SHUNT AND SERIES CONDITIONING OF HYBRID MATRIX CONVERTER

By

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ABSTRACT

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Hybrid matrix converters can potentially enable matrix converters in high power applications that conventional matrix converters will not be able to attain. It uses a conventional nine-switch matrix converter in conjunction with an auxiliary back-to-back ac-dc-ac converter that conditions the current and voltage waveforms on the input and output side of the matrix converter. The matrix converter processes the main power at low switching frequency to enable significant reduction of switching losses and to allow for adoption of high-power semiconductor devices such as integrated gate commutated thyristors (IGCTs). The auxiliary ac-dc-ac converter is dedicated to improving the power quality at the input and output terminals of the matrix converter by minimizing harmonic currents drawn from the source and harmonic voltages applied to the load. Essentially, the auxiliary back-to-back converter functions as a shunt-and-series active filter (AF).

Several AF control techniques have been presented in the literature. Based on the operating principle, these techniques can be categorized into two groups. The first group of methods are based on instantaneous reactive power theory (IRPT) and extract the reactive component of the power and the oscillatory component of the real power. The other methods are based on filtering techniques and extract the fundamental component of the current or voltage such as notch filter and fast Fourier transform (FFT) methods.

The main limitation for IRPT based method lies in its ineffectiveness when the harmonics are concurrently present in voltage and current while the limitation for FFT based method is its inability to compensate the fundamental component. To address these limitations of

the aforementioned methods, a new control strategy based on power averaging has been proposed. This proposed control method is able to effectively obtain the correct active component of current or voltage in cases where both the current and the voltage are non-sinusoidal and provide full control over the power factor.

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Chapter 1

Introduction

Matrix converters have several advantages over traditional frequency converters. They are able to provide sinusoidal input currents and output voltages with smaller size and no large energy storage components required, and to achieve full control over the input power factor for any load [1]. Due to the qualities that the matrix converter provides, it has been a very attractive area of research over the recent years. The real development was started by Venturini and Alesina published in 1980 [2][3], they first introduced the name "Matrix Converter" and provided a detailed mathematical model describing the behavior of the converter. Another modulation technique is based on a fictitious DC link connecting a current source bridge and voltage source bridge presented by Rodriguez [4]. This approach is known as the indirect transfer function approach. In 1989 the method of space vector modulation technique for matrix converters introduced by Huber [5].

In 1992 it was practically confirmed that the nine switches matrix converter shown in Fig.1.1, could be used effectively in vector control of induction motor [1]. However, the usage of the matrix converter was still limited due to the difficult current communication of bidirectional switches. Recently, many solutions have been presented to solve the communication problem such as the four step method [6].

Matrix converters generate current harmonics that are injected back into the AC system.

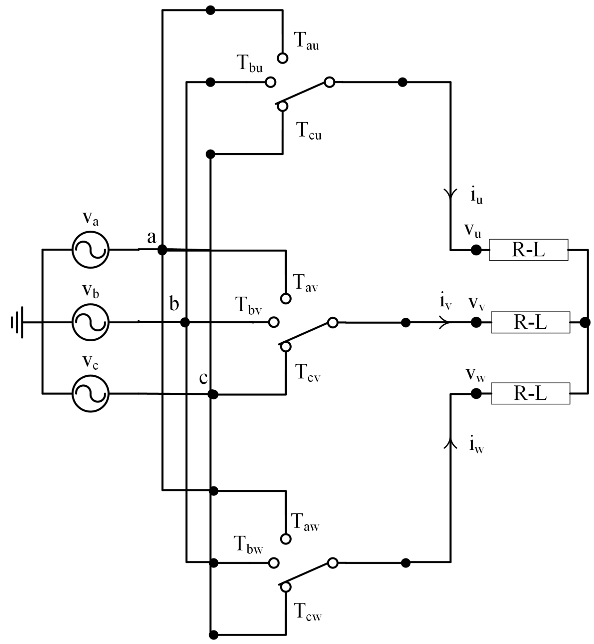

Figure 1.1: Conventional nine bidirectional switch matrix converter.

These current harmonics can result in voltage distortions that further affect the operation of the entire system. On the other hand the voltage harmonics on the output side will cause a disturbance to the load which in most cases is inductive. In order to reduce these harmonics in the source current and the output voltage, passive filters are typically used. Different configurations of low pass passive filters have been proposed [7], The size and design of these filters depend on many factors such as:

1- power quality requirements;

2- power system harmonic content;

3- converter switching frequency;

4- converter modulation technique [8].

Passive filters seem to be an preferred solution in low power applications of matrix converters. However, this is not the case when a matrix converter is introduced for high power interfaces, the optimal design of the passive filter would be a major challenge.

The first high power matrix converter is introduced by Yaskawa for wind power applications [9], in which the modular concept is used to cope with the various demands in the power grid. However, the modular concept is presented to be used in a very high voltages applications such as wind mill and large problems. In this paper a hybrid matrix converter will be introduced to operate at high power medium voltage demands. The proposed topology employ a conventional nine-switch matrix converter coupled with shunt active filter connected to the input side to reduce the source current harmonics, and series active filter connected to the output side to reduce the output voltage harmonics. The two active filters are connected through a small DC link capacitor.

The main matrix converter is modulated using the low frequency modulation algorithm [10] to allow maximum voltage transfer ratio, and the active filters are controlled based on the instantaneous reactive power theory [11] [12].

The thesis is organized as follows, The next chapter give a brief discussion about the high frequency modulation techniques and explains the low frequency modulation technique. The third chapter introduce the hybrid matrix converter and its operation. The fourth chapter discuss the operation of hybrid matrix converter in abnormal conditions. The sixth chapter discuss other conditioning method for hybrid matrix converter and introduce a new conditioning technique to assess the evaluation of the previous methods.

Chapter 2

Background

This chapter aim to give a general description of the main features of matrix converter. Matrix converter is a one stage AC-AC converter. It has several advantages over the conventional AC-DC-AC converters such as sinusoidal input and output waveform, it has inhered the bidirectional flow of power, and minimal energy storage requirements. Matrix converter has also disadvantages. The maximum output voltage is 0.866% for sinusoidal input and output waveform. It requires more semiconductor devices than the conventional AC-DC-AC converters.

2.1 The Conventioanl Topology of Matrix Converter

The nine bidirectional switch three phase matrix converter is shown in Fig. 1.1. The input terminal of the matrix converter is connected to a three phase voltage source. while the output side is connected to a three phase load. The configuration of matrix converter theoretically assumes (512) switching configuration. However, if we take in the consideration the constrains of the input being a voltage source and the output being a current source, such that the input side of the matrix converter can not be short circuited and the output side of the matrix converter can not be open circuited. This yield of only 27 feasible switching combinations.

2.2 High Frequency Modulation Techniques of Matrix Converter

In order to analysis the modulation techniques of matrix converter, its valid to consider ideal switching and that the switching frequency is much higher that the input and the output frequencies[15].

The input output relationships of voltages and currents are as follows.

$$v_{out} = H \cdot v_{in}, \tag{2.1}$$

$$i_{in} = H^T \cdot i_{out}, \tag{2.2}$$

where v_{out} is the output voltage vector $[v_u \;\; v_v \;\; v_w]$, v_{in} is the input voltage vector $[v_a \;\; v_b \;\; v_c]$, i_{in} is the input current vector $[i_a \;\; i_b \;\; i_c]$, and i_{out} is the output current vector $[i_u \;\; i_v \;\; i_w]$. H is the switching matrix of the matrix converter expressed as

$$H = \begin{bmatrix} h_{au} & h_{bu} & h_{cu} \\ h_{av} & h_{bv} & h_{cv} \\ h_{aw} & h_{bw} & h_{cw} \end{bmatrix}. \tag{2.3}$$

Each element in the transformation matrix (2.3) corresponds to one switching function for the direction matrix converter. Considering that assumptions we made earlier and furthermore neglecting all of the high frequency components we can replace the switching function in the switching matrix in to the modulation function represented by the duty cycle.

The constrains of the input and the output sides of the matrix converter can be maintained via the following equations

$$h_{au} + h_{bu} + h_{cu} = 1 \tag{2.4}$$

$$h_{av} + h_{bv} + h_{cv} = 1 \tag{2.5}$$

$$h_{aw} + h_{bw} + h_{cw} = 1 \tag{2.6}$$

The determination of any modulation strategy for the matrix converter is basically determining the appropriate duty cycle that the input-output relationships equ.(2.1) and (2.2)

2.2.1 Alesina-Venturini 1981

The modulation of matrix converter can be stated as follows. Given a set of input voltages and a set of output currents

$$v_{in} = V_{in} \begin{bmatrix} \cos(\omega_i t) \\ \cos(\omega_i t - \frac{2\pi}{3}) \\ \cos(\omega_i t + \frac{2\pi}{3}) \end{bmatrix}. \tag{2.7}$$

$$i_{out} = I_{out} \begin{bmatrix} \cos(\omega_o t + \phi_o) \\ \cos(\omega_o t - \frac{2\pi}{3} + \phi_o) \\ \cos(\omega_o t + \frac{2\pi}{3} + \phi_o) \end{bmatrix}. \tag{2.8}$$

find the modulation matrix $H(t)$ such that

$$v_{out} = qV_{in} \begin{bmatrix} \cos(\omega_o t) \\ \cos(\omega_o t - \frac{2\pi}{3}) \\ \cos(\omega_o t + \frac{2\pi}{3}) \end{bmatrix}. \tag{2.9}$$

and

$$i_{in} = q\cos(\phi_o) I_{out} \begin{bmatrix} \cos(\omega_i t + \phi_o) \\ \cos(\omega_i t - \frac{2\pi}{3} + \phi_o) \\ \cos(\omega_i t + \frac{2\pi}{3} + \phi_o) \end{bmatrix}. \tag{2.10}$$

where q is the voltage transfer ratio of the output and the input voltages. There are two basic solution for this as shown below

The first method obtained by using the duty cycle of the matrix converter [16] represented in the following equation

$$H1 = \begin{bmatrix} 1 + 2qcos(\omega_m t) & 1 + 2qcos(\omega_m t - \frac{2\pi}{3}) & 1 + 2qcos(\omega_m t + \frac{2\pi}{3}) \\ 1 + 2qcos(\omega_m t + \frac{2\pi}{3})) & 1 + 2qcos(\omega_m t) & 1 + 2qcos(\omega_m t - \frac{2\pi}{3}) \\ 1 + 2qcos(\omega_m t - \frac{2\pi}{3} & 1 + 2qcos(\omega_m t + \frac{2\pi}{3})) & 1 + 2qcos(\omega_m t)) \end{bmatrix}. \tag{2.11}$$

and

$$H2 = \begin{bmatrix} 1 + 2qcos(\omega_m t) & 1 + 2qcos(\omega_m t - \frac{2\pi}{3}) & 1 + 2qcos(\omega_m t + \frac{2\pi}{3}) \\ 1 + 2qcos(\omega_m t - \frac{2\pi}{3} & 1 + 2qcos(\omega_m t + \frac{2\pi}{3})) & 1 + 2qcos(\omega_m t)) \\ 1 + 2qcos(\omega_m t + \frac{2\pi}{3})) & 1 + 2qcos(\omega_m t) & 1 + 2qcos(\omega_m t - \frac{2\pi}{3}) \end{bmatrix}. \tag{2.12}$$

For $H1$, $\omega_m = \omega_o - \omega_i$ and for $H2$, $\omega_m = -(\omega_o + \omega_i)$

So for giving the same phase displacement at the input and the output ports $\phi_i = \phi_o$

gives (2.11), where $\phi_i = -\phi_o$ gives (2.12). Combining the two solutions provides full control over the input power factor.

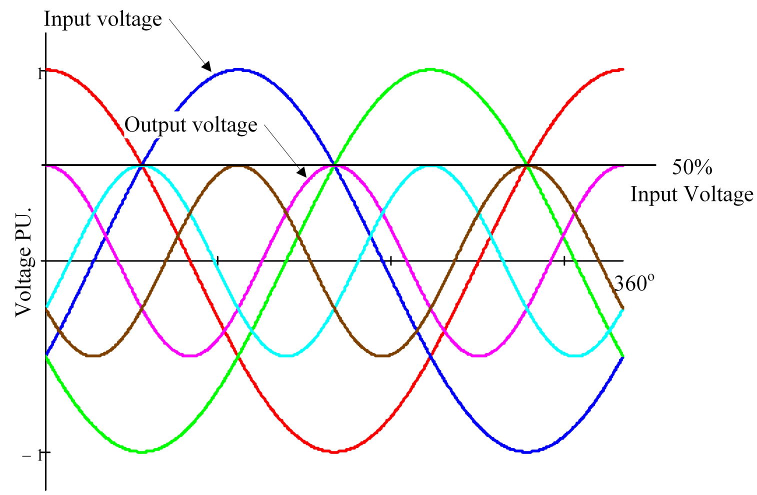

Figure 2.1: Illustration of maximum voltage transfer ratio.

The basic solution represents a direct transfer function approach. during each switch sequence time the avarge output voltage equal to the target voltage. For this to be achived, the output voltage must fit withen the invelope of the input voltage as shown in Fig. 2.1. This modulation solution limit the output voltage to %50 of the input voltage.

It is possible to add the common mode voltage to increase the voltage transfer ration to 0.866 as shown in Fig. 2.2.

It is important to mention that the common mode voltage does not effect the output line-to-line voltage. It only allow that output voltage to be within the input voltage envelope with 0.87 voltage transfer ratio.

2.2.2 Alesina-Venturini 1989

Calculating the switching timing directly from the first method is quite difficult. Its possible to express the the modulation function as

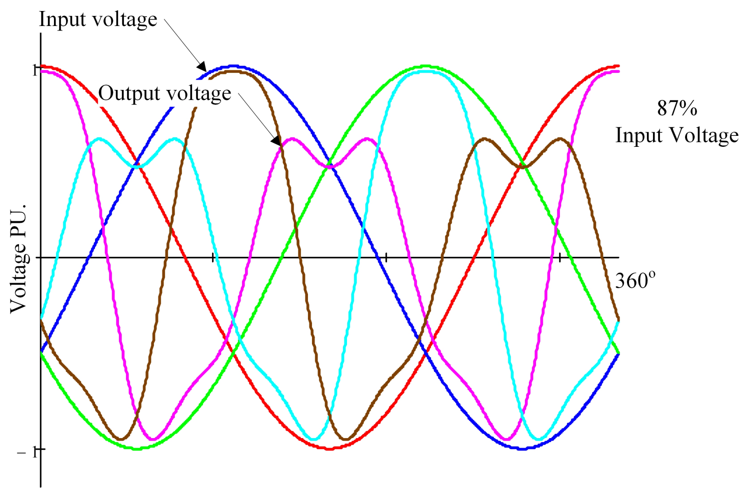

Figure 2.2: Illustration of maximum voltage transfer ratio improved to 87%.

$$h_{Kj} = \frac{t_{Kj}}{T_{seq}} = \frac{1}{3}\Big(1 + \frac{2v_K v_j}{V_{in}^2}\Big) \tag{2.13}$$

where $K = a, b, c$ and $j = u, v, w$. Equation (2.13) correspond to 50% voltage transfer ratio. to achieve the maximum voltage transfer ration we may consider the common mode voltage as follows

$$h_{Kj} = \frac{t_{Kj}}{T_{seq}} = \frac{1}{3}\Big(1 + \frac{2v_K v_j}{V_{in}^2} + \frac{4q}{3\sqrt{3}}\sin(\omega_i t + \beta_K)\Big) \tag{2.14}$$

where β_K is 0, $\frac{-2\pi}{3}$, $\frac{2\pi}{3}$ for $K = a, b, c$

The input displacement factor can be controlled by inserting a phase shift between the measured input voltage and the voltage v_K in (2.14). However, controlling the input displacement angle is on the expense of voltage transfer ratio.

2.2.3 Space-Vector Approach

An other approach is by defining the available vectors and for a space vector representation for the input and output voltages and current. Among the 27 available vectors there are only 21 voltage vector can be usefully employed to the space vector algorithm. The first 18 vector determine the output voltage and the input current and the other 3 are the zero vectors. The 6 vector that completes the 21 vector are basically connecting each output phase to a different input phase [15]. SVPWM method can also achieve 0.87 voltage transfer ratio. Since we are not interested in high frequency modulation techniques, we omit the details of SVPWM.

2.3 Low Frequency Modulation Technique of Matrix Converter

All of the proposed switching techniques except the programmable one introduced in [17] and the Z-source matrix converter family presented in [18] were not able to break $\frac{\sqrt{3}}{2}$ voltage transfer ratio. This internsic limitation of the matrix converter reduces the output power of the induction motor to $\frac{2}{3}$ when it is fed from the matrix converter using a conventional power source. Low switching technique, six step method was presented by Bingsen Wang [10] that is able to break the limit of voltage transfer ratio. The six-step method will be used as the modulation technique for the nine-switch matrix converter for the flowing reasons:

1- Greater than unity voltage transfer ratio, namely 1.05%

2- Low switching frequency with minimum switching losses compared to high-frequency synthesis modulation techniques

3- Easy to control and implement.

The conventional matrix converter in Fig. 1.1 could be realized to an indirect matrix converter as shown in Fig. 2.3. Where the input-output relationship can be expressed as

$$v_{out} = H_{VSB} H_{CSB} \cdot v_{in}, \tag{2.15}$$

$$i_{in} = H_{CSB}^T H_{VSB} \cdot i_{out}, \tag{2.16}$$

where H_{CSB} is the switching matrix for the current source bridge connected to the stiff

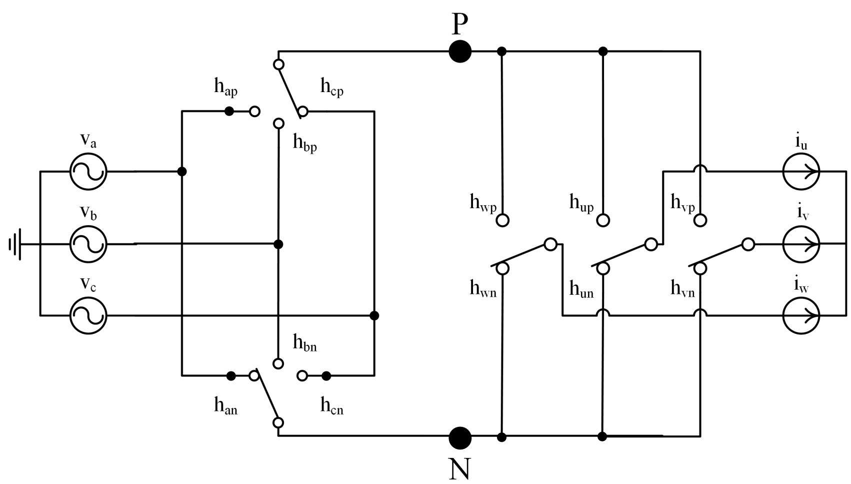

Figure 2.3: Conventional nine bidirectional switch matrix converter.

voltage and H_{VDB} is the switching matrix for the voltage source bridge connected to the current source and they are given by

$$H_{CSB} = \begin{bmatrix} h_{ap} & h_{bp} & h_{cp} \\ h_{an} & h_{bn} & h_{cn} \end{bmatrix}. \tag{2.17}$$

$$H_{CSB} = \begin{bmatrix} h_{up} & h_{un} \\ h_{vp} & h_{vn} \\ h_{wp} & h_{wn} \end{bmatrix}. \tag{2.18}$$

where

$$H = H_{CSB} \cdot H_{CSB} \tag{2.19}$$

Any modulation technique that provides solution to the indirect realization of the matrix converter can be applied to the direct matrix converter via (2.19). In our case the CSB will be treated as a rectifier and the VSB will be treated a voltage source inverter, and both converter are operating in full six-steps mode.

The switching signlas and the input-output voltage ans current are shown in 2.4.

Each element in the transformation matrix corresponds to one switching function for the direction matrix converter. By an appropriate selection of the switching functions, output voltages and the input currents similar to those of the voltage source inverter and current source inverter, respectively, can be achieved. The resulting output voltages and input currents for unity displacement factor operation at the input and output, and the corresponding Fourier spectra are shown in Fig. 2.4 and 2.5, respectively.

The spctrum of the input current includes low frequency harmonic 5^{th} and 7^{th}. The load voltage include the 5^{th} and 7^{th} as shown in Fig. 2.5.

The total harmonic distortion in both the input current and the output voltage is around 31%. The input current and output voltages waveform contain a low frequency harmonics, and for high power application its difficult to design low pass filter that is able mitigate these

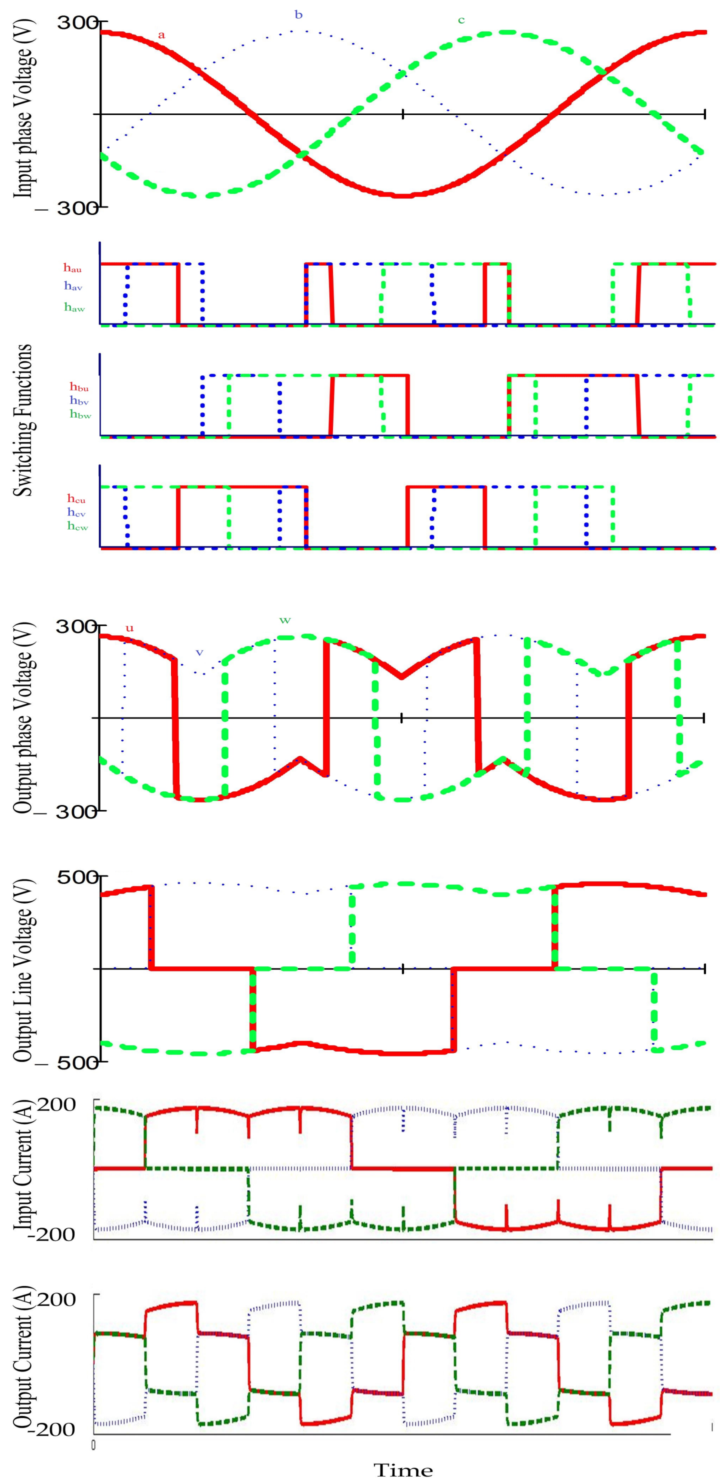

Figure 2.4: Typical input/output voltages/currents of a matrix converter with low-frequency modulation scheme.

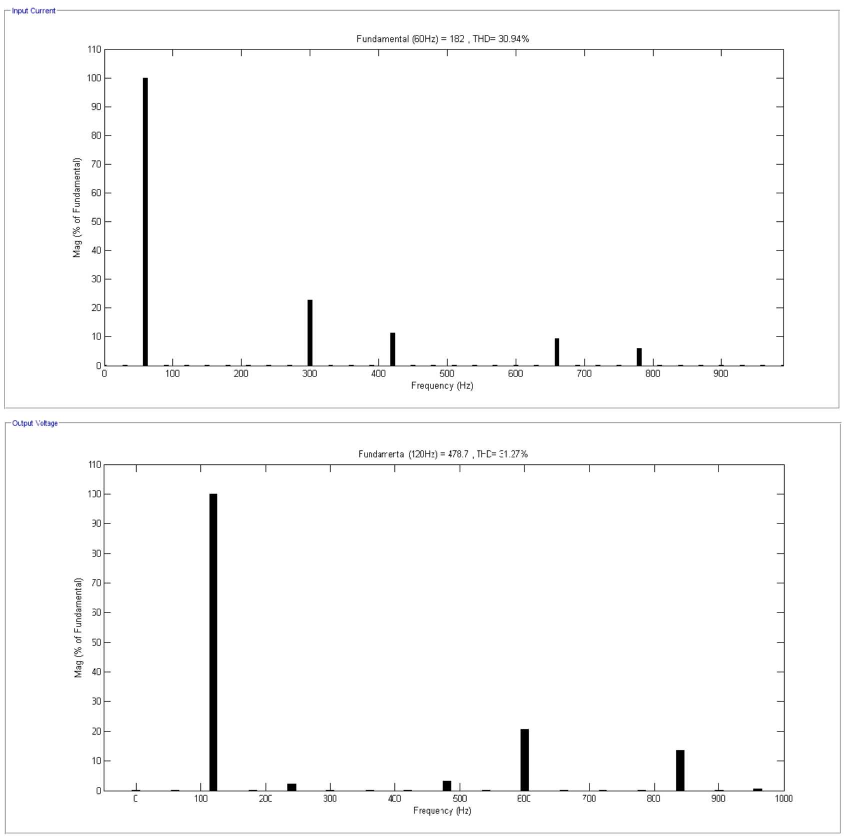

Figure 2.5: The spectra of input currents and output voltages.

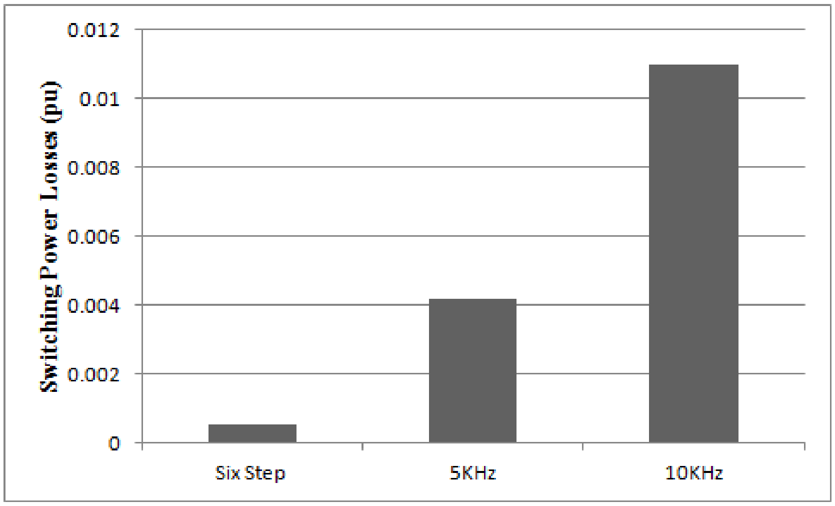

Figure 2.6: Switching losses of IGBTs and diodes at various switching frequencies.

low frequency harmonics.

For illustration, the switching losses have been calculated analytically for matrix converter connected to 100 kW RL load in cases of low frequency modulation and high frequency modulation techniques. Fig. 2.6 shows that the switching power losses of the low frequency modulation matrix converter are very small compared to high frequency synthesis matrix converter, The switch on power losses for the diode is neglected because they are very small.

Chapter 3

The Proposed Hybrid Matrix Converter

Matrix converters would inject a significant amount of current harmonics and voltage harmonics back in to the power source and the load, respectively, if filtering measures are not properly implemented. These harmonics cause distortion and adversely impact the operation of the whole system. The increasing demand for high power makes the conventional solution of passive filters no longer sufficient to solve the power quality problem in matrix converters, beside the other drawbacks of using bulky passive components such as, short life time, big size, and high cost. This chapter presents a hybrid matrix converter for medium voltage high power applications. By utilizing a low-frequency modulation techniques, combined with a shunt active filter implemented to the input side of the matrix converter to eliminate current harmonics and a series active filter is applied to the output side of the matrix converter to eliminate the voltage harmonics. This approach is more efficient in terms of reducing the total number of passive components used in the system, reducing the switching power losses, and improving voltage transfer ratio. The explanation of predicting the compensation current and the compensation voltage waveforms is based on the instantaneous reactive power theory. The feasibility and effectiveness proposed topology have been verified by the simulation results.

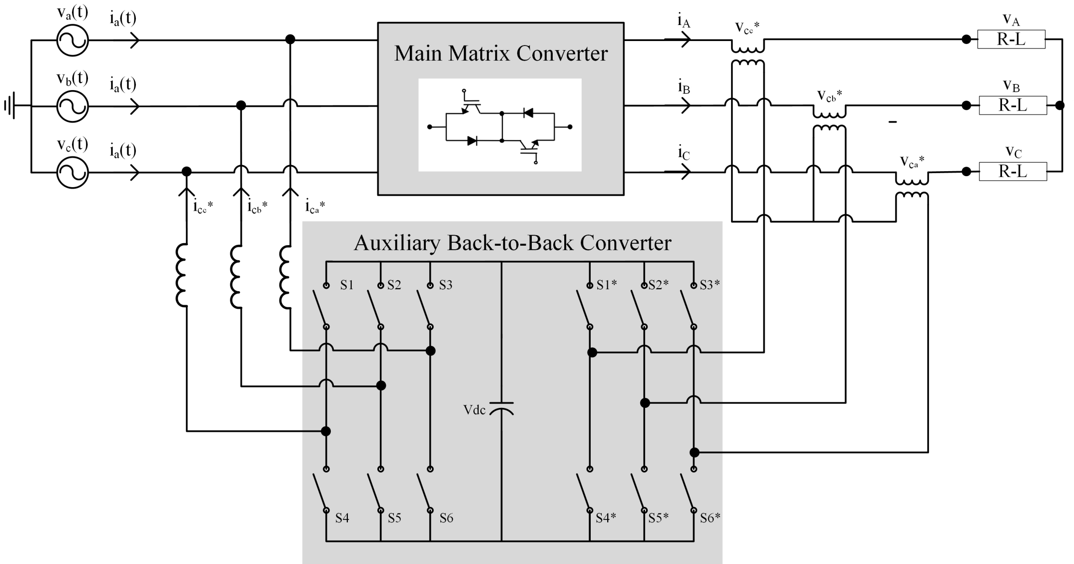

Figure 3.1: Hybrid Matrix Converter.

Fig. 3.1 shows the proposed topology of the hybrid matrix converter. The topology consist of the conventional nine bidirectional switches matrix converter coupled with a shunt active filter implemented on the input side and a series active filter on the output side. Both active filters consist of three-phase inverters connected by a common DC link. The shunt active filter compensates the input current harmonics produced by the switching operation of matrix converter. The series active filter compensates the output voltage harmonics. From a system-modeling point of view the matrix converter is considered as dual harmonics sources, current harmonic source in the input side, and voltage harmonic source in the output side. The small DC-link capacitor connecting the shunt and the series active filter is needed to compensate the oscillatory component of the instantaneous active power as will be explained next.

3.1 Active Filter Modulation

3.1.1 Input Current Conditioning

Considering the two waveforms at the input side, the three-phase input voltage is always sinusoidal and the three-phase current is extremely distorted due to the switching action of the matrix converter. The instantaneous three-phase active power at the input side p, can be given by

$$p_{in} = v_{in} \cdot v_{in}, \tag{3.1}$$

where "·" denotes the internal product of the two vectors. Equation (3.1) can also expressed in the conventional detention of power,

$$p_{in} = i_i v_a + i_b v_b + i_c v_c. \tag{3.2}$$

The instantaneous input reactive power vector of the three-phase system can be expressed as

$$\mathbf{q}_{in} = v_{in} \times i_{in}, \tag{3.3}$$

where "×" denotes the cross product of the voltage and current vectors. q can also be expressed

$$\mathbf{q}_{in} = \begin{bmatrix} q_a \\ q_b \\ q_c \end{bmatrix} = \begin{bmatrix} \begin{vmatrix} v_b & v_c \\ i_b & i_c \end{vmatrix} \\ \begin{vmatrix} v_c & v_a \\ i_c & i_a \end{vmatrix} \\ \begin{vmatrix} v_a & v_b \\ i_a & i_b \end{vmatrix} \end{bmatrix}, \tag{3.4}$$

the norm of the reactive power vector is

$$q_{in} = \|\mathbf{q}_{in}\| = \sqrt{q_a^2 + q_b^2 + q_c^2}. \tag{3.5}$$

We can decompose the source current into an active component i_{in-p}, and reactive component I_{in-q} in which

$$i_{in-p} = \begin{bmatrix} i_{a-p} \\ i_{b-p} \\ i_{c-p} \end{bmatrix} = \frac{p_{in} \cdot v_{in}}{v_{in} \cdot v_{in}}, \tag{3.6}$$

$$i_{in-q} = \begin{bmatrix} i_{a-q} \\ i_{b-q} \\ i_{a-q} \end{bmatrix} = \frac{q_{in} \times v_{in}}{v_{in} \cdot v_{in}}, \tag{3.7}$$

It is worth noting that the resultant current vector from the addition of the two currents i_{in-p} and i_{in-q} is always equal the source current i_{in}. Another observation is that the instantaneous power produced from $v_{in} \cdot i_{in-p}$ equal to the input power p_{in} , and the instantaneous power produced from $v_{in} \cdot i_{in-q}$ is always equal to zero. From this observation we can tell that i_{in-q} is not contributing to any power transmission from the source to the load. In fact if the instantaneous reactive current is kept $i_{in-q} \equiv 0$, the current i_{in} will be transmitting the same instantaneous active power p_{in} with unity power factor.

Consequently, the instantaneous active power p_{in} can be analyzed into two components,

$$p_{in} = \bar{p}_{in} + \tilde{p}_{in}, \tag{3.8}$$

where $\bar{p}_{in}$ is the direct component of the instantaneous active power and it represents the energy flow in the direction from the source to the matrix converter. And $\tilde{p}_{in}$ represents the oscillatory component of the instantaneous active power which is the energy exchanged between the source and the matrix converter. By eliminating the current component that produces $\tilde{p}_{in}$, the source current will be sinusoidal. It is important to note that we do not need an energy storing component to compensate the reactive power $\mathbf{q}_{in}$, because $\mathbf{q}_{in}$ represents the energy exchange among the three-phases. While an we energy storing component is required for compensating the oscillatory real power $\tilde{p}_{in}$, because $\tilde{p}_{in}$ is the real power exchanged between the source and the matrix converter.

Knowing this, we can generate the compensation current by selecting the appropriate power portion to be eliminated. The three-phase compensation current could be expressed as

$$i^*_{comp} == \frac{p^*_{comp} \cdot v_{in}}{v_{in} \cdot v_{in}} + \frac{q^*_{comp} \times v_{in}}{v_{in} \cdot v_{in}}, \tag{3.9}$$

where p^*_{comp} and q^*_{comp} can be determined from p_{in} and $\mathbf{q}_{in}$ depending on the compensation objectives.

3.1.2 Output Voltage Conditioning

The continuous current requirement of the matrix converter at the output side makes it a perfect fit to apply the series active filter to compensate the output voltage harmonics. By using the dual approach of the instantaneous reactive power theory we can define an instantaneous active output power and instantaneous reactive output power as

$$p_{out} = v_{out} \cdot i_{out}, \tag{3.10}$$

$$q_{out} = v_{out} \times i_{out}. \tag{3.11}$$

In turn, we define the instantaneous active voltage vector V_{out-p} and instantaneous reactive voltage vector V_{out-q} as

$$v_{out-p} = \begin{bmatrix} v_{A-p} \\ v_{B-p} \\ v_{C-p} \end{bmatrix} = \frac{p_{out} \cdot i_{out}^1}{i_{out}^1 \cdot i_{out}^1}, \tag{3.12}$$

$$v_{out-q} = \begin{bmatrix} v_{A-q} \\ v_{B-q} \\ v_{C-q} \end{bmatrix} = \frac{q_{out} \times i_{out}^1}{i_{out}^1 \cdot i_{out}^1}, \tag{3.13}$$

the superscript "^{1}" denotes the fundamental component of the output current. In most cases, series active filters is used in applications where the current is sinusoidal. However, this is not the case in the matrix converter and it requires further control effort to extract the fundamental component of the output current.

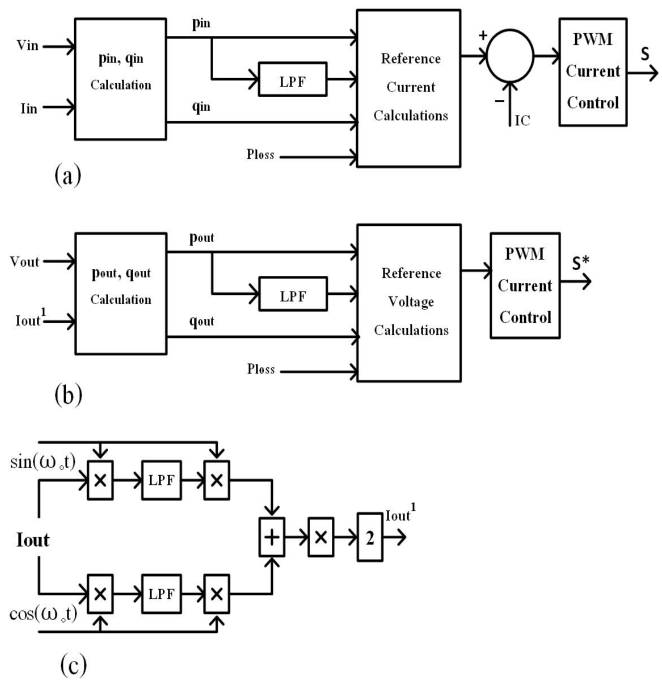

Figure 3.2: (a) Control circuit for the shunt active filter. (b) Control circuit for the series active filter. (c) Output current fundamental component extraction.

Similar observation can be made in the series active filter. The addition of the two voltage vectors v_{out-p} and v_{out-q} always equal the output voltage v_{out}. The instantaneous power produced from $v_{out-p} \cdot i_{out}$ is equal to the to the output power p_{out}, and the instantaneous power produced from $v_{out-q} \cdot i_{out}$ always equals zero.

The instantaneous output active power has a similar envelope to the instantaneous input active power. The only difference is the switching losses. We can also define an oscillatory component of the instantaneous output active power and the corresponding component of output voltage that causes this oscillation. By selecting the appropriate portions of power to be compensated we can write the equation of the compensating voltage as

$$v^*_{comp} == \frac{p^*_{comp} \cdot i^1_{out}}{i^1_{out} \cdot i^1_{out}} + \frac{q^*_{comp} \times i^1_{out}}{i^1_{out} \cdot i^1_{out}}, \tag{3.14}$$

where p^*_{comp} and q^*_{comp} can be assigned from p_{out} and q_{out} according to our one's compensation objectives.

The control circuits of the two active filters are shown in Fig. 3.2(a) includes computational circuits for the instantaneous input reactive power q_{in}, instantaneous oscillatory component of the input active power $\tilde{p}_{in}$, and instantaneous reactive component of the input current i_{in-q}, instantaneous oscillatory component of the input active current $\tilde{i}_{in-p}$. circuit Fig. 3.2(b) includes computational circuits for the instantaneous output reactive power q_{out}, instantaneous oscillatory component of the output active power $\tilde{p}_{out}$, and instantaneous reactive component of the output reactive voltage v_{out-q}, instantaneous oscillatory component of the output active voltage $\tilde{v}_{out-p}$. Fig. 3.2(c) shows the fundamental component extraction from the output current.

3.2 Simulation Results

A simulation model of the hybrid matrix converter as shown in Fig. 3.1 is build using MATLAB Simulink. In the shunt active filter a hysteresis current controller is used to track the instantaneous change of the inverter current and compare it back with the reference current.

It is necessary to control the voltage of the DC link capacitor by adding the power loss caused by the inverter switches. The active filter generates harmonics at its switching frequency, and it is necessary to filter out these harmonic, typically, small coupling inductor connected in series with the inverter output to eliminate these high frequency harmonics.

In our case the compensation powers are the reactive power and the oscillatory component of the the active power. The compensation of reactive power will guarantee that there is

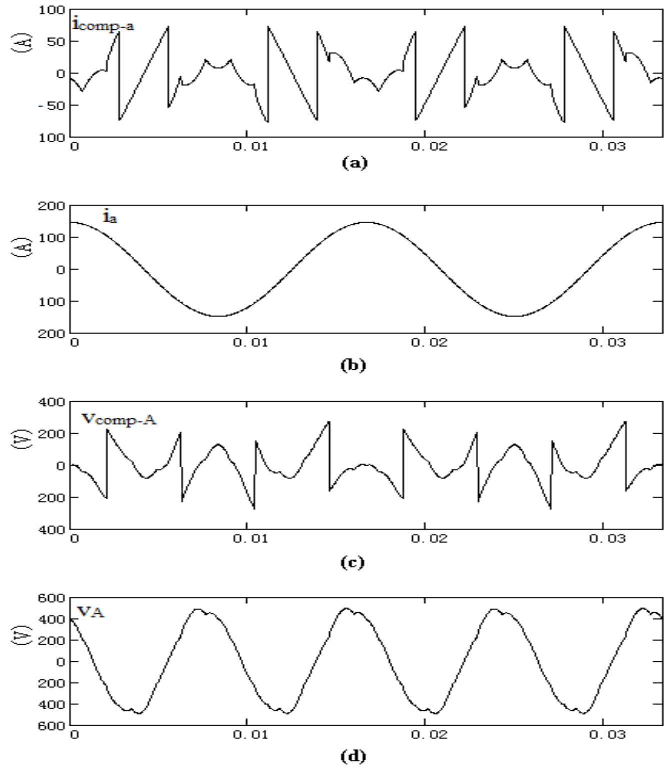

Figure 3.3: (a) Input compensation current injected by the shunt active filter. (b) Input current after using the shunt active filter. (c) Output compensation voltage injected by the series active filter. (d) Output line voltage after using the series active filter.

no phase shift between the current and the voltage in the input and output side, and the compensation of the oscillatory component of the active power will guarantee that the input current and the output voltage are sinusoidal.

Fig.3.3 shows the input current and the output voltage after the compensation. in the same phase with the input voltage which means that all the reactive power has been compensated effectively. The compensation of the oscillatory component of the input active power result in the sinusoidal shape of the input current. The same explanation can be made for the output voltage.

3.3 conclusion and remarks

In this chapter, shunt and series active filters have been implemented on the input and the output sides of the low-frequency modulated matrix converter, respectively. The analysis of the input and output power shows that the instantaneous reactive power theory can be applied in determining the compensation current of the input side and the compensation voltage for the output side of the matrix converter. The proposed topology is very efficient in medium voltage high power applications in which the conventional solution of passive filters is not effective. The hybrid matrix converter reduces the size of energy storage components, and provides higher reliability. The proposed topology could be utilized for mitigating the effect of voltage sag, especially when the matrix converter is used to drive sensitive loads. More detailed investigations will be reported in the future publications The DC link connecting the two active filters can provide the same voltage stiffness in the AC-DC-AC back to back inverters.

Chapter 4

Control Scheme of Hybrid Matrix Converter Operating Under Unbalanced Conditions

Hybrid matrix converters can potentially enable matrix converter in high-power applications that conventional matrix converters would not be able to attain. The hybrid matrix converter consists a main matrix converter that processes the bulk power conversion and an auxiliary back-back voltage source converter that improves the terminal power quality. Prior simulation study has successfully demonstrated that the superior spectral performance can be achieved. This chapter is focused on the control scheme of the hybrid matrix converter operating under balanced and unbalanced conditions.

4.1 Unbalanced Voltage Source

In the previous chapter the assumption is made that the source voltage is balanced, meaning the amplitudes of the three phase voltages are equal to each other and there is a 120^o phase shift among them. In case of unbalanced voltage source, further analysis needs to be considered to obtain the correct compensation current for the input stage of the matrix converter.

The unbalanced voltage source may include positive, negative, and zero sequence components according to the symmetrical component theory. The symmetrical component transformation is applied on both the input voltage and current to determine the sequence components.

$$\begin{bmatrix} v_{in0} \\ v_{in+} \\ v_{in-} \end{bmatrix} = \frac{1}{3} \begin{bmatrix} 1 & 1 & 1 \\ 1 & \alpha & \alpha^2 \\ 1 & \alpha^2 & \alpha \end{bmatrix} \begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix}, \tag{4.1}$$

The subscripts "0", "+", and "-" correspond to the zero, positive, and negative sequences, respectively. The complex number in the transformation matrix corresponds to the phase shift in the three phase system, $\alpha = 1\angle 120^o = e^{j\frac{2\pi}{3}}$,

$$\begin{bmatrix} i^n_{in0} \\ i^n_{in+} \\ i^n_{in-} \end{bmatrix} = \frac{1}{3} \begin{bmatrix} 1 & 1 & 1 \\ 1 & \alpha & \alpha^2 \\ 1 & \alpha^2 & \alpha \end{bmatrix} \begin{bmatrix} i^n_a \\ i^n_b \\ i^n_c \end{bmatrix}, \tag{4.2}$$

where "n" denotes the harmonic component order.

The time domain equivalent voltage and current can be derived from the phasors given by (4.1) and (4.2). By synthesizing the symmetrical components, the $a-b-c$ input voltages can be written as:

$$
\begin{cases}
v_a = \overbrace{\sqrt{2}V_{in0}\sin(\omega_i t + \phi_0)}^{v_{in0}} + \overbrace{\sqrt{2}V_{in+}\sin(\omega_i t + \phi_+)}^{v_{in+}} \\
\qquad\qquad + \overbrace{\sqrt{2}V_{in-}\sin(\omega_i t + \phi_-)}^{v_{in-}} \\
v_b = \sqrt{2}V_{in0}\sin(\omega_i t + \phi_0) + \sqrt{2}V_{in+}\sin(\omega_i t + \phi_+ - \frac{2\pi}{3}) \\
\qquad\qquad + \sqrt{2}V_{in-}\sin(\omega_i t + \phi_- + \frac{2\pi}{3}) \\
v_c = \sqrt{2}V_{in0}\sin(\omega_i t + \phi_0) + \sqrt{2}V_{in+}\sin(\omega_i t + \phi_+ + \frac{2\pi}{3}) \\
\qquad\qquad + \sqrt{2}V_{in-}\sin(\omega_i t + \phi_- - \frac{2\pi}{3}),
\end{cases} \tag{4.3}
$$

Similarly, the instantaneous input line currents are found to be

$$
\begin{cases}
i_a^n = \overbrace{\sqrt{2}I_{in0}^n\sin(\omega_i t + \phi_0)}^{i_{in0}^n} + \overbrace{\sqrt{2}I_{in+}^n\sin(\omega_i t + \phi_+)}^{i_{in+}^n} \\
\qquad\qquad + \overbrace{\sqrt{2}I_{in-}^n\sin(\omega_i t + \phi_-)}^{i_{in-}^n} \\
i_b^n = \sqrt{2}I_{in0}^n\sin(\omega_i t + \phi_0) + \sqrt{2}I_{in+}^n\sin(\omega_i t + \phi_+ - \frac{2\pi}{3}) \\
\qquad\qquad + \sqrt{2}I_{in-}^n\sin(\omega_i t + \phi_- + \frac{2\pi}{3}) \\
i_c^n = \sqrt{2}I_{in0}^n\sin(\omega_i t + \phi_0) + \sqrt{2}I_{in+}^n\sin(\omega_i t + \phi_+ + \frac{2\pi}{3}) \\
\qquad\qquad + \sqrt{2}I_{in-}^n\sin(\omega_i t + \phi_- - \frac{2\pi}{3}),
\end{cases} \tag{4.4}
$$

The input current is the result of adding the results of all the time domain currents from each harmonic.

$$
i_k = \sum_{n=1}^{\infty} I_k^n \qquad k = (a, b, c). \tag{4.5}
$$

The above description allows us to analyze the three phase unbalanced system in to two three phase balanced systems pulse zero sequence component. In the matrix converter case

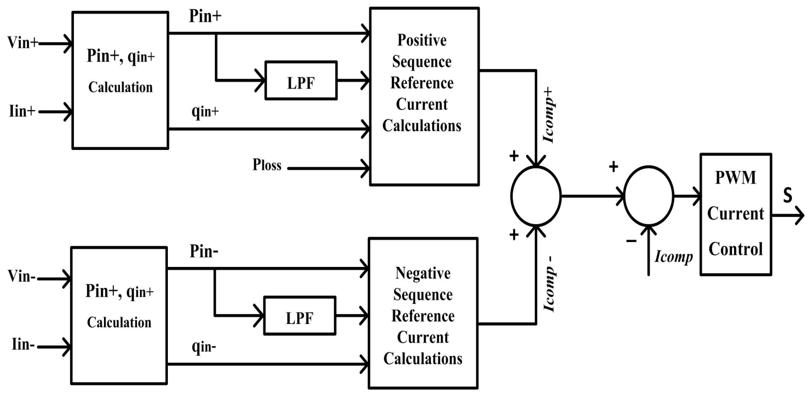

Figure 4.1: Control circuit of the shunt active filter in the case of unbalanced source voltage

we will not consider the zero sequence component to be compensated.

Fig.4.1 shows the two balanced systems (positive sequence system $v_{in+} = [v_{a+}\ v_{b+}\ v_{c+}]^T$, $i_{in+} = [i_{a+}\ i_{b+}\ i_{c+}]^T$ and the negative sequence system $v_{in-} = [v_{a-}\ v_{b-}\ v_{c-}]^T$, $i_{in-} = [i_{a-}\ i_{b-}\ i_{c-}]^T$) can be compensated in two different control loops, then the total compensating current will be the addition of the compensating current of the positive sequence system and the compensating current of the negative sequence system.

$$i^*_{comp+} = \frac{p^*_{comp+} \cdot v_{in+}}{v_{in+} \cdot v_{in+}} + \frac{q^*_{comp+} \times v_{in+}}{v_{in+} \cdot v_{in+}}, \tag{4.6}$$

$$i^*_{comp-} = \frac{p^*_{comp-} \cdot v_{in-}}{v_{in-} \cdot v_{in-}} + \frac{q^*_{comp-} \times v_{in-}}{v_{in+} \cdot v_{in+}}, \tag{4.7}$$

where

$$I^*_{comp} = I^*_{comp+} + I^*_{comp-}. \tag{4.8}$$

4.2 Unbalanced Load Current

In case of when different loads are connected to the matrix converter, each load will draw a different amount of current leading to a linearly independent three phase output current. The decomposition of the output voltage and current into it's symmetrical component is as follows:

$$\begin{bmatrix} v^n_{out0} \\ v^n_{out+} \\ v^n_{iout-} \end{bmatrix} = \frac{1}{3} \begin{bmatrix} 1 & 1 & 1 \\ 1 & \alpha & \alpha^2 \\ 1 & \alpha^2 & \alpha \end{bmatrix} \begin{bmatrix} v^n_A \\ v^n_B \\ v^n_C \end{bmatrix}, \tag{4.9}$$

$$\begin{bmatrix} i^1_{in0} \\ i^1_{in+} \\ i^1_{in-} \end{bmatrix} = \frac{1}{3} \begin{bmatrix} 1 & 1 & 1 \\ 1 & \alpha & \alpha^2 \\ 1 & \alpha^2 & \alpha \end{bmatrix} \begin{bmatrix} i^1_A \\ i^1_B \\ i^1_C \end{bmatrix}, \tag{4.10}$$

The time domain equivalent voltage and current can be derived from the phasors given by (4.9) and (4.10). By synthesizing the symmetrical components, the $A - B - C$ voltage and current can be written as:

$$
\begin{cases}
v_A^n = \overbrace{\sqrt{2}V_{out0}\sin(\omega_i t + \phi_0)}^{v_{out0}^n} + \overbrace{\sqrt{2}V_{out+}\sin(\omega_i t + \phi_+)}^{v_{out+}^n} \\
\qquad\qquad + \overbrace{\sqrt{2}V_{out-}\sin(\omega_i t + \phi_-)}^{v_{out-}^n} \\
v_B^n = \sqrt{2}V_{out0}\sin(\omega_i t + \phi_0) + \sqrt{2}V_{out+}\sin(\omega_i t + \phi_+ - \frac{2\pi}{3}) \\
\qquad\qquad +\sqrt{2}V_{out-}\sin(\omega_i t + \phi_- + \frac{2\pi}{3}) \\
v_C^n = \sqrt{2}V_{out0}\sin(\omega_i t + \phi_0) + \sqrt{2}V_{out+}\sin(\omega_i t + \phi_+ + \frac{2\pi}{3}) \\
\qquad\qquad +\sqrt{2}V_{out-}\sin(\omega_i t + \phi_- - \frac{2\pi}{3}),
\end{cases}
\tag{4.11}
$$

Similarly, the instantaneous line currents are found to be

$$
\begin{cases}
i_A^1 = \overbrace{\sqrt{2}I_{out0}^n\sin(\omega_i t + \phi_0)}^{i_{out0}^1} + \overbrace{\sqrt{2}I_{out+}^n\sin(\omega_i t + \phi_+)}^{i_{out+}^1} \\
\qquad\qquad + \overbrace{\sqrt{2}I_{out-}^n\sin(\omega_i t + \phi_-)}^{i_{out-}^1} \\
i_B^1 = \sqrt{2}I_{out0}^n\sin(\omega_i t + \phi_0) + \sqrt{2}I_{out+}^n\sin(\omega_i t + \phi_+ - \frac{2\pi}{3}) \\
\qquad\qquad +\sqrt{2}I_{out-}^n\sin(\omega_i t + \phi_- + \frac{2\pi}{3}) \\
i_C^1 = \sqrt{2}I_{out0}^n\sin(\omega_i t + \phi_0) + \sqrt{2}I_{out+}^n\sin(\omega_i t + \phi_+ + \frac{2\pi}{3}) \\
\qquad\qquad +\sqrt{2}I_{out-}^n\sin(\omega_i t + \phi_- - \frac{2\pi}{3}),
\end{cases}
\tag{4.12}
$$

then the output voltage is the result of adding all the harmonics together

$$
V_j = \sum_{n=1}^{\infty} V_j^n \qquad j = (A, B, C). \tag{4.13}
$$

The same consideration of zero sequence voltages will be made here. The control process is similar to the one we have in case of unbalanced source voltage, utilizing the output

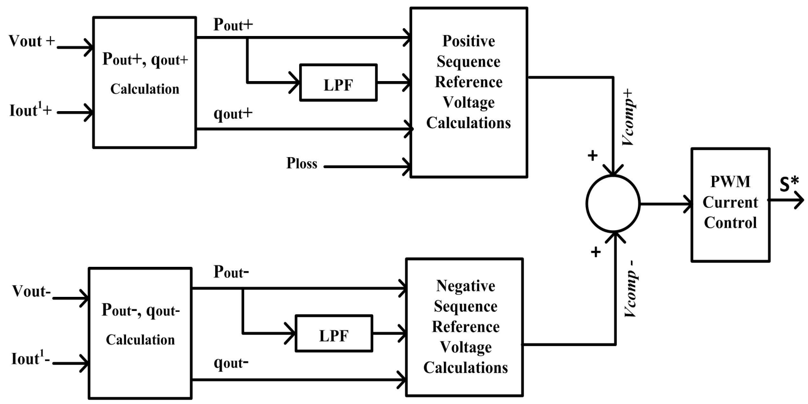

Figure 4.2: Control circuit of the series active filter in the case of unbalanced load

voltage and current into it's symmetrical component leaving us with two balanced systems, (positive sequence system $v_{out+} = [v_{A+}\ v_{B+}\ v_{C+}]^T$, $i^1_{out+} = [i^1_{A+}\ i^1_{B+}\ i^1_{C+}]^T$ and the negative sequence system $v_{out-} = [v_{A-}\ v_{B-}\ v_{C-}]^T$, $i^1_{out-} = [i^1_{A-}\ i^1_{B-}\ i^1_{C-}]^T$. The two balanced systems can be compensated into two different control loops as shown in Fig.4.2. The total compensating voltage will be the addition of the compensating voltage of the positive sequence system and the compensating voltage of the negative sequence system.

$$v^*_{comp+} = \frac{p^*_{comp+} \cdot I^1_{out+}}{I^1_{out+} \cdot I^1_{out+}} + \frac{q^*_{comp+} \times I^1_{out+}}{I^1_{out+} \cdot I^1_{out+}}, \tag{4.14}$$

$$v^*_{comp-} = \frac{p^*_{comp-} \cdot I^1_{out-}}{I^1_{out-} \cdot I^1_{out-}} + \frac{q^*_{comp-} \times I^1_{out-}}{I^1_{out-} \cdot I^1_{out-}}, \tag{4.15}$$

$$V^*_{comp} = V^*_{comp+} + V^*_{comp-}. \tag{4.16}$$

4.3 Simulation Results

Fig. 4.3, shows the output result of compensating the source current when the supply voltage is not balanced. The simulation shows that a sinusoidal input current can be achieved.

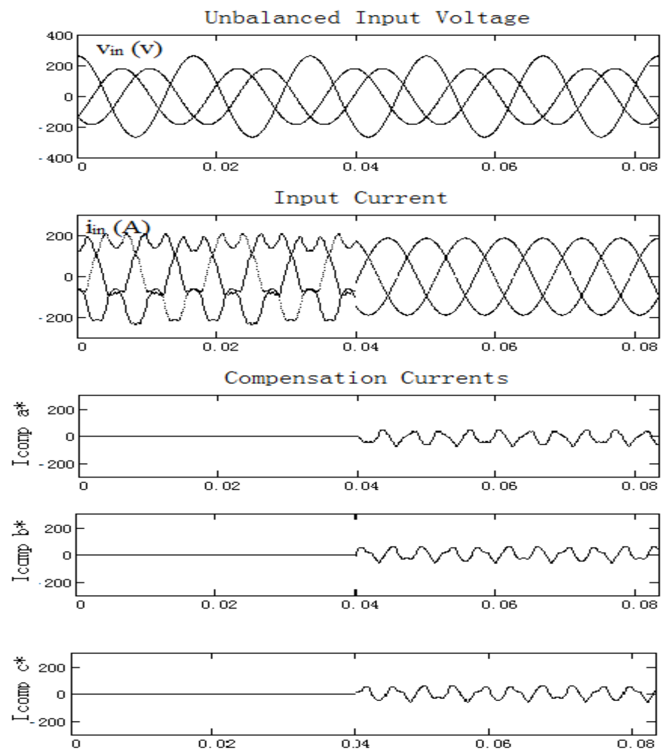

Figure 4.3: Output results under unbalanced voltage source conditions

Fig 4.4, shows the simulation in the case of an unbalanced load being fed by the matrix converter.

4.4 Conclusion

In this chapter, a general control scheme for the hybrid matrix converter control operating under different conditions (normal condition, unbalanced source voltage, and unbalanced load.

The analysis of the input and output power shows that the instantaneous reactive power theory can be applied in determining the compensation current of the input side and the compensation voltage for the output side of the matrix converter. Further analysis needs to be considered when the Hybrid Matrix Converter operates under abnormal conditions, those

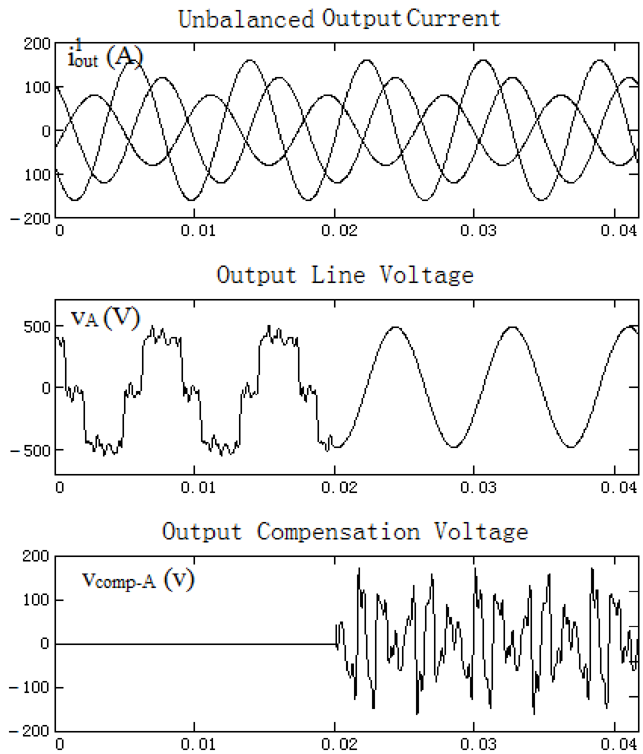

Figure 4.4: Output results under unbalanced load conditions

analyses include the symmetrical component theory.

The proposed topology can be utilized for mitigating the effect of voltage sag, especially when the matrix converter is used to drive sensitive loads.

Chapter 5

Critical Evaluation

This chapter is focused on conditioning the voltage and current waveform quality of a hybrid matrix converter that consists of a conventional nine-switch matrix converter and a back-to-back voltage source converter shoen in Fig. 3.1. Upon critical evaluation of the existing methods for shunt and series compensation, the fundamental limitations for achieving superior results have been identified. A new strategy based on power averaging for obtaining the reference compensating current and voltage has been proposed. The effectiveness of the proposed method has been evidenced by the simulation results for both the shunt and series compensation with concurrent presence of the harmonic components in voltages and currents.

Several AF control techniques have been presented in the literature [19][20][21]. Based on the operating principle, these techniques can be categorized into two groups. The first group of methods are based on instantaneous reactive power theory (IRPT) [22] and extract the reactive component of the power and the oscillatory component of the real power. The other methods are based on filtering techniques and extract the fundamental component of the current or voltage such as notch filter and fast Fourier transform (FFT) methods [23][24][25][26].

In this chapter different AF control strategies are presented and critically evaluated. The limitations of the existing control approaches have been clearly identified.

In frequency-varying environment, the best solution is to use adaptive approach. However, notch filter is only able for harmonic detection and can not extract the reactive components [26]. Further, such adaptive approaches might have convergence and robustness problems [27]. Other filtering techniques such Fourier methods are widely used. Fast Fourier transform requires high computational effort and discreet Fourier transform DFT requires synchronization tool such phase-locked-loop (PLL). Furthermore, they share the same limitation with adaptive notch filter of being unable to extract the reactive component. Therefore, only the analysis of adaptive notch filter is presented in the this paper.

Instantaneous reactive power theory is a time domain method that based on the law of conservation of energy.

It utilizes the concept that the non-active component of the voltage and current do not contribute in any energy transfer from the source to the load. The Instantaneous reactive power theory could undergo coordinate transformation to the synchronous reference frame. This transformation changes the oscillating AC variables to a DC variables and the harmonics appear in form of ripple in the DC signal. Although this approach provides extra filtering to the system, its required synchronizing tool such as PLL.

The instantaneous reactive power theory seems to be a good solution in obtaining the reactive components, it fails when both the current and the voltage contain harmonics. This because of the overlap in the frequencies of the current and the voltage harmonics. Therefore, it is incapable to obtain the correct compensating current or voltage when harmonics are present in both the current and the voltage.

To address these limitations of the aforementioned methods, a new control strategy is proposed. This proposed control method is able to effectively obtain the correct active component of current or voltage in cases where both the current and the voltage are non-sinusoidal and provide full control over the power factor.

5.1 Adaptive Notch Filter Method

Adaptive notch filtering is the technique that selectively extracts of a harmonic component of certain frequency. It features very powerful qualities where elimination of certain harmonics is required. Notch filter can estimate information embedded in the signal such as the amplitude, frequency and phase angle, the signal frequency. The dynamic behavior of the adaptive notch filter (ANF) can be characterized by the following deferential equations.

$$x^{''} + i^2\omega^2 x = 2\zeta\omega(u(t) - x^{'}) \tag{5.1}$$

$$\omega^{'} = -\gamma x\omega(u(t) - x^{'}) \tag{5.2}$$

where x is the integral of the fundamental component of the input signal $u(t)$, ω is the estimated frequency of the fundamental component; ζ and γ are adjustable real positive parameters that determine the accuracy and convergence speed of the ANF; $u(t)$ is the signal from which the fundamental component is to be extracted. For a sinusoidal input signal, the system described by (1) and (2) has a unique periodical orbit located at

$$\begin{pmatrix} x \\ x^{'} \\ \omega \end{pmatrix} = \begin{pmatrix} \frac{-A}{\omega_1}\cos(\omega_1 t + \phi_1) \\ A\sin(\omega_1 t + \phi_1) \\ \omega_1 \end{pmatrix}, \tag{5.3}$$

where the estimated component of the frequency ω is identical to its actual value ω_1, which is mathematically explained in [26]. A detailed implementation of the ANF is shown in Fig.5.1.

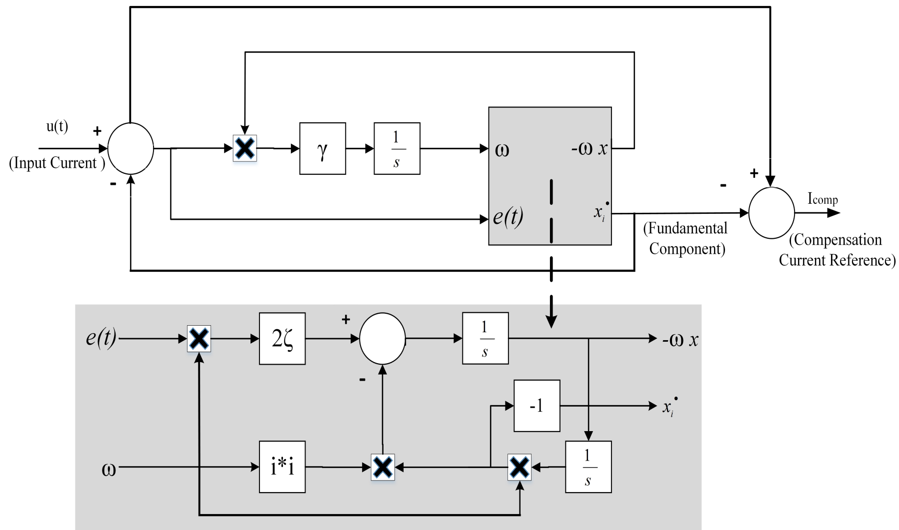

Figure 5.1: Detailed implementation of the adaptive notch filter.

The same control circuit is used to obtain the fundamental component of the output voltage.

5.2 Instantaneous Reactive Power Method

The instantaneous reactive power theory provides significant insight for understanding the power transferred from the source to the load and among the three phases. By elimination of the power component that does not contribute to the energy transmission from the source to the load, sinusoidal input current and sinusoidal output voltage can be achieved. The compensation system in instantaneous reactive power theory consist of only passive components and switches. Therefore, the net energy added or drown by the compensating system is zero.

$$P_C = 0; \quad P_S = P_L = P \tag{5.4}$$

where P_C is the real power from the compensator, P_S and P_L are the source average power, load average power, respectively. The average power is given by

$$P = P_X = \frac{1}{T_X} \int_{t-T_X}^{t} p(\tau) d\tau. \tag{5.5}$$

where T_X denote the averaging interval that can be zero, one fundamental cycle, one-half cycle, or multiple cycles, depending on compensation objectives and the passive components energy storage capacity; $p(\tau)$ is the instantaneous real power.

The instantaneous reactive power theory work can be explained from early definition of non-active current by Fryze [28].

$$i_p(t) = \frac{(v, i)}{(v, v)} \cdot v(t), \;\; i_q(t) = i(t) - i_p(t), \tag{5.6}$$

where i_p is the active current component, $v(t)$ and $i(t)$ is the reference voltage and current, respectively, and i_q is the non-active current component; (v, i) is the inner product of the voltage and the current over the interval $\{t - T_X, t\}$ with respect to weighting factor equal one, (v, v) is the inner product of the voltage and itself over the interval $\{t - T_X, t\}$ with respect to weighting factor equal one. They can be expressed as follows

$$(v, i) = \|v(t) \cdot i(t)\| = \frac{1}{T_X} \int_{t-T_X}^{t} v(\tau) \cdot i(\tau) d\tau = P \tag{5.7}$$

$$(v, v) = \|v(t)\|^2 = \frac{1}{T_X} \int_{t-T_X}^{t} v^2(\tau) d\tau = v_{rms}^2 \tag{5.8}$$

5.2.1 Active Current on the Input Side

On the input side of the matrix converter the source voltage is sinusoidal and the source current is not sinusoidal as it shown in Fig. 5.2(a). Therefore, they can be expressed by the following equations

$$v_a(t) = V_{Sf}\sin(\omega_i t) \tag{5.9}$$

$$i_a(t) = I_{Sf}\sin(\omega_i t - \alpha) + I_{Sh}\sin(\omega_{in} t + \beta_h) \tag{5.10}$$

where V_{Sf} is the amplitude of the voltage fundamental component; I_{Sf} is the amplitude of the current fundamental component; I_{Sh} is the amplitude of the h-th order harmonic, the corresponding average power is

$$P = \frac{V_{Sf} I_{Sf}}{2}\cos\alpha \tag{5.11}$$

It can be verified that the set of all harmonics in the current waveform $i_a(t)$ are orthogonal with respect to the voltage $v_a(t)$ over the interval of the integral. Therefore the corresponding average power is only a result of the fundamental components of the current i_a and the voltage $v_a(t)$.

Likewise, substituting (5.9) and (5.10) in (5.8) yields

$$v_{a-rms} = \frac{V_{Sf}}{\sqrt{2}} \tag{5.12}$$

By substituting (5.7),(5.9),and (5.12)in (5.6), we can obtain the active current component

$$i_p(t) = I_{Sf}\cos(\alpha)\sin(\omega t) \tag{5.13}$$

This proves that the instantaneous reactive power theory is able to obtain the active component of the current when the source voltage is sinusoidal and the source current contains harmonics.

5.2.2 Active Voltage on the Output Side

On the output side of the matrix converter, both load voltage and current include harmonics as shown in Fig. 5.2(a). Using a dual analogy to the active current defined in (6), we can define the active component v_p and the non-active component v_q of the voltage as

$$v_p(t) = \frac{(v,i)}{(i,i)} \cdot i(t), \;\; v_q(t) = v(t) - v_p(t), \tag{5.14}$$

where (i,i) is the inner product of the current and itself over the interval $\{t - T_X, t\}$ with respect to weighting factor equal one. It is expressed as follows

$$(i,i) = \|i(t)\|^2 = \frac{1}{T_X} \int_{t-T_X}^{t} i^2(\tau) d\tau = i_{rms}^2 \tag{5.15}$$

The load voltage $v_A(t)$ and the load current $i_A(t)$ can be expressed by the following equations

$$v_A(t) = V_{Lf} \sin(\omega_o t) + V_{Lh} \sin(\omega_{oh} t + \beta_h) \tag{5.16}$$

$$i_A(t) = I_{Lf} \sin(\omega_o t - \alpha) + I_{Lh} \sin(\omega_{on} t + \beta_h) \tag{5.17}$$

where V_{Lf} is the amplitude of the voltage fundamental component; V_{Lh} is the amplitude of the h-th order harmonic; I_{Lf} is the amplitude of the current fundamental component; I_{Lh} is the amplitude of the h-th order harmonic, the corresponding average power is

$$P = \frac{V_{Lf} I_{Lf}}{2} \cos\alpha + \frac{V_{Lh} I_{Lh}}{2} \cos\alpha_h \tag{5.18}$$

and the corresponding rms current is

$$i_{A-rms} = \sqrt{\frac{i_{Lf}^2 + i_{Lh}^2}{2}} \tag{5.19}$$

By substituting (5.15), (5.17), and (5.18) in (5.14) we can obtain the active component of the voltage

$$v_p(t) = \frac{V_{Lf} I_{Lf} \cos\alpha + V_{Lh} I_{Lh} \cos\alpha_h}{I_{Lf}^2 + I_{Lh}^2} \\ (I_{Lf} \sin(\omega_o t) + I_{Lh} \sin(\omega_{on} t + \beta_h)) \tag{5.20}$$

It is observed from (5.20) that the active component of the current is not sinusoidal. Therefore, the instantaneous reactive power theory is not able to achieve a sinusoidal load voltage. An FFT filter is used to obtain the fundamental component of the load current before it is used in the control loop [29]. This limitation of the instantaneous reactive power theory will be resolved by the proposed method in the next section.

By expanding the same approach for a three phase system we can define the compensation current and the compensation voltage as follows

$$i_{comp} = \frac{\tilde{p} \cdot v_{in}}{v_{in} \cdot v_{in}} + \frac{q \times v_{in}}{v_{in} \cdot v_{in}} \tag{5.21}$$

$$v_{comp} = \frac{\tilde{p} \cdot i_{out}^1}{i_{out}^1 \cdot i_{out}^1} + \frac{q \times i_{out}^1}{i_{out}^1 \cdot i_{out}^1} \tag{5.22}$$

where $v_{in} = [v_a \ v_b \ v_c]^T$ is the input voltage vector; $i_{out}^1 = [i_A \ i_B \ i_C]^T$ is the fundamental output current vector; $\tilde{p}$ is the oscillatory component of the real power; q is the reactive power. The control block diagrams for the input current compensation and the output voltage compensation are shown in Fig. 3(a). Both output current and voltage of the

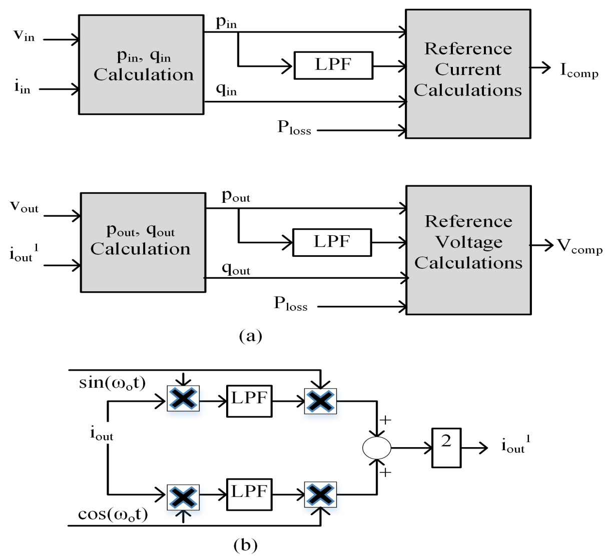

Figure 5.2: Block diagrams for (a) determining the reference signals for compensation current and voltage using IRPT; (b) extracting the fundamental component of the output current.

matrix converter contain a significant amount of harmonics, in such cases the typical control approach of instantaneous reactive power theory will not be able to obtain the correct active voltage component. For such reason the fundamental component of the output current has to be extracted first before it is employed in the control circuit as shown in Fig. 5.2(b).

5.3 Proposed Average Power Method

Average power compensation (APC) is based on the assumption that the total averaged power drawn from the source is equal to the power obtained by the sinusoidal active components of the current or the voltage. This assumption is mathematically represented in (5.23) for single-phase system,

$$\int_{t-T_X}^{t} v(t)i(t)dt = \int_{t-T_X}^{t} v_p(t)i(t)dt \tag{5.23}$$

where v_p is the active component of the voltage. Assuming the the active component of the voltage is sinusoidal, therefore,

$$v_p(t) = A\sin(\omega_o t) \tag{5.24}$$

We can estimate the amplitude of the active voltage component by substituting (5.16) and (5.17) in (5.23), the amplitude A is determined by

$$A = v_{af}\cos(\alpha) + \frac{v_{ah} i_{ah}\cos(\alpha_h)}{i_{af}} \tag{5.25}$$

From the active voltage amplitude in (5.25), it can be observed that the active component of the current v_p is not equal to the fundamental component, which is the fundamental reason that the ANF methods will not achieve maximum power transfer. A dual approach can be easily implemented to obtain the active component of the input current. The block diagram of averaged power method for the input current and the output voltage compensation are shown in Fig. 5.3(a) and (b), respectively.

5.4 Simulation Results and Evaluation

To evaluate the performance of the presented methods detailed simulation models have been constructed based on the three phase hybrid matrix converter shown in Fig. **??**. In each simulation model the control method is applied to compensate both the input current and the output voltage harmonics. The typical input and output current and voltage waveforms of the main matrix converter are shown in Fig. 5(a) and (b). Fig. 8(c) shows the input source current after compensation using adaptive notch filter method, by which the fundamental component of the input current is calculated. The harmonic content is obtained by

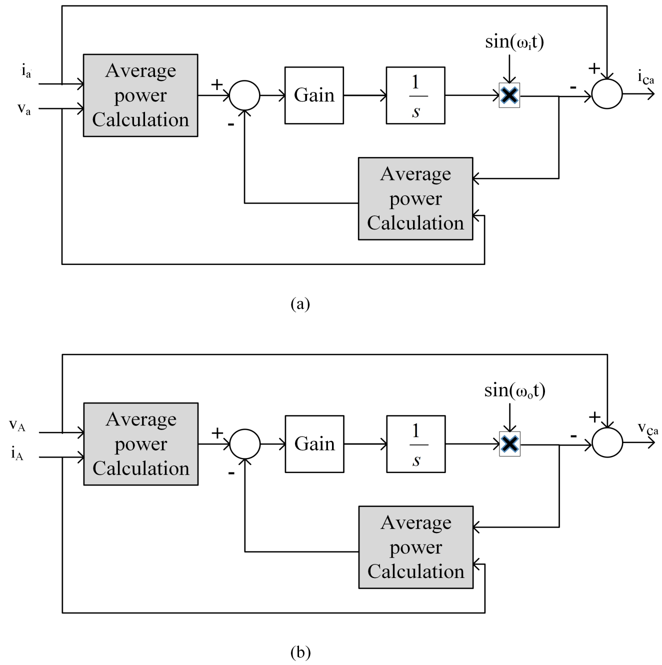

Figure 5.3: Average power compensation method control for (a) input current; (b) output voltage.

subtracting the fundamental component from the original current waveform.

Although the notch filter is easy to implement, it is unable to correct the phase shift between the input voltage and the input current, Therefore, the ANF method is ineffective when power factor correction is required. Another disadvantage for ANF is its ineffectiveness when both the voltage and the current contain harmonics such as the output voltage and current of the matrix converter. In this case some of the power transferred to the load takes place at harmonic frequencies. Compensation to achieve output voltage equal to its fundamental component simply will not achieve maximum power transfer. The output voltage using ANF is shown in Fig. 5(d).

The instantaneous reactive power theory allow for full control of the phase shift of the input current and subsequently a unity power factor could be achieved. However, IRP method is still unable to compensate the output voltage because both output current and

Table 5.1: A summary of the comparative evaluation of the three methods

Methods	ANF	IRP	APC
Power factor control	Unable to control the PF, Because it is only obtains the fundamental component with its phase shift.	Can achieve unity PF by compensating all the reactive component q.	Fully control the PF by selecting the required phase shift in the reference signal.
Obtaining the active component of current and voltage	Only able to obtain the fundamental component of the current or the voltage, therefore it fails to achieve maximum power transfer when both the current and voltage contain harmonics.	Fail to obtain the active component of load voltage, because, both load voltage and current contain harmonics. Therefore, it is required to filter the load current first before it is employed in the control circuit. ,	Able to successfully obtain the active component of the output voltage without any need of filtering the output current.
Transient Response (due to input voltage increase by by 0.3 pu).	Requires at least one cycle to reach the steady state.	current drops during the transient period.	Requires one cycle to reach the steady state without any overshoot in the current.
Complexly of Implementation.	Many design parameters need to be optimized to achieve low THD. Higher order than IRP and APC.	Obtaining the load current fundamental component increases the computational efforts significantly.	Requires less computational effort. The same control circuit can be used for all cases.
Total harmonic distortion THD	7.8%	2.5%	2%

voltage present harmonic. The only way to achieve sinusoidal output load voltage is to first obtain the fundamental component of the current before it is employed in the control circuit as shown in Fig. 3(b). The compensated input source current and the output load voltage using IRP are shown in Fig. 5(e) and (f).

The average power method in contrast can provide unity power factor when compensating the input current. It is also able to obtain the active component of the output voltage without any need of filtering the output current. The input source current and the output load voltage using APC are shown in Fig. 5(g) and (h). A summary of the comparative evaluation of the methods is listed in Table 1.

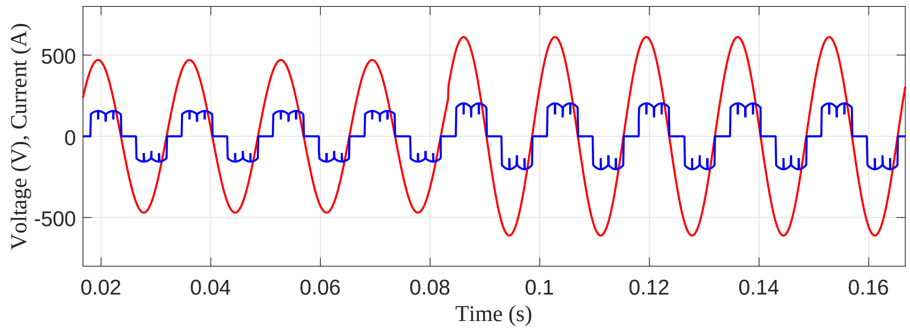

Figure 5.4: Input line voltage v_{ab} and current i_a.

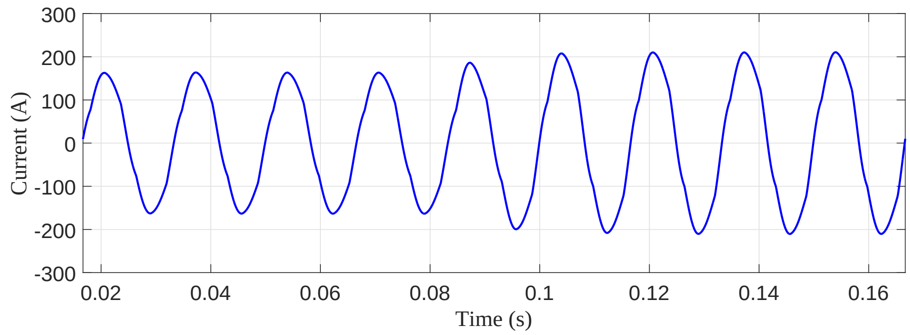

Figure 5.5: Input current i_a using Notch Filter.

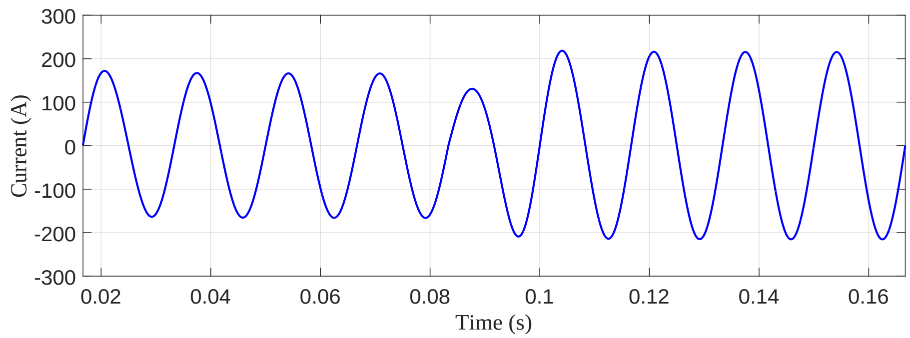

Figure 5.6: Input current i_a using IRPT.

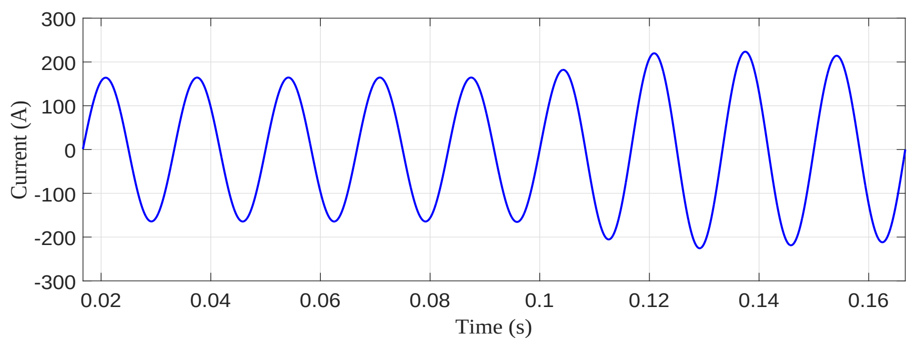

Figure 5.7: Input current i_a using AP.

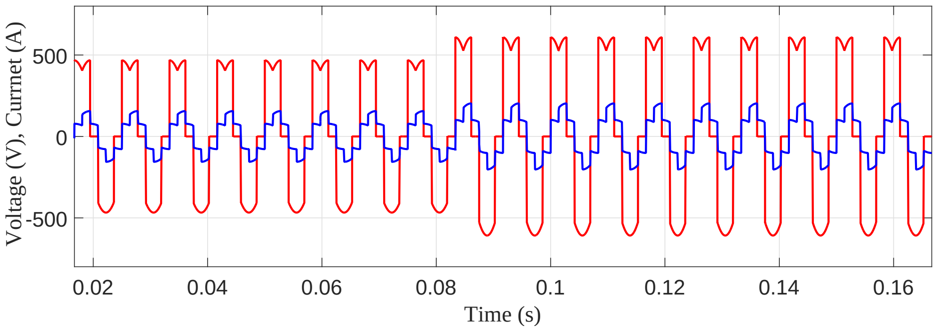

Figure 5.8: Output line voltage v_{AB} and current i_A.

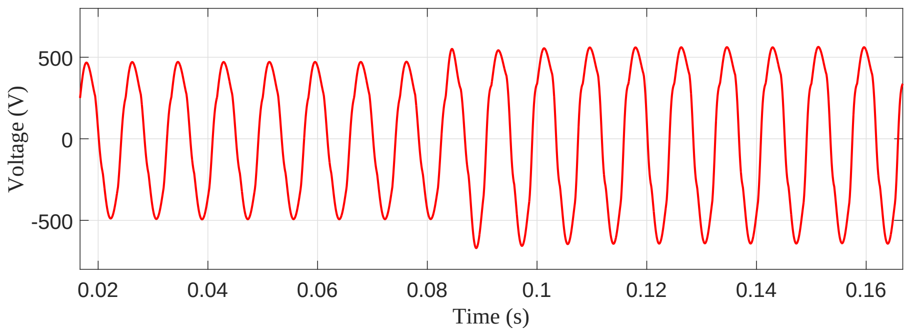

Figure 5.9: Output voltage v_A using Notch Filter.

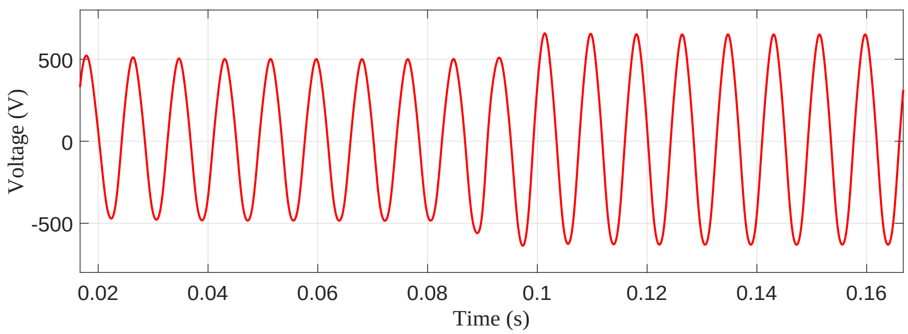

Figure 5.10: Output voltage v_A using IRPT.

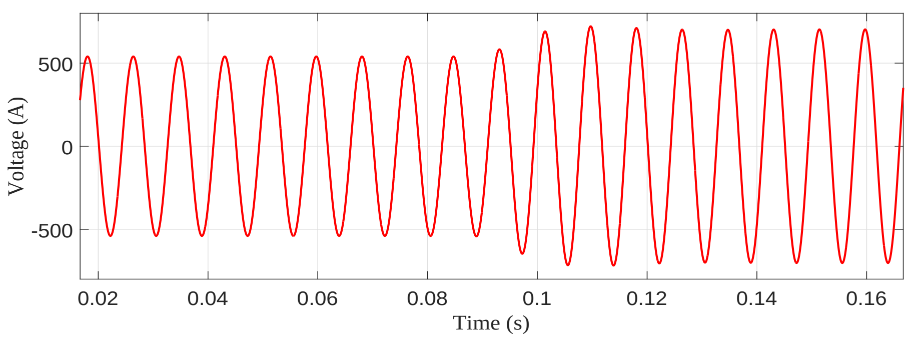

Figure 5.11: Output voltage v_A using AP.

5.5 Conclusion

This paper has presented the preliminary results for the evaluation of existing approaches to shunt and series conditioning of the currents and voltages. The critically evaluated methods are based on either instantaneous reactive power theory or fast Fourier transform. The main limitation for IRPT based method lies in its ineffectiveness when the harmonics are concurrently present in voltage and current while the limitation for FFT based method is its inability to compensate the fundamental component. To address the limitations associated with the existing methods, a new method based on power averaging has been proposed. The effectiveness of the proposed method has been verified by the simulation results obtained from a detailed MathCAD/Simulink model. Furthermore, experimental work has been planned and the experimental results will be included in the final manuscript.

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