Matrix Converter Based High Power High Frequency Modular Transformers for Traction Conversion Systems

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Abstract—A new system based on a Power Electronic Transformer has been proposed in this thesis. It is installed in the traction substation and regulates the voltage, to the characteristics of the train crossing that section of the rail. It consists of a High Frequency Transformer with a Three Phase Matrix Converter in its input, to guarantee controllable output voltage and frequency, as well as bidirectional power flow. The Matrix Converter uses Space Vector Modulation, which has important advantages such as; a simplified algorithm control, maximum voltage transfer ratio without adding third harmonic components, and an innovative feature developed in this thesis, which also guarantees the non-saturation of the high frequency transformer. Finally, in the output of the transformer are three Single Phase Matrix Converters that restore the original waveform determined by the Space Vector Modulation. By combining the advantages of the Matrix Converters with a high frequency transformer, it is possible to produce controllable voltage, galvanic isolation and power quality improvements without any extra devices. Several features such as instantaneous current regulation and voltage sag compensation are combined with the Power Electronic Transformer. The proposed new Power Electronic Transformer configuration has been modelled using MATLAB/SIMULINK and the main advantages mentioned above have been verified by the simulation results.

Index Terms—Current Regulator, Electric Traction, Matrix Converter, Power Conversion, Power Electronic Transformer, Space Vector Modulation.

I. INTRODUCTION

This research has been motivated by the several electrification railway systems implemented in Europe. Some locomotives use multiple transformers and converters to be able to run in different supply systems, which requires on board extra heavy equipment and in some cases, interruptions in train operation. Therefore, a power transformer converter was designed to be installed in the traction substations in order to provide suitable AC power for the train that crosses the rail sector.

The proposed Power Electronic Transformer for Electric Traction (PETET) [1] is presented in Fig. 1 and it is based on one three phase High Frequency Transformer (HFT) supported by power electronic converters. The input is connected with a three phase Matrix Converter controlled by Space Vector Modulation (SVM) with an innovative feature that guarantees the non-saturation of the HFT. The output consists of three Single Phase Matrix Converters with an output filter. These three MCs are used to restore the original SVM signal.

Figure 1 – Power Electronic Transformer for Electric Traction

The Matrix Converter (MC), which received significant improvements in recent years [3], is a power electronic converter with high frequency switching that generates a variable frequency in the output voltage.

Currently MCs may be used in electrical substations to regulate Distribution Grid voltages [4], in high power applications to regulate the power flow in Transmission Grids [5], in the renewable energy applications where they provide the electrical connection between the power generator and the electric grid [6], in the transportation industry ranging from the aerospace sector to the railway sector. Besides the low distortion of the input/output waveforms, the lower weight and volume of MCs when compared to back-to-back structures, and the bidirectional power flow are a great advantage for the transportation sector [7], allowing regenerative braking.

Also, the utilization of reactive power compensators in substations to compensate losses in the rail network and old locomotives with low power factors [8], is no longer needed since the MC provides controllable input power factor.

II. POWER ELECTRONIC TRANSFORMER FOR ELECTRIC TRACTION

A. Matrix Converter

The three-phase Matrix Converter, which is represented in Fig 2. It consists of nine controlled bidirectional switches making a 3x3 matrix (1) that allows a connection between two three-phase systems; the input with voltage source characteristics and the output system with characteristics of current source. These converters allow direct AC-AC conversion, without an intermediate but with a high efficiency guaranty.
By assuming ideal semiconductors, each switch can be mathematical represented by \( S_{ij} \) \( (k,j \in \{1,2,3\}) \), a binary variable with two possible states: “\( S_{ij}=1 \)” if the switch is ON, and “\( S_{ij}=0 \)” if it OFF. Due to electrical limitations of the MC topology, each line of the matrix can only have one switch “ON”.

\[
S = \begin{bmatrix}
S_{11} & S_{12} & S_{13} \\
S_{21} & S_{22} & S_{23} \\
S_{31} & S_{32} & S_{33}
\end{bmatrix} \quad \sum_{j=1}^{3} S_{kj} = 1 \quad k \in \{1,2,3\}
\]  

(1)

The \( S \) matrix represents the states of the switches and enables a mathematical correlation between the line-to-neutral output voltages \( V_A, V_B, V_C \) and the line-to-neutral input voltages \( V_a, V_b, V_c \). Still, the transpose of matrix \( S \) correlates the input currents \( i_a, i_b, i_c \) with the output currents (2).

\[
[V_a V_b V_c]^T = S^T [i_a i_b i_c]^T
\]

(2)

Finally, the MC has now 27 possible combinations to represent the input currents and output voltages.

B. Space Vector Modulation

SVM approach, including Indirect SVM and Direct SVM proposed in [9] and [10] were often used in MC for it’s appropriate in operation. Conventional SVM approach is used to synchronize the input voltage by zero cross detecting of the input phase voltage, which can be seen in Fig. 3 a), and assuming a balanced input voltage in rated value.

Fig.3 shows MC’s output voltage space vectors and output voltage synthesis, where 0, I, II, III, IV, V and VI stand for six vector sectors, and V1–V6 stand for active voltage vector, \( V_0 \) and \( V_7 \) stand for zero.

The reference vector \( V_{\text{ref}\alpha\beta} \) can be obtained using the adjacent space vector \( V_a, V_b \) and \( V_0 \) represented in Fig 3 c) and the duty-cycles associated to the vectors \( d_a, d_b \) and \( d_0 \). Considering that the switching frequency is much higher than the input frequency \( f_i > f_s \), it is possible for each commutation period to define the reference vector \( V_{\text{ref}\alpha\beta} \) as (3).

\[
V_{\text{ref}\alpha\beta} \approx V_a d_a + V_b d_b + V_0 d_0
\]

(3)

The reference vector \( V_{\text{ref}\alpha\beta} \) of the line-to-line output voltage (4) where the RMS value \( V_{\text{oc}} \) and the line-to-line voltage magnitude \( V_{\text{ocmax}} \), describes a circular path in the plane \( \alpha\beta \) and is synthesized using the space vectors represented in Fig. 3 c).

\[
V_{\text{ref}\alpha\beta}(t) = \sqrt{3} V_{\text{oc}} e^{j\omega t} = \frac{\sqrt{3}}{2} V_{\text{ocmax}} e^{j\omega t}
\]

(4)

Therefore, the maximum of the output voltage reference (4) is the maximum of the available voltage in the intermediate stage \( V_{\text{dc}} \), which concludes that, the limitation of the output voltage of the rectifier-inverter association model is imposed by the rectifier.

Additionally, similar procedure is implemented for the MC rectifier stage, which leads to nine active vectors. It is assumed that the adjacent vectors \( I_1-I_6 \) are \( I_s, I_y \) and zero vectors \( I_8, I_9 \) and \( I_0 \) with the respective duty cycles \( d_s \) (for \( I_s \)), \( d_y \) (for \( I_y \)) and \( d_0 \) (for one of the zero vector).

Considering that the switching frequency is much higher than the input frequency \( f_i > f_s \), it is possible for each commutation period to define the reference vector \( I_{\text{ref}} \) as (5).

\[
I_{\text{ref}} \approx I_y d_y + I_s d_s + I_0 d_0
\]

(5)
From (3) and (5) the duty cycles \(d_\alpha\), \(d_\beta\) and \(d_0\) can be calculated by using a trigonometric analysis. Since the rectifier stage needs two non-zero vectors to realize the input current modulation and the inverter stage needs also two non-zero vectors to realize the output voltage modulation, the modulation function will need five state vectors: four non-zero and one zero vector. The switch time (6) for each selected vector to control the input current and output voltage is obtained by the multiplication of the duty cycles calculated for the rectifier and inverter. [11].

\[
\begin{align*}
\{d_\alpha d_\beta &= m_c m_v \sin\left(\frac{\pi}{3} - \theta_i\right) \sin\left(\frac{\pi}{3} - \theta_v\right) \\
\{d_\alpha d_\beta &= m_c m_v \sin\left(\frac{\pi}{3} - \theta_i\right) \sin(\theta_v) \\
\{d_\delta d_\alpha &= m_c m_v \sin(\theta_i) \sin\left(\frac{\pi}{3} - \theta_v\right) \\
\{d_\delta d_\beta &= m_c m_v \sin(\theta_i) \sin(\theta_v) \\
\{d_0 &= 1 - d_\alpha d_\beta - d_\delta d_\beta - d_\delta d_\alpha - d_\delta d_{\alpha}
\end{align*}
\]

Based on the sector of the line-to-line output reference voltage, and on the sector of the input reference current, it is possible to choose one vector in Table I during a certain time frame in order to control output voltages and input currents.

Table I - Matrix Converter’s vectors used in the modulation of line-to-line output voltages and input currents

<table>
<thead>
<tr>
<th>(V_o)</th>
<th>(I)</th>
<th>(d_\alpha d_\beta)</th>
<th>(d_\alpha d_\beta)</th>
<th>(d_\delta d_\alpha)</th>
<th>(d_\delta d_\beta)</th>
</tr>
</thead>
<tbody>
<tr>
<td>I</td>
<td>1</td>
<td>-4</td>
<td>+1</td>
<td>+6</td>
<td>-3</td>
</tr>
<tr>
<td></td>
<td>2</td>
<td>+6</td>
<td>-3</td>
<td>-5</td>
<td>+2</td>
</tr>
<tr>
<td></td>
<td>3</td>
<td>-5</td>
<td>+2</td>
<td>+4</td>
<td>-1</td>
</tr>
<tr>
<td></td>
<td>4</td>
<td>+4</td>
<td>-1</td>
<td>-6</td>
<td>+3</td>
</tr>
<tr>
<td></td>
<td>5</td>
<td>-6</td>
<td>+3</td>
<td>+5</td>
<td>-2</td>
</tr>
<tr>
<td></td>
<td>6</td>
<td>+5</td>
<td>-2</td>
<td>-4</td>
<td>+1</td>
</tr>
<tr>
<td>IV</td>
<td>1</td>
<td>+4</td>
<td>-1</td>
<td>-6</td>
<td>+3</td>
</tr>
<tr>
<td></td>
<td>2</td>
<td>-6</td>
<td>+3</td>
<td>+5</td>
<td>-2</td>
</tr>
<tr>
<td></td>
<td>3</td>
<td>+5</td>
<td>-2</td>
<td>-4</td>
<td>+1</td>
</tr>
<tr>
<td></td>
<td>4</td>
<td>-4</td>
<td>+1</td>
<td>+6</td>
<td>-3</td>
</tr>
<tr>
<td></td>
<td>5</td>
<td>+6</td>
<td>-3</td>
<td>+5</td>
<td>-2</td>
</tr>
<tr>
<td></td>
<td>6</td>
<td>-5</td>
<td>+2</td>
<td>+4</td>
<td>-1</td>
</tr>
<tr>
<td>V</td>
<td>1</td>
<td>+1</td>
<td>-1</td>
<td>-7</td>
<td>+3</td>
</tr>
<tr>
<td></td>
<td>2</td>
<td>-3</td>
<td>+9</td>
<td>+2</td>
<td>-8</td>
</tr>
<tr>
<td></td>
<td>3</td>
<td>+2</td>
<td>-8</td>
<td>-1</td>
<td>+7</td>
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<tr>
<td></td>
<td>4</td>
<td>-1</td>
<td>+7</td>
<td>+3</td>
<td>-9</td>
</tr>
<tr>
<td></td>
<td>5</td>
<td>+3</td>
<td>-9</td>
<td>-2</td>
<td>+8</td>
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<tr>
<td></td>
<td>6</td>
<td>-2</td>
<td>+8</td>
<td>+1</td>
<td>-7</td>
</tr>
<tr>
<td>VI</td>
<td>1</td>
<td>+1</td>
<td>-7</td>
<td>+3</td>
<td>-9</td>
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<tr>
<td></td>
<td>2</td>
<td>-3</td>
<td>+9</td>
<td>+2</td>
<td>-8</td>
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<tr>
<td></td>
<td>3</td>
<td>+2</td>
<td>-8</td>
<td>-1</td>
<td>+7</td>
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<td>4</td>
<td>-1</td>
<td>+7</td>
<td>+3</td>
<td>-9</td>
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<td>5</td>
<td>+3</td>
<td>-9</td>
<td>-2</td>
<td>+8</td>
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<tr>
<td></td>
<td>6</td>
<td>-2</td>
<td>+8</td>
<td>+1</td>
<td>-7</td>
</tr>
</tbody>
</table>

The selection of the vector that specifies the commutations of the three-phase matrix converter switches is not only based on the regions of Fig. 4, but also in Table I, as the vectors are different for every location of the input currents and output voltages. For each current and voltage sector, Table I presents the four vectors that will be used in the modulation process, thus producing the \(\alpha\) or \(\beta\), and \(\gamma\) or \(\delta\) components necessary to follow the reference voltage and current vectors. The selected vector is finally sent to the three-phase matrix converter to command the switches state.

A summary of this process can be seen using one example shown in Fig. 2. The components \(d_\delta d_\beta\) are selected during \(T_{d_\delta d_\beta}\), and with Table I, which receives information about the location of the input current and output voltage, it is possible to select the vector “-2”. The sequence continues in order to represent also the reference vectors (3), (5) by the components \(d_\alpha\) and \(d_\beta\).

![Figure 4](image-url)  
**Figure 4** – Modulation process used to select the space vectors and the time interval when they are applied

![Figure 5](image-url)  
**Figure 5** – Selection scheme for the SVM vectors

C. Modified Space Vector Modulation

The conventional SVM presented in the previous chapter is improved in order to produce an output voltage with an average value equal to zero in each switching period. Therefore, the duty-cycles of Fig. 4 should be divided in two time frames; the first one corresponding to the vector selected
by SVM, and the second one to the vector that produces a symmetric voltage (7). Note that $T_{d\delta\beta}$ is variable since the duty cycles are not constant.

$$H_{3\delta}(t) = \begin{cases} 
\text{SVM Vector} & 0 \leq T < T_{d\delta\beta}/2 \\
-1 \times \text{SVM Vector} & T_{d\delta\beta}/2 \leq T < T_{d\delta\beta} 
\end{cases}$$

(7)

As a result, in the modified modulation process, four more duty cycle signals are created (8), when compared to the original SVM (7).

$$\begin{align*}
&d_\delta d_\alpha = \frac{m}{2} \sin(\frac{\pi}{3} - \theta_i) \sin(\frac{\pi}{3} - \theta_v) \\
&d_\delta d_{\alpha -} = \frac{m}{2} \sin(\frac{\pi}{3} - \theta_i) \sin(\frac{\pi}{3} - \theta_v) \\
&d_\delta d_\beta = \frac{m}{2} \sin(\theta_i) \sin(\frac{\pi}{3} - \theta_v) \\
&d_\delta d_{\beta -} = \frac{m}{2} \sin(\theta_i) \sin(\frac{\pi}{3} - \theta_v) \\
&d_0 = 1 - d_\delta d_\alpha - d_\delta d_\beta - d_\delta d_{\alpha -} - d_\delta d_{\beta -}
\end{align*}$$

(8)

Due to the nomenclature of the vectors established, the vector that produces a symmetric voltage corresponds to the vector named with a symmetric number. Therefore, in the example of Fig 5 the three phase MC is controlled by $H_{3\delta}(t)$ (7), whereby the vector selected for $0 \leq T < T_{d\delta\beta}/2$ is the vector “-2” and for $T_{d\delta\beta}/2 \leq T < T_{d\delta\beta}$ the selected vector is “+2”.

Fig. 6 presents the modified modulation process with the extra four duty cycle signals (8). It can be seen that the intervals are now divided in two (except for $d_0$); the first is defined by the original SVM as in Fig 4, and the second interval is created by the extra duty cycle signals from the modified SVM.

This process demands a higher switching frequency but provides a zero mean value for the MC output voltage. The modulator will produce an output voltage waveform similar to the one represented in Figure 7 (considering only one of the input voltages).

Fig. 7 shows that the mean value of the voltage during each switching period is not zero, which means that for transformers that operate at higher frequencies (1 kHz) the voltage is continuously increasing during one half of the grid period or decreasing during the other half of the grid period. Consequently, this is a problem for HFTs as they can easily saturate. To avoid this problem, a new strategy is developed in this thesis in order to ensure the non-saturation of the HFT.

D. Single-Phase Matrix Converter

As mentioned above, the system also contains Single Phase Matrix Converters (SPMC) that consists of four controlled bidirectional switches. As in the Three-Phase Matrix Converter the switching conditions in the SPMC, have to avoid the problems of short circuits and open circuits. Therefore, there are four possible states, which are described in Table II with the respective correlations between the electrical variable combinations.
Since the non-saturation of the transformer is assured by the modified SVM strategy used in the Three-Phase Matrix Converter, the purpose of the SPMC is to restore the output voltage applied to the load into the voltage synthesized by the original SVM.

Considering Table II, the SPMC is able to produce at its output the same voltage applied at the input or its symmetric voltage. In the current example, $H_{10}(t)$ determines the state “1” for the SPMCs, during $0 \leq T < T_{\text{dab}}/2$ and the state “-1” for $T_{\text{dab}}/2 \leq T < T_{\text{dab}}$ as indicated in (9).

$$H_{10}(t) = \begin{cases} \text{State } "1" & 0 \leq T < T_{\text{dab}}/2 \\ \text{State } "-1" & T_{\text{dab}}/2 \leq T < T_{\text{dab}} \end{cases}$$

As an example of the modulation process, a sequence of vectors is presented in Fig. 8. According to the conventional SVM approach, vector -2 should be applied during time interval $T_{\text{dab}}$. However, to avoid the HFT saturation, it is necessary to use the modified SVM. Then:

- In the first half of the considered interval, where the vector is “-2”, the SPMCs reproduces the input voltage at the output.
- In the second half of the considered interval, the three phase MC modified SVM changes the vector to “+2”. Consequently, there is the need to restore the vector selected by the original SVM. This can be done guaranteeing that the SPMCs are in the state “-1” (9), thus assuring that the voltage applied to their output, is symmetric to their input voltage.

![Table II - Possible switch combinations for a Single-Phase Matrix Converter](image)

<table>
<thead>
<tr>
<th>State</th>
<th>$S_{11}$</th>
<th>$S_{12}$</th>
<th>$S_{13}$</th>
<th>$S_{14}$</th>
<th>$V_A$</th>
<th>$V_B$</th>
<th>$i_A$</th>
<th>$i_B$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>$V_a$</td>
<td>$V_b$</td>
<td>$i_a$</td>
<td>$i_b$</td>
</tr>
<tr>
<td>2</td>
<td>0</td>
<td>1</td>
<td>1</td>
<td>0</td>
<td>$V_b$</td>
<td>$V_a$</td>
<td>$i_b$</td>
<td>$i_a$</td>
</tr>
<tr>
<td>3</td>
<td>1</td>
<td>0</td>
<td>1</td>
<td>0</td>
<td>$V_b$</td>
<td>$V_a$</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>4</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>1</td>
<td>$V_a$</td>
<td>$V_b$</td>
<td>0</td>
<td>0</td>
</tr>
</tbody>
</table>

The developed SVM approach ensures that the final output is the one that would be obtained by the conventional SVM method, while avoiding the saturation of the HFT.

### III. CONTROL OF THE OUTPUT CURRENT

The current regulator block diagram is represented in Fig. 9, where $I_{\text{load}}$ is the reference current and $I_{\text{load}}$ the load current. Both are multiplied by $a_i$, the current sensor gain, and the difference between the two currents, i.e., the current error is applied to the controller $C(s)$. This controller generates the modulating voltage used by the SVM.

![Figure 9 – Output current regulator block diagram](image)

C(s) is a Proportional-Integral (PI) Controller, which ensures a dynamic second order closed chain. This compensator ensures a null static error and an acceptable rise time.

To design the current regulator, it is possible to represent the three phase matrix converter as a function of the first order with a delay $T_d$ and a unitary gain. This transfer function is designated by $G(s)$, and given by (10).

$$G(s) = \frac{1}{1 + sT_d} \tag{10}$$

To calculate the parameters $T_r$ and $T_p$, it is considered that the zero of $C(s)$ cancels the lowest frequency pole, introduced by the output LR filter. $T_r$ is given by (11) where $R_{\text{load}}$ is the sum of the coil intern resistance with the load resistance.

![Figure 8 – Overall process of vectors selection](image)
\[ T_p = \frac{L_{out}}{R_{out}} \]  

(11)

The value of \( T_p \) is calculated by (12), where \( \alpha_i \) is the current gain and \( T_d \) is the average delay of the system.

\[ T_p = \frac{2 \alpha_i T_d}{R_{out}} \]  

(12)

IV. CONTROL OF THE OUTPUT VOLTAGE

The voltage regulator is shown in Fig. 10, where the capacitor (\( C_{fout} \)) was changed to the SPMCs output, next to the load.

![Figure 10 - Load voltage regulator](image)

The voltage regulator has to ensure that the load voltage, which is the same as the capacitor voltage, (13).

\[ V_{load} = \frac{i_c}{sC_{fout}} \]  

(13)

In the design of the controller, it is considered that the load current (\( I_{load} \)) is a disturbance of the system [12], [13]. As the current output of the matrix converters is controlled, it is also possible to consider that the matrix converters, filters and transformer leakage inductances can be represented by the current source \( I_{system} \).

In Fig. 11 is presented the voltage regulator block diagram, wherein the block \( \frac{1/\alpha_i}{sT_d + 1} \) represents the matrix converter controlled by current, [14].

![Figure 11 - Block diagram of the voltage regulator](image)

Finally, the proportional gain \( K_p \) and the integral gain \( K_i \) are obtained by:

\[
\begin{align*}
K_p &= \frac{2.15C_{fout}\alpha_i}{\alpha_i T_d(1.75)^2} \\
K_i &= \frac{C_{fout}\alpha_i}{\alpha_i T_d^2(1.75)^3}
\end{align*}
\]  

(14)

V. RESULTS

The developed system was implemented in MATLAB/SIMULINK software in order to evaluate its performance.

The simulations presented aim to confirm the adequate operation of this system in several operation scenarios. These scenarios are intended to simulate different electric traction systems with different voltage levels and frequencies, as described in Chapter 2.

A. Scenario 1 - 50Hz operation

The first scenario aims to determine the performance of the PETET in a 1x25 System (25 kV at 50 Hz). Due to the semiconductor’s limitations described above, the simulation was designed for a single PETET module, to provide 500kV in the output, a small fraction of 25 kV.

Fig. 12 presents the line-to-neutral load voltages, which are close to their sinusoidal waveform references, with a fundamental frequency of 50 Hz. Despite the small disturbances created by the MCs switching and the HFT, the waveforms are nearly sinusoidal. The same can be seen in Fig. 13, where the load currents are shown.

![Figure 12 - Line-to-neutral load voltages](image)

![Figure 13 - Load currents](image)

Fig. 14 presents the input current in one phase of the HFT. Since the transformer operates at 1 kHz, the 50 Hz component, which is easily observed, is not considered for the HFT saturation issue, as the average value of the current is nearly equal to zero in each switching period.
With a closer zoom of Fig. 14, it can be seen in Fig. 15 that the average value of the current is zero in each switching period, since the applied voltage changes from maximum to minimum and vice versa for identical time intervals. Consequently, the non-saturation of the HFT is assured.

Fig 16 presents one phase of the input current and line-to-neutral input voltage of the Three-Phase MC. The slightly leading current indicates that the system injects reactive power. With higher load system currents, the input power factor will be closer to 1.

Fig 17 presents one phase of the line-to-neutral load voltage and the voltage regulator performance. At t=2s, the magnitude of the output voltage reference is decreased with a good response of the voltage regulator as shown in Fig. 17 and Fig. 18.

The same situation is possible to observe with the errors of the components $v_d$ and $v_q$ in Fig 18. At t=2s, when the voltage reference is decreased, it is easily noted that there is a high peak for a small time instant, but the controller quickly guarantees that the steady-state is reached, and the output voltages are equal to their reference values.

B. Scenario 2 – 16.7Hz operation
Fig. 19 presents the line-to-neutral load voltages, which are close to their sinusoidal waveform references, with fundamental frequency of 16.7 Hz. Despite the small disturbances created by the MCs switching and the HFT, the waveforms are nearly sinusoidal.

Fig. 20 shows the load currents. As expected, they are sinusoidal.

Fig. 21 presents one phase of the input HFT current. As seen in the previous scenario test, for the 50Hz grid, the average value of the current in each switching period is equal to zero as well.
Fig. 22 presents one phase of the input current and line-to-neutral input voltage of the Three-Phase MC. The slightly leading current indicates that the system injects reactive power.

![Figure 22 - One phase of the input current (red), one phase of line-to-neutral input voltage (blue)](image)

VI. CONCLUSION

The PETET was tested in MATLAB/SIMULINK environment, with good results. The new SVM control of the Three-Phase Matrix Converter was successful proved, which guarantees the non-saturation of the HFT. This technique was described and later demonstrated by the input current waveform of the HFT. At the same time, the reputation of this modulation technic remains unbroken since both input currents and output voltages are sinusoidal. However, it was understood that for the PETET implementation in each railway electrification system, different input and output filters have to be designed as well as the HFT.

The simulations were realized using lower voltages and currents. As explained above, this decision was made to provide more truthful simulations since semiconductors have voltage limitations. Therefore, it is possible to conclude that the system developed here has other uses, e.g., a voltage regulator for the distribution grid. The advantage of an electric grounded system is also an important advantage in power distribution systems as well due to safety human reasons.

Also, this system guarantees the redundancy due to voltage margin required for any defect in the power electronic devices.

The current controller, based in PI controller, was tested with a good performance, with a static error close to zero, and allowing a quick response while ensuring system stability for various load scenarios. The power factor might not be unitary since it depends on the input filter and the load conditions.

It is possible to conclude that the work achieved the proposed objectives. As all the main features of the project were confirmed by the simulations, it can be conclude that the system developed in this thesis has conditions for being implemented in the electric traction, but also in other areas.

REFERENCES


