO) algorithm and two versions of the particle swarm optimization (PSO) algorithm. The motivation for doing this is as follows:

Gradient-based algorithms such as the MGBO can suffer from problems related to computing/approximating the gradient. Particle swarm optimization (PSO)/metaheuristic approaches have been used to circumvent issues related to the gradient. Both MGBO and PSO approaches may be affected due to solutions being trapped in local optima. The literature has several randomized approaches to avoid such issues. However, such randomized approaches often disregard/do not explicitly incorporate checking for parameter bounds and also do not necessarily check if a randomized candidate solution actually helped achieve the objective of not being trapped in a local optimum. Furthermore, it is possible that due to the nature of the optimization problem or the algorithm or both, such issues related to entrapment in local optima, and parameter-bound constraint violation happen more frequently during a certain stage (early iterations/late iterations) of the iterative optimization cycle. Moreover, encoding such information in the optimization algorithm itself may help achieve better results.

In addition to improving parameter estimation accuracy, which requires a reduced number of algorithm iterations, alleviating the above-mentioned issues is the objective of this research. To alleviate the above issues, this work develops the M-MGBO algorithm, which incorporates an appropriate local escaping operator (LCEO) algorithm, the details of which are available in Section 3.3.1. Further, results exist in the literature showing the benefits of generating hybrid versions of the PSO algorithm [53]. Therefore, this work combines the MBGO algorithm with the PSO and also further combines the advanced

LCEO algorithm mentioned earlier with the PSO algorithm, using them for supercapacitor parameter estimation. The relevant details are available in Sections 3.3.2 and 3.3.3.

2. Supercapacitor (SC) Model and Experimental Setup

2.1. SC Model

Various models have been devised to describe the behavior of supercapacitors. Nevertheless, electrical equivalent circuit models have emerged as a practical and prevalent method for simulating the electrical characteristics of supercapacitors (SCs). In characterizing and simulating the electrical behavior of an SC, equivalent circuit models use parameterized RC networks. These electrical models serve to determine the SC capacitance and its correlation with voltage drop, bias voltage, self-discharge, power loss due to internal resistance, leakage current effects, and the electric dynamic behavior induced by ion diffusion [1,2].

Furthermore, the SC model can be conceptualized as an electrochemical double-layer capacitor (DLC) [3]. An exemplary model proposed by [4] is the Zubieta model, which, through a comprehensive consideration of the device's behaviors under various operating conditions and states, has demonstrated an accurate reflection of the physics of the SC. The Zubieta model comprises three parallel RC branches with a leakage resistance (EPR), as illustrated in Figure 1.

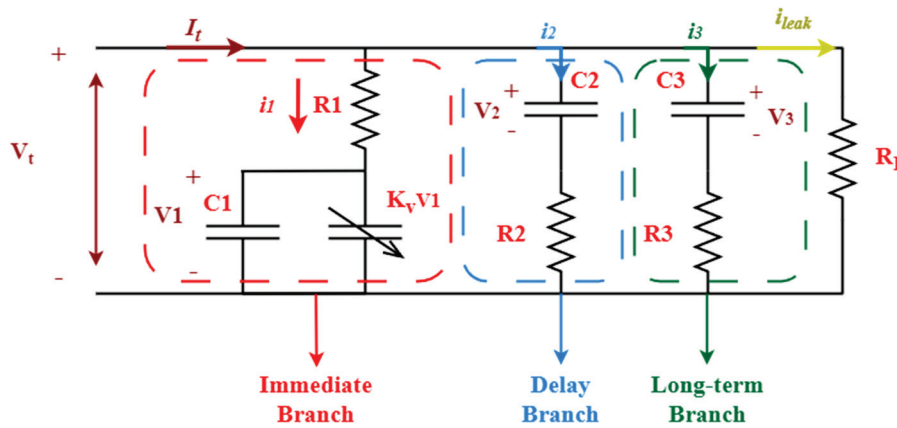


Figure 1. Supercapacitor Zubieta model.

Each branch is characterized by a unique time constant indicative of SC behavior across a specific frequency range. The Zubieta model includes three RC branches, namely the immediate branch, the delayed branch, and the long-term branch. The immediate (first) branch is characterized by a voltage-dependent differential capacitance with a constant capacitor (C_1) and a parallel voltage-dependent capacitor ($K_v \times V_1$). This first branch significantly influences the supercapacitor's behavior during the initial transient time of operation. The second and third branches (delayed branch and long-term branch) represent SC behavior over a longer period of time (minutes and hours, respectively). The equivalent series resistance (EPR) is denoted by R_L .

The incorporation of three RC branches allows the model to capture SC characteristics on distinct timescales, with the immediate branch providing a nonlinear response emulating the effects of the internal porous double-layer electrode structure of the supercapacitor. The model parameters are determined using voltage–current measurements, ensuring an accurate representation of SC behavior. This model is a valuable tool for optimizing SC performance and improving energy storage systems.

Referring to Figure 1, the total capacitance of the first branch can be calculated as follows:

$$C_0 = C_1 + \frac{K_v}{2} \times V_1(t), \quad (1)$$

where C_1 represents the electrostatic capacitance of the capacitor and K_v represents the effects of the diffused layer of the SC, and it is the slope of the capacitance as a function of voltage in F/V .

The dynamics of the immediate branch can be found as follows

$$\frac{dV_1}{dt} = \frac{V_t(t) - V_1(t)}{\left(C_1 + \frac{K_v}{2} \cdot V_1(t)\right) \cdot R_1} \quad (2)$$

Similarly, the dynamic of the second and third branches (delayed branch and long-term branch) are given by the following:

$$\frac{dV_2}{dt} = \frac{V_t(t) - V_2(t)}{C_2 \cdot R_2} \quad (3)$$

$$\frac{dV_3}{dt} = \frac{V_t(t) - V_3(t)}{C_3 \cdot R_3} \quad (4)$$

The current resulting from self-discharge can be represented using the following equation:

$$i_{leak} = \frac{V_t}{R_L} \quad (5)$$

Hence, the current input to the supercapacitor can be expressed as the summation of all currents passing through all four branches in the Zubieta model representation, $I_t = i_1 + i_2 + i_3 + i_{leak}$. Computing the currents in terms of the voltages, further allows the development of a mathematical relation that expresses the SC's terminal voltage concerning all branches as follows:

$$V_t \times \left(1 + \frac{R_1}{R_2} + \frac{R_1}{R_3} + \frac{R_1}{R_L}\right) = V_1 + \left(\frac{R_1}{R_2}\right)V_2 + \left(\frac{R_1}{R_3}\right)V_3 + I_t R_1 \quad (6)$$

By rearranging (6), the following relation is obtained for the SC terminal voltage:

$$V_t = \left(\frac{R_2 R_3 R_L}{R_2 R_3 R_L + R_2 R_1 R_L + R_1 R_3 R_L + R_1 R_2 R_3}\right) \left(V_1 + \left(\frac{R_1}{R_2}\right)V_2 + \left(\frac{R_1}{R_3}\right)V_3 + I_t R_1\right) \quad (7)$$

Hence, employing the derived dynamics for the three branches and having knowledge of the input/total current, the Simulink representation depicted in Figure 2 is employed. This approach enables the calculation of the estimated terminal voltage once the electrical parameters of the supercapacitor (SC) are identified. Subsequently, the computed terminal voltage is cross-referenced with the measured terminal voltage, acting as a validation step to verify the accuracy of the obtained model.

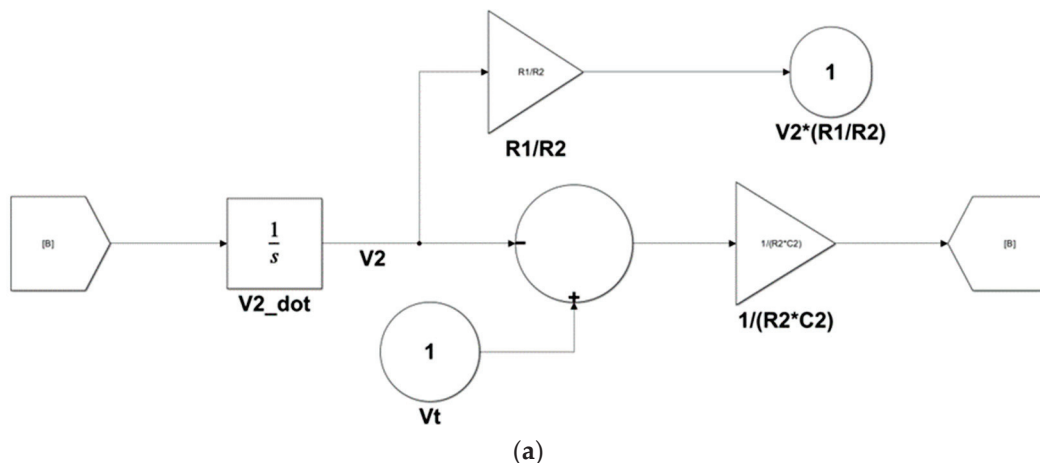


Figure 2. Cont.

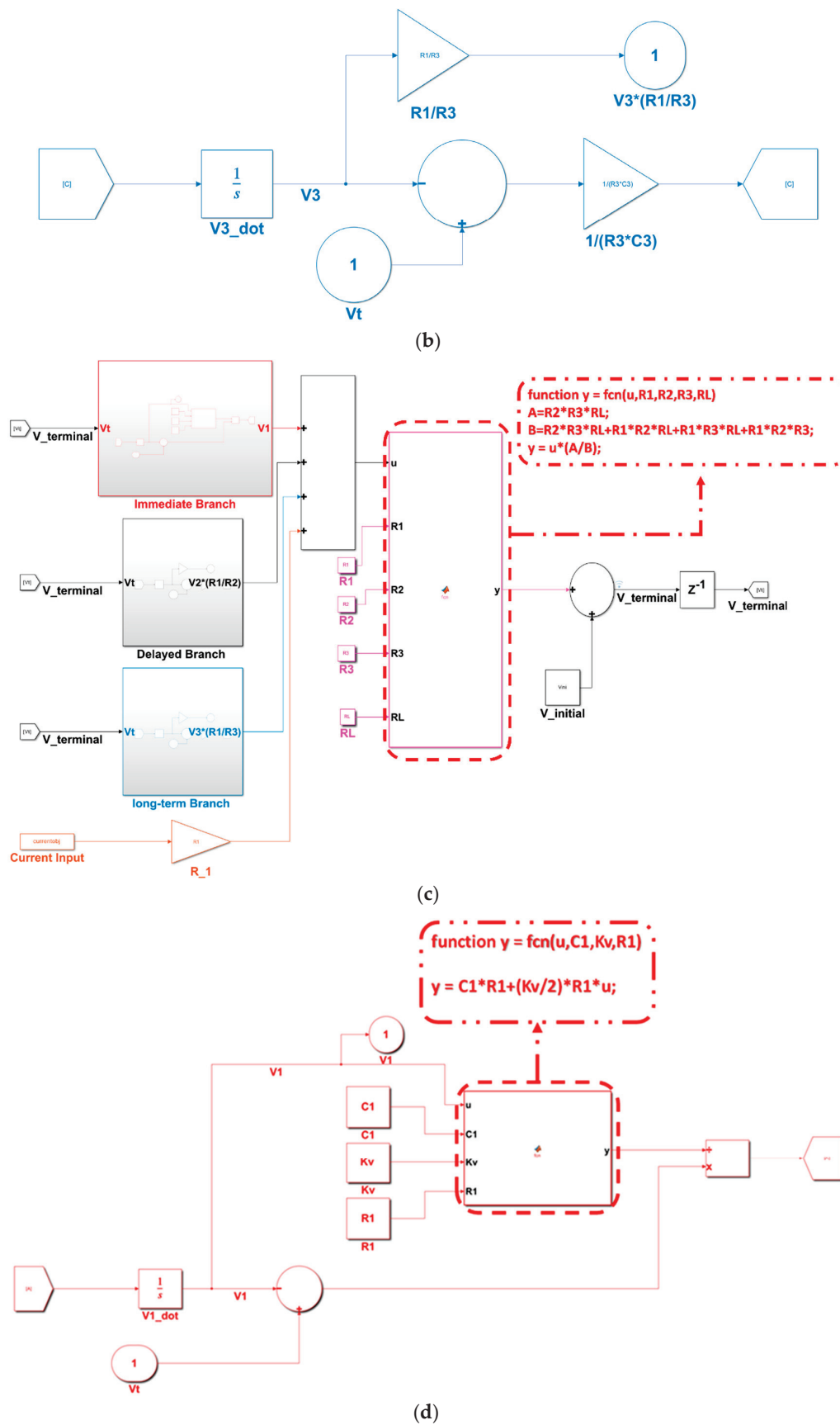


Figure 2. Supercapacitor Zubieta model Simulink representation: (a) long-term branch; (b) delayed branch; (c) main model; (d) immediate branch.

2.2. Experimental Setup

The experimental setup configuration comprises a supercapacitor bank, a controlled current source, current and voltage sensors, and a dSPACE controller, as illustrated in Figure 3. The controlled current source was modulated using an analog signal through dSPACE. A voltage signal ranging from 0 V to 5 V governed the remote control of the power supply. The interface of the power supply with the dSPACE controller is depicted in Figure 3. The dSPACE controller monitored the terminal voltage of the supercapacitor bank, and the charging current was measured using the LA25-NP current transducer. To ensure that the analog signal read by the dSPACE controller remains below 10 V (1 after scaling), a voltage divider was employed.

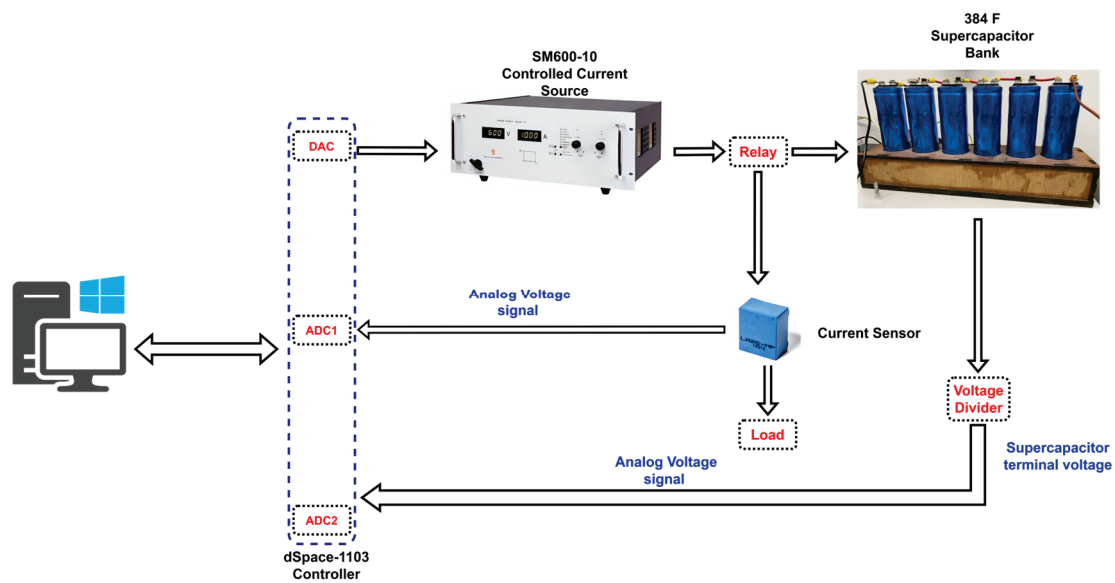


Figure 3. Experimental setup block diagram.

The current source was programmed to generate three distinct charging profiles: Profile 1: a 10 A step input; Profile 2: a 4 A step input; and Profile 3: a variable frequency input, as illustrated in Figure 4, respectively. For each of these input scenarios, the supercapacitor bank undergoes charging until it reaches its rated voltage of 15 V. At this point, the relay between the current source and the supercapacitor bank is deactivated, allowing the SC bank to settle for a duration of 30 min. Throughout both the charging and settling processes, the terminal voltage of the SC bank is continuously monitored by the dSPACE controller.

Following the 30 min settling period, the SC bank was discharged to a load. In this analytical study, particular emphasis was placed on the voltage response of the SC bank during the charging process as a key element in the optimization problem. The model generated was designed to emulate the behavior of the supercapacitor during the charging process.

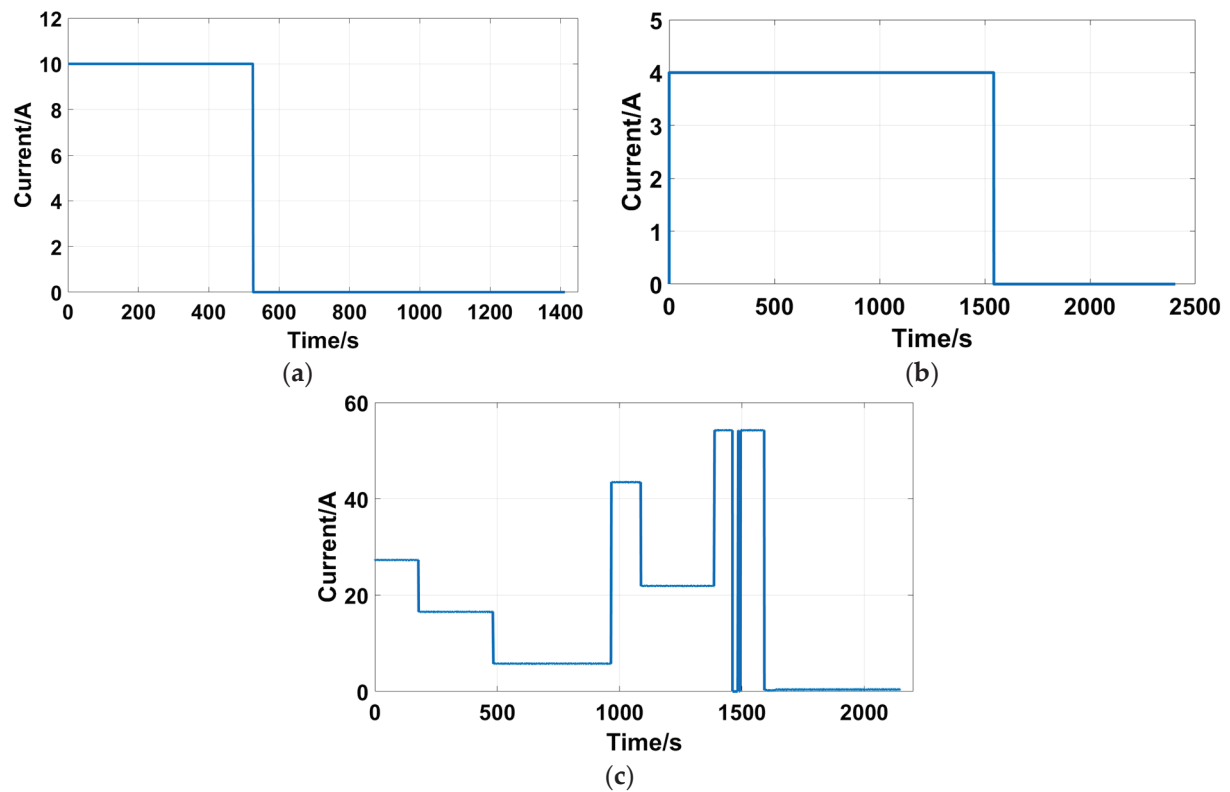


Figure 4. Experimental current profiles: (a) Profile 1: 10 A step input; (b) Profile 2: 4 A step input; (c) Profile 3: Variable frequency input.

2.3. SC Model Validation

The determination of equivalent circuit model parameters is a crucial aspect of the optimization of SC operation, and various approaches have been proposed in the literature for this purpose. One such approach is the circuit analysis/analytical method, which establishes relationships between different circuit components, allowing for the straightforward identification of the equivalent circuit model parameters. This method has the advantage of having a clear physical interpretation and not requiring extensive experimental equipment [5].

In the case of the three-branch Zubieta model, a procedure was proposed by Zubieta et al. [4] to determine the parameters of the SC model. This method involves charging a fully discharged SC with a high constant current. The immediate branch parameters are identified by measuring the voltage response at a short time constant, as it is assumed that this branch has a smaller time constant compared with the other two branches. This results in all of the charges being initially stored in the immediate branch. Once the capacitor reaches its rated voltage, the charging current is set to zero, and the parameters of the delayed branch and long-term branch are then estimated using the voltage response at the appropriate time constants of each branch.

The parameterization procedure for the Zubieta model is summarized in Table 1. By employing this method, it is possible to reliably determine the parameters of each branch of the equivalent circuit model using the SC voltage response at the appropriate time constants.

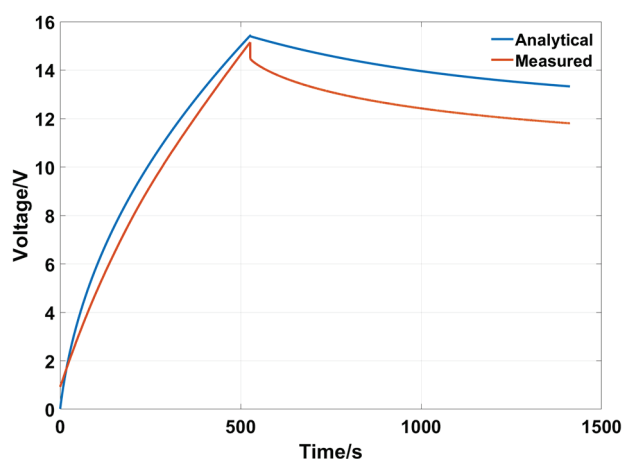
Table 1. Zubieta model analytical parameterization procedure.

Parameter	Relation	Comments
R_1	$R_1 = \frac{V_1}{i_{sc}}$	The SC is assumed to start charging at zero initial conditions. V_1 is the voltage at t_1 where $20 \text{ ms} < t_1 < 50 \text{ ms}$ Where t_2 is the time when $\Delta V = V_2 - V_1 = 50 \text{ mV}$ Where the amount of charge stored $Q_t = i_{sc}(t_4 - t_1)$, and $t_4 = t_3 + 20 \text{ ms}$. Where t_3 is when $V_{sc} = V_{rated}$
C_1	$C_1 = \frac{i_{sc}(t_2 - t_1)}{\Delta V}$	
K_v	$K_v = \frac{2}{V_4} \left(\frac{Q_t}{V_4} - C_0 \right)$	
R_2	$R_d = \frac{(V_4 - \frac{\Delta V_2}{2})(t_5 - t_4)}{(C_0 + k(V_4 - \frac{\Delta V_2}{2}))\Delta V_2}$	Where t_5 is at $V_5 = V_4 - \Delta V_2$. Where $\Delta V_2 = 50 \text{ mV}$
C_2	$C_2 = \frac{Q_t}{V_6} - \left(C_0 + \frac{kV_6}{2} \right)$	V_6 is the voltage at $t_6 = t_5 + 300 \text{ s}$
R_3	$R_3 = \frac{(V_7 - \frac{\Delta V_2}{2})(t_7 - t_6)}{(C_0 + k(V_6 - \frac{\Delta V_2}{2}))\Delta V_2}$	Where t_7 is at $V_7 = V_6 - \Delta V_2$. Where $\Delta V_2 = 50 \text{ mV}$
C_3	$C_3 = \frac{Q_t}{V_7} - \left(C_0 + \frac{kV_7}{2} \right) - C_2$	Where V_8 is at $t_8 = 30 \text{ min}$

Accordingly, the procedure in Table 1 is used to find the parameters of the Zubieta model for a 384 F SC bank rated at 15 V using Profile 1 (Figure 4a). The identified parameters are presented in Table 2, and the corresponding voltage response is graphically represented in Figure 5.

Table 2. Identified SC parameters using Zubieta model analytical parameterization procedure.

Parameter	Value
$K_v \left(\frac{F}{V} \right)$	53.4472
$R_1 (\Omega)$	0.0041
$C_1 (F)$	0.8516
$R_2 (\Omega)$	0.0024
$C_2 (F)$	69.7146
$R_3 (\Omega)$	4.9780
$C_3 (F)$	126.8252
T_1	0.00349156
T_2	0.167315
T_3	631.4593

**Figure 5.** Experimental voltage response vs. the analytical response using current Profile 1.

Despite its simplicity, the analytical method is not suitable for real-life applications, as it does not provide accurate parameter estimates for the Zubieta model with inputs other than a step input. However, the parameters obtained using this method can serve as an initial estimate for parameterization. A study by researchers in [6] compared the Zubieta model's behavior to that of a real supercapacitor. The research found that there is a discrepancy between the simulated and actual behavior due to the model's inability

to account for self-discharge phenomena. Additionally, the quality of data and charging current amplitude significantly affect the parameterization process.

3. Proposed Identification Methods

This section introduces our proposed procedure for identifying the SC Zubieta model parameters. Each phase of this process is explained in the subsequent subsections.

3.1. Metaheuristic Gradient-Based Optimization (MGBO)

The metaheuristic gradient-based optimization (MGBO) algorithm was first introduced in [7]. This method was further applied for the parameter identification of single and two-diode models of solar cells [8]. The MGBO method incorporates both gradient and population-based techniques to determine the search direction via Newton's method (Given by (8)). MGBO uses a set of vectors and two key operators, the gradient search rule and local escaping operators, to traverse the search domain and minimize the objective function in optimization problems.

$$x_{n+1} = x_n - \frac{f(x_n)}{f'(x_n)} = x_n - \frac{2\Delta x \times f(x_n)}{f(x_n + \Delta x) - f(x_n - \Delta x)} \quad (8)$$

The MGBO algorithm applied a variant of the Newton's method in its formation proposed by [9]. The methodology followed by this variant improves the iterative process of the method by approximating the indefinite integral by a trapezoid instead of a rectangle. This reduces the error in the approximation and uses the average vector of z_{n+1} and x_n , instead of using x_n only. The enhanced representation is given by the following:

$$x_{n+1} = x_n - \frac{f(x_n)}{f'\left(\frac{[z_{n+1} + x_n]}{2}\right)} \text{ where,} \quad (9)$$

$$z_{n+1} = x_n - \frac{f(x_n)}{(f')} \quad (10)$$

3.1.1. Gradient Search Rule (GSR) and Algorithm Formulation

The formulation of the gradient search rule (GSR) is based on an approach that incorporates the use of Newton's gradient. The use of this methodology entails regulating the exploration of the solution vector within the search space with the aim of augmenting the rate at which the metaheuristic GBO achieves convergence. The GSR, using (8), is formulated as follows:

$$\text{GSR} = 2\Delta x \cdot \mu_1 \cdot r_n \frac{X_i^{it}}{x_{worse} - x_{best} + \varepsilon}, \quad (11)$$

where in (11), r_n is a normal distributed random number, X_i^{it} is the current approximation (x_n), Δx is the derivative incrementation value, and x_{worse} and x_{best} are the worst and best positions neighboring X_i^{it} in the optimization process. The search capabilities of the MGBO algorithm are enhanced by adding a random parameter μ_1 to the GSR, where it is given by the following relation:

$$\mu_1 = 2 \cdot r \cdot \alpha - \alpha, \quad (12)$$

where r is a random number between 0 and 1 and

$$\alpha = \left| \beta \sin\left(\frac{3\pi}{2} + \sin\left(\frac{\beta 3\pi}{2}\right)\right) \right|, \quad (13)$$

$$\beta = \beta_{min} + (\beta_{max} - \beta_{min}) \left(1 - \left(\frac{m}{M}\right)^3\right)^2, \quad (14)$$

where β_{max} and β_{min} are 0.2 and 1.2, respectively, m is the number of the current iteration, and M is the total number of iterations. The optimization process involves searching for the optimal solution in the given search space. The parameter α is introduced to enhance the algorithm's balance between local and global search.

In contrast, Δx is defined as the difference between the best position and a randomly selected position from the population of the current iteration.

$$\Delta x = rand(1 : N) \times |S|, \quad (15)$$

$$S = \frac{(x_{best} - x_{r1}^m) + \gamma}{2}, \quad (16)$$

$$\gamma = 2 \cdot r \left(\left| \frac{\sum_{n=1}^4 x_{r_n}^m}{4} - x_n^m \right| \right), \quad (17)$$

where r_a and r_b are random variables in the range $[0, 1]$.

3.1.2. Direction of Movement

The direction of movement (DM) in (18) is introduced into the algorithm to enhance the exploitation of the area near the solution X_i^{it} , moving the current solution vector in the direction of $(x_{best} - X_i^{it})$.

$$DM = rand \times \mu_2 (x_{best} - X_i^{it}) \quad (18)$$

Hence, the GSR and DM are used to update the current position vector (X_i^{it}) and the following relation is derived:

$$X1_n^m = X_i^{it} - rand \cdot \mu_1 \left(\left(\frac{2\Delta x \cdot X_i^{it}}{(yp_n^m - yq_n^m + \varepsilon)} \right) \right) + rand \cdot \mu_2 (x_{best} - X_i^{it}) \quad (19)$$

Moreover, by replacing the position vector of the best solution with the current solution vector (X_i^{it}) , a newly generated vector can be generated as follows:

$$X2_n^m = x_{best} - rand \cdot \mu_1 \left(\left(\frac{2\Delta x \cdot X_i^{it}}{(yp_n^m - yq_n^m + \varepsilon)} \right) \right) + rand \cdot \mu_2 (x_{r1}^m - x_{r2}^m), \quad (20)$$

where, in the given context, the variables yp_n^m and yq_n^m are two positions centered/averaged with respect to z_{n+1} and X_i^{it} and are given by the following:

$$yp_n^m = rand \cdot \left(\frac{(z_{n+1} + X_i^{it})}{2} + rand \cdot \Delta x \right) \quad (21)$$

$$yq_n^m = rand \cdot \left(\frac{(z_{n+1} + X_i^{it})}{2} - rand \cdot \Delta x \right) \quad (22)$$

Moreover, DM for $X2_n^m$ is given as follows:

$$DM_{X2} = rand \cdot \mu_2 (x_{r1}^m - x_{r2}^m) \quad (23)$$

Hence, the GSR can be expressed as follows:

$$GSR_{X1} = GSR_{X2} = rand \times \frac{\mu_1 (2\Delta x \times X_i^{it})}{(y_p - y_q + \varepsilon)} \quad (24)$$

The solution $X2_n^m$ is defined as a local search direction, where although it is good for local search, it is limited to global search. Moreover, the solution $X1_n^m$ is defined for global search (exploration), but it is limited to local search. Hence, the MGBO algorithm is developed to take advantage of both solutions ($X1_n^m$ and $X2_n^m$), as well as take advantage of both exploitation and exploration. The combination of both solutions results in the following updated solution (X_i^{it+1}) , derived from the current solution X_i^{it} :

$$X_i^{it+1} = r_a(r_b X1_n^m + (1 - r_b)X2_n^m) + (1 - r_a)X3_n^m, \quad (25)$$

where

$$X3_n^m = X_i^{it} - \mu_1(X2_n^m - X1_n^m) \quad (26)$$

and r_a and r_b are two random numbers between 0 and 1.

3.1.3. Local Escaping Operator LCEO

The MGBO algorithm can be susceptible to getting stuck in local optima, which can limit its ability to find the global optimum. To address this issue, the local escaping operator (LCEO) has been proposed as a way to significantly change the position of the solution x_n , leading to improved search capability. The LCEO operator generates a new solution, X_{LCEO} , by combining several solutions, including the best position (x_{best}), the solutions $X1_n^m$ and $X2_n^m$, two randomly selected solutions from the population, x_{r1} and x_{r2} , and a new randomly generated solution, X_z . This approach effectively enables the MGBO algorithm to break away from local optima, exploring new regions within the solution space, thereby enhancing the likelihood of discovering a global optimum. Hence, the LCEO is defined via the following scheme (Algorithm 1) [8]:

Algorithm 1 Local ESCAPING OPERATOR (LCEO)

Require: $\mu_1, \mu_2, X_i^{it}, x_{best}, x_{best2}, x_{best3}, x_{r1}, x_{r2}, nP, MaxIT$ and LC

```

1   $LC = 4 \times LC \times (1 - LC)$ 
2  if  $rand < LC$  then
3       $k = integer(rand(1, nP))$  // Randomly generated integer;  $1 \leq k \leq nP$ 
4       $L = rand < 0.5$  // Either 1 or 0
5       $X_z = (1 - L) \times x_{best2} + L \times X_i^{it}$ 
6      if  $rand < 0.5 \times (1 - \frac{it}{MaxIT})$  then
7           $X_{LCEO} = x_{best3} + \mu_1 \times (x_{best} - X_z) + \mu_2 \times (x_{r1} - x_{r2})$ 
8      else then
9           $X_{LCEO} = x_{best} + \mu_1 \times (x_{best} - X_z) + \mu_2 \times (x_{r1} - x_{r2})$ 
10     end if
11 end if

```

In Algorithm 1, nP is the number of members in the population, LC is a changing variable, x_{r1} and x_{r2} are two random members from the population, and X_i^{it} is the i^{th} member current solution in iteration it .

3.1.4. MGBO Algorithm

To start the optimization process, the first crucial step is to initialize the population. This step involves randomly generating a set of individuals that can be used to represent potential solutions to the optimization problem. Within the framework of the MGBO algorithm, N vectors are used as the population in a d -dimensional search space. In practice, the initial population is generated by selecting N points within the search space at random, as presented in the following equation:

$$X_i = X_{min} + r \times (X_{max} - X_{min}) \quad (27)$$

In (27), X_{max} and X_{min} are the upper and lower bounds, and r is a random number between 0 and 1.

The full algorithm flow proposed by [7] is presented in Algorithm 2, which provides a detailed overview of the steps involved in the optimization process. It is worth noting that the optimization process may require multiple iterations before the best solution is found. Each iteration involves updating the positions of the population in the search space based on their individual and collective behavior. Through this process, the algorithm can systematically navigate the search space, facilitating more effective exploration and convergence towards the optimal solution.

Algorithm 2 Metaheuristic gradient-based optimization

Require: ε , nP , nj , LC , $MaxIt$, X_{min} , and X_{max}

```

1   Set  $i = 1$ , and  $it = 1$ , randomly initialize the initial population  $X_{nP \times nj}$ 
2   Evaluate the objective function for the initially generated parameters
3   Update  $x_{best1}$ ,  $x_{best2}$ ,  $x_{best3}$ 
4   for  $it < MaxIt$  do
5       Define  $\alpha$  and  $\beta$  using (13) and (14)
6       while  $i \leq nP$  do
7           Define  $X_R = [x_{r1}, x_{r2}, x_{r3}, x_{r4}]$ 
8           Find  $\mu_1$ , and  $\mu_2$  using (12)
9           Evaluate  $DM_{X1}$ ,  $DM_{X2}$  and the GSR operators using (18), (23), and (24)
10          Evaluate  $X1_n^m = X_i^{it} - GSR + DM_{X1}$ 
11          Evaluate  $X2_n^m = X_i^{it} - GSR + DM_{X2}$ 
12          Evaluate  $X3_n^m$  using (26)
13          Update  $X_i^{it}$  using (25) and compute the objective function
14          if  $rand < LC$  then; Go to Algorithm 1
15              Compute  $X_{LCEO}$  and the objective function of  $X_{LCEO}$ 
16               $X_i^{it} = X_{LCEO}$ 
17          end if
18          Update  $x_{best1}$ ,  $x_{best2}$ ,  $x_{best3}$ 
19      end while
20  end for

```

In Algorithm 2, ε is a small number ($0 \leq \varepsilon \leq 0.1$), nj is the number of parameters, X_{min} , and X_{max} are the parameters upper and lower bounds, and $MaxIt$ is the maximum number of iterations.

3.2. Particle Swarm Optimization (PSO)

Particle swarm optimization (PSO) (Figure 6) is a metaheuristic computational method that emulates the social behavior of birds flocking or fish schooling. It is a population-based optimization algorithm that was first introduced by Kennedy and Eberhart in 1995 [10]. The algorithm starts with a population of particles, each representing a potential solution to the optimization problem. Each particle has a position in the search space and a velocity that determines how it moves through the space. The particles update their positions and

velocities based on their own experience and the experience of their neighbors, with the goal of moving toward the best solution found so far by the entire population [11]. The PSO algorithm starts with a population of particles, each representing a set of parameter values. Each particle has a position in the parameter space and a velocity that determines how it moves through the space. The particles update their positions and velocities based on their own experience and the experience of their neighbors, intending to move towards the best set of parameters found so far by the entire population [10].

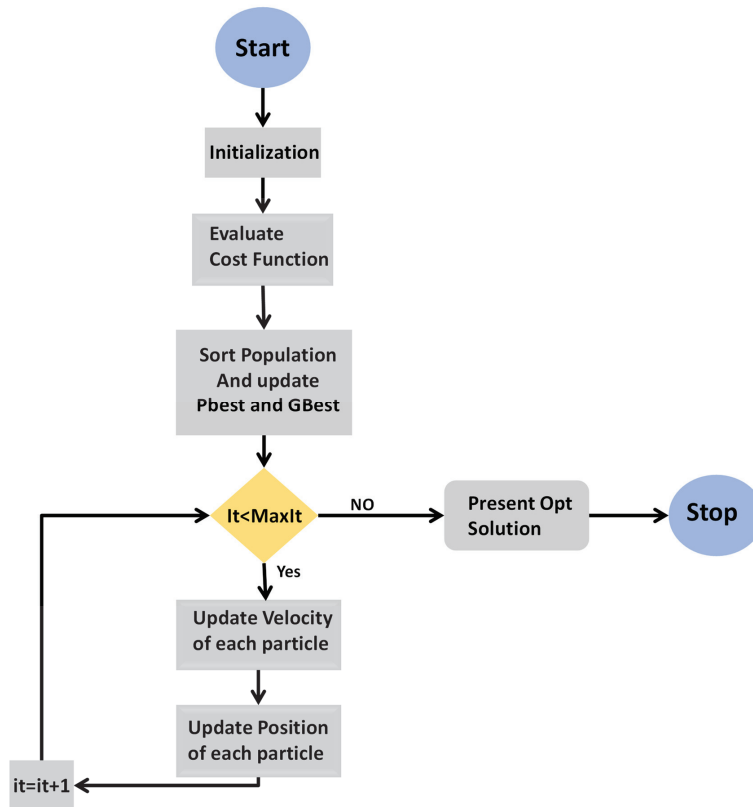


Figure 6. PSO flow chart.

The movement of a particle is determined by its current velocity, which is updated based on its personal best and global best positions, as well as a random value called the “inertia weight”. The velocity update rule is given by the following:

$$V(t+1) = W \times V(t) + c1 \times r \times (P_{best} - P(t)) + c2 \times r \times (G_{best} - P(t)), \quad (28)$$

where t is the current iteration, r is a random number between 0 and 1, W is the inertia weight, P_{best} is the best set of parameters (position values) a particle has encountered so far, G_{best} is the best set of parameters (position values) encountered by any particle in the population, and $c1$ and $c2$ are constants that control the relative importance of the personal best and global best positions. The position of the particle is updated based on its current velocity using the following equation:

$$Position(t+1) = Position(t) + Velocity(t+1) \quad (29)$$

3.3. Proposed Algorithms Variants

3.3.1. Modified Metaheuristic Gradient-Based Optimization Algorithm (M-MGBO) Application for SC Parameterization

The original MGBO algorithm, as described in [7], has been augmented with certain modifications. A noteworthy addition has been the introduction of a cross-over operator that assesses the adherence of the solution generated by (25) to the upper and lower

bounds. In the MGBO algorithm, should a solution exceed these bounds, a novel solution that remains within these confines is produced for the respective population member. This approach mirrors the methodology used during the initial population generation, as presented in (27). The modified cross-over operator functions by contrasting the solution X_i^{it} of the current iteration with the upper and lower bounds. If any parameter of the solution breaches these bounds, it is substituted with its previous value from the previous iteration, X_i^{it-1} .

An additional noteworthy constraint of the original MGBO algorithm becomes apparent during its transition into the LCEO phase. In this phase, the solution obtained by the MGBO approach, marked as X_i^{it} , is replaced with the solution obtained through the LCEO (X_{LCEO}^i). The basis for this change is grounded in the belief that the LCEO procedure often produces superior solutions, thereby enabling an escape from local minima. However, this advantage is mostly evident in the initial stages of the optimization process. The MGBO algorithm has an inherent tendency to guide the whole population towards the globally optimum solution as long as the gradient is not too small. The aforementioned tendency ultimately limits the effectiveness of the LCEO when the number of iterations increases.

Moreover, the incorporation of randomness by the LCEO algorithm serves the purpose of aiding the MGBO in converging toward the optimal solution. However, it is important to note that this randomness might also restrict the algorithm's capacity to escape from a local minimum. To tackle these concerns, several alterations have been suggested for the LCEO. The modified approach includes a comparator to evaluate and compare the responses produced by the LCEO and MGBO methods. This allows the algorithm to select the superior solution based on the fitness function. The comparator serves the additional purpose of mitigating the algorithm's tendency to become trapped in a local minimum. This is achieved by identifying instances in which the LCEO solution surpasses the MGBO solution in terms of cost function.

Secondly, the LCEO has been modified to reduce randomness and control the direction of the generated solution. The LCEO now moves towards the second and third best solutions, in addition to the best solution. This modification aims to increase population diversity and accuracy while also improving the ability of the algorithm to escape from a local minimum. By moving towards other good solutions, the LCEO is less likely to get stuck at a local minimum and more likely to find the global minimum. Moreover, to further facilitate the investigation of the LCEO technique, an additional variable denoted as C is introduced.

The value of C is used to replace α in (12). The parameter α is employed to determine the equilibrium between exploration and exploitation, while the variable C is incorporated to augment this dynamic within the LCEO and to increase the weight of the best solution movement. As depicted in Figure 7, the C value exhibits an increase with the decrease of α (initially). In the event that the algorithm gets confined to a local minimum within the initial 100 iterations, the incorporation of the C value amplifies the LCEO's capacity to provide a diverse solution, hence augmenting its exploitative potential. As a result, a new variable is defined (μ_3):

$$\mu_3 = 2 \cdot r \cdot C - C \quad (30)$$

Hence, the modifications to the LCEO aim to improve the performance of the MGBO by addressing some of the limitations of the original LCEO method. By introducing a comparator and modifying the LCEO to reduce randomness and control the direction of the generated solution, the algorithm is enhanced in its ability to locate the global minimum and avoid being trapped in a local minimum.

The comprehensive pseudocode detailing the modified MGBO (M-MGBO) algorithm, inclusive of the model's constraints and the modifications, is provided in Algorithm 3. Corresponding functions associated with this algorithm can be found in Algorithm 4 and Algorithm 5.

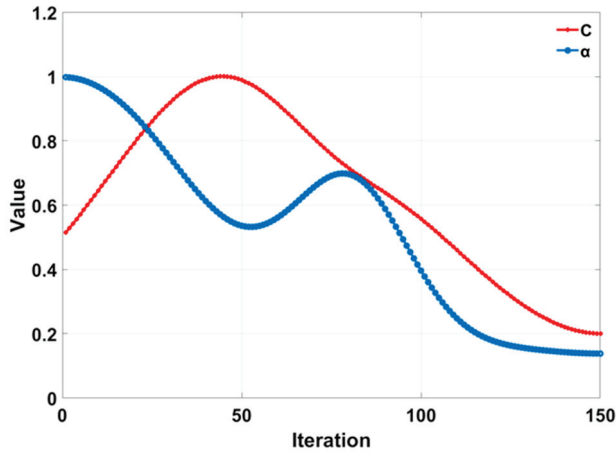


Figure 7. Comparison between α and C .

The LCEO modifications have the potential to significantly enhance the performance of optimization algorithms and lead to better solutions across a broad spectrum of problems. The updated LCEO algorithm is given in Algorithm 6.

Algorithm 3 Modified MGBO

Require: $\mathcal{E}$, nP , nj , LC , $MaxIt$, X_{min} , and X_{max}

- 1 Set $i = 1$, and $it = 1$, randomly initialize the initial population $X_{nP \times nj}$
 - 2 Evaluate the MSE for the population initially; $MSE = \frac{1}{\sum_{i=1}^n (X_i - \tilde{X}_i)^2}$
 - 3 Update x_{best1} , x_{best2} , x_{best3}
 - 4 **for** $it < MaxIt$ **do**
 - 5 Define α and β using (13) and (14)
 - 6 **while** $i \leq nP$ **do**
 - 7 $X_{old} = X_i^{it}$ // $X_{old} = [K_v(i), R_1(i), C_1(i), R_2(i), C_2(i), R_3(i), C_3(i), R_L(i)]_j$
 - 8 Define $X_R = [x_{r1}, x_{r2}, x_{r3}, x_{r4}]$ // **Random members from the population**
 - 9 Evaluate μ_1 , and μ_2 using (12)
 - 10 Evaluate DM_{X1} , DM_{X2} and the GSR operators using (18), (23), and (24)
 - 11 Evaluate $X1_n^m = X_i^{it} - GSR + DM_{X1}$
 - 12 Evaluate $X2_n^m = X_i^{it} - GSR + DM_{X2}$
 - 13 Evaluate $X3_n^m$ using (26)
 - 14 Update X_i^{it} using (25) // $X_i^{it} =$
 - 15 $X_i^{it} = \text{limit}(X_i^{it}, X_{old})$ // **Check for upper and lower bounds, Algorithm 4**
 - 16 Let $A = [X_i^{it}(C_1), X_i^{it}(C_2), X_i^{it}(C_3)]_j$ and $B = [X_i^{it}(R_1), X_i^{it}(R_2), X_i^{it}(R_3)]_j$
 - 17 $X_i^{it} = \text{Check}(A, B, X_{old})$ // **Checking for constraints, Algorithm 5**
 - 18 Evaluate cost function of X_i^{it} using Equation in line 2.
 - 19 **if** $\text{rand} < LC$ **then** // **Local escaping operator, Algorithm 6**
 - 20 Compute X_{LCEO} and the objective function of X_{LCEO}
 - 21 $X_{LCEO}^i = \text{limit}(X_{LCEO}^i, X_i^{it})$ // **Check for upper and lower bounds**
 - 22 Let $C = [X_{LCEO}^i(C_1^{LCEO}), X_{LCEO}^i(C_2^{LCEO}), X_{LCEO}^i(C_3^{LCEO})]_j$ and $D =$
 - 23 $X_{LCEO}^i = \text{Check}(C, D, X_i^{it})$ // **Checking for constraints**
 - 24 Evaluate cost function for X_{LCEO}^j using Equation in line 2.
 - 25 $X_i^{it} = \text{Compare}(X_i^{it}, X_{LE}^i, \text{Cost}(X_i^{it}), \text{Cost}(X_{LE}^i))$ // **To update X_i^{it} if**
 - 26 **end if**
 - 27 Update x_{best1} , x_{best2} , x_{best3} , and x_{worst}
 - 28 **end while**
 - 29 **end for**
-

Algorithm 4 Checking upper and lower boundsRequire: nj, X_{min} and X_{max}

```

1  function  $X = \text{limit}(X1, X2)$ 
2      for  $j = 1:nj$ 
3          if  $(X1(j) < X_{min}(j) \parallel X1(j) > X_{max}(j))$  then
4               $X1(j) = X2(j)$ 
5          end if
6      end for
7  end function

```

Algorithm 5 Comparing two members of the population and keeping the one with the lowest cost function

```

1  function  $X = \text{compare}(X1, X2, \text{Cost1}, \text{Cost2})$ 
2      if  $(\text{Cost1} > \text{Cost2})$  then
3           $X = X2$ 
4      else
5           $X = X1$ 
6      end if
7  end function

```

Algorithm 6 MODIFIED Local ESCAPING OPERATOR (LCEO)Require: $\mu_1, \mu_2, X_i^{it}, x_{best}, x_{best2}, x_{best3}, x_{r1}, x_{r2}, nP, \text{beta}, \alpha$, and LC

```

1   $LC = 4 \times LC \times (1 - LC)$ 
2  if  $\text{rand} < LC$  then
3       $k = \text{integer}(\text{rand}(1, nP))$  // Randomly generated integer;  $1 \leq k \leq nP$ 
4       $L = \text{rand} < 0.5$  // Either 1 or 0
5       $X_z = (1 - L) \times x_{best2} + L \times (x_{best3} - X_i^{it})$ 
6       $C = \text{beta} - \alpha \left(1 - \frac{it}{MaxIt}\right)^2$  and  $\mu_3 = 2 \cdot r \cdot C - C$ 
7      if  $\text{rand} < 0.5 \times \left(1 - \frac{it}{MaxIt}\right)$  then
8           $X_{new}^{it} = x_{best2} + \mu_3 \times (x_{best} - X_z) + \mu_2 \times (x_{best3} - x_{r2})$ 
9      else then
10          $X_{new}^{it} = x_{best} + \mu_3 \times (x_{best} - X_z) + \mu_2 \times (x_{best2} - x_{r1})$ 
11     end if
12 end if

```

3.3.2. Hybrid PSO-MGBO

The hybrid PSOGBO algorithm leverages the strengths of both the PSO and M-MGBO algorithms to achieve a better performance. The PSO algorithm is the primary algorithm used for solving the optimization problem, while the M-MGBO algorithm is used to direct

the population toward the optimal solution. Initially, at the start of the iteration, the PSO algorithm updates the velocity and position of each particle in the population. Next, the ratio of the cost (MSE) of the current solution to the global best solution cost is compared to a threshold. If this ratio exceeds the threshold, the particle is recorded and saved into a vector. This process is repeated for each particle in the population, and the newly generated vector is used as the initial population for the M-MGBO algorithm. The M-MGBO algorithm is then executed for a specified number of iterations, generating new members of the population that are compared to the original members to determine if the newly generated positions are improvements. Following this step, the global and personal best solutions are updated for the hybrid algorithm.

By combining the PSO algorithm's efficiency in exploring the search space with the M-MGBO algorithm's ability to guide the search towards the optimal solution, the hybrid approach achieves a better performance than either algorithm on its own. The algorithm flow chart is shown in Figure 8.

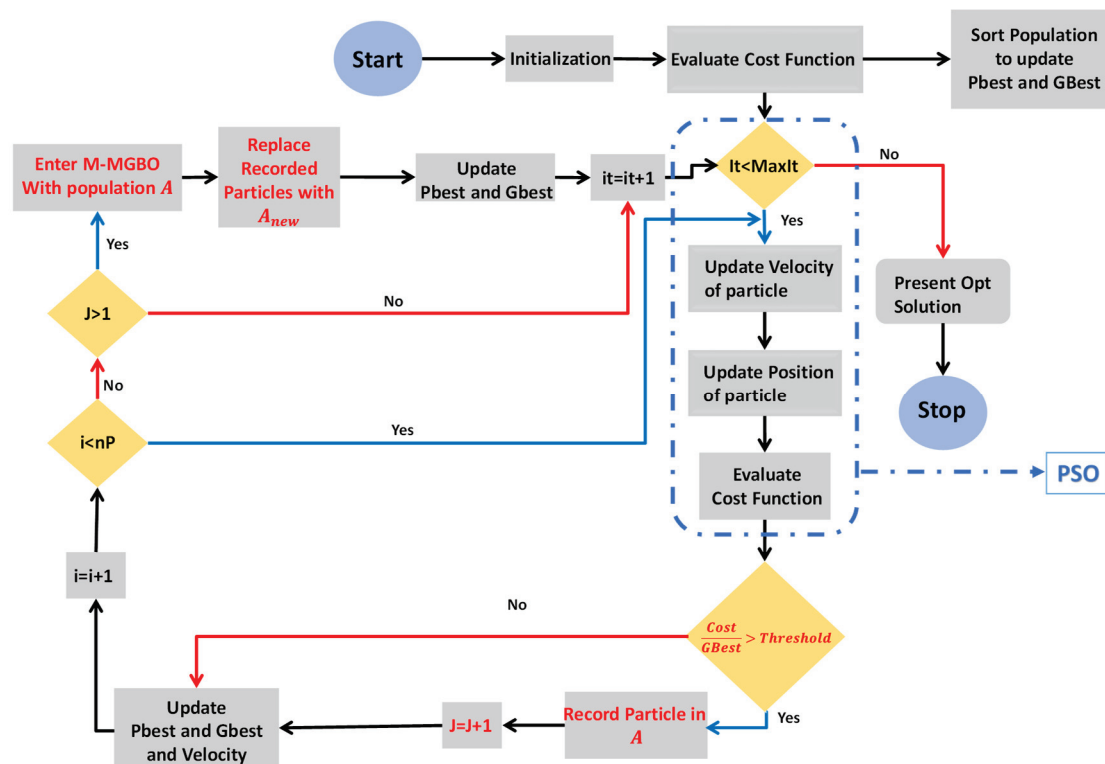


Figure 8. Flow chart hybrid PSO-MGBO method developed in this work. P_{best} is the personal best solution, and G_{Best} is the global best solution. Text in red indicates the related steps to the M-MGBO algorithm.

3.3.3. Adding the Local Escaping Operator to the PSO

Another variant of the PSO algorithm introduced in this research involves the incorporation of the local escaping operator (LCEO) into the standard PSO algorithm. The LCEO generates a superior particle that is guided by the global best solution (x_{best}), x_{best2} , and x_{best3} , hence enhancing the diversity of the population. Similar to the MGBO and M-MGBO algorithms, the LCEO is used to overcome the problem of particles getting stuck in local optima by allowing the particle to escape from its current local optima and move toward the global optima.

After generating the LCEO solution X_{LCEO} (Algorithm 6), it is compared to the PSO-generated solution, the global best solution, and the personal best solution. If the performance of the X_{LCEO} solution in terms of MSE is better than the PSO-generated solution, the particle's current position is changed to the X_{LCEO} position, and the particle's velocity is computed again. Otherwise, the X_{LCEO} solution is discarded, and the particle's current

position remains unchanged. This process is repeated until the optimal solution is found. Algorithm 7 provides the pseudocode of the algorithm applied for SC parameterization. The use of LCEO in the PSO algorithm has exhibited a promising performance in tackling optimization problems, particularly in scenarios where the search space is complex or where particles tend to get trapped in local optima.

Algorithm 7 PSO-LCEO

Require: nP , nj , LC , $MaxIt$, X_{min} , X_{max} , $C1$, $C2$, W_{max} , W_{min} , V_{max} , and V_{min}

```

1  Initialize set of Random population,  $P = [K_v, R_1, C_1, R_2, C_2, R_3, C_3, R_L]_{np \times nj}$ .
2  Evaluate the MSE for the population initially;  $MSE = \frac{1}{n} \sum_{i=1}^n (P_i - \tilde{P}_i)^2$ 
3  Select  $[G_{best}, P_{best}]$  from population // Such that  $G_{best} = \min(Cost)$  over entire
   population, and  $P_i^{best} = \operatorname{argmin}(Cost^i)$  at the  $i^{\text{th}}$  iteration.
4  for  $j < MaxIt$  do
5      Evaluate  $W$ ;  $W = W_{max} - \left( (W_{max} - W_{min}) * \frac{j}{MaxIt} \right)$ 
6      while  $i \leq nP$  do
7           $X_{old} = P_i^j$  //  $X_{old} = [K_v(i), R_1(i), C_1(i), R_2(i), C_2(i), R_3(i), C_3(i), R_L(i)]_j$ 
8          Update  $V_i^j$  using (28)
9           $V_i^j = \operatorname{limit}(V_i^j)$  // Check for upper and lower bounds; Algorithm 4
10         Update  $P_i^j$  using (29)
11         Let  $A = [P_i^j(C_1), P_i^j(C_2), P_i^j(C_3)]_j$  and  $B = [P_i^j(R_1), P_i^j(R_2), P_i^j(R_3)]_j$ 
12          $P_i^j = \operatorname{limit}(P_i^j)$  // Check for upper and lower bounds; Algorithm 4
13          $P_i^j = \operatorname{Check}(A, B, X_{old})$  // Checking for constraints; Algorithm 5
14         Evaluate  $Cost(P_i^j)$  using Equation in line 2
15         if  $\operatorname{rand} < LC$  then // Local escaping operator Algorithm 6
16             Compute  $X_{LCEO}$  and the objective function of  $X_{LCEO}$ 
17              $X_{LCEO}^i = \operatorname{limit}(X_{LE}^i, P_i^j)$  // Check for upper and lower bounds
18             Let  $C = [X_{LCEO}^i(C_1^{LCEO}), X_{LCEO}^i(C_2^{LCEO}), X_{LCEO}^i(C_3^{LCEO})]_j$  and  $D =$ 
                $[X_{LCEO}^i(R_1^{LCEO}), X_{LCEO}^i(R_2^{LCEO}), X_{LCEO}^i(R_3^{LCEO})]_j$ 
19              $X_{LCEO}^i = \operatorname{Check}(C, D, P_i^j)$  // Checking for constraints
20             Evaluate cost function for  $X_{LCEO}^j$  using Equation in line 2.
21              $P_i^j = \operatorname{Compare}(P_i^j, X_{LE}^i, Cost(P_i^j), Cost(X_{LE}^i))$  // To update  $P_i^j$  if
                $Cost(X_{LE}^i) < Cost(P_i^j)$ 
22         end if
23         Update  $[G_{best}, P_i^{best}]$  // By comparing  $Cost(P_i^j)$  to  $Cost(G_{best})$  and
                $Cost(P_i^{best})$ 
24     end while
25 end for

```

4. Results and Discussion

In this study, the Zubieta model parameters were determined for an SC bank comprising six capacitors in series, each with a rating of 2300 F, resulting in a total bank capacitance of 383.33 F. The estimation process involved using three different input signals: a step input with amplitudes of 10 A (Figure 4a) and 4 A (Figure 4b), along with a variable frequency

input (Figure 4c), corresponding to profiles 1, 2, and 3 respectively. For each input, four distinct algorithms (M-MGBO, PSO, PSOLCEO, and PSO-MGBO) were executed in three separate runs. The Zubieta model, characterized by three branches, requires the estimation of a total of eight parameters, namely R_1 , R_2 , R_3 , C_1 , C_2 , C_3 , Kv , and R_L . Each of the algorithms (M-MGBO, PSO, PSOLCEO, and PSO-MGBO) requires initial guesses, which are formed using the analytical Zubieta model parameters and also using any initial guess selection rules the algorithms may have. These are detailed in the listings of each algorithm—for example, certain algorithms use the random initialization of the parameters' values in the given bounds. The bounds themselves are derived using the parameters' values derived using Zubieta's analytical method. In each iteration of the respective algorithm, the parameters are estimated following the respective algorithmic procedures as outlined in Section 3. These estimated parameters are then used to estimate the supercapacitor terminal voltage at the given value of current, following Zubieta's model equations. The mean square error between the measured and estimated supercapacitor terminal voltage, which is obtained using the estimated parameters, is used as the cost/fitness function to evaluate the fitness of the estimated parameters.

It is worth noting that the existence of a voltage-dependent capacitance in one of the Zubieta model's branches makes this a nonlinear model. It is this nonlinear model that is used to evaluate the fitness of the estimated model parameters. Moreover, the MGBO algorithm used incorporates both gradient and population-based techniques to determine the search direction using Newton's method. This search direction is based on a linear estimation of the gradient of the cost/fitness function. However, such approximation can make the algorithm prone to being stuck in a local optimum. To avoid this, the local escaping operator (LCEO)-based routine is used to augment the MGBO algorithm into the M-MGBO algorithm.

4.1. Parameters Estimation Results

The evaluation of the algorithms' performance considered various criteria, including MSE and the computation time. Optimization took place on a personal computer equipped with an Intel(R) Xeon(R) E-2124 central processing unit operating at a frequency of 3.30 gigahertz and 16 gigabytes of random-access memory.

The cost function for the optimization problem was set to minimize the difference between the simulated SC voltage from the derived model and the actual measured voltage.

$$\text{Minimize } SSE = \sum_{t=1}^T [V_t(t) - V_m(t)]^2, \quad (31)$$

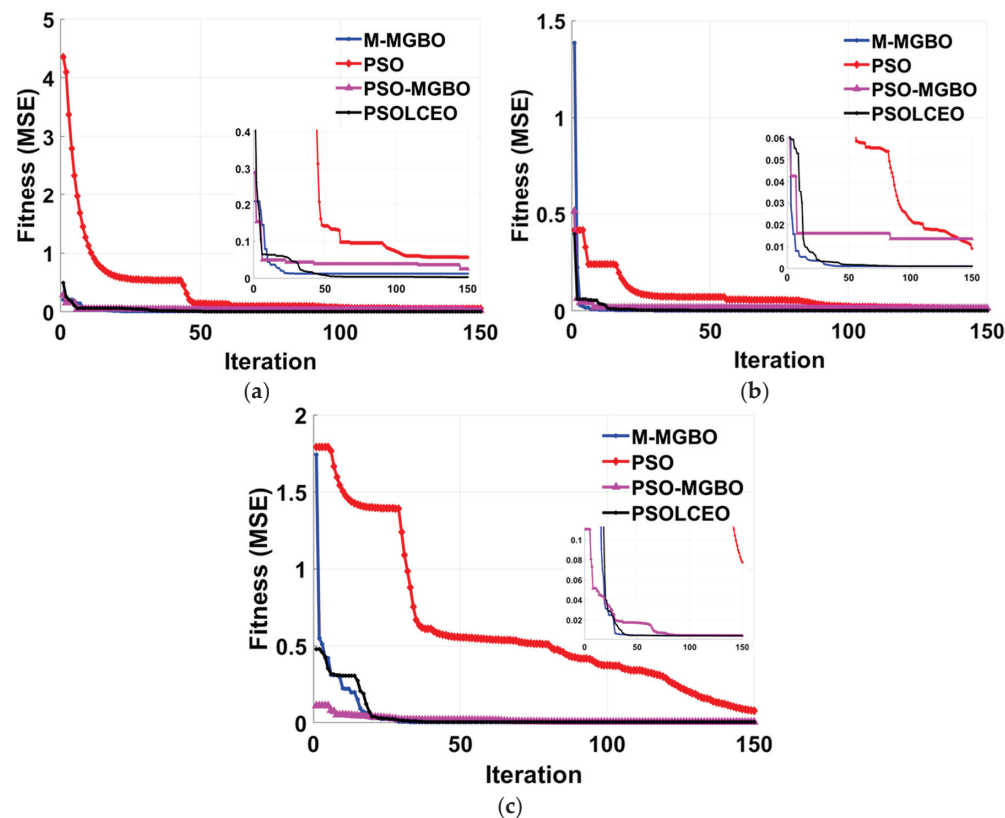
where $V_t(t)$ is the capacitor terminal voltage given by Equation (7) obtained using estimated model parameters, and $V_m(t)$ is the measured supercapacitor voltage.

The M-MGBO algorithm demonstrated superior performance compared with the other algorithms. In the case of Profile 1 (Table 3), M-MGBO and PSO-LCEO exhibited MSE values of 0.003927 and 0.003493, respectively. Notably, M-MGBO boasted the shortest computational time at 20,141.028 s, outperforming PSO-LCEO (26,782.93 s) and the hybrid PSO-MGBO (77,735.494 s). Additionally, parameters obtained by the M-MGBO algorithm were closer to analytically derived parameters (Table 2). For instance, the first branch capacitor C_1 had a value of 0.7723 F with M-MGBO compared with 14.607195 F with PSO-LCEO.

Table 3. Experimental results for the parameters estimated using current Profile 1 with 150 iterations (T1, T2, and T3 are the respective branches' time constants).

Parameter	M-MGBO	PSO	PSO-LCEO	PSO-MGBO
$K_v \left(\frac{F}{V} \right)$	38.45929984	14.07398	33.22398	33.266903
$R_1 (\Omega)$	0.013163046	0.127028	0.014034	0.04586
$C_1 (F)$	0.77232877	48.13062	14.60719	0.95056
$R_2 (\Omega)$	0.487789546	5.190033	0.4396	1.61732
$C_2 (F)$	224.0938314	154.1119	168.6782	175.7074
$R_3 (\Omega)$	36.734606504	21.07912	4.807739	42.54308
$C_3 (F)$	394.158958	331.8499	214.2253	243.417
$R_l (\Omega)$	25.70462833	68.13273	62.47676	37.94408
MSE	0.003927391	0.057314	0.003494	0.025034
Elapsed Time	20141.028	16564.040	26782.930	77735.494
T1	0.010166	6.113936	0.204997	0.043593
T2	109.3106	799.8458	74.15094	284.1751
T3	14479.27	6995.104	1029.939	10355.71
Average MSE	0.006813	0.07581693	0.008789473	0.03842393

The convergence curve plays a vital role in assessing optimization algorithms, offering insights into the speed and efficiency of their convergence to the optimal solution. Notably, the convergence curve for the M-MGBO algorithm stands out for its swift and stable convergence with minimal oscillation. While the PSO-LCEO algorithm also showed stable convergence, it did so at a slower pace compared with the M-MGBO algorithm. For Profile 1, Figure 9a illustrates the convergence curve of the best run for all four algorithms, showcasing the M-MGBO algorithm's rapid convergence. Simultaneously, Figure 10a presents the voltage response for the same input.

**Figure 9.** (a) Convergence curve for Profile 1, (b) Convergence curve for Profile 2, and (c) Convergence curve for Profile 3.

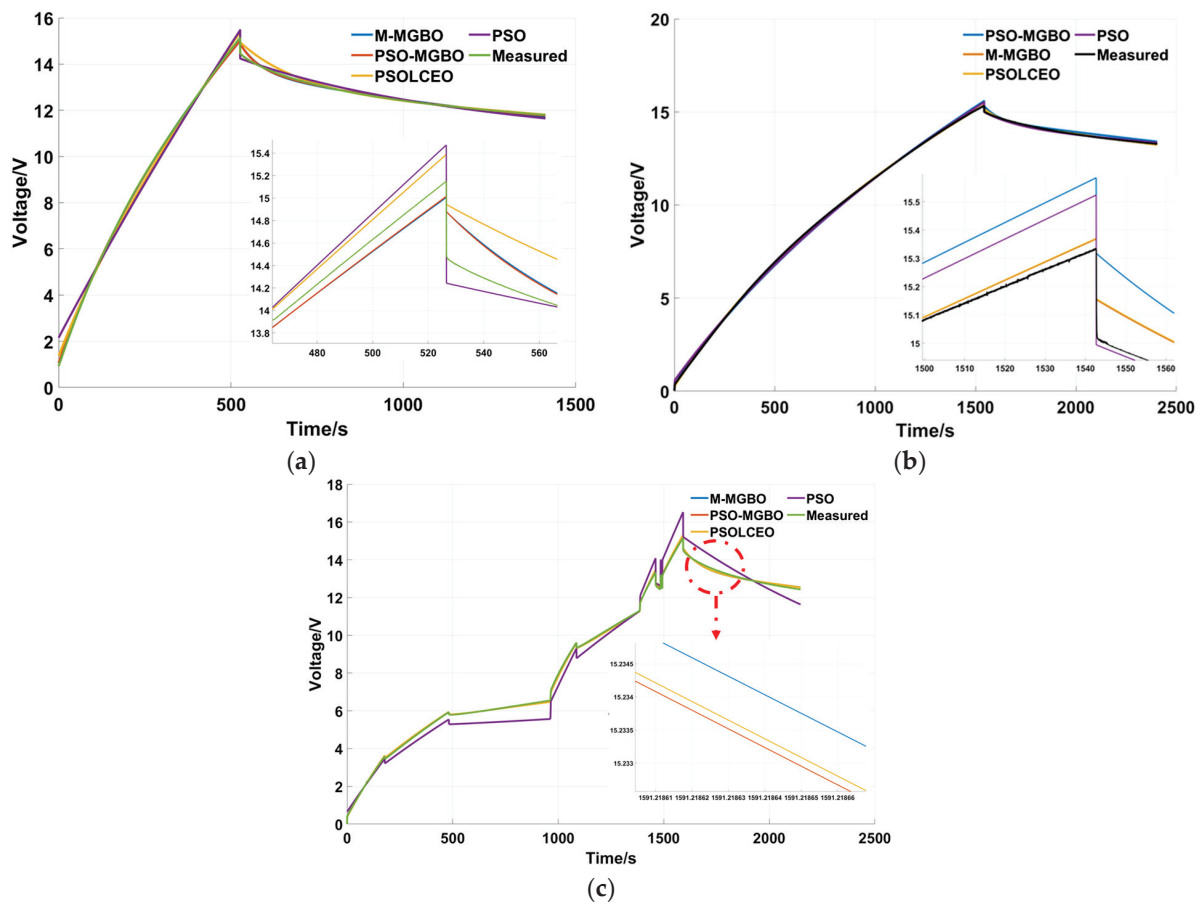


Figure 10. (a) Voltage response for Profile 1, (b) Voltage response for Profile 2, and (c) Voltage response for Profile 3.

Table 4 demonstrates the performance of the four optimization algorithms in estimating the parameters of a Zubieta supercapacitor model using Profile 2 as input. The M-MGBO algorithm exhibited the lowest MSE of 0.001001 and a significantly shortened computational time of 31,320.07 s. In contrast, the PSO-LCEO algorithm yielded a very similar MSE of 0.001017 compared with the M-MGBO algorithm. However, it exhibited a longer execution time of 36,297.98 s.

Table 4. Experimental results for the parameters estimated using current Profile 2 with 150 iterations (T1, T2, and T3 are the respective branches' time constants).

Parameter	M-MGBO	PSO	PSO-LCEO	PSO-MGBO
$K_v \left(\frac{F}{V} \right)$	37.2805859	27.8131	36.4359	36.86813
$R_1 (\Omega)$	0.05835297	0.07556	0.056184	0.149894
$C_1 (F)$	0.03307588	17.957	0.627231	0.105358
$R_2 (\Omega)$	0.69575519	0.5943	0.701702	5.188335
$C_2 (F)$	228.197898	163.674	226.8816	198.3325
$R_3 (\Omega)$	24.4703793	31.6255	18.62833	37.35831
$C_3 (F)$	394.767419	330.281	273.4351	251.036
$R_l (\Omega)$	52.358648	56.0493	94.7854	63.34033
MSE	0.0010013	0.01342	0.001017	0.009034
Elapsed Time	31320.07	109260	36297.98	22324.22
T1	0.00193	1.356831	0.03524	0.015793
T2	158.7699	97.27146	159.2033	1029.015
T3	9660.108	10445.3	5093.639	9378.281
Average MSE	0.001001	0.103347	0.0031693	0.107982

The hybrid PSO-MGBO and standard PSO algorithms exhibited higher MSE values of 0.01342 and 0.009034, respectively. In terms of average MSE, the M-MGBO algorithm outperformed others with a value of 0.001001, followed by the PSOL-CEO algorithm at 0.003169. The hybrid PSO-MGBO and regular PSO algorithms had higher average MSE values at 0.103347 and 0.107982, respectively. It is noteworthy that the LCEO operator, which is employed in the PSO-LCEO algorithm, serves as a method for escaping local optima during the optimization process. Despite this feature, the PSO-LCEO algorithm could not provide superior parameter estimates compared with the M-MGBO algorithm in this study. For the convergence curve of the best run for the four algorithms, refer to Figure 9b, while Figure 10b presents the corresponding voltage response.

The parameters obtained using Profile 3 input are presented in Table 5. The M-MGBO algorithm exhibited an MSE of 0.0042315, while the PSO-LCEO showed a comparable performance with an MSE of 0.004231. The hybrid PSO-MGBO had a similar MSE of 0.004231 but still outperformed the PSO-LCEO in terms of computational time, completing the task in only 32,206 s compared with 39,252 s. Notably, the M-MGBO algorithm showcased the best performance in terms of simulation time, with a runtime of 28,622 s.

Table 5. Experimental results for the parameters estimated using a variable frequency current profile with 150 iterations (T1, T2, and T3 are the respective branches' time constants).

Parameter	M-MGBO	PSO	PSO-LCEO	PSO-MGBO
$K_v \left(\frac{F}{V} \right)$	51.74533692	125.415	51.72809	51.74644
$R_1 (\Omega)$	0.080265842	0.1346	0.080451	0.015515
$C_1 (F)$	0.012783609	71.44	0.002166	0.572341
$R_2 (\Omega)$	0.572682701	8.90377	0.571799	0.080359
$C_2 (F)$	223.6858074	186.193	223.5703	223.6356
$R_3 (\Omega)$	49.99947713	24.526	50	49.99997
$C_3 (F)$	399.9692153	292.147	400	400
$R_l (\Omega)$	20.90658189	31.4604	20.86642	20.90162
MSE	0.004231544	0.30036	0.004231	0.004231
Elapsed Time	28622.51195	19408.00527	39252.59362	32206.02127
T1	0.001026	9.615824	0.000174	0.00888
T2	128.101	1657.82	127.8373	17.97113
T3	19998.25	7165.197	20000	19999.99
Average MSE	0.004233505	0.226921781	0.010902	0.004257128

It can be noted from the results that the M-MGBO and PSO-LCEO algorithms showed the best performance in terms of MSE, while the PSO-MGBO algorithm had a good balance between performance and simulation time. The PSO algorithm had the lowest computational time, but it had a significantly worse performance in terms of MSE. The convergence curve for the best run of the four algorithms is presented in Figure 9c, and the voltage response is presented in Figure 10c.

4.2. M-MGBO Results Validation

The results of 30 iterations of the M-MGBO method used to estimate the parameters of the Zubieta model of supercapacitors using a variable frequency input are presented in Table 6. The primary objective of this analysis is to assess the algorithm's performance in terms of the consistency of parameter estimations, the accuracy of results pertaining to resistance and capacitance values, and any other noteworthy observations obtained from the table.

Table 6. Experimental results for the parameters using M-MGBO and variable frequency with 30 runs (T1, T2, and T3 are the respective branches' time constants).

Parameter	Standard Deviation	Average
$K_v\left(\frac{F}{V}\right)$	0.346049	51.95607
$R_1(\Omega)$	0.000967	0.079637
$C_1(F)$	0.005086	0.01552
$R_2(\Omega)$	0.02133	0.555367
$C_2(F)$	0.129872	223.7431
$R_3(\Omega)$	0.070908	49.96596
$C_3(F)$	0.912619	399.2558
$R_l(\Omega)$	0.405767	21.18428
T1	0.000393	0.001233
T2	4.766871	124.2593
T3	72.75695	19949.26

It is also important to consider the parameters' standard deviation while analyzing the algorithm's performance, as it provides a measure of the algorithm's consistency. The standard deviation is a statistical measure used to assess the variability in parameter estimations across the 30 runs. Lower values of standard deviation indicate a higher level of consistency in the estimations. It is worth mentioning that certain parameters, including $R_1(\Omega)$ and $R_2(\Omega)$, exhibit remarkably low standard deviation values, specifically 0.0009 and 0.02, respectively. The observed low values of standard deviation suggest that the M-MGBO algorithm exhibits a consistent pattern of convergence, yielding parameter estimations that are highly similar across every iteration. The level of consistency demonstrated in this context is of the utmost importance for practical applications, as it ensures the reliability and predictability of outcomes.

4.3. MGBO and M-MGBO Comparison

To verify the reliability and assess the effectiveness of the modifications applied to the original MGBO algorithm, three separate runs were executed using the unaltered MGBO algorithm but with varying frequency inputs. The MGBO algorithm produced an average MSE of 0.0064, with a standard deviation of 5.7735×10^{-7} over the course of these three runs. Table 7 provides a comparative analysis between the M-MGBO and MGBO. Additionally, Figure 11 illustrates the convergence curve for the best-performing runs.

Table 7. Comparison of M-MGBO and M-MGBO best run 150 iterations (T1, T2, and T3 are the respective branches' time constants).

Parameter	MGBO	M-MGBO
$K_v\left(\frac{F}{V}\right)$	53.00867	51.74533692
$R_1(\Omega)$	0.229319	0.080265842
$C_1(F)$	0.0488278	0.012783609
$R_2(\Omega)$	0.872346	0.572682701
$C_2(F)$	223.6373	223.6858074
$R_3(\Omega)$	50	49.99947713
$C_3(F)$	399.9993	399.9692153
$R_l(\Omega)$	20.90988	20.90658189
T1	127.9978	17.9543
T2	19999.97	19998.2516
T3	53.00867	51.74533692
MSE	0.0063571	0.004231544

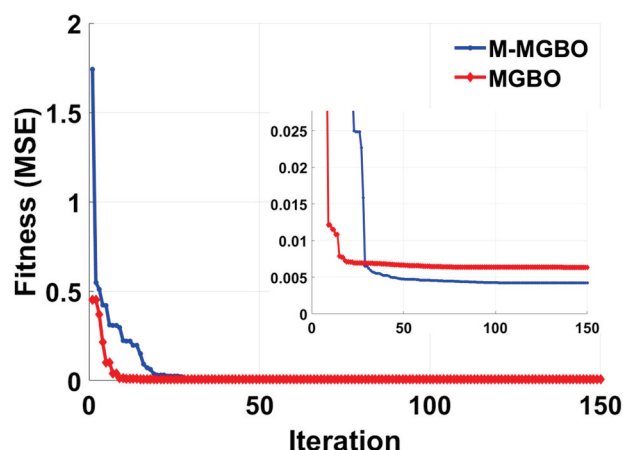


Figure 11. Best run convergence curve comparison between M-GBO and M-MGBO for variable frequency input.

For Kv the MGBO yielded a value of 53.0087 F/V, while the M-MGBO resulted in 51.7453 F/V. Despite the standard M-MGBO exhibiting a greater MSE when compared to the M-MGBO, it demonstrated a lower standard deviation for Kv . Specifically, the standard deviation for Kv was 0.0702 for the standard MGBO and 0.2311 for M-MGBO. This pattern is consistent for both $R3$ and RL .

4.4. Parameters Testing

As demonstrated earlier, the M-MGBO method provides a more precise estimation of parameters concerning accuracy and MSE. To conduct additional testing on parameters from the four distinct methods, the optimal parameters acquired from the three runs (as presented in Tables 4–6) undergo further evaluation by interchanging and applying them to the remaining inputs. For example, parameters obtained using Profile 1 input are tested using Profile 2 and Profile 3 as inputs, and their mean squared error (MSE) is recorded.

Table 8 presents a scenario where parameters, initially estimated via Profile 1, were subsequently evaluated with Profile 2 and Profile 3 inputs. In the experiment with Profile 2, it is evident that the parameters of the M-MGBO algorithm yield the lowest MSE value of 1.4291. Furthermore, the PSO, PSO-MGBO, and PSO-LCEO had similar MSE values. Specifically, the PSO-LCEO algorithm demonstrated the second lowest MSE of 1.6312, followed by the PSO-MGBO method, with an MSE of 1.6396, and the PSO algorithm, with an MSE of 1.6442. When considering the use of parameters with Profile 3, it was seen that the M-MGBO algorithm exhibited the lowest MSE of 0.1195. Subsequently, the PSO-LCEO program showed an MSE of 0.1469, while the PSOGBO algorithm yielded an MSE of 0.1482. The PSO method demonstrated the most significant MSE value, measuring 0.4614.

Table 8. MSE for Profile 1 parameters tested with Profiles 2 and 3 as inputs.

Algorithm	Profile 2 (4A Input)	Profile 3 (Variable Frequency Input)
MGBO	----	0.1489
PSO-GBO	1.6396	0.1482
M-MGBO	1.4291	0.1195
PSO-LCEO	1.6312	0.1469
PSO	1.6442	0.4614

Table 9 presents the results of the parameter testing conducted using parameters estimated using Profile 3 but tested with Profile 1 and Profile 2. The M-MGBO parameters yielded the lowest MSE in all tests. Specifically, the MSE was found to be 0.7448 for the Profile 2 input and 0.1754 for the Profile 1 input. The parameters of the PSO-LCEO

algorithm exhibited the second-lowest MSE for both tests. Specifically, the MSE was found to be 0.7464 for the 4 A step input and 0.2001 for the 10 A step input.

Table 9. MSE for Profile 3 parameters tested with Profiles 1 and 2 as inputs.

Algorithm	Profile 2 (4 A Input)	Profile 1 (10 A Input)
MGBO	----	----
PSO-MGBO	0.8144	0.5252
M-MGBO	0.7448	0.1754
PSOL-CEO	0.7464	0.2001
PSO	0.8645	0.5794

The PSO-MGBO algorithm parameters have been determined to be the third-best performing algorithm, whereas the PSO algorithm exhibited the greatest mean MSE for both test cases. In a similar manner, while employing parameters estimated using Profile 2 tested with Profile 1 as input (Table 10), the same pattern emerged. The M-MGBO algorithm parameters yielded the lowest MSE value of 1.4149, followed by the PSO-LCEO and the PSO-MGBO, with MSE values of 1.5438 and 1.5349, respectively. In the instance of the Profile 3 input, the PSOL-CEO parameters yielded the lowest MSE of 0.9877, while the M-MGBO resulted in a slightly higher MSE of 0.9902. Furthermore, the particle swarm optimization (PSO) method yielded the highest MSE for both test scenarios.

Table 10. M-MGBO variable frequency 30 runs parameters standard deviation and average.

Algorithm	Profile 1 (10 A Input)	Profile 3 (Variable Frequency Input)
MGBO	----	1.2619
PSO-MGBO	1.5349	0.9896
M-MGBO	1.4149	0.9902
PSO-LCEO	1.5438	0.9877
PSO	1.6930	1.2362

To perform a more exhaustive evaluation of the algorithm's parameters, two supplementary tests were conducted. These tests involved using a step input of 8 A (See Figure 12a) and a controlled current sequence as inputs. The optimal run parameters derived from prior experiments were used in both tests. Additionally, the performance is also assessed by comparing it to the analytical technique in order to validate the suitability of the parameters derived from the four algorithms.

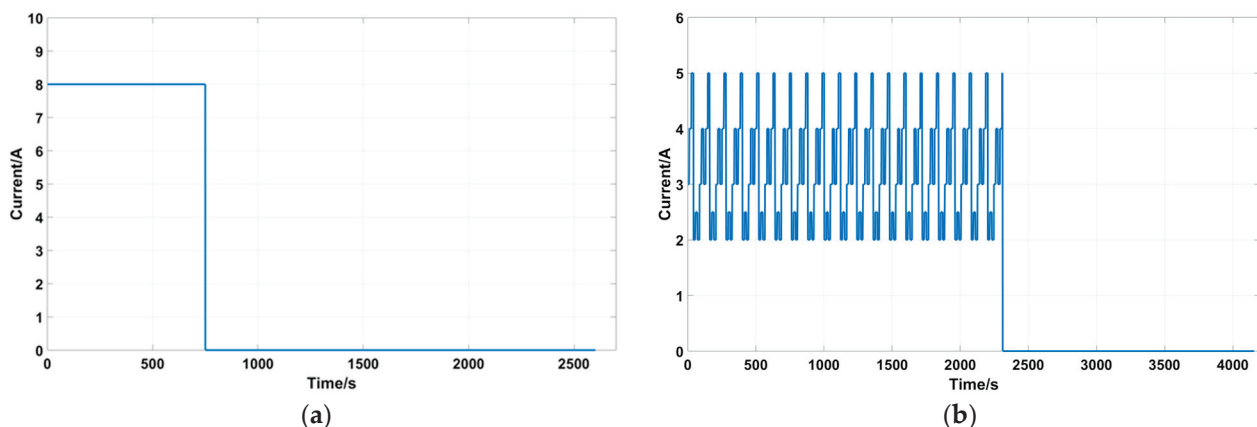


Figure 12. (a) 8 A step input, (b) Controlled current sequence.

As evident from Table 11, concerning the Profile 2 parameters, the M-MGBO algorithm demonstrates the lowest MSE at 0.9619, followed by the PSOLCEO with the second lowest

MSE of 0.9793, and subsequently, the PSO-MGBO algorithm. In the case of the Profile 1 parameters, the PSO-LCEO showcased the lowest MSE of 0.1418, followed by the PSO-MGBO algorithm, while the M-MGBO ranked third, with an MSE of 0.2699. Profile 3 parameters exhibited a closely aligned response among M-MGBO, PSO-LCEO, and PSO-MGBO, displaying minimal deviation in MSE values. Conversely, the PSO algorithm, across various inputs, performed the least favorably, indicating higher MSE values and implying lower accuracy and reliability in the estimated parameter.

Table 11. Terminal voltage MSE for test using 8A profile, with parameters estimated using profiles 1, 2, and 3 and the analytical approach parameters.

Algorithm	Profile 2 (4 A Parameters)	Profile 1 (10 A Parameters)	Profile 3 (Variable Frequency Parameters)
PSO-MGBO	0.9848	0.1617	0.0942
M-MGBO	0.9619	0.2699	0.0943
PSO-LCEO	0.9793	0.1418	0.0945
PSO	1.1325	0.2824	0.3063
Analytical		7.7745	

The controlled current sequence profile used for the following tests is illustrated in Figure 12b. According to the data presented in Table 12, it can be observed that the M-MGBO algorithm exhibited superior performance compared with the other algorithms and the analytical approach in terms of Profile 2 parameters. Specifically, the M-MGBO algorithm achieved an MSE value of 1.5037.

Table 12. Terminal voltage MSE for test using controlled current sequence profile, with parameters estimated using profiles 1, 2, and 3 and the analytical approach parameters.

Algorithm	Profile 2 (4 A Parameters)	Profile 1 (10 A Parameters)	Profile 3 (Variable Frequency Parameters)
PSO-MGBO	2.1681	0.5573	0.0571
M-MGBO	1.5037	0.0562	0.0483
PSO-LCEO	1.5778	0.2005	0.05765
PSO	2.1251	0.3547	0.0570
Analytical		7.364942	

In the case of the Profile 1 parameters, it is evident that the M-MGBO algorithm exhibited superior performance compared with the other algorithms. The M-MGBO algorithm achieved an MSE of 0.0562, while the PSO-LCEO algorithm achieved an MSE of 0.2005, and the PSO-MGBO algorithm achieved an MSE of 0.5573. Nevertheless, the PSO-MGBO algorithm had the lowest performance in this particular test, as the PSO algorithm outperformed it, with an MSE value of 0.3547. Moreover, using Profile 3 parameters with the controlled current sequence profile, it was observed that the M-MGBO algorithm demonstrated superior performance in comparison to the other algorithms, achieving an MSE of 0.0483. On the other hand, the PSO, PSO-LCEO, and PSO-MGBO algorithms demonstrated a similar performance, with minimal deviation in MSE values.

5. Conclusions

The importance of energy storage for grid stability and power backup in both conventional and renewable systems is evident. Particularly, supercapacitors stand out for their excellence in high-power density applications, such as regenerative braking in electric vehicles. This research concentrates on the parameterization of the Zubieta model tailored for supercapacitors, employing a hybrid metaheuristic gradient-based optimization (MGBO) strategy. The Zubieta model, featuring three RC branches and a self-discharge branch, necessitates the identification of seven parameters. The investigation systematically

compared the performance of the modified MGBO (M-MGBO) against particle swarm optimization (PSO) and two variations—one integrating PSO with M-MGBO and the other introducing a local escaping operator (LCEO). The evaluation encompasses key metrics like convergence rate, accuracy, and processing/computational time. The results indicate that both hybrid PSO-MGBO and PSO-LCEO surpass conventional PSO, exhibiting enhancements of 51% and 94%, respectively. Remarkably, these variants proved to be comparable to M-MGBO. These methodologies effectively contribute to the accurate estimation of Zubieta model parameters. The study underscores the considerable potential of hybrid optimization strategies in advancing the precision and effectiveness of supercapacitor model parameterization.

The findings showed M-MGBO's adaptability to diverse operational contexts, with specific examples of a computational time of 31,320.07 s under a 4 A step input. Delving into the consistency and reliability of M-MGBO, the study scrutinized standard deviation values across 30 runs, revealing a remarkably stable pattern. This consistency is pivotal for practical applications, ensuring predictability and reliability in outcomes. The meticulous examination of the algorithm's performance underscored its reliability, with a computational time of approximately 28,610 s across 30 runs, adhering to the expected order of magnitude in resistance and capacitance values with respect to the analytical approach. As reported in the literature, when using metaheuristic algorithms the search for the optimal solution can become caught in local optima in most nonlinear systems with more than one minimum. Moreover, they do not always guarantee the global optimum solution and only provide nearly optimal solutions. This paper attempts to overcome these challenges by proposing some novel updates to the metaheuristic algorithms, and further improvements are left for future efforts.

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Article

An Experimental Analysis of Three-Phase Low-Voltage Power Factor Controllers Used in a Deforming Regime

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Abstract: In industry, to improve the power factor in low-voltage power substations, a power factor controller (the most used method) can be installed which connects capacitors banks (connected, or not, with coils) in the electrical installation. The most important parameters of power engineering are the power factors that indicate the efficiency of energy use. Currently, many non-linear consumers (more single-phase than three-phase) are used on low voltages. Harmonics (currents and/or voltages) are the most important dynamic component of power quality, affecting electrical equipment performance and also reducing power factors. The purpose of this analysis is to increase the displacement power factor and to decrease the total harmonic distortion (for the current) in the conditions where there are linear and non-linear consumers, where the power factor must be improved with capacitors banks. Relevant different consumers have been selected for both the industry and the home sector, as follows: inductive motors that are linear, inductive consumers, compact fluorescent lamps that are non-linear, and capacitive consumers. This analysis was carried out depending on the number of steps used for the power factor controller, the values of the capacitors banks, the AC reactor (connected in series with all consumers), and the LC shunt filters. For a slight deformation regime, a large number of capacitors banks with different values ensure a finer adjustment of the displacement factor. The maximum number of steps that regulators can command should not necessarily be used (the reliability of the installation decreases with the increase in the number of capacitors banks), but a reduced number of steps can be used, which can lead to higher values of displacement power factors. To improve the deforming regime and to increase the displacement power factor, the use of LC shunt filters, connected to a small number of steps, will also increase the displacement power factor (over 0.9) and decrease the total harmonic distortion (up to 7–10%) for the current. Weaker results were obtained with AC reactors connected to the power supply phases of consumers and, if a larger number of stages were used, to which LC shunt filters were connected, these filters become difficult to calibrate (resonances occur).

Keywords: power factor controller; displacement power factor; capacitors; coils; LC shunt filters

1. Introduction

Improved power factor means that, in a defined part of the installation, the necessary steps are taken to increase the power factor by locally delivering the necessary power to reduce current values and power flows through the upstream network with the same output power required. In this way, lines, generators, and transformers can be sized to lower power. The size of generators and transformers depends on the apparent power. At the same active power, the smaller the reactive power provided, the smaller the apparent power. Therefore, by improving the power factor, this equipment can be sized for less apparent power, but still provide the same active power. The devices that produce reactive power are synchronous generators, synchronous condensers, static VAR compensators, and capacitors banks [1,2].

The shunt compensation with capacitors banks of the reactive power can be used at load level, power substations, or transmission (at various voltage levels from low voltage to high voltage). Compensation must be provided as close to the consumption point as possible, in order to avoid the distribution of this power elsewhere in the network.

From a strictly technical point of view, even a suitable size installation can work properly in the case of low power factors. However, improving the power factors is a solution that can achieve technical and economic benefits and, in fact, managing low-cost installations will increase costs for power generation [3–6].

The main advantages of power factor correction can be summarized as follows: better use of electrical machines, better use of electrical lines, loss reduction, and lower voltage drops.

The correction of the power factor also allows us to obtain benefits for the size of the cables. In fact, at the same output power, the current decreases by increasing the power factor. This current reduction can enable the selection of conductors with lower cross section areas.

There are no general rules for all types of installations and, in theory, capacitors can be installed at any place, but the appropriate practical and economic feasibility must be assessed. The main methods of improving power factor correction depend on the position of the capacitors banks, as follows: the distribution power factor correction, the group power factor correction, the central power factor correction, the combined power factor correction, and the automatic power factor correction. According to the literature, improving the power factors in power engineering can lead to cost savings of 10 to 30% [7–11].

Power factor is related to the power quality. Currently, linear and non-linear consumers are connected to power substations (due to switching sources), which, together with the capacitors (used to improve the power factor), cause current distortion (power quality is affected). In practice, the voltage is not perfectly sinusoidal. The harmonic currents generated by the load will circulate in the circuit through the impedance of the source and all other branches in parallel. Consequently, harmonic voltages will appear on the impedance of the power supply and they will be present everywhere in the installation. The main equipment that produce harmonics are personal computers, LED lamps, fluorescent and gas expanding lamps, static converters, variable speed drives, welding machines, and arc and induction furnaces [12].

The presence of harmonics (currents or/and voltages) in electrical networks may cause equipment failure, such as the overloading of neutral conductors, increasing transformer losses, and causing torque disturbances in motors. More important problems occur when distortions reach high values and the danger of resonances between power factor correction systems and electrical networks are present [13,14].

In three-phase systems, the capacitors bank is composed of three capacitors with the same capacity and can have wye or delta connections. The capacitors banks can be combined with the inductors to limit the harmonic on the network and reduce inrush currents. In fact, the combination capacitor–inductor forms a harmonic filter. In this way, the component inductors–capacitors provide a very low reaction in correspondence to the elimination of harmonics that circulate in the component without affecting the entire network. Also, to limit the distortion of the current, AC reactors can be used that are connected in series with the electrical consumers [15–19].

Normally, two options are used to solve power quality problems (non-sinusoidal current), as follows: passive power factor correction (to improve power factor by filtering harmonics with passive filters) and active power factor correction (using a switching converter to modulate distorted waves and shape them into sine waves). For ordinary electrical consumers who have a switching source, a power factor correction circuit is added to the energy supply circuit to increase the power factor and/or reduce the harmonics. There are three power factor correction techniques: passive with reactors, active that controls a current at high frequency using a switching device, and partial switching with

switching devices to regulate currents several times per mains cycle and whose applications are limited [20,21].

Power passive filters can be used to improve power factors in electrical systems. It is difficult to use passive filters to increase power factor and to decrease deformation (currents) because they must also take into account consumer specificities (linear or non-linear, capacitive or inductive characteristics). Reactive power and negative sequence currents generated by inductive unbalanced loads not only increase line losses, but also cause relay protection devices to fail due to negative sequence components of the power grid, threatening the safe operation of the power system. Reactive power compensation and unbalance suppression were achieved in three-phase wye connectors, with complex electronic structures for inductive unbalanced electric consumers [7,14,22,23].

Real-time control of the reactive power is currently the most efficient solution to avoid voltage fluctuations and maintain the standard limits. For this purpose, capacitors banks, reactors, synchronous generators, synchronous condensers, and FACTS (Flexible Alternating Current Transmission Systems) installations are used. Types of FACTS installations are SVC (Static VAR Compensator), STATCOM (Static Synchronous Compensator), SSCR (Switched Shunt-Capacitor and Reactor), UPFC (Unified Power Flow Controller) and D-STATCOM (Distribution STATCOM) that provide control of reactive power to obtain voltage stabilization and ensure mitigation solutions for voltage changes, swelling, interruptions, and flashes. These solutions are usually used for medium and high voltages. For the power substation, the most common solution for improving the power factor is the use of capacitors banks controlled by a power factor controller (which makes the adjustment according to the displacement power factor) mounted in the point of common coupling [24–26].

This paper is divided into five sections. The second section is about materials and methods and presents theoretical aspects about electrical parameters used in the deforming regime and the power factor controller with capacitors banks used for improving power factors. The third section presents the experimental results (20 groups of experiments) with capacitors (without coils connected in series), AC reactors, and LC shunt filters. The fourth section is a detailed discussion about the use of power factor controllers in different conditions. Conclusions on the use of the power factor controller are presented in the last section.

2. Materials and Methods

2.1. Theoretical Aspects

The RMS values for voltage (U) and current (I) in a deforming mode are calculated with [14]:

$$U = \sqrt{U_0^2 + \sum_{k=1}^{50} U_k^2} \quad (1)$$

$$I = \sqrt{I_0^2 + \sum_{k=1}^{50} I_k^2} \quad (2)$$

where U_0 and I_0 are the continuous components of voltage and current and U_k and I_k are fundamental ($k = 1$) and higher harmonics of voltage and current.

In periodic non-sinusoidal mode, the following powers can be defined [12,26]:

- apparent power:

$$S = U \cdot I, \quad (3)$$

- active power:

$$P = U_0 \cdot I_0 + \sum_{k=1}^{50} U_k \cdot I_k \cdot \cos \varphi_k \quad (4)$$

where φ_k represents the phase shift between the k component of the voltage and the k component of the current.

- reactive power:

$$Q = \sum_{k=1}^{50} U_k \cdot I_k \cdot \sin \varphi_k \quad (5)$$

- deforming power:

$$D = \sqrt{S^2 - P^2 - Q^2} \quad (6)$$

Based on the powers S , P , Q , and D , the power factor is defined in non-sinusoidal mode:

$$PF = \frac{P}{S} = \frac{P}{\sqrt{P^2 + Q^2 + D^2}} \quad (7)$$

The displacement power factor is calculated with:

$$DPF = \cos \varphi_1 \quad (8)$$

where φ_1 represents the phase shift between the voltage and current fundamentals (50 Hz).

The total harmonic distortion factor (THD, Total Harmonic Distortion) is a ratio between the deforming residue and the RMS value of the fundamental:

$$THD = \frac{\sqrt{\sum_{k=2}^{50} G_k^2}}{G_1} \quad (9)$$

where G_k is the RMS value for harmonics of rank k and G_1 is the RMS value of the fundamental harmonic. THD can be calculated for both current and voltage. As can be seen, the total harmonic distortion factor does not take into account the phases of the harmonics, only their RMS values.

In order to decrease the reactive energy, all electrical energy consumers are obliged to operate at a power factor higher than the neutral power factor. In three-phase alternating voltage, the capacitors can be connected to the network in wye or delta connections (Figure 1). It is denoted by C_y —the capacitor on a branch of the capacitors bank in wye connection; C_d —the capacitor on a branch of the bank at the delta connection; and Q_{CB} —total reactive power for the capacitors bank.

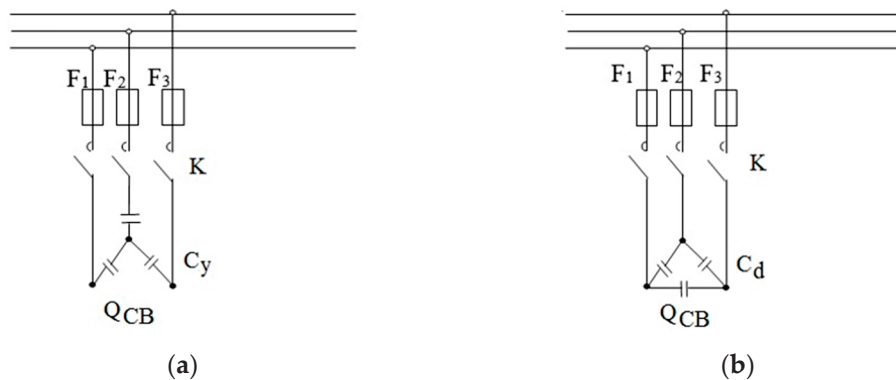


Figure 1. (a) The wye connection of a capacitors bank. (b) The delta connection of a capacitors bank.

For a consumer balanced on the three-phases powered from a symmetrical three-phase voltage, if Q_{CB} is the total reactive power of the three-phase capacitors bank [7]:

$$Q_{CB} = P \cdot (\operatorname{tg} \varphi_1 - \operatorname{tg} \varphi_2) \quad (10)$$

where P is the active power of the consumer (on all phases), φ_1 is the initial phase shift between voltage and current, and φ_2 is the final phase shift (it is usually imposed by the neutral power factor) between voltage and current after the introduction of the capacitors bank.

Depending on the phase voltage, the capacitor for the wye connection is determined using:

$$C_y = \frac{P}{2 \cdot \pi \cdot f \cdot U_1^2} \cdot \left(\frac{\sqrt{1 - \cos^2 \varphi_1}}{\cos \varphi_1} - \frac{\sqrt{1 - \cos^2 \varphi_2}}{\cos \varphi_2} \right) \quad (11)$$

The capacitor for the delta connection is determined using:

$$C_d = \frac{P}{3 \cdot 2 \cdot \pi \cdot f \cdot U_1^2} \cdot \left(\frac{\sqrt{1 - \cos^2 \varphi_1}}{\cos \varphi_1} - \frac{\sqrt{1 - \cos^2 \varphi_2}}{\cos \varphi_2} \right) \quad (12)$$

For the same reactive power of the capacitor bank:

$$C_d = \frac{C_y}{3} \quad (13)$$

So, at the same reactive power of the capacitors bank, the capacitors in the delta connection are three times less than the capacitors in the wye connection, but the capacitors in the delta connection must work at the line voltage, compared to the capacitors in the wye connection, which must work at the phase voltage.

To reduce the deforming regime of the current determined by connecting the capacitors to real supply voltages, which differ from the sinusoidal form, LC filters can be used, whose frequency is calculated with:

$$f_c = \frac{1}{2 \cdot \pi \cdot \sqrt{L \cdot C}} \quad (14)$$

2.2. Classical Power Factor Controller

The power factor improvement with classical power factor controllers (PFCs; also known as reactive power controllers) is known (Figure 2) [4]. Normally, using a single current transformer, CT measures the line current and the PFC measures the phase voltage, and displacement power factor is calculated and, depending on the set value of displacement power factor (usually the neutral power factor), capacitors banks (CB1, CB2, CB3, CB4, CB5, and CB6) are timed to switch on or switch off through electromagnetic relays (K1, K2, K3, K4, K5, and K6). Choosing the number of banks steps depends on the fineness of the displacement power factor adjustment and the values of banks, depending on the reactive power. PFCs monitor the presence of harmonics in the network, incorrect operation of capacitors banks, and the temperature level of the space where the capacitors banks (CBs) are located.

The disadvantage of this installation is that when connected to a capacitors bank, the high values of currents peaks are generated by the capacitors bank (depending on the values of capacitors, the remanent voltages on the capacitors, and the initial voltage applied on capacitors). With these types of PFCs, a fine adjustment of the displacement power factor cannot be achieved (the change is in steps). Using the classical PFC, premature wear of the electromagnetic relay contacts occurs (reducing their lifetime).

When connecting capacitors banks in an installation with inductive consumers, the power factor is improved, but the current will be distorted. To improve the waveforms of the currents, AC reactors can be used that are connected in series with the consumers (Figure 3). Basically, the higher their value, the closer they are obtained to the sinusoidal currents. It is obvious that their value cannot be any higher because there will be a significant voltage drop on the AC reactors [14].

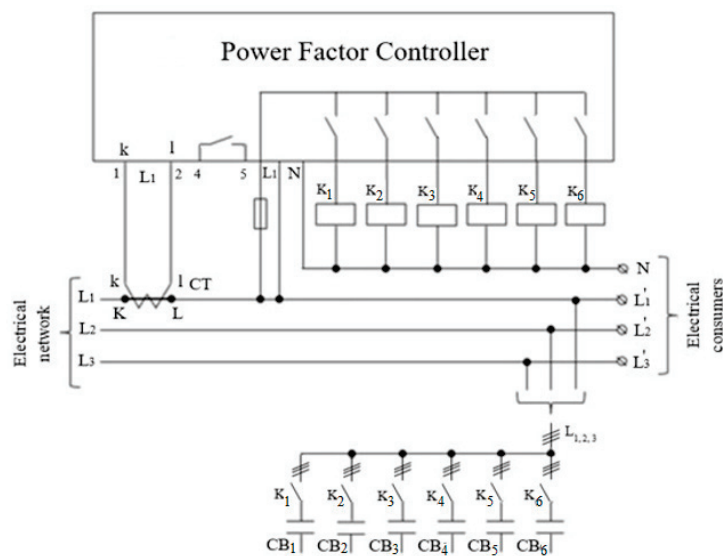


Figure 2. Block diagram of a power factor controller in low-voltage three-phase installations with capacitors banks connected through electromagnetic contactors.

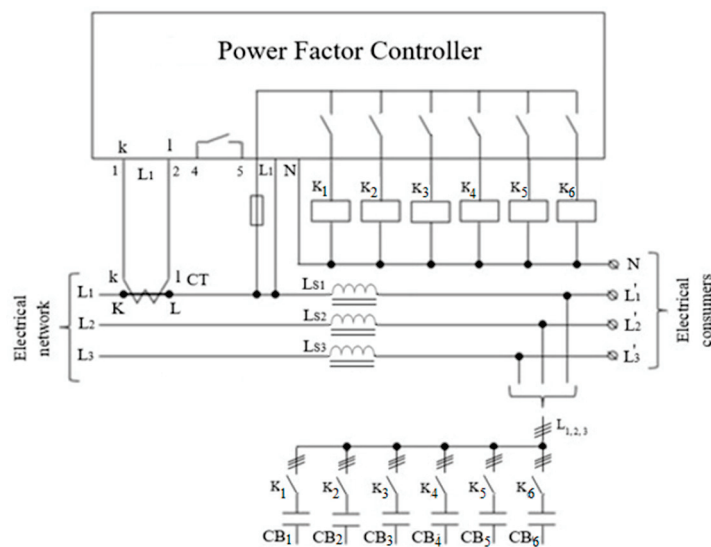


Figure 3. Block diagram of a power factor controller in low-voltage three-phase installations with capacitors banks with AC reactors.

In power engineering, a synchronous condenser is a DC-driven synchronous motor, whose shaft is not connected to anything but rotates freely. Its objective is to adjust the conditions (adjustment of power factor or voltage) of the electric network. Its field is controlled by a voltage regulator to generate or absorb reactive power, as needed, to adjust the voltage of the network or to improve the power factor. The installation and operation of the synchronous condenser is identical to that of the large electric motor and generator (some generators are designed to work as synchronized condensers with the main motor disconnected). However, due to high installation and maintenance costs, they are not suitable for most low-voltage applications (e.g., power substations). However, synchronous condensers have higher energy losses than static capacitors banks. Most synchronous condensers connected to electrical networks have a rating of between 20 and 200 MVAR, but are not used in ordinary low-voltage substations (usually, up to 1600 KVA maximum power), and many of them are hydrogen-cooled. An advantage of using synchronous condensers is that a fine adjustment of the power factor can be achieved,

which is very difficult to achieve with installations with condenser banks (many capacitors banks are controlled by power factor controllers) [22].

The mean cost for a synchronous condenser is between USD $(10 \div 40)$ /kVAR with high maintenance costs, and for an installation with capacitors banks is between USD $(10 \div 30)$ /kVAR, but with very low maintenance costs [24]. For example, for a reactive power of 5 MVAR, the costs (capital cost, preventive maintenance cost, and voltage regulator maintenance) when using an installation with capacitors banks, the estimated values are USD 34,600, and for an installation with a synchronous condenser the estimated costs are USD 56,800 (almost double), under the conditions in which the same annual financial benefits are obtained. However, it should be mentioned that the average life span for an installation with capacitors banks is 10 years, while an installation with a synchronous condenser is 20 years [24].

On medium voltages, for high-power electrical industrial consumers, it is beneficial to use a synchronous condenser that can adjust both the voltage level and the power factor.

3. Results

The purpose of the experimental measurements is to analyze the operation of the PFC with different numbers of capacitors banks (from 2 to 6 stages), when using the same values or different values for the capacitors banks, and when using the capacitors banks connected to coils (each terminal of the capacitors bank is in series with a coil).

To experimentally analyze the PFC (Roederstein ESTAmat RPR type, Vishay), two induction motors (inductive character, linear) and two compact fluorescent lamps (CFLs, capacitive character, non-linear consumers) were used (Figure 4) which connected to the installation in Figure 2. The characteristics of consumers are as follows:

- M1, three-phase induction motor with the following data (ASI 90L-24-2 type): 2.2 kW; 2780 rpm; 400 V; 4.95 A; PF = 0.855;
- M2, three-phase induction motor with the following data (N 80 L type): 750 W; 1450 rpm; 400 V; 2.13 A; PF = 0.72;
- Compact fluorescent lamps, connected in parallel on the phase where CT (15/5 A, SM 5VA, Metra type) is also connected with the following data: CFL1: 220 V, 50 Hz, 85 W, 6400 K; CFL2: 220–240 V, 50/60 Hz, 120 W, 580 mA, 4000 K.

Induction motors work without mechanical loads to have a power factor as low as possible and the two CFLs work in parallel. Consumers can be connected manually through electrical separators (S1, S2, S3—Figure 4). The measurements were made with a power quality analyzer (CA 8334B type, Chauvin-Arnoux).

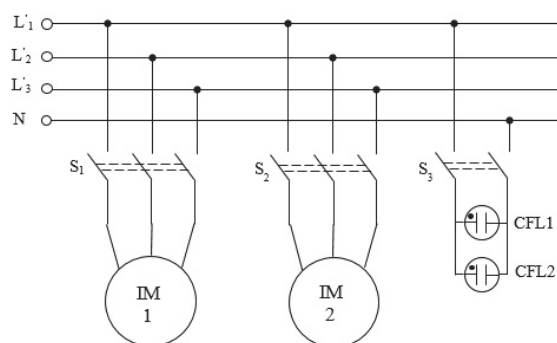
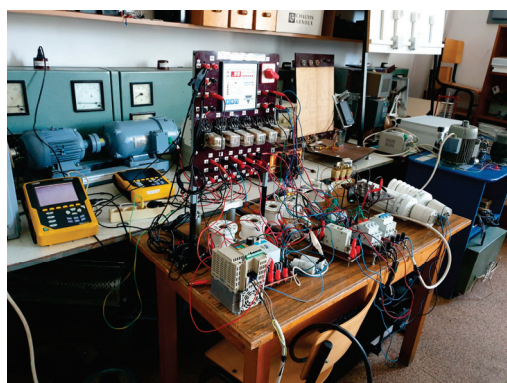


Figure 4. The electrical consumers used in the experiments depicted in the diagram in Figure 2 [27].

The main electrical characteristics of the consumers used in the experiments are listed in Table 1. In all the tables, the quantities have the following meanings: P—active power; Q—reactive power; S—apparent power; PF—power factor; DPF—displacement power factor; THDc—total harmonic distortion for current. In all the tables, for Q, PF, and DPF, for the numerical values, i represents the inductive character and c represents the capacitive character. Figure 5 shows a view of the experimental setup used in the experiments.

Table 1. The main electrical characteristics of the consumers used in the experiments, measured on the L1 phase.

Type	P (W)	Q (W)	S (VA)	PF (-)	DPF (-)	THDc (%)
M1	75.7	573.1 i	578.2	0.13 i	0.13 i	7.8
M2	44.8	296 i	299.3	0.15 i	0.15 i	3.5
CFLs	183.4	118 c	213.3	0.84 c	0.986 c	58.9

**Figure 5.** Experimental setup used for experiments.

In the following tables i represents the inductive character, and c represents the capacitive character of the loads.

Although it also measures the deforming regime of voltage and current, PFC ensures the regulation of capacitors banks according to the displacement power factor (DPF). It would be impossible for the PFC to improve the PF in the installation by introducing CBs, under the conditions that, by connecting the CBs, it determines the pronounced deformation of the current and the PF decreases (all the more as the deforming regime increases). From a practical point of view, it is not possible to ensure a fine adjustment of the DPF in the electrical installation (due to the stages to which the CBs are connected), but it is possible to ensure an adjustment in stages (depending on the combination of CBs).

During the experiments, all the capacitors used were new (they have very low dielectric loss) and were working at 450 V, 50 Hz. The neutral power factor for low voltages is 0.9, value adjusted at PFC. In experiments, wye connection for CBs were used.

With Equation (11) and with the data from Table 1, the (theoretical) capacities of the capacitors (on a branch of the wye-connected bank) that must be inserted in parallel with the inductive consumers were determined as follows:

- for $U_f = 220$ V, $f = 50$ Hz: for electric motor M1: $C_{M1} = 35.57$ μ F; for the electric motor M2: $C_{M2} = 18.01$ μ F, and when the two motors are operating, the total capacity $C_M = 53.58$ μ F;
- for $U_f = 230$ V, $f = 50$ Hz: for electric motor M1: $C_{M1} = 32.55$ μ F; for the electric motor M2: $C_{M2} = 16.47$ μ F, and when the two motors are operating, the total capacity $C_M = 49.02$ μ F.

Using Equation (11), the equivalent capacity was determined for CFLs:

- for $U_f = 220$ V, $f = 50$ Hz: $C_{CFLs} = 7.76$ μ F;
- for $U_f = 230$ V, $f = 50$ Hz: $C_{CFLs} = 7.1$ μ F.

Table 2 shows some examples of combinations of electrical consumers (M1, M2, and/or CFLs) together with capacitors (on phase L1) that ensure a certain DPF.

Table 2. Combinations of consumers used in experiments and different capacitors (a branch of the wye-connected banks) that ensure a certain DPF, without PFC, determined on the L1 phase.

Type	Capacitors (μF)	DPF (-)
M1 + M2	20 + 30	0.76 i
M1 + M2	10 + 20 + 30	0.78 c
M1 + M2	4 + 20 + 30	0.93 i
M1 + M2 + CFLs	30	0.6 i
M1 + M2 + CFLs	10 + 30	0.78 i
M1 + M2 + CFLs	4 + 10 + 30	0.86 i
M1 + M2 + CFLs	$2 \times 4 + 10 + 30$	0.94 i
M1 + M2 + CFLs	20 + 30	0.99 i

The power factor controller (Roederstein ESTAmat RPR type, Vishay; software version 4.1.2) used in the experiments is a typical regulator (measurement of current on a single phase through a CT; with identification of CBs values; DPF adjustment using a single CT; timed switch on/switch off of CBs; identification of the pronounced deforming regime in the case of connecting a certain CB; possibilities to adjust the DPF according to a set value; multiple display), which is commonly used in power substations (where inductive consumers are connected, usually in industry) for improving the power factor in low-voltage electrical installations. Depending on the DPF value on phase L1 (the phase difference between the fundamental voltage and the fundamental current), the regulator timed switch on or switch off of one or several CBs. From a practical point of view, the PFC can have between 6 and 14 stages and the CBs can be identical or different (on one CB there are identical capacitors in wye or delta connection). An example of the worsening of the deforming regime when the displacement power factor is improved is presented in Figure 6.

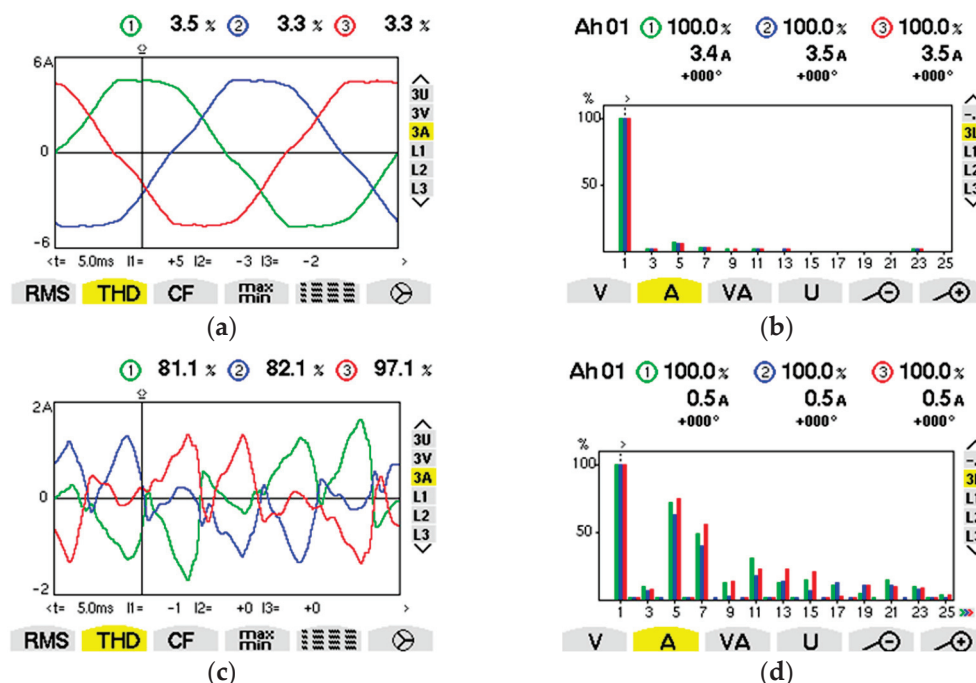


Figure 6. (a) Currents through electric motors M1 + M2; (b) the harmonic spectra of the currents through the electric motors M1 + M2; (c) currents through electric motors M1 + M2 and a capacitors bank (in wye connection) with $C = 54 \mu\text{F}$ (on one branch); and (d) the harmonic spectra of the currents through the electric motors M1 + M2 and a capacitors bank (in wye connection) with $C = 54 \mu\text{F}$ (on one branch).

When supplying voltage to the PFC, a calibration (value, symmetry) of the CBs on each stage is performed. Obviously, if there is no CB connected to a certain step, it is not taken into account in the adjustment. The identification of the connected CBs is carried out three times, in an interval of approx. 3 min.

In order to carry out the calibration, it is necessary to connect an electrical consumer in the installation. During several experiments, calibrations were made with linear consumers (e.g., induction motors) or with non-linear consumers (e.g., CFLs). It was found that the capacitors banks are correctly identified, regardless of the type of electrical consumer (linear or non-linear) used for calibration (at the beginning of the PFC's operation).

The transformation ratio of the current transformer (CT) must be chosen in such a way (depending on the consumers and the CBs) that the current in the secondary CT is not less than 150 mA and does not exceed 5 A. A CT 15A/5A (Metra) was used for the experiments.

Several experiments were carried out to identify the operation of the PFC with different numbers of CBs (identical or not), when using coils in series per phase (Figure 3) and when using coils in series with capacitors banks (LC shunt filters, Figure 7). The main quantities monitored during the experiments were DPF (quantity according to which the adjustment is made by the PFC) and total harmonic distortion for the current (THDc).

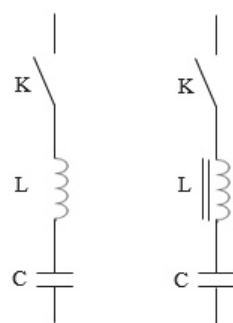


Figure 7. LC shunt filters without core and with core used in some experiments.

In the following, the experiments refer to the capacitors on a branch from the bank in wye connection. In the vast majority of the cases analyzed, capacitors were used to compensate the reactive power of the consumers (in the worst case).

To make it easier to understand the results of the experiments, measurements are presented only for phase L1 for the current and for the voltage between L1 and null (N). A current transformer (CT, Figure 2) is mounted on the L1 phase, which has a secondary connection to the PFC, where the current is measured, which ensures the connection of the capacitors banks (wye connections).

3.1. Connecting Consumers without PFC and without Capacitors Banks

A first experiment was carried out without PFC and without CBs, only with electrical consumers connected in a certain order (Table 3).

Table 3. Experimental measurements for Section 3.1.

Consumer Status	DPF (-)	I (A)	THDc (%)	Figures
Connection M1	0.13 i	2.6	8.2	Figure 8
Connection M2	0.12 i	3.9	7	Figure 9
Connection CFLs	0.32 c	3.94	17.3	Figure 10
Disconnection M2	0.44 i	2.67	23.3	Figure 11

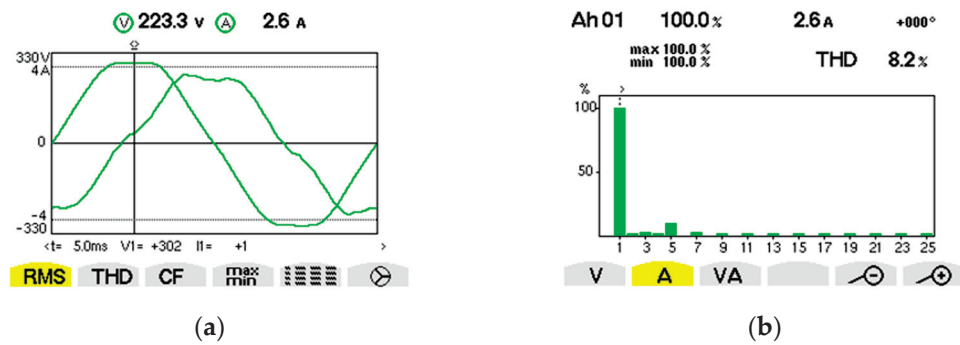


Figure 8. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motor M1.

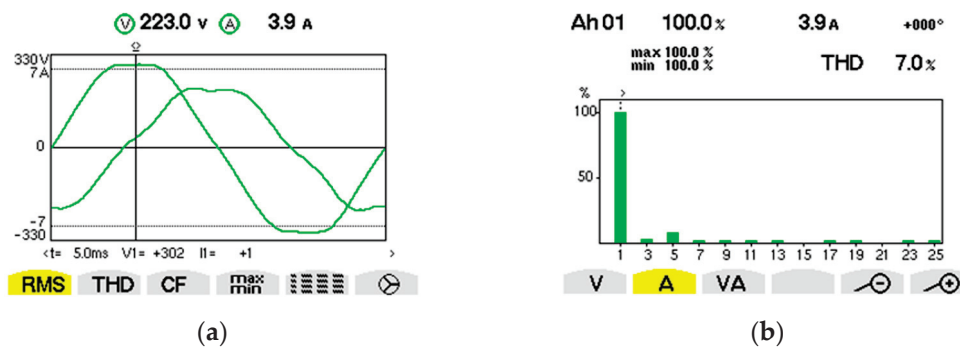


Figure 9. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1 and M2.

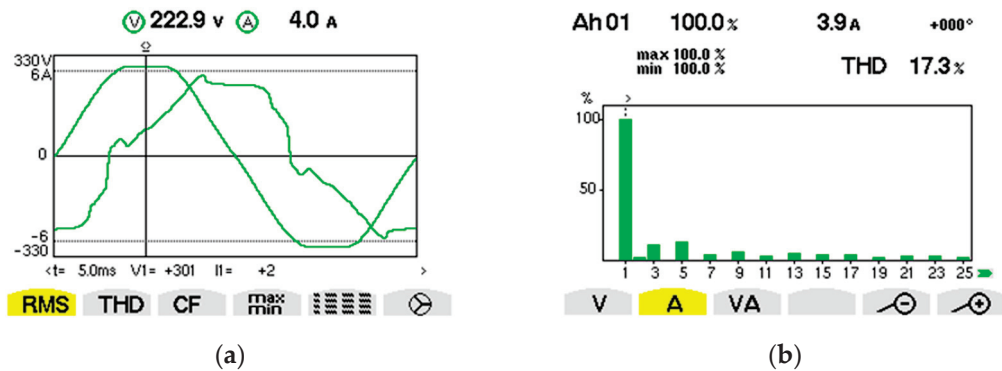


Figure 10. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1, M2, and compact fluorescent lamps.

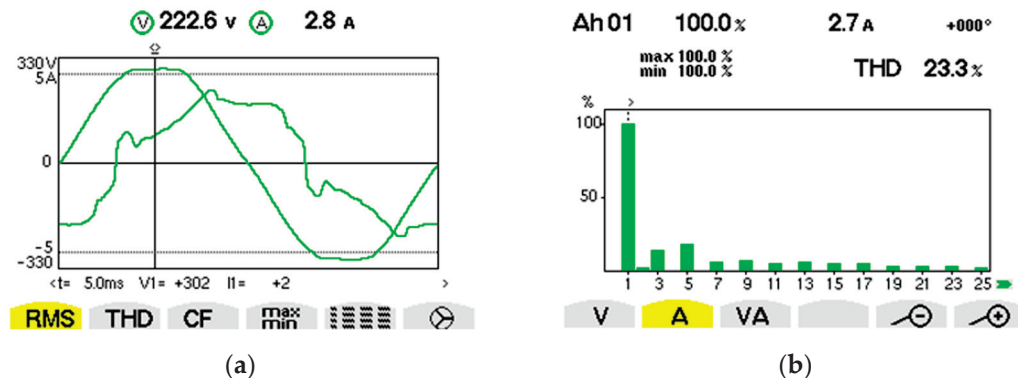


Figure 11. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motor M1 and compact fluorescent lamps.

Induction motors, being inductive and linear consumers, will cause a large phase shift (DPF has very small values) between voltage and current (current follows the voltage), and THDc has values of the order of percentages (current is more distorted than voltage).

If CFLs are introduced in parallel with IMs, which have a capacitive and non-linear character, they will cause a slightly smaller phase shift between voltage and current (DPF has small values), and the current will be more distorted (compared to the previous situation).

Next, when the value of a capacitor is specified, it refers to the capacitor on a branch of a capacitors bank in wye connection.

3.2. Single-Step PFC, $C1 = 54 \mu F$

In these experiments, a single stage was connected with a capacitor (on the branch) of $54 \mu F$ (Table 4).

Table 4. Experimental measurements for Section 3.2.

Consumer Status	Connected Steps	DPF(-)	I (A)	THDc (%)	Figures
Connection M1	-	0.16 i	2.64	9.3	Figure 12
Connection M2	1	0.91 c	0.68	101.1	Figure 13
Connection CFLs	1	1	1.4	56.1	Figure 14
Disconnection M2	-	0.44 i	2.71	22.5	

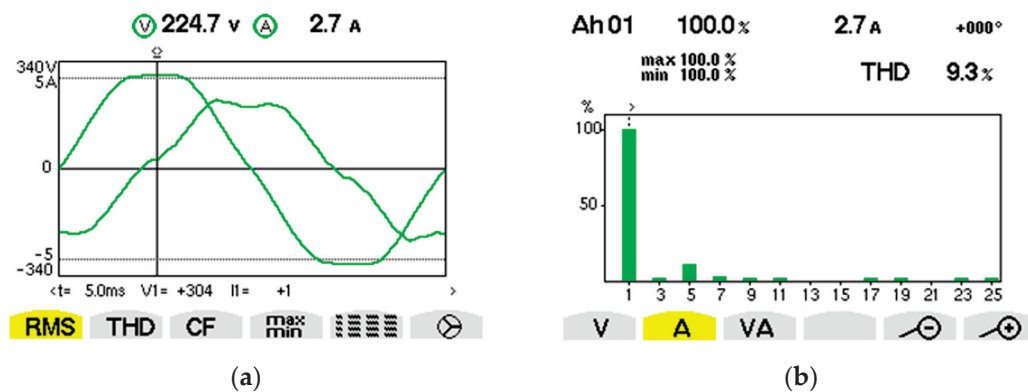


Figure 12. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motor M1.

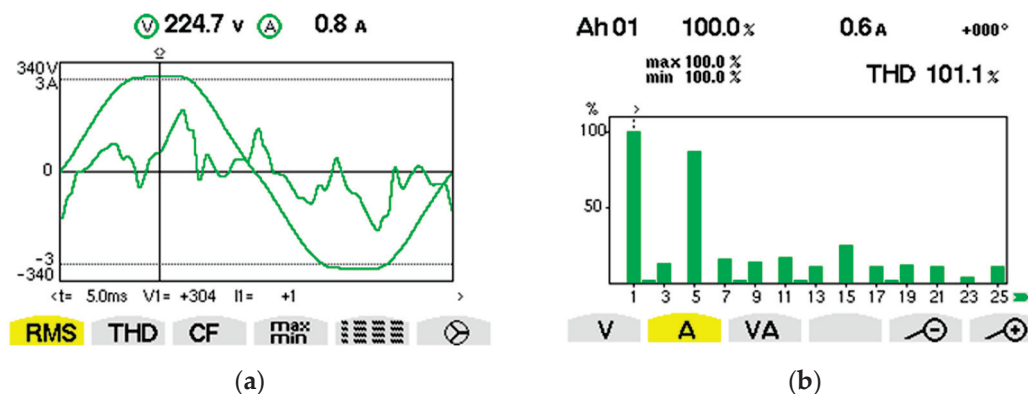


Figure 13. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1 and M2.

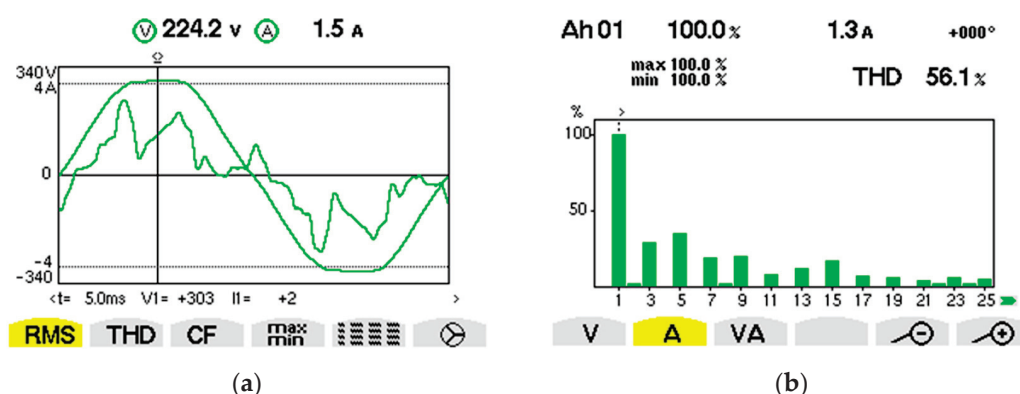


Figure 14. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1, M2, and compact fluorescent lamps.

The use of a single CB leads to the impossibility of adjusting the DPF in the installation where the DPF is modified within wide limits. A high-capacity CB also introduces a greater distortion of the current (high THDc) in the situation where only inductive consumers are connected. If non-linear and capacitive consumers are connected, THDc decreases.

If the capacitive reactive power is too high, the bank is not connected into the circuit.

If the two induction motors are used, the PFC connects the capacitors bank. The capacity, being large, will determine a large DPF, the current drops considerably, but the current will be highly distorted (high THDc).

By introducing CFLs, the DPF increases and the THDc decreases almost by half (and due to the diversity affect that can determine this effect).

3.3. Single-Step PFC, $C1 = 54 \mu\text{F}$ and AC Reactor of 9 mH

In the following experiments, a phase coil was used, inserted with consumers and CB (Table 5). A single capacitors bank was used.

Table 5. Experimental measurements for Section 3.3.

Consumer Status	Connected Steps	DPF(-)	I (A)	THDc (%)	Figures
Connection M1	1	0.34 c	1.24	52.3	Figure 15
Connection M2	1	0.95 i	0.66	113.5	Figure 16
Connection CFLs	1	1	1.41	62.6	Figure 17
Disconnection M2	1	0.68 c	1.79	44.5	Figure 18

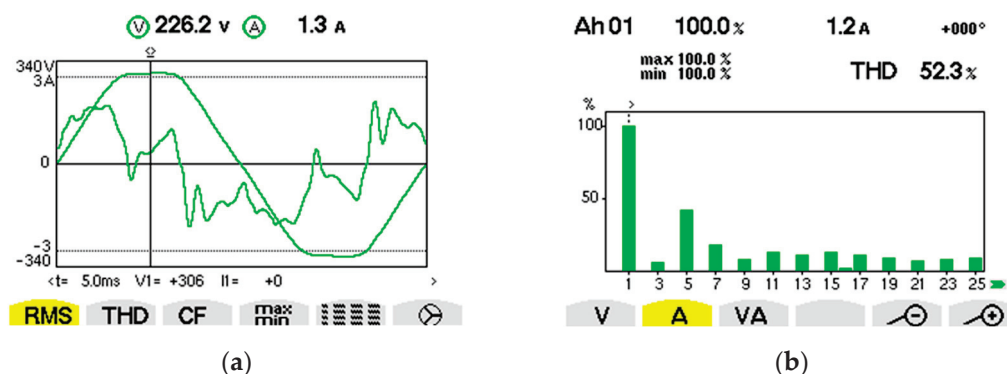


Figure 15. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motor M1.

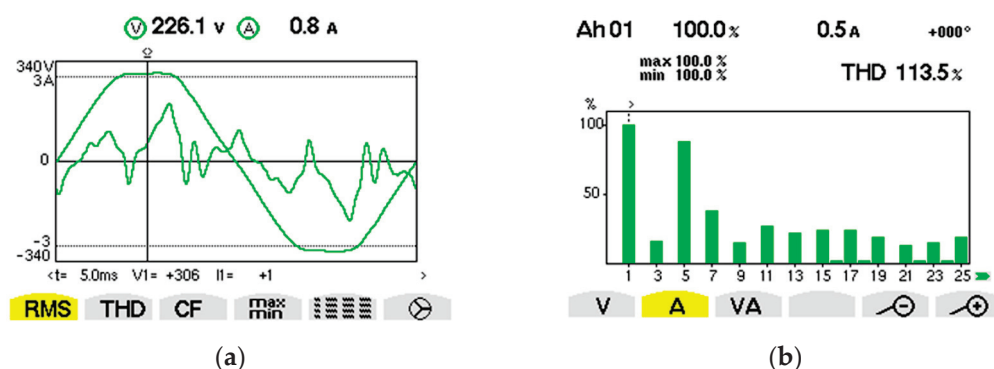


Figure 16. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1 and M2.

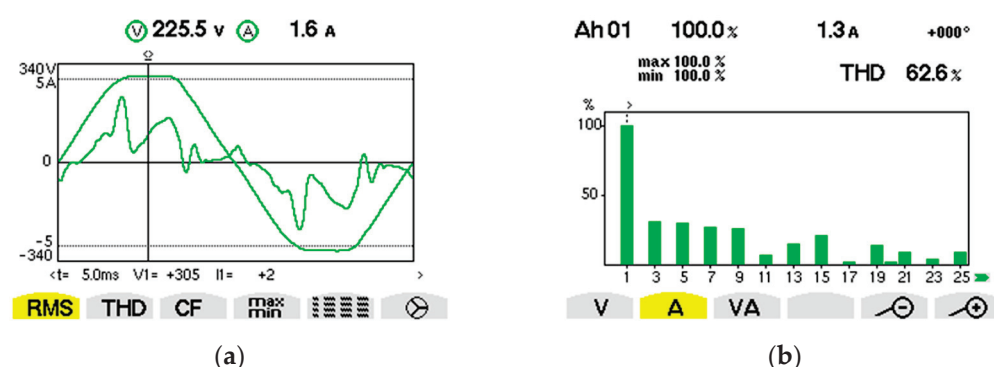


Figure 17. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1, M2, and compact fluorescent lamps.

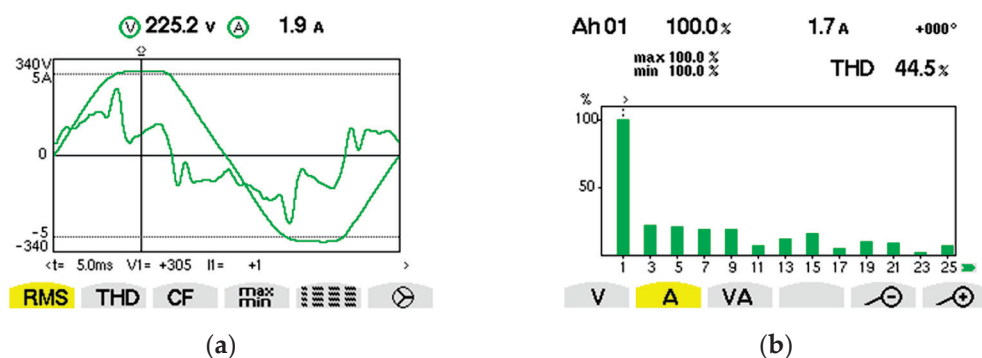


Figure 18. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motor M1 and compact fluorescent lamps.

If it is connected into circuit M1, by inserting CB (using the PFC), this having high capacitors, the character of the consumers will be capacitive and the current will be distorted (THDc = 52.3%).

If M2 is also connected, the current becomes lower, the DPF will be high (inductive character), but the current will be very distorted.

If CFLs are also connected, the DPF will have a maximum value, and the current will be very distorted (THDc = 62.6%).

By disconnecting M2, the remaining consumers will have a capacitive character and the current will be distorted. When using some coils inserted with the consumers, the absorbed currents will be more distorted (higher THDc) compared to the experiments from Section 3.2. It is mentioned that the usual adjustment of the DPF is carried out in the inductive range (e.g., if the DPF has been set to 0.9, the PFC adjusts the DPF in the inductive range 0.9–0.99).

3.4. Single-Step PFC (Stage 1), LC Shunt Filter LC, $C = 54 \mu F$, $L = 9 mH$

The PFC does not immediately connect (case a, Table 6) the LC shunt filter ($f = 228.3 \text{ Hz}$) but does so after a period of time (case b in Table 6). If a filter is used and connected in parallel with the consumers, it is found that THDc increases and DPF decreases.

Table 6. Experimental measurements for Section 3.4.

Consumer Status	Connected Steps	DPF(-)	I (A)	THDc (%)
Connection M1 + M2 a	-	0.12 i	4.14	6.4
Connection M1 + M2 b	1	0.82 i	0.75	116.1

3.5. PFC with Two Identical Steps, $C1 = 20 \mu F$, $C2 = 20 \mu F$

If two stages are used, with identical CBs, a finer adjustment of the DPF can be achieved (it has both inductive and capacitive values) compared to the previous situation (Table 7).

Table 7. Experimental measurements for Section 3.5.

Consumer Status	Connected Steps	DPF(-)	I (A)	THDc (%)
Connection M1	1,2	0.92 c	0.46	110.8
Connection M2	1,2	0.38 i	1.2	30.2
Connection CFLs	1,2	0.76 i	1.62	39.5
Disconnection M2	1,2	0.97 c	1.25	50.7

By using two identical CBs, finer adjustment of the DP is not obtained and the current will be equally distorted (high THDc). Larger capacities will lead to greater current distortion. However, in this situation there is an under-dimensioning of the CBs.

3.6. PFC with Two Different Steps, $C1 = 20 \mu F$, $C2 = 40 \mu F$

If two different steps are used (Table 8), an adjustment of the DPF can be made (it has values in the inductive as well as capacitive character), but THDc has higher values compared to Section 3.5.

Table 8. Experimental measurements for Section 3.6.

Consumer Status	Connected Steps	DPF(-)	I (A)	THDc (%)
Connection M1	2	0.85 c	0.47	100.8
Connection M2	1,2	0.86 c	0.7	117.1
Connection CFLs	1,2	0.93 c	1.43	59.5
Disconnection M2	1	0.72 i	1.51	36.1

3.7. PFC with Three Identical Steps, $C1 = 20 \mu F$, $C2 = 20 \mu F$, $C3 = 20 \mu F$

When three identical CBs are used (Table 9), a DPF of high values is not obtained (Table 9). Since the CBs are identical, a finer adjustment cannot be achieved compared to Section 3.6.

Table 9. Experimental measurements for Section 3.7.

Consumer Status	Connected Steps	DPF(-)	I (A)	THDc (%)	Figure
Connection M1	1,2	0.7 i	0.54	80.6	Figure 19
Connection M2	1,2	0.49 i	0.94	37.9	Figure 20
Connection CFLs	1,2	0.86 c	1.46	45.6	Figure 21
Disconnection M2	1,2	0.91 c	1.29	51.9	Figure 22

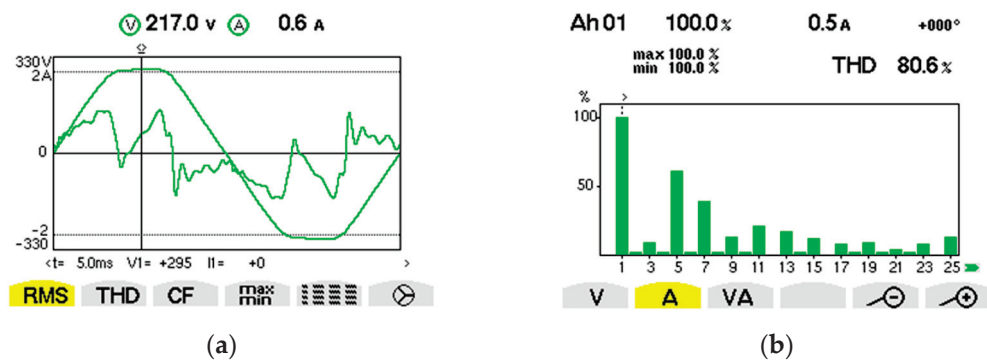


Figure 19. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motor M1.

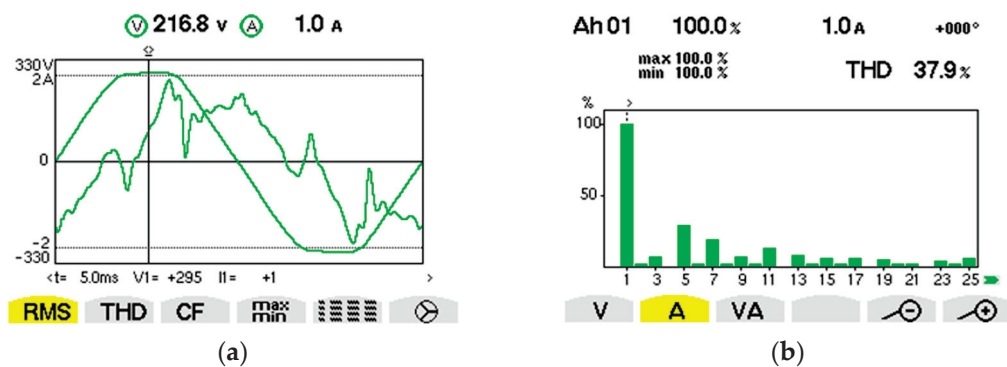


Figure 20. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1 and M2.

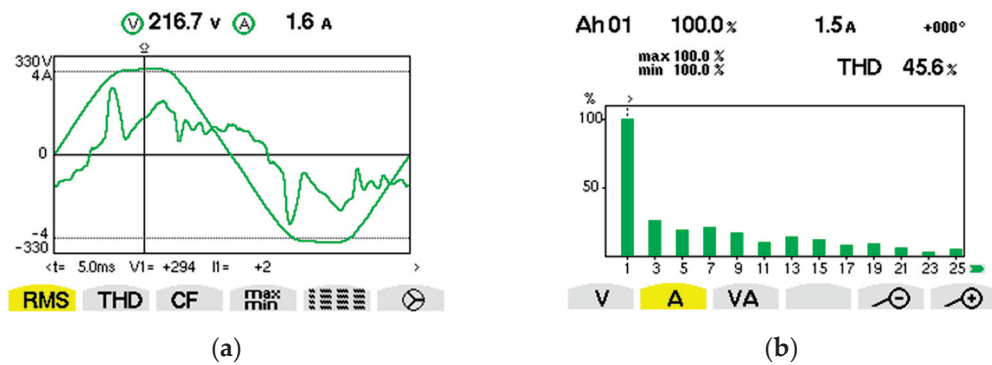


Figure 21. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1, M2, and compact fluorescent lamps.

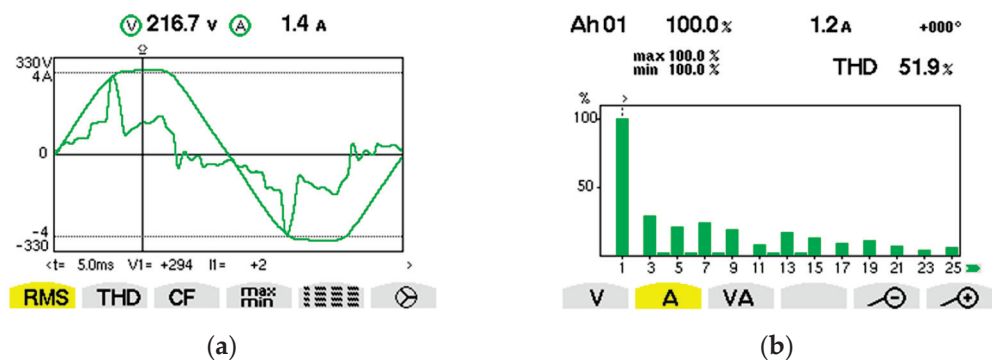


Figure 22. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motor M1 and compact fluorescent lamps.

Surprisingly, by using three identical CBs, the DPF has lower values than in Section 3.6, but the current will be less distorted.

In all situations, the current is distorted, having large harmonics, the largest being 3, 5, 7, 9, 13, etc. When using only electric motors, the third harmonic is much reduced.

3.8. PFC with Three Different Steps, $C1 = 10 \mu F$, $C2 = 20 \mu F$, $C3 = 30 \mu F$

When CBs of different values are used (but which have the same reactive power as in Section 3.7), the PFC works better, the DPF has an inductive character only (Table 10). When CBs of large values are introduced and electric motors are used, then the current is distorted the most (high THDc).

Table 10. Experimental measurements for Section 3.8.

Consumer Status	Connected Steps	DPF (-)	I (A)	THDc (%)	Figure
Connection M1	1,2	0.59 i	0.5	54.5	Figure 23
Connection M2	2,3	0.77 i	0.64	90.3	Figure 24
Connection CFLs	2,3	0.99 i	1.34	59.4	Figure 25
Disconnection M2	3	0.93 i	1.15	51	Figure 26

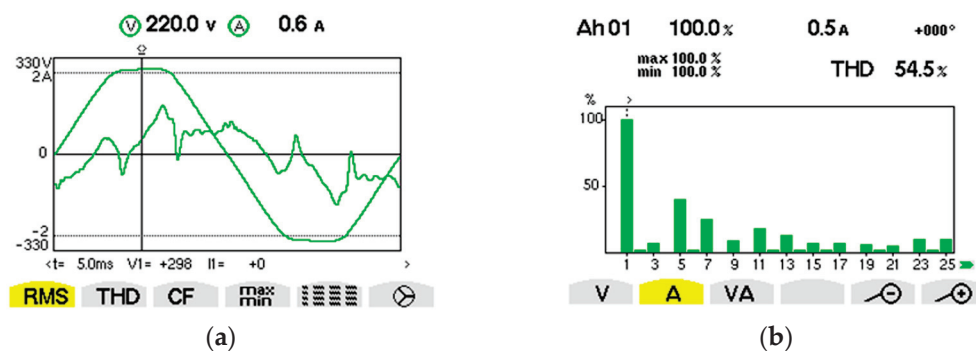


Figure 23. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motor M1.

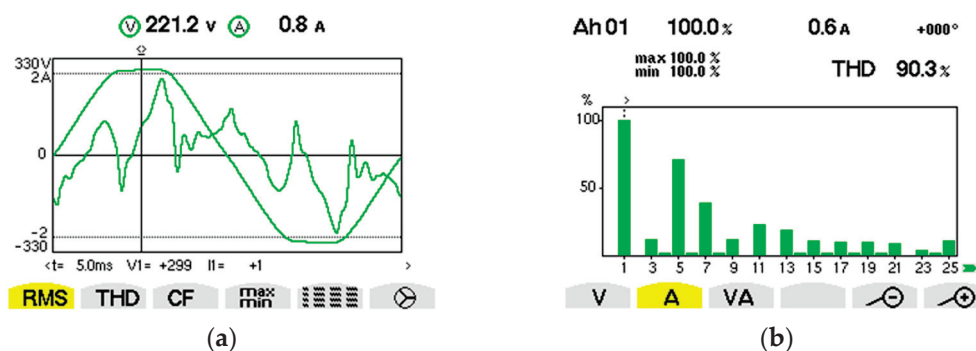


Figure 24. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1 and M2.

Three CBs of different values will result in better DPF tuning, but THDc will be higher.

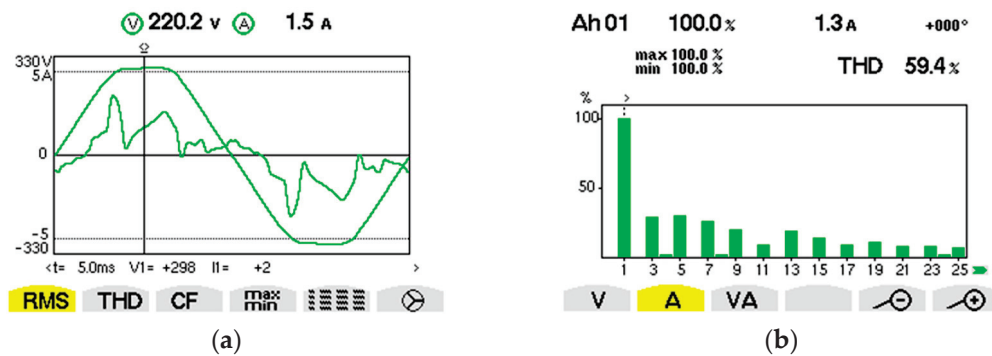


Figure 25. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1, M2, and compact fluorescent lamps.

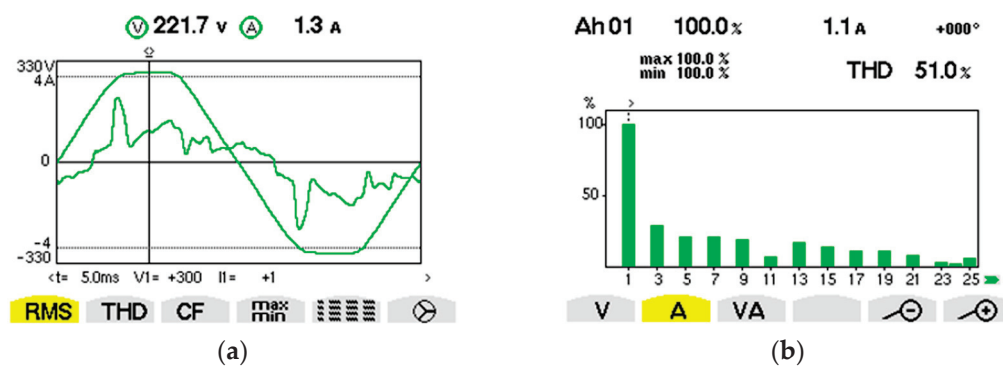


Figure 26. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motor M1 and compact fluorescent lamps.

3.9. PFC with Three Identical Steps, $C1 = 20 \mu F$, $C2 = 20 \mu F$, $C3 = 20 \mu F$ and AC Reactor of 9 mH

When using the PFC with three identical CBs, together with AC reactors, a DPF is obtained that is both inductive and capacitive (Table 11).

Table 11. Experimental measurements for Section 3.9.

Consumer Status	Connected Steps	DPF(-)	I (A)	THDc (%)
Connection M1	1	0.34 i	1.044	15.1
Connection M2	1,2	0.51 i	0.978	41.7
Connection CFLs	1,2	0.87 i	1.516	53
Disconnection M2	1,2	0.9 c	1.333	56
Disconnection M1	-	0.99 c	0.86	73.2

The DPF has higher values compared to the case in Section 3.7 and the THDc has lower values (by 3–5%).

3.10. PFC with Three Different Steps, $C1 = 10 \mu F$, $C2 = 20 \mu F$, $C3 = 30 \mu F$ and with AC Reactor of 9 mH

The combined use of PFC with three different stages and with AC reactors leads to DPF with only inductive character (Table 12). THDc will have higher values when only electric motors are used and will have lower values when combinations of consumers are used compared to Section 3.7.

Table 12. Experimental measurements for Section 3.10.

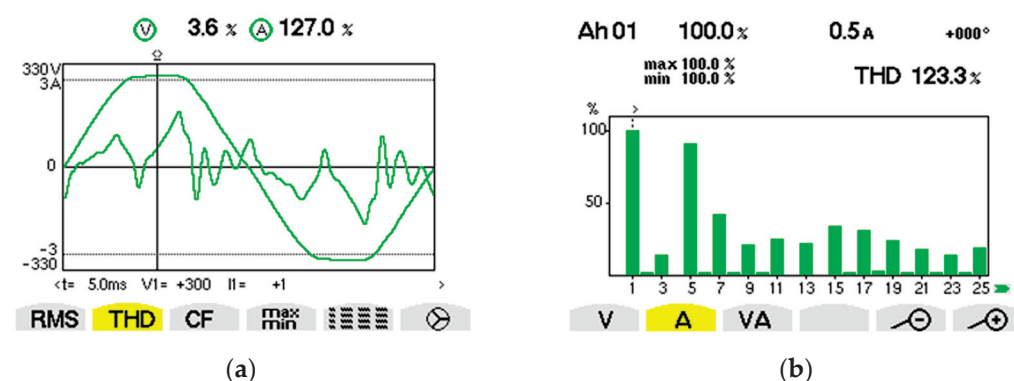
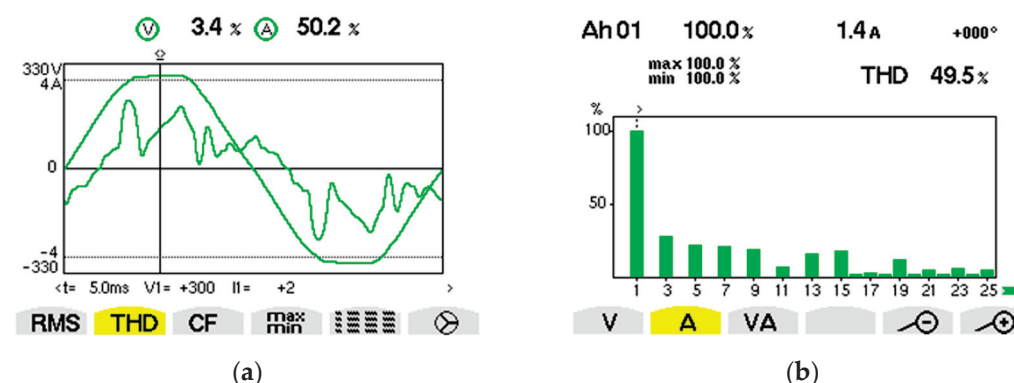
Consumer Status	Connected Steps	DPF (-)	I (A)	THDc (%)
Connection M1	3	0.75 i	0.43	65.5
Connection M2	2,3	0.98 i	0.58	104.8
Connection CFLs	2,3	1	1.35	69.7
Disconnection M2	2	0.8 i	1.36	45.6

3.11. PFC with Five Different Steps, $C1 = 1.9 \mu F$, $C2 = 4 \mu F$, $C3 = 8 \mu F$, $C4 = 16.9 \mu F$, $C5 = 31.9 \mu F$

Five different capacitors banks connected at five stages to the PFC are used. The values of the capacitors were chosen so that they are (approx.) twice the other, with the aim of obtaining a smaller number of steps. In this situation, inductive DPFs were obtained with values above 0.92 (Table 13). When the M2 motor was disconnected, steps 2,3,4 were initially connected (case a), then, after a period of time (case b), steps 1,2,3,4 were connected.

Table 13. Experimental measurements for Section 3.11.

Consumer Status	Connected Steps	DPF (-)	I (A)	THDc (%)	Figures
Connection M1	2,5	0.97 i	0.381	106.5	
Connection M2	1,2,4,5	0.97 i	0.605	123.3	Figure 27
Connection CFLs	1,2,3,5	0.92 i	1.465	49.5	Figure 28
Disconnection M2 a	2,3,4	0.94 i	1.251	47.9	
Disconnection M2 b	1,2,3,4	0.97 i	1.2	49.6	Figure 29

**Figure 27.** (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1 and M2.**Figure 28.** (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1, M2, and compact fluorescent lamps.

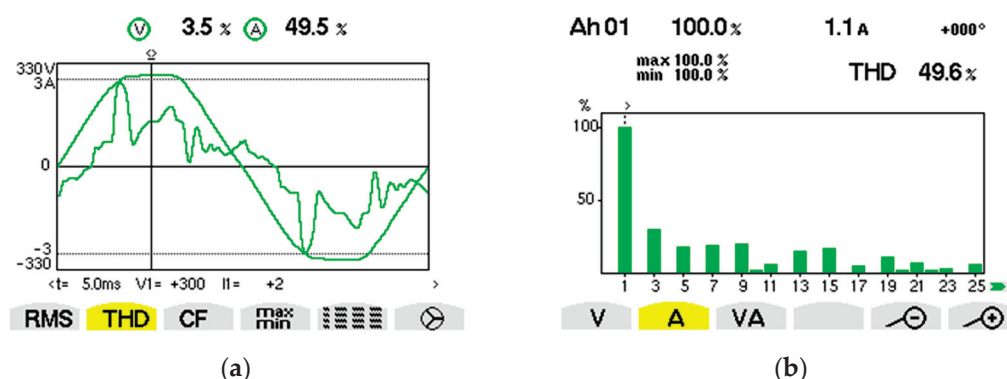


Figure 29. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motor M1 and compact fluorescent lamps.

THDc has high values when using electric motors and values are reduced by half when using a combination of electrical consumers. Higher capacity CBs lead to higher current distortion.

When only electric motors are used, high harmonics are recorded, fitting with the 5th order.

When a combination of consumers is used, the amplitudes of the harmonics decrease (due to the diversity effect).

3.12. PFC with Six Identical Steps, $C1 = C2 = C3 = C4 = C5 = C6 = 10 \mu F$

The use of six identical CBs leads to an inductive DPF adjustment with values above 0.93. THDc has 10–15% lower values compared to using five-stage PFC (Table 14).

Table 14. Experimental measurements for Section 3.12.

Consumer Status	Connected Steps	DPF (-)	I (A)	THDc (%)	Figure
Connection M1 + M2	1–6	0.96 i	0.637	91.4	Figure 30
Connection CFLs	1,3,4,5,6	0.96 i	1.342	42.4	Figure 31
Disconnection M2	3,4,5	0.93 i	1.192	41.4	Figure 32
Disconnection CFLs a	1,5,6	0.52 i	0.657	30.7	
Disconnection CFLs b	1,2,5,6	0.97 i	0.4	88.3	

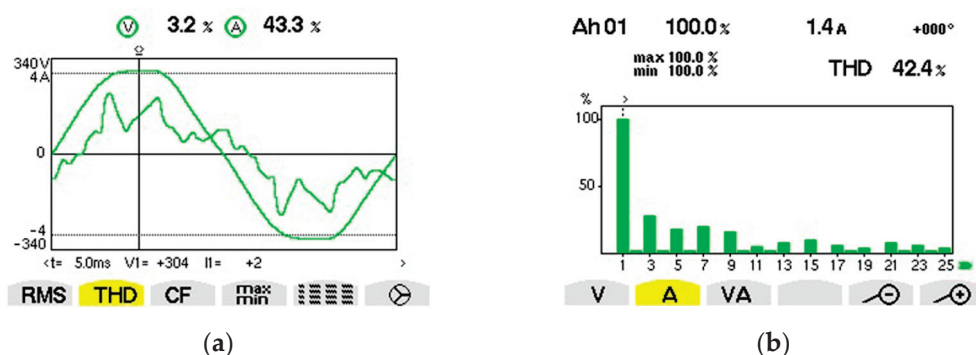


Figure 30. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1 and M2.

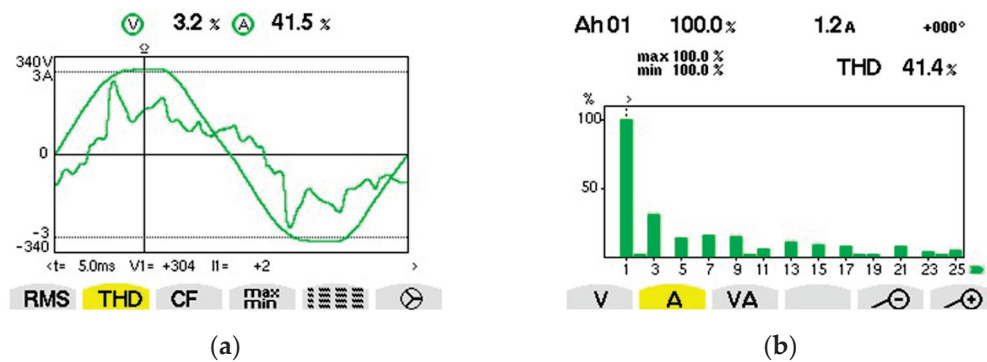


Figure 31. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1, M2, and compact fluorescent lamps.

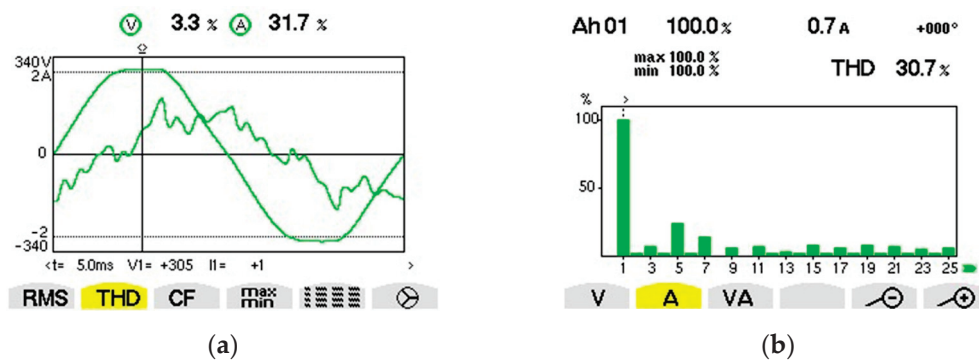


Figure 32. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motor M1 and compact fluorescent lamps (steps 1, 5, 6).

3.13. PFC with Six Steps with Different Capacitors, $C1 = 1.9 \mu F$, $C2 = 2.5 \mu F$, $C3 = 4 \mu F$, $C4 = 10 \mu F$, $C5 = 15 \mu F$, $C6 = 20 \mu F$

When using the PFC with six different CBs, an adjustment of the inductive DPF over 0.9 was obtained (Table 15). THDc has high values, compared to the measurements made in Section 3.11.

Table 15. Experimental measurements for Section 3.13.

Consumer Status	Connected Steps	DPF (-)	I (A)	THDc (%)	Figure
Connection M1	5,6	0.96 i	0.383	105.1	Figure 33
Connection M2	1–6	0.9 i	0.634	105.9	Figure 34
Connection CFLs	3–6	0.97 i	1.424	54.5	Figure 35
Disconnection M2	2–5	0.98 i	1.219	48.8	Figure 36

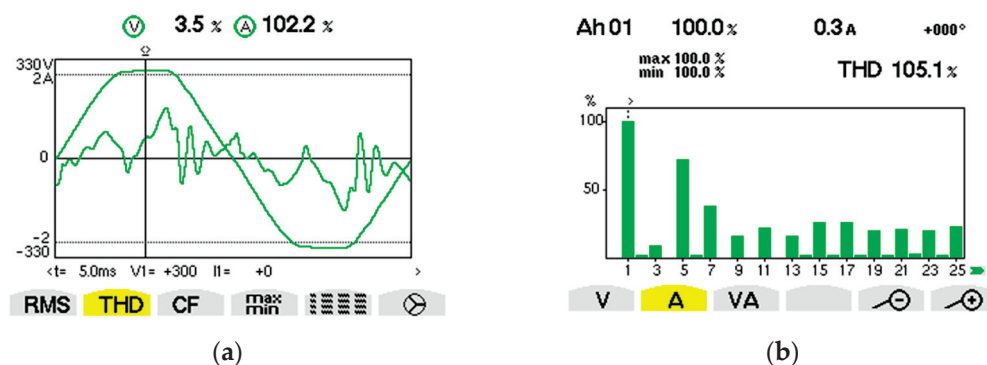


Figure 33. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motor M1.

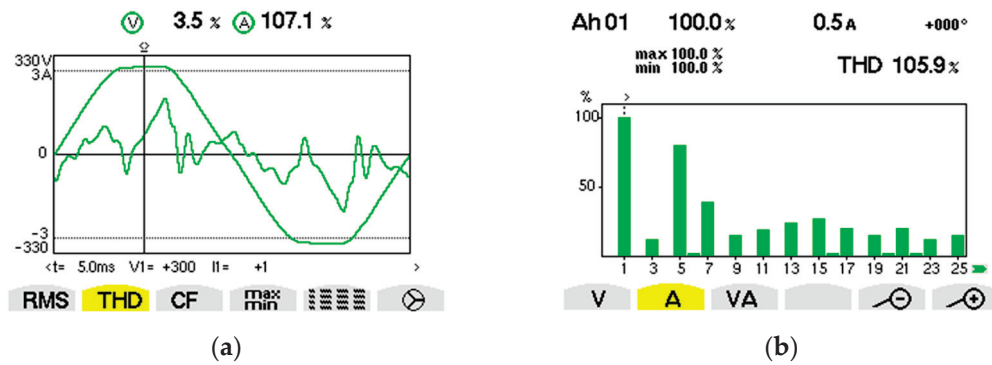


Figure 34. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1 and M2.

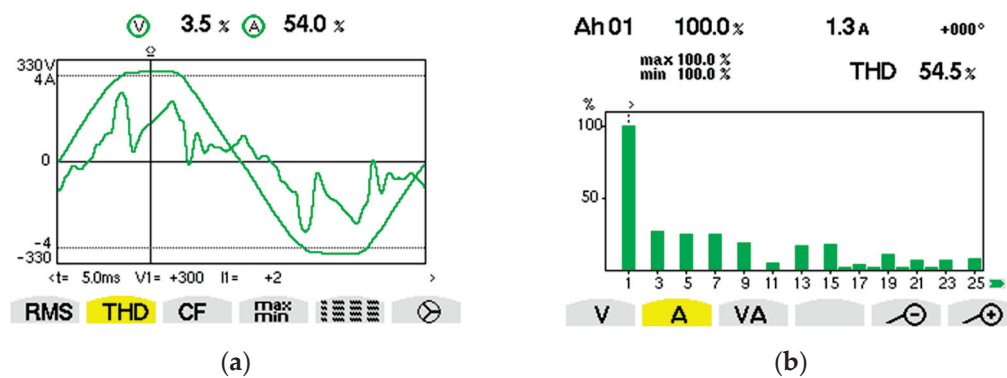


Figure 35. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1, M2, and compact fluorescent lamps.

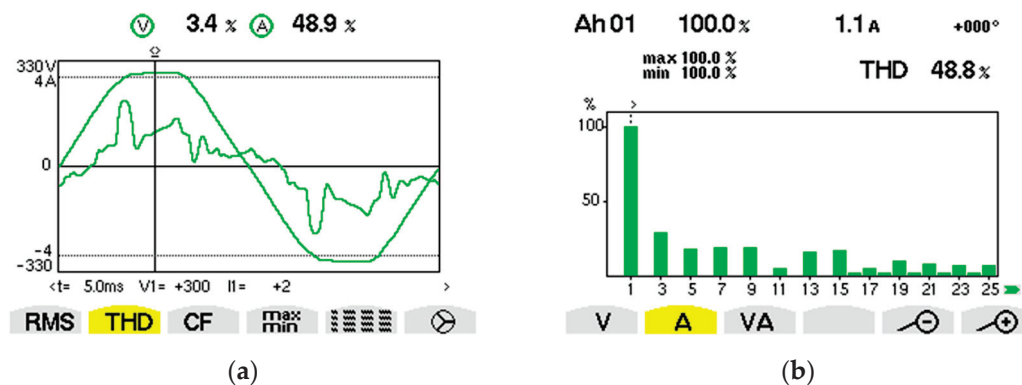


Figure 36. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motor M1 and compact fluorescent lamps.

By using six stages with different CBs, DPF adjustment is obtained, but THDc will have high values when IM is connected.

When electric motors are used, the harmonics starting with rank 5 will have high values.

With the combined use of electrical consumers, the third order harmonic increases, but the higher order ones decrease.

3.14. PFC with Three Steps with LC Shunt Filters, Case 1

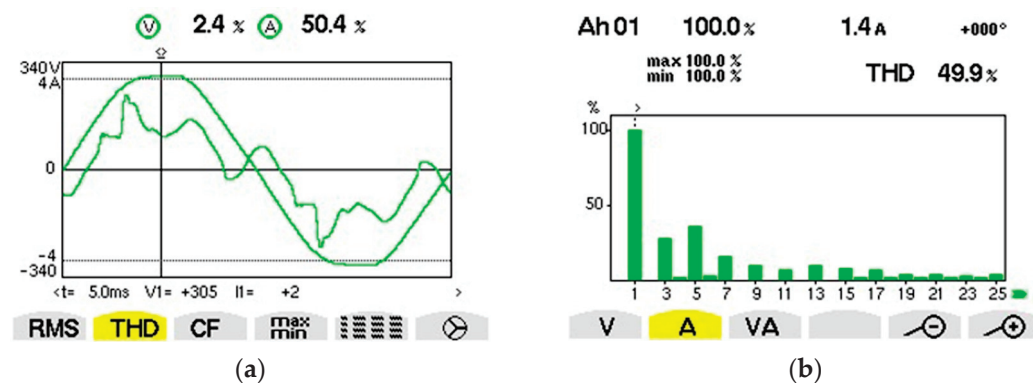
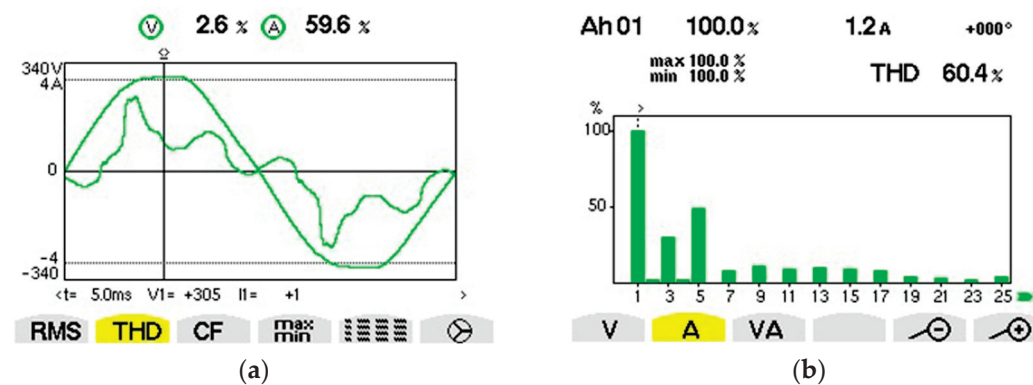
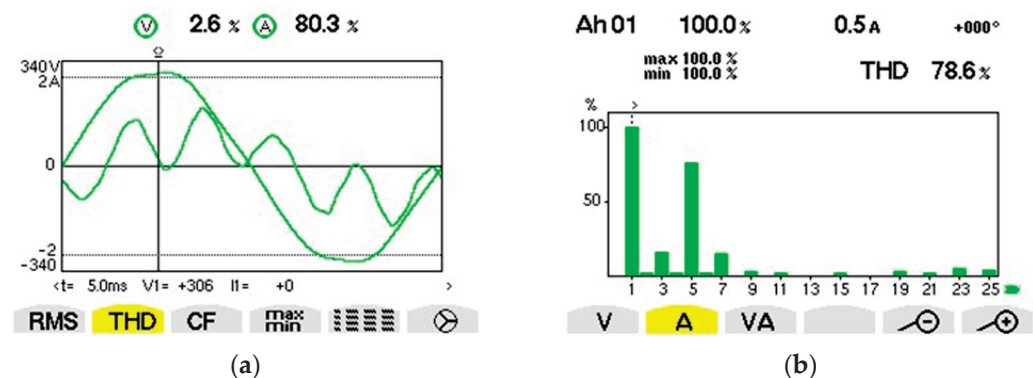
In the following experiments, LC filters were introduced instead of CBs (Table 16). The values are as follows:

- Step 1 (Filter LC 1): $L1 = 97.07$ mH; $C1 = 10$ μ F; $If1 = 0.74$ A; $f = 161.54$ Hz;
- Step 2 (Filter LC 2): $L2 = 26.42$ mH; $C2 = 20$ μ F; $If2 = 1.43$ A; $f = 218.94$ Hz;

- Step 3 (Filter LC 3): $L3 = 15.48 \text{ mH}$; $C3 = 30 \text{ }\mu\text{F}$; $I_{f3} = 2.24 \text{ A}$; $f = 233.54 \text{ Hz}$.
Inductive DPF was obtained in all cases, but the filters did not reduce THDc.

Table 16. Experimental measurements for Section 3.14.

Consumer Status	Connected Steps	DPF (-)	I (A)	THDc (%)	Figure
Connection M1 + M2	2,3	0.86 i	0.992	103.7	Figure 37
Connection CFLs	2,3	1	1.42	49.9	Figure 38
Disconnection M2	2,3	0.98 i	1.22	60.4	Figure 39
Disconnection CFLs	3	0.73 i	0.583	78.6	

**Figure 37.** (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1 and M2.**Figure 38.** (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1, M2 and compact fluorescent lamps.**Figure 39.** (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motor M1 and compact fluorescent lamps.

A decalibration of the coils and, implicitly, of the filters will lead to a lower value DPF and a THDc of over 100%.

Current harmonics are large for ranks 3, 5, 7, 9, etc.

3.15. PFC with Three Steps with LC Shunt Filters, Case 2

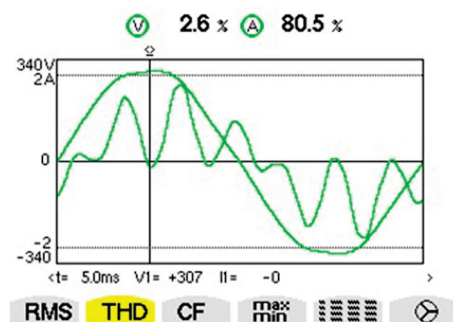
In the following experiments, LC filters were introduced instead of CBs. The values are as follows:

- Step 1 (Filter LC 1): $L1 = 97.07 \text{ mH}$; $C1 = 10 \text{ }\mu\text{F}$; $I_{f1} = 0.74 \text{ A}$; $f = 161.53 \text{ Hz}$;
- Step 2 (Filter LC 2): $L2 = 10.81 \text{ mH}$; $C2 = 20 \text{ }\mu\text{F}$; $I_{f2} = 1.42 \text{ A}$; $f = 342.29 \text{ Hz}$;
- Step 3 (Filter LC 3): $L3 = 24.26 \text{ mH}$; $C3 = 30 \text{ }\mu\text{F}$; $I_{f3} = 2.3 \text{ A}$; $f = 186.55 \text{ Hz}$.

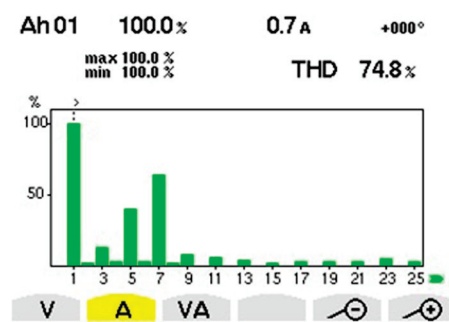
In all experiments, the inductive DPF is adjusted. When only electric motors are used, THDc is reduced considerably (by tens of %)—Table 17.

Table 17. Experimental measurements for Section 3.15.

Consumer Status	Connected Steps	DPF (-)	I (A)	THDc (%)	Figure
Connection M1 + M2	2,3	0.9 i	0.7	74.8	Figure 40
Connection CFLs	2,3	1	1.39	53.9	Figure 41
Disconnection M2	3	0.98 i	1.286	63.8	Figure 42
Disconnection CFLs	3	0.7 i	0.65	85.1	Figure 43

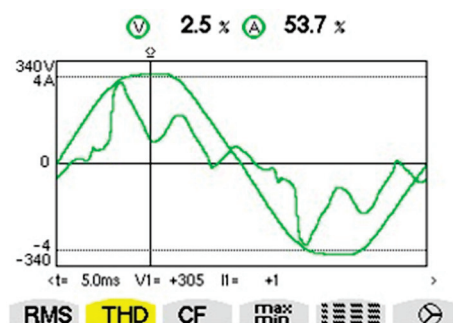


(a)

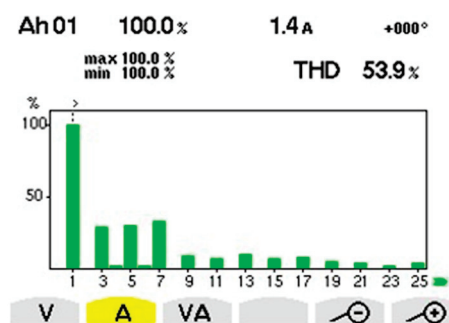


(b)

Figure 40. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1 and M2.



(a)



(b)

Figure 41. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motors M1, M2, and compact fluorescent lamps.

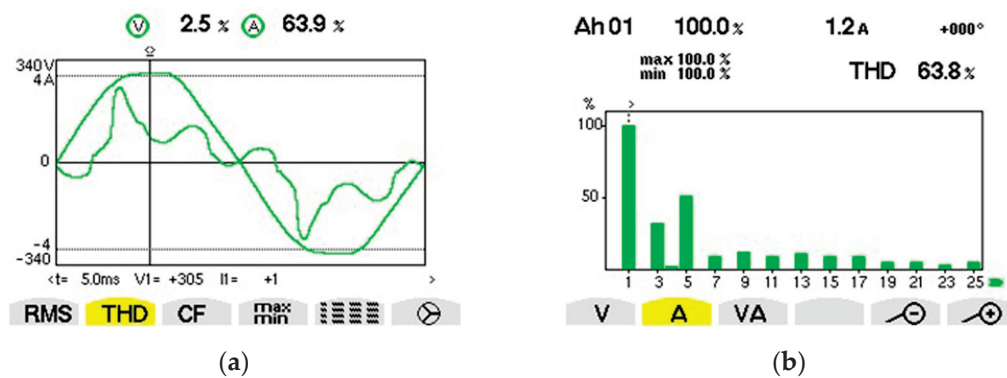


Figure 42. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motor M1 and compact fluorescent lamps.

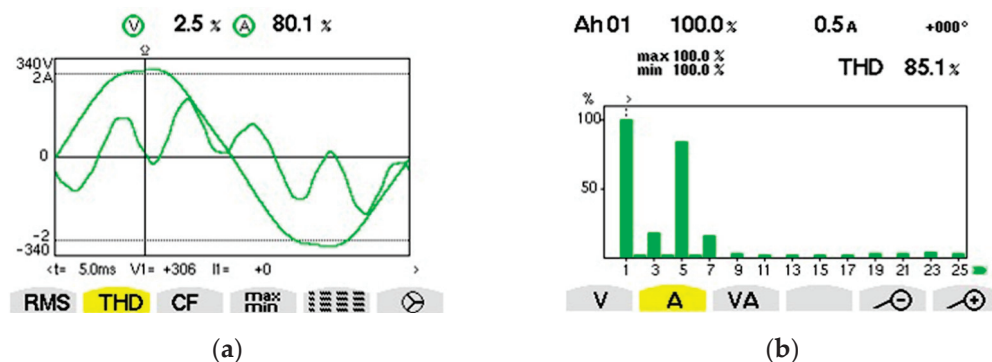


Figure 43. (a) The voltage and the current. (b) The spectrum of harmonics for the current supplying the induction motor M1.

When using consumers, the filters can amplify some harmonics (e.g., of the 5th order).

3.16. PFC with Three Steps with LC Shunt Filters, Case 3

In the following experiments, LC filters were introduced instead of CBs (Table 18). The values are as follows:

- Step 1 (Filter LC 1): $L1 = 10.78 \text{ mH}$; $C1 = 20 \text{ }\mu\text{F}$; $I_{f1} = 1.45 \text{ A}$; $f = 342.76 \text{ Hz}$;
- Step 2 (Filter LC 2): $L2 = 10.78 \text{ mH}$; $C2 = 20 \text{ }\mu\text{F}$; $I_{f2} = 1.45 \text{ A}$; $f = 342.76 \text{ Hz}$;
- Step 3 (Filter LC 3): $L3 = 10.78 \text{ mH}$; $C3 = 20 \text{ }\mu\text{F}$; $I_{f3} = 1.45 \text{ A}$; $f = 342.76 \text{ Hz}$.

Lower values of coil inductances can lead to a decrease in DPF (inductive and capacitive character) and a slight increase in THDc (Table 18).

Table 18. Experimental measurements for Section 3.16.

Consumer Status	Connected Steps	DPF (-)	I (A)	THDc (%)
Connection M1 + M2	1,3	0.42 i	1.31	65.9
Connection CFLs	1,3	0.76 i	1.69	54.1
Disconnection M2	1,3	0.97 c	1.3	73.9
Disconnection CFLs	1,3	0.92 c	0.6	142.7

3.17. PFC with Six Different Steps with LC Shunt Filters

In the following experiments, LC filters were introduced instead of CBs (Table 19). The values are as follows:

- Step 1 (Filter LC 1): $L1 = 97.07 \text{ mH}$; $C1 = 4 \text{ }\mu\text{F}$; $f = 255.41 \text{ Hz}$;
- Step 2 (Filter LC 2): $L2 = 97.07 \text{ mH}$; $C2 = 5 \text{ }\mu\text{F}$; $f = 228.45 \text{ Hz}$;
- Step 3 (Filter LC 3): $L3 = 26.42 \text{ mH}$; $C3 = 11.9 \text{ }\mu\text{F}$; $f = 283.84 \text{ Hz}$;

- Step 4 (Filter LC 4): $L_4 = 15.48 \text{ mH}$; $C_4 = 16.9 \text{ }\mu\text{F}$; $f = 311.16 \text{ Hz}$;
- Step 5 (Filter LC 5): $L_5 = 24.26 \text{ mH}$; $C_5 = 7.5 \text{ }\mu\text{F}$; $f = 373.11 \text{ Hz}$;
- Step 6 (Filter LC 6): $L_6 = 47 \text{ mH}$; $C_6 = 10 \text{ }\mu\text{F}$; $f = 232.15 \text{ Hz}$.

An inductive DPF was obtained, but the THDc has very high values (Table 19).

Table 19. Experimental measurements for Section 3.17.

Consumer Status	Connected Steps	DPF (-)	I (A)	THDc (%)
Connection M1 a	4–6	0.7 i	0.63	101.3
Connection M1 b	2,4–6	0.99 i	0.58	116.6
Connection M2 a	2–6	0.75 i	1	105.6
Connection M2 b	1–6	0.92 i	0.979	108.5
Connection CFLs	2–6	0.97 i	1.62	88.1

The use of a greater number of stages, which also fulfill the function of LC filters, will not lead, at the same time, to the increase in the DPF and to the decrease in the deforming regime.

3.18. PFC with Six Different Steps with LC Shunt Filters

In the following experiments, LC filters were introduced instead of CBs (Table 20). The values are as follows:

- Step 1 (Filter LC 1): $L_1 = 11.45 \text{ mH}$; $C_1 = 10 \text{ }\mu\text{F}$; $f = 470.58 \text{ Hz}$;
- Step 2 (Filter LC 2): $L_2 = 11.45 \text{ mH}$; $C_2 = 10 \text{ }\mu\text{F}$; $f = 470.58 \text{ Hz}$;
- Step 3 (Filter LC 3): $L_3 = 11.45 \text{ mH}$; $C_3 = 10 \text{ }\mu\text{F}$; $f = 470.58 \text{ Hz}$;
- Step 4 (Filter LC 4): $L_4 = 63.6 \text{ mH}$; $C_4 = 10 \text{ }\mu\text{F}$; $f = 199.66 \text{ Hz}$;
- Step 5 (Filter LC 5): $L_5 = 63.6 \text{ mH}$; $C_5 = 10 \text{ }\mu\text{F}$; $f = 199.66 \text{ Hz}$;
- Step 6 (Filter LC 6): $L_6 = 63.6 \text{ mH}$; $C_6 = 10 \text{ }\mu\text{F}$; $f = 199.66 \text{ Hz}$.

When adjusting the DPF, only inductive characteristics were obtained and THDc decreased (by 15–20%) compared to the previous case (Table 20).

Table 20. Experimental measurements for Section 3.18.

Consumer Status	Connected Steps	DPF (-)	I (A)	THDc (%)
Connection M1 + M2	1–5	0.86 i	0.393	80.4
Connection CFLs	1–5	0.99 i	1.49	61.9
Disconnection M2	3–5	0.99 i	1.35	69
Disconnection CFLs	3–5	0.69 i	0.581	95.5

The results are better than in Section 3.18. DPF is higher (above 0.69) and THDc is lower (below 96%).

3.19. PFC with Six Steps with Capacitors and AC Reactor of 3 mH

CBs were used in the following experiments and AC reactors were introduced in series with the consumers. The values are as follows:

- Step 1: $C_1 = 4 \text{ }\mu\text{F}$;
- Step 2: $C_2 = 5 \text{ }\mu\text{F}$;
- Step 3: $C_3 = 11.9 \text{ }\mu\text{F}$;
- Step 4: $C_4 = 16.9 \text{ }\mu\text{F}$;
- Step 5: $C_5 = 7.5 \text{ }\mu\text{F}$;
- Step 6: $C_6 = 10 \text{ }\mu\text{F}$.

In the analyzed situations, DPFi had an inductive character (with values over 0.96), and THDc had lower values compared to the use of LC filters (Table 21).

Table 21. Experimental measurements for Section 3.19.

Consumer Status	Connected Steps	DPF (-)	I (A)	THDc (%)
Connection M1	3,4,6	0.99 i	0.365	94.1
Connection M2	1–6	0.96 i	0.618	84.2
Connection CFLs	2–6	0.98 i	1.441	45.6

It is a combination between the coil inserted with consumers and capacitors C. The capacity of the capacitors when using L-C filters must be chosen 10–15% higher compared to the case of using only single capacitors banks.

3.20. PFC with Six Steps with LC Shunt Filters and AC Reactor of 3 mH

In the following experiments, LC filters were introduced instead of CBs and AC reactors were introduced in series with the consumers. The values are as follows:

- Step 1 (Filter LC 1): $L1 = 47 \text{ mH}$, $C1 = 4 \text{ }\mu\text{F}$, $f = 367.25 \text{ Hz}$;
- Step 2 (Filter LC 2): $L2 = 26.29 \text{ mH}$, $C2 = 5 \text{ }\mu\text{F}$, $f = 439.2 \text{ Hz}$;
- Step 3 (Filter LC 3): $L3 = 15.48 \text{ mH}$, $C3 = 11.9 \text{ }\mu\text{F}$, $f = 371 \text{ Hz}$;
- Step 4 (Filter LC 4): $L4 = 26.42 \text{ mH}$, $C4 = 16.9 \text{ }\mu\text{F}$, $f = 238.3 \text{ Hz}$;
- Step 5 (Filter LC 5): $L5 = 24.26 \text{ mH}$, $C5 = 7.5 \text{ }\mu\text{F}$, $f = 373.11 \text{ Hz}$;
- Step 6 (Filter LC 6): $L6 = 97.07 \text{ mH}$, $C6 = 10 \text{ }\mu\text{F}$, $f = 161.62 \text{ Hz}$.

In the analyzed situations, DPF had an inductive character, but THDc has higher values compared to the previous cases (Table 22).

Table 22. Experimental measurements for Section 3.20.

Consumer Status	Connected Steps	DPF (-)	I (A)	THDc (%)
Connection M1	2–4	0.94 i	0.54	112.6
Connection M2	2–6	0.95 i	0.686	108.6
Connection CFLs	3–6	0.97 i	1.489	69.7

The combined use of AC reactors and LC filters is not recommended for a PFC with a large number of stages.

4. Discussion

The most common PFCs are those that have a large number of stages (6–14) of capacitors that can be connected in the electrical installation and that are intended for DPF regulation in three-phase installations. Typically, these PFCs measure (via a CT) the line current and a phase voltage to determine the DPF. The improvement of the power factor of these PFCs is achieved by obtaining a DPF above a certain preset value (e.g., the neutral power factor). In deforming mode, things are more complicated because there are differences between the power factor and DPF (the more deforming the regime, the bigger the difference between PF and DPF). Obviously, the number of steps is chosen depending on the number, type, and operating mode of the consumers that are supplied from the power station. A large number of steps can ensure a finer adjustment of the DPF, but it decreases the reliability of the entire installation (it becomes more complex, with more switching and protection equipment and more CBs). Usually, the PFC works properly when an inductive DPF (not capacitive) is obtained with values higher than the preset value (0.9).

In the following (Tables 23–25), the experimental measurements from Section 3 are presented, depending on the connection of consumers used in the experiments, DPF and THDc, after which PFC introduced CBs or LC filters.

Table 23. Connecting the M1 motor.

Experiment Number	DPF (-)	THDc (%)
Section 3.1	0.13 i	8.2
Section 3.2	0.16 i	9.3
Section 3.3	0.34 c	52.3
Section 3.5	0.92 c	110.8
Section 3.6	0.85 c	100.8
Section 3.7	0.7 i	80.6
Section 3.8	0.59 i	54.5
Section 3.9	0.34 i	15.1
Section 3.10	0.75 i	65.5
Section 3.11	0.97 i	106.5
Section 3.13	0.96 i	105.1
Section 3.17	0.99 i	116.6
Section 3.19	0.99 i	94.1
Section 3.20	0.94 i	112.6

Table 24. Connecting the M1 and M2 motors.

Experiment Number	DPF (-)	THDc (%)
Section 3.1	0.12 i	7
Section 3.2	0.91 c	101.1
Section 3.3	0.95 i	113.5
Section 3.4	0.82 i	116.1
Section 3.5	0.38 i	30.2
Section 3.6	0.86 c	117.1
Section 3.7	0.49 i	37.9
Section 3.8	0.77 i	90.3
Section 3.9	0.51 i	41.7
Section 3.10	0.98 i	104.8
Section 3.11	0.97 i	123.3
Section 3.12	0.96 i	91.4
Section 3.13	0.9 i	105.9
Section 3.14	0.86 i	103.7
Section 3.15	0.9 i	74.8
Section 3.16	0.42 i	65.9
Section 3.17	0.92 i	108.5
Section 3.18	0.86 i	80.4
Section 3.19	0.96 i	84.2
Section 3.20	0.95 i	108.6

A relatively fine adjustment of the DPF can be obtained even if there are fewer steps (e.g., three steps with different CBs). With one stage (with a high capacity per stage), two stages (identical or different), or three identical stages, DPF can be obtained several times. Under these conditions, the use of coils inserted when one, two, or three stages are used does not result in a decrease in THDc (it can even result in an increase of 5 to 7%).

The DPF is best adjusted when using 5 and 6 steps, where, in all situations, an inductive DPF is obtained. If 6 stages are used, all different, with capacities to compensate the inductive reactive power, in the most disadvantageous case (inductive consumers), slightly higher inductive DPFs are obtained, but also a higher THDc compared to the use of six identical stages.

An interesting case is the use of five different stages, where each stage has double the capacity of the other. In this case, inductive DPFs have slightly higher values than in the case of using the six-step regulator, but THDc is also higher. So, if it uses steps with double the capacity compared to each other, it can obtain an even better adjustment of the DPF than in the case of six steps (with different or identical values).

Table 25. Connecting the M1 and M2 motors and CFLs.

Experiment Number	DPF (-)	THDc (%)
Section 3.1	0.32 c	17.3
Section 3.2	1	56.1
Section 3.3	1	62.6
Section 3.5	0.76 i	39.5
Section 3.6	0.93 c	59.5
Section 3.7	0.86 c	45.6
Section 3.8	0.99 i	59.4
Section 3.9	0.87 i	53
Section 3.10	1	69.7
Section 3.11	0.92 i	49.5
Section 3.12	0.96 i	42.4
Section 3.13	0.97 i	54.5
Section 3.14	1	49.9
Section 3.15	1	53.9
Section 3.16	0.76 i	54.1
Section 3.17	0.97 i	88.1
Section 3.18	0.99 i	61.9
Section 3.19	0.98 i	45.6
Section 3.20	0.97 i	69.7

It is important to choose the values of the capacitors to improve the DPF when all the inductive consumers are connected. The use of deforming consumers that also have a capacitive character determines a slight improvement of the DPF.

If an LC shunt filter is used instead of a single stage, DPF can be lower and THDc higher, compared to the situation of a single stage (with high capacity) or a coil inserted with consumers and a single stage.

If PFC is used with three different stages, each stage having an LC shunt filter, it is possible to increase the DPF (above the value of 0.9) and decrease the THDc (by 5–7%). It is important to properly adjust the filters (quite difficult in practice), otherwise they can cause a slight increase in the DPF (does not reach 0.9) and an increase in the THDc (higher values compared to using only capacitors for the PFC). The use of LC filters, with identical capacitors leads to inductive and capacitive DPF, and the THDc can have higher values (compared to the use of LC shunt filters, with different capacities connected to the PFC).

It has been determined experimentally that the use of a large number of steps (e.g., 6) of LC shunt filters or combinations of coils inserted with consumers and LC shunt filters does not lead, at the same time, to the increase in DPF and to the decrease in THDc. With a large number of steps, it is more efficient to use AC reactors, compared to the use of LC filters on each step. Practically, it is difficult to tune several LC filters, and resonances can appear when the LC filters are operating in parallel. The best performances were achieved with three stages with LC filters, where the capacities have different values.

Some electrical consumers can be inductive, while others can have a capacitive character. Depending on the place where PFC is used in the power substations, there are, usually, three-phase linear electrical consumers (e.g., induction motors) or single-phase consumers (electric heaters), but there can also be three-phase non-linear consumers (e.g., variable frequency drives) and non-linear single-phase consumers (e.g., compact fluorescent lamps). Through the intelligent distribution of the various consumers on phases, an improvement of the DPF can be ensured, and perhaps a reduction in the THDc (the phenomenon of diversity also appears, which can reduce the THDc). For example, at a transformer substation that supplies many non-linear capacitive consumers (e.g., transformer substations that supply households and residential and educational institutions [20]), a classic PFC with capacitors banks no longer proves its usefulness because, overall, consumers have a capacitive character with a high value DPF (0.9–1). At the same time, it must be taken into account that most consumers are single-phase ones that determine an unbalance on the three-phases in the point of common coupling of the installation and even a significant

current through the neutral conductor (due to both the unbalance and the consumers non-linearity).

For classic PFCs that measure a single line current, it is important to choose the phase on which the CT is mounted (usually, it is mounted on the most loaded phase). Another important aspect concerns the choice of an optimal transformation ratio for the CT. A too-low value of the transformation ratio may exceed the maximum value of the current measured by the PFC, and a too-high value of the transformation ratio may determine a current below the sensitivity limit of the PFC (e.g., below 150 mA, the current is no longer measured correctly and the DPF adjustment is no longer performed correctly).

For unbalanced electrical networks that have slightly deforming consumers, it is recommended to use high-performance PFCs with a large number of capacitors banks (of different values, appropriately sized) that perform the regulation of the reactive power independently on each phase separately (by using three CTs mounted on each phase separately). The installation can also have AC reactors to reduce the deforming regime. The use of different steps, where each step has double the value of the other, can achieve the DPF adjustment with a smaller number of steps, and the reliability of the installation increases. A large number of steps reduces the reliability of the entire installation. For unbalanced electrical networks that have high deformation consumers, it is recommended to use a small number of PFCs (e.g., three) of high-performance LC filters (with different capacities in each phase) to control the reactive power independently in each phase (using three CTs mounted in each phase and measuring the DPF independently in each phase), while reducing (to some extent) the deformation regime.

To reduce the deforming regime, the combined use of AC reactors and LC filters is not recommended (even if they are sized appropriately for common operation, there is a high chance that the deforming regime will be amplified due the resonance).

5. Conclusions

Classical power factor controllers with capacitors banks, which use a single current transformer per phase and measures a phase voltage (in order to determine the phase shift between the voltage and current fundamentals to calculate the displacement power factor) have been used for a long time. Such a regulator cannot provide regulation according to the power factor. They are used more often in industry in transformer stations, where there are many single-phase and three-phase inductive consumers.

Usually, a small number of steps (below three) is not recommended because a fine adjustment of the displacement power factor will not be obtained (of course there are exceptions). Using a smaller number (than the maximum number of adjustment stages of the power factor controller) of capacitors banks, where the capacity doubles from one bank to another, can determine the same results as using the power factor controller with a maximum number of steps.

By connecting capacitors banks, a higher displacement power factor can be obtained, but the deforming regime increases (the current becomes more distorted, and the total harmonic distortion increases). If the deforming regime worsens, there will be a considerable difference between the displacement power factor and the power factor (which is smaller). To improve the deforming regime (to some extent) AC reactors can be used, which can be used in a power factor controller with several capacitors banks, or LC shunt filters can be used if a power factor controller with fewer capacitors banks is required. The use of several LC shunt filters and the combination of AC reactors and LC shunt filters leads to the amplification of the deforming regime (unwanted resonances appear).

Future research will be carried out on the analysis of the operation of the classical power factor controller that controls the displacement power factors, on all phases (even if the displacement power factor is adjusted on a single-phase) in unbalanced three-phase electrical networks, with linear and non-linear electrical consumers. Research will be carried out to use new power factor controllers that independently control the operation of improving the power factor and improving currents at each phase.

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