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Enhanced Singular Value Decomposition Modulation Technique to Improve Matrix Converter Input Reactive Power Control

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Abstract

Matrix converters (MC) offer a compact, bidirectional solution for power conversion; however, achieving precise reactive power control at the input terminals remains challenging under varying operating conditions. This paper presents an enhanced Singular Value Decomposition modulation technique (e-SVD) as a solution tailored to optimize reactive power management on the MC input side, enabling both active and reactive power control regardless of the power factor. The proposed method achieves input reactive power control based on a reactive power gain, a quantity derived from the apparent output power and defined by a mathematical expression involving electrical parameters and control variables. Experimental tests carried out on a low-power MC prototype to validate the proposal show that the measured reactive power gain closely aligns with theoretical predictions from the mathematical expressions. Overall, the proposed e-SVD modulation technique lays the foundation for more reliable reactive power regulation in applications such as microgrids and distributed generation systems, contributing to the development of smarter and more resilient energy infrastructures.



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1. Introduction

The matrix converter (MC) is a power electronics device that performs direct AC–AC conversion without bulky energy storage components. This device offers several advantages, including high power density, high-quality low-frequency signals, bidirectional power flow, and a controllable input power factor, among others [1–5]. Several modulation techniques have been proposed for MC control [6], and new methods continue to be developed, highlighting the current validity of the power converter as a research topic [7–9].

A notable feature of the MC is its ability to control the input power factor, which has been explored to enhance the capabilities of the device. Several studies have proposed controlling the MC to achieve unitary power factor operation upstream of the input filter [10–12], where the reactive power at the input terminals of the MC compensates for the filter’s reactive power demand. Another approach enables operation with a controllable input power factor over a limited range using space vector modulation (SVM) [13]. Reactive power management is critical for MC operation and its interaction with the power grid. As a result, research has extended beyond power factor control to directly investigate

reactive power behavior at the input terminals. Ref. [14] presents an MC with an auxiliary circuit designed to improve input reactive power control by introducing an additional degree of freedom, albeit with increased control complexity to cope with the additional hardware. Research [15] focuses on developing an input reactive power controller with an active damping strategy for MCs used in wind power generation systems. In [16], a modulation technique is proposed for a three-input, five-output (3X5) MC to maximize input reactive power based on load conditions. The work presented in [17], introduces a modulation strategy to achieve maximum input reactive power, based on the active or reactive output power and the input–output voltage ratio. Similarly, ref. [18] proposes several indirect modulation techniques that consider the output power factor to enhance the MC's input reactive power capacity. In [19], two additional modulation methods are presented, showing improved performance compared to those in [18]. The last three proposals aim to maximize input reactive power under specific load conditions. However, they do not provide explicit mathematical expressions or clearly defined operating conditions for precise reactive power control. Finally, Ref. [20] explores the use of Singular Value Decomposition modulation (SVD-M) to improve the reactive power performance of a permanent magnet synchronous wind generator by solving an optimization problem.

The MC has been proposed for various applications due to its advanced features, including outstanding input reactive power control. Since its invention, the use of the MC as a motor driver has been the most prominent and, consequently, the most extensively studied [21–25]. Another important application is as a power-conditioning device installed between distributed generation (DG) sources and the main power grid [26,27]. In this context, the MC has been proposed as a controller for wind energy conversion systems (WECS), managing the operation of different types of generators, such as in [28], where it is applied to a doubly fed induction generator (DFIG); in [29], to an induction generator; and in [20], to a permanent magnet synchronous machine. Similarly, the MC has been proposed for power quality applications [30–32], as well as a power controller within the electrical system [33], a static synchronous compensator [34–36], a unified power flow controller [37], and a hybrid transformer [38].

To support the integration of the MC in power control applications within electrical systems, this work proposes an enhanced Singular Value Decomposition (e-SVD) modulation technique. This technique consists of a set of algorithms derived from the conventional SVD modulation strategy, designed to improve its performance. To the best of the authors' knowledge, the SVD-M has not been thoroughly investigated in input reactive power control, apart from the results presented in [20]. The proposed method explores the capabilities of SVD-M to manage reactive power at the input terminals while simultaneously supplying both active and reactive power at the output terminals. Accordingly, a formulation for a reactive power gain is developed, relating the apparent output power to the input reactive power through electrical variables and controllable parameters.

The structure of the paper is as follows. Section 2 introduces the fundamentals of the SVD-M. Section 3 details the principles behind the e-SVD modulation algorithm, with a focus on voltage generation and reactive power management. Section 4 presents experimental results that validate the effectiveness of the proposal. Finally, Section 5 summarizes the main conclusions of the work.

2. SVD Modulation Technique (SVD-M)

The SVD modulation proposed in [39] is a duty cycle modulation technique that models the relationship between current and voltage at the MC terminals using space vector signals. The topology of the MC is illustrated in Figure 1, where the modulation strategy operates on signals at the input (abc), after the input filter elements, and at the

output terminals (ABC), before the output filter. The modulation technique defines the procedure for constructing the transformation matrix $\mathbf{M}$, which links the output voltages ($V_{ABC} = [v_A \ v_B \ v_C]^T$) and input currents ($I_{abc} = [i_a \ i_b \ i_c]^T$) to the input voltages ($V_{abc} = [v_a \ v_b \ v_c]^T$) and output currents ($I_{ABC} = [i_A \ i_B \ i_C]^T$), as shown in Equations (1) and (2). The matrix $\mathbf{M}$ is subject to the constraints described in Equation (3), which ensure the safe operation of the matrix converter. In the SVD-M approach, the matrix $\mathbf{M}$ explicitly depends on four parameters θ_i , θ_o , q_d , and q_q , and is defined in Equation (4) [39].

$$V_{ABC} = \mathbf{M} V_{abc} \quad (1)$$

$$I_{abc} = \mathbf{M}^T I_{ABC} \quad (2)$$

$$\mathbf{M} = \begin{bmatrix} m_{aA} & m_{aB} & m_{aC} \\ m_{bA} & m_{bB} & m_{bC} \\ m_{cA} & m_{cB} & m_{cC} \end{bmatrix} \quad (3)$$

$$\begin{cases} 0 \leq m_{ij} \leq 1 \\ m_{iA} + m_{iB} + m_{iC} = 1 \end{cases} \quad i = a, b, c; j = A, B, C$$

$$\begin{aligned} \mathbf{M} &= \frac{q_d - q_q}{3} \begin{bmatrix} \cos(\theta_p) & \cos(\theta_{p-}) & \cos(\theta_{p+}) \\ \cos(\theta_{p-}) & \cos(\theta_{p+}) & \cos(\theta_p) \\ \cos(\theta_{p+}) & \cos(\theta_p) & \cos(\theta_{p-}) \end{bmatrix} \\ &+ \frac{q_d + q_q}{3} \begin{bmatrix} \cos(\theta_m) & \cos(\theta_{m+}) & \cos(\theta_{m-}) \\ \cos(\theta_{m-}) & \cos(\theta_m) & \cos(\theta_{m+}) \\ \cos(\theta_{m+}) & \cos(\theta_{m-}) & \cos(\theta_m) \end{bmatrix} \\ &+ \frac{1}{3} \begin{bmatrix} 1 & 1 & 1 \\ 1 & 1 & 1 \\ 1 & 1 & 1 \end{bmatrix} + \begin{bmatrix} c_1 & c_2 & c_3 \\ c_1 & c_2 & c_3 \\ c_1 & c_2 & c_3 \end{bmatrix} \\ &= M_1 + \begin{bmatrix} c_1 & c_2 & c_3 \\ c_1 & c_2 & c_3 \\ c_1 & c_2 & c_3 \end{bmatrix}, \end{aligned} \quad (4)$$

where angles θ and coefficients c_k ($k = \{1, 2, 3\}$) are defined as follows,

$$\theta_p = \theta_o + \theta_i, \quad \theta_{p+} = \theta_o + \theta_i - \frac{2\pi}{3}, \quad \theta_{p-} = \theta_o + \theta_i,$$

$$\theta_m = \theta_o - \theta_i, \quad \theta_{m+} = \theta_o - \theta_i - \frac{2\pi}{3}, \quad \theta_{m-} = \theta_o - \theta_i,$$

$$c_k = -\min_i \{M(i, k)\} + \sum_j \min_i \{M(i, j)\} \times \frac{1 - \max_i \{M_1(i, k)\} + \min_i \{M_1(i, k)\}}{3 - \sum_j \max_i \{M_1(i, j)\} + \sum_j \min_i \{M_1(i, j)\}}.$$

In the SVD modulation scheme, the three-phase signals (abc) are converted into the $\alpha\beta$ frame by the Clarke transformation, as indicated in Equation (5), to represent the signals on the complex plane. The vector $\alpha + j\beta$ can be expressed in polar form as $re^{j\theta}$, where $r = \sqrt{\alpha^2 + \beta^2}$ and $\theta = \text{atan2}\left(\frac{\beta}{\alpha}\right)$. The three-phase input and output voltages correspond to the complex vectors $v_i e^{j\phi_i}$ and $v_o e^{j\phi_o}$, while the three-phase input and output currents correspond to the vectors $i_i e^{j\gamma_i}$ and $i_o e^{j\gamma_o}$, respectively.

$$V_{\alpha\beta} = K V_{abc}; V_{\alpha\beta} = \begin{bmatrix} v_\alpha \\ v_\beta \end{bmatrix}, V_{abc} = \begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix} \quad (5)$$

$$K = \frac{2}{3} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \quad (6)$$

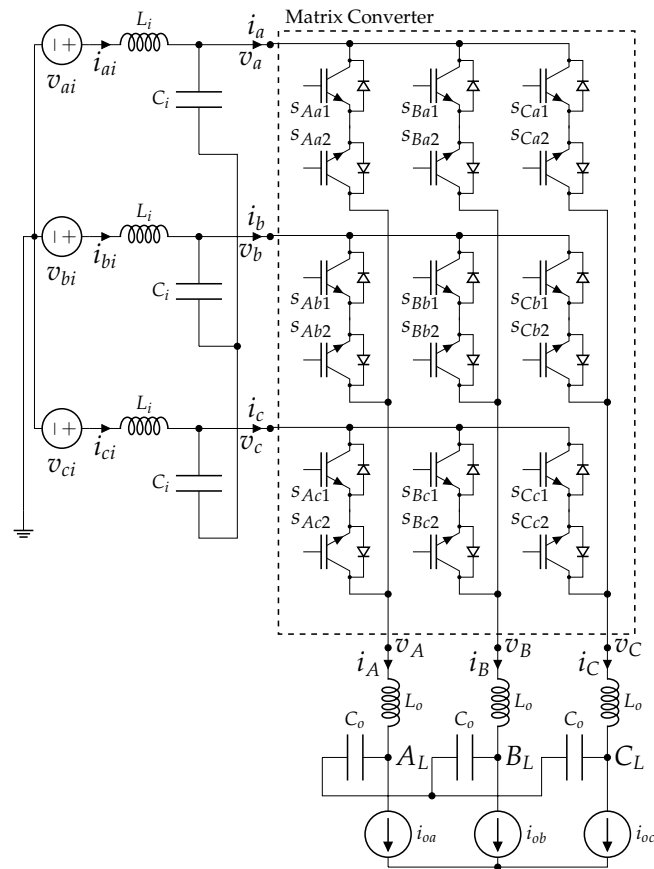


Figure 1. Matrix converter (MC) circuit and peripherals.

In order to generate the output voltages and input currents, the input voltages and output currents are projected onto a rotating dq reference frame defined by the angles θ_i and θ_o . This transformation yields the components v_{di} and v_{qi} for the input voltages, as shown in Equation (7), and the components i_{do} and i_{qo} for the output currents, as shown in Equation (8).

$$\begin{bmatrix} v_{di} \\ v_{qi} \end{bmatrix} = \begin{bmatrix} \cos(\theta_i) & -\sin(\theta_i) \\ \sin(\theta_i) & \cos(\theta_i) \end{bmatrix} \begin{bmatrix} v_{ai} \\ v_{bi} \end{bmatrix} \quad (7)$$

$$\begin{bmatrix} i_{do} \\ i_{qo} \end{bmatrix} = \begin{bmatrix} \cos(\theta_o) & -\sin(\theta_o) \\ \sin(\theta_o) & \cos(\theta_o) \end{bmatrix} \begin{bmatrix} i_{ao} \\ i_{bo} \end{bmatrix} \quad (8)$$

Based on this set of signals, the dq components of the output voltages (v_{do} , v_{qo}) and input currents (i_{di} , i_{qi}) are synthesized using the parameters q_d and q_q as described in Equation (9). These components must satisfy the constraints defined by expression (10) (See Figure 2).

$$\begin{aligned} v_{do} &= q_d v_{di}, \\ v_{qo} &= q_q v_{qi}, \\ i_{di} &= q_d i_{do}, \\ i_{qi} &= q_q i_{qo}. \end{aligned} \quad (9)$$

$$\begin{cases} |q_d| + |q_q| \leq 1 \\ \max\{|q_d|, |q_q|\} \leq \frac{\sqrt{3}}{2} \end{cases} \quad (10)$$

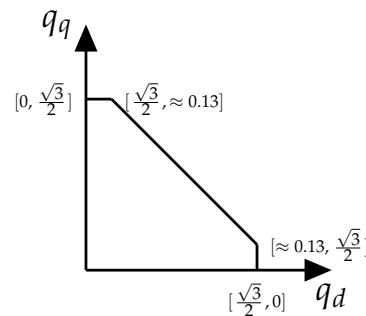


Figure 2. Accepted values for q_d and q_q for the SVD-M.

Figure 3 illustrates the signals at the input and output terminals of MC involved in the SVD-M scheme. The figure shows both the stationary reference frame ($\alpha\beta$) and the rotating reference frame (dq), which is defined by the angle θ . In this context, the input voltage vector $v_i e^{j\phi_i}$ transitions to the output voltage vector $v_o e^{j\phi_o}$, while the output current vector $i_o e^{j\gamma_o}$ transitions to the input current vector $i_i e^{j\gamma_i}$. These transformations are governed by the parameters $\theta_i, \theta_o, q_d, q_q$ as described in Equation (9).

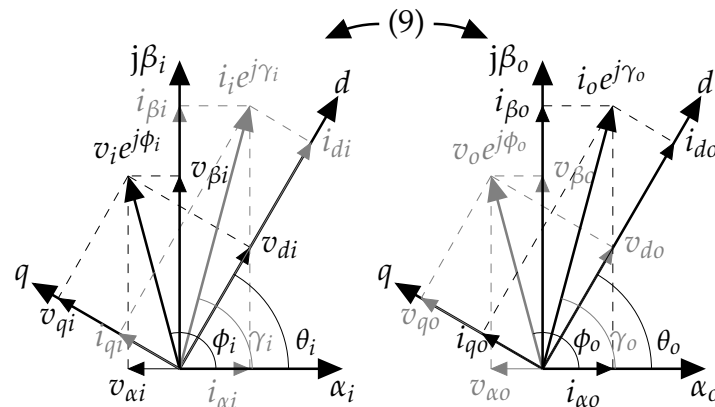


Figure 3. Input and output signal representation for the SVD-M.

3. Enhanced Singular Value Decomposition (e-SVD)

The SVD-M technique controls the operation of the MC through the parameters θ_i, θ_o, q_d , and q_q , enabling the transfer of voltage from the input to the output terminals and the flow of current from the output to the input terminals. The proposed e-SVD modulation defines these parameters to synthesize the desired output voltages and to regulate the transfer of input reactive power from the output apparent power, as detailed below.

3.1. Generation of Output Voltages

The components v_{di}, v_{qi} and i_{do}, i_{qo} depend on the signals $v_i e^{j\phi_i}$ and $i_o e^{j\gamma_o}$, in addition to the SVD-M parameters θ_i and θ_o , as described in Equations (11) and (12).

$$\begin{aligned} v_{di} &= v_i \cos(\phi_i - \theta_i), \\ v_{qi} &= v_i \sin(\phi_i - \theta_i). \end{aligned} \quad (11)$$

$$\begin{aligned} i_{do} &= i_o \cos(\gamma_o - \theta_o), \\ i_{qo} &= i_o \sin(\gamma_o - \theta_o). \end{aligned} \quad (12)$$

To establish the output voltage vector $v_o e^{j\phi_o}$ in the $\alpha\beta$ reference frame, the dq components of the output voltage, computed using Equation (9), are rotated by the angle θ_o . This transformation is represented mathematically in Equation (13).

$$\begin{bmatrix} v_{\alpha o} \\ v_{\beta o} \end{bmatrix} = \begin{bmatrix} \cos(\theta_o) & -\sin(\theta_o) \\ \sin(\theta_o) & \cos(\theta_o) \end{bmatrix} \begin{bmatrix} v_{do} \\ v_{qo} \end{bmatrix} \quad (13)$$

By expanding and rearranging Equation (13), Equation (14) is obtained, which explicitly demonstrates that the output voltage depends on the input voltage v_i , the input phase angle ϕ_i and the SVD-M parameters θ_i , θ_o , q_d and q_q .

$$\begin{bmatrix} v_{\alpha o} \\ v_{\beta o} \end{bmatrix} = \begin{bmatrix} v_i \cos(\phi_i - \theta_i) \cos(\theta_o) & -v_i \sin(\phi_i - \theta_i) \sin(\theta_o) \\ v_i \cos(\phi_i - \theta_i) \sin(\theta_o) & v_i \sin(\phi_i - \theta_i) \cos(\theta_o) \end{bmatrix} \begin{bmatrix} q_d \\ q_q \end{bmatrix} \quad (14)$$

Finally, by rewriting Equation (14) considering that the desired output voltages and angles θ_i and θ_o are now input parameters, the required values for variables q_d and q_q are obtained, as seen in expression (15).

$$\begin{bmatrix} q_d \\ q_q \end{bmatrix} = \frac{1}{v_i} \begin{bmatrix} \frac{\cos(\theta_o)}{\cos(\phi_i - \theta_i)} & \frac{\sin(\theta_o)}{\cos(\phi_i - \theta_i)} \\ \frac{-\sin(\theta_o)}{\sin(\phi_i - \theta_i)} & \frac{\cos(\theta_o)}{\sin(\phi_i - \theta_i)} \end{bmatrix} \begin{bmatrix} v_{\alpha o} \\ v_{\beta o} \end{bmatrix} \quad (15)$$

Therefore, the evaluation of the parameters q_d and q_q requires the desired output voltage, expressed through its components $v_{\alpha o}$ and $v_{\beta o}$, along with the specification of the angles θ_i and θ_o . To avoid singularities, the difference $(\phi_i - \theta_i)$ must not be equal to 0 or $\frac{\pi}{2}$, as this would cause the input voltage vector to align exactly with either the d -axis or the q -axis. The following conditions assess the constraints on the output voltage without imposing limitations on the parameter θ_o as detailed below.

3.2. Output Voltage Restrictions

The maximum achievable output voltage magnitude is determined by the constraints imposed on the parameters q_d and q_q , as well as the components v_{di} and v_{qi} . These components are functions of the angle θ_i and the known input voltage angle ϕ_i , specifically through the term $(\phi_i - \theta_i)$. In this work, the angle resulting from this difference is treated as a fixed value used to compute v_{di} and v_{qi} , making them dependent solely on the magnitude, rather than the phase angle of the input voltage vector. Consequently, by evaluating q_d and q_q the achievable output voltage magnitude can be determined, with θ_o treated as an unrestricted, free parameter. This study considers two specific cases for evaluating the output voltage magnitude: when $(\phi_i - \theta_i) = \pi/4$ and when $(\phi_i - \theta_i) = \pi/6$.

3.2.1. Case 1, $(\phi_i - \theta_i) = \frac{\pi}{4}$

In this case, the evaluation of the dq components of the input voltage yields the values presented in Equation (16), as illustrated in Figure 4a. Taking into account the constraints defined by Equation (10), the corresponding output voltage is shown in Figure 4b. The achievable output voltage lies within the region bounded by the continuous gray line. It can be observed that the voltage magnitude is greater near the dq axes compared to the regions between them. To ensure a constant maximum output voltage magnitude, the region enclosed by the dashed gray line, representing a radius equal to $\frac{v_i}{2}$, is considered.

Within this region, any output voltage vector with a magnitude less than or equal to $\frac{v_i}{2}$ can be achieved for any phase angle, without imposing restrictions on the value of θ_o .

$$v_{di} = \frac{\sqrt{2}}{2} v_i, v_{qi} = \frac{\sqrt{2}}{2} v_i. \quad (16)$$

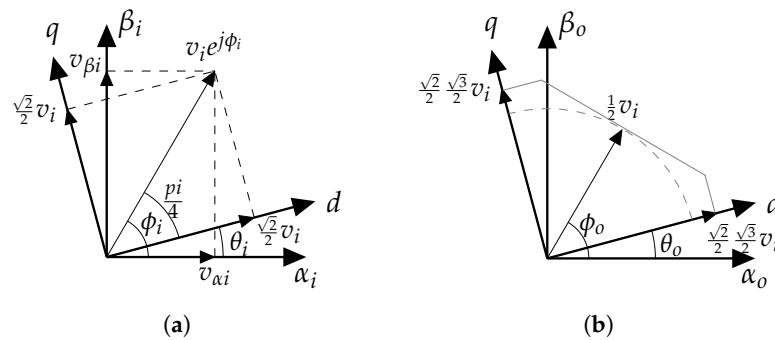


Figure 4. Voltage vectors for $(\phi_i - \theta_i) = \pi/4$, (a) input, (b) output.

3.2.2. Case 2, $(\phi_i - \theta_i) = \frac{\pi}{6}$

This case considers a scenario in which the input voltage vector is positioned closer to the dq axis compared to the previous case. As a result, the dq components of the input voltage are given by Equation (17), as shown in Figure 5a. The corresponding output voltage is presented in Figure 5b, where the achievable output voltage region is identified. The area near the d axis yields the highest voltage magnitude, reaching up to $0.75v_i$. However, as in the previous case, it is preferable to define a region with a fixed maximum output voltage magnitude. Therefore, the region bounded by the dashed gray line, representing a radius of approximately $0.433v_i$, is considered. Within this region, any output voltage vector with a magnitude less than or equal to $0.433v_i$ can be generated for any phase angle, regardless the value of θ_o .

$$v_{di} = \frac{\sqrt{3}}{2}v_i, v_{qi} = \frac{1}{2}v_i. \quad (17)$$

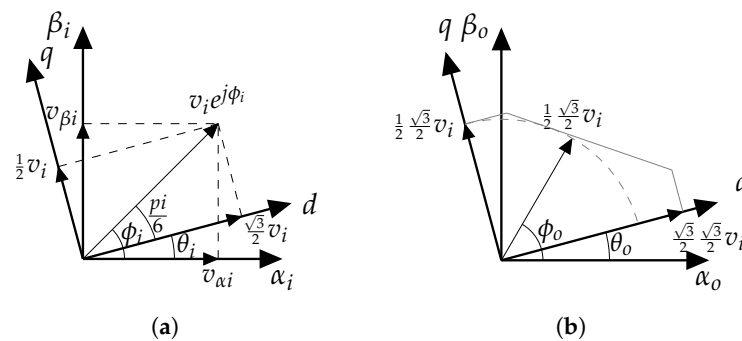


Figure 5. Voltage vectors for $(\phi_i - \theta_i) = \pi/6$, (a) input, (b) output.

3.3. Reactive Power Control at the Input Terminals.

The manipulation of reactive power at the input terminals of a MC depends on the selected modulation technique. Some approaches are based on the input–output voltage ratio (G_v), while others utilize the output active and reactive power in conjunction with modulation parameters.

A number of studies have focused on determining the maximum achievable input reactive power. For well-established MC modulation strategies, such as optimum-amplitude modulation [2] and space vector modulation (SVM), the input reactive power could be synthesized from the output reactive power, as expressed in Equation (18), and form the active power, as shown in Equation (19) [20].

$$Q_{in} \leq \frac{1 - G_v}{G_v} Q_{out} \quad (18)$$

$$Q_{in} \leq \frac{\sqrt{\frac{3}{4} - G_v^2}}{G_v} P_{out} \quad (19)$$

Additionally, the maximum value of the input reactive power can be determined through the optimization problem formulated in Equation (20), although this formulation does not yield an explicit analytical expression. Input reactive power can also be synthesized from the output apparent power, as proposed in [17], where Equation (21) was developed based on the input–output voltage ratio. Furthermore, in [39], Equation (22) was derived, showing that the input reactive power also depends on the modulation parameters. It is worth noting that these studies do not explicitly define the conditions and constraints required to determine the valid operating range for the last two expressions.

$$\begin{aligned} \max |I_i| &= \max \{\mathbf{M}^T I_o\} \\ \text{Subject to } &\begin{cases} V_o = \mathbf{M} V_i \\ G_v \leq k = |q_d| + |q_q| \leq 1 \\ \max\{|q_d|, |q_q|\} \leq \frac{\sqrt{3}}{2} \end{cases} \end{aligned} \quad (20)$$

$$|Q_i| \leq \begin{cases} 1.5 U_i I_o [q_{max} - q \sin(\phi_L)], \\ \quad \text{if } q \leq q_{max} \sin(\phi_L) \\ 1.5 U_i I_o \cos(\phi_L) \sqrt{q_{max}^2 - q^2}, \\ \quad \text{if } q_{max} \sin(\phi_L) \leq q \leq q_{max} \end{cases} \quad (21)$$

$$\begin{aligned} Q_{in} \leq V_L I_L & \left(\tan(\delta_s) \cos(\delta_L) \cos(\delta_L - \phi_L) \right. \\ & \left. - \frac{1}{\tan(\delta_s)} \sin(\delta_L) \sin(\delta_L - \phi_L) \right) \end{aligned} \quad (22)$$

3.4. e-SVD Reactive Power

According with the instantaneous power theory, the relationships governing the input and output power of the MC in the dq reference frame are expressed by Equations (23) and (24), respectively. Likewise, assuming an ideal (lossless) MC, the active power at the input and output terminals is equal, i.e., $p_{out} = p_{in}$.

$$\begin{aligned} p_{in} &= \frac{3}{2} (v_{di} i_{di} + v_{qi} i_{qi}), \\ q_{in} &= \frac{3}{2} (v_{qi} i_{di} - v_{di} i_{qi}). \end{aligned} \quad (23)$$

$$\begin{aligned} p_{out} &= \frac{3}{2} (v_{do} i_{do} + v_{qo} i_{qo}), \\ q_{out} &= \frac{3}{2} (v_{qo} i_{do} - v_{do} i_{qo}) \end{aligned} \quad (24)$$

Subsequently, the expression for the input reactive power q_{in} is reformulated using the expressions introduced in Section 3.1. After applying appropriate simplifications, this leads to Equation (25). The resulting expression depends on the modulation parameters (θ_i, θ_o) , the electrical output variables $(v_o, \phi_o, i_o, \gamma_o)$, and the input voltage phase angle (ϕ_i) .

$$\begin{aligned} q_{in} &= \frac{3}{2} [v_o i_o [\tan(\phi_i - \theta_i) \cos(\gamma_o - \theta_o) \cos(\phi_o - \theta_o) \\ & \quad + \cot(\phi_i - \theta_i) \sin(\gamma_o - \theta_o) \sin(\phi_o - \theta_o)]] \end{aligned} \quad (25)$$

In order to derive a simplified expression and define the reactive power gain (G_q), Equation (25) is rewritten as Equation (26), which establishes the relationship between the input reactive power and the output apparent power. This gain can be adjusted through

the nonlinear dependence between the angles θ_i and θ_o . It is important to note that (G_q) is a periodic function of θ_o with a period of π .

$$G_q = \frac{q_{in}}{s_{out}} = \tan(\phi_i - \theta_i) \frac{\cos(\gamma_o + \phi_o - 2\theta_o) + \cos(\gamma_o - \phi_o)}{2} + \cot(\phi_i - \theta_i) \frac{\cos(\gamma_o - \phi_o) - \cos(\gamma_o + \phi_o - 2\theta_o)}{2} \quad (26)$$

Accordingly, by incorporating the output voltage constraints discussed in Section 3.2, and treating the parameter θ_o as an unrestricted variable, the expression for the reactive power gain in Equation (26) can be modified as detailed below.

3.4.1. Case 1, $(\phi_i - \theta_i) = \frac{\pi}{4}$

By applying the conditions outlined in Section 3.2.1 to Equation (26), a simplified analytical expression for the reactive power gain is obtained, as shown in Equation (27). To further refine this expression, a change of variable is introduced in Equation (28), where the new variable Θ_o is defined in Equation (29). Substituting this into Equation (27) yields Equation (30), which expresses the reactive power gain in terms of Θ_o . This final expression reaches its maximum and minimum values of 1 and -1 , respectively, under the conditions $\Phi_o - 2\Theta_o = 0$ and $\Phi_o - 2\Theta_o = -\pi$. Since θ_o is treated as a free and unrestricted variable, and the parameters ϕ_o and γ_o are known, any value within the range of $[-1, 1]$ can be assigned to G_q .

The graphical representation of Equation (30) is provided in Figure 6, where Figure 6a shows a three-dimensional view of the model and Figure 6b presents the front view. These figures illustrate the feasible region for G_q across all values of Θ_o .

$$G_q = \cos(\phi_o + \gamma_o - 2\theta_o) \quad (27)$$

$$\theta_o = \Theta_o + \phi_o \quad (28)$$

$$\Phi_o = \phi_o - \gamma_o \quad (29)$$

$$G_q = \cos(\Phi_o - 2\Theta_o) \quad (30)$$

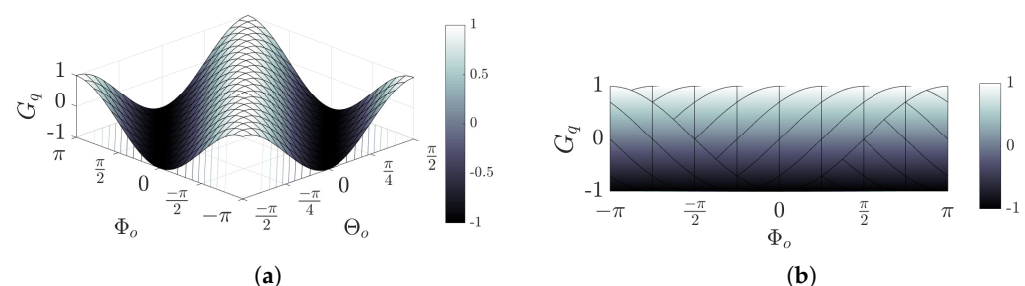


Figure 6. Reactive power gain for $G_q = \cos(\Phi_o - 2\Theta_o)$, Equation (30). (a) 3D view, (b) front view.

In the case where $(\phi_i - \theta_i) = -\frac{\pi}{4}$, the reactive power gain represented by Equation (30) changes to expression (31), resulting in the inverted version of the case where $(\phi_i - \theta_i) = \frac{\pi}{4}$.

$$G_q = -\cos(\Phi_o - 2\Theta_o) \quad (31)$$

3.4.2. Case 2, $(\phi_i - \theta_i) = \frac{\pi}{6}$

Applying the conditions established in Section 3.2.2, Equation (26) can be reformulated as Equation (32). Similarly, by using the change of variable introduced in Equation (28),

the reactive power gain G_q can be expressed as Equation (33). In this case, the maximum and minimum values of G_q are $\sqrt{3}/3$ and $-\sqrt{3}/3$, respectively. The maximum value is achieved when $\theta_o = \pm\pi/2$ and $\Phi_o = \pm\pi$, and the minimum occurs when $\theta_o = \pm\pi/2$ and $\Phi_o = 0$.

Expression (32) is presented in a three-dimensional representation in Figure 7a and in a frontal projection in Figure 7b. These figures illustrate that the maximum and minimum values of G_q occur only under specific conditions, rather than across the entire domain.

$$G_q = \frac{2 \cos(\phi_o + \gamma_o - 2\theta_o) - \cos(\phi_o - \gamma_o)}{\sqrt{3}} \quad (32)$$

$$G_q = \frac{2 \cos(\Phi_o - 2\Theta_o) - \cos(\Phi_o)}{\sqrt{3}} \quad (33)$$

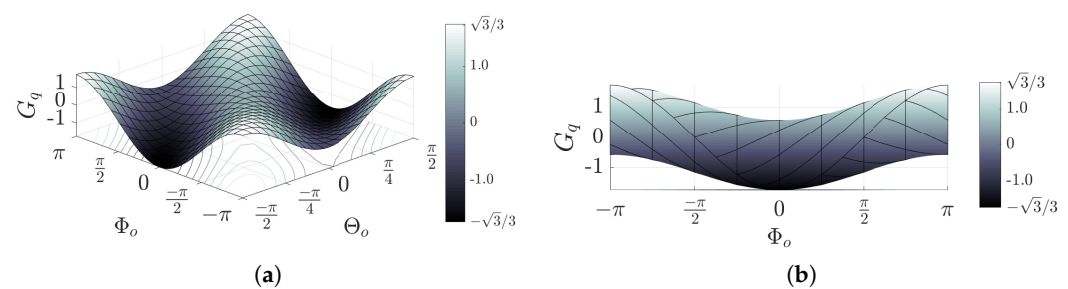


Figure 7. Reactive power gain $G_q = \frac{2 \cos(\Phi_o - 2\Theta_o) - \cos(\Phi_o)}{\sqrt{3}}$ (33). (a) 3D view, (b) front view.

3.4.3. Case 3, $(\phi_i - \theta_i) = -\frac{\pi}{6}$

This case exhibits the same voltage constraints as case 2; however, for reactive power gain, expression (26) is reformulated as shown in Equation (34). The corresponding three-dimensional and frontal views of expression (34) are presented in Figures 8a and 8b, respectively. This visualizations clearly reveal an inversion in comparison to those depicted in Figure 7.

$$G_q = - \left(\frac{2 \cos(\Phi_o - 2\Theta_o) - \cos(\Phi_o)}{\sqrt{3}} \right) \quad (34)$$

An analysis of the G_q expressions derived for cases 2 and 3 reveals that they can complement each other to achieve higher G_q values, considering that the parameter $(\phi_i - \theta_i)$ is adjustable. Figure 9 displays the overlapping regions generated by expressions (33) and (34), highlighting that a broader set of conditions for Φ_o can be considered to enhance G_d . However, in this scenario, it is necessary to know the specific value of Φ_o in order to determine the corresponding angular difference $(\phi_i - \theta_i)$.

Based on the preceding analysis, it can be concluded that the proposed e-SVD modulation strategy effectively defines the conditions necessary for generation a controllable output voltage and achieving a reactive power gain. This is accomplished by assigning appropriate values to q_d, q_q, θ_i , and θ_o derived from the electrical signals of the MC and the desired output voltage. The implementing procedure for the e-SVD strategy is detailed in Algorithm 1 below.

The modulation techniques presented demonstrate the capability to control reactive power at the input terminal, albeit with limitations, as this control depends on the voltage ratio and electrical parameters such as the power factor, and the active and reactive components of the output power. In contrast, the proposed approach enables input reactive power control based on the output apparent power, utilizing a mathematically defined reactive power gain.

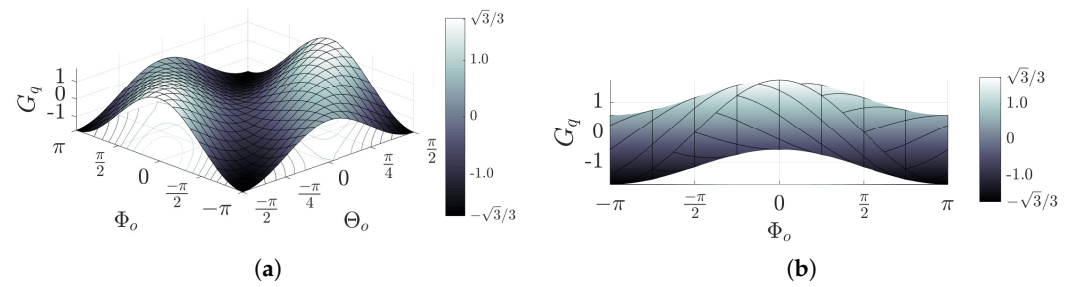


Figure 8. Reactive power gain $G_q = -\left(\frac{2\cos(\Phi_o - 2\Theta_o) - \cos(\Phi_o)}{\sqrt{3}}\right)$, Equation (34). (a) 3D view, (b) front view.

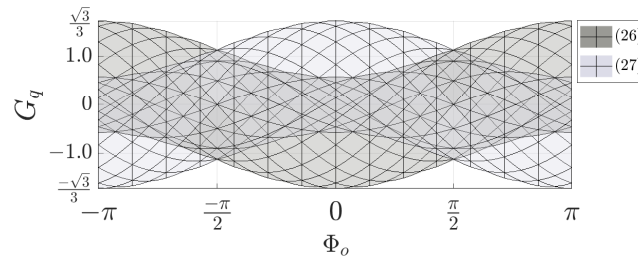


Figure 9. Overlay of the expressions (33) and (34).

Algorithm 1 e-SVD modulation

1: Inputs:

$v_{i\alpha}, v_{i\beta} \leftarrow v_a, v_b, v_c$
 $v_{o\alpha}, v_{o\beta} \leftarrow v_A, v_B, v_C$
 $i_{o\alpha}, i_{o\beta} \leftarrow i_A, i_B, i_C$

▷ Input voltage measurement
 ▷ Output voltage setting
 ▷ Output current measurement

2: Initialize:

3: $\phi_i \leftarrow \text{atan2}(v_{\alpha i}, v_{\beta i})$
 4: $v_i \leftarrow \sqrt{v_{\alpha i}^2 + v_{\beta i}^2}$
 5: $\phi_o \leftarrow \text{atan2}(v_{\alpha o}, v_{\beta o})$
 6: $v_o \leftarrow \sqrt{v_{\alpha o}^2 + v_{\beta o}^2}$
 7: $\gamma_o \leftarrow \text{atan2}(i_{\alpha o}, i_{\beta o})$
 8: $i_o \leftarrow \sqrt{i_{\alpha o}^2 + i_{\beta o}^2}$

9: Procedure:

10: $\theta_i \leftarrow$ Set value for $(\phi_i - \theta_i)$
 11: $\theta_o \leftarrow$ Set value for θ_o
 12: $q_d, q_q \leftarrow$ evaluating Equation (15)
 13: $M \leftarrow q_d, q_q, \theta_i, \theta_o$

▷ e.g., $\theta_i = \phi_i + \pi/6$
 ▷ evaluating G_q with γ_o, ϕ_o

▷ Conventional SVD-M

4. Experimental Results

To validate the effectiveness of the proposed method, experiments were conducted using a low-power MC prototype integrated into a small-scale test power system. The objective of the test was to demonstrate control of the reactive power at the MC input terminals based on the apparent output power, while maintaining constant output voltage and current waveforms.

The experimental setup emulates a low-scale three-phase transmission system comprising a line impedance, a coupling transformer, and the MC equipped with input and output filters. The MC prototype was assembled using HGT1S1260A4DS IGBT modules. The filter components were selected as follows: $L_i = 1.0$ [mH], $C_i = 6.8$ [μ F], $L_o = 0.56$ [mH], and $C_o = 6.8$ [μ F]. The peak input and output voltages were 100 [V] and 30 [V], respectively,

with a transformer turns ratio of 2.3. A schematic of the experimental setup is shown in Figure 10, while Figures 11 and 12 display the physical implementation and the MC prototype, respectively. Power measurements were acquired using the Labvolt Data Acquisition and Control Interface (DACI), with signals recorded at the input terminals of the MC input filter and at the secondary side of the coupling transformer.

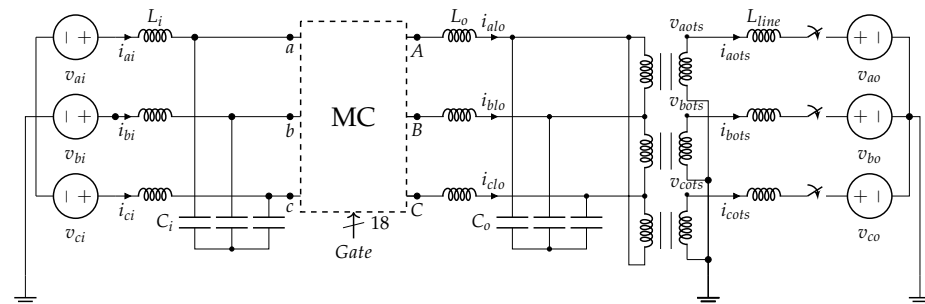


Figure 10. Experimental test system setup for proposal validation.

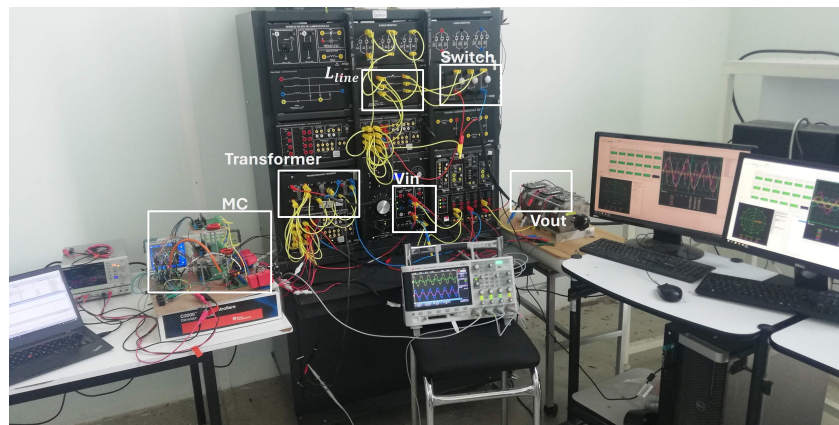


Figure 11. Experimental implementation of the power system under test.

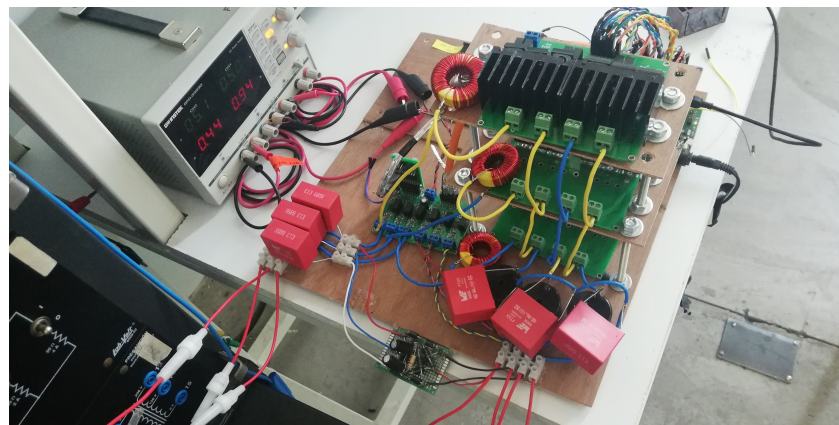


Figure 12. MC prototype.

MC control algorithms are implemented in the microcontroller TMS320F28379D, produced by Texas Instruments (Dallas, TX, USA), along with the four-step current commutation strategy within the same device as described in [40]. The voltages v_{od} and v_{oq} are set by a proportional integral controller to achieve a voltage at the transformer secondary terminals with a deviation from the output voltage of $\Delta v_d = 3$ and $\Delta v_q = -2$, resulting in an active and reactive power of 23.85 [w] and -13.93 [var]. Consequently, the resulting power factor causes Φ_o have a value of -0.3877 [rad] from the experimental data.

The experimental validation of the proposed method comprises two tests. The first evaluates the input reactive power response to variations in the control parameter, aiming

to validate the concept of reactive power gain. The second test examines the transient behavior of the voltage and current signals in response to changes in the control parameter. These tests are described below.

4.1. Test 1

The first experimental test considers a variation in the variable $\Delta\theta_i$, assigning the values of $\frac{\pi}{4}$, $-\frac{\pi}{4}$, $\frac{\pi}{6}$, and $-\frac{\pi}{6}$ during four time periods defined as cycles I, II, III, and IV, respectively. For each cycle, $\Delta\theta_o$ changes in steps of $\frac{\pi}{16}$ from $-\frac{\pi}{2}$ to $\frac{\pi}{2}$, corresponding to one reactive gain period. Figure 13 presents the active, reactive, and apparent input powers, while Figure 14 shows the active, reactive, and apparent output powers at the secondary terminals of the coupling transformer. It is worth noting that the MC low efficiency in terms of active power is due to its operation at the low power level set by the transistors. Additionally, reactive input power exhibits an offset due to the capacitor of the input filter, which must be eliminated in order to evaluate the reactive power at the input terminals of the MC, Q_{inmc} .

In Figures 13 and 14, during cycle I, $\Delta\theta_o$ starts with a value of $-\frac{\pi}{2}$, then it changes for a whole period. For the other cycles, variable $\Delta\theta_o$ changes as indicated until completing two periods. It must be noted that during cycles I and II, the reactive input power signals show opposite behavior concerning the mean value. Therefore, the controllable input reactive power range is the same in both cycles. Finally, during cycles III and IV, the input reactive power curves are scaled and shifted up and down with respect to cycles I and II. On the other hand, it can be observed that unitary power factor operation is reached only in cycle IV, and a lower value of -100 [var] is achieved only in cycle III.

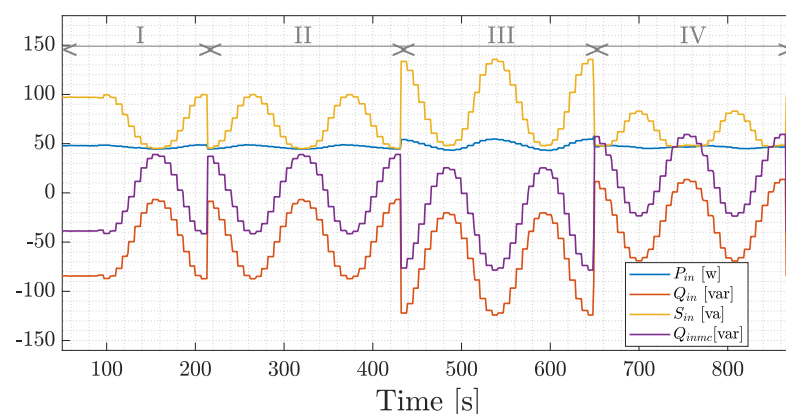


Figure 13. Experimental measurements for active, reactive, and apparent input power.

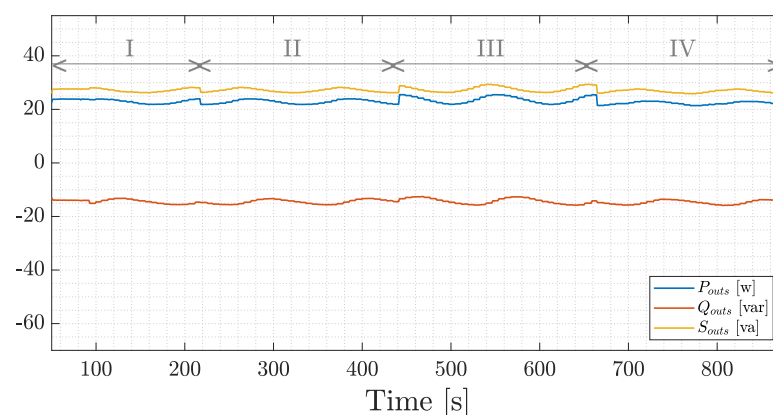


Figure 14. Experimental measurements for active, reactive, and apparent input power at the transformer secondary terminals.

The evaluation for each section considers the theoretical gain and the normalized experimental gain to validate the proposal, where the normalization helps to represent the gain variation from $\Delta\theta_o$.

4.1.1. Section I

For $\Delta\theta_i = \frac{\pi}{4}$ and $\Phi_o = -0.3877$ [rad], the reactive power gain in Equation (30) takes the values shown in Figure 15. The experimental powers in the system are shown in Figure 16, where the input–output active power waveforms are similar, while the input reactive and apparent power changes as does the parameter $\Delta\theta_o$. Calculating the ratio $\frac{Q_{inmc}}{S_{ots}}$ to normalize the gain by dividing it by the max value (1.48), results in the values shown in Figure 17, which corroborate that the gain exhibits a similar behavior as the theoretical result.

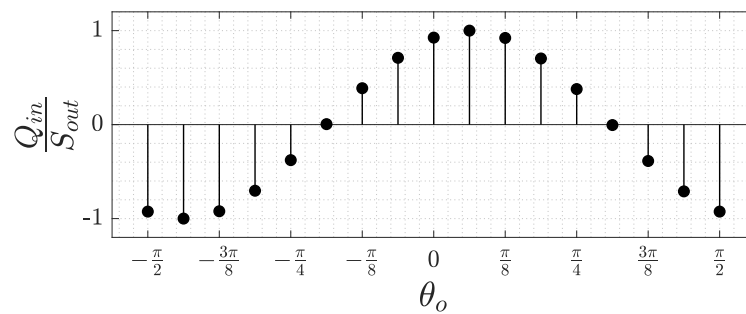


Figure 15. Theoretical reactive power gain at $\Delta\theta_i = \frac{\pi}{4}$ and $\Phi_o = -0.3877$ [rad].

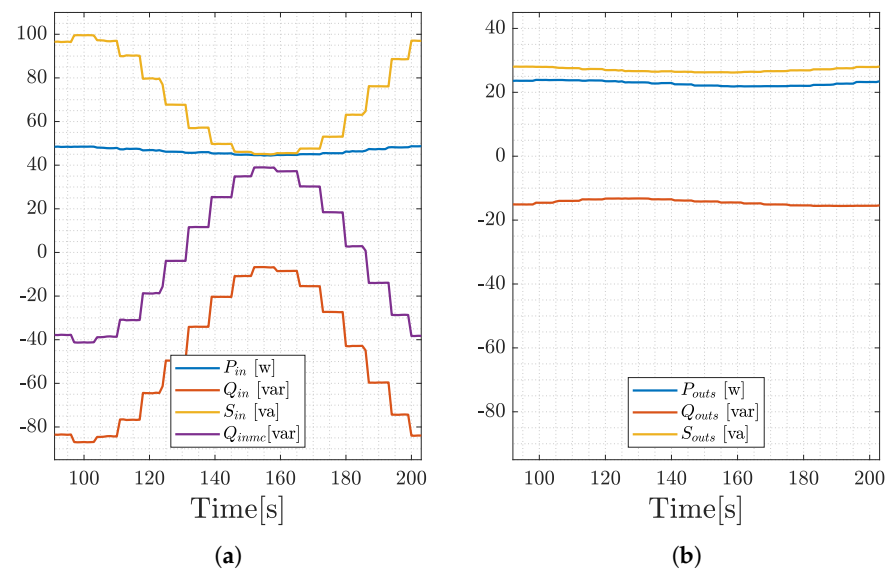


Figure 16. Active, reactive and apparent powers while $\Delta\theta_i = \frac{\pi}{4}$ at the (a) input and (b) output.

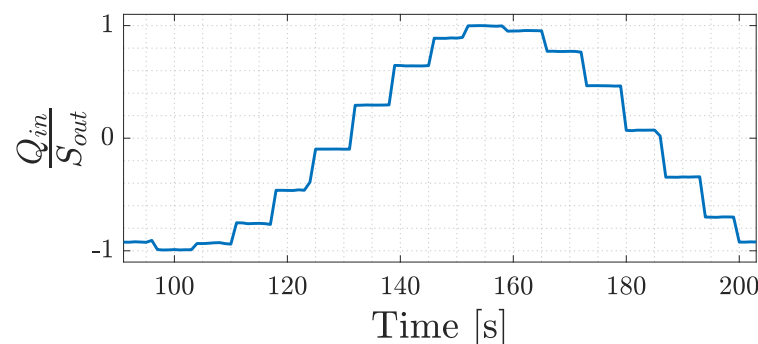


Figure 17. Experimental normalized reactive power gain at $\Delta\theta_i = \frac{\pi}{4}$ and $\Phi_o = -0.3877$ [rad].

4.1.2. Section II

For $\Delta\theta_i = -\frac{\pi}{4}$ and $\Phi_o = -0.3877$ [rad], the reactive power gain takes the values shown in Figure 18, resulting in a negative version of the previous case. The experimental powers in the system are shown in Figure 19, where the input reactive and apparent power present the opposite behavior to that shown in the previous case. Again, calculating the ratio $\frac{Q_{inmc}}{S_{ots}}$ to normalize the gain by dividing it by the max value (1.48), indicated in Section 4.1.1, results in the gain shown in Figure 20, validating the theoretical result.

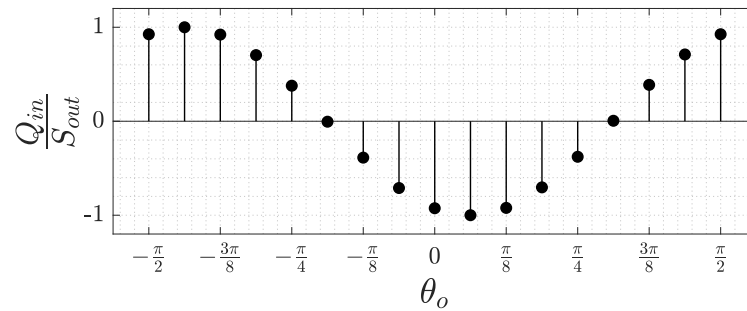


Figure 18. Theoretical reactive power gain at $\Delta\theta_i = -\frac{\pi}{4}$ and $\Phi_o = -0.3877$ [rad].

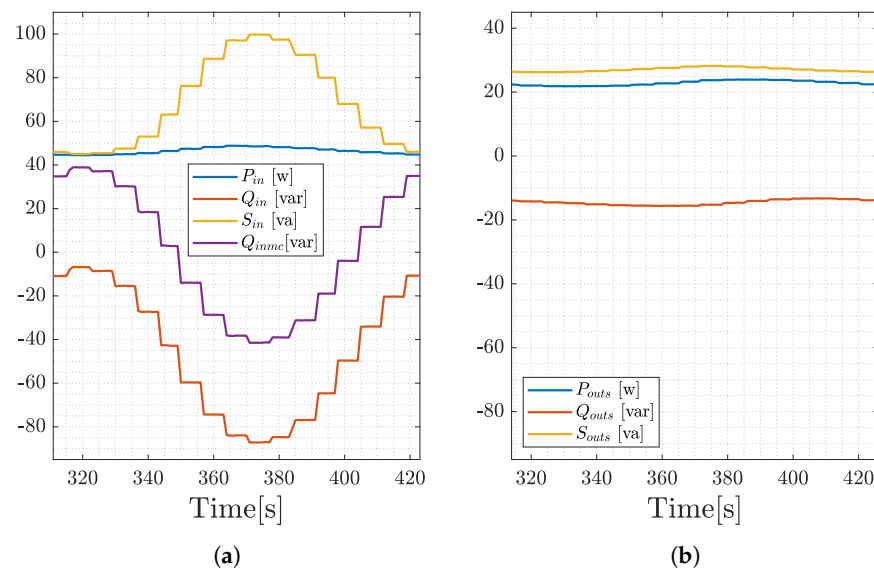


Figure 19. Active, reactive, and apparent powers while $\Delta\theta_i = -\frac{\pi}{4}$ at the (a) input and (b) output.

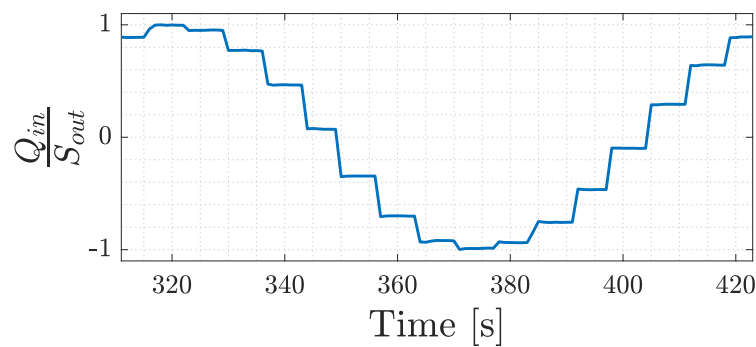


Figure 20. Experimental normalized reactive power gain at $\Delta\theta_i = -\frac{\pi}{4}$ and $\Phi_o = -0.3877$ [rad].

4.1.3. Section III

Considering for this case that $\Delta\theta_i = -\frac{\pi}{6}$ and $\Phi_o = -0.3877$ [rad], the reactive power gain in Equation (32) takes the values shown in Figure 21. The experimental powers in the system are shown in Figure 22b, where the input reactive power is not centered at zero vars due to the presence of a negative offset. As in the previous cases, the gain is normalized using the ratio value $\frac{Q_{inmc}}{S_{ots}}$ and dividing it by 1.48. The result is the gain curve shown in Figure 23, which presents a waveform similar to the theoretical prediction with only a small difference in the maximum values.

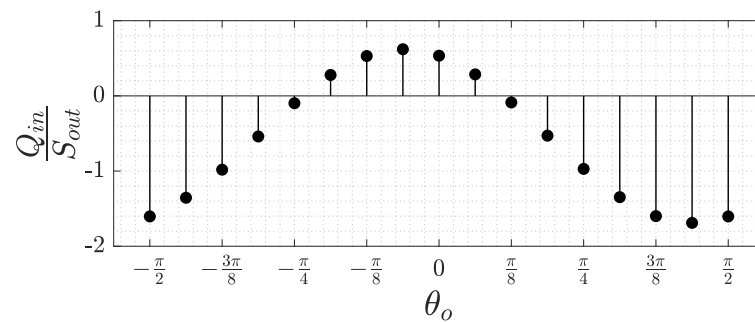


Figure 21. Theoretical reactive power gain at $\Delta\theta_i = \frac{\pi}{6}$ and $\Phi_o = -0.3877$ [rad].

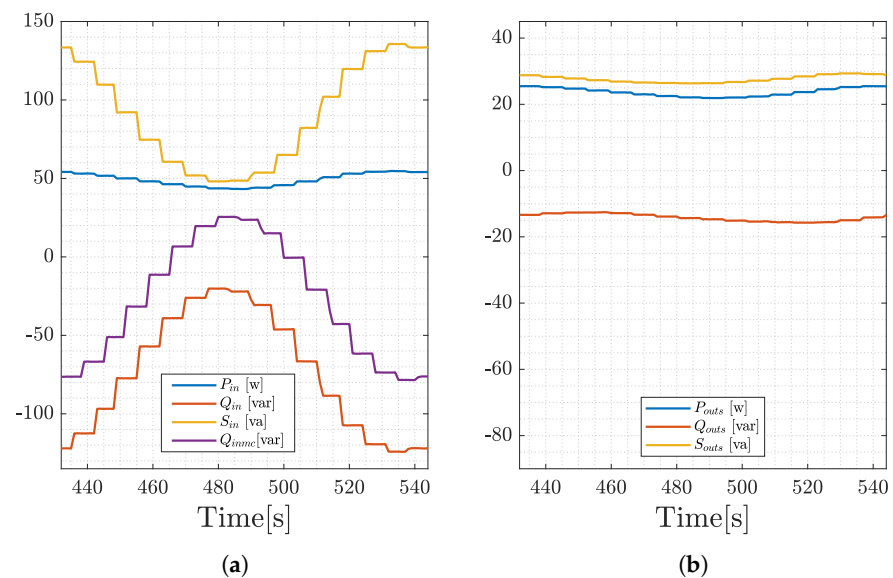


Figure 22. Active, reactive, and apparent powers while $\Delta\theta_i = \frac{\pi}{6}$ at the (a) input and (b) output.

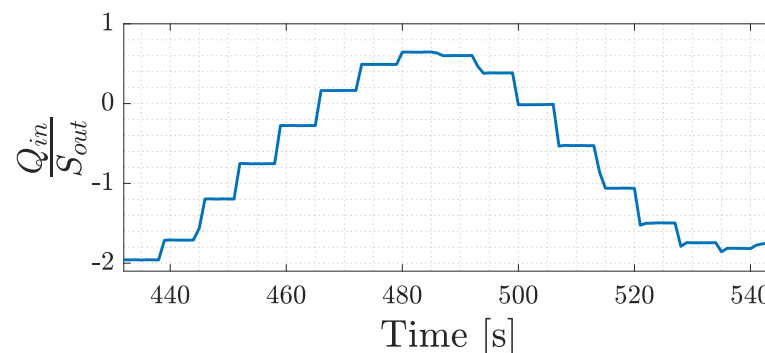


Figure 23. Experimental normalized reactive power gain at $\Delta\theta_i = \frac{\pi}{6}$ and $\Phi_o = -0.3877$ [rad].

4.1.4. Section IV

In this last case of test 1, $\Delta\theta_i = -\frac{\pi}{6}$ and $\Phi_o = -0.3877$ [rad], the reactive power gain in Equation (32) takes the values shown in Figure 24. Figure 25b shows the experimental powers in the system, where the input reactive power has a positive offset value, achieving negative and positive values. Normalizing the gain as in all previous cases, Figure 26 is obtained. Again, this curve presents a waveform similar to the theoretical gain, except for small differences in maximum values.

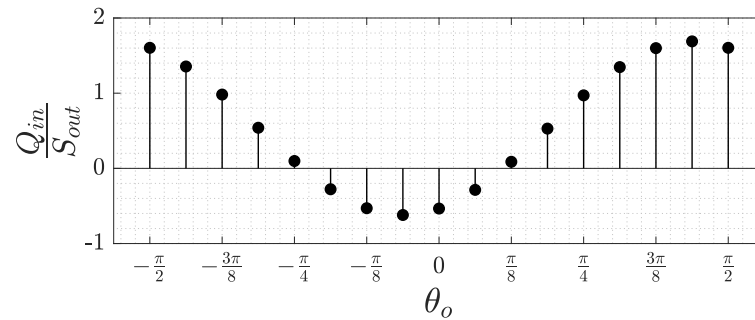


Figure 24. Theoretical reactive power gain at $\Delta\theta_i = -\frac{\pi}{6}$ and $\Phi_o = -0.3877$ [rad].

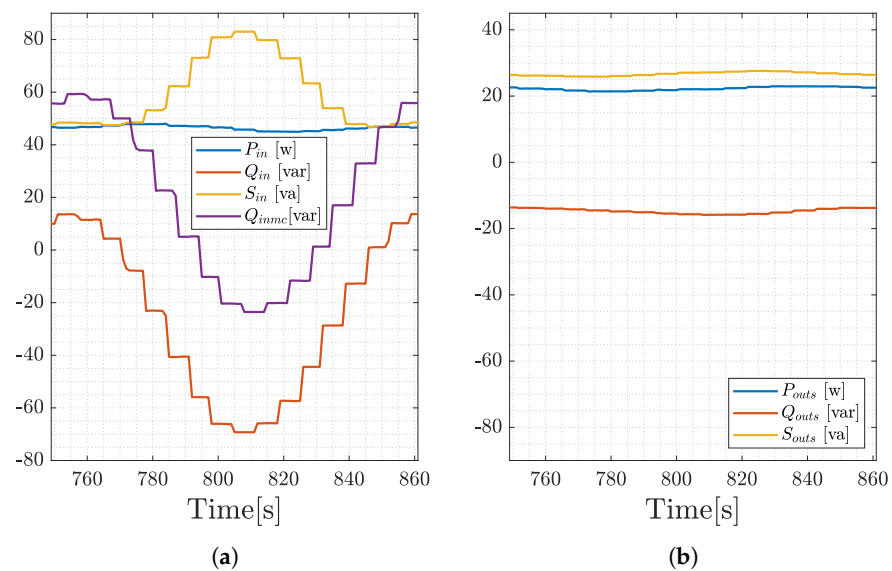


Figure 25. Active, reactive, and apparent powers while $\Delta\theta_i = -\frac{\pi}{6}$ at the (a) input and (b) output.

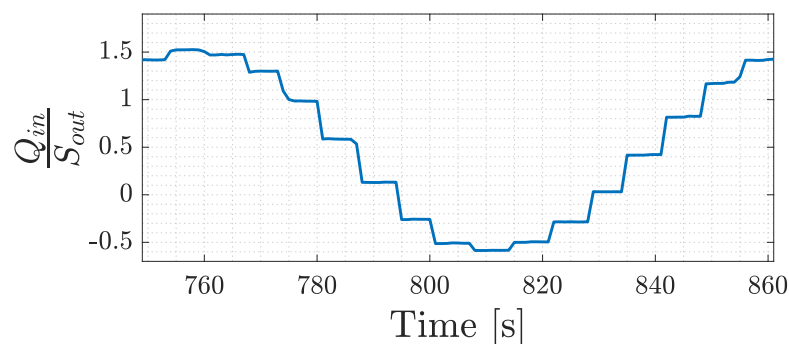


Figure 26. Experimental normalized reactive power gain at $\Delta\theta_i = -\frac{\pi}{6}$ and $\Phi_o = -0.3877$ [rad].

As a general conclusion from the first test, the results confirm that the proposed technique effectively regulates the input reactive power by adjusting the reactive power gain, as detailed in Section 3.4. This evaluation required extended measurement periods

to accurately capture the active, reactive, and apparent power values. To complement this analysis and assess the system's instantaneous response, a second test was conducted over a shorter time frame, focusing on the transition between two operating conditions previously examined in Section 4.1.4. during the first test.

4.2. Test II

As a second test, voltage and current signals at the input and output terminals of the experimental power system are measured under the variation in $\Delta\theta_o$ at values $-\frac{\pi}{2}$ and $\frac{\pi}{16}$, while $\Delta\theta_i$ has the fixed value of $-\frac{\pi}{6}$. The value of $-\frac{\pi}{2}$ in $\Delta\theta_o$ corresponds to the start of cycle IV in Figure 13, with a positive value of input reactive power. On the other hand, the value of $\frac{\pi}{16}$ corresponds to the time of 700 [s] in Figure 13, where there is a negative value in the input reactive power. Hence, test II evaluates the system's response to variations between positive and negative reactive power values.

Figures 27 and 28 show phase *c* voltage signals of the experimental power system, when the value of $\Delta\theta_o$ changes from $\frac{\pi}{16}$ to $-\frac{\pi}{2}$ and from $-\frac{\pi}{2}$ to $\frac{\pi}{16}$, respectively. The value $-\frac{\pi}{2}$ corresponds to the operation where the current lags the voltage, i.e., inductive behavior, while for $\frac{\pi}{16}$ the current leads the voltage, i.e., capacitive behavior.

In these figures, it can be observed that the variation in the parameter $\Delta\theta_o$ instantly modifies the input current waveform, while the output signals remain invariant. Additionally, it is appreciated that the magnitude of the input current also changes, as does the controllable value. Similarly, the power factor changes according to the phase angle difference between the input voltage and current signals.

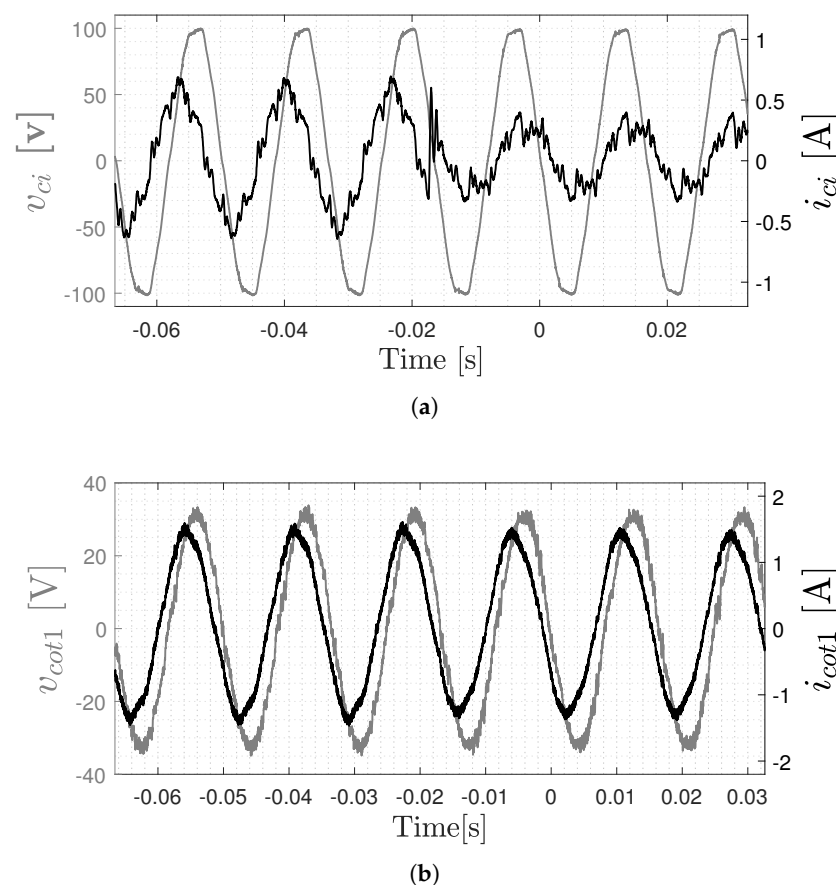


Figure 27. Transition for $\Delta\theta_o$ from $\frac{\pi}{16}$ to $-\frac{\pi}{2}$ (voltage on the left axis (gray), current on the right axis (black)). (a) Input and (b) output.

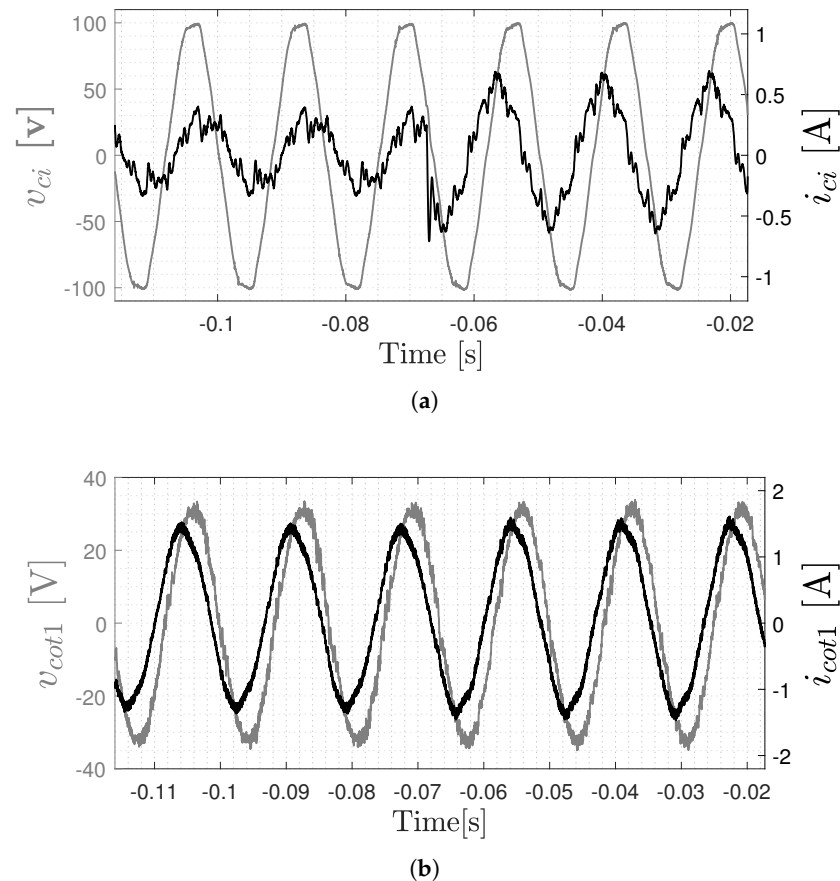


Figure 28. Transition for $\Delta\theta_o$ from $-\frac{\pi}{2}$ to $-\frac{\pi}{16}$ (voltage on the left axis (gray), current on the right axis (black)). (a) Input and (b) output.

5. Conclusions

This paper proposed the enhanced Singular Value Decomposition modulation technique (e-SVD), a set of algorithms that determine the SVD modulation parameters required to control the MC operation. The e-SVD strategy allows the generation of a controlled output voltage in the MC, with limited magnitude but for any phase angle, in addition to a controllable input reactive power dependent on the apparent output power. Furthermore, an important contribution of this work has been the deduction of a deterministic mathematical expression for the relationship between the input reactive power and the output apparent power in the MC, which allows precise control of the input reactive power value within a continuous region with established limits. This contribution is expected to facilitate the integration of the MC into a wider range of power compensation applications.

Finally, experimental tests were performed on a low-scale prototype of a power transmission system to validate the effectiveness of the proposal. Through a series of tests, it was verified that it is possible to achieve the operational objectives set for the MC in terms of input reactive power control and output voltage generation.

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References

1. Alesina, A.; Venturini, M. Solid-state power conversion: A Fourier analysis approach to generalized transformer synthesis. *IEEE Trans. Circuits Syst.* **1981**, *28*, 319–330. [\[CrossRef\]](#)
2. Alesina, A.; Venturini, M. Analysis and design of optimum-amplitude nine-switch direct AC-AC converters. *IEEE Trans. Power Electron.* **1989**, *4*, 101–112. [\[CrossRef\]](#)
3. Wheeler, P.; Rodriguez, J.; Clare, J.; Empringham, L.; Weinstein, A. Matrix converters: A technology review. *IEEE Trans. Ind. Electron.* **2002**, *49*, 276–288. [\[CrossRef\]](#)
4. Simon, O.; Mahlein, J.; Muenzer, M.N.; Bruckmann, M. Modern solutions for industrial matrix-converter applications. *IEEE Trans. Ind. Electron.* **2002**, *49*, 401–406. [\[CrossRef\]](#)
5. Gong, Z.; Zheng, X.; Zhang, H.; Dai, P.; Wu, X.; Li, M. A QPR-Based Low-Complexity Input Current Control Strategy for the Indirect Matrix Converters with Unity Grid Power Factor. *IEEE Access* **2019**, *7*, 38766–38777. [\[CrossRef\]](#)
6. Zhang, J.; Li, L.; Dorrell, D.G. Control and Applications of Direct Matrix Converters: A Review. *Chin. J. Electr. Eng.* **2018**, *4*, 18–27. [\[CrossRef\]](#)
7. Ichiki, S. New Modulation Method for Three-Phase Matrix Converter: Mathematical Theory and Demonstration Test Results. *IEEE Trans. Ind. Electron.* **2022**, *69*, 3959–3966. [\[CrossRef\]](#)
8. Xiong, W.; Xie, Q.; Lin, J.; Yuan, W.; Dan, H.; Sun, Y.; Su, M. Algebraic Modulation Strategy for Direct Matrix Converter with Adaptation to Unbalanced Grids. *IEEE Trans. Power Electron.* **2024**, *39*, 6204–6214. [\[CrossRef\]](#)
9. Raju, S.; Erickson, M.; Mohan, N. Direct Modulation of Matrix Converter with Optimized Switch Transitions. *IEEE Trans. Ind. Electron.* **2025**, *72*, 6809–6820. [\[CrossRef\]](#)
10. Malekjamshidi, Z.; Jafari, M.; Zhu, J.; Xiao, D. Comparative analysis of input power factor control techniques in matrix converters based on model predictive and space vector control schemes. *IEEE Access* **2019**, *7*, 139150–139160. [\[CrossRef\]](#)
11. Hamouda, M.; Blanchette, H.F.; Al-Haddad, K. Unity Power Factor Operation of Indirect Matrix Converter Tied to Unbalanced Grid. *IEEE Trans. Power Electron.* **2016**, *31*, 1095–1107. [\[CrossRef\]](#)
12. Tuyen, N.D.; Dzung, P.Q. Space Vector Modulation for an Indirect Matrix Converter with Improved Input Power Factor. *Energies* **2017**, *10*, 588. [\[CrossRef\]](#)
13. Nguyen, H.M.; Lee, H.H.; Chun, T.W. Input Power Factor Compensation Algorithms Using a New Direct-SVM Method for Matrix Converter. *IEEE Trans. Ind. Electron.* **2011**, *58*, 232–243. [\[CrossRef\]](#)
14. Sun, Y.; Li, X.; Su, M.; Wang, H.; Dan, H.; Xiong, W. Indirect matrix converter-based topology and modulation schemes for enhancing input reactive power capability. *IEEE Trans. Power Electron.* **2015**, *30*, 4669–4681. [\[CrossRef\]](#)
15. Mondal, S.; Kastha, D. Input Reactive Power Controller with a Novel Active Damping Strategy for a Matrix Converter Fed Direct Torque Controlled DFIG for Wind Power Generation. *IEEE J. Emerg. Sel. Top. Power Electron.* **2020**, *8*, 3700–3711. [\[CrossRef\]](#)
16. Qiu, L.; Li, Y.; Member, S.; Huang, X. Modulation Technique for a 3×5 Matrix Converter Achieving a Maximum Input Reactive Power Range Based on Load Information. *IEEE Trans. Power Electron.* **2021**, *36*, 10409–10419. [\[CrossRef\]](#)
17. Li, X.; Su, M.; Sun, Y.; Dan, H.; Xiong, W. Modulation Strategy Based on Mathematical Construction for Matrix Converter Extending the Input Reactive Power Range. *IEEE Trans. Power Electron.* **2014**, *29*, 654–664. [\[CrossRef\]](#)
18. Schafmeister, F.; Kolar, J.W. Novel hybrid modulation schemes significantly extending the reactive power control range of all matrix converter topologies with low computational effort. *IEEE Trans. Ind. Electron.* **2012**, *59*, 194–210. [\[CrossRef\]](#)
19. Li, X.; Sun, Y.; Zhang, J.; Su, M.; Huang, S. Modulation Methods for Indirect Matrix Converter Extending the Input Reactive Power Range. *IEEE Trans. Power Electron.* **2017**, *32*, 4852–4863. [\[CrossRef\]](#)
20. Hojabri, H.; Mokhtari, H.; Chang, L. Reactive power control of permanent-magnet synchronous wind generator with matrix converter. *IEEE Trans. Power Deliv.* **2013**, *28*, 575–584. [\[CrossRef\]](#)
21. Schauder, C.D. Theory and Design of a 30-hp Matrix Converter. *IEEE Trans. Ind. Appl.* **1992**, *28*, 546–551. [\[CrossRef\]](#)
22. Nielsen, P.; Blaabjerg, F.; Pedersen, J.K. New protection issues of a matrix converter: Design considerations for adjustable-speed drives. *IEEE Trans. Ind. Appl.* **1999**, *35*, 1150–1161. [\[CrossRef\]](#)
23. Podlesak, T.F.; Katsis, D.C.; Wheeler, P.W.; Clare, J.C.; Empringham, L.; Bland, M. A 150-kVA vector-controlled matrix converter induction motor drive. *IEEE Trans. Ind. Appl.* **2005**, *41*, 841–847. [\[CrossRef\]](#)

24. Orser, D.; Mohan, N. A Matrix Converter Ride-Through Configuration Using Input Filter Capacitors as an Energy Exchange Mechanism. *IEEE Trans. Power Electron.* **2015**, *30*, 4377–4385. [\[CrossRef\]](#)
25. Stella, F.; Yousefi-Talouki, A.; Odhano, S.; Pellegrino, G.; Zanchetta, P. An Accurate Self-Commissioning Technique for Matrix Converters Applied to Sensorless Control of Synchronous Reluctance Motor Drives. *IEEE J. Emerg. Sel. Top. Power Electron.* **2019**, *7*, 1342–1351. [\[CrossRef\]](#)
26. Liu, X.; Loh, P.C.; Wang, P.; Blaabjerg, F.; Tang, Y.; Al-Ammar, E.A. Distributed generation using indirect matrix converter in reverse power mode. *IEEE Trans. Power Electron.* **2013**, *28*, 1072–1082. [\[CrossRef\]](#)
27. Zhang, J.; Wang, H.; Liu, G.; Tian, G.; Rivera, M.; Wheeler, P. Virtual Synchronous Generator Control Strategy of Grid-Forming Matrix Converter for Renewable Power Generation. *IEEE Access* **2024**, *12*, 106672–106684. [\[CrossRef\]](#)
28. Zhang, L. A matrix converter excited doubly-fed induction machine as a wind power generator. In Proceedings of the Seventh International Conference on Power Electronics and Variable Speed Drives, London, UK, 21–23 September 1998; IEEE: New York, NY, USA, 1998; Volume 1998, pp. 532–537. [\[CrossRef\]](#)
29. Cardenas, R.; Pena, R.; Tobar, G.; Clare, J.; Wheeler, P.; Asher, G. Stability analysis of a wind energy conversion system based on a doubly fed induction generator ded by a matrix converter. *IEEE Trans. Ind. Electron.* **2009**, *56*, 4194–4206. [\[CrossRef\]](#)
30. Babaei, E.; Farhadi Kangarlu, M. A new topology for dynamic voltage restorers without dc link. In Proceedings of the 2009 IEEE Symposium on Industrial Electronics & Applications, Kuala Lumpur, Malaysia, 4–6 October 2009; IEEE: New York, NY, USA, 2009; Volume 2, pp. 1016–1021. [\[CrossRef\]](#)
31. Lozano, J.; Ramirez, J.; Correa, R. A novel Dynamic Voltage Restorer based on matrix converters. In Proceedings of the Modern Electric Power Systems (MEPS), 2010 Proceedings of the International Symposium, Wrocław, Poland, 20–22 September 2010; pp. 1–7.
32. Merchan-Villalba, L.; Lozano-Garcia, J.; Avina-Cervantes, J.; Estrada-Garcia, H.; Pizano-Martinez, A.; Carreno-Meneses, C. Linearly decoupled control of a dynamic voltage restorer without energy storage. *Mathematics* **2020**, *8*, 1794. [\[CrossRef\]](#)
33. Amrr, S.M.; Asghar, M.S.J.; Ashraf, I.; Meraj, M. A comprehensive review of power flow controllers in interconnected power system networks. *IEEE Access* **2020**, *8*, 18036–18063. [\[CrossRef\]](#)
34. Shadmand, M.B.; Mosa, M.; Balog, R.S.; Abu-Rub, H. Model predictive control of a capacitorless matrix converter-based STATCOM. *IEEE J. Emerg. Sel. Top. Power Electron.* **2017**, *5*, 796–808. [\[CrossRef\]](#)
35. Rohouma, W.; Metry, M.; Balog, R.S.; Peerzada, A.A.; Begovic, M.M.; Zhou, D. Analysis of the Capacitor-Less D-STATCOM for Voltage Profile Improvement in Distribution Network with High PV Penetration. *IEEE Open J. Power Electron.* **2022**, *3*, 255–270. [\[CrossRef\]](#)
36. Merchan-Villalba, L.R.; Lozano-Garcia, J.M.; Pizano-Martinez, A.; Gonzalez-Ramirez, X.; Estrada-Garcia, H.J.; Zamora-Cardenas, E.A. Static Synchronous Compensator based on the Matrix Converter. In Proceedings of the 2019 IEEE CHILEAN Conference on Electrical, Electronics Engineering, Information and Communication Technologies (CHILECON), Valparaiso, Chile, 13–27 November 2019; IEEE: New York, NY, USA, 2019; pp. 1–6. [\[CrossRef\]](#)
37. Monteiro, J.; Silva, J.F.; Pinto, S.F.; Palma, J. Matrix Converter-Based Unified Power-Flow Controllers: Advanced Direct Power Control Method. *IEEE Trans. Power Deliv.* **2011**, *26*, 420–430. [\[CrossRef\]](#)
38. Szczesniak, P.; Kaniewski, J. Hybrid Transformer with Matrix Converter. *IEEE Trans. Power Deliv.* **2016**, *31*, 1388–1396. [\[CrossRef\]](#)
39. Hojabri, H.; Mokhtari, H.; Chang, L. A generalized technique of modeling, analysis, and control of a matrix converter using SVD. *IEEE Trans. Ind. Electron.* **2011**, *58*, 949–959. [\[CrossRef\]](#)
40. Merchan-Villalba, L.R.; Lozano-Garcia, J.M.; Gutierrez-Torres, D.A.d.J.; Avina-Cervantes, J.G.; Pizano-Martinez, A. Four-Step Current Commutation Strategy for a Matrix Converter Based on Enhanced-PWM MCU Peripherals. *Electronics* **2019**, *8*, 547. [\[CrossRef\]](#)

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