

Improved operation and design of Modular multi-level converters Through Harmonic Injection

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Abstract

Modular multilevel converters (MMC), amongst all multi-level converters, have received significant interest recently due to their modularity, capability, availability and excellent performance characteristics. The MMC can be scaled to very high voltage levels and power ratings while providing good power quality and the opportunity to operate with redundancy to enable high availability. Operation at high voltage levels and power rating requires a high number of submodules (SMs) each with its own capacitor assembly of significant size to store the required energy. The required energy storage and hence capacitor volume can be influenced and reduced using harmonic injection techniques employing either voltage and/or current injection. Approaches using 3rd harmonic voltage injection and 2nd harmonic current injection have been reported, but, the combination of the 3rd harmonic injection, i.e. voltage and/or current, and 2nd harmonic current together have not been fully explored (including the freedom to adjust the relative time displacements).

This thesis evaluates the influence of the classical and phase shifted 3rd harmonic voltage and current injection in combination with the 2nd harmonic current on the energy storage requirements including energy pulsation reduction and SM capacitor sizing. Mathematical models are derived. The phase-shifted 3rd harmonic voltage injection concept is presented considering its influence on reducing the peak synthesised voltage and two possible methods are proposed utilising the reduced peak synthesised voltage. Increasing the output voltage at the MMC terminals using the phase-shifted 3rd harmonic is considered, and a lookup table method is presented allowing further increase of the output voltage and reducing the line current. Power losses are reduced since the AC voltage can be increased allowing a reduction in current with an appropriate transformer design. The influence of the reduced peak synthesised voltage is further investigated considering its impact on resizing the SM capacitor size by allowing higher voltage ripple. A new design is achieved using phase shifted 3rd harmonic voltage injection. The continuous mathematical models are validated using a model with a full representation of switching, and selected methods are then validated experimentally in the laboratory.

List of Terms

V_{dc} - Converter DC voltage

I_{dc} - Converter DC current

V_S - Grid AC voltage

V_C - Converter AC voltage

v_C - Converter instantaneous AC voltage

$\hat{V}_C, \hat{V}_{C1}$ - Converter AC voltage amplitude; fundamental

$V_{C.a}, V_{C.b}, V_{C.c}$ - Converter AC voltage phase A, B and C

$v_{C.a}, v_{C.b}, v_{C.c}$ - Converter AC instantaneous voltage of phase A, B and C

$\hat{V}_{C.a}, \hat{V}_{C.b}, \hat{V}_{C.c}$ - Converter voltage amplitude of phase A, B and C; fundamental

$\hat{V}_{C1}$ - Converter voltage amplitude; fundamental

$\hat{V}_{C2}$ - Converter voltage amplitude; 2nd harmonic

$\hat{V}_{C3}$ - Converter voltage amplitude; 3rd harmonic

I_C - Converter and grid AC current

i_C - Converter and grid instantaneous AC current

$\hat{I}_{C1}$ - Converter current amplitude; fundamental

$\hat{I}_{C2}$ - Converter current amplitude; 2nd harmonic component

$\hat{I}_{C3}$ - Converter current amplitude; 3rd harmonic component

θ_{v1} - Phase angle of fundamental component converter voltage

θ_{v3} - Phase angle of the 3rd harmonic voltage component

θ_{i1} - Phase angle of fundamental component converter current

θ_{i2} - Phase angle of 2nd harmonic component converter current

$L_{limb}, X_{L_{limb}}$ - Equivalent limb inductor and its impedance of each limb
 $C_{eq.SM}, X_{C_{eq.SM}}$ - Equivalent capacitance and its impedance of the limb SM capacitors
 N_{SM} - Number of submodules
 C_{SM} - Submodule capacitance
 $V_{C_{SM}}$ - Submodule capacitor average voltage
 $v_{C_{SM}}$ - Submodule capacitor instantaneous voltage
 $V_{C_{SM}.n}$ - Submodule capacitor nominal voltage
 $\Delta V_{C_{SM}}, \Delta V$ - Capacitor voltage ripple
 $\Sigma V_{C_{SM}}$ - Total submodule capacitor average voltage
 $\Sigma V_{C_{SM}.n}$ - Total submodule capacitor nominal voltage
 $\Sigma v_{C_{SM}}$ - Total submodule capacitor instantaneous voltage
 $V_{red}, k_{V_{red}}$ - Reduced peak synthesised voltage and its percentage in per unit
 $V_{red,p.u}$ - Reduced peak synthesised voltage in per unit
 $v_{pos.a}, v_{pos.b}, v_{pos.c}$ - Positive limb instantaneous voltage of phase A, B and C
 $v_{neg.a}, v_{neg.b}, v_{neg.c}$ - Negative limb instantaneous voltage of phase A, B and C
 $i_{pos.a}, i_{pos.b}, i_{pos.c}$ - Positive limb instantaneous current of phase A, B and C
 $i_{neg.a}, i_{neg.b}, i_{neg.c}$ - Negative limb instantaneous current of phase A, B and C
 P_{dc} - Converter DC side average power
 S - AC apparent power
 P_{ac} - Grid AC side average (active) power
 Q - Grid AC side reactive power
 $\emptyset$ - Power factor

T - Grid period

$p_{pos.a}, \bar{P}_{pos.a}$ - Positive limb instantaneous and average power of phase A

$p_{pos.b}, \bar{P}_{pos.b}$ - Positive limb instantaneous and average power of phase B

$p_{pos.c}, \bar{P}_{pos.c}$ - Positive limb instantaneous and average power of phase C

$p_{neg.a}, \bar{P}_{neg.a}$ - Negative limb instantaneous and average power of phase A

$p_{neg.b}, \bar{P}_{neg.b}$ - Negative limb instantaneous and average power of phase B

$p_{neg.c}, \bar{P}_{neg.c}$ - Negative limb instantaneous and average power of phase C

$e_{pos.a}, E_{pos.a}$ - Positive limb instantaneous and average energy of phase A

$e_{pos.c}, E_{pos.c}$ - Positive limb instantaneous and average energy of phase C

$e_{neg.a}, E_{neg.a}$ - Negative limb instantaneous and average energy of phase A

$e_{neg.c}, E_{neg.c}$ - Negative limb instantaneous and average energy of phase C

$E_{pos.n}, E_{neg.n}$ - Positive and negative limb nominal energy

$E_{SM.n}$ - SM capacitor nominal energy

ΔE - Energy deviation

ΔE_{max} - Maximum energy deviation

W_{rated} - Rated stored energy capability

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Chapter 1

1 Introduction

1.1 Motivation

Many environmental and infrastructural challenges are introduced due to the growing demand for electricity. This demand promotes the construction of new transmission lines to improve the security of electricity supply as well as considering the environment requirements. High voltage direct current (HVDC) transmission promises to secure electrical supply while keeping environmental impact low. It is also more efficient especially for transmitting bulk power over long distances since there are lower losses associated with DC transmission compared to high alternating voltage (HVAC). In the case of VSC-HVDC, considered in this thesis, the ratings have reached almost 1GW at a voltage of $\pm 320\text{kV}$.

The recent development of power semi-conductor devices and digital controllers has made high voltage direct current (HVDC) technology more attractive especially for certain applications such as long distance power transmission. This application requires operation at voltages far beyond the rating of available power semi-conductor devices [1]. Multilevel converter topologies are used as one way to meet these voltage requirements. Besides the high voltage capability, multilevel converters such as modular multilevel converters (MMC), can be used to provide lower distortion due to their ability to generate multi-step output voltages compared to the conventional two-level voltage source converters (VSCs). Such properties make the use of multilevel converters widely recognised in HVDC transmission applications. The multilevel-based power converters are becoming the core of HVDC transmission particularly AC/DC or DC/AC converters.

The multilevel topologies patented in 1971 such as a transformer-less power supply [2]. The three classical multilevel converter topologies have been widely studied and developed by academia and industry. These three converters are the neutral point clamped (NPC)[3], the cascaded H-bridge (CHB)[4], and the flying capacitor (FCC) [5].

Newer topologies have also been built and used in practical applications [6] such as the three-level active NPC (3L-ANPC) [7] and the modular multilevel converter (MMC) [8, 9]. Currently, the MMC is considered as the most promising topology in the area of HVDC transmission [10] and is recognised as state-of-the-art among existing topologies [11]. The compact and high efficient hybrid modular multilevel VSC converter is also being considered particularly desirable for HVDC applications [12, 13] but it is currently under development.

MMC converters are based on the modularity concept [9]. Each module comprises a capacitor and power semi-conductor devices. Generally, the most voltage sensitive components in the MMC are the capacitors and semi-conductor devices [14, 15]. In a practical MMC converter, a large number of submodules (SM) are used which comprise a large number of power semi-conductor devices and capacitors. For instance, to achieve 320kV using an IGBT rating of 1500V, the required number SMs is 2×214 SM/leg. Considering the redundancy and to ensure the reliability, this number might be increased. This large number of SMs, and therefore their capacitors and semi-conductor devices, can lead to an increase in the size of the MMC and its operation requirements such as its auxiliary systems, for example, the cooling system. It is reported that harmonic injection has its impact on the energy pulsation and the synthesised voltage by the MMC limbs [16, 17]. 3rd harmonic current has also been considered in the literature for alternative topologies including the Alternate-Arm multilevel converter (AAC) [18]. The impact of the optimum 2nd harmonic current injection in the circulating current on reducing the SM capacitor voltage ripple was also considered in the literature [19].

In this thesis work, the power converters are considered the core of HVDC transmission which will be discussed considering multilevel converters particularly AC to DC or vice versa. Therefore, firstly a literature review on the existing and the most recognisable converter topologies is presented highlighting the reasons behind the vast growth demand towards HVDC. Then, the possible solutions that might be applied in order to improve the operation or design of the state-of-the-art MMC is considered by applying harmonic injections. Therefore, an investigation on the impact of classical 3rd harmonic voltage, current and 2nd harmonic current injection on energy storage requirement is first presented. The focus is then moved to a detailed and extensive investigation of classical and phase-shifted 3rd harmonic voltage injection. The benefits of the phase-shifted 3rd

harmonic in terms of improving the design and operation of the MMC are analysed and validated in simulation and experimentally.

1.2 Contribution

As discussed above, 3rd harmonic voltage and current injections have already been considered individually for the MMC and for the AAC. However, a detailed study has not been done yet considering the impact of all these harmonics at the same time. The first part of the work done in this thesis aimed at filling this gap, providing a generalised study of the impact of current and voltage harmonics. Comparisons are presented investigating the impact of harmonics on the operation and design of the MMC including energy storage requirements. A review of the impact of 3rd harmonic voltage based on an analytical study is first presented.

The main work presented by this thesis can be recognised by two main parts. Each part covers analytical study and simulation results are achieved (in the case where there are also experimental results, it will be indicated clearly). These parts are summarised as follow;

- An extensive study considering the impact of harmonic injection on energy storage requirements.
- The impact of the harmonic injection, in particular 3rd harmonic voltage injection on the development of the MMC operation and design including the operating voltage and capacitor sizing.

With regards to the first part of the thesis, e.g. the impact of using harmonic injection on energy storage requirements, the thesis is aimed to filling the areas that have not been considered by previous studies. The achievements of the first part listed above are:

- For the first time, the impact of 3rd harmonic current injection for the MMC is investigated with and without 3rd harmonic voltage injection and compared to the 3rd harmonic voltage injection method.
- The thesis proposed the phase shift of the 3rd harmonic injection concept, including voltage and current and investigating their impact the operation of the MMC including, voltages, currents, power and energy.
- The thesis also presents for the first time a comparison between 3rd harmonic voltage injection and 2nd harmonic current injection with and without 3rd

harmonic voltage injection on the operation of the MMC. In particular, energy storage requirements and rated voltage are considered. The result of the analysis indicates some applications that can benefit from applying 3rd harmonic voltage and/or current injection.

It has already been reported in the literature that using 2nd harmonic current is beneficial with regards to reducing energy deviation and therefore the capacitor voltage ripple [18]. This thesis gives further insights by considering both 3rd harmonic voltage and 2nd harmonic current injection.

In the second part of the thesis, two new methods based on using phase-shifted 3rd harmonic voltage injection are proposed for improving the operation or design of the MMC. These methods utilise the advantage of optimal 3rd harmonic voltage injection and the proposed phase-shift of the 3rd harmonic in reducing peak synthesised voltage. Limitations and other aspects due to using phase-shifted 3rd harmonic injection are discussed and lookup tables are implemented for the optimal phase shift. Due to this reduction, a first option is to reduce the AC current by increasing the AC voltage. The other option that can be exploited is the reduction of the size of SM capacitors by allowing a larger voltage ripple. This is considered beneficial because the submodule capacitor size dominates the physical size and weight of a submodule. The work achieved by the second part of the thesis can be summarised by;

- A literature review on the optimal 3rd harmonic voltage injection and its impact on the reduced peak of the synthesised voltage.
- For the first time, limits of classical and phase-shifted 3rd harmonic voltage injection, positive and negative phase shift, is presented due to applying the optimal solution of the classical 3rd harmonic injection.
- Discussion about the proposed phase-shifted 3rd harmonic voltage injection (positive and negative phase shift) by this thesis on the synthesised voltage.
- Discussing the limits and other aspects due to using phase-shifted 3rd harmonic injection.
- Proposing lookup table method for applying the proposed phase shift in order to develop a method to improve the operation of the MMC by increase the output voltage at the MMC terminals.

- Proposing two methods, which can be applied to benefit from the reduced peak synthesised voltage due to using the first time proposed phase-shifted 3rd harmonic voltage injection, including, developing a method of increasing the output voltage at the MMC terminals and the, for the first time, capacitor sizing.
- Discussing other aspects including loss reduction and converter transformers design due to using phase-shifted 3rd harmonic injection.

The two different methods proposed by the thesis to improve the operation and design of the MMC due to the reduced peak synthesised voltage can be summarised as follows:

- Applying the proposed method of the phase-shifted 3rd harmonic can be used to develop increasing the voltage at MMC terminals. This can be allow further increasing of the output voltage at the MMC terminals due to the increase of the fundamental component of the synthesised voltage. This increase can lower the line currents with proper scaling of the transformer turns ratio. As a result, current through the converter will be reduced and conduction loss will be reduced.
- A reduced peak synthesised voltage can allow higher voltage ripple. This can be used to reduce the size of the SM capacitor by allowing higher voltage ripple. This second option considers a given design of the MMC and only revisits the size of the SM capacitors by exploiting the increased available voltage margin after phase-shifted 3rd harmonic injection.

1.3 Thesis Structure

To achieved the thesis aims, the thesis is structured as follows.

Chapter 2 introduces a literature review on HVDC transmission systems and their advantage over HVAC systems. Some of the HVDC applications are discussed considering the conventional line-commutated current source converters (CSC) and self-commutated voltage source converters (VSC). This chapter ends with an introduction to the modularity concept presenting the most recent and recognised topologies.

In Chapter 3 a literature review on the operating principles of 3-phase MMC is presented. The operating principle of an MMC is introduced in an intuitive way and then the detailed mathematical model is discussed.

In Chapter 4 an analytical study is presented considering the impact of 3rd harmonic voltage, current and 2nd current injection on energy storage requirement including energy deviation. First a review of the 3rd harmonic voltage injection is considered. For the first time, 3rd harmonic current injection is presented considering two possible conditions including without and with 3rd harmonic voltage injection. Moreover, a comparison is presented comparing both cases with the operation with conventional 3rd harmonic voltage injection. The 2nd harmonic current injection is also considered and a comparison between conventional and 3rd harmonic voltage injection is presented. A generalised mathematical model is also presented. This chapter ends with a summary including a proposed application based on the considered harmonic injection. For power transmission applications, the analysis shows that 3rd harmonic voltage injection with 2nd harmonic current injection has the most promising performances. The 2nd harmonic currents reduces capacitor voltage ripples whilst 3rd harmonic reduces the peak of the synthesised voltage.

In chapter 5 further investigation on the impact of 3rd harmonic voltage injection on SM capacitor size requirements is presented. This investigation is based on a comparison between traditional operation and 3rd harmonic injection. First a review of the capacitor sizing based on the energy deviation is presented. The chapter ends with a discussion about two important aspects including the energy storage requirements and the peak voltage synthesised by the MMC limbs. It is observed that 3rd harmonic voltage injection is more effective when considered to reduce peak of the synthesised voltage.

In Chapter 6 there is further investigation about using 3rd harmonic voltage injection on the reduced peak synthesised voltage. The optimum 3rd harmonic voltage injection is discussed including the voltage margin. The phase-shifted 3rd harmonic voltage injection concept is introduced which then is used in two different ways to improve the operation and design of the MMC. The result of the analysis presented in this chapter proposes for the first time two different methods that can benefit from the reduced peak synthesised voltage due to phase-shifted 3rd harmonic voltage injection. The first method leads to increased output voltage at the MMC terminal while the second method reduces the synthesised voltage in order to reduce the size of the SM capacitors.

Chapter 7 discusses the MMC modelling and control design. This chapter validates the analysis and mathematical model and the proposed methods in this thesis. This chapter ends with simulation results based on a switching model.

In Chapter 8 the practical implementation of one of the proposed methods in Chapter 6 is presented. An experimental rig is used to validate the switching model and the control schemes presented in Chapter 7. The chapter ends with comparison between the simulation results and the experimental results.

In Chapter 9 conclusions and further work are discussed.

Chapter 2

2 Multilevel Converters for HVDC Transmission - Literature Review

This chapter presents a brief literature review on the existing HVDC power transmission technologies and current practices in the industry particularly multilevel converter topologies. The benefits of HVDC on the HVAC transmission is first discussed. The power converters are considered the core of HVDC transmission which will be discussed in Section 2.4 considering the multilevel converters particularly AC to DC or vice versa converters. A literature review of the modular multilevel converter concept and particularly the generalised modular multilevel converter (MMC) is introduced in Section 2.6.

2.1 Introduction

Since inaugurating the first HVDC transmission based on VSC with PWM for HVDC in 1997 [20], the development of High Voltage Direct Current (HVDC) technology has attracted the movement towards more applications for an HVDC transmission system. This has not only been motivated by the original advantages of HVDC being more efficient in terms of power transmission, but also by other technical and environmental aspects [21]. This has led to making it recognized more widely especially for certain applications, for instance power transmission of bulk energy using long-distance overhead lines and submarine cables. In fact, this is can be attributed to its advantages over the AC transmission system of there being no constraints or technical limitations with respect to the transmission distance [21-23]. Moreover, as synchronism is not required for DC transmission systems, it is considered to be a desirable solution in some applications where interconnection between asynchronous systems are required. Usually back-to-back converters are used allowing interconnection and providing a buffer between the two systems such as the interconnection between two systems (50Hz and 60Hz) by Shin-Shinano II using back-to-back converter [24]. This will allow linkage between systems with different frequencies. By considering voltage, frequency and

phase angle, and also impedance fluctuations, the HVDC system has immunity from such fluctuations [22].

Since the first HVDC mercury arc valve systems, the development of HVDC technology has increased significantly. Furthermore, this has led to successes and rapid progression to the use of thyristor valves. The development increased rapidly until recently when the applications of the HVDC system become more desirable especially when using voltage-source converters (VSC) [21, 22]. This has its influence on broadening the range of HVDC transmission applications, VSC based as opposed to the conventional HVDC, based on thyristor technology and known as current-source converters (CSC). A HVDC transmission system enables linking remote generation locations of renewable energy sources such as offshore wind farms [22, 23].

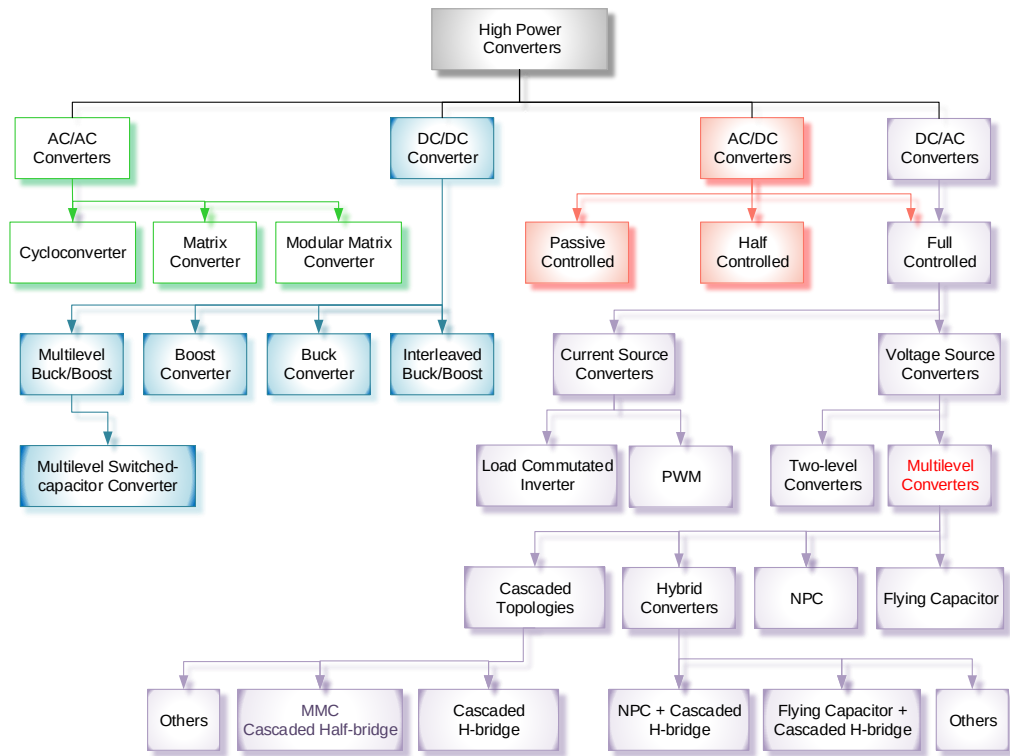


Figure 2.1: The Classification of high power converters.

The core component of the HVDC transmission system is the power converter which converts from AC to DC or vice versa. Recent study however proposed a multilevel DC/DC converter topology based on multilevel switched capacitor converter as an indication of the growth interest towards HVDC systems. This also means that the technology of power electronic devices has developed rapidly. The classification of the power converters is illustrated by Figure 2.1 [6, 25]. However, our interest is on

discussing multilevel converters connecting between high voltage alternating current (HVAC) and HVDC systems especially with high power and high voltages ratings (Gigawatts and hundreds of kilovolts) especially in the case of weak AC interconnections or offshore wind farms. Therefore, AC/AC or DC/DC converters are not considered in this thesis which might be considered in future work.

The demand for HVDC systems has increased due to the many limitations and other aspects of the AC transmission systems. These aspects are not only technical but also economic and environmental [22]. The improvement of the chip characteristics of the technology of Insulated-Gate Bipolar Transistors (IGBT), which led to reduction not only in power loss but also the cost, IGBTs has influence on increasing the range of the applications based on voltage-source converters (VSC) in the HVDC system. The history of the IGBT development is reported by [26]. In comparison to the conventional CSC based HVDC system, pulse-width modulation (PWM) can also be used in the VSC system. However, VSCs have the ability to control both active and reactive power [23, 27, 28]. This can be achieved by controlling the phase and magnitude of the voltage. More recently, new HVDC technologies based on VSCs have been introduced to the market and have already been constructed for many projects and applications. The most common application is known as a back-to-back converter VSC based technology, commercially known as HVDC Light, as a next step from using the conventional CSC based technology [22, 27].

Since high power conversion has contributed new techniques, many projects have been constructed and some of them are under construction [10, 22]. The first generation of VSC power converters was based on a series connection of forced commutated devices. Multilevel converters followed and then the modular-multilevel converters (MMC). This can be attributed to the variation of the applied PWM schemes that exist which have been introduced and discussed in the literature in more detail [28]. This led to an increase in demand for pushing the HVDC technology further to be used widely by introducing the MMC concept, known as HVDC Plus, which was introduced in 2003. Recently, many alternative topologies have been introduced in the literature, some of which have been examined not only with simulation analysis but also with experimental validation [29-32]. Further reading about the history of the HVDC transmission and its technology can found in literature, text books and journals [21-23, 27, 33].

2.2 Benefits of HVDC over HVAC transmission

Recently, certain applications such as power transmission have become especially attracted by the characteristics and the development of HVDC technology compared to the HVAC system where some of the recent HVDC projects based on VSCs are listed in Table 2.1 [10]. The most significant advantage of the HVDC transmission is that there is no limit with regard to the stability related to the length of the transmission system or the power transmitted. Therefore, it is an attractive choice in case of long distance bulk power transmission. This is because of the AC limitations as regards long distances. These limitations are economic, stability issue and voltage variation [23]. Furthermore, it is an economical alternative choice to AC for to remote resources such as large-scale wind farms. However, this depends on the distance and amount of power to be transferred [21], but a DC system can transmit for longer distances using fewer lines compared to the AC system with the same rated power [22].

Table 2.1: Recent commercial HVDC projects based on VSC (MMC).

Project Name	Power (MW)	Voltage ($\pm$ kV)	Cable length (km)	Commissioning (company-year)
Helwin 1	576	+/- 250	130	Siemens-2015
Dolwin 1	800	+/-320	165	ABB-2015
Borwin 2	800	+/-300	200	Siemens-2015
Sylwin	864	+/-320	205	Siemens-2015
Dolwin 2	900	+/-320	135	ABB-2015
Dolwin 3	900	+/-320	160	GE (Alstom)-2018

Other advantages of HVDC that can also be considered are asynchronous interconnections between systems operating with different frequencies and, therefore, no synchro-checking is required. Furthermore, black start up is possible in the case of VSC based technology [22]. Moreover, improving the AC system stability is one of the major achievements that can be gained. In contrast with the AC transmission system, limitation of short circuit levels and limitation of faults is also achieved. It is also beneficial in terms of voltage control and power flow control, in which, both active and reactive power can be controlled independently. Furthermore, the total cost of the transmission can be reduced as there are fewer lines than with AC for the same power rating. Furthermore, it can be beneficial to the environment [22, 23].

2.3 HVDC Applications

The applications of HVDC transmission can be divided into different categories. Although the underlying principle for the selection is often economically based, there may be other reasons for its selection. HVDC may be the only feasible way to connect asynchronous networks, utilise long underground and submarine cable circuits, reducing fault currents, bypass network congestion and the mitigation of environmental concerns such as physical, pollution, electrical and magnetic fields, and also visual impacts [22, 23].

One of the basic categories is the bulk energy transmission through long distance. This can be applied to either overhead lines or submarine cables, meaning that offshore energy can be easily and effectively delivered. It is reported that, the interconnection of asynchronous networks and multi-terminal systems can also be possible and power delivery can be made for large urban areas [22, 23, 34]. Some other applications that have been reported in literature are, for instance, reactive power compensation, Static Synchronous Compensation (STATCOM), and protection purposes. These applications are based on MMC and other alternative topologies [35-37].

2.4 Multilevel converters for HVDC Transmission

The main components of a HDVC power transmission system are reviewed in this section. Of particular interest will be the two main converter technologies used in HVDC transmission systems as this research is on the investigation of a converter circuit for HVDC systems.

Modern HVDC transmission systems are based on two converter technologies. The first is the conventional line-commutated current source converters (CSC). The other is the self-commutated voltage source converter (VSC) [22].

2.4.1 Line-Commutated Current Source Converters CSCs

In this section, the use of the CSC in HVDC schemes is discussed. The advantages and disadvantages of having CSC as the main converter in HVDC schemes will be presented.

Line-commutated current source converters (CSC) have been used in conventional HVDC transmission for a while. The CSC is based on thyristor valves. In order for it to operate, a synchronous voltage source is required. The conventional CSC valve based

thyristor is a three-phase controllable full-wave bridge converter. It also referred to as a six-pulse controllable bridge. Figure 2.2 illustrates the conventional HVDC with current source converters based on six-pulse [22, 23].

The characteristic of the CSC operation due to its commutation is a harmonic ripple of six times the fundamental frequency in the dc output voltage. In order to achieve the required voltage and current rating, each bridge comprises six controlled switching thyristor valves. The thyristor valves are a number of series-connected thyristors. Moreover, for better harmonic characteristic profile, two six-pulse bridges can be connected via a YYΔ three-phase three-winding transformer to obtain twelve-pulse converters displaced by 30° . Therefore, the connection of two CSCs can be used to increase the voltage rating and eliminate some of the AC and DC harmonics. For instance, the AC current harmonics have frequencies of $12n - 1$ whilst the DC voltages contain harmonic of order $12n$. This results in the reduction of the harmonic filtering requirements such as the size which reduces the overall cost [22, 38].

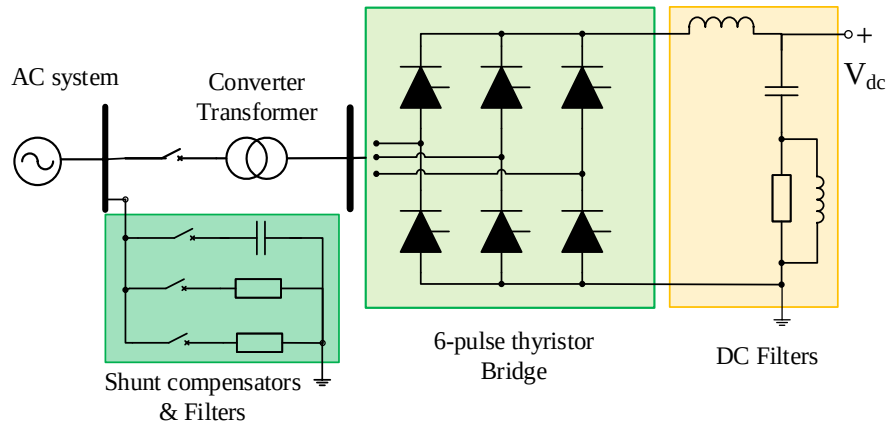


Figure 2.2: Conventional HVDC based on current source converters (CSC).

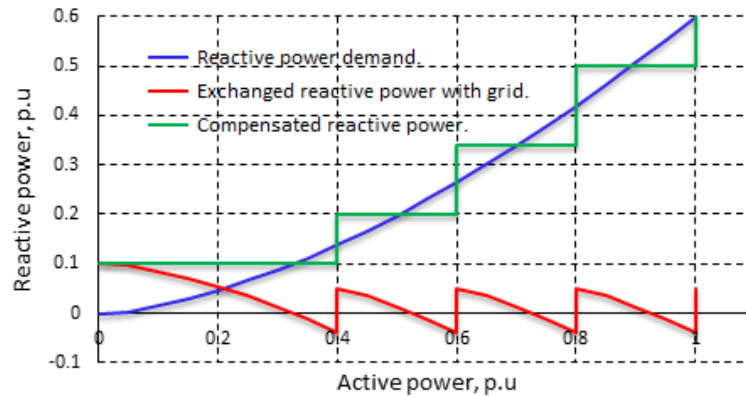


Figure 2.3: Variation and compensation of reactive power for a conventional HVDC converter station CSC based [22].

CSCs, however, require a strong synchronous voltage source in order to commute. Moreover, as the commutation is the transfer of the current from one phase to another, it is reported that, the network should have a symmetrical short circuit capacity at the point of connection with the converter of at least twice the converter rating. This is to ensure the operation of the CSC. Moreover, during the conversion process of the CSCs, reactive power is required. To operate the CSC, reactive power is required for the conversion process where this demand of reactive power is provided using an AC filter. The AC filter comprises shunt banks, series capacitors or any suitable shunt compensation devices such as a STATCOM which is integrated as part of the converter station. The CSC demands only the reactive power of the conversion process and any overflow of the reactive power from the AC filter affects the AC voltage. Therefore, to keep the AC voltage within the desired voltage tolerance, the difference in reactive power as shown by the red line in Figure 2.3 needs to be kept within the limits. The reactive power demand, the compensated reactive power and the exchanged reactive power with the AC grid are illustrated by Figure 2.3 [22, 23]. As can be seen, the demand of reactive power is approximately 60% of the power transmitted in the case of a full load [21].

2.4.2 Self-Commutated Voltage Source Converters VSCs

Since 1997 [20], the use of VSCs for HVDC power transmission has become popular due to the many advantages of voltage source converters. In this section, the benefits of using VSC in HVDC systems will be presented and some of the VSC circuits considered for HVDC systems will be reviewed.

The self-commutated Voltage Source Converters (VSC), known also as forced – commutated converters, have been used in HVDC transmission [27]. The VSC uses insulated-gate bipolar transistors (IGBT) with a PWM switching technique. This is conventionally marketed as HVDC Light by ABB [22, 32]. In comparison to the CSCs, the VSCs have benefits for the whole system. For instance, active and reactive power can be controlled rapidly and independently. Furthermore, the reactive power can be controlled in the AC side without affecting the DC link. So, this gives the flexibility for placement anywhere in the grid [21, 23, 28]. Many HVDC links based on VSC are in operation worldwide some of them are reported by Table 2.1. The self-commutated Voltage Source Converters (VSC), known also as forced –commutated converters, have

been used in HVDC transmission [27]. The VSC uses insulated-gate bipolar transistors (IGBT) with a PWM switching technique. This is conventionally marketed as HVDC Light by ABB [22, 32]. Figure 2.4 illustrates single line diagram of an HVDC transmission system based on HVDC Light. In comparison to the CSCs, the VSCs have benefits for the whole system. For instance, active and reactive power can be controlled rapidly and independently. Furthermore, the reactive power can be controlled in the AC side without affecting the DC link. So, this gives the flexibility for placement anywhere in the grid [21, 23, 28].

Another characteristic is that there is no restriction of limitation of the AC symmetrical short circuit capacity. Moreover, it can be used as a synchronous generator, so a black start can be permitted. It also provides dynamic support to the AC system at each converter terminal so that voltage stability can also be improved [21, 22]. Figure 2.4 shows a conventional HVDC system based on VSC (HVDC Light). The voltage source converter, as described above, is based on IGBTs which can be connected in series to achieve the required voltage level. There are however some issues related to this approach, which multilevel and modular multilevel converters overcome [39].

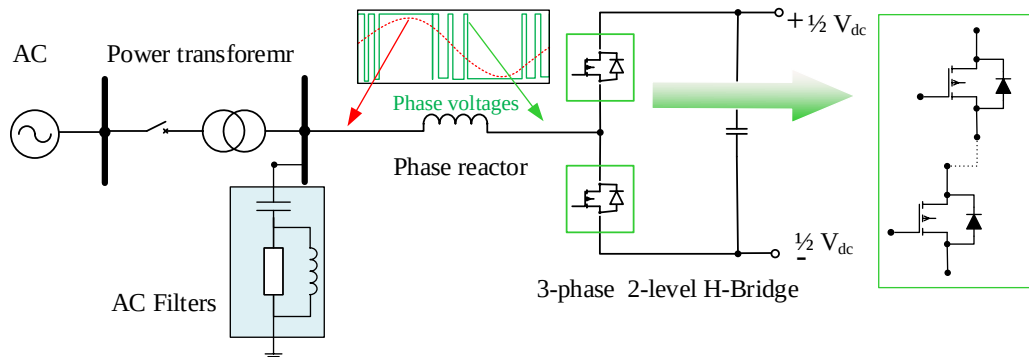


Figure 2.4: Conventional HVDC Light based on two-level H-bridge VSC.

There is no limitation to the transferred power due to reactive power requirements as there is in the case of the CSC in which about 60% of transmitted at full load is due to reactive power. Therefore, this leads to increased transfer capacity not only for sending, but for receiving as well. Also, there is no reactive power demand for the commutation as the IGBTs are force commutated. In fact, not only can reactive power can be controlled to regulate AC system voltage, but also active power as well. This is actually similar to a synchronous generator. Figure 2.5 illustrates the power operating range of

the MMC where the reactive power is limited by upper and lower voltage limits [21, 23, 28].

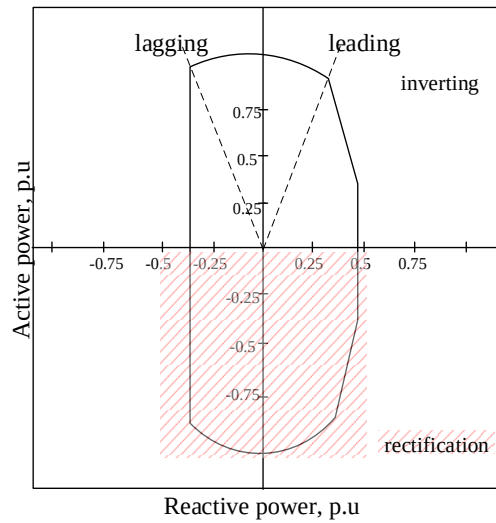


Figure 2.5: Power operation range of HVDC based on VSC [22].

2.5 Conventional Multilevel Voltage Source Converters

The IGBT power devices have become suitable and reliable for high power and high voltage applications. Some new projects were presented in Table 2.1 where the converter is intended to operate with 900MW and ± 320 kV. To meet such HVDC application, the IGBT modules used in such converters must be set the standards for quality and reliability and manufactured for high power. They can be connected in series to achieve the desired voltage rating for the two-level scheme where the rating of the individual IGBT module, for example, from 1.7kV-6.5kV and current from 600A-2400A which meets the requirements for high power and high voltage converters. For instance, the Toshiba MG1200FXF1US53C IGBT is rating 3.3 kV, 1.2 kA. However, this has disadvantages of causing voltage sharing problems. Moreover, voltage stress on the devices, dv/dt will be high [21, 39].

Alternative multilevel schemes and topologies have advantages over the two-level conversion. To overcome the issues associated with voltage sharing, for instance, clamping devices can be used. Furthermore, the output waveforms can also be improved since number of levels is increased which has the impact on reducing the stress on the devices by reducing the dv/dt . This also has an impact on eliminating the low-order harmonics. However, the number of levels is limited by many factors such as, for instance, the converter structure becoming more complicated as the voltage rating and

converter levels are increased. There are three classical topologies offering the multilevel solution including; diode-clamped converters often known as Diode-clamped converter (DCC), e.g. 3-level DCC known as neutral-point clamped (NPC) [3], flying-capacitor converters (FCC) [5] and isolated H-bridge circuit, cascaded H-bridge circuit (CHB) [4].

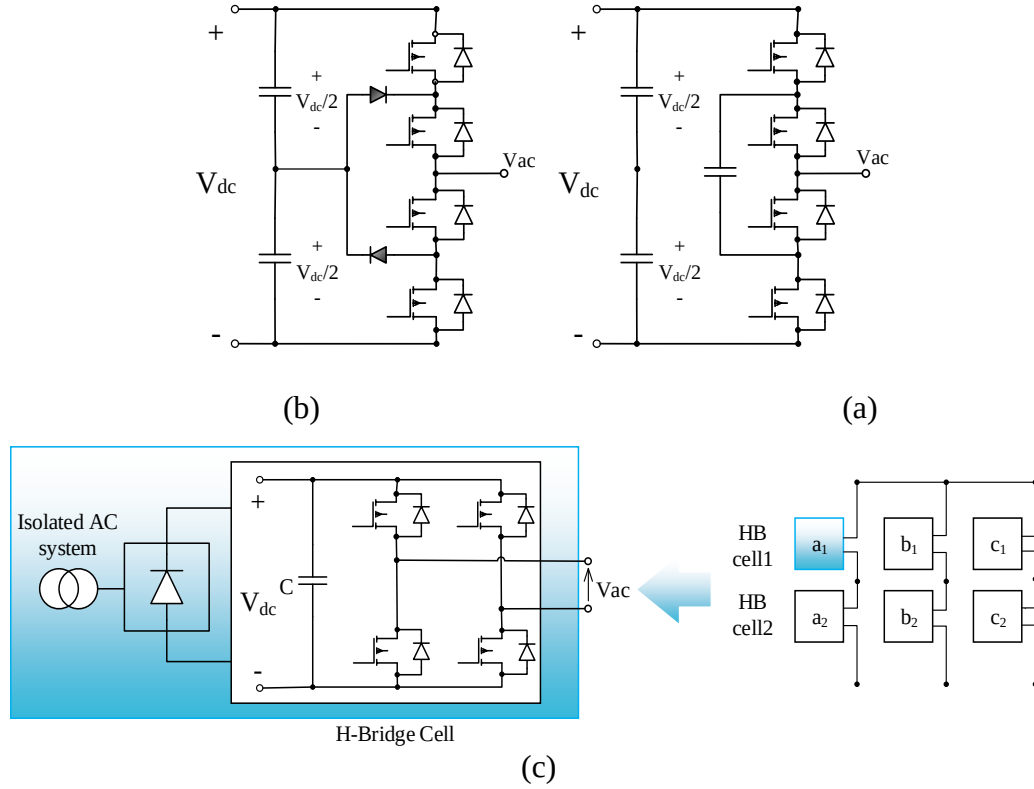


Figure 2.6: Conventional multilevel converters (a) FCC, (b) NPC, (c) CHB.

However, the applications of previous multilevel converters are limited due to many drawbacks such as the complexity of the structure and maintenance. Moreover, some of them require extra control circuits in order to balance the voltages of the large capacitors when their number is increased and they are series connected. Furthermore, there are issues related to the availability due to the difficulty of enabling redundancy in the operation [21, 39]. Figure 2.6 illustrates the structure of FCC, NPC and CHB converters.

Theoretically, FCC and DCC can be extended to any number of levels. However, this requires an increase in control and design, i.e. structure, complexity where a series connection of switches, capacitors, or diodes are required. To increase the number of levels, for the FCC which has a modular structure for instance, an addition of FCC cell with increased capacitor voltage rating is required. This means that the capacitors are

not rated at the same voltage levels. To overcome this issue, a stack of capacitors with the same ratings can be used[40]. While in case of DCC, with the increase of number of the levels, not only the number of clamping diodes increases dramatically, but also it requires to balance more DC link capacitor voltages [1].

Due to these issues, the two topologies, the FCC and DCC, cannot be extended to higher number of levels and therefore are used in their 4 level (FC converter) and 3-level (DCC; for 3-level know as NPC converter) configurations, respectively, in industrial applications such as medium voltage (MV) drives applications [6, 40].

Compared to the FCC or NPC, the CHB converter with its cascaded modular structure can be extended to any number of levels. this can be achieved by adding additional, identical or asymmetrical converter cells. However, the CHB requires multiple isolated DC sources. Alternatively, a complex phase-shifting transformer is required especially when applied with multi-pulse diode rectifiers [6, 40]. A capacitor voltages balancing mechanism has to be implemented compared to FCC or NPC converters. Despite these issues, the CHB converter successes in many applications such as active filters (AFs), static synchronous compensator (STATCOM), and other parts of flexible AC transmission systems (FACTS) [6]. In addition, the major role played by the CHB converter can be seen in the Universal Flexible Power Management system (UNIFLEX-PM) [41, 42].

Despite the fact that the previously mentioned topologies had succeeded in industry, they could not meet the requirements of HVDC transmission. To overcome the drawbacks of the previously mentioned converters, modular multilevel converter topology (MMC) was first introduced in 2003 [9] and the first project based on this MMC topology went into operation in 2010 [43]. The authors in [9] presented a new topology based on VSC suitable for HVDC transmission. Figure 2.8 illustrates its structure and the concept of the modular construction which can be easily recognized. It is reported that MMC can be built using similar submodules (SM) with standard components. Moreover, it is believed that the level of voltage and power can be increased by using a suitable number of SMs.

2.6 Modular Multilevel Converter Concept

The MMC is a multilevel VSC that has recently gained attention and is being considered extensively for HVDC power transmission and other grid applications such as STATCOM, energy storage systems etc. The modular multilevel converters are based on two key parts including the “director switches” (DS) and “wave-shaping” circuits (WSC) [44]. The DS and WSC functionality can be illustrated by Figure 2.7 where the DS controls the direction of the current from AC/DC while the WSC is responsible for shaping the output voltage waveform.

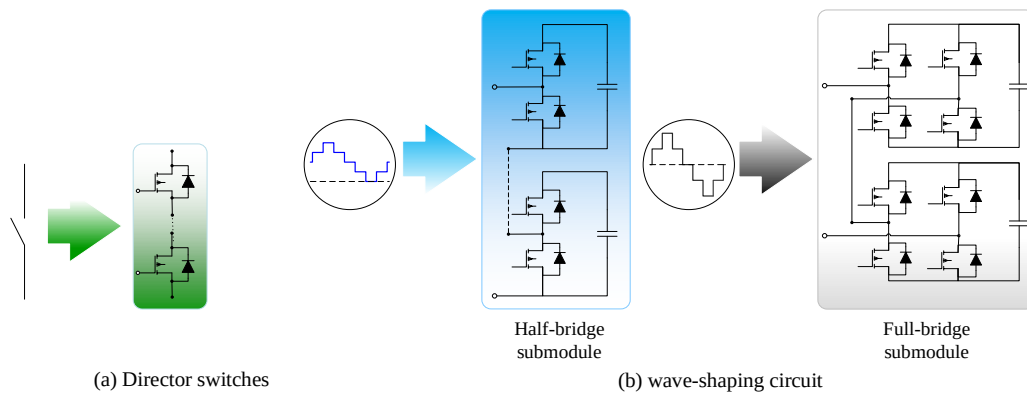


Figure 2.7: Series switch and wave-shaping circuits [44].

The WSC based on half-bridges has only two outputs, i.e. zero or positive voltage, whilst the full-bridge based WSC can output negative, zero or positive voltage. This means that the voltage polarity can be inverted using the full-bridge based WSC. Based on the connection of these two circuits with each other, the converter can be based on two connection strategies, series or parallel connection. However, based on the position of the WSC, modular multilevel converters can generally be divided into two categories including a “wave-shaping circuit” on AC side or on DC side [44].

2.6.1 Generalised MMC Topology

The first presentation of the MMC concept was in 2003[45] and it was first marketed in 2010 as ‘HVDC Plus’ by Siemens [43]. This topology has already been examined and evaluated by both simulation and experimental validation. Also it has already been built for industrial application [30, 31, 43, 46-48]. It is concluded that the MMC topology is suitable for both AC/AC conversion applications and AC/DC application [8, 43, 46]. This topology has been considered widely and different schemes of its control have been introduced using different algorithms [30, 49]. Simulation results and prototype modules

are presented considering both the dynamic performance and its steady state operation [49, 50]. This topology provides characteristics to enable most of the factors required for further HVDC applications such as allowing asynchronous AC networks to be connected and allowing transmission of bulk power over a long distance compared to the HVAC[22]. Therefore, mathematical models have been represented to reduce the calculation time of the simulations [51]. Figure 2.8 illustrates the three-phase MMC converter. It is reported that the overall efficiency of a HVDC station based on the MMC is approximately 99% [52].

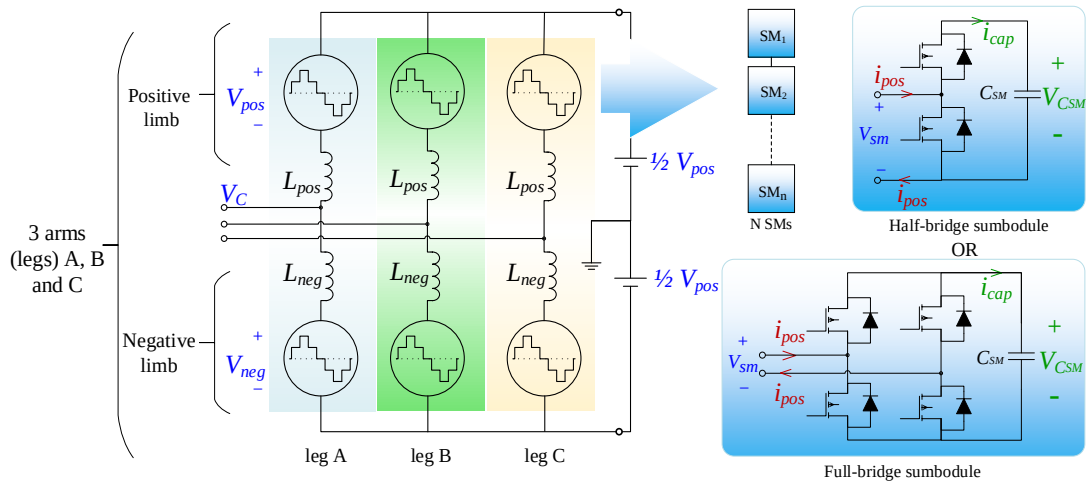


Figure 2.8: HVDC transmission system based on generalized MMC.

The main characteristics of the three-phase MMC topology as shown in Figure 2.8 have already been discussed in the literature such as its scalability, its modular structure and better harmonic profile. For these and other reasons, it is agreed that this topology has the flexibility to design converters for different ranges of voltages and power levels with a variety of applications [9, 43].

As illustrated in Figure 2.8 MMC converter topology consists of three legs, sometimes called arms. Each leg has two limbs, positive and negative limbs which content identical submodules, n SMs per limb. The SM can either be a half-bridge or full-bridge as illustrated in Figure 2.9. The output voltage of the converter can be synthesised by controlling the switching pattern of the SMs in each leg, the upper and lower legs for each phase [53].

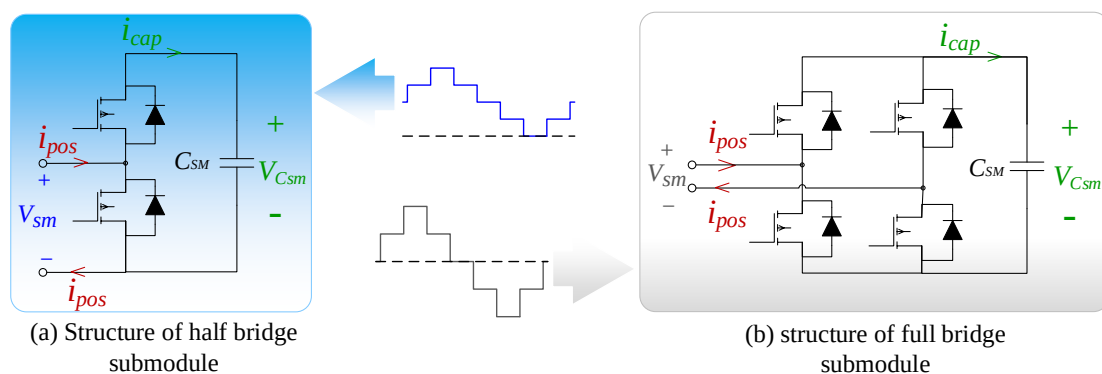


Figure 2.9: Structure of submodules. (a) Half-bridge SM. (b) Full-bridge SM.

2.6.2 Parallel Hybrid Modular Multilevel VSC

Recently, a novel hybrid modular multilevel VSC was presented in the literature [54]. Since it was presented by Alstom Grid, the hybrid modular multilevel VSC, which also known as parallel hybrid modular multilevel VSC (PH-M2L-VSC) [29], has been widely examined by a research group at the University of Nottingham in partnership with Alstom Grid. The hybrid modular multilevel VSC is a compact and high efficient converter being considered for HVDC applications. Its compact structure makes it particularly desirable for offshore applications. This section will present a brief overview of the converter operating principles [12].

Considering Figure 2.10, this topology has three main parts. Firstly, and starting from the AC grid side, the three-phase transformer which in fact is a configured using a single-phase transformer for each phase. The three-phase transformer configuration is based on three single-phase transformers which provides the required phase isolation between the three H-Bridge converters, HB a, HB b and HB c, preventing any current path between the full-HB converters. The authors in [55] also reported that this can be beneficial in terms of the transportation of the transformers and also cancels the 3rd harmonic from the grid side. Secondly, series IGBTs combined to obtain the H-bridge converters for each phase which is connected to cascaded half-bridge cells. These cells are intended to provide the wave-shaping function where the three-phases then are connected in series to obtain the DC link voltage via an inductive filter (DC link filter) as illustrated in Figure 2.10. However, to achieve the desired characteristic operation, a suitable control scheme is required some of which has already been presented in [12].

One of the characteristics of the PH-M2L-VSC is that the series IGBTs operate at the fundamental frequency which reduces the switching losses as soft switching can be

2.6.3 Hybrid Cascaded Modular Multilevel (HCMC)

In the previous discussed topology, PH-M2L-VSC topology, the wave-shaping circuit (WSC) was in parallel. A new topology based on a wave-shaping circuit has been recently presented, also by Alstom Grid [37]. Unlike the PH-M2L-VSC, the wave-shaping circuit is connected in series. In fact, the wave-shaping circuit can be connected either in the AC side or the DC side as discussed later. According to [37], the presented topology provides the ability to clear the DC-side faults without any exchange of power between the AC and the DC sides. This unique feature eliminates the consideration of needing the DC side breakers network.

The basic idea of HCMC topology is the usage of a wave-shaping circuit (WSC) which is connected in the AC side with the series director switches (DS) which are connected in the DC side. Figure 2.11 illustrates the equivalent circuit of the HCMC. As can be seen, the WSC is simply a series of full H-bridge submodules providing the wave shaping to the director switches. The WSCs based on a full-bridge have the ability to suppress the fault current by inserting an inversed polarity voltage in the fault path [44]. The sequence of operation of the director switches operates as full bridge converter modulation. The main features of the HCMC are reported in [37, 57] which include fault ride-through capability, low conversion losses, low harmonic filtering requirements.

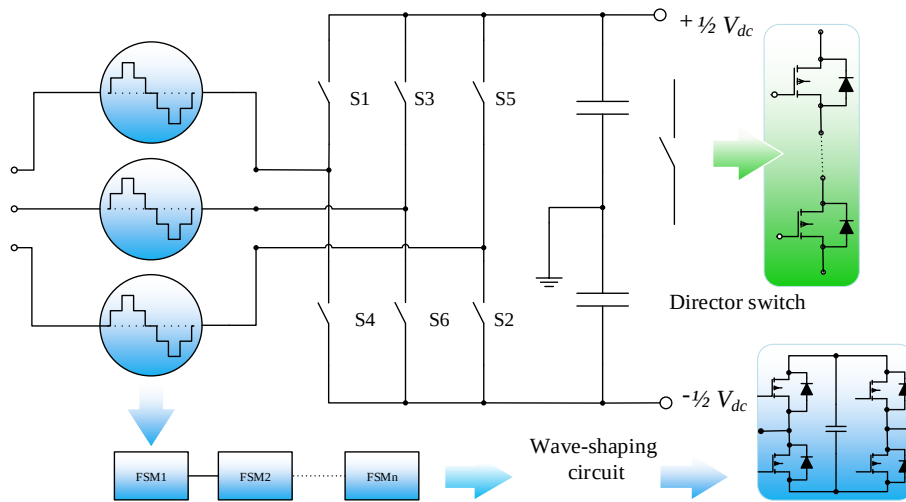


Figure 2.11: The hybrid cascaded modular multilevel VSC topology.

Despite the fact that the hard-switching occurs due to putting the WSC on the AC side, the switches operate at low switching frequency which lead to minimizing the switching losses. In fact, it is reported that the topology offering low conversion losses and low

harmonic filtering requirements. This topology also has the capability of fault ride-through for both AC and DC which means no DC breaker as discussed earlier [37].

2.6.4 Hybrid Alternate-Arm Multilevel Converter (AAC)

The previous presented topologies by Alstom Grid are classified as a series circuit. However, a new parallel circuit topology has been presented by Imperial College, London in cooperation with GE (previously Alstom Grid) [58]. Figure 2.12 illustrates the schematic of the AAC topology. It can be seen that the AAC topology is a hybrid of the generalised MMC with series director switches [44]. This hybrid multilevel VSC is known as an Alternate-Arm Multilevel Converter (AAC) [58]. The alternate arm converter is another modular converter with the capability of handling DC side faults in HVDC systems [35] and [59]. This capability makes this converter more desirable for multi-terminal HVDC systems [60].

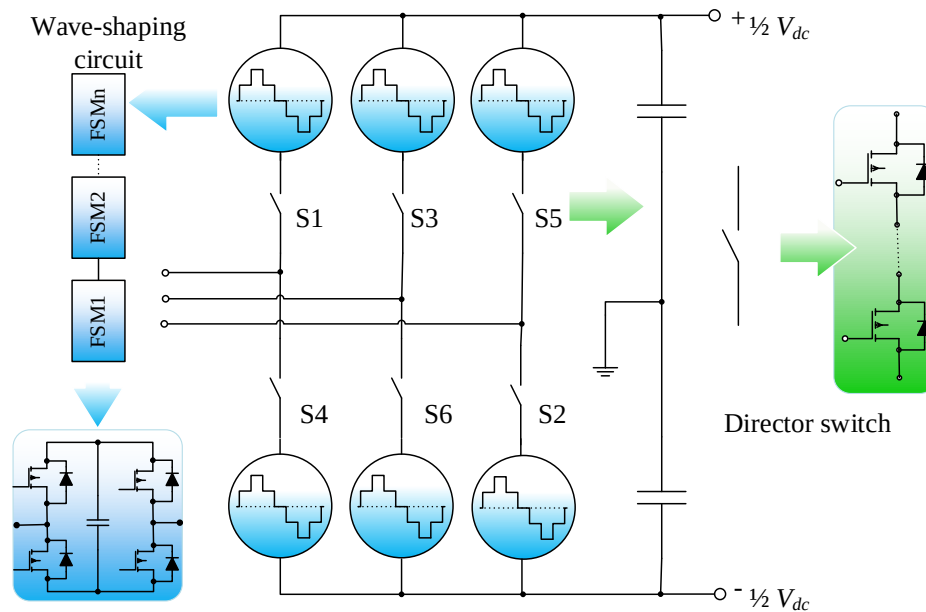


Figure 2.12: The hybrid Alternate-Arm Multilevel Converter.

The AAC has many features some of which are low harmonic content and low switching losses as the topology provides soft-switching operation [58]. Moreover, it is reported that the topology has the ability to control the fault currents on the DC. Previous work has reported that the topology is capable for DC fault ride-through and can be used as STATCOM operation [35, 58]. It is reported that a simulation model of twenty MW AAC has an efficiency of 98.85% [22].

2.7 Conclusion

In this chapter, a literature review on the HVDC transmission was presented. The benefits of HVDC over HVAC transmission were discussed. Based on the converting technology, the components of HVDC transmission technologies was discussed presenting two classic converters, line commutated current and self-commutated voltage source converters, CSCs and VSCs respectively. The conventional multilevel converters based on VSCs were discussed including CHB, NPC, FCC and MMC.

The Concept of modularity first introduced by [8] was discussed in more details considering other alternative well-known topologies based on modular and hybrid structure. The focus started from the classic topology (MMC), and ended with modular hybrid solutions. The modular concept can ensure increasing the number of level up to hundreds and for a power rating and voltage level of gigawatts and hundreds of kilovolts. Therefore, depending on the application, power and voltage level, different multilevel topologies can be selected, starting from the classical FCC, NPC and CHB and ended with modular and hybrid concept, e.g. MMC. For HVDC transmission, the focus of this thesis will be on the improvement of the operation and design of the MMC topology as it is currently the state of the art for HVDC applications. However, similar approaches can be applied also to other modular and hybrid structures.

In Chapter 3 a literature review on the system description and principle operation of the generalised MMC is presented including limbs voltages, currents, power and energy.

Chapter 3

3 The MMC Principle of Operation

3.1 Introduction

In Chapter 2, the concept of modular multilevel converters for HVDC systems was discussed. The generalised modular multilevel converter (MMC) and some other alternative topologies for HVDC systems were discussed. An introduction to the generalised MMC and its control schemes and challenges was presented in the previous chapter.

In the previous chapter, the MMC has been considered as the state of the art for HVDC applications. There are many new HVDC transmission projects, e.g. in Table 2.1, based on half-bridge MMC which highlights how successful the MMC is today. This growth demand helps to increase the development of the MMC, improving its design and operation. This thesis participates in this development providing an extensive study investigating using 3rd harmonic injection to improve the operation and design of the generalised MMC.

Therefore, in this chapter, the principle of operation of the generalised MMC will be presented including the system description and the theoretical limb voltages, currents and the instantaneous power and energy. The study in this thesis is based on some assumptions which will be presented in this chapter. The operating principles of an MMC are firstly introduced and then the mathematical model of the MMC is resented.

3.2 System Description

The MMC circuit considered in this study is illustrated in Figure 3.1 where a half bridge submodule configuration is adopted for populating the converter legs. This is a standard configuration of the converter that utilises fewer semiconductor devices but does not support DC fault ride through which can be achieved with the full bridge implementation [61]. The figure also illustrates the concept of the synthesised positive and negative limb voltages and currents of a phase, using phase A for the illustration.

The MMC consists of three legs (arms) which are responsible for synthesising the required output voltage. Each leg consists of two limbs, positive and negative, where each limb includes identical submodules (SMs). In this study, each SM consists of two IGBT switches and one DC capacitor (C_{SM}). The capacitor size must be sufficient to prevent voltage distortion in the output voltage waveform. This means that the required inserted voltage must always be less than the available voltage in a limb, (i.e. the sum of inserted capacitors' voltages must be able to provide the required reference converter voltage). The limitation of the MMC terminal voltage respective to the number of the SMs per limb is shown by Equation (3.1) which can be ensured by pre-charging the SM capacitors with a nominal voltage given by (3.2) [9].

$$0 \leq v_{pos}(t) \leq N_{SM} \cdot v_{cap}(t) \quad (3.1)$$

$$V_{cap} = \frac{V_{dc}}{N_{SM}} \quad (3.2)$$

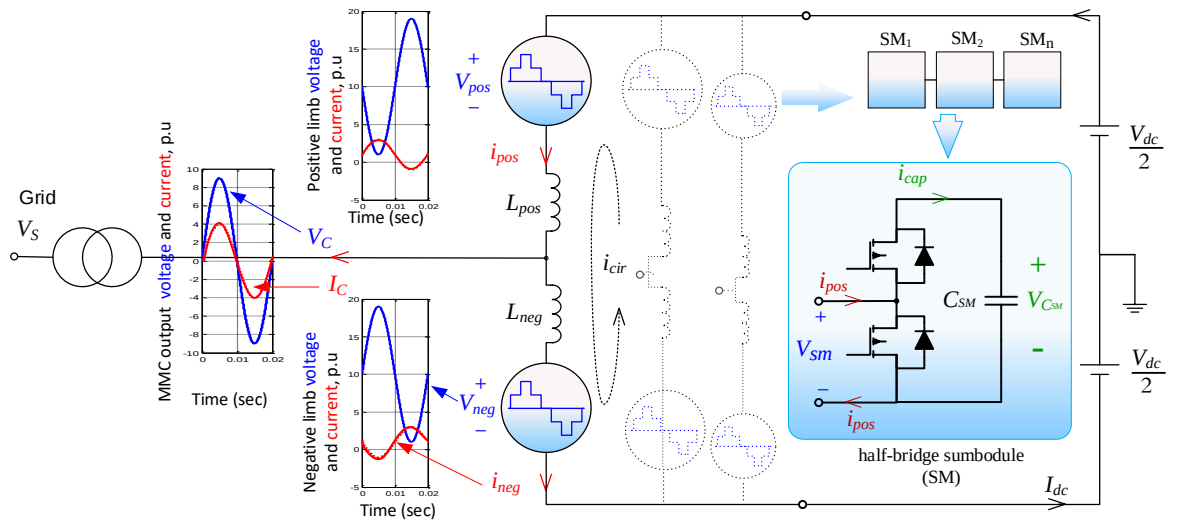


Figure 3.1: The generalised Modular Multilevel Converter (MMC).

The inductance in each limb works as a buffer inductor and is inserted in order to reduce the rate of current rise due to the possible voltage difference between two converter limbs. The buffer inductor can also limit fault current. Unlike the HVDC Light, the phase inductor not required for the MMC. However, the buffer inductor needs to be considered when considering the voltage control.

The legs of the MMC intended to synthesis the required AC voltage at the MMC terminals to meet the demand of active and reactive power. This can be done by switching ON or OFF the SMs of each limb where the synthesised voltages by the

positive and negative limbs is illustrated by Figure 3.1 and Figure 3.2.b. The relationship of the one leg voltages (two limbs, positive and negative) and the MMC terminal voltage is illustrated by Figure 3.2. To ensure that, the SMs capacitors are inserted in or bypassed while charging or discharging to keep their voltage at the permissible level. In fact, the SM capacitors act as a DC source. Therefore, for the half-bridge based MMC, the SM has two output voltages based on its states, a positive voltage when it is inserted (ON) and a zero voltage when it is bypassed (OFF). During ON and OFF, each SM capacitor will discharge or charge to maintain their voltage at the permissible level based on the direction of the limb current. The DC-link voltage, V_{dc} , is adjusted by switching a number of SMs in the positive and negative limbs where the total number of the required switched ON or OFF SMs for one leg (positive and negative limbs) has always to be equal to the number of the SMs per one limb, N_{SM} . This means that the SM capacitor voltage is given by Equation (3.2). In a similar manner, the grid voltage, V_s , can be adjusted to a desired value. To achieve that, the MMC control is required which consider in Chapter 7 while the relationship of the voltages, currents, power and energy of the MMC terminals and its limbs are presented in next sections.

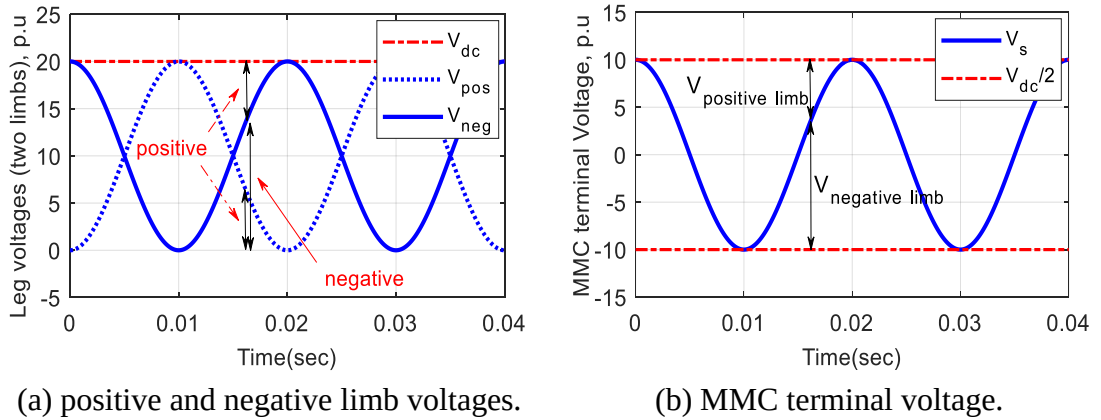


Figure 3.2 : Relationship of leg voltages (i.e. positive and negative limbs voltages) and terminal voltage.

3.3 Assumptions

The detailed converter phase arm and one phase system shown in Figure 3.1 is used to illustrate the concept of the synthesised voltage and operating mechanism of the MMC where the AC system voltage and current are denoted by V_C and I_C , respectively, while the limb voltage and currents are denoted by v_{pos} and i_{pos} , respectively. DC voltage and current are denoted by V_{dc} and I_{dc} , respectively.

In the ensuing analysis, the following assumptions have been made to simplify the complexity of the analysis without compromising the accuracy of the investigation:

1. The MMC is assumed to have enough available voltage to operate without over modulation or any other voltage profile reduction. To achieve this, it is assumed that the MMC operates with ideal voltage sources or with large local storage capacitors with a capacitor voltage ripple compensation technique such as an effective SM rotating algorithm [62].
2. In this investigation, the MMC circuit considered is based on half-bridge submodules.
3. Losses in the converter are neglected. This is reasonable since typical large MMC converters have an efficiency of $>98\%$ [63].
4. It is assumed that the converter synthesises pure sinusoidal voltage and current waveforms as would be achieved with an infinite number of converter submodules.
5. In this investigation, to consider the influence of the power factor, it is assumed that the converter is operating with rated apparent power $|S|$ with variation of the active (P) and reactive power components (Q) transferred from/to the network. Therefore, P and Q are adjusted accordingly to achieve the desired operation point as illustrated by Figure 3.3. It can be seen that the power convention used here is the Generator Power Convention as described below.

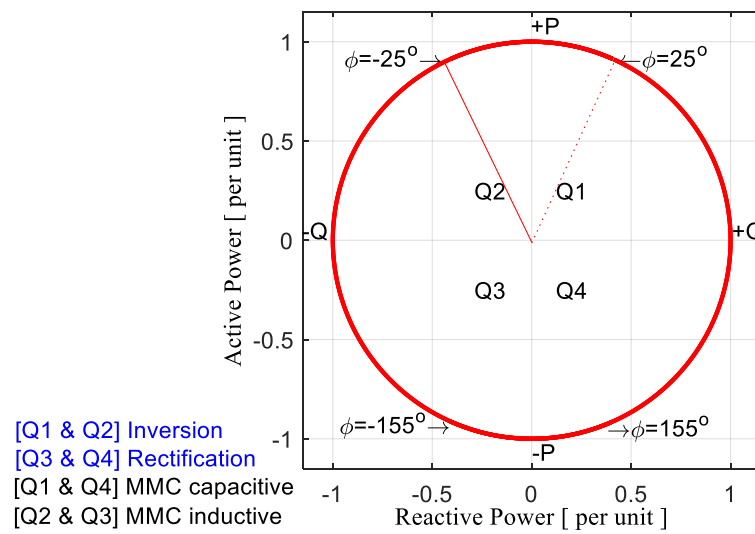


Figure 3.3: PQ curve of an ideal MMC and the Generator Power convention.

The power convention used in this thesis is presented in Appendix A. The PQ capability curve of an ideal MMC is illustrated where positive P and Q are taken to mean that the MMC is injecting P and Q into the grid. Conversely, negative P and Q means the grid transferring P and Q to the MMC. Moreover, positive P means the MMC operates as an inverter, whilst negative P means the MMC operates as a rectifier. Positive Q means the MMC acts like a capacitor injecting Q into the AC system, whilst negative Q means the MMC acts like an inductor absorbing Q from the grid.

6. It is assumed that the equivalent impedance of each limb due to the SMs capacitors is greater than that of the buffer inductors, $X_{C_{eq.SM}} \gg X_{L_{limb}}$. Thus; $C_{eq.SM} \ll \frac{1}{\omega_o^2 L_{limb}}$, where $C_{eq.SM}$ is the equivalent capacitance of the limb SM capacitors and L_{limb} is the limb inductor. ω_o is the system nominal frequency given in rad/sec.

7. It is assumed that the AC supply and the converter are symmetrical.

3.4 MMC Theoretical Voltage, Current, Power and Energy

This section presents the theoretical limb voltages, currents and also discusses the instantaneous power and energy. Subsections 3.4.1 and 3.4.2 are an introduction of the MMC theoretical voltages, currents, instantaneous power and energy variation without considering any possible active harmonic injection concepts. However, the stored energy in a limb and its available voltage concept is introduced. MMC limb peak-peak energy (energy deviation) is presented in Section 3.5. This study can be expanded to account for additional harmonics including 3rd harmonic and 2nd harmonic components which are discussed in the analysis in the next Chapter.

3.4.1 MMC Theoretical Voltages and Currents

The converter terminal voltage of phase A of the MMC can be defined by (3.3).

$$v_{Ca} = \hat{V}_{C1} \cos(\omega t) \quad (3.3)$$

For a symmetrical AC system, the voltages and currents amplitude are equal with 120° displacement between the 3 phases, therefore, the voltages and currents amplitude is denoted by $\hat{V}_C$ and $\hat{I}_C$ since $\hat{V}_{Ca} = \hat{V}_{Cb} = \hat{V}_{Cc} = \hat{V}_C$ and $\hat{I}_{Ca} = \hat{I}_{Cb} = \hat{I}_{Cc} = \hat{I}_C$. The fundamental components are denoted by $\hat{V}_{C1}$ and $\hat{I}_{C1}$. The positive and negative voltages are related to the DC and terminal voltages and are given by (3.4) and (3.5), respectively.

The currents, in the positive and negative limbs due to the DC and AC currents can be expressed by (3.4) and (3.5), respectively. The expressions given are for phase A while the same principle can be applied for the other phases, B and C.

Due to the interaction between the limb AC current and voltage on the submodule capacitors, the two limbs in a phase experience a circulating current between them denoted by i_{cir} [30] which can be seen by the capacitor voltage ripple, where the frequency of the circulating current is twice the output frequency.

$$v_{pos.a} = \frac{V_{dc}}{2} - v_{Ca} - v_{cir.a} \quad (3.4)$$

$$v_{neg.a} = \frac{V_{dc}}{2} + v_{Ca} - v_{cir.a} \quad (3.5)$$

$$i_{pos.a} = \frac{I_{dc}}{3} + \frac{i_{Ca}}{2} + i_{cir.a} \quad (3.6)$$

$$i_{neg.a} = \frac{I_{dc}}{3} - \frac{i_{Ca}}{2} + i_{cir.a} \quad (3.7)$$

where $v_{pos.a}$ and $v_{neg.a}$ denote the synthesised voltage of positive and negative limbs respectively. $i_{pos.a}$ and $i_{neg.a}$ denote the synthesised current of the positive and negative limbs respectively. V_{dc} and I_{dc} is the DC link voltage and DC current respectively. $v_{cir.a}$ and $i_{cir.a}$ are voltage components with their corresponding circulating currents. Note that $v_{cir.a}$ may be introduced intentionally (as discussed later) but it can also arise unintentionally due to ripple on the capacitor voltages causing errors in the desired synthesised voltage.

As half-bridge submodules are unable to synthesise a negative voltage, the synthesised AC voltage of the half-bridge based MMC is always restricted by (3.8) [9]. Where N_{SM} is the SM number for each limb while $|\hat{V}_{CSM}|$ is the capacitor voltage for one SM.

$$0 \leq |\hat{V}_{pos}| \leq N_{SM} \cdot |\hat{V}_{CSM}| \quad (3.8)$$

The circulating current is normally prevented due to an appropriate modulating algorithm which is able to mitigate the effects of the capacitor voltage ripple on the synthesised voltage. However, the circulating current can also be controlled using an appropriate controller such as a proportional-resonant controller (PR) to zero [64].

Considering the amplitude, phase shift and time variance, the periodic positive and negative limb voltage and current of equations (3.4) to (3.7) can be given by (3.9) and (3.10), respectively where $v_{cir} = 0$ and $i_{cir} = 0$. In this analysis, the phase voltage is considered the reference with a phase angle reference of zero. Therefore, the current phase shift (lag) with respect to the phase voltage is denoted by the power factor angle, ϕ .

$$v_{pos.a}(t) = \frac{V_{dc}}{2} - \hat{V}_{c1} \cos(\omega_0 t) \quad (3.9)$$

$$i_{pos.a}(t) = \frac{I_{dc}}{3} + \frac{\hat{I}_{c1}}{2} \cos(\omega_0 t - \phi) \quad (3.10)$$

Similarly, the negative limb voltage and current are given by (3.11) and (3.12)

$$v_{neg.a}(t) = \frac{V_{dc}}{2} + \hat{V}_{c1} \cos(\omega_0 t) \quad (3.11)$$

$$i_{neg.a}(t) = \frac{I_{dc}}{3} - \frac{\hat{I}_{c1}}{2} \cos(\omega_0 t - \phi) \quad (3.12)$$

The ideal available voltage within the submodules, however, equals the DC voltage when it is assumed that each submodule has a large capacitor so the effect of ripple from the energy storage components can be neglected. The waveforms of positive and negative limb voltages and currents for an ideal MMC system in Table 3.1 are given in Figure 3.4 for an operating point of $+P=1.0$ p.u and $Q=0$. The P and Q are defined based on the power convention discussed before and presented in Appendix A.

Table 3.1: Ideal modular multilevel converter specifications.

Rated Power	S	20.0	MVA
DC link voltage	V_{dc}	20.0	kV
Grid AC voltage (rms)	V_s	11.0	kV
Modulation index	m_i	1.0	
Number of submodules	N_{SM}	10	SMs
SM capacitor nominal voltage	V_{CSM}	2000	V
Capacitor voltage ripple	x	$\approx 0.0\%$	
Displacement power factor	ϕ	$-90^\circ \leq \phi \leq 90^\circ$	degrees

The relationship between the DC-link voltage denoted by V_{dc} and the amplitude of the AC voltage, V_c , on the MMC terminals is denoted by $\hat{V}_c$, is given by the modulation index (m_i) where V_s in Table 3.1 denotes the grid AC voltage as shown by Figure 3.1;

$$\hat{V}_c = m_i \frac{V_{dc}}{2} \quad (3.13)$$

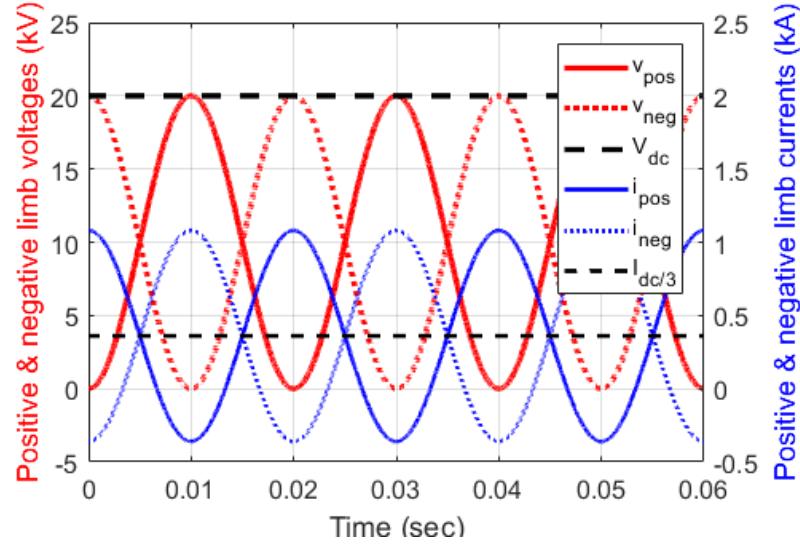


Figure 3.4: Positive and negative limb voltages and currents of an ideal Modular Multilevel Converter (MMC).

3.4.2 Instantaneous Power and Energy

The positive limb instantaneous power, $v_{pos.a}i_{pos.a}$, is expressed in (3.14). Substituting (3.9) and (3.10) into (3.14) and using the trigonometry identity;

$$\cos a \cos b = \frac{1}{2} \{ \cos(a - b) + \cos(a + b) \}$$

The instantaneous limb power is given by (3.15).

$$p_{pos}(t) = v_{pos.a}(t)i_{pos.a}(t) \quad (3.14)$$

$$p_{pos.a}(t) = \left[\begin{aligned} & \frac{V_{dc}I_{dc}}{6} - \frac{\hat{V}_{c1}\hat{I}_{c1}}{4}\cos(\phi) \\ & + \left(\frac{V_{dc}\hat{I}_{c1}}{4}\cos(\omega_0 t - \phi) - \frac{\hat{V}_{c1}I_{dc}}{3}\cos(\omega_0 t) \right) \\ & - \frac{\hat{V}_{c1}\hat{I}_{c1}}{4}\cos(2\omega_0 t - \phi) \end{aligned} \right] \quad (3.15)$$

Similarly, the instantaneous power of negative limb is given by;

$$p_{neg.a}(t) = v_{neg.a}(t)i_{neg.a}(t) \quad (3.16)$$

$$p_{neg.a}(t) = \left[\begin{array}{c} \frac{V_{dc}I_{dc}}{6} - \frac{\hat{V}_{c1}\hat{I}_{c1}}{4}\cos(\emptyset) \\ - \left(\frac{V_{dc}\hat{I}_{c1}}{4}\cos(\omega_0 t - \emptyset) - \frac{\hat{V}_{c1}I_{dc}}{3}\cos(\omega_0 t) \right) \\ - \frac{\hat{V}_{c1}\hat{I}_{c1}}{4}\cos(2\omega_0 t - \emptyset) \end{array} \right] \quad (3.17)$$

The expression for the average power is given by integrating the instantaneous power in equations (3.15) and (3.17) for one periodic cycle. The terms produced by the product of the voltage and current with the same frequency will remain while the others are eliminated. Hence the average powers are given by (3.18) and (3.19).

$$\bar{P}_{pos.a} = \frac{V_{dc}I_{dc}}{6} - \frac{\hat{V}_{c1}\hat{I}_{c1}}{4}\cos(\emptyset) \quad (3.18)$$

$$\bar{P}_{neg.a} = \frac{V_{dc}I_{dc}}{6} - \frac{\hat{V}_{c1}\hat{I}_{c1}}{4}\cos(\emptyset) \quad (3.19)$$

To obtain capacitor charge balancing in the MMC it is required that the average power of the limbs must be zero during one periodic cycle. Therefore, equations (3.18) and (3.19) equal zero yielding (3.20). This is considered a particularly important relationship because it establishes the point where there can be power balance between the AC and DC sides and therefore sustainable converter operation can be achieved [44, 45]. Therefore,

$$V_{dc}I_{dc} = \frac{3}{2}\hat{V}_{c1}\hat{I}_{c1}\cos(\emptyset) \quad (3.20)$$

$$P_{dc} = P_{ac} \quad (3.21)$$

Substituting (3.13) into (3.20), we obtain (3.22).

$$I_{dc} = \frac{3}{4}m_i\hat{I}_{c1}\cos(\emptyset) \quad (3.22)$$

From (3.9),(3.11) and (3.13) we can obtain:

$$v_{pos.a}(t) = \frac{V_{dc}}{2} [1 - m_i \cos(\omega_0 t)] \quad (3.23)$$

$$v_{neg.a}(t) = \frac{V_{dc}}{2} [1 + m_i \cos(\omega_0 t)] \quad (3.24)$$

Using (3.10) and (3.22), we can obtain:

$$i_{pos.a}(t) = \frac{\hat{I}_{C1}}{2} \left[\frac{1}{2} m_i \cos(\phi) + \cos(\omega_0 t - \phi) \right] \quad (3.25)$$

Similarly, the positive and negative limb instantaneous power in (3.15) and (3.17) can be rearranged and expressed by the modulation index and rated power $|S|$. The latter is given by $\frac{3}{2} |\hat{V}_{C1} \hat{I}_{C1}|$.

$$p_{pos.a}(t) = \frac{|S|}{6} \left[\left(\frac{2}{m_i} \cos(\omega_0 t - \phi) - m_i \cos(\phi) \cos(\omega_0 t) \right) - \cos(2\omega_0 t - \phi) \right] \quad (3.26)$$

$$p_{neg.a}(t) = \frac{|S|}{6} \left[- \left(\frac{2}{m_i} \cos(\omega_0 t - \phi) - m_i \cos(\phi) \cos(\omega_0 t) \right) - \cos(2\omega_0 t - \phi) \right] \quad (3.27)$$

The energy variations of the positive and negative limbs are given by integrating the instantaneous power. That is

$$E(t) = \int p(t) \cdot dt \quad (3.28)$$

$$e_{pos.a}(t) = \frac{|S|}{6\omega_0} \left[\left(\frac{2}{m_i} \sin(\omega_0 t - \phi) - m_i \cos(\phi) \sin(\omega_0 t) \right) - \frac{1}{2} \sin(2\omega_0 t - \phi) \right] \quad (3.29)$$

$$e_{neg.a}(t) = \frac{|S|}{6\omega_0} \left[- \left(\frac{2}{m_i} \sin(\omega_0 t - \phi) - m_i \cos(\phi) \sin(\omega_0 t) \right) - \frac{1}{2} \sin(2\omega_0 t - \phi) \right] \quad (3.30)$$

From equations (3.29) and (3.30) it can be seen that the energy variations within the converter limbs increase with increasing power $|S|$ and vice versa for given power factor

and modulation index. Considering the change of power factor angle, all the three terms of (3.29) and (3.30) are affected. This means that not only will the energy peak be affected by the power factor but also the time at which the maximum and minimum value occurs. This is particularly important as the shape of the energy variation has a direct impact on the available voltage for each individual SM and for the whole limb.

From equations (3.29) and (3.30) it can be seen that the energy variations within the converter limbs do not depend on the capacitor sizes denoted by C_{SM} . It is assumed that the energy storage in each submodule is the same using a balancing method such as an effective sorting algorithm [65]. So, the nominal energy stored in one SM capacitor is given by (3.31) while the total energy stored in one converter limb, e.g. negative limb, is given by (3.32) where $V_{C_{SM}.n}$ and N_{SM} are the nominal voltage of one SM capacitor and number of SMs respectively. Therefore, the available voltage across SM capacitors is the same. Thus;

$$E_{SM.n} = \frac{1}{2} C_{SM} V_{C_{SM}.n}^2 \quad (3.31)$$

$$E_{pos.n} = E_{neg.n} = \sum_{m=1}^{N_{SM}} E_{SM.n,m} = N_{SM} E_{SM.n} \quad (3.32)$$

$$E_{pos.n} = E_{neg.n} = \frac{1}{2} C_{SM} N_{SM} V_{C_{SM}.n}^2 \quad (3.33)$$

The periodic energy variation of one SM capacitor $e_{SM}(t)$ can also be expressed by knowing the capacitor size and its voltage variation $v_{C_{SM}}(t)$. The total energy variation for one limb, e.g. positive limb in our case, expressed by (3.29) can also be re-expressed using the size of capacitor:

$$e_{SM}(t) = \frac{1}{2} C_{SM} v_{C_{SM}}^2(t) - E_{SM.ave} \quad (3.34)$$

$$e(t) = \frac{1}{2} C_{eq} \Sigma v_{C_{SM}}^2(t) - E_{ave} \quad (3.35)$$

where $E_{SM.ave}$ and E_{ave} is the average energy stored in one SM capacitor and one limb, respectively. C_{eq} is the equivalent capacitor of the serially connected SMs. This is equal to the capacitance of one SM divided by the number of submodules. The voltage

variation of each SM capacitor and one limb is denoted by $v_{C_{SM}}(t)$ and $\Sigma v_{C_{SM}}(t)$ respectively. It should be noted that the AC voltage variation has a ripple and an average value. The average value is represented by $V_{C_{SM}}$ for one SM capacitor and for one limb it equals the dc-link voltage V_{dc} .

$$\Sigma v_{C_{SM}}(t) = \sum_{m=1}^{N_{SM}} v_{C_{SM},m}(t) \quad (3.36)$$

The energy variation expressed by (3.35) can be rearranged. Thus

$$e(t) = \frac{1}{2} \frac{C_{SM}}{N_{SM}} \Sigma v_{C_{SM}}^2(t) - E_{ave} \quad (3.37)$$

or

$$e(t) = \frac{1}{2} C_{SM} N_{SM} v_{C_{SM}}^2(t) - E_{ave} \quad (3.38)$$

And

$$E_{ave} = \frac{1}{2} C_{SM} N_{SM} V_{C_{SM}}^2 \quad (3.39)$$

$$E_{ave} = \frac{1}{2} C_{SM} \frac{V_{dc}^2}{N_{SM}} \quad (3.40)$$

Therefore, (3.36) becomes;

$$\Sigma v_{C_{SM}}(t) = N_{SM} v_{C_{SM}}(t) \quad (3.41)$$

From the positive and negative energy variations represented by (3.29) and (3.30), it can be seen that the energy variation depends on the operation point of the MMC expressed by its power |S|, power factor and the modulation index. Therefore, whatever the SM capacitor size, the behaviour of the energy variation will not be affected. From equations (3.37) or (3.38) the available voltage from each SM capacitor or limb can be derived. This is expressed by (3.42). It can be seen that voltage of each SM is affected by the energy variation not vice versa.

$$\Sigma v_{c_{SM}}(t) = \sqrt{\frac{2N_{SM}}{C_{SM}}(e(t) + E_{ave})} \quad (3.42)$$

Equation (3.42) is particularly important as it describes the behaviour of the available voltage as a function of energy variation, total nominal or average energy stored in the SM capacitors, size of capacitor per SM and also the number of submodules per limb. The energy variation is described by (3.29) and (3.30) for positive and negative limbs respectively. There is no relationship between the energy variation and the SM's design except that the SMs should be designed to have the ability to handle the energy variation within the individual SM and the entire limb. Whatever the size of the SM capacitors, the energy variation remains the same depending on the PQ operating point. However, the available voltage expressed by (3.42) is a function of the SM capacitor size, number of SMs and also the energy variation.

For instance, if the N_{SM} is increased, the size of capacitor needs to be reduced by the same factor to achieve similar results for the available voltage waveforms and keep average energy constant. Considering increasing N_{SM} while C_{SM} is kept constant, the available voltage variation is increased while the average energy is kept constant according to Equation (3.42). Increasing N_{SM} leads to a reduction in the average energy while energy variation is never affected by size or number of SMs.

3.5 MMC Peak-to-peak Energy (Energy Deviation)

A review of the relationships between limb voltages, currents, instantaneous power and energy variation for standard MMC operation has been presented. Now maximum and minimum energy analyses are considered together with a review of the peak-to-peak energy (energy deviation) concept for fundamental mode operation of the MMC.

3.5.1 MMC Energy Deviation Analysis

To determine the energy deviation, the local maximum and minimum values of energy variation $e_{pos}(t)$ in equation (3.29) are obtained as illustrated in Figure 3.5. A review of the technique used to determine the maximum and minimum energy and their location within a cycle is presented. By solving equation (3.43) defining the turning points, the local maximum and minimum values can be obtained. Energy deviation can be determined from the difference between the maximum and minimum values.

$$\frac{de_{pos.a}(t)}{dt} = p_{pos.a}(t) = 0 \quad (3.43)$$

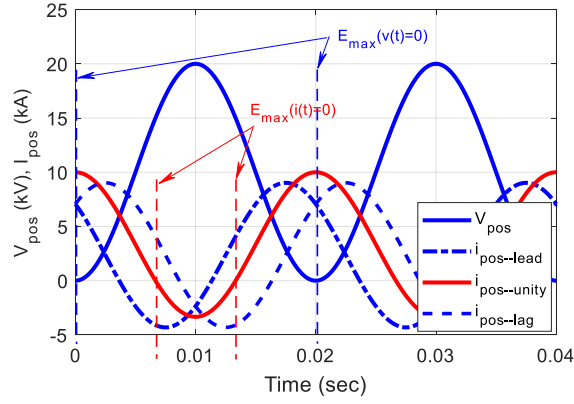
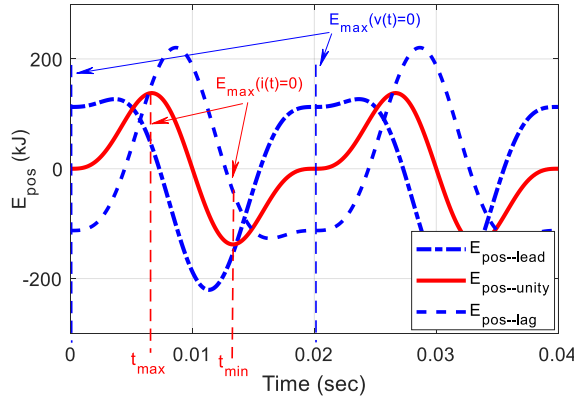

 (a) Voltage and currents when roots $v(t) = 0$ and $i(t) = 0$

 (b) Energy roots when $v(t) = 0$ and $i(t) = 0$

 Figure 3.5: E_{max} and E_{min} relationships with limb voltage and current for leading, unity and lagging power factor.

Solving (3.43) analytically cannot be achieved easily as the instantaneous power expression in equation (3.26) is quite complicated. This becomes more complicated as more harmonic components are included and therefore a numerical method could be used. However, it is observed that the maximum and minimum energy occurs when the power equals zero. Of course power equals zero when either periodic voltage and/or current equal zero as shown in Figure 3.5. Therefore, the algebraic solution becomes more convenient and turning points then can be easily obtained.

$$p_{pos.a}(t) = \begin{cases} 0, & v_{pos.a}(t) = 0 \\ 0, & \text{or } i_{pos.a}(t) = 0 \\ 0, & \text{or } v_{pos.a}(t) \text{ and } i_{pos.a}(t) = 0 \end{cases} \quad (3.44)$$

Using equations , the periodic positive limb voltage and currents are expressed by (3.23) and (3.25) respectively. Applying the first condition in (3.44) with the periodic voltages of the positive limb there the phase angle of the voltage is zero, yields:

$$v_{pos.a}(t) = 0 \quad (3.45)$$

$$t|_{v_{pos.a}(t)=0} = \frac{\cos^{-1}\left(\frac{1}{m_i}\right)}{\omega_o} \quad (3.46)$$

Equation (3.46) gives the time at which the instantaneous energy becomes zero, however, for a half-bridge based MMC this cannot occur as the modulation index is always less than unity and therefore the minimum synthesised voltage cannot hit the lower limit of zero. Considering the current equation, there are two possible roots, t_{max} and t_{min} which can be given as below.

$$i_{pos.a}(t) = 0 \quad (3.47)$$

$$t_{max} = \left(\pi + \phi - \cos^{-1}\left(\frac{m_i}{2} \cos(\phi)\right)\right)/\omega_o \quad (3.48)$$

$$t_{min} = \left(\pi + \phi + \cos^{-1}\left(\frac{m_i}{2} \cos(\phi)\right)\right)/\omega_o \quad (3.49)$$

Considering Figure 3.5, the t_{max} always occurs in the first quarter of the cycle, $t_{max} < \frac{1}{4}T$, for all the operating point, leading, lagging or unity power factor which indicated by the minus sign. On the other hand, t_{min} occurs always in the second quarter of the period T , $\frac{1}{4}T < t_{min} < \frac{2}{4}T$ which indicated by the plus sign. The According to (3.48) and (3.49), local maximum and minimum roots depend on the power factor and modulation index. Substituting equations (3.48) and (3.49) in the limb energy variation in (3.29), e.g. the positive limb, the maximum and minimum values of energy $e_{pos.a}(t)$ are presented by equations (3.50) and (3.51) respectively.

$$E_{max} = \frac{|S|}{6\omega_o} \left[\left(\frac{2}{m_i} \sin(\omega_o t_{max} - \phi) - m_i \cos(\phi) \sin(\omega_o t_{max}) \right) - \frac{1}{2} \sin(2\omega_o t_{max} - \phi) \right] \quad (3.50)$$

$$E_{min} = \frac{|S|}{6\omega_o} \left[\left(\frac{2}{m_i} \sin(\omega_o t_{min} - \phi) - m_i \cos(\phi) \sin(\omega_o t_{min}) \right) - \frac{1}{2} \sin(2\omega_o t_{min} - \phi) \right] \quad (3.51)$$

Maximum energy deviation depends on the variation of the load (active and reactive power). Considering (3.50), obviously it can be seen that the dominant term is the first one where the energy variation is inversely proportional to the modulation index. $E_{max} \propto \frac{1}{m_i}$. Using equations (3.50) and (3.51) with variations of power factor, the energy deviation is denoted by ΔE and expressed below in (3.52).

$$\Delta E = E_{max} - E_{min} \quad (3.52)$$

Considering (3.50) and the range of P and Q required, the maximum energy deviation (ΔE_{max}) that the converter can experience can be determined.

3.5.2 Simulation Results and Discussion

For a given MMC with the parameters in Table 3.2 and with variation of P and Q results are obtained based on the mathematical analysis above. The given parameters are for a practical MMC where voltage limits are neglected (reactive power limits) to examine the energy deviation with rated capacity and variation of P and Q. Compared to Table 3.1 where an ideal MMC are considered.

Maximum and minimum energy varies with the variation of power factor. Therefore, energy deviation also changes in the same manner. Figure 3.6 illustrates the variation of the operating point from -Q to +Q whilst P is always positive ($-90^\circ \leq \phi \leq 90^\circ$). However, both $|S|$ and m_i are kept constant. It has been observed (not shown) that the same change is experienced with the inversion and rectification modes of operation, so both modes are represented in Figure 3.6.

Table 3.2: Modular multilevel converter specifications.

Rated Power	S	20.0	MVA
DC link voltage	V_{dc}	20.0	kV
Grid voltage	V_S	11.0	kV
Modulation index	m_i	0.8	
Number of submodules	N_{SM}	10	SMs
Capacitor voltage ripple	$\Delta V_{C_{SM}}$	10%	
Power factor	ϕ	$-90^\circ \leq \phi \leq 90^\circ$	degrees

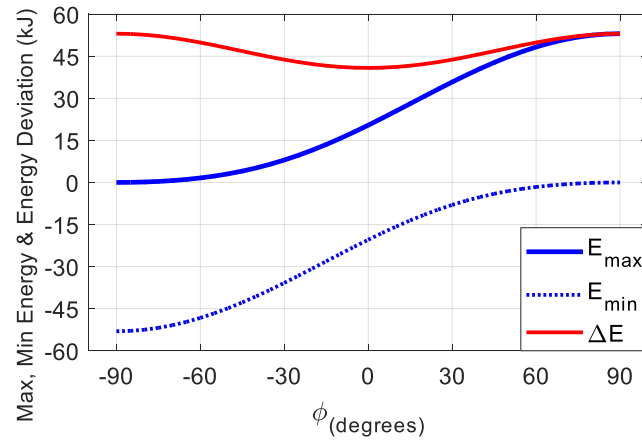


Figure 3.6: Minimum, maximum and energy deviation relationship with power factor.

With regards to energy deviation and power factor variation, Figure 3.7 represents the impact of the power factor variation on energy deviation. Figure 3.7 shows a zoomed version of the red trace in Figure 3.6. It can be seen that the deviation is a minimum for purely resistive operation and is a maximum for purely inductive or capacitive operation.

The previous waveforms and simulation results are based on a constant modulation index. However, if the modulation index is varied, the behaviour of the energy deviation changes as seen in Figure 3.8. Figure 3.8 shows the impact of the power factor and modulation index variation on energy deviation. The presented simulation results show the impact of variation of modulation index, m_i , (of course the variation of the MMC AC current) on the energy deviation. It is assumed that the secondary windings of the converter transformer have a variable turns ratio to meet the requirements of the m_i and PQ variation, whilst maintaining constant $|S|$.

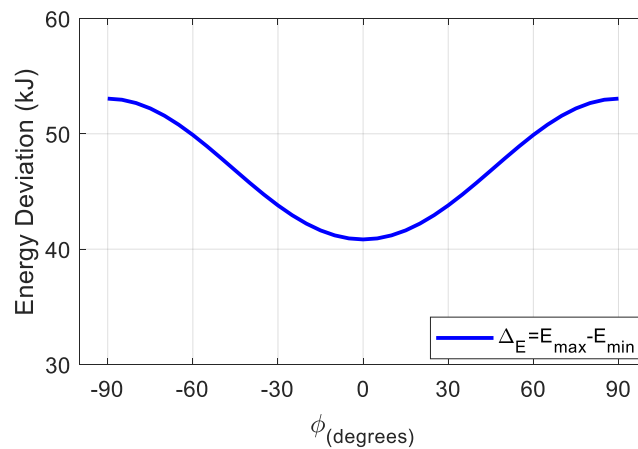
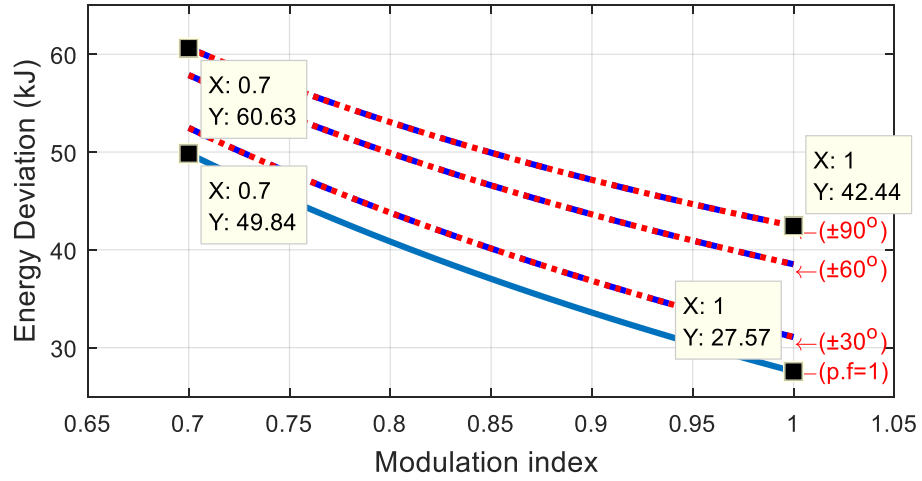
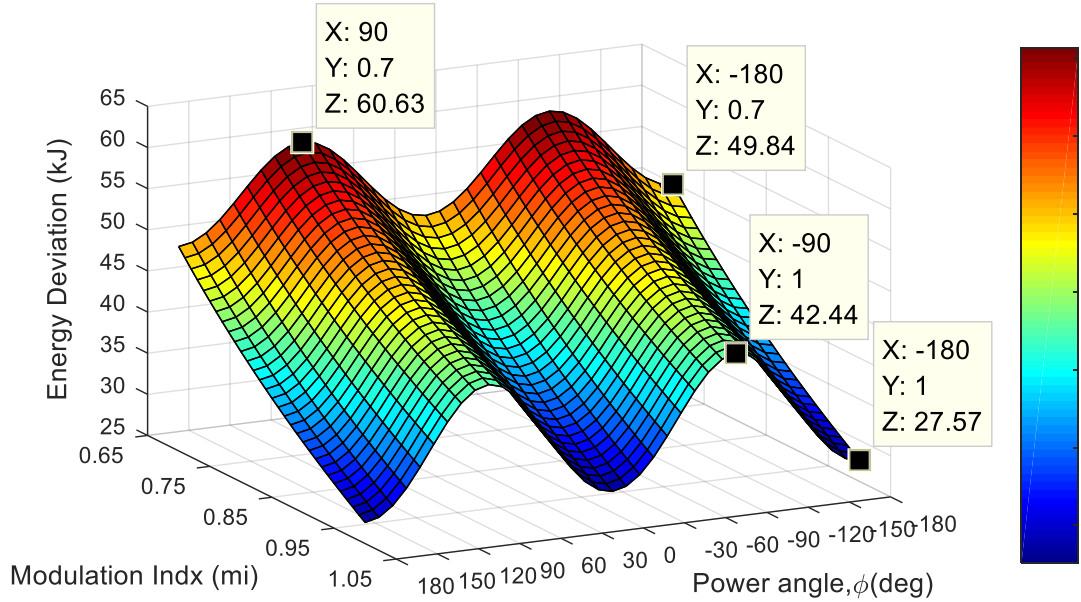


Figure 3.7: V curve of maximum and minimum energy deviation with power factor ($-90^\circ \leq \phi \leq 90^\circ$).



(a) Energy deviation with modulation index and power factor ($-90^\circ \leq \phi \leq 90^\circ$).



(b) M curve of Energy deviation, modulation index and p.f ($-180^\circ \leq \phi \leq 180^\circ$).

Figure 3.8: Energy deviation and power factor relationship with modulation index.

The m_i envelope gives the ability to determine the energy deviation envelope at any operating point. This can help understand the MMC design and operation requirements at any particular operating point. However, for a practical MMC, the MMC operates such that the modulation index varies depends on PQ with a fixed transformer turns ratio while the AC current is constant for a constant $|S|$. It can be seen that, as the modulation index increases (reducing the AC current), the energy deviation is reduced.

3.6 Conclusion

In this chapter, the system description and principle operation of the generalised MMC was discussed. The MMC operates so that each SM can be switched to either 0 V or to

the voltage $V_{c_{SM}}$ where the SM capacitors act like a DC source. The DC-link voltage, V_{dc} , is adjusted by switching a number of SMs in the positive and negative limbs. In a similar manner, the AC voltage can be adjusted to a desired value especially when there are 100s of SMs per limb.

The theoretical voltage and current of the MMC limbs were presented. Using the limb voltage and currents, the instantaneous power and energy was discussed where the focus was on the energy deviation which will be further investigated in the next chapters considering harmonic injection including, 3rd harmonic voltage and current and 2nd harmonic current injection.

In chapter 4, an analytical study on the impact of harmonic injection on the operation and design of the MMC will be presented. A mathematical model of the MMC will be presented considering fundamental, 2nd and the 3rd harmonic components.

Chapter 4

4 Energy Storage Requirements for the MMC with Generalised Harmonic Injection

4.1 Introduction

In Chapter 3, the generalised modular multilevel converter (MMC) for HVDC systems was discussed. The operating principle of the generalised MMC was presented considering the instantaneous power and energy. In this chapter, an analytical study on the storage energy requirements of the MMC is given. Steady state operation with fundamental frequency and additional harmonic injection is considered including using the 3rd harmonic of voltage or/and current and the 2nd harmonic of current.

In [9], 3rd harmonic voltage injection was first proposed to improve the utilization of the DC voltage in an MMC for high voltage applications. The classical 3rd harmonic voltage injection was also implemented for MMC by [49] to extend the operating range of the converter where the dynamic performance of the MMC was considered. The authors presented a phase-disposition sinusoidal pulsewidth modulation (PD-SPWM) strategy and a voltage balancing method of the SM capacitor voltages and devolving a mathematical model under balanced and unbalanced grid operation modes.

The authors in [15] investigated the impact of the classical 3rd harmonic voltage injection on the energy storage requirements of the MMC, presenting total energy storage needs per transferred VA. Operation of the MMC with the addition of harmonic voltage injection has been presented previously considering the use of 3rd harmonic injection and its effect on modulation, control [49] and the energy storage requirement [16]. However, the effect of the phase angle of the 3rd harmonic voltage component on both the energy storage requirement and the synthesised AC voltage has not been considered yet. Moreover, the limitations of using 3rd harmonic voltage injection and also the applicable range with regards to the active power (P) and reactive power (Q) capability were not considered by either of the studies presented by [49] and [16]. Therefore, some

of these limitations related to maximum modulation index and the range of P and Q are considered by the study presented in this thesis.

The operation of MMC under voltage constraints was considered by [17] where the benefit of using the classical 3rd harmonic voltage injection to extend the operating voltage of the MMC was discussed. The limitations of using the 3rd harmonic voltage injection, however, was not considered by the authors which will be investigated in more detail in this thesis considering both classical and also phase shifted 3rd harmonic injection. The influence of classical 3rd harmonic injection on system power losses, size of SM capacitance, circulating current, and also fault current are evaluated considering station conduction losses which are reduced by 11%, yielding higher efficiency and lowering cooling system capacity [66].

3rd harmonic current injection alone was considered for both the AAC and MMC in [18] where the injected 3rd harmonic currents are used to produce a DC power component into the SMs. This 3rd harmonic current component can be controlled in magnitude and phase to produce a net flow of energy into the SM capacitors. Therefore, controlling the injected 3rd harmonic current can lead to control the SM capacitor voltages. Considering the phase angle (phase sifted), 3rd harmonic voltage injection was not considered by any of the mentioned references.

Other low order harmonic components that could be considered to improve the performance of the MMC are the 2nd harmonic voltages and currents. The 2nd harmonic current in the MMC arm is an inherent phenomenon caused by the non-ideal modulation that leads to a 100Hz component in the circulating current. Various aspects of the MMC performance are affected by the 2nd harmonic current such as capacitor voltage ripple (and therefore available voltage), power losses and the required current rating of the semiconductors. For the reasons above, this current is normally controlled to be zero. However, it can be observed that using 2nd harmonic current injection can lead to a minimization of the capacitor voltage ripple [67]. To inject a 2nd harmonic current, a current controller is used to generates the 2nd harmonic voltage demand which is required to circulate the requested amount of 2nd harmonic current. In [19], 2nd and 4th harmonic current injection were investigated considering the optimal current injection required to reduce the available voltage ripple depending on the P and Q operating point. It is noted that the phase angle of the investigated currents was considered. The authors

in [68] investigated the impact of 2nd harmonic current injection on extending the P and Q operating envelope of the converter where a 2nd harmonic component was added to the modulation signals to drive the required 2nd harmonic current. The same was done with [69] where the controller generates the 2nd harmonic voltage reference to inject the required 2nd harmonic current. An improved steady-state analysis of the MMC operation with 2nd harmonic current injection was presented in [69]. It can be observed that a small value of 2nd harmonic added to the modulation signals can lead to the injection of a relatively high 2nd harmonic current. Therefore, the addition of the 2nd harmonic current has a negligible impact on the voltage waveforms generated by the converter.

However, the effect of the 2nd harmonic current injection has been extensively investigated and reported in [19, 67-69] without considering the impact of co-existing 3rd harmonic injection. Therefore, as part of this work, the effect of the 2nd harmonic currents with the existence of 3rd harmonic injection will be studied. In studying the effect of the 2nd harmonic current in the limbs, the attendant 2nd harmonic voltage due to the interaction of the current and the limb impedance is expected to be minimal and is therefore neglected throughout the studies.

Normally, the purpose of 3rd harmonic voltage injection in VSCs is to reduce the peak synthesised voltage and therefore improve the voltage capability of the converters, generating larger fundamental voltage. The same concept can also be applied to the MMC where using 3rd harmonic voltage injection can lead to increase the margin between the available and the requested voltage. As opposed to 2-level VSCs, in the MMC the 3rd harmonic injection can also be used to maintain the same voltage capability but reducing the SM capacitors, allowing larger ripple. Therefore, analysis considering classical and phase shifted 3rd harmonic voltage injection was done to investigate the impact on energy storage requirements. To provide a comprehensive analysis of the potential solutions to achieve higher voltage capability or smaller energy storage in the MMC, this thesis considers different scenarios including 3rd harmonic voltage, 3rd harmonic current and 2nd harmonic current injections. The 3rd harmonic voltage is injected with the modulation voltage while the 3rd harmonic current can be injected using an external circuit as will be discussed later and shown Figure 4.25. The 2nd harmonic current can be injected using a PR controller where its output can be added to the modulation signal.

For HVDC applications where high power transmission requires to operate with high voltage to meet the operating conditions, low limb currents (and therefore device) help to reduce the converter station losses and therefore improve the system efficiency. Therefore, it is desirable to operate the MMC with high voltage and lower limb current taking into the account the energy storage requirements. In fact, each SM in the MMC based on half bridges has only two outputs, i.e. 0 or $+V_{CSM}$, whilst the full bridge based MMC has SMs that can output $-V_{CSM}$, 0 or $+V_{CSM}$. Therefore, the synthesised voltage of the MMC based on half bridge SMs must always be positive while the synthesised voltage of the full bridge based MMC can be positive or negative. Therefore, to increase the AC output voltage of the MMC using the classical modulation without 3rd harmonic injection cannot be achieved for the MMC based on half bridge SMs, and a full bridge or an hybrid solution would be required. Of course, other solution is using a bigger SM capacitor size. However, this will impact on the size and the cost of the MMC design while the synthesised voltage still limited by the P and Q operating points.

Second harmonic and 3rd harmonic injection has already been presented in the literature. It has been shown that these harmonics can have an effect on improving the operating performance for the modular multilevel based converters [16, 18, 19, 69]. The idea of 2nd harmonic injection is normally referred to as the circulating current and has already been discussed in the literature including its control [64] and its performance optimisation. It has been shown that not only the 2nd harmonic can improve operating performance with regards to capacitor voltage ripple but also that 4th harmonic current injection can be advantageous [19, 69]. Minimising the capacitor voltage ripple can help to optimise the energy storage requirements including the capacitor size. Also, it can increase the margin between the available voltage and the synthesised voltage. The latter then can be increased by increasing the fundamental component.

However, using 2nd harmonic current injection has its limitations with regards to reducing the current capability due to current stress on the devices. Moreover, more 2nd harmonic current means more 2nd harmonic voltage component injected with the synthesised voltage which can cause an error by increasing the peak of the synthesised voltage. While the focus here is to reduce the peak synthesised voltage in order to increase the fundamental component. Therefore, increasing the 2nd harmonic component in fact has a negative impact on increasing the peak of the synthesised voltage as shown

by Figure 4.1. Compared with 3rd harmonic voltage injection, the latter has the ability to reduce the peak of the synthesised voltage has been reported for VSC based converters [70, 71].

Please note that Figure 4.1 is given only as an example where the amplitudes of the 2nd harmonic voltage and current have been exaggerated on purpose to make them clearly visible. In practical applications, the 2nd harmonic current is typically controlled to zero or to a small amount of 2nd harmonic current. As a consequence, the 2nd harmonic voltage component needed to drive the current can be neglected compared to the fundamental component.

However, the operation with a high synthesised voltage can be limited by various factors such as: available limb voltage, the MMC rated apparent power, and PQ operation point. The available voltage should be big enough with low ripple to meet the voltage target. Unfortunately, operating with as high a synthesised voltage as possible, can be limited especially for a half-bridge based MMC without using 3rd harmonic voltage injection. It is reported that the overmodulation strategies can be achieved for VSCs using 3rd harmonic injection allowing reducing peak synthesised voltage whilst increasing the fundamental by up to 15% [70, 71]. The MMC, like the traditional VSC, can benefit from the improved voltage range with the 3rd harmonic injection as explored in [17].

It is noted that V_1 is the synthesised voltage with no 2nd harmonic injection (fundamental component). V_{12} is the synthesised voltage based on the combination of the fundamental and 2nd harmonic voltage components. θ_{v2} is the phase angle of the 2nd harmonic component.

It can be seen that for high value of the 2nd harmonic component for different phase angles can lead to increase the peak of the synthesised voltage. In fact this can lead to increase the peak of the synthesised voltage. This can lead to an overmodulation when considering the available voltage or the synthesised voltage can hit the lower limit (zero). It is worth noting that in typical designs the condition shown in Figure 4.1 is not a concern, since the amplitude of the 2nd harmonic voltage will be small. However, in different designs with different requirements for the 2nd harmonic voltage, the overmodulation limit could be an issue and in general it must be taken into account.

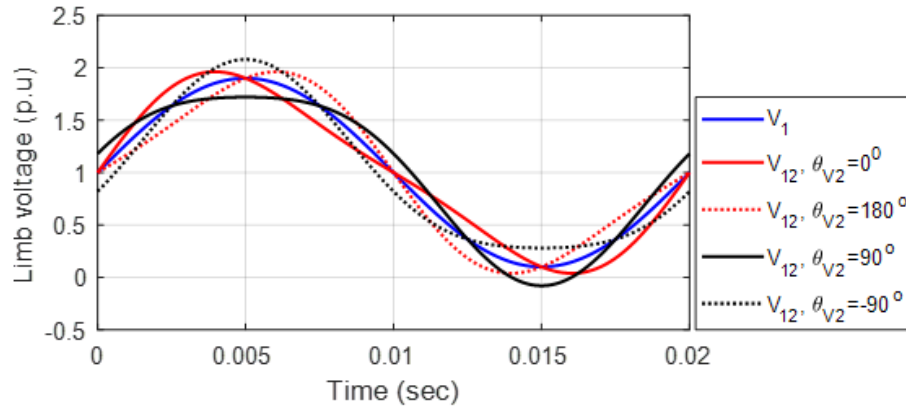


Figure 4.1: The impact of 2nd harmonic component on the inserted voltage

Although the MMC offers significant advantages for HVDC system applications, the energy storage requirement significantly affects the size of the converter station. In offshore and other similar applications where size and weight are critical, a reduction in the energy storage components in the MMC circuit could be economically beneficial. The impact of 3rd harmonic voltage injection on energy storage requirement has already been reported [16]. However, considering both 3rd harmonic voltage and current has not been considered yet. Moreover, the impact of the phase angle of the 3rd harmonic injection has not been considered yet.

The combination between any of the mentioned methods has not been considered yet (classical and phase-shifted) which will be considered by the investigation in this thesis. Therefore, the study presented by this thesis aims at filling this gap, presenting a comparative study of the impact of 3rd harmonic voltage injection, 3rd harmonic current and 2nd harmonic current injection on the MMC operation including the energy storage requirements. To date, no other studies have considered the phase shifted 3rd harmonic voltage injection to improve the operation and design of the MMC. Therefore, within this Chapter, research into the investigation of the effect of harmonic component injection on the energy storage requirement of a converter submodule is presented. In addition, the impact of the 2nd harmonic current injection is investigated with the present of 3rd harmonic voltage injection.

The investigation considers an MMC system based on half-bridge submodules as shown in Figure 3.1. A mathematical model is presented for the MMC considering steady state operation which can help to predict MMC behaviour for different operating conditions. The model is also used to help the design of the MMC including its SM capacitor size and operation voltage level as will be discussed in Chapter 5 and 6 respectively.

4.2 Generalised Injection Analysis

In the previous chapter, the analysis of standard operation of the MMC was presented. In this subsection, however, a generalised analysis based on additional harmonic injection (m th and n th harmonic order) under steady state operation is presented. A generalised form of limb voltage, current, instantaneous power and energy are presented.

For a periodic function $f(t)$, the function $f(t)$ can be represented by the Fourier function expressed by (4.1) [72].

$$f(t) = a_v + \sum_{n=1}^{\infty} (a_n \cos n\omega_0 t + b_n \sin n\omega_0 t) \quad (4.1)$$

Where a_v , a_n and b_n are the Fourier coefficients, n is equal to 1,2,3,.. and ω_0 is the fundamental frequency of the function $f(t)$. However, the Fourier function given above can be rewritten as equation (4.2) below:

$$f(t) = a_v + \sum_{n=1}^{\infty} A_n \cos(n\omega_0 t - \theta_n) \quad (4.2)$$

where a_v and A_n are Fourier coefficients. n is equal 1,2,3,.. and ω_0 is the fundamental frequency of the function $f(t)$. Where, $A_n = \sqrt{a_n^2 + b_n^2}$ and $\theta_n = \tan^{-1} \frac{b_n}{a_n}$. Thus the limbs voltage $v_p(t)$ and current $i_p(t)$ of MMC can be represented as a single phasor quantity.

The Fourier coefficients of (4.1) can represent the MMC limb voltages and currents. Hence, a_v represents the DC offset while A_n represents the harmonic contents of the limb voltage or current which has a fundamental frequency of ω_0 . Therefore, the MMC converter upper and lower limb voltage and the current of phase A can be given by equations (4.3) and (4.4).

$$v_{pos}(t) = \frac{V_{dc}}{2} - \sum_{n=1}^{\infty} \hat{V}_{cn} \cos(n\omega_0 t - \theta_{vn}) \quad (4.3)$$

$$i_{pos}(t) = \frac{I_{dc}}{3} + \sum_{m=1}^{\infty} \frac{\hat{I}_{cm}}{2} \cos(m\omega_0 t - \theta_{im}) \quad (4.4)$$

Similarly, for the negative limb;

$$v_{neg}(t) = \frac{V_{dc}}{2} + \sum_{n=1}^{\infty} \hat{V}_{cn} \cos(n\omega_0 t - \theta_{vn}) \quad (4.5)$$

$$i_{neg}(t) = \frac{I_{dc}}{3} - \sum_{m=1}^{\infty} \frac{\hat{I}_{cm}}{2} \cos(m\omega_0 t - \theta_{im}) \quad (4.6)$$

Therefore, the positive limb instantaneous power at the limb terminals is $v_{pos}i_{pos}$.

$$p_{pos}(t) = v_{pos}(t)i_{pos}(t) \quad (4.7)$$

By substituting (4.5) and (4.6) into (4.7), the instantaneous and average power of the upper limb can be obtained. The positive limb instantaneous power can then be given by (4.8). The instantaneous energy then can be obtained by integrating the instantaneous power in equation (4.8). This yields;

$$p_{pos}(t) = \left[\frac{V_{dc}I_{dc}}{6} + \frac{V_{dc}}{4} \sum_{m=1}^{\infty} \hat{I}_{cm} \cos(m\omega_0 t - \theta_{im}) - \frac{I_{dc}}{3} \sum_{n=1}^{\infty} \hat{V}_{cn} \cos(n\omega_0 t - \theta_{vn}) \right. \\ \left. - \sum_{n=1}^{\infty} \sum_{m=1}^{\infty} \frac{\hat{V}_{cn}\hat{I}_{cm}}{4} \left\{ \cos((n-m)\omega_0 t + (-\theta_{vn} + \theta_{im})) \right\} \right. \\ \left. + \cos((n+m)\omega_0 t + (-\theta_{vn} - \theta_{im})) \right] \quad (4.8)$$

$$E_{pos}(t) = \int p_{pos}(t) \cdot dt \quad (4.9)$$

$$e_{pos}(t) = \left[\frac{V_{dc}I_{dc}}{6} t + \frac{V_{dc}}{4\omega_0} \sum_{m=1}^{\infty} \frac{\hat{I}_{cm}}{m} \sin(m\omega_0 t - \theta_{im}) - \frac{I_{dc}}{3\omega_0} \sum_{n=1}^{\infty} \frac{\hat{V}_{cn}}{n} \sin(n\omega_0 t - \theta_{vn}) \right. \\ \left. - \frac{1}{4\omega_0} \sum_{n=1}^{\infty} \sum_{m=1}^{\infty} \hat{V}_{cn}\hat{I}_{cm} \left\{ \frac{1}{n-m} \sin((n-m)\omega_0 t + (-\theta_{vn} + \theta_{im})) \right\} \right. \\ \left. + \frac{1}{n+m} \sin((n+m)\omega_0 t + (-\theta_{vn} - \theta_{im})) \right] \quad (4.10)$$

The equations above are the general expression of the positive arm instantaneous power and energy including the m and n harmonic components.

The average power is also can be calculated from Equation (4.8). Only terms where the voltage and current have the same frequency, ($m=n$), give average power. This yields;

$$\bar{P}_{pos}^{ave} = \left[\frac{V_{dc}I_{dc}}{6} - \sum_{m=n=1}^{\infty} \frac{\hat{V}_{cn}\hat{I}_{cn}}{4} \cos((- \theta_{vn} + \theta_{in})) \right] \quad (m = n) \quad (4.11)$$

Using the generalised equations above, the impact of any particular harmonic injection on energy variation can be observed. The previous analysis gives the general case. In the following analysis, the investigation is focused on the impact of some specific harmonic components on energy deviation reduction, including 2nd harmonic current and 3rd harmonic voltage and current.

4.3 Third Harmonic Voltage and Current Injection Analysis

In Chapter 3 a theoretical analysis considering the fundamental component of the converter limb voltages and currents was presented. The importance of using 3rd harmonic voltage injection on energy storage requirements has already been presented in [16]. However, a detailed analysis on the impact of both 3rd harmonic voltage and current considering their phase angle has not been considered before and is therefore discussed below.

Third harmonic voltage injection can be achieved by adding the 3rd harmonic component to the synthesised limb voltage using an appropriate modulation scheme. To allow a 3rd harmonic current, a path for the 3rd harmonic current needs to be provided. This can be obtained either naturally from the neutral point of the converter transformer or by an auxiliary external circuit. However, using the first method requires high impedance connected to the neutral point to control the zero sequence current. The auxiliary injection circuit is required to control the zero sequence current if the second method is considered [18]. For both methods the neutral point of the Wye/Delta transformer needs to be connected to the converter side of the transformer while the Delta is connected to the grid side. Figure 4.2 compares two converter transformer configurations (a) with a path of the 3rd harmonic while in (b) zero sequence current is eliminated.

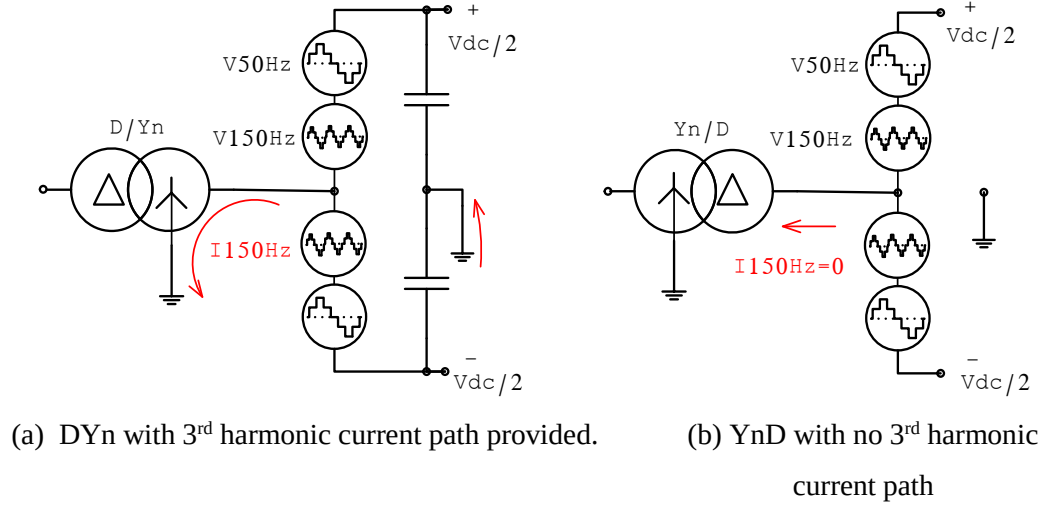


Figure 4.2: Converter transformer configurations for MMC.

Generally, both the triple harmonic current and voltage can be present and the periodic voltages and currents of the positive limb are given by equations (4.12) and (4.13) where the amplitude of the 3rd harmonic voltages and current components are expressed by $\hat{V}_{C3}$ and $\hat{I}_{C3}$ with their phase shift of θ_{v3} and θ_{i3} , respectively. θ_{v1} is used to distinguish between the phase angle of the fundamental, 3rd voltage and current components.

$$v_{pos.a}(t) = \frac{V_{dc}}{2} - \hat{V}_{C1} \cos(\omega_0 t - \theta_{v1}) + \hat{V}_{C3} \cos(3\omega_0 t - \theta_{v3}) \quad (4.12)$$

$$i_{pos.a}(t) = \frac{I_{dc}}{3} + \frac{\hat{I}_{C1}}{2} \cos(\omega_0 t - \theta_{i1}) + \frac{\hat{I}_{C3}}{2} \cos(3\omega_0 t - \theta_{i3}) \quad (4.13)$$

Due to the symmetry between the positive and negative limbs, the negative limb periodic voltage and current is given by (4.14) and (4.15).

$$v_{neg.a}(t) = \frac{V_{dc}}{2} + \hat{V}_{C1} \cos(\omega_0 t - \theta_{v1}) - \hat{V}_{C3} \cos(3\omega_0 t - \theta_{v3}) \quad (4.14)$$

$$i_{neg.a}(t) = \frac{I_{dc}}{3} - \frac{\hat{I}_{C1}}{2} \cos(\omega_0 t - \theta_{i1}) - \frac{\hat{I}_{C3}}{2} \cos(3\omega_0 t - \theta_{i3}) \quad (4.15)$$

4.3.1 Classical and third harmonic phase-shift

Before starting the analysis for specific harmonic content, let's first define the relationship between the position of the 3rd harmonic voltage waveform and the fundamental in the time domain.

In order to avoid any ambiguity in describing the relative positions of the fundamental and the injected harmonics on the time axis, the phase shift parameter is always used so that it represents the time shift between positive zero crossings of the respective waveforms. Thus

$$\theta_{v3} = 3\omega_0 t_{shift} \quad (4.16)$$

where t_{shift} is the time displacement between the positive zero crossing of the 3rd harmonic and an un-shifted fundamental. Note that when the waveforms are expressed as cosines (as in 4.11 for example) this convention requires opposite signs in the summation of the fundamental and 3rd harmonic components.

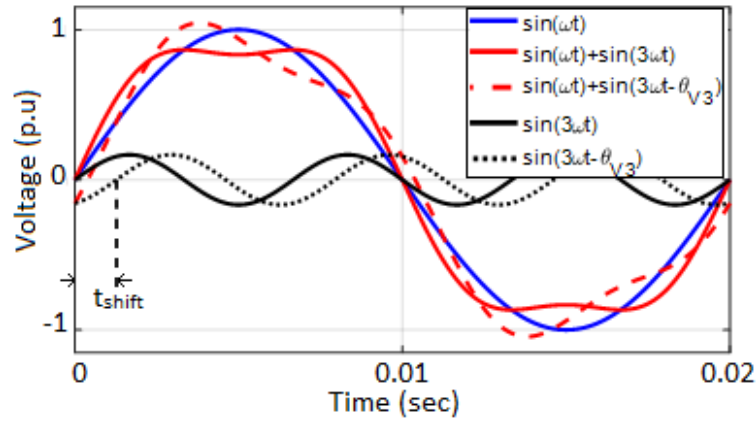


Figure 4.3: 3rd harmonic phase shift.

4.4 Third Harmonic Voltage Injection (V_{C3})

The focus on this section is on the analysis of the effect of the 3rd harmonic voltage injection. This is done by eliminating the 3rd harmonic current component (triplen current injection) in the previous section in Equations (4.13) and (4.15). In the following analysis and results, a 3rd harmonic voltage injection equal to 15% of the fundamental will generally be considered as an example to evaluate the impact of the injection technique on the converter. However, the impact of increasing the 3rd harmonic injection is also considered.

4.4.1 Limb Voltage, Current, Power and Energy

It is assumed that the (YD) converter transformer configuration is connected as given by Figure 4.2.b otherwise a path of zero sequence current is provided. Therefore, such a configuration will prevent a 3rd harmonic current path and the positive limb voltage

and current in (4.12) and (4.13) are re-expressed in (4.17) and (4.18). Substituting them into (4.19) yields the instantaneous power of the positive limb given by (4.20):

$$v_{pos.a}(t) = \frac{V_{dc}}{2} - \hat{V}_{C1} \cos(\omega_0 t - \theta_{v1}) + \hat{V}_{C3} \cos(3\omega_0 t - \theta_{v3}) \quad (4.17)$$

$$i_{pos.a}(t) = \frac{I_{dc}}{3} + \frac{\hat{I}_{C1}}{2} \cos(\omega_0 t - \theta_{i1}) + (I_{cir} = 0) \quad (4.18)$$

$$p_{pos}(t) = v_{pos}(t) \cdot i_{pos}(t) \quad (4.19)$$

$$p_{pos.a}(t) = \left[\begin{aligned} & \left(\frac{V_{dc}I_{dc}}{6} - \frac{\hat{V}_{C1}\hat{I}_{C1}}{4} \cos(\theta_{i1} - \theta_{v1}) \right) \\ & + \left(\frac{V_{dc}\hat{I}_{C1}}{4} \cos(\omega_0 t - \theta_{i1}) - \frac{\hat{V}_{C1}I_{dc}}{3} \cos(\omega_0 t - \theta_{v1}) \right) \\ & \quad - \frac{\hat{V}_{C1}\hat{I}_{C1}}{4} \cos(2\omega_0 t - \theta_{v1} - \theta_{i1}) \\ & + \left(\frac{\hat{V}_{C3}I_{dc}}{3} \cos(3\omega_0 t - \theta_{v3}) + \frac{\hat{V}_{C3}\hat{I}_{C1}}{4} \cos(2\omega_0 t + \theta_{i1} - \theta_{v3}) \right) \\ & \quad + \frac{\hat{V}_{C3}\hat{I}_{C1}}{4} \cos(4\omega_0 t - \theta_{i1} - \theta_{v3}) \end{aligned} \right] \quad (4.20)$$

The total average power of (4.20) equals the sum of the average powers of the products of the currents and voltages of the same frequency and is expressed in (4.21). Clearly there is no contribution to the average power due to the 3rd harmonic because of the YD converter transformer configuration which eliminates the 3rd harmonic current component. For simplicity, it is assumed that the phase angle of the fundamental component of the alternating voltage θ_{v1} is zero and therefore the power angle, $\emptyset$ represents only the current phase angle θ_{i1} .

$$\bar{P}_{pos.a} = \frac{V_{dc}I_{dc}}{6} - \frac{\hat{V}_{C1}\hat{I}_{C1}}{4} \cos(\emptyset) \quad (4.21)$$

To obtain capacitor charge balancing in the MMC and therefore achieve sustainable converter operation, the net power into the converter must be zero. By using the relationship between the direct and alternating voltages expressed in (3.13) the relationship between the currents is similar to the fundamental operation expressed in (3.22).

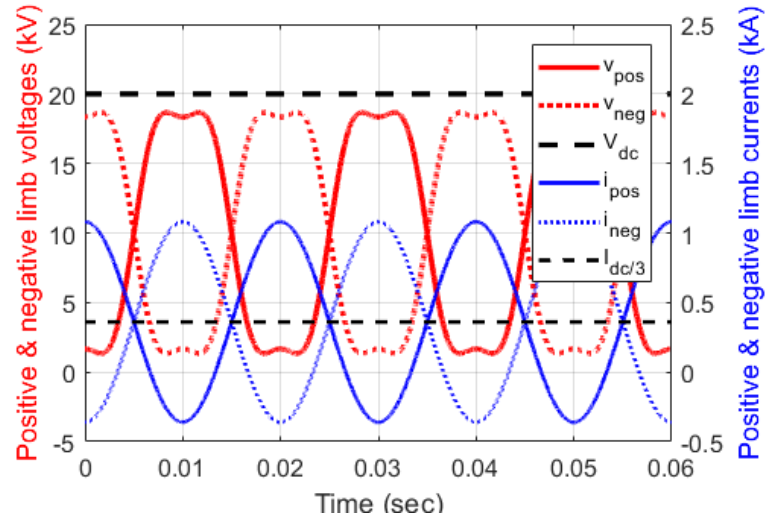
$$I_{dc} = \frac{3}{4} m_i \hat{I}_{C1} \cos(\emptyset) \quad (4.22)$$

The energy variation of the positive limb is expressed by (4.23). Due to symmetry, the negative limb is analysed in the same manner. Therefore the negative limb energy variation $e_{neg}(t)$ is equal to the positive limb energy variation of the previous half cycle.

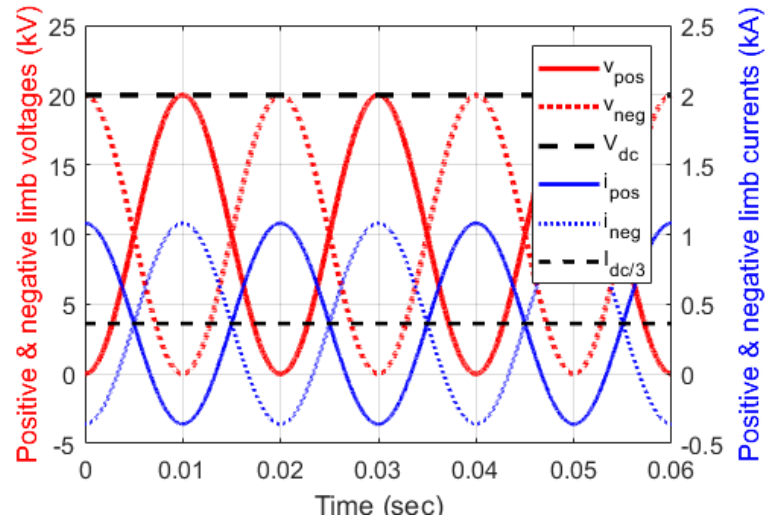
$$e_{pos.a}(t) = \left[\begin{aligned} & \left(\frac{V_{dc} I_{dc}}{6} - \frac{\hat{V}_{C1} \hat{I}_{C1}}{4} \cos(\theta_{i1} - \theta_{v1}) \right) t \\ & + \left(\frac{V_{dc} \hat{I}_{C1}}{4\omega_0} \sin(\omega_0 t - \theta_{i1}) - \frac{\hat{V}_{C1} I_{dc}}{3\omega_0} \sin(\omega_0 t - \theta_{v1}) \right. \\ & \quad \left. - \frac{\hat{V}_{C1} \hat{I}_{C1}}{8\omega_0} \sin(2\omega_0 t - \theta_{v1} - \theta_{i1}) \right) \\ & + \left(\frac{\hat{V}_{C3} I_{dc}}{9\omega_0} \sin(3\omega_0 t - \theta_{v3}) + \frac{\hat{V}_{C3} \hat{I}_{C1}}{8\omega_0} \sin(2\omega_0 t + \theta_{i1} - \theta_{v3}) \right. \\ & \quad \left. + \frac{\hat{V}_{C3} \hat{I}_{C1}}{16\omega_0} \sin(4\omega_0 t - \theta_{i1} - \theta_{v3}) \right) \end{aligned} \right] \quad (4.23)$$

For better understanding of the concept, the positive limb voltage and current given by equations (4.17) and (4.18) is illustrated in Figure 4.4 where a comparison is presented between fundamental operation and classical 3rd harmonic voltage injection with zero phase angle (classical 3rd harmonic).

Using 3rd harmonic voltage injection allows a reduction of the peak synthesised voltage of the positive and negative limbs in Figure 4.4.a compared with Figure 4.4.b while the fundamental is still the same for both. The synthesised voltage is still less than the available voltage, $\sum V_{cap} = V_{dc}$, which in this case is assumed to be for an ideal MMC with a large size capacitor. As can be seen, the fundamental limb voltage can be increased compared with the synthesised limb voltage in Figure 4.4.b. Therefore, for a given rated power, by increasing the modulation index using the advantage of 3rd harmonic voltage injection, the AC current can be reduced. This can lead to a reduction of converter conduction losses.



(a) Due to classical 3rd harmonic voltage injection



(b) Due to standard operation

Figure 4.4: Waveforms of limb voltages and currents of an ideal MMC with 3rd harmonic voltage and without 3rd harmonic current injection.

4.4.2 Instantaneous Power and Energy as a Function of the Apparent Power $|S|$

To achieve sustainable converter operation, the total average energy of one limb for one periodic cycle is zero [44, 45]. This means that once the MMC has been charged with the nominal energy, the capacitor nominal voltages should be maintained at the same level for one periodic cycle. This also means that active power absorbed by the grid at the AC side equals the direct power at the DC-link side considering the losses. Making the total average power in equation (4.21) equal zero, the direct power equals the

average AC active power. This can be rearranged where the power factor is represented by $\phi = \theta_{i1}$ and the phase angle of voltage $\theta_{v1} = 0$.

$$V_{dc}I_{dc} = \frac{3}{2} \hat{V}_{c1} \hat{I}_{c1} \cos(\phi) \quad (4.24)$$

The relationship between the alternating and direct voltages has been given in (3.13) but can be rewritten here as (4.25). The relationship between the direct and alternating line current is given by (4.22) also rewritten here by (4.26). The Absolute Value of the apparent power is denoted by $|S|$ while the amplitude of the fundamental phase voltage and line current are denoted by $\hat{V}_{c1}$ and $\hat{I}_{c1}$. The 3rd harmonic component can be expressed as a percentage of the fundamental component, β , as (4.28).

$$\hat{V}_{c1} = m_i \frac{V_{dc}}{2} \quad (4.25)$$

$$I_{dc} = \frac{3}{4} m_i \hat{I}_{c1} \cos(\phi) \quad (4.26)$$

$$|S| = \left| \frac{3}{2} \hat{V}_{c1} \hat{I}_{c1} \right| \quad (4.27)$$

$$\hat{V}_{c3} = \beta \hat{V}_{c1} \quad (4.28)$$

By substituting equations from (4.25) to (4.28) into (4.20) and (4.23), the instantaneous power and energy equations can be rearranged and expressed by.

$$p_{pos.a}(t) = \frac{|S|}{6} \left[\left(\frac{2}{m_i} \cos(\omega_0 t - \phi) - m_i \cos(\phi) \cos(\omega_0 t) - \cos(2\omega_0 t - \phi) \right) + \left(m_i \beta \cos(\phi) \cos(3\omega_0 t - \theta_{v3}) + \beta \cos(2\omega_0 t + \phi - \theta_{v3}) + \beta \cos(4\omega_0 t - \phi - \theta_{v3}) \right) \right] \quad (4.29)$$

$$e_{pos.a}(t) = \frac{|S|}{6\omega_0} \left[\left(\frac{2}{m_i} \sin(\omega_0 t - \phi) - m_i \cos(\phi) \sin(\omega_0 t) - \frac{1}{2} \sin(2\omega_0 t - \phi) \right) + \left(\frac{m_i \beta \cos(\phi)}{3} \sin(3\omega_0 t - \theta_{v3}) + \frac{\beta}{2} \sin(2\omega_0 t + \phi - \theta_{v3}) + \frac{\beta}{4} \sin(4\omega_0 t - \phi - \theta_{v3}) \right) \right] \quad (4.30)$$

Due to symmetry, the negative limb power and energy is equal to positive limb at the previous cycle, $p_{neg.a}(t) = p_{pos.a}\left(t - \frac{T}{2}\right)$.

$$p_{neg.a}(t) = \frac{|S|}{6} \left[\begin{aligned} &\left(-\frac{2}{m_i} \cos(\omega_0 t - \phi) + m_i \cos(\phi) \cos(\omega_0 t) - \cos(2\omega_0 t - \phi) \right) \\ &- \left(m_i \beta \cos(\phi) \cos(3\omega_0 t - \theta_{v3}) - \beta \cos(2\omega_0 t + \phi - \theta_{v3}) \right. \\ &\quad \left. - \beta \cos(4\omega_0 t - \phi - \theta_{v3}) \right) \end{aligned} \right] \quad (4.31)$$

$$e_{neg.a}(t) = \frac{|S|}{6\omega_0} \left[\begin{aligned} &\left(-\frac{2}{m_i} \sin(\omega_0 t - \phi) + m_i \cos(\phi) \sin(\omega_0 t) - \frac{1}{2} \sin(2\omega_0 t - \phi) \right) \\ &- \left(\frac{m_i \beta \cos(\phi)}{3} \sin(3\omega_0 t - \theta_{v3}) - \frac{\beta}{2} \sin(2\omega_0 t + \phi - \theta_{v3}) \right. \\ &\quad \left. - \frac{\beta}{4} \sin(4\omega_0 t - \phi - \theta_{v3}) \right) \end{aligned} \right] \quad (4.32)$$

The available voltage within the SM capacitors then can be determined using equation (3.42). Thus;

$$\Sigma v_{C_{SM}.a}(t) = \sqrt{\left(\frac{2N_{SM}}{C_{SM}} (e_a(t) + E_{ave}) \right)} \quad (4.33)$$

For a zero phase-shifted 3rd harmonic, the phase angle θ_{v3} equal zero. Therefore, equations (4.29) to (4.33) can be further simplified and the positive and negative limb voltages and currents of Phase A are illustrated by Figure 4.5. and 4.6.

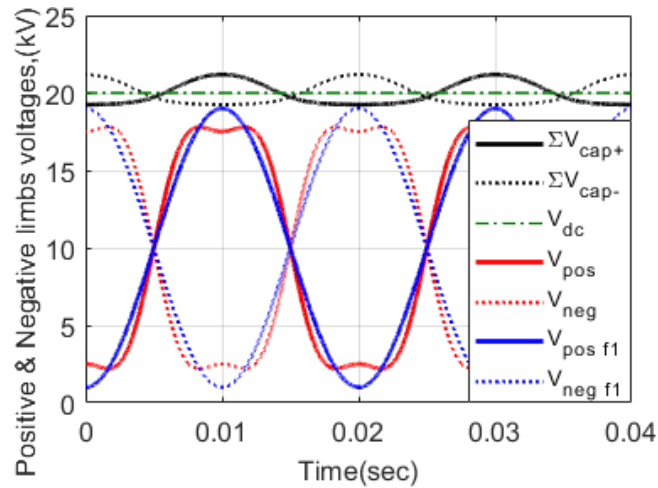


Figure 4.5: The impact of 3rd harmonic injection on limb currents for $S=0+jQ$.

Figure 4.5 also illustrates the impact of 3rd harmonic voltage injection on the synthesised limb voltages and currents for a non-ideal MMC when finite capacitor voltage ripple is considered. It can be seen that the 3rd harmonic voltage injection can lead to reduce peak synthesised voltage whilst the limb currents are not affected. This simulation results are based on a large size of capacitor of $12.0mF$ and operating point of rated $|S|$ with $P=0$

and +Q. It can be seen that for modulation index of 0.9, also the available voltages are not affected with using 15% of 3rd harmonic voltage injection.

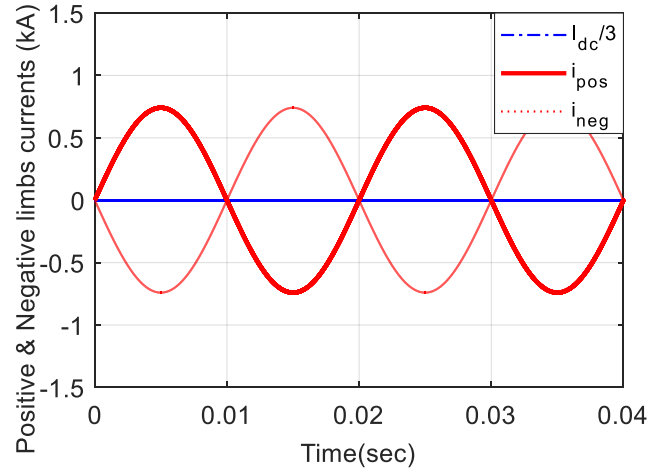


Figure 4.6: The impact of 3rd harmonic injection on limb currents for S=0+jQ.

4.4.3 Peak-to-peak Energy (Energy Deviation)

This subsection presents an investigation on the effect of 3rd harmonic voltage injection on energy deviation. Using the same principle of the analysis in section 3.5, maximum energy deviation, considering the impact of harmonic components, can be obtained here. The time at which maximum and minimum energy occurs remains the same as it depends only on the current equation since the 3rd harmonic injection affects only the synthesised voltage waveform. Energy variation has already been discussed in section 4.4.2 and expressed by equation (4.30). It is observed that energy variation using 3rd harmonic injection is accompanied by two terms, one caused by the fundamental component while the other is caused by the 3rd harmonic.

$$e_{pos.a}(t) = e_{pos.f1}(t) + e_{pos.f3}(t) \quad (4.34)$$

Energy deviation can be expressed by (4.35) where maximum and minimum energy both have two components. This can be given by equations (4.36) and (4.37)

$$\Delta E = E_{max} - E_{min} \quad (4.35)$$

$$E_{max} = E_{pos.t_{max}}^{f1} + E_{pos.t_{max}}^{f3} \quad (4.36)$$

$$E_{min} = E_{pos.t_{min}}^{f1} + E_{pos.t_{min}}^{f3} \quad (4.37)$$

where $E_{pos.t_{max}}^{f1}$ and $E_{pos.t_{max}}^{f3}$ are the maximum energy due to fundamental frequency and 3rd harmonic voltage injection, respectively. Thus;

$$E_{pos.t_{max}}^{f_1} = \frac{|S|}{6\omega_0} \left[\left(\frac{2}{m_i} \sin(\omega_0 t_{max} - \phi) - m_i \cos(\phi) \sin(\omega_0 t_{max}) \right) - \frac{1}{2} \sin(2\omega_0 t_{max} - \phi) \right] \quad (4.38)$$

$$E_{pos.t_{max}}^{f_3} = \frac{\beta|S|}{6\omega_0} \left[\begin{aligned} &\frac{m_i \cos(\phi)}{3} \sin(3\omega_0 t_{max} - \theta_{v3}) \\ &+ \frac{1}{2} \sin(2\omega_0 t_{max} + \phi - \theta_{v3}) \\ &+ \frac{1}{4} \sin(4\omega_0 t_{max} - \phi - \theta_{v3}) \end{aligned} \right] \quad (4.39)$$

Similarly;

$$E_{pos.min.t_{min}}^{f_1} = \frac{|S|}{6\omega_0} \left[\left(\frac{2}{m_i} \sin(\omega_0 t_{min} - \phi) - m_i \cos(\phi) \sin(\omega_0 t_{min}) \right) - \frac{1}{2} \sin(2\omega_0 t_{min} - \phi) \right] \quad (4.40)$$

$$E_{pos.t_{min}}^{f_3} = \frac{\beta|S|}{6\omega_0} \left[\begin{aligned} &\frac{m_i \cos(\phi)}{3} \sin(3\omega_0 t_{min} - \theta_{v3}) \\ &+ \frac{1}{2} \sin(2\omega_0 t_{min} + \phi - \theta_{v3}) \\ &+ \frac{1}{4} \sin(4\omega_0 t_{min} - \phi - \theta_{v3}) \end{aligned} \right] \quad (4.41)$$

By substituting (4.36) and (4.37) into (4.35):

$$\Delta E = (E_{pos.t_{max}}^{f_1} - E_{pos.t_{min}}^{f_1}) + (E_{pos.t_{max}}^{f_3} - E_{pos.t_{min}}^{f_3}) \quad (4.42)$$

4.4.4 Simulation Results and Discussion

For the given MMC system parameters in Table 3.2, Figure 4.7 illustrates the impact of classical and phase-shifted 3rd harmonic voltage injection on the peak synthesised voltage. It can be seen that the peak synthesised voltage can be increased or reduced while using the advantage of the phase-shifted 3rd harmonic voltage injection. This can be implemented in order to meet special operating conditions.

Considering Figure 4.7, the dotted red waveform is the positive limb synthesised voltage with the phase angle of the 3rd harmonic component set to zero. The MMC operates such that high modulation index, $m_i = 1.05$, with SM capacitor size of $12.0mF$. The operating point of P and Q is $P = +0.94$ p.u and $Q = -0.33$ p.u with the rated apparent power, S . It is noted that the 3rd harmonic component, β , used in this figure is 15% of the fundamental. The peak synthesised voltage is approximately equal to the minimum

available voltage. It can be seen that phase-shifted 3rd harmonic voltage injection can provide a sustainable operation by increasing this margin. The following simulation results are obtained using 3rd harmonic voltage with phase shift of 15° degrees. The available voltage, ΣV_{cap} , is calculated using equation (4.33).

Positive limb voltage waveforms in the analysis above can be illustrated by Figure 4.8. Unlike Figure 4.4, a non-ideal MMC with a reasonably sized capacitor is considered here in order to introduce the impact of the capacitor ripple on the available voltage.

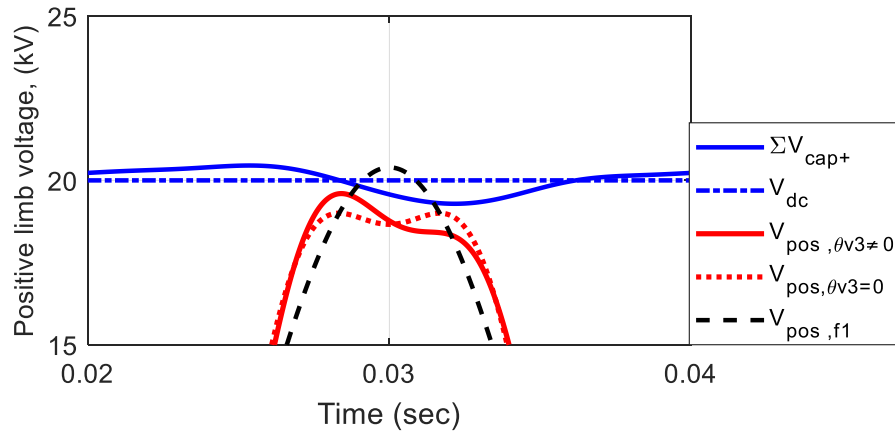


Figure 4.7: The impact of phase-shifted 3rd harmonic injection ($\theta_{v3} = 15^\circ$) on limb voltage ($S=+0.94-j0.33$ p.u.).

Only the positive limb waveforms are shown, because the negative limb voltage is equal to the positive limb voltage of the previous half cycle, $v_{neg}(t) = v_{pos}(t - \frac{T}{2})$. This also is applied with the limb currents, power and energy. An operating point with m_i 0.9 is considered and the results below are obtained based on Table 3.2 with 3rd harmonic component of 15% of the fundamental.

Figure 4.8, 4.8, 4.9 and 4.10 illustrate the positive limb voltages including, the available voltage and the limb synthesised voltage where both the fundamental and 3rd harmonic components are considered. All the plots have been generated using the mathematical models given by the analysis and then validated by the average PLECS model and by the switching model presented in Chapter 7. It can be seen that using 15% 3rd harmonic can lead to a reduction in the peak synthesised voltage while the available voltage is not affected even for different P and Q operating points. The impact of the 3rd harmonic injection on the available voltage cannot be seen clearly due to the insignificant impact

on the energy variation itself. This can be seen considering Figure 4.14 and Figure 4.17 particularly as a small 3rd harmonic voltage injection of 15% of fundamental is used.

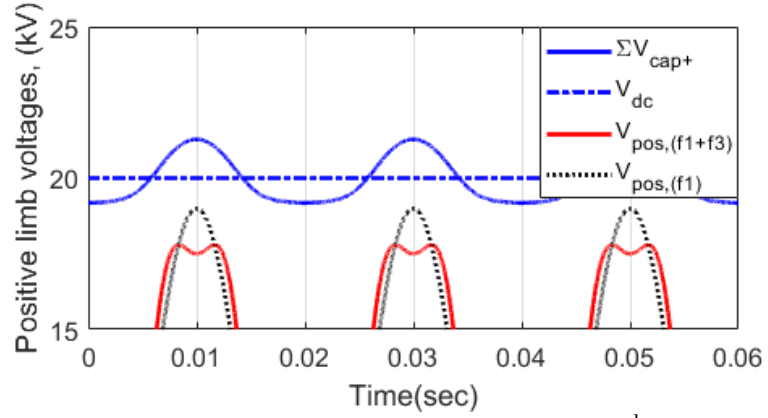


Figure 4.8: Positive limb voltages for fundamental and 3rd harmonic injection operation, ($\theta_{v3} = 0^\circ$), for $P=0$ and $Q=+1.0$ p.u.

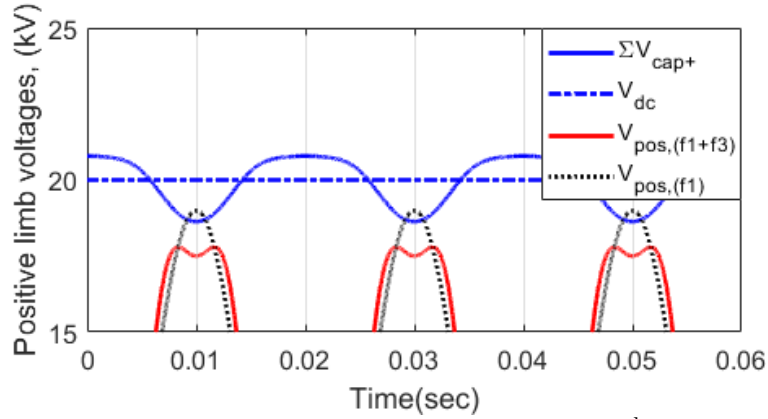


Figure 4.9: Positive limb voltages for fundamental and 3rd harmonic injection operation ($\theta_{v3} = 0^\circ$) for $P=0$ and $Q=-1.0$ p.u.

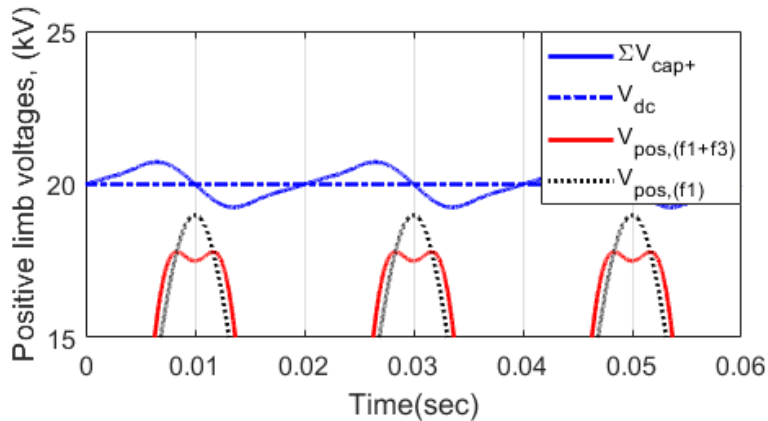


Figure 4.10: Positive limb voltages for fundamental and 3rd harmonic injection operation ($\theta_{v3} = 0^\circ$) for $P=+1.0$ p.u and $Q=0$ (inverting mode).

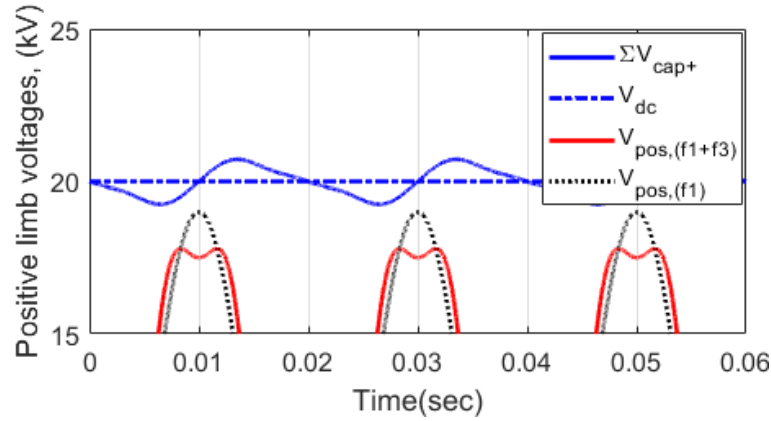


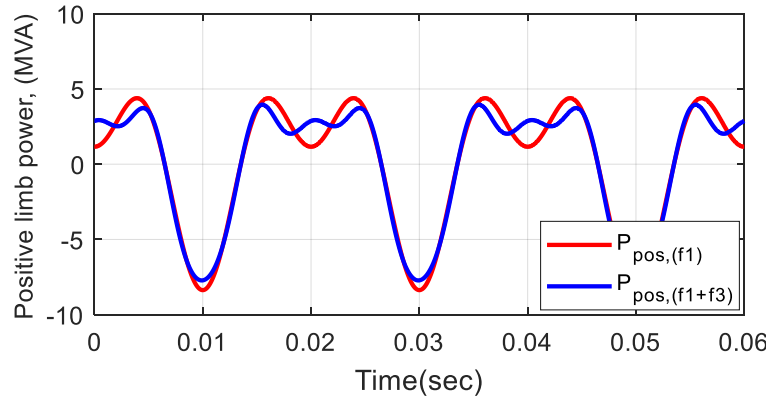
Figure 4.11: Positive limb voltages for fundamental and 3rd harmonic injection operation ($\theta_{v3} = 0^\circ$) for $P=-1.0$ p.u and $Q=0$ (rectifying mode).

The instantaneous power and energy variation waveforms of the positive limb are shown from Figure 4.12 to Figure 4.17 where different operating points of P and Q are considered including $S=+P+j0$ and $S=0+jQ$. A comparison with and without 3rd harmonic injection is presented considering both classical and phase-shifted 3rd harmonic voltage.

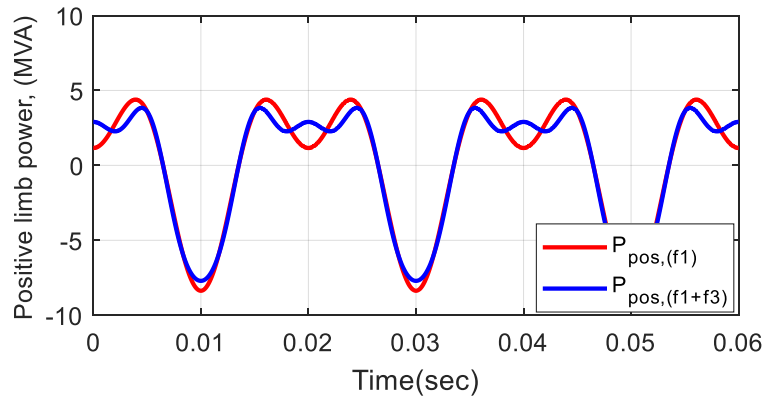
The red waveforms correspond to the operation with fundamental operation while the blue waveforms are due to using 3rd harmonic injection, classical or phase-shifted as will be indicated. It can be seen that, in general, due to 3rd harmonic voltage injection, peak-to-peak power is reduced at unity power factor, $\pm P$, while it increases with purely reactive power operation, $\pm Q$. However, the peak-to-peak energy (deviation) reduction is observed particularly with 3rd harmonic voltage injection and with active power operation. Therefore, the impact of 3rd harmonic voltage injection on energy deviation reduction can be neglected as the operation moves towards pure reactive power. Therefore the energy deviation reduction depends on the power angle range, ϕ .

Figure 4.12, 4.12 and 4.13 illustrates the positive limb instantaneous power and energy variation when operating with unity power factor. While in Figure 4.15, 4.15 and 4.16, the consideration for pure capacitive operation, $\pm Q$, where only $+Q$ is considered. The following simulation results are obtained using 15% of 3rd harmonic injection with phase shift of 15° .

With regards to classical and phase-shifted 3rd harmonic, and from Figure 4.12 to Figure 4.17, it is observed that the phase-shifted 3rd harmonic injection influence on energy deviation reduction can be neglected as illustrated by Figure 4.13, 4.13, 4.15 and 4.16.



(a) Positive limb power for fundamantal and phase-shifted 3rd harmonic ($\theta_{v3} = 15^\circ$).



(b) Positive limb power for fundamantal and clasical 3rd harmonic.

Figure 4.12: Positive limb power at +P; (a) phase-shifted 3rd harmonic ($\theta_{v3} = 15^\circ$), (b) classical 3rd harmonic.

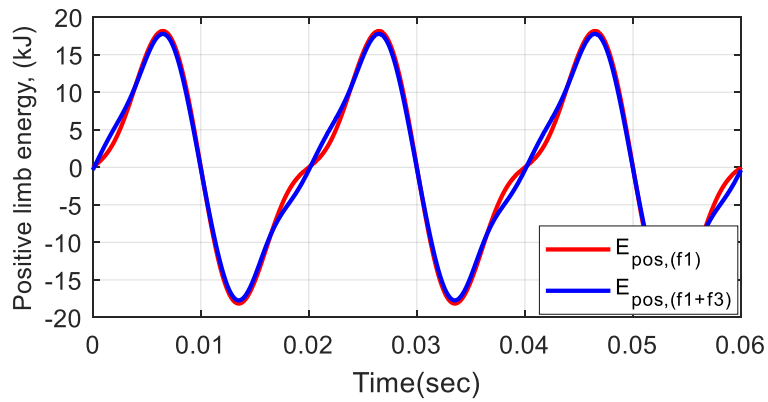


Figure 4.13: Positive limb energy variation at +P for fundamental and phase-shifted 3rd harmonic ($\theta_{v3} = 15^\circ$).

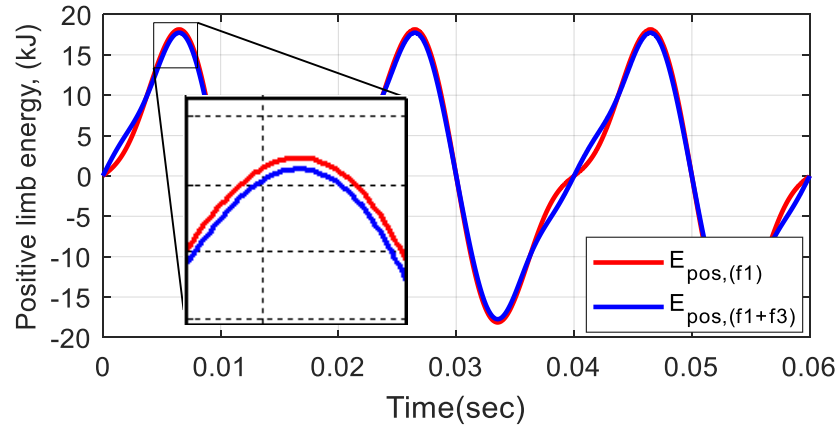
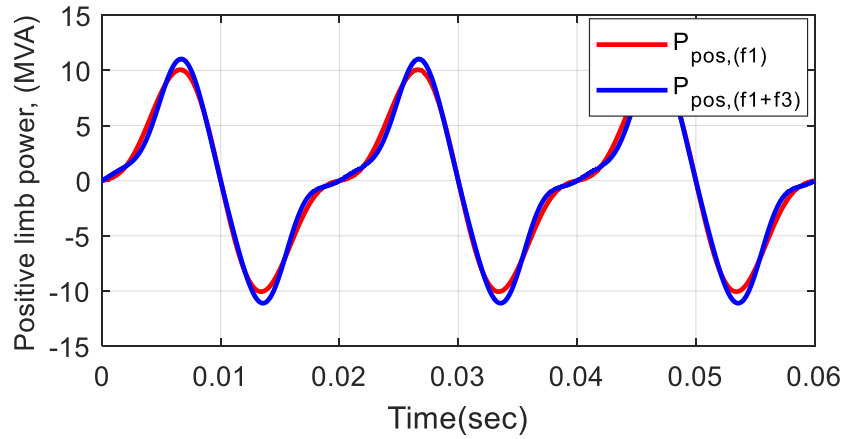
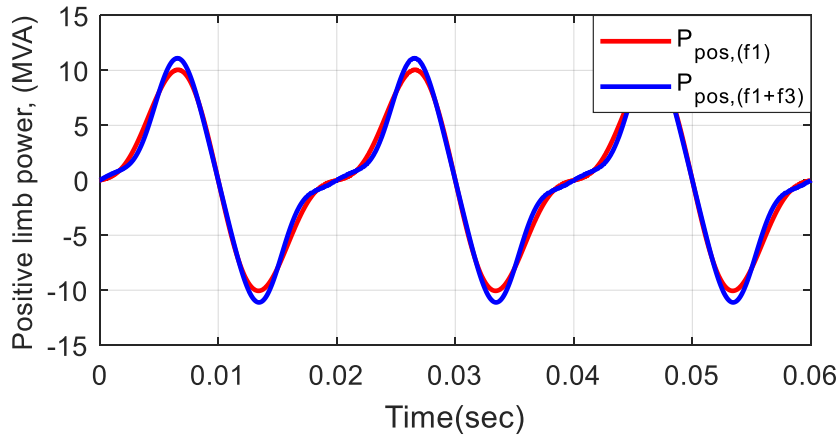


Figure 4.14: Positive limb energy variation at +P for fundamental and classical 3rd harmonic.



(a) Positive limb power for fundamental and phase-shifted 3rd harmonic ($\theta_{v3} = 15^\circ$).



(b) Positive limb power for fundamental and classical 3rd harmonic.

Figure 4.15: Positive limb power at +Q; (a) phase-shifted 3rd harmonic ($\theta_{v3} = 15^\circ$), (b) classical 3rd harmonic.

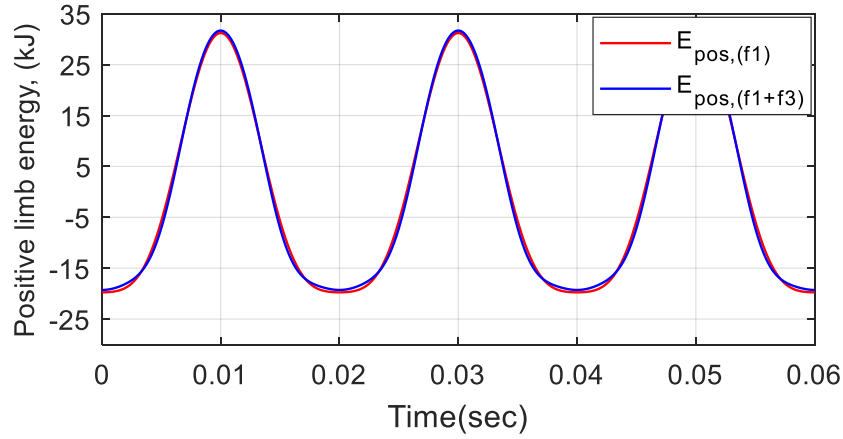


Figure 4.16: Positive limb energy variation at +Q for fundamental and phase-shifted 3rd harmonic, ($\theta_{v3} = 15^\circ$).

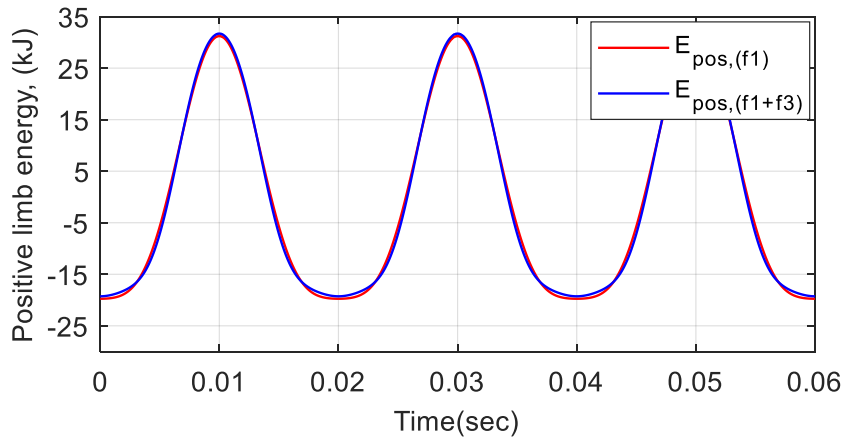
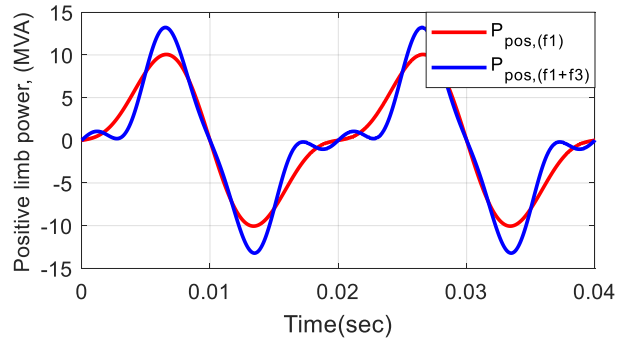


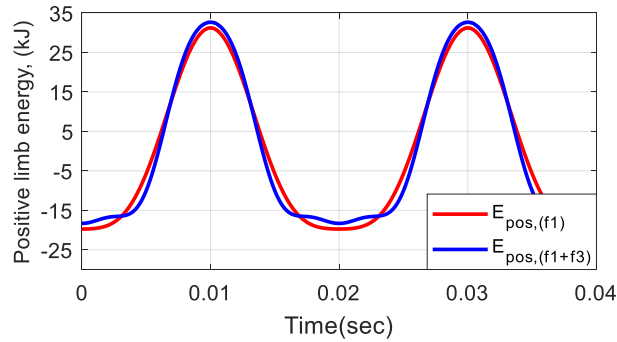
Figure 4.17: Positive limb energy variation at +Q for fundamental and classical 3rd harmonic voltage injection.

Figure 4.18 investigates the impact of injecting higher 3rd harmonic component, e.g. 50% instead of 15%. It is observed that the peak-to-peak power is increased. However, comparing Figure 4.18.b ($\beta = 50\%$) to Figure 4.16 or Figure 4.17 ($\beta = 15\%$), energy deviation at $\pm Q$ is still the same for any value of 3rd harmonic voltage injection however, using higher 3rd harmonic injection can lead to increasing limb energy average. . Peak-to-peak power is reduced due to 3rd harmonic voltage injection when operating with only $\pm P$ as shown by Figure 4.12. However, it is increased when the operating point is only $\pm Q$ as shown by Figure 4.15. It is observed that this depends on the 3rd harmonic

component percentage of the fundamental which increases by increasing the 3rd harmonic percentage while energy deviation is decreased as illustrated in Figure 4.18.



(a) Positive limb power.



(b) Positive limb energy variation.

Figure 4.18: Positive limb power and energy deviation at +Q for the classical 3rd harmonic, β , 50% of the fundamental.

The maximum and minimum energy variation and energy deviation is illustrated in Figure 4.19. The impact of the 3rd harmonic on energy deviation is further examined and presented in Figure 4.20 and Figure 4.21. Figure 4.20 shows a zoomed version of the red trace in Figure 4.19, highlighting the effect of different β values. It can be seen that the energy deviation reduction is a maximum for purely resistive operation and it is a minimum as the operation moves towards inductive or capacitive operation, -Q and +Q respectively. It is observed that energy deviation reduction is zero when power factor angle is $\pm 30^\circ$ degrees (leading and lagging) or/and pure capacitive or inductive operation, $\pm Q$. The worst case scenario of energy deviation is when at pure reactive power, $\pm Q$ as shown by Figure 4.20 and Figure 4.21. Again, m_i is considered to be 0.9 whilst the other parameters are based on Table 3.2.

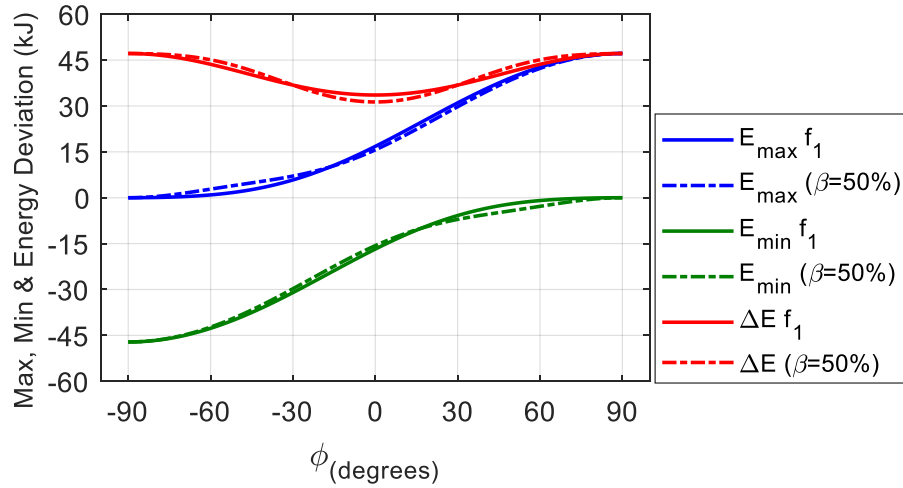


Figure 4.19: Maximum and minimum energy and energy deviation for $\beta=50\%$ of the fundamental.

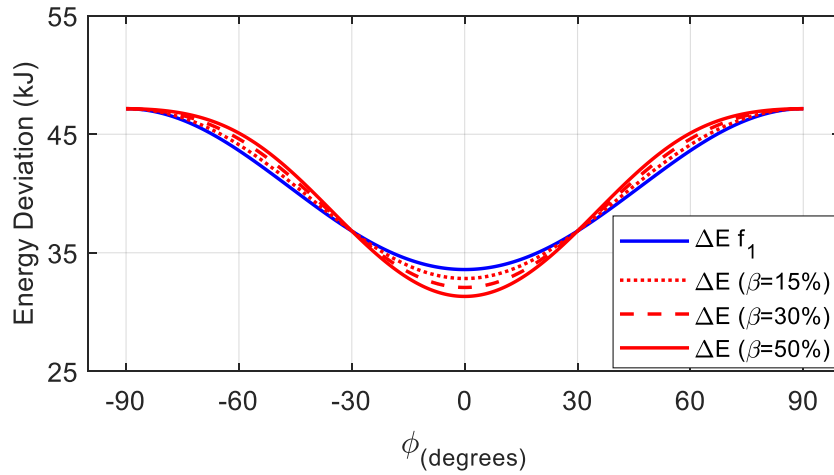


Figure 4.20: Energy deviation with power angle due to fundamental and 3rd harmonic injection operation.

It can be seen that as 3rd harmonic injection, β , increases from 15% to 50%, energy deviation reduces compared to the fundamental operation as illustrated in Figure 4.20 and Figure 4.21. Furthermore, this reduction occurs only between the power angle region of $(-30^\circ \leq \phi \leq +30^\circ)$ with a significant reduction at unity power factor. However, the concept does not yield significant improvement across all operating regions as can be observed from Figure 3.20.

Considering Table 3.2, for the given region of operation $-90^\circ \leq \phi \leq +90^\circ$, four significant operation points can be identified include -90° , -30° , $+30^\circ$ and $+90^\circ$. This idea is further illustrated by Figure 4.22. It is also observed that energy deviation can be reduced for a higher modulation index. The resulting M curve is presented in Figure

4.22 where a wider range of power factor, $-180^\circ \leq \phi \leq 180^\circ$, and modulation index is considered.

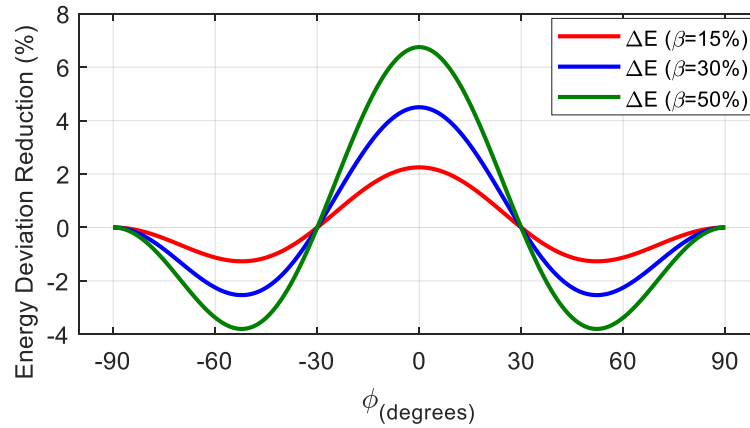


Figure 4.21: Energy deviation reduction with power factor angle due to fundamental and 3rd harmonic injection operation.

From the previous analysis it can be seen that, 3rd harmonic voltage injection can lead to reduced energy deviation. This reduction increases by increasing the 3rd harmonic component. However, not all the operating points of P and Q experience this reduction. Considering the PQ curve in Figure 4.21 and Figure 4.22, six operating points can be identified where 3rd harmonic voltage injection has no effect on energy reduction including $(\pm 90^\circ, \pm 30^\circ$ and $\pm 150^\circ)$. However, the highest energy deviation when the MMC operates with purely reactive power, $\pm Q$. It also can be seen that, the maximum energy deviation reduction occurs when the MMC operates with a pure active power, $\pm P$. The 3rd harmonic voltage injection is set at 50%, zero phase. Considering the energy deviation, it is noted that there is no impact of the phase-shifted on the energy deviation. Therefore, only classical 3rd harmonic injection is considered.

Using high value of 3rd harmonic voltage injection (50%) can lead to 7% of energy deviation reduction. This reduction is relatively small compared with the value of 3rd harmonic injection. Therefore, this is not considered practical, since the figure (7%) occurs only when operating at unity power factor, see Figure 4.21. Moreover, using high value of 3rd harmonic leads to increased peak of the synthesised voltage, potentially leading to overmodulation. This is discussed in more details in Chapter 6, Figure 6.1. For the reasons above, the 3rd harmonic component injection has limitations and for this example of (50%) is considered as upper limit for the injection.

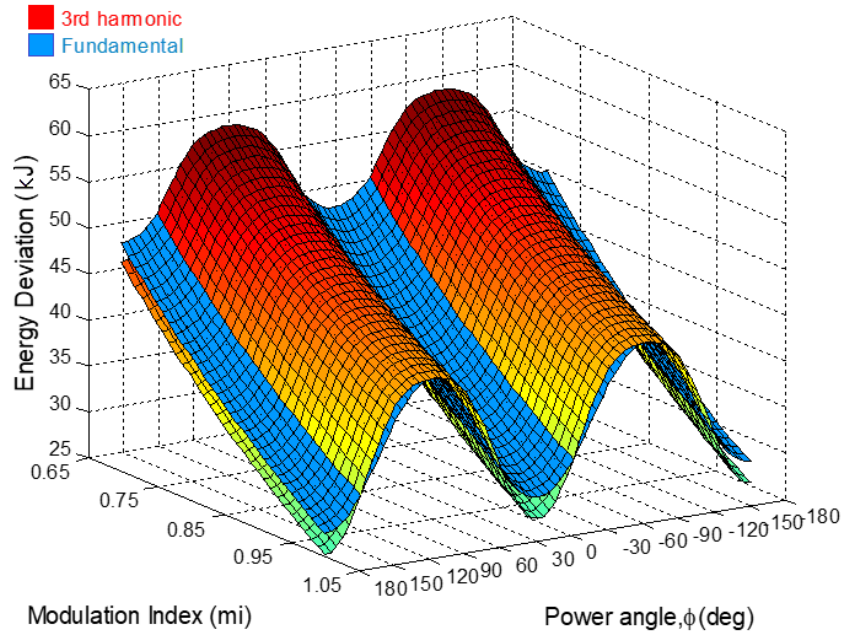


Figure 4.22: M curve of Energy deviation and modulation index and power factor for both classical and 3rd harmonic ($\theta_{v3} = 0^\circ$ and $\beta = 50\%$).

The analysis presented in this section shows the impact of 3rd harmonic voltage injection on the limb voltage instantaneous power and energy deviation. A comparison was presented between fundamental and 3rd harmonic voltage injection that shows that 3rd harmonic voltage injection leads to insignificant energy deviation reduction for theoretical 3rd harmonic voltage injection of 50%. This effect can be neglected when the normal practical injection of 15% is considered. However, the synthesised voltage can be reduced.

The next analysis considers 3rd harmonic current injection and its impact on energy storage requirement. Situations involving both 3rd harmonic voltage and current are also considered.

4.5 Third Harmonic Current Injection (Triplen Current, I_{C3})

In the previous section, the impact of 3rd harmonic voltage injection above was discussed including, classical and phase-shifted. In this section, the impact of the triplen current injection is presented. The triplen current injection needs a path for the current to follow. In general this can be achieved as discussed in section 4.3 and illustrated in Figure 4.2.a.

Third harmonic current injection can be implemented by two methods. First, an external circuit can be used as illustrated in Figure 4.23 presented by the dashed line. This was first presented by [18] for an AAC converter. However, the same principle can be used

for the MMC. The second option is using 3rd harmonic voltage injection to produce the required 3rd harmonic current injection as shown by Figure 4.24. This, however, might require neutral impedance connected to the neutral point of the transformer. Moreover, with regards to the neutral impedance, variable impedance can be used to limit the current. The 3rd harmonic current can be produced as illustrated by Figure 4.25 below which depends on the 3rd harmonic magnitude and also the impedance path.

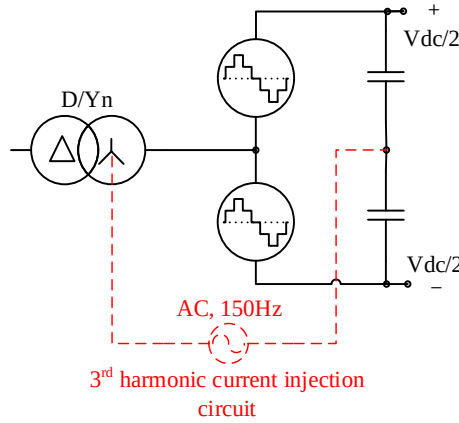


Figure 4.23: 3rd harmonic current injection using an auxiliary circuit for an MMC [18].

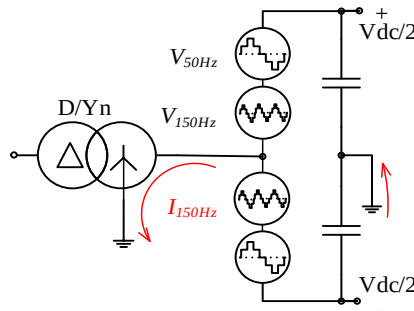


Figure 4.24: 3rd harmonic current injection providing a neutral path by DYn transformer, Solution 1.

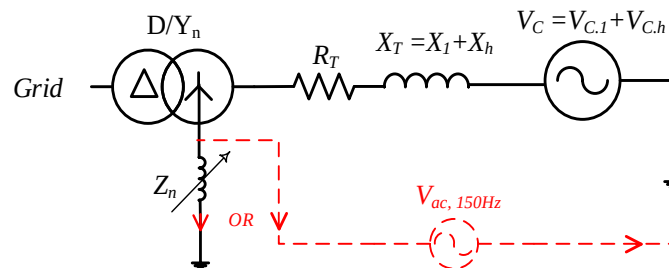


Figure 4.25: 3rd harmonic current injection providing a neutral path by DYn transformer, Solution 2.

The zero sequence current through the Y connection, however, needs to be controlled as the zero sequence impedance is significantly smaller. The small zero sequence impedance can cause a high 3rd harmonic current even for small 3rd harmonic voltage injection. Therefore, a reasonably high zero sequence impedance, Z_n , must be used. In fact, the zero sequence current (3rd harmonic current) must be maintained at permissible level. The 3rd harmonic current phasor can be written as;

$$I_{C3} = \left(\frac{V_{C3}}{Z_T + \frac{1}{2}Z_{limb} + Z_n} \right) \quad (4.43)$$

where V_{C3} is the 3rd harmonic voltage component and Z_T and Z_{limb} are transformer and limb impedance respectively.

4.5.1 Impact of only Triplen Current Injection on Instantaneous Power and Energy

The periodic voltage and current of the positive and negative limbs can be expressed by (4.44) through (4.47) where the 3rd harmonic current component is denoted by $\hat{I}_{C3}$ with phase shift of θ_{i3} . Assuming that the arrangement of Figure 4.23 is considered, the 3rd harmonic component of the limb voltages $\hat{V}_{C3}$ is eliminated.

$$v_{pos.a}(t) = \frac{V_{dc}}{2} - \hat{V}_{C1} \cos(\omega_0 t - \theta_{v1}) \quad (4.44)$$

$$i_{pos.a}(t) = \frac{I_{dc}}{3} + \frac{\hat{I}_{C1}}{2} \cos(\omega_0 t - \theta_{i1}) - \frac{\hat{I}_{C3}}{2} \cos(3\omega_0 t - \theta_{i3}) \quad (4.45)$$

$$v_{neg.a}(t) = \frac{V_{dc}}{2} + \hat{V}_{C1} \cos(\omega_0 t - \theta_{v1}) \quad (4.46)$$

$$i_{neg.a}(t) = \frac{I_{dc}}{3} - \frac{\hat{I}_{C1}}{2} \cos(\omega_0 t - \theta_{i1}) + \frac{\hat{I}_{C3}}{2} \cos(3\omega_0 t - \theta_{i3}) \quad (4.47)$$

Due to the symmetry, the instantaneous power of positive and negative limb is given by;

$$p_{pos.a}(t) = v_{pos.a}(t) \cdot i_{pos.a}(t) \quad (4.48)$$

$$p_{pos.a}(t) = \left[\begin{aligned} &\left(\frac{V_{dc}I_{dc}}{6} - \frac{\hat{V}_{c1}\hat{I}_{c1}}{4} \cos(\theta_{i1} - \theta_{v1}) \right) \\ &+ \left(\frac{V_{dc}\hat{I}_{c1}}{4} \cos(\omega_0 t - \theta_{i1}) - \frac{\hat{V}_{c1}I_{dc}}{3} \cos(\omega_0 t - \theta_{v1}) \right) \\ &\quad - \frac{\hat{V}_{c1}\hat{I}_{c1}}{4} \cos(2\omega_0 t - \theta_{v1} - \theta_{i1}) \\ &- \left(\frac{V_{dc}\hat{I}_{c3}}{4} \cos(3\omega_0 t - \theta_{i3}) - \frac{\hat{V}_{c1}\hat{I}_{c3}}{4} \cos(-2\omega_0 t + \theta_{i3} - \theta_{v1}) \right) \\ &\quad - \frac{\hat{V}_{c1}\hat{I}_{c3}}{4} \cos(4\omega_0 t - \theta_{v1} - \theta_{i3}) \end{aligned} \right] \quad (4.49)$$

The average power for one periodic cycle is given by (4.50) and (4.51). As it can be seen, average power is only produced by the product of the fundamental terms of the current and voltage. Assuming the phase angle of the fundamental voltage is zero then the phase angle of the current will be the power factor angle, ϕ . This will yield;

$$\bar{P}_{pos.a} = \left(\frac{V_{dc}I_{dc}}{6} - \frac{\hat{V}_{c1}\hat{I}_{c1}}{4} \cos(\phi) \right) \quad (4.50)$$

$$\bar{P}_{neg.a} = \left(\frac{V_{dc}I_{dc}}{6} - \frac{\hat{V}_{c1}\hat{I}_{c1}}{4} \cos(\phi) \right) \quad (4.51)$$

Considering the last two equations, total average power is not affected by the triplen current injection. However, the triplen current can increase power loss and magnetic stress of the converter transformer as the total RMS line current is increased [73].

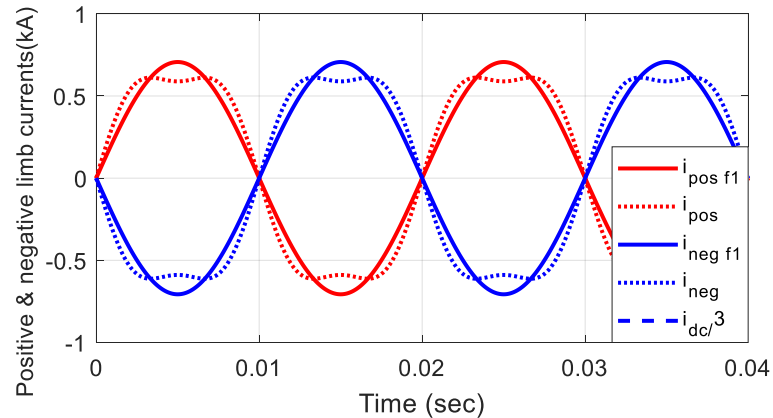
Using third harmonic current injection can also have an impact on the peak alternating current. Injecting the third harmonic current in phase with the fundamental current reduces the peak line current. However, peak current can be increased if the two currents are out of phase by 180° degrees as illustrated by Figure 4.26. This phenomenon is the key factor to determine the power pulsation and therefore the energy variation which will be discussed later. The simulation results are based on the rated current of the MMC parameters given in Table 3.2 with 15% of the 3rd harmonic current with respect to fundamental. As before, setting (4.50) and (4.51) to zero yields (4.52) and (4.53).

$$V_{dc}I_{dc} = \frac{3}{2} \hat{V}_{c1}\hat{I}_{c1} \cos(\phi) \quad (4.52)$$

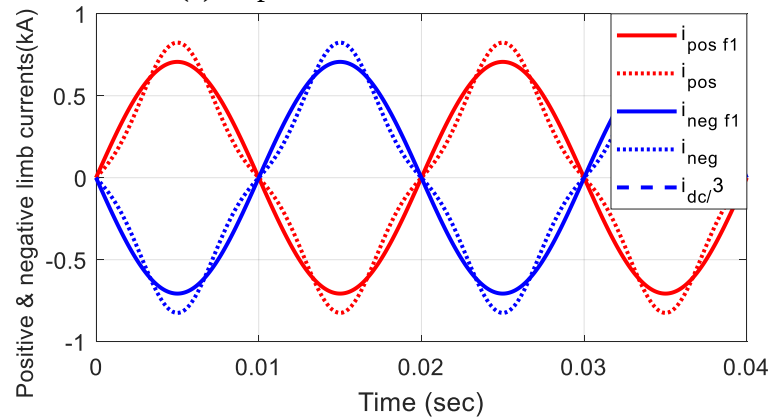
$$I_{dc} = \frac{3}{4} m_i \hat{I}_{c1} \cos(\phi) \quad (4.53)$$

As presented by equation (4.52), $P_{dc} = P_{ac}$, where; $P_{dc} = \frac{V_{dc}I_{dc}}{6}$ and $P_{ac} = \frac{V_{c1}I_{c1}}{4} \cos(\phi)$. Therefore, the term $\frac{V_{dc}I_{dc}}{6} - \frac{\hat{V}_{c1}\hat{I}_{c1}}{4} \cos(\phi) = 0$. It is noted that, this explains the principle operation of the MMC with regards to the energy exchanged from the AC and DC side. Following integration, positive limb energy variation then is expressed by (4.54).

$$e_{pos.a}(t) = \begin{bmatrix} \left(\frac{V_{dc}\hat{I}_{c1}}{4\omega_0} \sin(\omega_0 t - \theta_{i1}) - \frac{\hat{V}_{c1}I_{dc}}{3\omega_0} \sin(\omega_0 t - \theta_{v1}) \right) - \frac{\hat{V}_{c1}\hat{I}_{c1}}{8\omega_0} \sin(2\omega_0 t - \theta_{v1} - \theta_{i1}) \\ - \left(\frac{V_{dc}\hat{I}_{c3}}{12\omega_0} \sin(3\omega_0 t - \theta_{i3}) + \frac{\hat{V}_{c1}\hat{I}_{c3}}{8\omega_0} \sin(-2\omega_0 t + \theta_{i3} - \theta_{v1}) \right) - \frac{\hat{V}_{c1}\hat{I}_{c3}}{16\omega_0} \sin(4\omega_0 t - \theta_{v1} - \theta_{i3}) \end{bmatrix} \quad (4.54)$$



(a) in-phase 3rd harmonic current.



(b) phase-shifted 3rd harmonic current.

Figure 4.26: The impact of in-phase and 180° phase-shifted 3rd harmonic injection on peak currents.

Using (4.25) (4.27) (4.28) and (4.53) and substituting them into the previous equations of instantaneous power and energy variation (4.49) and (4.54), respectively, this yields:

$$p_{pos.a}(t) = \frac{|S|}{6} \left[\begin{array}{l} + \left(\frac{2}{m_i} \cos(\omega_0 t - \phi) - m_i \cos(\phi) \cos(\omega_0 t) \right) \\ - \cos(2\omega_0 t - \phi) \\ + \gamma \left(\frac{2}{m_i} \cos(3\omega_0 t - \phi_3) - \cos(2\omega_0 t - \phi_3) \right) \\ - \cos(4\omega_0 t - \phi_3) \end{array} \right] \quad (4.55)$$

$$e_{pos.a}(t) = \frac{|S|}{6\omega_0} \left[\begin{array}{l} + \left(\frac{2}{m_i} \sin(\omega_0 t - \phi) - m_i \cos(\phi) \sin(\omega_0 t) \right) \\ - \frac{1}{2} \sin(2\omega_0 t - \phi) \\ + \gamma \left(\frac{2}{3m_i} \sin(3\omega_0 t - \phi_3) - \frac{1}{2} \sin(2\omega_0 t - \phi_3) \right) \\ - \frac{1}{4} \sin(4\omega_0 t - \phi_3) \end{array} \right] \quad (4.56)$$

Where $\phi_3 = \theta_{i3}$ and the 3rd harmonic component is expressed by;

$$\gamma = \frac{\hat{I}_{c3}}{\hat{I}_{c1}} \quad (4.57)$$

Considering the analysis above, it can be seen that the 3rd harmonic current has an impact on only the power pulsation and therefore the energy variation, whilst the average power is not affected. Moreover, the power pulsation can therefore affect the SM capacitor voltages. This effect can be increased by increasing the 3rd harmonic current.

However, the RMS line current is increased where high additional current can limit the AC and limb current and cause extra power stress on the power electronic devices. Therefore, this needs to be considered.

4.5.2 Impact of Third Harmonic Voltage and Current Injection on Instantaneous Power and Energy

In this subsection, the impact of both, third harmonic voltage and current is considered. This sample case can be considered as a general case of injecting 3rd harmonic. It is assumed that the periodic voltage and current of the positive limb are given by equations

(4.58) and (4.59). The amplitude of the 3rd harmonic voltage and current components are expressed by $\hat{V}_{C3}$ and $\hat{I}_{C3}$ with their phase angle of θ_{v3} and θ_{i3} , respectively. The negative limb periodic voltage and current are given by (4.60) and (4.61).

$$v_{pos.a}(t) = \frac{V_{dc}}{2} - \hat{V}_{C1} \cos(\omega_0 t - \theta_{v1}) + \hat{V}_{C3} \cos(3\omega_0 t - \theta_{v3}) \quad (4.58)$$

$$i_{pos.a}(t) = \frac{I_{dc}}{3} + \frac{\hat{I}_{C1}}{2} \cos(\omega_0 t - \theta_{i1}) - \frac{\hat{I}_{C3}}{2} \cos(3\omega_0 t - \theta_{i3}) \quad (4.59)$$

$$v_{neg.a}(t) = \frac{V_{dc}}{2} + \hat{V}_{C1} \cos(\omega_0 t - \theta_{v1}) - \hat{V}_{C3} \cos(3\omega_0 t - \theta_{v3}) \quad (4.60)$$

$$i_{neg.a}(t) = \frac{I_{dc}}{3} - \frac{\hat{I}_{C1}}{2} \cos(\omega_0 t - \theta_{i1}) + \frac{\hat{I}_{C3}}{2} \cos(3\omega_0 t - \theta_{i3}) \quad (4.61)$$

By substituting limb periodic voltage and current (4.58) and (4.59) into the instantaneous power equation (4.62), the instantaneous power of positive limb is given by (4.63).

$$p_{pos.a}(t) = v_{pos.a}(t) \cdot i_{pos.a}(t) \quad (4.62)$$

$$p_{pos.a}(t) = \left[\begin{aligned} & \left(\frac{V_{dc}I_{dc}}{6} - \frac{\hat{V}_{C1}\hat{I}_{C1}}{4} \cos(\theta_{i1} - \theta_{v1}) - \frac{\hat{V}_{C3}\hat{I}_{C3}}{4} \cos(+\theta_{i3} - \theta_{v3}) \right) \\ & + \left(\frac{V_{dc}\hat{I}_{C1}}{4} \cos(\omega_0 t - \theta_{i1}) - \frac{\hat{V}_{C1}I_{dc}}{3} \cos(\omega_0 t - \theta_{v1}) \right) \\ & \quad - \frac{\hat{V}_{C1}\hat{I}_{C1}}{4} \cos(2\omega_0 t - \theta_{v1} - \theta_{i1}) \\ & - \left(\frac{V_{dc}\hat{I}_{C3}}{4} \cos(3\omega_0 t - \theta_{i3}) - \frac{\hat{V}_{C1}\hat{I}_{C3}}{4} \cos(-2\omega_0 t + \theta_{i3} - \theta_{v1}) \right) \\ & \quad - \frac{\hat{V}_{C1}\hat{I}_{C3}}{4} \cos(4\omega_0 t - \theta_{v1} - \theta_{i3}) \\ & + \left(\frac{\hat{V}_{C3}I_{dc}}{3} \cos(3\omega_0 t - \theta_{v3}) + \frac{\hat{V}_{C3}\hat{I}_{C1}}{4} \cos(2\omega_0 t + \theta_{i1} - \theta_{v3}) \right) \\ & \quad + \frac{\hat{V}_{C3}\hat{I}_{C1}}{4} \cos(4\omega_0 t - \theta_{i1} - \theta_{v3}) \\ & \quad - \left(\frac{\hat{V}_{C3}\hat{I}_{C3}}{4} \cos(6\omega_0 t - \theta_{i3} - \theta_{v3}) \right) \end{aligned} \right] \quad (4.63)$$

The average power expression is the integration of the instantaneous power in (4.63) for one periodic cycle. As can be seen, the terms produced by the product of the same frequency between the first brackets will survive. The others however integrate to zero.

Assuming that the phase angle of the fundamental voltage is zero, then the phase angle of the current will be the power angle, ϕ .

$$\bar{P}_{pos.a} = \left(\frac{V_{dc}I_{dc}}{6} - \frac{\hat{V}_{c1}\hat{I}_{c1}}{4} \cos(\phi) - \frac{\hat{V}_{c3}\hat{I}_{c3}}{4} \cos(\theta_{i3} - \theta_{v3}) \right) \quad (4.64)$$

Similarly;

$$\bar{P}_{neg.a} = \left(\frac{V_{dc}I_{dc}}{6} - \frac{\hat{V}_{c1}\hat{I}_{c1}}{4} \cos(\phi) - \frac{\hat{V}_{c3}\hat{I}_{c3}}{4} \cos(\theta_{i3} - \theta_{v3}) \right) \quad (4.65)$$

For one periodic cycle, the average power of the limbs must be zero in order to obtain capacitor charge balancing. Considering equation (4.64) and (4.65), it is a particularly important result because it states that where there is interaction between the periodic voltage of the third harmonic and its corresponding periodic current, an average power be seen at the output of the converter terminals. This is presented by the third term of the equation. The first three terms denote the total average power which is the sum of the average powers obtained from the interaction of the currents and voltages of the same frequency.

Equations (4.64) and (4.65) equal zero establishing the point where there can be power balance between the AC and DC sides for a lossless converter and where $\phi_3 = \theta_{i3} - \theta_{v3}$.

$$V_{dc}I_{dc} = \frac{3}{2} (\hat{V}_c \hat{I}_c \cos(\phi) + \hat{V}_{c3} \hat{I}_{c3} \cos(\phi_3)) \quad (4.66)$$

$$P_{dc} = P_{ac.f_1} + P_{ac.f_3} \quad (4.67)$$

Comparing (3.20) which has the same result as (4.52) and (4.66), the relationship between the direct and the alternating current given by (3.22) is not valid in this case. According to (4.66), the total average power seen at the converter terminals is the sum of the average powers associated with the fundamental and third harmonic voltages and currents. This can be further explained by (4.67).

In the case of Figure 4.23, the power goes to/comes from the additional AC source. It can be positive or negative. While in the case of Figure 4.25 it must be the i^2R loss in

the network parts. In general, this power is not absorbed by the AC system and therefore it can only be a loss. [16] reported that due to the extra power losses, the impact of triplen current was not considered further in their investigation.

The relationship between the direct and alternating currents can be given by;

$$I_{dc} = \frac{3}{4} m_i \hat{I}_{c1} \{\cos(\phi) + \beta \gamma \cos(\phi_3)\} \quad (4.68)$$

The energy variation then can be given using the power integral. Due to the symmetry, the positive energy variations are given by (4.69).

$$e_{pos.a}(t) = \left[\begin{aligned} & \left(\frac{V_{dc} I_{dc}}{6} - \frac{\hat{V}_{c1} \hat{I}_{c1}}{4} \cos(\theta_{i1} - \theta_{v1}) - \frac{\hat{V}_{c3} \hat{I}_{c3}}{4} \cos(+\theta_{i3} - \theta_{v3}) \right) t \\ & + \left(\frac{V_{dc} \hat{I}_{c1}}{4\omega_0} \sin(\omega_0 t - \theta_{i1}) - \frac{\hat{V}_{c1} I_{dc}}{3\omega_0} \sin(\omega_0 t - \theta_{v1}) \right) \\ & - \frac{\hat{V}_{c1} \hat{I}_{c1}}{8\omega_0} \sin(2\omega_0 t - \theta_{v1} - \theta_{i1}) \\ & - \left(\frac{V_{dc} \hat{I}_{c3}}{12\omega_0} \sin(3\omega_0 t - \theta_{i3}) + \frac{\hat{V}_{c1} \hat{I}_{c3}}{8\omega_0} \sin(-2\omega_0 t + \theta_{i3} - \theta_{v1}) \right) \\ & - \frac{\hat{V}_{c1} \hat{I}_{c3}}{16\omega_0} \sin(4\omega_0 t - \theta_{v1} - \theta_{i3}) \\ & + \left(\frac{\hat{V}_{c3} I_{dc}}{9\omega_0} \sin(3\omega_0 t - \theta_{v3}) + \frac{\hat{V}_{c3} \hat{I}_{c1}}{8\omega_0} \sin(2\omega_0 t + \theta_{i1} - \theta_{v3}) \right) \\ & + \frac{\hat{V}_{c3} \hat{I}_{c1}}{16\omega_0} \sin(4\omega_0 t - \theta_{i1} - \theta_{v3}) \\ & - \left(\frac{\hat{V}_{c3} \hat{I}_{c3}}{24\omega_0} \sin(6\omega_0 t - \theta_{i3} - \theta_{v3}) \right) \end{aligned} \right] \quad (4.69)$$

The previous equations of instantaneous power and energy variation can be given by the rated apparent power and modulation index where the first terms in the equations above are cancelled due to power balance.

$$p_{pos.a}(t) = \frac{|S|}{6} \left[\begin{aligned} & + \left(\frac{2}{m_i} \cos(\omega_0 t - \phi) - m_i ((\cos(\phi) + \beta \gamma \cos(\phi_3))) \cos(\omega_0 t) \right) \\ & - \cos(2\omega_0 t - \phi) \\ & + \gamma \left(\frac{2}{m_i} \cos(3\omega_0 t - \phi_3) - \cos(2\omega_0 t - \phi_3) \right) \\ & - \cos(4\omega_0 t - \phi_{i3}) \\ & + \beta \left(m_i (\cos(\phi) + \beta \gamma \cos(\phi_3)) \cos(3\omega_0 t - \theta_{v3}) \right. \\ & \left. + \cos(2\omega_0 t + \phi - \theta_{v3}) + \cos(4\omega_0 t - \phi - \theta_{v3}) \right) \\ & \left. + (2\beta \gamma \cos(6\omega_0 t - \phi_3 - \theta_{v3})) \right] \quad (4.70) \end{aligned} \right]$$

$$e_{pos.a}(t) = \frac{|S|}{6\omega_0} \left[\begin{aligned} &+ \left(\frac{2}{m_i} \sin(\omega_0 t - \phi) - m_i ((\cos(\phi) + \beta\gamma \cos(\phi_3))) \sin(\omega_0 t) \right) \\ &\quad - \frac{1}{2} \sin(2\omega_0 t - \phi) \\ &+ \gamma \left(\frac{2}{3m_i} \sin(3\omega_0 t - \phi_3) - \frac{1}{2} \sin(2\omega_0 t - \phi_3) \right) \\ &\quad - \frac{1}{4} \sin(4\omega_0 t - \phi_3) \\ &+ \beta \left(\frac{m_i}{3} (\cos(\phi) + \beta\gamma \cos(\phi_3)) \sin(3\omega_0 t - \theta_{v3}) \right. \\ &\quad \left. + \frac{1}{2} \sin(2\omega_0 t + \phi - \theta_{v3}) + \frac{1}{4} \sin(4\omega_0 t - \phi - \theta_{v3}) \right) \\ &\quad \left. + \left(\frac{\beta\gamma}{3} \sin(6\omega_0 t - \phi_3 - \theta_{v3}) \right) \right] \quad (4.71) \end{aligned} \right.$$

The obtained mathematical equations provide a mathematical model which can be used to determine limb voltages, currents and the instantaneous power and energy variation. Compared with the previous case when only 3rd harmonic current injection is considered, it can be seen that the 3rd harmonic voltage and current injection has its impact not only on power pulsation and energy variation. But also the average power is affected as well. Moreover, the power pulsation can therefore affect the SM capacitor voltages. Therefore, the impact of 3rd harmonic current on energy deviation is considered in the next analysis.

The limb synthesised voltage can be reduced while the MMC terminal voltage is not affected. Moreover, the RMS line current is increased where the high current can limit the AC and limb current and cause extra power stress on the power electronic devices. Therefore, this needs to be considered.

In the next section, a comparison is presented considering the impact of both 3rd harmonic voltage and current injection on energy deviation.

4.5.3 The Impact of Third Harmonic Voltage and Current Injection on Peak-to-peak Energy

In this section, the impact of 3rd harmonic voltage and current injection on MMC energy deviation is discussed. In Section 3.5 and 4.4.3, energy deviation was determined directly using an algebraic solution. However, when the 3rd harmonic component of the limb current is considered, direct solution cannot be easily achieved. This becomes more difficult as the phase angle of the 3rd harmonic component is considered, ϕ_3 , as

expressed by the Equation (4.72) below. In these cases energy deviation is more conveniently determined numerically.

$$i_{pos.a}(t) = \frac{I_{dc}}{3} + \frac{\hat{I}_{c1}}{2} \cos(\omega_o t - \phi) + \frac{\hat{I}_{c3}}{2} \cos(3\omega_o t - \phi_3) \quad (4.72)$$

With regards to third harmonic current and its phase angle, peak-to-peak limb current can be increased when the third harmonic current is out of phase with the fundamental component as illustrated by Figure 4.26 in Section 4.5.1. However, peak-to-peak current and power pulsation are reduced when the third harmonic component and fundamental are in phase with each other.

The simulation results are obtained based on the MMC parameters given in Table 3.2. A comparison has been made considering different operating conditions. It can be seen that classical and phase-shifted 3rd harmonic current can be used here, $\phi_{i3} = 0^\circ$ and 180° as shown by Figure 4.27.

Figure 4.28 and Figure 4.29 illustrate the impact of 3rd harmonic current injection on peak-to-peak power and energy variation for different operating points of P and Q including $+P$ and $+Q$, whilst $|S|=1$ p.u. The blue line represents the positive limb power and energy variation for the fundamental component. The black line is based on the 3rd harmonic voltage injection while dashed ones based on both 3rd harmonic voltage and current injection. It is noted that modulation index of 0.9 is used while 3rd harmonic voltage, β , and current injection, γ , are 15% and 15% of the fundamental respectively.

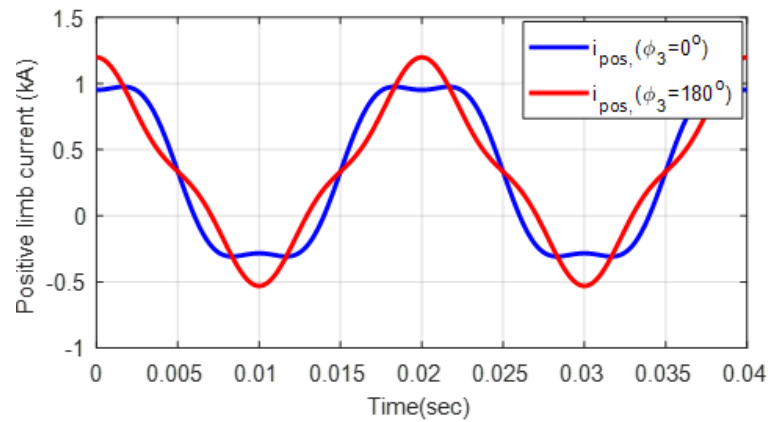


Figure 4.27: Positive limb current considering classical and phase-shifted 3rd harmonic current injection.

Considering P and Q curve presented by Figure 3.3 and the power conversion presented in the Appendix A and used in this thesis, the next simulation results were obtained based on the variation of the operating points of P and Q, whilst $|S|$ remains constant. The considered quadrants are Q1 and Q2 where +P and -Q and +Q are considered, thus $-90^\circ \leq \phi \leq 90^\circ$. The response in other quadrants is the same considering the same assumptions made above including m_i , β and γ .

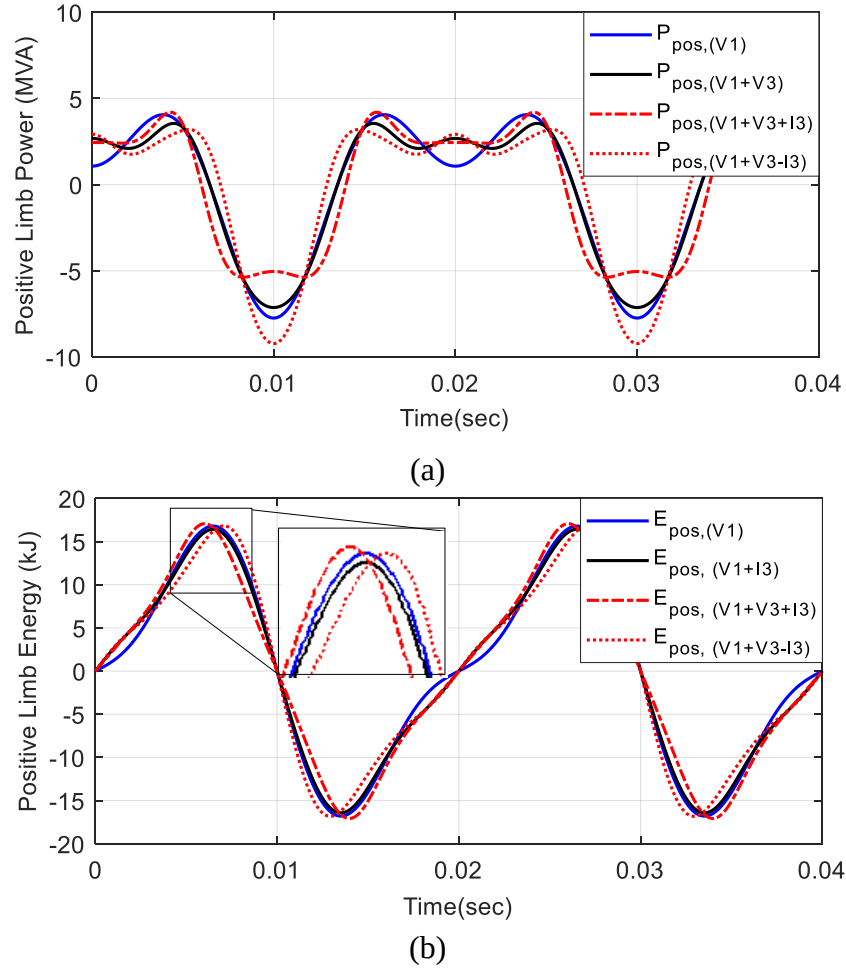


Figure 4.28: Positive limb power and energy variation when $|S|=1$ p.u and $P=+1$, ($+P$, ($\phi = 0^\circ$, β and $\gamma = 15\%$, $\theta_{v3} = 0^\circ$ and 3rd current $\phi_3 = 0^\circ$ and 180°).

Figure 4.30.a presents a comparison of the energy deviation using both 3rd harmonic voltage and current with the variation of power factor while Figure 4.30.b uses only 3rd harmonic current. Figure 4.30.c and Figure 4.30.d show the energy deviation reduction based on Figure 4.30.a and Figure 4.30.b, respectively.

Considering energy deviation, three significant behaviours can be observed as shown in Figure 4.30. Firstly, using in-phase third harmonic current injection can lead to increasing the energy deviation for the whole range of power factor variation $-90^\circ \leq$

$\phi \leq 90^\circ$. The maximum increase of energy deviation occurs at $\pm 90^\circ$ degrees while it reduces when moving towards more active power. Secondly, power factor ranges of $-90^\circ \leq \phi < 30^\circ$ and $30^\circ < \phi \leq 90^\circ$ experience significant reduction of energy deviation with a maximum reduction at pure reactive power operation due to using out of phase third harmonic current. However, in the range from $-30^\circ \leq \phi \leq 30^\circ$ 3rd harmonic voltage is the best choice to use to reduce energy deviation compared with fundamental operation and 3rd harmonic current injection.

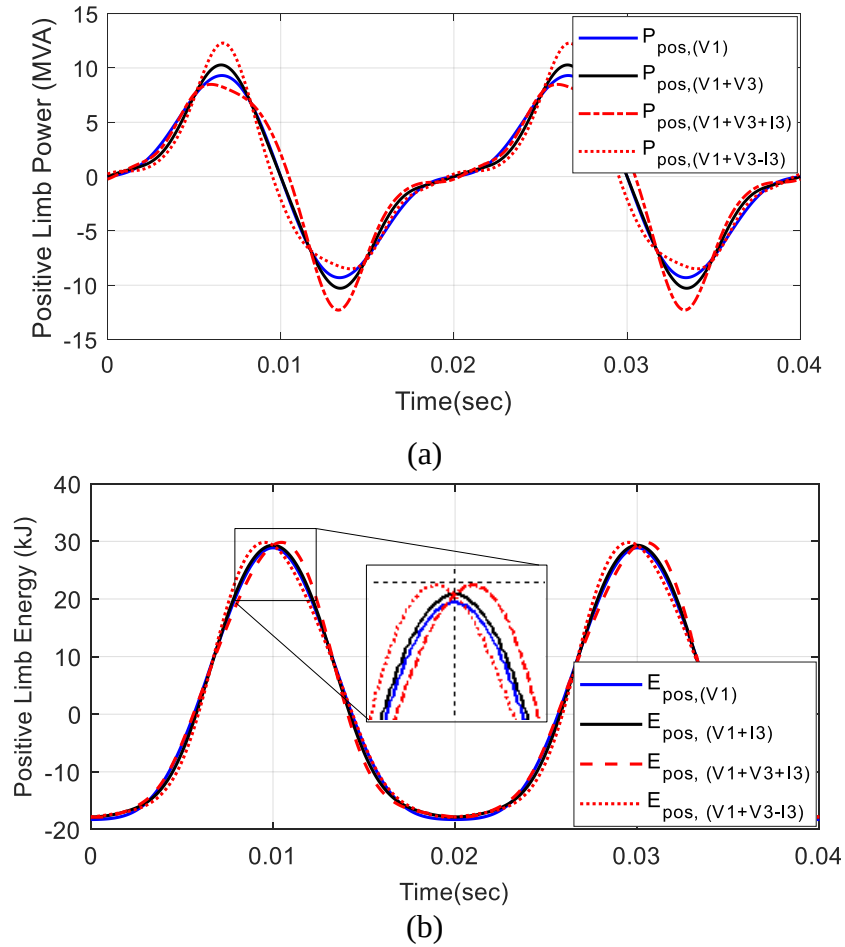


Figure 4.29: Positive limb power and energy variation when $|S|=1$ p.u, +Q, ($\phi = +90^\circ$, β and $\gamma = 15\%$, $\theta_{v3} = 0^\circ$ and 3rd current $\phi_3 = 0^\circ$ and 180°).

Comparing the usage of 3rd harmonic current injection with third harmonic voltage injection, it is observed that the latter can provide a significant energy deviation reduction when operating with more active power operation. Unlike the 3rd harmonic current the region $-30^\circ \leq \phi \leq 30^\circ$ experiences a reduction while the maximum reduction occurs at unity power factor. In the other two regions of $-90^\circ < \phi < -30^\circ$

and $30^\circ < \phi < 90^\circ$, however, energy deviation is increased. The operation at power angle of $\pm 30^\circ$ and $\pm 90^\circ$ energy deviation reduction is zero.

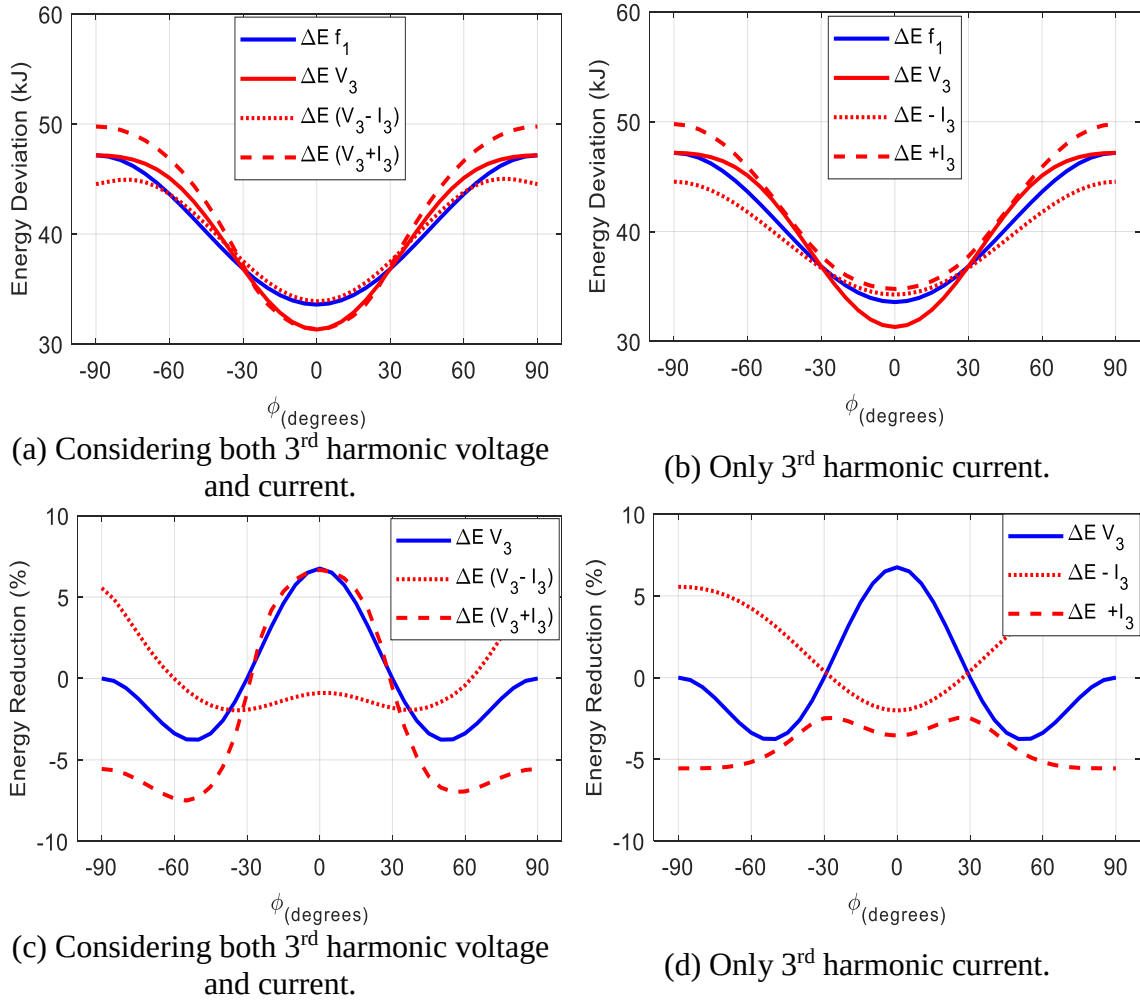


Figure 4.30: The impact of 3rd harmonic voltage and current injection on energy deviation reduction ($-90^\circ \leq \phi \leq +90^\circ$).

4.6 Second Order Current Injection

In this section 2nd harmonic current injection is considered. In the previous sections it is assumed that this current was controlled to be zero using an appropriate controller such as a proportional-resonant (PR) controller. However, the PR controller is now assumed to be used to inject a 2nd harmonic current. To achieve 2nd harmonic injection, the 2nd harmonic voltage also must appear on the limb voltage. However, the required 2nd harmonic voltage generated by the controller can be neglected compared with the fundamental and 3rd harmonic component. This is because its amplitude would be small for any sensible current. Therefore, and for simplicity, it is neglected and assumed to be zero in this analysis.

The 2nd harmonic component current appears naturally in the limb currents due to the interaction between the SM capacitor voltage and current during switching of the capacitor voltage ON/OFF. This is known by the circulating current which has already been widely discussed earlier. The dominant component of the circulating current is the 2nd harmonic component which so far is assumed to be zero in all previous analysis.

The 2nd harmonic current is often considered to be a potential issue associated with the MMC which can introduce additional conduction losses to the system. The 2nd harmonic current has been widely discussed in the literature and is known as the circulating current. In fact, not only the circulating current control but also its impact on the energy pulsation has been presented including its impact on limb available voltage, the synthesised voltage and the range of PQ operation [19, 64, 68, 69]. A steady state analysis method was also presented considering this current and an MMC model based on this current was presented [74]. Furthermore, the optimal injection has also been reported where both 2nd and 4th harmonic current injection was considered [19]. However, considering both 2nd harmonic current and 3rd harmonic voltage is not considered before. Therefore, a review on the 2nd harmonic is presented in this section. Then a comparison is presented considering the performance due to the 2nd harmonic current and 3rd harmonic voltage

In this section, the impact of only 2nd harmonic current is considered. Then, the performance is presented taking into consideration the existence of both 2nd harmonic current and 3rd harmonic voltage injection. However, for more convenience, the analysis of the positive limb voltage and current includes both. To consider the 2nd harmonic current alone, the 3rd harmonic term is set to zero.

The positive limb voltage and current are given by (4.73) and (4.74). As it can be seen, the triplen current is eliminated due to using a Delta transformer connection.

$$v_{pos.a}(t) = \frac{V_{dc}}{2} - \hat{V}_{C1} \cos(\omega_0 t - \theta_{v1}) + \hat{V}_{C3} \cos(3\omega_0 t - \theta_{v3}) \quad (4.73)$$

$$i_{pos.a}(t) = \frac{I_{dc}}{3} + \frac{\hat{I}_{C1}}{2} \cos(\omega_0 t - \theta_{i1}) + \hat{I}_{C2} \cos(2\omega_0 t - \theta_{i2}) \quad (4.74)$$

Here, V_{dc} is the DC link voltage. $\hat{V}_{C1}$ and $\hat{V}_{C3}$ are the amplitude of the fundamental voltage and 3rd harmonic content with their phase angle θ_{v1} and θ_{v3} , respectively. I_{dc}

is the DC current while $\hat{I}_{C1}$ and $\hat{I}_{C2}$ are the AC and the circulating currents with their phase angles θ_{i1} and θ_{i2} respectively, where θ_{i2} can be either 0° or 180° . It can be seen that θ_{v1} is used to distinguish between the phase angle of the fundamental and the 3rd harmonic voltage and current components.

4.6.1 The impact of only second harmonic current

Using (4.73) and (4.74) and after eliminating 3rd harmonic voltage component, positive limb instantaneous power can be given by (4.75).

$$p_{pos.a}(t) = \left[\begin{aligned} & \left(\frac{V_{dc}I_{dc}}{6} - \frac{\hat{V}_{C1}\hat{I}_{C1}}{4} \cos(\theta_{i1} - \theta_{v1}) \right) \\ & + \left(\frac{V_{dc}\hat{I}_{C1}}{4} \cos(\omega_0 t - \theta_{i1}) - \frac{\hat{V}_{C1}I_{dc}}{3} \cos(\omega_0 t - \theta_{v1}) \right) \\ & \quad - \frac{\hat{V}_{C1}\hat{I}_{C1}}{4} \cos(2\omega_0 t - \theta_{v1} - \theta_{i1}) \\ & + \left(\frac{V_{dc}\hat{I}_{C2}}{2} \cos(2\omega_0 t - \theta_{i2}) - \frac{\hat{V}_{C1}\hat{I}_{C2}}{2} \cos(\omega_0 t - \theta_{i2} + \theta_{v1}) \right) \\ & \quad - \frac{\hat{V}_{C1}\hat{I}_{C2}}{2} \cos(3\omega_0 t - \theta_{v1} - \theta_{i2}) \end{aligned} \right] \quad (4.75)$$

Assuming the reference phase angle of limb voltage is zero and therefore, the phase angle is represented by the power angle. Therefore, the average power is expressed by the following equation.

$$\bar{P}_{pos.a} = \left(\frac{V_{dc}I_{dc}}{6} - \frac{\hat{V}_{C1}\hat{I}_{C1}}{4} \cos(\emptyset) \right) \quad (4.76)$$

In order to achieve a sustainable operation of the MMC with balanced SM capacitors, the average power must equal zero and therefore the relationship between the direct power and average AC power is obtained by setting (4.76) to zero, the DC current can be expressed by Equation (4.77) which is the same as (4.22).

$$I_{dc} = \frac{3}{4} m_i \hat{I}_{C1} \cos(\emptyset) \quad (4.77)$$

It can be seen that the 2nd harmonic current has no impact on the active power flow. However, the existence of this current can be seen on the power pulsation which increases power losses of the MMC. Moreover, in general, injecting harmonics with the

circulating current requires reconsideration of the current rating of the power electronic devices. Considering the assumption made with regards to the 2nd harmonic voltage the existence of both 2nd harmonic voltage and current can lead to power transfer between the limbs. As a result this can be considered as an advantage and used to control the energy in the SM capacitors and therefore reduce capacitors ripple voltage as accordingly.

Positive limb instantaneous power can then be expressed in terms of the nominal apparent power. The modulation index denoted by m_i is still given by (3.13) while the amplitude of the 2nd harmonic component is re-presented by δ .

$$\delta = \frac{\hat{I}_{C2}}{\hat{I}_{C1}} \quad (4.78)$$

Thus;

$$p_{pos.a}(t) = \frac{|S|}{6} \left[\begin{array}{c} \left(\frac{2}{m_i} \cos(\omega_0 t - \phi) - m_i \cos(\phi) \cos(\omega_0 t) \right) \\ - \cos(2\omega_0 t - \phi) \\ + 2\delta \left(\frac{2}{m_i} \cos(2\omega_0 t - \theta_{i2}) - \cos(\omega_0 t - \theta_{i2} + \theta_{v1}) \right) \\ - \cos(3\omega_0 t - \theta_{v1} - \theta_{i2}) \end{array} \right] \quad (4.79)$$

From the instantaneous power equation above and by considering the amplitude of the fundamental and other harmonic components, it can be said that the instantaneous power is produced from two terms. The first is due to the interaction between the DC and fundamental component of the voltage and current. The second is produced due to the interactions between DC, fundamental and the 2nd harmonic component.

$$p_{pos.a}(t) = p_{pos.f1}(t) + p_{pos.f2}(t) \quad (4.80)$$

Positive limb energy variation can be given by (4.81).

$$e_{pos.a}(t) = \frac{|S|}{6\omega_o} \left[\begin{array}{c} \left(\frac{2}{m_i} \cos(\omega_0 t - \phi) - m_i \cos(\phi) \sin(\omega_0 t) - \frac{1}{2} \sin(2\omega_0 t - \phi) \right) \\ + 2\delta \left(\frac{1}{m_i} \sin(2\omega_0 t - \theta_{i2}) - \sin(\omega_0 t - \theta_{i2} + \theta_{v1}) \right) \\ - \frac{1}{3} \sin(3\omega_0 t - \theta_{v1} - \theta_{i2}) \end{array} \right] \quad (4.81)$$

$$e_{pos.a}(t) = e_{pos.f1}(t) + e_{pos.f2}(t) \quad (4.82)$$

From the previous analysis, it can be seen that the instantaneous power and energy variation of a limb has two main terms. The first is due to the fundamental component while the other is due to the 2nd harmonic current injection. The simulation results and discussion of the previous analysis are presented later on in Section 4.6.3.

4.6.2 Second harmonic current and 3rd harmonic voltage injection

Using (4.73) and (4.74), positive limb instantaneous power can be given by (4.83).

$$p_{pos.a}(t) = \left[\begin{aligned} & \left(\frac{V_{dc}I_{dc}}{6} - \frac{\hat{V}_{c1}\hat{I}_{c1}}{4} \cos(\theta_{i1} - \theta_{v1}) \right) \\ & + \left(\frac{V_{dc}\hat{I}_{c1}}{4} \cos(\omega_0 t - \theta_{i1}) - \frac{\hat{V}_{c1}I_{dc}}{3} \cos(\omega_0 t - \theta_{v1}) \right) \\ & - \frac{\hat{V}_{c1}\hat{I}_{c1}}{4} \cos(2\omega_0 t - \theta_{v1} - \theta_{i1}) \\ & + \left(\frac{V_{dc}\hat{I}_{c2}}{2} \cos(2\omega_0 t - \theta_{i2}) - \frac{\hat{V}_{c1}\hat{I}_{c2}}{2} \cos(\omega_0 t - \theta_{i2} + \theta_{v1}) \right) \\ & - \frac{\hat{V}_{c1}\hat{I}_{c2}}{2} \cos(3\omega_0 t - \theta_{v1} - \theta_{i2}) \\ & + \left(\frac{\hat{V}_{c3}I_{dc}}{3} \cos(3\omega_0 t - \theta_{v3}) + \frac{\hat{V}_{c3}\hat{I}_{c1}}{4} \cos(2\omega_0 t + \theta_{i1} - \theta_{v3}) \right) \\ & + \frac{\hat{V}_{c3}\hat{I}_{c1}}{4} \cos(4\omega_0 t - \theta_{i1} - \theta_{v3}) \\ & + \left(\frac{\hat{V}_{c3}\hat{I}_{c2}}{2} \cos(\omega_0 t + \theta_{i2} - \theta_{v3}) + \frac{\hat{V}_{c3}I_{dc}}{2} \cos(5\omega_0 t - \theta_{i2} - \theta_{v3}) \right) \end{aligned} \right] \quad (4.83)$$

The average power then can be given by the equation below.

$$\bar{P}_{pos.a} = \left(\frac{V_{dc}I_{dc}}{6} - \frac{\hat{V}_{c1}\hat{I}_{c1}}{4} \cos(\emptyset) \right) \quad (4.84)$$

It can be seen that the average power obtained by (4.84) is the same result obtained by (4.76). Therefore, the DC current is expressed by (4.85) which the same result as (4.77).

$$I_{dc} = \frac{3}{4} m_i \hat{I}_{c1} \cos(\emptyset) \quad (4.85)$$

It can be seen that both the 2nd harmonic current and 3rd harmonic voltage have no impact on the active power. However, the existence of this current can still be seen on the power pulsation which increases power losses of the MMC.

Similar to the previous section, positive limb instantaneous power can then be expressed in terms of the nominal apparent power. The modulation index denoted by m_i is still

given by (3.13) while the amplitude of 3rd harmonic and 2nd harmonic components can be re-presented by β and δ respectively. Thus;

$$p_{pos.a}(t) = \frac{|S|}{6} \left[\begin{aligned} &\left(\frac{2}{m_i} \cos(\omega_0 t - \phi) - m_i \cos(\phi) \cos(\omega_0 t) - \cos(2\omega_0 t - \phi) \right) \\ &+ 2\delta \left(\frac{2}{m_i} \cos(2\omega_0 t - \theta_{i2}) - \cos(\omega_0 t - \theta_{i2} + \theta_{v1}) \right. \\ &\quad \left. - \cos(3\omega_0 t - \theta_{v1} - \theta_{i2}) \right) \\ &+ \beta \left(m_i \cos(\phi) \cos(3\omega_0 t - \theta_{v3}) + \cos(2\omega_0 t + \phi - \theta_{v3}) \right. \\ &\quad \left. + \cos(4\omega_0 t - \phi - \theta_{v3}) \right) \\ &+ 2\beta\delta (\cos(\omega_0 t + \theta_{i2} - \theta_{v3}) + \cos(5\omega_0 t - \theta_{i2} - \theta_{v3})) \end{aligned} \right] \quad (4.86)$$

The energy variation is given by (4.87). Then considering the energy variation expressed by (4.81) and (4.87), the two main dominant terms are the first two ones which including $\frac{2}{m_i} \sin(\omega_0 t - \phi)$ and $\frac{1}{m_i} \sin(2\omega_0 t - \theta_{i2})$. To examine the impact of 2nd harmonic current component on energy variation, simulation results are obtained taking into consideration both 2nd harmonic current and 3rd harmonic voltage injection.

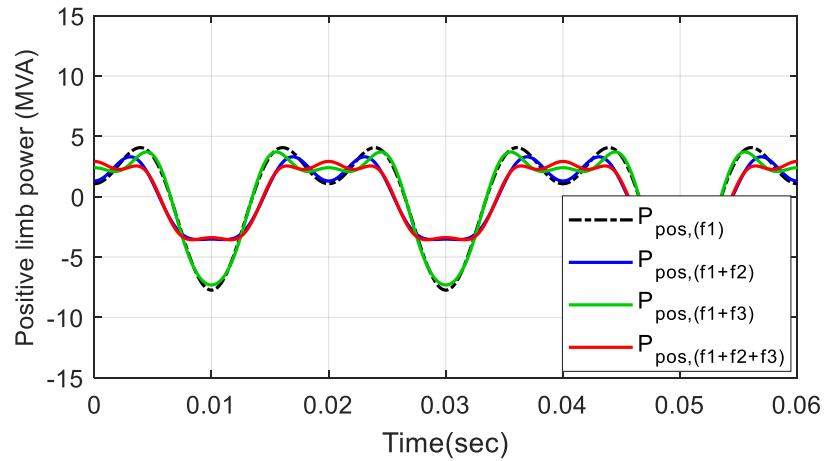
$$e_{pos.a}(t) = \frac{|S|}{6\omega_o} \left[\begin{aligned} &\left(\frac{2}{m_i} \sin(\omega_0 t - \phi) - m_i \cos(\phi) \sin(\omega_0 t) - \frac{1}{2} \sin(2\omega_0 t - \phi) \right) \\ &+ 2\delta \left(\frac{1}{m_i} \sin(2\omega_0 t - \theta_{i2}) - \sin(\omega_0 t - \theta_{i2} + \theta_{v1}) \right. \\ &\quad \left. - \frac{1}{3} \sin(3\omega_0 t - \theta_{v1} - \theta_{i2}) \right) \\ &+ \beta \left(\frac{m_i}{3} \cos(\phi) \sin(3\omega_0 t - \theta_{v3}) + \frac{1}{2} \sin(2\omega_0 t + \phi - \theta_{v3}) \right. \\ &\quad \left. + \frac{1}{4} \sin(4\omega_0 t - \phi - \theta_{v3}) \right) \\ &+ 2\beta\delta \left(\sin(\omega_0 t + \theta_{i2} - \theta_{v3}) + \frac{1}{5} \sin(5\omega_0 t - \theta_{i2} - \theta_{v3}) \right) \end{aligned} \right] \quad (4.87)$$

4.6.3 Simulation Results

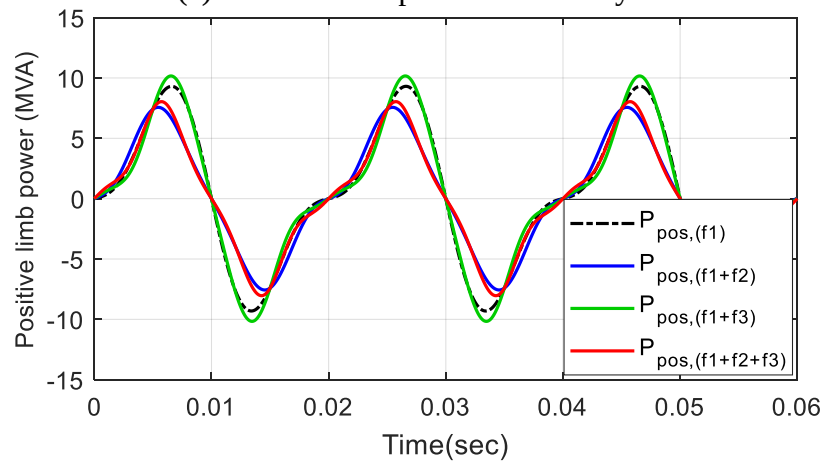
The simulation results obtained are based on the MMC parameters given by Table 3.2. The results are based on operation at rated power $|S|$. The 3rd harmonic voltage injection shows that maximum energy deviation reduction occurs when operating with only $\pm P$. While for the operation with only $\pm Q$ there is no energy deviation reduction. Therefore, only the operating points with $+P$ and $+Q$ are considered. The 2nd harmonic current and 3rd harmonic voltage injections are set at 15% of the respective fundamental components are used and the modulation index is 0.9. Please note that the choice of 15% injection

for both voltage and current has been made only as an example of values that could be acceptable in practice.

Positive limb power pulsation is illustrated by Figure 4.31. It can be seen that peak to peak power is reduced due to the 2nd harmonic current injection. This can be further reduced using both 2nd harmonic current and 3rd harmonic voltage injection in both cases of the operating points +P and +Q. In the figures, $P_{pos,(f1)}$, $P_{pos,(f1+f2)}$, $P_{pos,(f1+f3)}$ and $P_{pos,(f1+f2+f3)}$ are the limb instantaneous power due to fundamental, combination of fundamental and 2nd harmonic current injection, combination of fundamental and 3rd harmonic voltage and combination of fundamental, 2nd harmonic current and 3rd harmonic voltage injection respectively. Similarly, the energy variations are denoted by $E_{pos,(f1)}$, $E_{pos,(f1+f2)}$, $E_{pos,(f1+f3)}$, $E_{pos,(f1+f2+f3)}$, respectively. $\theta_{v3} = 0^\circ$ since it has no impact on the energy variation.



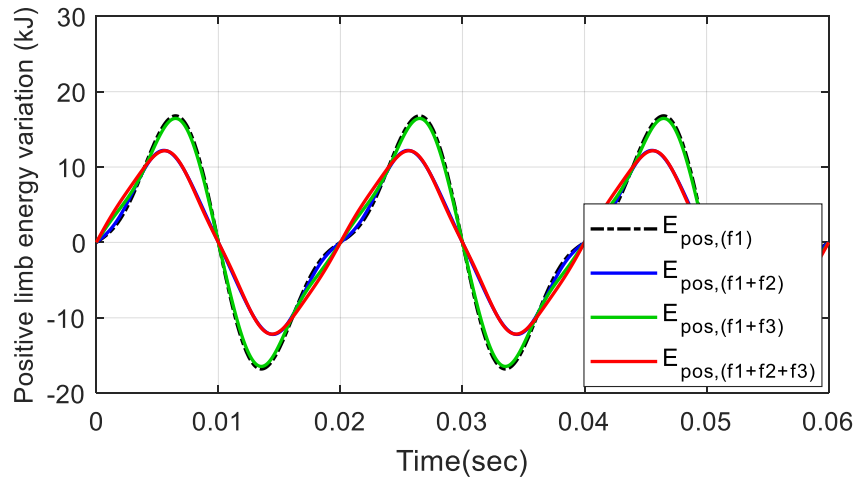
(a) Positive limb power due to only +P.



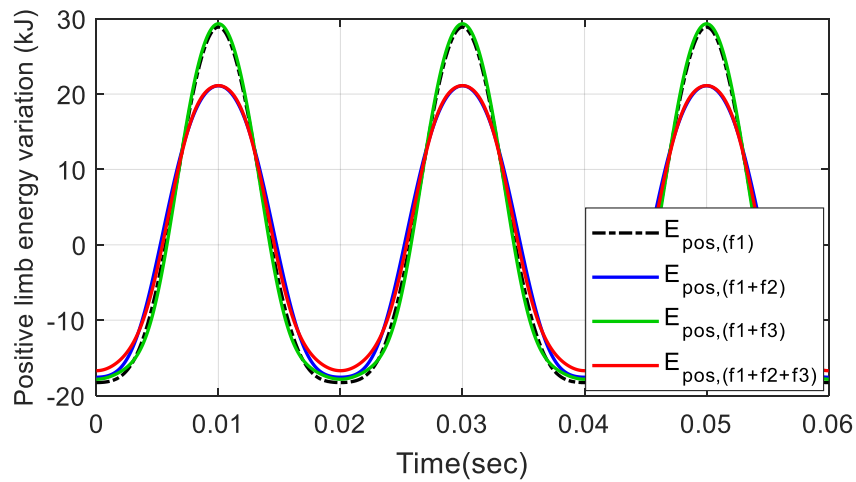
(b) Positive limb power due to only +Q.

Figure 4.31: Comparison of positive limb power due to 2nd and 3rd harmonic injection for two operating points ; (a) only +P and (b) only +Q.

Figure 4.32 shows the waveforms of the positive limb energy variation taking into the consideration the injection of both, 2nd harmonic current and 3rd harmonic voltage. The simulation results are based on modulation index of $m_i = 0.9$, $\beta = 50\%$ and $\delta = 15\%$ and $\theta_{i2} = 0^\circ$ and 90° for (a) and (b) respectively. The figure shows that using only 3rd harmonic voltage injection leads to insignificant reduction of energy deviation. However, using only 2nd harmonic injection or the combination of both 2nd harmonic current and 3rd harmonic voltage components can lead to a significant reduction in both cases of +P and +Q compared to the fundamental and only 3rd harmonic voltage injection. It can be said that the significant reduction is only caused by the 2nd harmonic current injection



(a) positive limb energy variation at only +P.



(b) positive limb energy variation at only +Q.

Figure 4.32: Comparison of positive limb energy variation due to 2nd and 3rd harmonic injection for two operating points of +P and +Q.

It is observed that the phase angle of the 2nd harmonic injection has a significant impact on the energy fluctuation as shown by the example given by Figure 4.33. A significant reduction is achieved and caused by the 2nd harmonic current when the latter phase angle equals the power factor angle, ϕ , with a range of $(-90^\circ \leq (\theta_{i2} = \phi) \leq 90^\circ)$ as illustrated by Figure 4.33.a confirming the results shown by Figure 4.32 where $\theta_{i2} = \phi$, $(0^\circ \text{ and } 90^\circ)$. However, fixing the phase angle of the 2nd harmonic component to zero as illustrated by Figure 4.33.b, the energy reduction decreases when $\phi = \pm 90^\circ$ for the same operating conditions mentioned above. It is observed that the 3rd harmonic phase angle has no impact on energy deviation reduction. Therefore, it is set to zero when considering the presence of the 3rd harmonic injection. Moreover, there is no recognisable improvement on the energy reduction when considering both 2nd and 3rd harmonic components. Generally, it can be said that the significant reduction is only caused by the 2nd harmonic current injection.

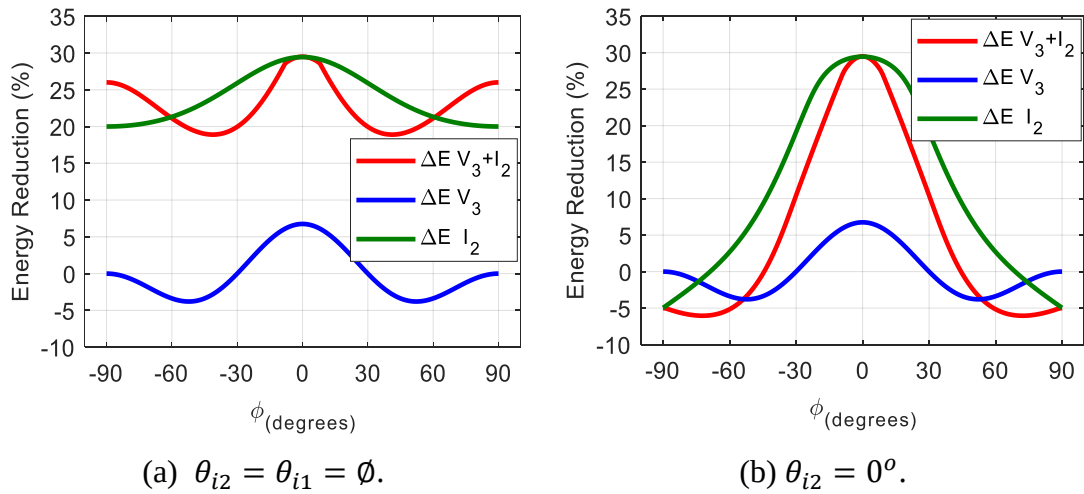


Figure 4.33: The impact of 2nd and 3rd harmonic components ($\gamma = 15\%$ and $\beta = 50\%$) on energy deviation reduction for different phase angle of the 2nd harmonic; (a) ($\theta_{i2} = \theta_{i1} = \phi$) equals the fundamental component phase angle, (b) constant $\theta_{i2} = 0^\circ$.

From the previous analysis, it can be seen that 2nd harmonic injection can be used with 3rd harmonic voltage injection to improve the operation of the MMC. Considering Figure 4.33, the table below represents a comparison of energy deviation reduction due to 2nd harmonic current and 3rd harmonic voltage injection based on 15% of 2nd and 50% of 3rd harmonic components of the fundamental as shown by the previous figures. It can be seen that using the combination of both 2nd and 3rd harmonic injection can lead to

insignificant energy reduction which can be neglected especially if the focus was on reducing energy deviation. Moreover, 2nd harmonic current injection has better performance than the 3rd harmonic voltage injection. However, both can be used to reduce energy deviation while the 3rd harmonic voltage can be used to reduce also the peak synthesised voltage. This can lead to further improvement in operation of the MMC.

It can be seen that a significant energy deviation reduction can be achieved due to 2nd harmonic current injection for $\pm P$ and $\pm Q$. However, the energy deviation reduction is greater when operating with only $\pm P$ than when operating with only $\pm Q$. Table 4.1 summarised the effect of different harmonic injections on the energy fluctuation illustrated by Figure 4.33. Considering that the focus of the chapter is on energy fluctuation only, discussion about improvements in the available voltage is presented in Chapter 6.

Table 4.1: Energy deviation reduction due to 2nd harmonic current and 3rd harmonic voltage injection $\gamma = 15\%$ and $\beta = 50\%$.

	P and Q operating point							
	ϕ							
	0°	$\pm 30^{\circ}$	$\pm 60^{\circ}$	$\pm 90^{\circ}$	0°	$\pm 30^{\circ}$	$\pm 60^{\circ}$	$\pm 90^{\circ}$
2 nd harmonic phase angle, θ_{i2}	$\theta_{i2} = \phi$				$\theta_{i2} = 0^{\circ}$			
Harmonic order	Energy Deviation Reduction (%)							
3 rd harmonic	7	0	$\cong 3$	0	7	0	$\cong -3$	0
2 nd harmonic	$\cong 30$	25	$\cong 22$	20	$\cong 30$	20	$\cong -2$	-5
2 nd and 3 rd	$\cong 30$	20	$\cong 22$	$\cong 26$	$\cong 30$	10	-5	-5

4.7 Summary and Conclusion

Based on the previous analysis, it is observed that using third harmonic voltage and current has an impact on energy deviation reduction. This reduction, however, depends on the power factor and the modulation index and therefore can only be beneficial with specific range of power factor as discussed earlier. The reduction also depends on the amount of 3rd harmonic injection. For instance, 50% of 3rd harmonic injection from the fundamental can cause a reduction of about 7% while about 2.2% reduction can be caused by 15% of 3rd harmonic voltage injection as shown by Figure 4.21.

The outcome of the analysis can be summarised in Table 4.2 below where both third harmonic voltage and current are considered. Both approaches can lead to energy deviation reduction. However, considering the range of power factor and other aspects such as losses, complexity and also the current rating of the IGBTs, third harmonic current may only be valid for special applications such as STATCOM while using only third harmonic voltage injection is valid for HVDC transmission applications. Therefore, the latter is considered within the next chapters with more detail.

The table provides a qualitative summary of the range of power factor and the impact of 3rd harmonic voltage and current injection on energy deviation reduction without considering the optimal values. The maximum available energy reduction and the associated power factor are not shown since they would require specific value of 3rd harmonic injection which would be out of the scope of this Chapter.

Table 4.2: The MMC applications based on the energy deviation reduction using third harmonic injection voltage/current.

		Energy deviation Statue	Power factor range of reduction	Maximum reduction/increase at	Zero reduction/increase at	Applications
3 rd V	classic	Reduction	$-30 \leq \phi \leq 30$	$\phi = 0$	$\phi = \pm 30$ & $\phi = \pm 90$	HVDC power transmission
	Phase shifted	Reduction	$-30 \leq \phi \leq 30$	$\phi = 0$	$\phi = \pm 30$ & $\phi = \pm 90$	HVDC power transmission
3 rd I	Phase angle=0°	Only Increase	$90 - \phi \leq 90$	Non	Non	Not recommended
	Phase angle=180°	Reduction	$-90 \leq \phi \leq -30$ & $30 \leq \phi \leq 90$	$\phi = \pm 90$	$\phi = \pm 30$	Reactive power compensation (STATCOM)
Both	Phase angle=0°	Reduction	$-30 \leq \phi \leq 30$	$\phi = 0$	$\phi \approx \pm 30^\circ$	HVDC power transmission
	Phase angle=180°	Reduction	$-90 \leq \phi \leq -60$ & $60 \leq \phi \leq 90$	$\phi = \pm 90$	$\phi = \pm 60$	Reactive power compensation (STATCOM)

Third harmonic current injection is suitable for applications where reactive power compensation is required to support grid voltage such as STATCOM. HVDC transmission applications, especially long-distance bulk power transmission where the

concern is that more active power is required, 3rd harmonic voltage injection is applicable.

Unlike 3rd harmonic injection, a significant energy deviation reduction can be achieved due to 2nd harmonic current injection for $\pm P$ and $\pm Q$ as shown by Table 4.1. However, the energy deviation reduction is greater when operating point with only $\pm P$ than when operating with $\pm Q$. Generally, it can be said that the significant reduction is only caused by the 2nd harmonic current injection. Moreover, using both can improve the MMC performance where the current injection reduces the available voltage ripple while the voltage harmonic injection reduces the peak synthesised voltage. Therefore, the next chapter considers a further investigation on the effect of 3rd harmonic voltage injection on SM capacitor size.

In this chapter, an analytical study on the impact of harmonic injection on the operation and design of the MMC has been discussed. A mathematical model of the MMC has been presented considering both fundamental and additional harmonic injection including fundamental, 2nd and the 3rd harmonic components. Applications are suggested based energy variation and its reduction.

In Chapter 5, a review of capacitor sizing methods is discussed. The same method is implemented investigating the impact of third harmonic voltage injection on capacitor size reduction. A comparison is presented between fundamental and third harmonic voltage injection for the application of HVDC power transmission.

Chapter 5

5 MMC SM Capacitor Sizing Based on peak-to-peak Energy Deviation

5.1 Introduction

In the previous Chapter, it was concluded that peak-to-peak voltages and currents are affected using 3rd harmonic voltage and current injection. It was concluded that both 3rd harmonic voltage and current can lead to reducing energy deviation as presented by Table 4.2 depending on the operation point and the power factor. Therefore, these findings can be used to determine the applicable 3rd harmonic injection for particular applications. It is also observed that the impact on energy variation reduction due to 3rd harmonic current injection is negligible compared with 3rd harmonic voltage injection alone which has the advantages over 3rd harmonic current injection by reducing peak-to-peak synthesised voltage. Moreover, unlike 3rd harmonic voltage injection, the current rating of the semiconductors needs to be considered when 3rd harmonic current injection is used.

Considering the application of HVDC transmission system, the focus of this Chapter is on an analysis of 3rd harmonic voltage injection impact on energy storage requirements including SM capacitor size. First, a review of capacitor size based on intra-cycle energy deviation is discussed. The same technique is implemented with third harmonic voltage injection design mode.

One of the important aspects to be considered during MMC design is the energy storage requirements. Peak-to-peak energy deviation is the key factor in determining the SM capacitor size. It is generally reported that the size of the SM capacitors is proportional to the energy deviation (peak-to-peak) of the converter [16, 45, 75]. However, the energy variation itself within the MMC does not depend on the capacitor size as discussed in Chapter 3 and expressed by (3.50). The chosen capacitor size should have the ability to deal with rated energy variation, so, the available voltage from the capacitors is always

greater than the synthesised voltage which avoids undesirable over modulation. The reported studies so far consider storage energy requirements including capacitor size under fundamental frequency operation [16, 75]. However, the effect of 3rd harmonic injection yet has to be considered in detail.

Although the MMC offers significant advantages for HVDC system applications, the energy storage requirement significantly affects the size of the converter station especially for offshore and other similar applications where size and weight are critical. Therefore, reduction in the energy storage components in the MMC circuit could be economically beneficial especially the SM capacitor size. Within this Chapter, research into the investigation of the effect of third harmonic voltage component injection on the energy storage requirement of a converter submodule is presented.

5.2 SM Capacitor Size without Additional Harmonic Injection

In order to understand the impact of capacitor size on the operation of the MMC, let us first consider Figure 5.1, below, which illustrates its importance. This example is based on simulation results of two MMC units which have the same parameters and operate under the same conditions while their SMs have different capacitor sizes. In Figure 5.1 the available voltage of the positive limb with a bigger capacitor size is represented by the continuous blue line. The red dashed line, however, has a capacitor size one third of the previous one. The dashed line illustrates entry into overmodulation mode caused by an unsuitable size of capacitor where the peak of the synthesised voltage was not considered during capacitor selection. Overmodulation happens when the total available voltage (dashed blue line) falls below the desired voltage V_{pos} , meaning that the converter is not capable of generating the desired reference.

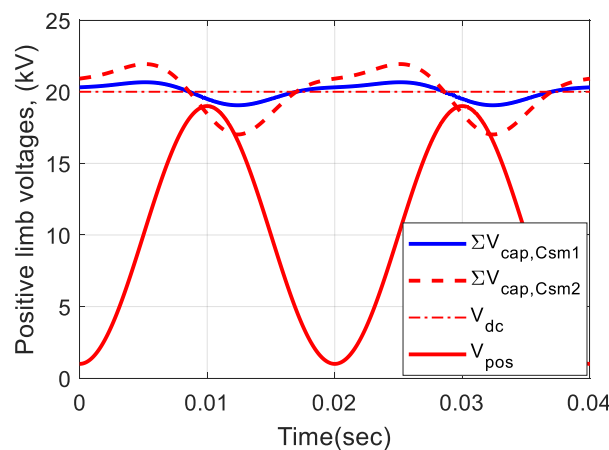


Figure 5.1 The importance of capacitor size to avoid overmodulation.

The method for analysing capacitor size is based on the local maximum and minimum of energy variations. This technique has already been introduced and implemented not only for the generalized MMC but also for alternative MMC topologies such as the Alternate-Arm converter (AAC) [16, 45, 76] .

The derivation for the instantaneous power and energy variation of the positive limb has already been considered in Section 3.4.2. The generalised positive limb instantaneous power and energy were discussed in Sections 3.4.2 and (5.1). It can be seen that the limb energy variation does not depend on the size of capacitor.

$$e_{pos.a}(t) = \frac{|S|}{6\omega_0} \left[\left(\frac{2}{m_i} \sin(\omega_0 t - \phi) - m_i \cos(\phi) \sin(\omega_0 t) \right) - \frac{1}{2} \sin(2\omega_0 t - \phi) \right] \quad (5.1)$$

where $e_{pos}(t)$ is the positive limb energy variation. Let us first determine the relationship between the SMs capacitor size and the energy stored in a limb. The energy stored in the positive limb can be given by equation below;

$$E_{pos} = \frac{1}{2} N_{SM} C_{SM} V_{C_{SM}}^2 \quad (5.2)$$

where $V_{C_{SM}}$ is the voltage of the SM which has a nominal voltage of $V_{C_{SM}.n}$ and a voltage ripple given below;

$$\Delta V_{C_{SM}} = V_{C_{SM}.max} - V_{C_{SM}.min} \quad (5.3)$$

$V_{C_{SM}.max}$ and $V_{C_{SM}.min}$ denote the peak maximum and minimum values of the capacitor voltage, respectively. The energy deviation of (5.1) was discussed in Chapter 3 and is given by equation (5.4) below.

$$\Delta E = \frac{1}{2} N_{SM} C_{SM} (V_{C_{SM}.max}^2 - V_{C_{SM}.min}^2) \quad (5.4)$$

Please note that equation 5.4 is true under the hypothesis that voltage ripple is symmetrical. As demonstrated in Appendix C, the capacitor voltage ripple is not

symmetrical. However, for small capacitor voltage ripple, the impact of the asymmetry can be neglected, i.e. $\frac{\Delta V_{C_{SM}}^2}{4} \ll V_{C_{SM}.max}$. Capacitor voltage ripple $\Delta V_{C_{SM}}$ then can be expressed by equation (5.5).

$$\delta = \frac{\Delta V_{C_{SM}}}{V_{C_{SM}.n}} \quad (5.5)$$

Considering Appendix C, it is assumed that the voltage ripple is symmetrical about the mean submodule voltage and using (5.5) so that equations (5.6) and (5.7) below are true. Therefore,

$$V_{C_{SM}.max} = V_{C_{SM}.n} + \frac{1}{2} \delta V_{C_{SM}.n} \quad (5.6)$$

$$V_{C_{SM}.min} = V_{C_{SM}.n} - \frac{1}{2} \delta V_{C_{SM}.n} \quad (5.7)$$

Substituting (5.6) and (5.7) into (5.4) and after manipulation, energy deviation can be re-expressed by;

$$\Delta E = \delta N_{SM} C_{SM} V_{C_{SM}.n}^2 \quad (5.8)$$

The total exact energy of the positive limb at any instant in time has already been derived in Section 3.4.2 by equation (3.37). This can be rewritten here by (5.9).

$$e_{pos}(t) + E_{ave} = \frac{1}{2} C_{SM} N_{SM} v_{C_{SM}}^2(t) \quad (5.9)$$

The total available voltage of one limb is presented in Section 3.4.2 and expressed by (3.42) which can be re-expressed by (5.10). The individual SM capacitor voltage, however, can be given by (5.11).

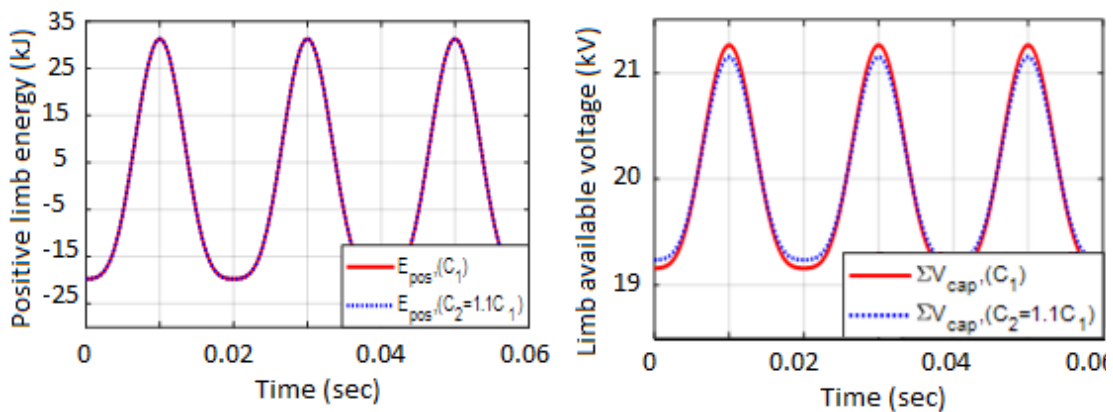
$$\Sigma v_{C_{SM}}(t) = \sqrt{\frac{2 N_{SM} e_{pos}(t)}{C_{SM}} + V_{dc}^2} \quad (5.10)$$

$$v_{C_{SM}}(t) = \sqrt{\frac{2e_{pos}(t)}{N_{SM}C_{SM}}} + V_{C_{SM}.n}^2 \quad (5.11)$$

As discussed earlier in Section 3.4.2, and also can be seen from (5.10) and (5.11), the total available voltage is produced by two components, an average denoted by V_{dc} and a ripple denoted by the first term under the root squared sign. The individual SM capacitor voltage has an average of $V_{C_{SM}.n}$ and a ripple associated by energy variation $e_{pos}(t)$.

The basic idea of using the intra-cycle energy technique (energy deviation) is simply based on using maximum energy deviation (maximum and minimum energy). To determine the proper size of capacitor, it requires using the maximum energy deviation which occurs during the whole active and reactive power envelope (PQ capability curve) to make sure that the worst case scenario is included. The maximum energy deviation represents the maximum available voltage ripple that the MMC can experience. Using (5.8), the minimum size of capacitor required to meet maximum energy deviation can be determined.

$$C_{SM.min} \geq \frac{\Delta E_{max}}{\delta N_{SM} V_{C_{SM}.n}^2} \quad (5.12)$$



(a) The Instantaneous energy $E_{post}(t)$

(b) Positive limb available voltage

$\Sigma V_{cap.pos}$

Figure 5.2 The available voltage and energy for MMC for a two size of capacitors,

$C_1=12.0mF$ and 110% of C_1 .

As mentioned before, it should be noticed that the maximum energy deviation does not depend on SM capacitor size. Figure 5.2 compares between using two different sizes of capacitors where the energy variation is the same while the waveforms of the available voltage are different.

5.3 The Impact of 3rd Harmonic Voltage Injection on Capacitor Sizing

An over view of capacitor sizing based on intra-cycle energy was presented considering only fundamental voltage and current components. This subsection, investigates the effect of 3rd harmonic voltage injection on the capacitor size based on the same method. A comparison is also presented here.

Using the same principle as in the previous section, capacitor size can be obtained using (5.13). Maximum energy deviation must consider the impact of harmonic components. Even though the 3rd harmonic voltage injection has an impact on the synthesised voltage, the energy deviation is not affected significantly. Energy variation has already been discussed in Section 4.4.3 and expressed by (4.42). It is observed that energy variation using 3rd harmonic injection is a combination of two terms, one caused by the fundamental component and the other caused by the 3rd harmonic. Thus;

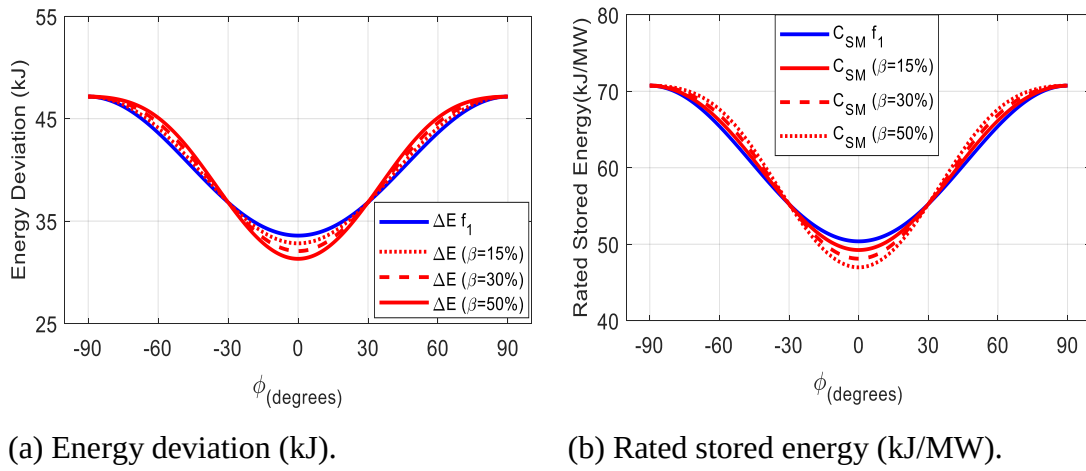
$$C_{SM.min} \geq \frac{\Delta E_{max}^{f1} + \Delta E_{max}^{f3}}{\delta N_{SM} V_{C_{SM}.n}^2} \quad (5.13)$$

From the energy variation discussed in previous Chapter, energy variation depends on the PQ operation point and third harmonic injection. It is observed that energy deviation can either be increased or decreased by using 3rd harmonic injection. This is further investigated later using a simulation model.

5.4 Simulation Results and Discussion.

For the given MMC system parameters in Table 3.2, Section 4.4.4, energy deviation depends on the variation of power factor. Therefore, capacitor size also changes in the same manner. Figure 5.3 illustrates the variation of energy deviation and the rated stored energy where its unit is given by (kJ/MW = ms) [16] for a range of power factors from $-90^\circ \leq \phi \leq 90^\circ$. It is observed that the same variation is experienced with the inverting and rectification mode of operation.

The energy deviation is given in Figure 5.3.a while capacitor size with regards to P and Q variation is given by the total stored energy as illustrated by Figure 5.3.b. It can be seen that, the minimum size of capacitor is required to meet the worst case of energy deviation occurs when the MMC operates with only $\pm Q$ (capacitive /reactive). However, if the size of capacitor was reselected based on operation with only real power, $\pm P$ then the capacitor size cannot provide sustainable converter operation. This becomes much worse especially for higher values of reactive power, Q. Furthermore, unlike the CSC converters, in practice VSC converters are chosen to operate with reasonable $\pm Q$ for AC voltage support.



(a) Energy deviation (kJ). (b) Rated stored energy (kJ/MW).

Figure 5.3: Energy deviation and capacitor size with variation of P and Q due to fundamental and 3rd harmonic injection operation.

It can be observed that even using third harmonic injection of 50% of the fundamental, energy deviation and capacitor size are reduced to a maximum of 7% while 30% of 3rd harmonic injection leads to a maximum capacitor size reduction of about 4.5% as illustrated in Figure 5.4. Furthermore, the reduction occurs only between power angle region of $(-30^\circ \leq \phi \leq +30^\circ)$ with its significant reduction at unity power factor. The other regions experience an increase of the required size of capacitor, $-90^\circ \leq \phi \leq -30^\circ$ and $+30^\circ \leq \phi \leq +90^\circ$.

Figure 5.5 illustrates a comparison between operating of the MMC where two different sizes of capacitors are considered. The first is designed based on the worst case scenario at -90° degrees with a modulation index of 0.8. The second is designed based on unity power factor. Comparing both cases and considering the operation range given in Table 3.2, the size of the capacitor designed based on the operating at only $\pm P$ cannot meet

sustainable operation of the MMC. This is because the MMC can experience the entry into overmodulation mode as illustrated by Figure 5.5.b.

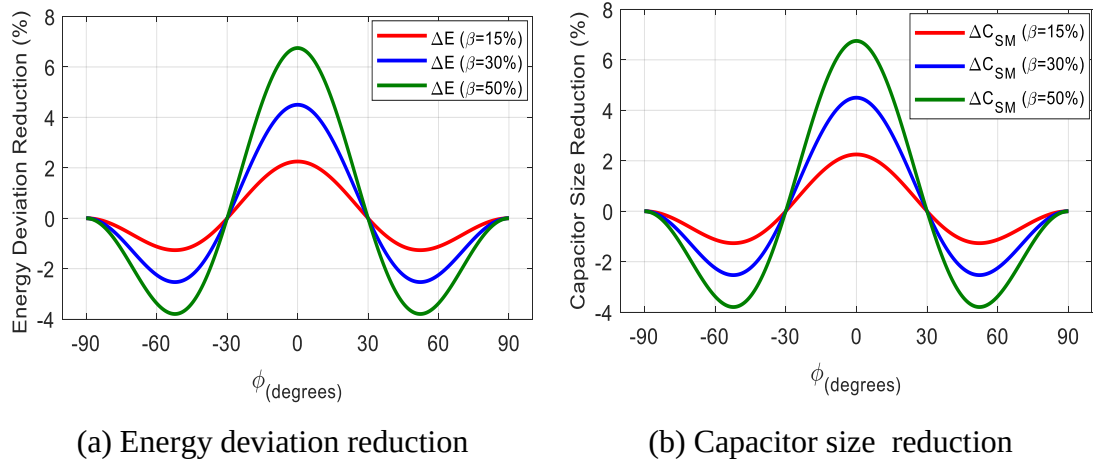


Figure 5.4: Energy and capacitor size reduction with variation of P and Q due to fundamental and 3rd harmonic injection operation.

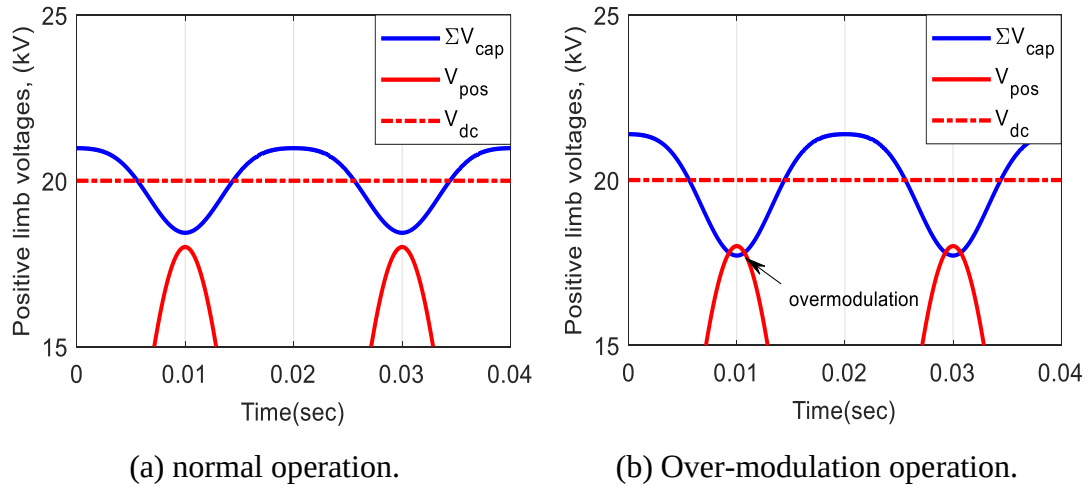
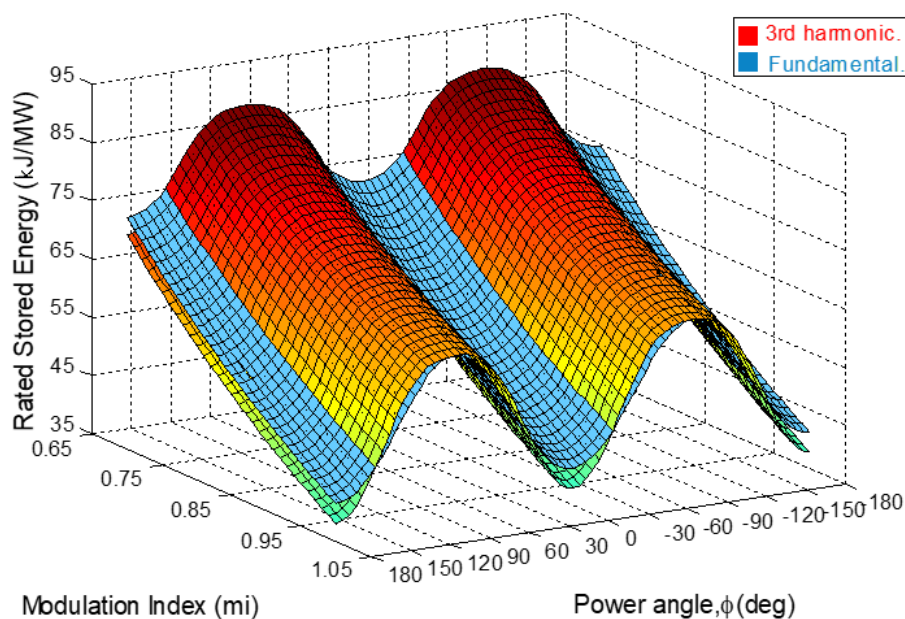


Figure 5.5: Over-modulated and normal operation of a MMC with two different capacitor sizes designed based on (a) $\phi=-90^\circ$ and $P=0$ and (b) $+P$ and $\phi=0^\circ$ degrees of modulation, index of 0.8.

The previous waveforms and simulation results are based on a constant modulation index of 0.8. However, if the modulation index is varied, the behaviour of energy deviation and therefore the required size of capacitor then can be given by Figure 5.6.

It is observed that when modulation index increases, the required capacitor size and total stored energy also can be reduced. However, increasing modulation index is limited by the available voltage which is influenced by the operating points of P and Q as shown by Figure 5.7.a for modulation index of 0.85 and 0.9 and $-Q$.



M curve of rated stored energy, modulation index and power factor

Figure 5.6: Relationship between rated stored energy, power factor and modulation index.

Considering available and synthesised voltages, Figure 5.7 highlights the issue related to operating the MMC with higher modulation index compared to Figure 5.5 where two different SM capacitor sizes are considered. Two possible solutions can be implemented to overcome the overmodulation issue considering the same size of the SM capacitor used in Figure 5.5.a. The first solution is to operate with reduced modulation index as demonstrated by the example in Figure 5.7.a where the worst case scenario of the available voltage is considered ($-Q$). In this example, the modulation index is reduced from 0.9 to 0.85 in order to achieve correct operation. However, this leads to increased line currents and so power losses can be increased for the same rated power.

The Second possible solution is that 3rd harmonic voltage injection can be used to increase the margin between the available and synthesised voltages by reducing the peak of the latter. Figure 5.7.b illustrates the second option where the available voltage due to fundamental alone and by applying 3rd harmonic component are denoted by $V_{pos,f1}$ and $V_{pos,f13}$ respectively. The waveforms of the example presented by Figure 5.7.b shows that peak of the synthesised voltage denoted by the dotted green line is reduced to the solid green line for the same modulation index ($m_i = 0.9$) while correct operation

is achieved. In fact, this can be used to increase the fundamental component above the nominal while maintaining a peak synthesised voltage less than the available voltage. This can lead to a reduction of line currents and a resultant power losses reduction. More discussion about this is presented in Chapter 6.

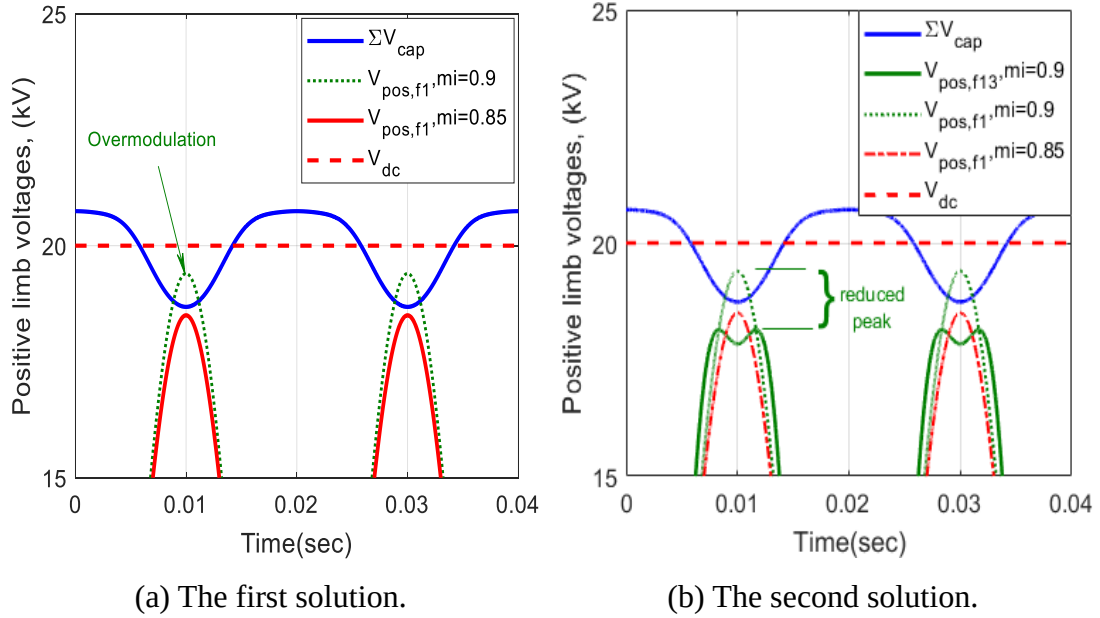


Figure 5.7: Two possible solution to prevent overmodulation due to high modulation index for the same size of SM capacitor; (a) reducing modulation index from 0.9 to 0.85, (b) use 3rd harmonic injection.

Based on the previous analysis, it is observed that using both classical and phase-shifted 3rd harmonic voltage injection can lead to reduction to SM capacitor size. This reduction, however, depends on the power factor and the modulation index and therefore can only be beneficial with a specific range of P and Q as discussed earlier.

5.5 Summary and Conclusion

From the previous analysis, it is observed that, an appropriate design of MMC also needs to consider the maximum modulation index which can be limited by the available voltage depending on the operation point of the MMC represented by the power factor variation, PQ envelope. Therefore, it is necessary to optimise between as high as possible modulation index and as small capacitor size as possible in order to meet the energy requirements. The outcome of the analysis can be summarised in Table 5.1 below where two values of 3rd harmonic voltage injection, β , is considered, 15% and 50% of the fundamental.

From an energy deviation point of view; the third harmonic voltage injection cannot be considered as an achievement as the capacitor size reduction is only approximately ($\approx 7\%$) due to 50% of third harmonic injection. This is because injecting such a percentage of 3rd harmonic can lead to an increase in the peak synthesised voltage above the limit. Moreover, considering the practical percentage of 3rd harmonic injection of approximately 15%, the total reduction is about 2.2%. Moreover, this reduction occurs only when operating with only active power, $\pm P$. However, there is a range of 3rd harmonic injection where the peak synthesised voltage can be limited where the peak synthesised voltage can be reduced whilst the fundamental component can be increased.

Table 5.1: The MMC applications based on the energy deviation reduction and capacitor sizing using 3rd harmonic injection voltage/current.

Energy deviation Statues			Range of reduction	Reduction/increase at ($\emptyset$)		3 rd harmonic	% reduction/increase
				Maximum	Zero	β	Maximum
3 rd V	classic	Reduction	$-30 \leq \emptyset \leq 30$	0^o	$\pm 30^o$	50% 15%	7% 2%
		Increase	$-90 < \emptyset < -30$ $+30 < \emptyset < +90$	$\approx 53^o$ & -53^o	$\pm 90^o$	50% 15%	3.9% 1.0%
	Phase-shifted	Reduction	$-30 \leq \emptyset \leq 30$	0	$\pm 30^o$	50% 15%	7% 2%
		Increase	$-90 < \emptyset < -30$ $+30 < \emptyset < +90$	$\approx 53^o$ & -53^o	$\pm 90^o$	50% 15%	3.9% 1.0%

Therefore, instead of considering 3rd harmonic to reduce energy storage requirement, the former can be used to reduce the peak synthesised voltage. This reduction can be beneficial in two possible ways including;

- First, reducing the peak synthesised voltage can be used to increase the fundamental component of voltage at the MMC terminals. Moreover, increasing the fundamental component can lead to reduce energy deviation as discussed in Chapter 3 and this chapter.
- The second possible option is using the reduced synthesised voltage to increase the safe margin between the synthesised and the available voltage. Moreover, the reduced synthesised voltage can be used to reconsider the peak to peak available voltage by reconsidering its ripple. This means that the SM capacitors can be replaced with smaller sized ones.

Therefore, the focus in Chapter 6 is on an improved operation and design analysis based on the two proposed methods.

In this chapter, a review of capacitor sizing methods was presented. An investigation on the impact of 3rd harmonic voltage injection on SM capacitor size reduction has been discussed.

In Chapter 6, further investigation of using both optimal classical and phase-shifted 3rd harmonic voltage injection on the reduction of the limb peak synthesised voltage are presented. The advantage of the reduced peak synthesised voltage are further considered proposing two different methods including increasing the output voltage at the MMC terminals or reducing the size of SM capacitor.

Chapter 6

6 Optimal Third Harmonic Voltage Injection

6.1 Introduction

In Chapter 5, a review of the submodule capacitor sizing method based on intra-cycle energy deviation was discussed considering first the standard operation of the MMC. The same approach was then applied to the MMC operating with 3rd harmonic voltage injection, and a comparison between the two cases was discussed. The outcome was that from energy deviation point of view 3rd harmonic voltage injection does not lead to a significant reduction in the size of the submodule capacitors.

In this Chapter, an investigation will be carried out to consider 3rd harmonic voltage injection and its impact on the operation of the MMC and on its design aspects. The advantage of using optimal 3rd harmonic voltage injection on reducing the peak of the synthesised voltage will be investigated for HVDC transmission applications. The injection of 3rd harmonic voltage can optimise different aspects of the converter operation, such as voltage margin, ripple or losses. Therefore, following the results from the previous chapters, showing that the significant impact of 3rd harmonic voltage injection is only on reducing the peak synthesised voltage, this Chapter provides an in-depth analysis of the impact of 3rd harmonic voltage injection on the available voltage margin.

The proposed investigation will consider the advantage of the peak synthesised voltage reduction from two different perspectives. Firstly, the advantage of reducing the peak value of the synthesised voltage will be used in order to increase the output voltage at the MMC terminals. As an alternative, for a given nominal output voltage of the MMC, the approach presented in Chapter 5 to select the size of capacitors can be revisited by considering a larger value of SM voltage ripple, leading to smaller capacitors. This larger ripple can be tolerated due to the reduction of the peak synthesised voltage.

In summary, the reduction of the peak synthesised voltage can be exploited in two different ways:

- Increasing the output voltage at the MMC terminals due to the increase of the fundamental component of the synthesised voltage. This will lead to lower current through the converter, with proper scaling of the transformer turns ratio. This approach requires a dedicated design of the conversion system, and its applicability is therefore limited to new installations.
- Reducing the size of the SM capacitor by allowing higher voltage ripple. This second option considers a given design of the MMC and only revisits the size of the SM capacitors by exploiting the increased available voltage margin after 3rd harmonic injection.

The main objective of this chapter is to define the optimum 3rd harmonic voltage component in order to obtain an optimal peak reduction of the synthesised voltage for the highest fundamental component as possible. Therefore, a review of the concept of optimal 3rd harmonic voltage injection based on the voltage margin for a generic VSC will first be presented. The same approach will be subsequently applied to the MMC considering classical and phase-shifted 3rd harmonic voltage injection.

6.2 Optimal Third Harmonic Injection for MMC

The use of harmonic component injection to increase the output voltage in power converters was first introduced for three-phase PWM inverters. It was reported that the use of 3rd harmonic injection for 2-level VSCs can lead to reduced peak voltage where the optimal value of 3rd harmonic is 1/6 of the fundamental component and therefore the fundamental component can be increased by approximately 15% [70].

In order to qualitatively highlight the importance of using optimal 3rd harmonic injection let us consider Figure 6.1. The corresponding waveforms of the synthesised and available voltages of the positive limb are shown by Figure 6.1 based on the MMC parameters given by Table 6.1 where the relationship between the DC voltage and SM capacitor voltage is given by (3.2), e.g. $V_{CSM} = \frac{V_{dc}}{N_{SM}} = 20kV/10 = 2000V$. According to the convention described in Appendix A, in Figure 6.1, the MMC operates with $P > 0$ and $Q < 0$ with values of $0.7 p.u$ and $-0.7 p.u$ respectively. To provide a visual representation of the impact of 3rd harmonic voltage injection, two injections are considered; 50% and 15%. It can be seen that when the amplitude of the 3rd harmonic is 50% of the fundamental, the MMC enters overmodulation mode due to the increased

peak synthesised voltage, causing $V_{pos}(t) > \sum V_{cap}$. It is worth remembering that the available voltage is less affected by the same amount of 3rd harmonic component due to the negligible reduction it causes in the energy deviation, as discussed in Chapter 4.

Table 6.1: Circuit parameters of MMC.

Rated output power	P	20 MW
Nominal AC system voltage	V_{ac}	11 kV
Nominal converter AC voltage	V_c	10.65 kV
Converter transformer turns ratio	r_t	0.596
DC voltage	V_{dc}	20 kV
Number of SM	N_{SM}	10
Nominal voltage of SM capacitance	V_{cap}	2000 V
SM capacitance	C_{SM}^*	10.07 mF
Inductance of limb inductor	L_{pos}^{**}	2.71 mH
Limb resistance	R_{pos}^{**}	0.0 Ω
AC system inductance	L_T	3.25 mH
AS system resistance	R_T	2.5 m Ω

*The capacitor nominal voltage is given by Equation (3.2).

**The inductance and resistance of positive and negative limb inductors are equal [77].

It can be seen that, using 15% 3rd harmonic injection in contrast to 50%, the operation can be assured with the same fundamental component for the same modulation index of 0.95. However, the energy deviation reduction is negligible which is about 2% in case of 15% 3rd harmonic voltage injection in contrast to 7% reduction in the second case of 50% injection as discussed in Chapter 4. Therefore, from the peak synthesised and available voltage point of view, using a higher percentage of 3rd harmonic voltage injection can cause conflict with the essential operating condition of the MMC given by (3.6).

Two possible solutions can be found to overcome the issue of entering overmodulation when the MMC operates with high percentage of 3rd harmonic voltage injection, as illustrated by Figure 6.2. These include:

- The operation with a lower modulation index.
- Using a smaller percentage of 3rd harmonic voltage injection.

It is observed that using a high 3rd harmonic voltage injection ratio would require operation with a low modulation index. Therefore, as a result, the fundamental component of the output voltage at the converter AC terminals would be reduced. The operation with a low modulation index means higher energy deviation as shown by Figure 4.17. This can lead to increase peak-to-peak available voltage variation as illustrated by the comparison shown in Figure 6.2 where modulation indices of 0.95 and 0.75 are used applying 15% and 50 3rd harmonic voltage injection respectively. The dotted line represents the operation with modulation index of 0.95 and 3rd harmonic injection of 15% while the solid line is for 0.75 and 50%. Of course it is assumed that the converter transformer ratio is adapted in order to generate the same AC voltage on the grid side.

Figure 6.1 and Figure 6.2 illustrate different operation and design scenarios while the rated power $|S|$ is kept constant and the same for both cases. For each modulation index, the transformer turns ratio is considered constant for all the different operating conditions with regards to the 3rd harmonic voltage injection, 15% and 50%. In the second figure, however, two values of the transformer turns ratio are considered based on m_i values of the fundamental component of 0.75 and 0.95, with two different operation conditions, 50% and 15%. It can be seen that the used 3rd harmonic injection used in both cases ensures operation where $V_{pos}(t) < \sum V_{cap}$.

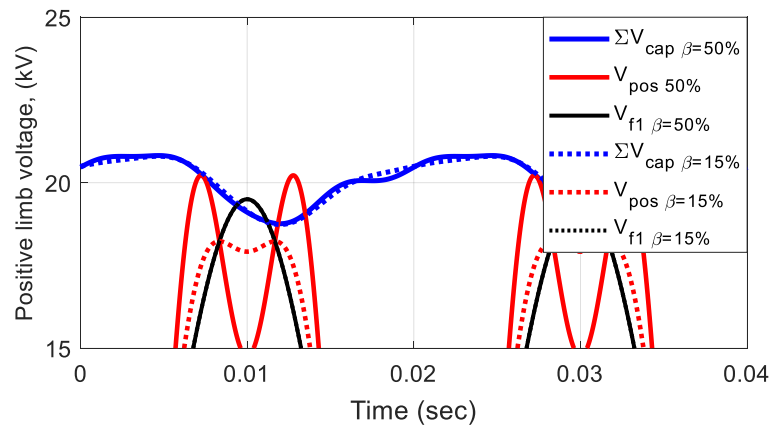


Figure 6.1: Overmodulation due to high 3rd harmonic injection.

Therefore, the optimum 3rd harmonic voltage component needs to be defined in order to obtain an optimal peak reduction with the highest fundamental component possible. The next section presents the optimal value of the 3rd harmonic voltage injection for the MMC considering both classical and phase-shifted 3rd harmonic injection. The concept

of classical and phase-shifted 3rd harmonic injection has already been introduced in Chapter 4 and in Appendix B.

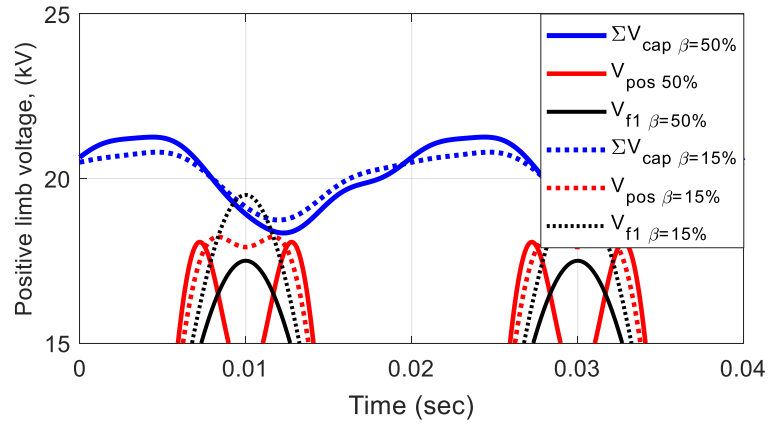


Figure 6.2: The synthesised voltage due to the reduction of modulation index by 21% (from $m_i = 0.95$ dashed red to 0.75 solid red) and by high 3rd harmonic component injection.

6.3 Calculation of Optimum Third Harmonic Injection

Although the in-phase 3rd harmonic components lead to a reduced peak of the phase voltage, they cancel each other in the line-to-line voltage. Due to this peak reduction, 3rd harmonic voltage injection can be used to increase the output voltage of VSC converters as reported in [71]. Since the use of the optimum amount of the 3rd harmonic injection was presented in [70], it was widely used also for MMCs [16, 17, 49, 54, 78]. Therefore, in this chapter, the optimal 3rd harmonic voltage injection for the MMC is discussed. The classical 3rd harmonic injection is first presented where the phase of the 3rd harmonic component is zero (referred later as Zero Phase Shifted 3rd harmonic). Then, the same analysis method is used to explore and discuss the advantages of a phase-shifted 3rd harmonic voltage injection. Due to the symmetry of the converter limbs, the analysis is considered for the positive limb of phase A.

The positive limb voltage is expressed by (4.11) in Chapter 4. After manipulation, it can be rewritten as (6.1) for the injection of a generic 3rd harmonic voltage where the amplitude of the 3rd harmonic component is given as a percentage of the fundamental component expressed by (4.28) while the fundamental component is expressed by (4.25). Figure 6.3 illustrates the synthesised positive limb voltage showing the three possible modulation states, fundamental (solid blue), zero phase-shifted (dotted red) and

phase-shifted (solid red) 3rd harmonic injection where the inequality in (6.2) should be applied to ensure the correct operation.

$$V_{pos.a}(t) = \frac{V_{dc}}{2} (1 - m_i \cos(\omega t) + \beta m_i \cos(3\omega t - \theta_{v3})) \quad (6.1)$$

$$0 \leq V_{pos}(t) \leq \sum v_{c_{SM}}(t) \quad (6.2)$$

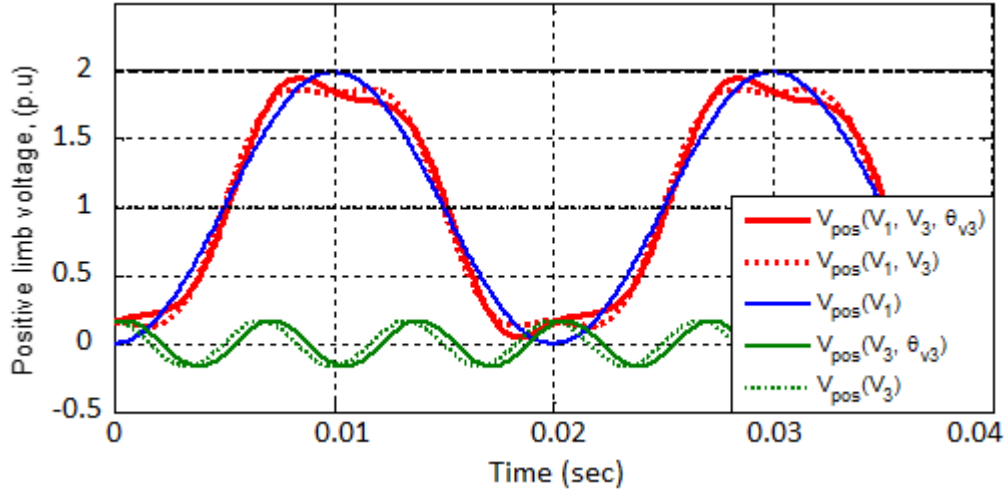


Figure 6.3: Synthesised limb voltage including; fundamental, and 3rd harmonic frequency.

6.3.1 Zero Phase Shifted Third Harmonic Injection

In this first analysis, the phase shift of the 3rd harmonic component is zero. Therefore, θ_{v3} in (6.1) is set to be zero. This represents the classical approach followed in 2-level VSCs, applied to the MMC. The analysis is well known but is reported here for completeness and to simplify the comparison with the phase shifted 3rd harmonic injection that will be investigated later.

$$V_{pos.a}(t) = \frac{V_{dc}}{2} (1 - m_i \cos(\omega t) + \beta m_i \cos(3\omega t)) \quad (6.3)$$

For more simplification, (6.3) is expressed by (6.4) where $\omega t = \theta$.

$$V_{pos.a}(\theta) = \frac{V_{dc}}{2} (1 - m_i \cos(\theta) + \beta m_i \cos(3\theta)) \quad (6.4)$$

Using the trigonometry identity (6.5) in (6.4), the positive limb periodic voltage then can be written as (6.6).

$$\cos 3\theta = 4 \cos^3 \theta - 3 \cos \theta \quad (6.5)$$

$$V_{pos.a}(\theta) = \frac{V_{dc}}{2} y(\theta) \quad (6.6)$$

where

$$y(\theta) = 1 - m_i(1 + 3\beta) \cos(\theta) + 4\beta m_i \cos^3(\theta) \quad (6.7)$$

The optimum value of β can be determined firstly by locating the turning points of the function in (6.7). This can be found by the differentiation of $y(\theta)$ with respect to θ and then equating the result to zero. Thus

$$\begin{aligned} \frac{dy(\theta)}{d\theta} &= (0 + (1 + 3\beta)m_i \sin(\theta) - 12m_i\beta \sin \theta \cos^2(\theta)) = 0 \quad (6.8) \\ (1 + 3\beta)m_i \sin(\theta) - 12m_i\beta \sin \theta \cos^2(\theta) &= 0 \end{aligned}$$

Two possible solutions for (6.8) with respect to θ can be given by the following equations. The first solution is given by the following equation:

$$\sin(\theta) = 0 \quad (6.9)$$

Subsequently, from (6.9), cosine angle of θ is given by;

$$\cos(\theta) = 1 \quad (6.10)$$

The second solution of Equation (6.8) is given by the following equation:

$$\cos(\theta) = \left(\frac{1 + 3\beta}{12\beta} \right)^{1/2} \quad (6.11)$$

In order to determine the maxima of function $y(\theta)$ given by (6.7), equations (6.12) and (6.13) are obtained by substituting (6.10) and (6.11) into (6.7), respectively.

$$\check{Y}_1 = 1 - (1 - \beta)m_i \quad (6.12)$$

$$\hat{Y}_2 = 1 - \left(8\beta \left(\frac{1 + 3\beta}{12\beta} \right)^{3/2} \right) m_i \quad (6.13)$$

where $\check{Y}_1$ and $\hat{Y}_2$ are the peak values of $y(\theta)$. Subsequently, the optimum value of β can be achieved by differentiation of equations (6.12) and (6.13) with respect to β and

equating the result to zero. As can be seen, no optimum value can be obtained from the first equation, $\check{Y}_1$, as it is a linear equation. Instead, from the second equation:

$$\frac{d\hat{Y}_2}{d\beta} = \frac{d}{d\beta} \left(1 - \left(8\beta \left(\frac{1+3\beta}{12\beta} \right)^{3/2} \right) m_i \right) = 0 \quad (6.14)$$

Solving (6.14) for β , two possible roots are obtained as shown (6.15) and (6.16). The negative value is considered more beneficial if also a 3rd harmonic current injection is performed, as analysed in Chapter 4 for the MMC and also for other topologies as reported by [56]. However, applying the negative value in case of 3rd harmonic voltage injection will lead to increase peak of the synthesised voltage while the required optimum value is to reduce the peak of the synthesised voltage. Hence, in the analysis proposed in this Chapter, the only value considered for β is equal to 1/6. This result is the same as expected and reported for 2-level VSC by [70].

$$\beta = -\frac{1}{3} \quad (6.15)$$

$$\beta = \frac{1}{6} \quad (6.16)$$

It can be seen that the optimum value of β does not depend on modulation index m_i . This means that the optimal value in (6.16) is always correct for all m_i range and also for all operation points of P and Q as soon as the essential condition given by Equation (3.6) is respected.

The peak values of (6.6) at the optimal value of β can be determined by substituting Equation (6.16) into (6.13). Figure 6.4 shows the maximum and minimum values of the positive limb synthesised voltage as expressed by (6.17) and (6.18) for the ideal MMC where the m_i is 1.0 and the base voltage is V_{dc} .

$$\check{V}_{pos} = \frac{V_{dc}}{2} \left(1 - \frac{\sqrt{3}}{2} m_i \right) \quad (6.17)$$

$$\hat{V}_{pos} = \frac{V_{dc}}{2} \left(1 + \frac{\sqrt{3}}{2} m_i \right) \quad (6.18)$$

The maximum peak due to the fundamental is given by;

$$\hat{V}_{pos.1} = \frac{V_{dc}}{2} (1 + m_i) \quad (6.19)$$

The maximum peak voltage reduction for any design point of m_i caused by the 1/6 of 3rd harmonic voltage component is given by subscribing (6.18) from (6.19) which is expressed by:

$$\hat{V}_{red} = \hat{V}_{pos}^1 - \hat{V}_{pos}^{13} = (1 - \frac{\sqrt{3}}{2})m_i \frac{V_{dc}}{2} \quad (6.20)$$

where $\hat{V}_{pos}^1$ and $\hat{V}_{pos}^{13}$ are the peak values of the synthesised voltages without and with 3rd harmonic voltage injection respectively.

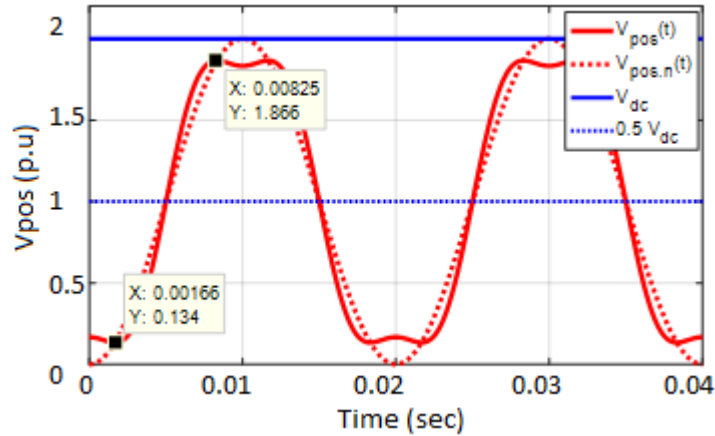


Figure 6.4: Maximum and minimum p.u. values of positive limb synthesised voltage due to 1/6 of 3rd harmonic injection and $V_{base} = V_{dc}$ and $m_i=1.0$.

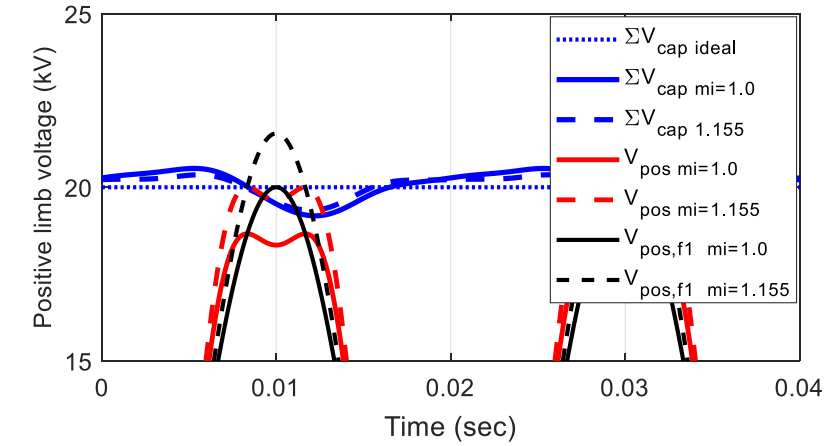
Ideally, this amount of reduction can be used to increase the peak synthesised voltage by the same amount, $0.134m_i \frac{V_{dc}}{2}$. As a result the new peak of the synthesised voltage will be equal to the previous peak of the fundamental, but the new fundamental voltage will increase. This leads to the definition of a new modulation index. Consider equations (6.18) and (6.19) where the modulation index in (6.18) can be increased and expressed by $m_{i.new}$. The new peak synthesised voltage after increasing the modulation index becomes equal to (6.19). After manipulation, this yields:

$$\hat{V}_{pos.new} = \hat{V}_{pos.1} \quad (6.21)$$

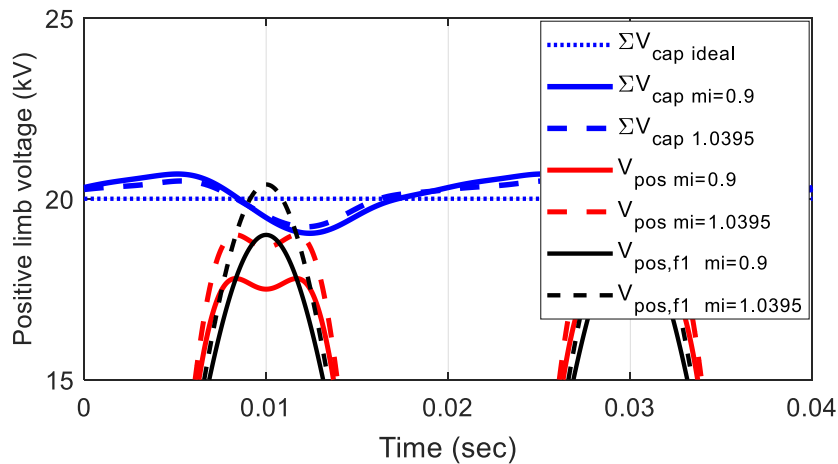
$$\frac{\sqrt{3}}{2} m_{i.new} = m_i \quad (6.22)$$

$$m_{i.new} = \frac{2}{\sqrt{3}} m_i \quad (6.23)$$

Using 1/6 of third harmonic can lead to an increase in the output AC voltage of the MMC terminals by 15.5% due to an increase of the modulation index by the same percentage. Again, not surprisingly, this is the same result given by [70]. However, this result can only be applied for the ideal MMC as discussed in Table 3.1 and assuming that the modulation can use the entire voltage range from zero to the available voltage corresponding to the DC link voltage. Therefore, ideally, the nominal modulation index m_i can be increased from 1.0 to 1.155.



(a) $m_i=1.0$ and 1.155 at $S=+0.91+j0.41$ p.u



(b) $m_i=0.9$ and 1.0395 at $S=+0.91+j0.41$ p.u

Figure 6.5: The effect of m_i on to MMC synthesised voltage compared to available voltage.

However, with regards to the practical MMC, the modulation index cannot reach unity and is also limited by the available voltage as shown by Figure 6.5.a. The corresponding waveform depends on many factors and the most important one is the PQ operating point which affects the available voltage ripple. Figure 6.5 presents a comparison of the positive limb voltages for the ideal and non-ideal MMC given by Tables 3.1 and Table 3.2 respectively.

The simulation results are obtained for a constant $|S|$ with $P=0.94$ p.u. and $Q=-0.33$ p.u with a size of capacitor of 10.07mF selected based on a voltage ripple of 10% (10% peak to peak). This represents the worst case as the MMC operates at rated $|S|$, maximum energy deviation and the minimum available voltage.

It can be seen that the factor of 1.155 cannot be applied for converters designed with high modulation index due to overmodulation as shown by Figure 6.5.a, especially for the same P and Q envelope. However, the modulation index of the fundamental can be increased by 15.5% for an MMC designed to operate with modulation indices as high as 0.9. It can be seen that the initial 0.9 can be increased 15% up to 1.0395 as shown in Figure 6.5.b. Phase-shifted 3rd harmonic can give better performance as illustrated by Figure 6.7.

Therefore, considering the waveforms in the figures below and to ensure operation, Equation (6.23) can then be rewritten as the inequality (6.24).

$$m_{i.new} < 1.155m_i \quad (6.24)$$

From Figure 6.5 it can also be seen that available voltage ripple is reduced in both cases due to the increase of the modulation index as discussed in Chapter 3 and Figure 3.7 due to the reduction of energy deviation. Figure 6.6 illustrates a comparison between five operating points considering the variation of m_i as discussed above alongside Figure 6.5 where $|S|$ is considered constant with $P=+0.94$ p.u and $Q=-0.33$ p.u. Obviously, to obtain such an operating condition it must be assumed that the transformer turns ratio is adapted to the modulation index in order to maintain the AC voltage of the grid. It is noted that the initial converter transformer turns ratio applied to achieve modulation index of 0.9 is grater than that of 1.15 ($r_{t.ini,0.9} > r_{t,1.15}$).

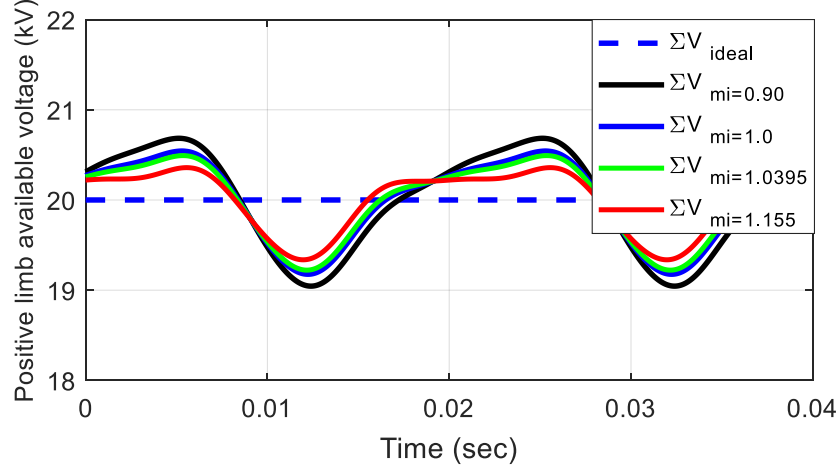


Figure 6.6: The available voltage of MMC for different m_i (0.9, 1.0, 1.039 and 1.15) with converter transformer turns ratios ($r_{t,ini,0.9} > r_{t,1.0} > r_{t,1.039} > r_{t,1.15}$) for $P = +0.94\text{p.u}$ and $Q = -0.33\text{p.u}$.

6.3.2 Phase-Shifted 3rd Harmonic Voltage Injection

Previous analysis has shown that the optimum zero phase shifted 3rd harmonic voltage injection is one-sixth of the fundamental component as expected and reported for conventional 2-level VSCs and for the MMC as reported in [17, 70]. The impact of the phase angle of the 3rd harmonic is examined in this section. The special solution of Equation (6.1) was previously discussed as the phase shift of the third harmonic was set to zero. This solution becomes more complicated if $\theta_{v3} \neq 0$ is considered.

Example of waveforms of the positive limb synthesised voltage in Figure 6.7 provide a qualitative understanding of the effect of both classical and phase-shifted 3rd harmonic voltage injection on increasing the fundamental component of the synthesised voltage. The waveforms are obtained based on the same operation specification used in Figure 6.5 and considering zero phase shifted and phase-shifted 3rd harmonic.

The zero phase shifted 3rd harmonic could satisfactorily allow correct operation with m_i when increased by 15.5% from 0.9. Using the phase-shifted 3rd harmonic can increase the safety margin between the synthesised and available voltages as shown by Figure 6.7. However, in another case where the nominal modulation index is greater than 0.9, then phase-shifted 3rd harmonic must be applied in order to operate with the maximum possible increase given by (6.24). On the other hand, not all 3rd harmonic phase angles

can be considered a solution to (6.24) as illustrated by Figure 6.8 where the upper and lower limits are exceeded.

The dashed black waveform is the fundamental component of the nominal voltage, whilst the solid black is the fundamental of the synthesised voltage due to the increase of 15% of the nominal. The solid red line is the waveform of the synthesised voltage with zero-phase due to the increase of m_i to 15%. The dashed red lines illustrate the waveforms of the synthesised voltage for the phases-shifted 3rd harmonic voltage injection due to increasing of the m_i to 15% whilst applying different values of phase angle of the 3rd harmonic component, θ_{v3} . It can be seen that the upper and lower limit of the synthesised voltage are exceeded due to excessive phase angle of the 3rd harmonic injection where the limit is given by Equation (3.6).

The modulation index, m_i , is determined by the MMC operating point of P and Q. For any given β , the optimum phase shift needs to be determined for each value of β with respect to different values of m_i which forms an equation with four variables, t , m_t , β and θ_{v3} . A solution of such a problem of one equation with four variables is difficult to be obtain. Therefore, for the simplicity, β is set to be 1/6 which represents the classical phase-shifted 3rd harmonic in the previous section despite the fact that this might not be the optimum solution.

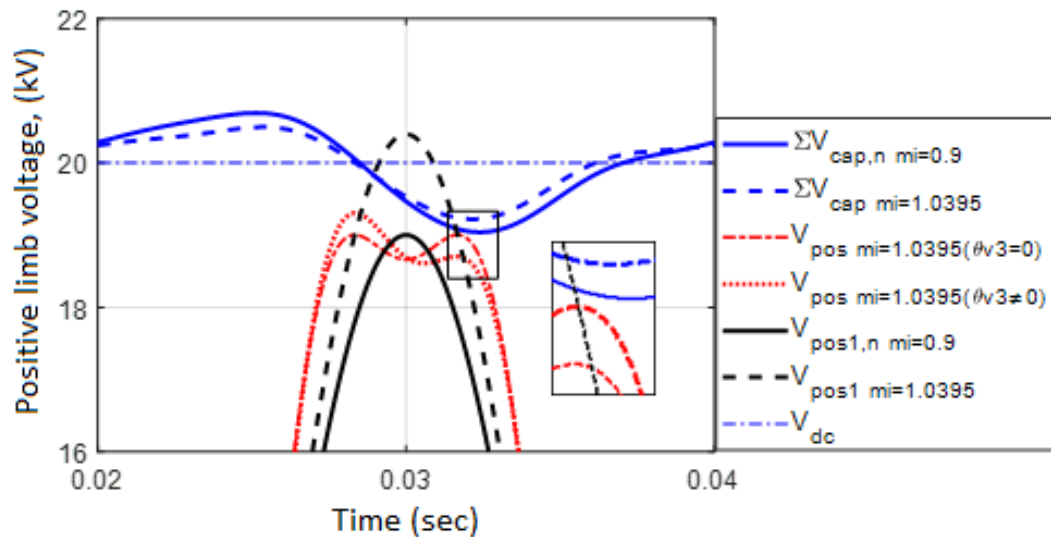


Figure 6.7: The effect of optimum zero phase shifted and phase-shifted 3rd harmonic components on increasing fundamental component.

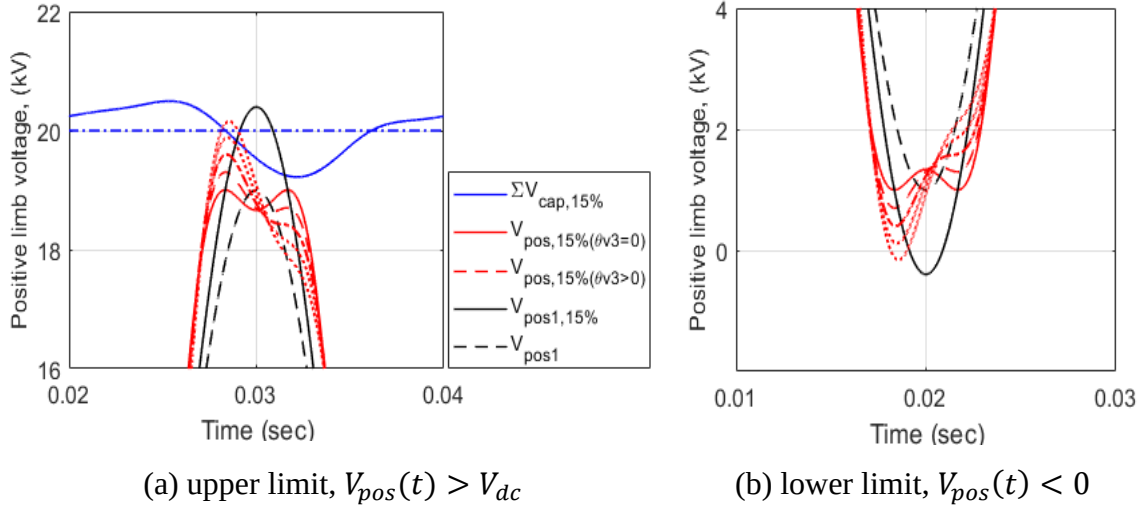


Figure 6.8: The impact of θ_{v_3} variation on the peak synthesised voltage limits for

$$\theta_{v_3} \geq 0^\circ, m_{i,1} = 0.9 \text{ and } m_{i,1,15\%} = 1.0395.$$

To evaluate the optimum phase for the phase shifted 3rd harmonic injection, both the fundamental component and 3rd harmonic amplitudes, expressed by m_i and β in (6.1), are considered known and therefore they are constant values. The limb voltage waveform in (6.1) can be expressed by $y(t)$ in (6.25).

$$y(t) = 1 - m_i \cos(\theta) + \beta m_i \cos(3\theta - \theta_{v_3}) \quad (6.25)$$

where $\theta = \omega t$. Using the identity:

$$\cos(3\theta - \theta_{v_3}) = \cos 3\theta \cos \theta_{v_3} + \sin 3\theta \sin \theta_{v_3} \quad (6.26)$$

Then applying the identities:

$$\cos 3\theta = 4 \cos^3 \theta - 3 \cos \theta \quad (6.27)$$

$$\sin 3\theta = 3 \sin \theta - 4 \sin^3 \theta \quad (6.28)$$

$$y(\theta) = 1 - m_i \cos \theta + m_i \beta ((4 \cos^3 \theta - 3 \cos \theta) \cos \theta_{v_3} + (3 \sin \theta - 4 \sin^3 \theta) \sin \theta_{v_3}) \quad (6.29)$$

Differently from (6.7), an explicit solution cannot be found for (6.29). Therefore, a numerical solution is considered using MATLAB. It is defined that the optimum value of the 3rd harmonic ratio β is the value which can be used to minimise the peak

synthesised voltage, $V_{pos}(t)$. This definition cannot be satisfactorily applied with (6.29). This is because the inclusion of θ_{v3} in this case leads to an increase of peak value of the synthesised voltage for the same values of m_i and β as shown in Figure 6.3. Moreover, in order to maintain the peak of the synthesised voltage to the previous one ($\hat{V}_{pos}(\theta_{v3} \neq 0) = \hat{V}_{pos}(\theta_{v3} = 0)$), then operation with a reduced modulation index is required, where $m_i(\theta_{v3} \neq 0) < m_i(\theta_{v3} = 0)$.

Therefore, the optimal phase-shift of the 3rd harmonic injection will be defined as the phase of the 3rd harmonic component which allows operating with as high modulation index as possible for the same PQ envelope and the same capacitor size, so that the essential operation condition in Equation (3.6) is still applied. In addition, a simplified assumption is made considering a fixed amplitude for the injected 3rd harmonic voltage, corresponding to the optimum value previously calculated with zero phase shift. It must be noted that this might not be the exact optimum when the variation of phase angle of the 3rd harmonic is considered, but results demonstrate that good improvements can be achieved also with this simplified assumption.

Considering the inequality in (6.2), Equation (6.29) is numerically solved for θ_{v3} , for each given m_i , considering the variation of P and Q so that the essential operation condition given by Equation (3.6) or the inequality (6.2) is still applied. This is in order to determine the optimum solution for the phase angle of the 3rd harmonic θ_{v3} with the variation of P, Q for each selected modulation index m_i .

In practice, the converter will be connected to the AC grid through a line reactor, X_L , typically corresponding to the transformer leakage inductance. The modulation index is generated by the controller which will control the current in the AC network in order to track the P and Q demands. The connection to the grid through an inductive impedance gives a direct relation that enables the calculation of the required modulation index for each set of P, Q and grid voltage amplitude. In this work, several modulation indexes are considered for each P and Q, so that a superset of operating conditions is explored. This guarantees generality of the analysis. Therefore, for any operating point of m_i and P and Q, the right θ_{v3} will be selected.

6.4 Phase Shifted Third Harmonic Limits

In this section, the ideal MMC converter is considered first where the MMC is assumed to operate with no ripple in the SM capacitors. Then the practical MMC with non-zero ripple is considered.

6.4.1 Ideal MMC Converter

Considering the active and reactive power curve of the MMC shown by Figure 3.3, the idea demonstrated by Figure 6.8 shows that not all phase shift of the 3rd harmonic component can be applied due to the upper and lower limits expressed by equation (6.3) being exceeded, especially when operating in inductive mode (Q₂ or Q₃).

It is observed that the upper limit is considered the most important factor to determine the correct operation of the MMC as the upper limit can be hit before the lower limit (the entry in overmodulation) due to the increase of θ_{v3} . On the other hand, for instance, the lower limit can be hit before exceeding the upper limit (entry in overmodulation) in case of operating the MMC in capacitive mode (Q₁ or Q₄) as illustrated by the example in Figure 6.9.

For instance, waveforms of the available and synthesised voltages in Figure 6.9 illustrates operating the MMC with $m_i=0.97$ and a rated apparent power while $\phi = 22.3^\circ$, i.e. Q₁, and two different phase shifts, $\theta_{v3} = 40^\circ$ and 30° . It can be seen that the peak synthesised voltage ($\check{V}_{pos} \leq 0$) exceeds the lower limit when $\theta_{v3} = 30^\circ$ whilst the MMC can still operate if only the upper limit ($\hat{V}_{pos} < \sum V_{cap}$) is considered as shown by Figure 6.9.a. However, the lower limit is exceeded in both cases when $\theta_{v3} = 30^\circ$ and 40° as shown in Figure 6.9.b. Therefore, not all phase shift, θ_{v3} , can be applied where the both limits need to be considered especially when investigating the optimum phase shift during the operating with high modulation index.

Assuming the operation with an ideal MMC, the upper and lower limits in Equation (6.2) can be re-expressed by (6.30) where the available voltage is given by (3.42). As mentioned earlier, for the ideal MMC the available voltage is given by V_{DC} and the synthesised voltage, $V_{pos}(t, m_i, \theta_{v3})$, is given by (6.1).

$$0 \leq V_{pos}(t, m_i, \theta_{v3}) \leq \sum v_{cSM}(t, m_i, \theta_{v3}) \quad (6.30)$$

$$0 \leq V_{pos}(t, m_i, \theta_{v3}) \leq V_{DC} \quad (6.31)$$

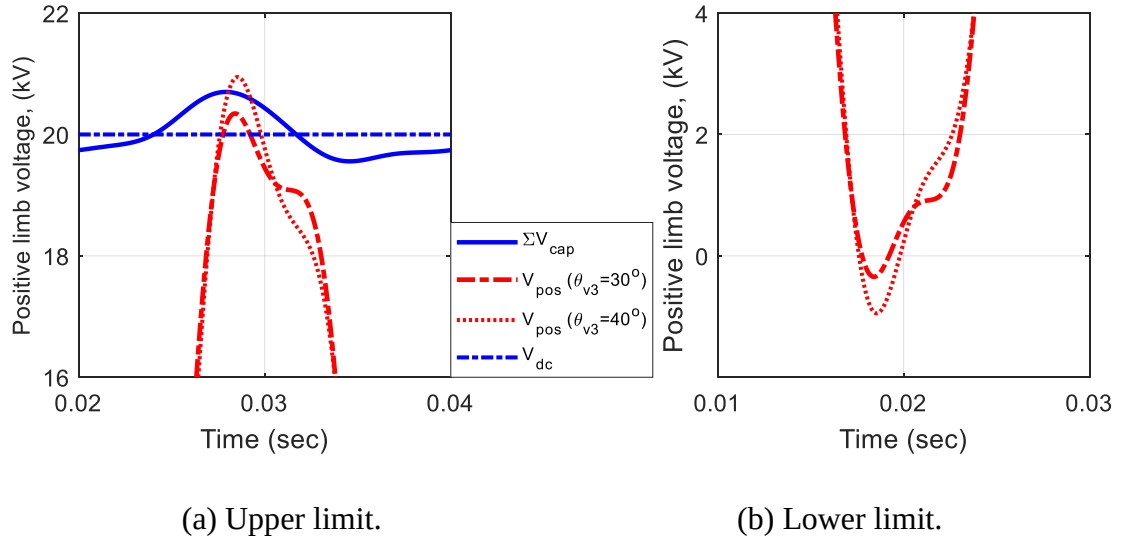
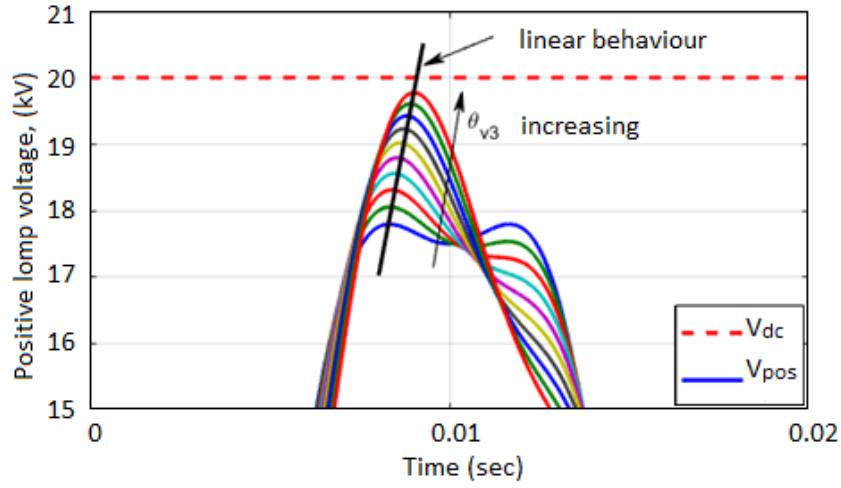
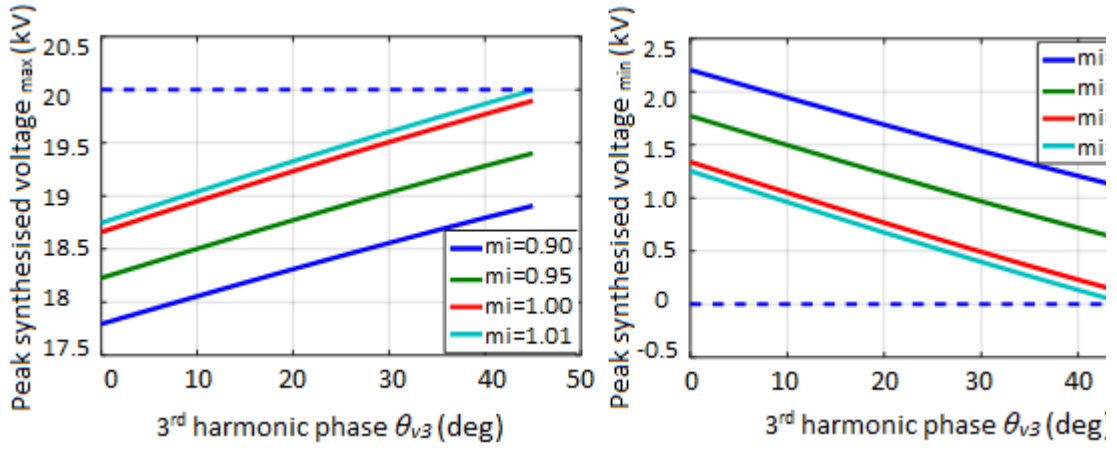


Figure 6.9: The impact of phase shift on exceeding the lower limitation before the entry in overmodulation when operating in Q_1 .

By solving the inequality in (6.31), the 3rd harmonic phase can be determined. To simplify the solution of determining the phase shift limits, first it is important to find a relationship between the modulation index, peak synthesised voltage and the phase shift of 3rd harmonic component as illustrated by Figure 6.10. This can be done by considering Figure 6.11 where the relationship between the maximum and minimum peak synthesised voltages (Figure 6.11.a and Figure 6.11.b) respectively and the 3rd harmonic phase shift can be determined for different values of modulation index using (6.1). The data used in Figure 6.11 represents the peak of the synthesised voltage with the variation of θ_{v3} as demonstrated by Figure 6.10.

It can be seen that as the modulation index is increased, the peaks of the synthesised voltage for the upper and lower limits are respectively increased and reduced as shown by Figure 6.11.a and Figure 6.11.b. It is noted that the relationship between the peak of the synthesised voltage and θ_{v3} is the same for both positive/negative phase shifted 3rd harmonic injection where the positive and negative 3rd harmonic phase shift is illustrated by Figure 6.12. Therefore, the following analysis considers the positive phase-shifted 3rd harmonic.


 Figure 6.10: Peak synthesised voltage in relation to θ_{v3} for m_i of 0.9.


$$(a) \hat{V}_{pos} = f(m_i, \theta_{v3})$$

$$(b) \check{V}_{pos} = f(m_i, \theta_{v3})$$

 Figure 6.11: Peak synthesised voltage in relation to θ_{v3} for different values of m_i .

Using the MATLAB Curve Fitting tool for each value of modulation index, each of the individual linear behaviour in Figure 6.11 can be expressed by an individual equation for one particular modulation index. For instance, the Red line in Figure 6.11 is obtained for m_i of 1.0 where the downstream and upstream behaviour is given in per unit by (6.34) and (6.35) respectively. For a zero phase-shift 3rd harmonic, the minimum and maximum values are 0.1327 p.u and 1.866 p.u which gives the same result using (6.17) and (6.18) for m_i of 1.0 where the $V_{base} = V_{dc}$.

$$\check{V}_{pos.pu} = -0.00275\theta_{v3} + 0.1327 \quad (6.32)$$

$$\hat{V}_{pos.pu} = 0.00275\theta_{v3} + 1.867 \quad (6.33)$$

The linear relationship between θ_{v3} and the peak synthesised voltages in (6.34) and (6.35) can be extended to consider the variations of m_i expressed by (6.36) and (6.37)

respectively as shown by Figure 6.13.a and Figure 6.13.b. The black dotted points are the data of the peak synthesised voltage obtained from Figure 6.11 using the same idea demonstrated by Figure 6.10 for different values of m_i .

Equations (6.36) and (6.37) are obtained using the MATLAB Curve Fitting tool which expresses the peak synthesised voltage in relationship with θ_{v3} and m_i . In order to determine the limit of the 3rd harmonic phase-shift for any given modulation index with regards to the previous solutions in (6.34) and (6.35), both equations (6.36) and (6.37) can be examined which give the same result. The given limit is expressed as a function of the variation of modulation index and phase-shift of the 3rd harmonic component based on the previous solution.

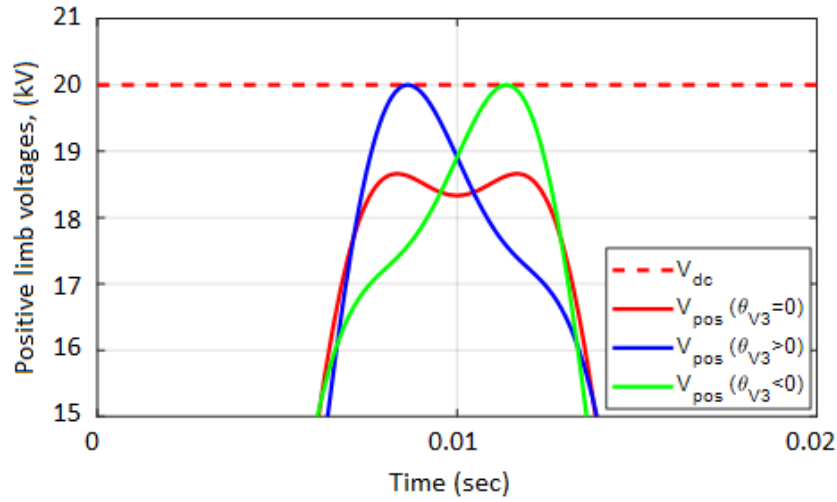


Figure 6.12: The impact of positive and negative 3rd harmonic phase shift on the synthesised voltage.

$$\check{V}_{pos.pu} = -0.00275\theta_{v3} + 0.1327 \quad (6.34)$$

$$\hat{V}_{pos.pu} = 0.00275\theta_{v3} + 1.867 \quad (6.35)$$

The linear relationship between θ_{v3} and the peak synthesised voltages in (6.34) and (6.35) can be extended to consider the variations of m_i expressed by (6.36) and (6.37) respectively as shown by Figure 6.13.a and Figure 6.13.b. The black dotted points are the data of the peak synthesised voltage obtained from Figure 6.11 using the same idea demonstrated by Figure 6.10 for different values of m_i .

Equations (6.36) and (6.37) are obtained using the MATLAB Curve Fitting tool which expresses the peak synthesised voltage in relationship with θ_{v3} and m_i . In order to determine the limit of the 3rd harmonic phase-shift for any given modulation index with

regards to the previous solutions in (6.34) and (6.35), both equations (6.36) and (6.37) can be examined which give the same result. The given limit is expressed as a function of the variation of modulation index and phase-shift of the 3rd harmonic component based on the previous solution.

$$\check{V}_{pos} = 1.06 - 0.9293m_i - 0.002657\theta_{v3} \quad (6.36)$$

$$\hat{V}_{pos} = 0.9402 + 0.9293m_i + 0.002657\theta_{v3} \quad (6.37)$$

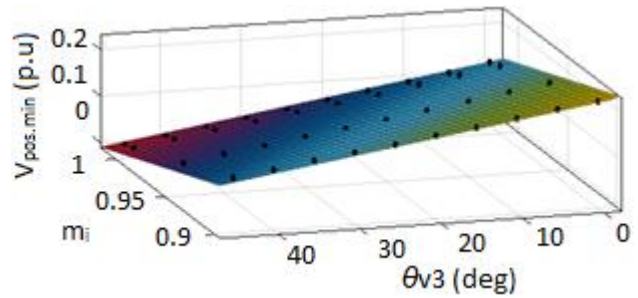
Considering the inequality in (6.31), $\check{V}_{pos}$ in equation (6.36) must be zero in order to determine the maximum 3rd harmonic as a function of the modulation index. Thus:

$$0 = 1.06 - 0.9293m_i - 0.002657\theta_{v3} \quad (6.38)$$

$$|\theta_{v3.max}| < \frac{1.06 - 0.9293m_i}{0.002657} \quad (6.39)$$

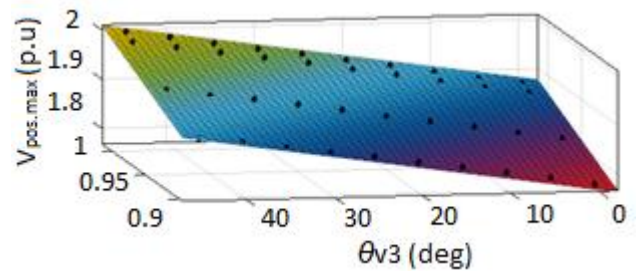
Equation (6.39) expresses a linear relationship between the maximum θ_{v3} of the 3rd harmonic component and modulation index. For any given modulation index by the controller, the maximum θ_{v3} is limited by (6.39) where the equation, (6.39), considers both positive and negative phase angle of the 3rd harmonic voltage shown in Figure 6.12.

Linear model Poly11:
 $f(x,y) = p00 + p10*x + p01*y$
 Coefficients (with 95% confidence bounds):
 p00 = 1.06 (1.046, 1.074)
 p10 = -0.9293 (-0.944, -0.9146)
 p01 = -0.002657 (-0.002702, -0.002613)



(a) minimum synthesised voltage, $\check{V}_{pos}$, in relationship to m_i and θ_{v3} .

Linear model Poly11:
 $f(x,y) = p00 + p10*x + p01*y$
 Coefficients (with 95% confidence bounds):
 p00 = 0.9402 (0.926, 0.9544)
 p10 = 0.9293 (0.9146, 0.944)
 p01 = 0.002657 (0.002613, 0.002702)



(b) peak, maximum, synthesised voltage, $\hat{V}_{pos}$, in relationship to m_i and θ_{v3} .

Figure 6.13: Peak synthesised voltage in relation to m_i and 3rd harmonic phase-shift using Curve fitting MATLAB.

6.4.2 Non-ideal MMC

The previous analysis presents a particularly important limit based on the ideal MMC. The lower limit of the inequality given by (6.30) was determined while the upper limit can be given by the overmodulation limit. However, the phase-shift limit presented can be applied even for the non-ideal MMC depending on the operating point according to the P and Q capability curve. To extend the analysis considering the upper limit for the non-ideal MMC where the ripple of the available voltage is considered, the available voltage given by (3.40) must be considered. However, for more convenience and after manipulation, it is rewritten here by (6.40).

$$\Sigma v_{C_{SM}} = \sqrt{\frac{2 N_{SM}}{C_{SM}} e_{pos}(t) + V_{dc}^2} \quad (6.40)$$

To investigate the maximum modulation index allowed due to the reduced available voltage, the m_i is varied for any given P and Q, whilst considering the rated $|S|$. Unfortunately, solving the inequality problem in (6.30) becomes more difficult even using a numerical solution. This is because the solution must consider three variables for different operating points where each particular operating point of PQ as given by Figure 6.14 should be examined with the variation of m_i and θ_{v3} . For the given P and Q envelope in Figure 6.14, the dotted red circle represents the PQ envelope for an ideal MMC whilst the blue curve represents a practical PQ envelope.

The analysis is carried out for the rated $|S|$ with the variation of P and Q where m_i is varied to examine the 3rd harmonic voltage phase shift starting from point A and moving back to A via B as shown by Figure 6.15.a. During this variation, the optimal phase shift of 3rd harmonic is selected at each operating point. This will create a lookup table that can be used to determine the required 3rd harmonic phase shift for any operating point at rated $|S|$.

Considering the operating point A in Figure 6.15.a, with the increase of the modulation index the AC voltage at the MMC terminals will increase while the AC current will reduce for a given rated $|S|$. To keep the AC voltage of the grid fixed, the transformer turns ratio is adapted accordingly. The new AC voltage at the MMC terminals and therefore the transformer turns ratio will be based on the maximum modulation index of this lookup table considering the worst case of the MMC operation with respect to its

synthesised voltage lower limit. Considering the synthesised voltage lower limit, it is noted that the worst case occurs when the MMC acts like a capacitor injecting reactive power $Q > 0$ to the AC system as demonstrated by point D in Figure 6.15.b. Of course the selected modulation index for the new design should also allow correct operation at the worst case of the available voltage at the opposite mode of operation during inductive mode $Q < 0$ presented by point A as shown by Figure 6.15.b. Therefore, the new design of the transformer turns ratio will be based on these new operation conditions so that the MMC can operate properly at rated operating points A and D.

Considering Figure 4.17, the dashed green line illustrates the operating with $+P$ or $-P$, i.e. inverter and rectifier, and zero reactive power. The operating in this point can be considered as an overlap point where the MMC moves from the operation from capacitive to inductive or vice versa. In fact there are four overlap points ($\phi = 0^\circ$, $\phi = +90^\circ$, $\phi = +180^\circ$ and $\phi = -90^\circ$) as shows in Figure 6.14. These four points can lead to interfering when solving the optimum 3rd harmonic phase shift. Therefore, these overlap points between the quadrants which has different characteristics of the operation, need to considered carefully in order to obtain the optimal solution. For the simplicity, four lookup tables are applied for each quadrant. The interfering due to the overlap points are considered so that the phase shift is chosen to meet the requirement of the worst case scenario. This provides a general result from which the practical case of an MMC grid connected through a line reactor can be extracted as shown by Appendix D.

The idea of the phase-shift is that it will change the shape of the peak of the synthesised voltage by increasing it in some point of the waveform and reducing it in another to insure a correct operation while not affecting the available voltage as discussed in the previous Chapters. Therefore, some of the issues related to these overlap points and other considerations should be taken into account when solving the problem of the required phase shift including:

- The interfering between the quadrants Q_1 , Q_2 , Q_3 and Q_4 .
- The chosen phase shift should provide a correct operation during any transient response such as step change of PQ .

- The phase shift has the ability to provide the correct operation to meet any demand of the apparent power ($S = \pm P \pm jQ$). Therefore, the rated power, S , is used to calculate the required phase shift.
- The high value of the phase shift which provides a correct operation for Q_2 and Q_3 is considered to be the optimum solution of θ_{v3} .

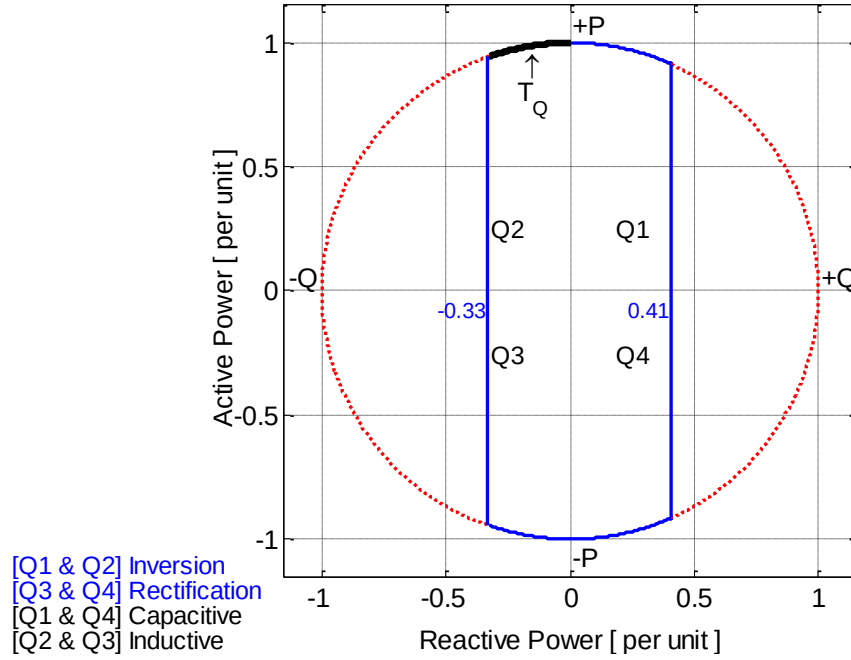
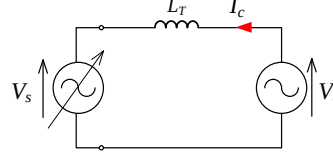


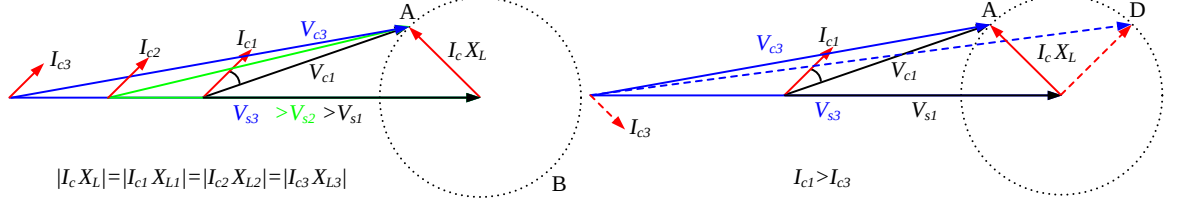
Figure 6.14: PQ curve for ideal and practical MMC.

Applying different values of the phase shift at the overlap points leads to duplicate the phase shift for one operating point especially when the power exchange varies from one quadrant to another, for example, the variation between Q_1 and Q_2 . Therefore, second option is applicable where the same value of the θ_{v3} at the overlap points is applied. The high value of the phase shift which provides a correct operation for Q_2 and Q_3 is considered to be the optimum solution of θ_{v3} at these operating points, ($\phi = 0^\circ$). Similarly, this is applied to the other overlap points.

Generally, an overmodulation can be occurred especially when operating with high modulation index. Therefore it is cortical to apply the phase shift, θ_{v3} , to avoid this overmodulation. It is noted that θ_{v3} should be positive, $+\theta_{v3}$, for (Q_1 and Q_2) and negative, $-\theta_{v3}$, for (Q_3 and Q_4) as shown by Figure 4.17.a and Figure 4.17.b respectively.

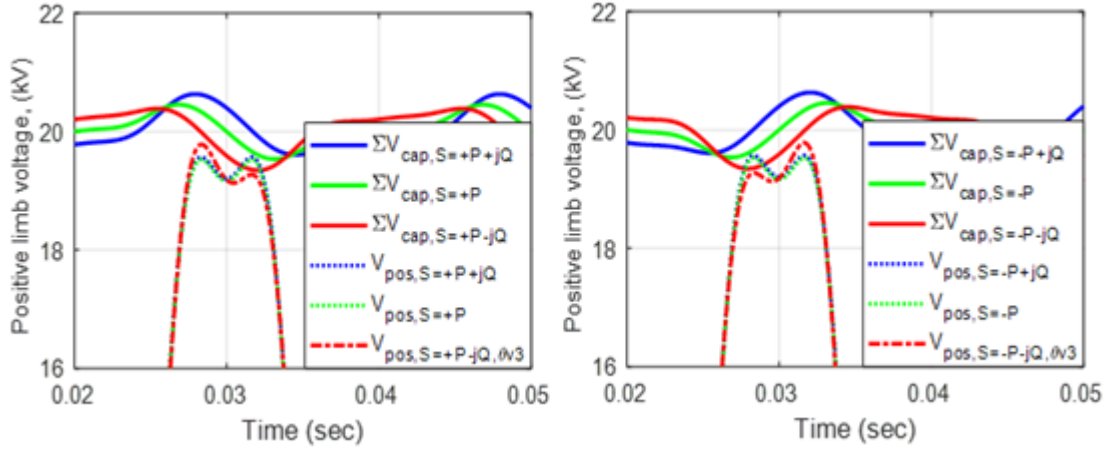


(a) MMC equivalent circuit with variable transformer turns ratio.


 (b) the increase of m_i to select optimal θ_{v3} .

 (c) the increased m_i applied for A and D using optimal θ_{v3}

Figure 6.15: Phasor diagrams of the MMC terminal voltage during increase of the modulating index.


 (a) operating in quadrants Q₁ and Q₂

 (b) operating in quadrants Q₃ and Q₄

 Figure 6.16: The required phase shift angle for different operating points; (a) $S = +P + jQ$, $S = +P + j0$ and $S = +P - jQ$, (b) $S = -P - jQ$, $S = -P - j0$ and $S = -P + jQ$.

Using the same approach presented in Section 6.4.1 and as shown in Figure 6.17, considering the lookup table in Appendix D, the relationship between m_i , θ_{v3} and the PQ operating points at rated S represented by $(\emptyset)$ could be obtained using the MATLAB Curve Fitting tool. It is noted that the dotted points represent the data obtained from the simulation results with the MMC operating properly. The non-dotted surface represents (red, blue and yellow) either limitation due to the overmodulation limit or by the lower limit as discussed in the previous section in (6.39). The red, blue and yellow areas represent operation with $\theta_{v3} < 0$, $\theta_{v3} = 0$ and $\theta_{v3} > 0$ respectively.

If the analysis considers only one particular m_i , then the investigation will be much easier. Therefore, let us consider the segment T_Q as shown in Figure 6.14. In this case, the MMC is operating at rated power $|S|$ and therefore maximum modulation index for each operating point of P and Q which presented in the lookup table in Figure 6.17.a. As an example, Figure 6.18 shows the fitting that can be obtained for the optimal phase shift of the 3rd harmonic using the MATLAB Curve fitting tool for a modulation index of 1.07 as shown by the segment T_Q in Figure 6.17.

Equation (6.41) is obtained using MATLAB Curve Fitting with the segment T_Q in Figure 6.18. It is noted that both angles ϕ and θ_{v3} , are measured in degrees.

$$\theta_{v3} = -5.255e^{0.352\phi} + 5.293e^{-0.139\phi} \quad (6.41)$$

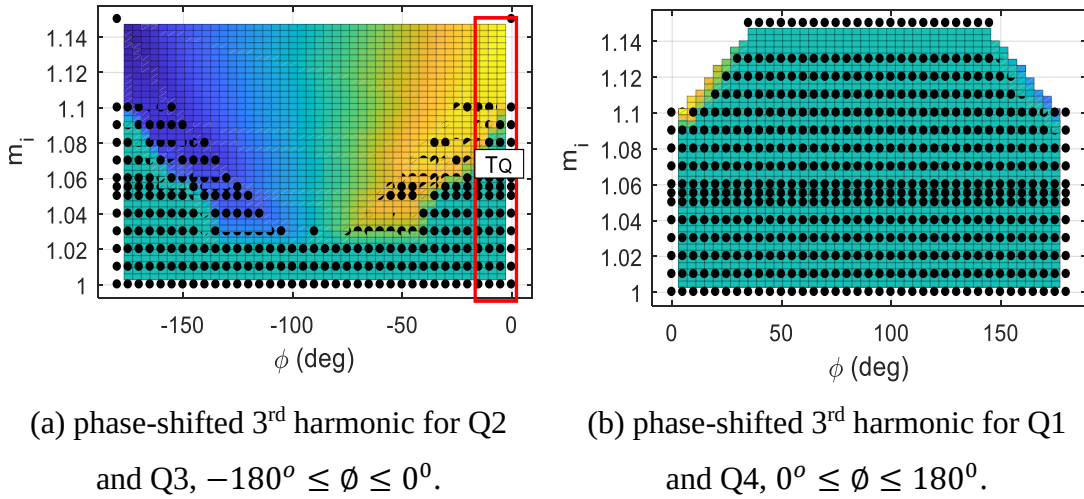


Figure 6.17: The required phase angle of the 3rd harmonic with variation of m_i and variation of ϕ at rated $|S|$ of ideal MMC.

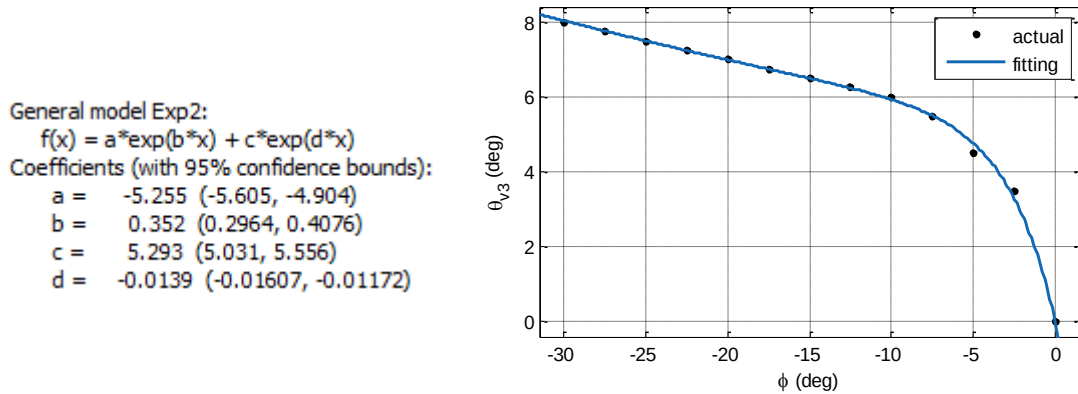


Figure 6.18: The relationship between θ_{v3} and rated $|S|$ and variation of P and Q ($-30^\circ \leq \phi \leq 0^\circ$) and $m_i=1.07$.

The solution given by (6.41) can be extended where the modulation index is considered as a variable. Based on the PQ envelope given by Figure 6.14, Figure 6.19 shows the relationship between m_i , ϕ and the 3rd harmonic phase-shift angle, where ϕ and θ_{v3} are measured in degrees.

Figure 6.17.a can be seen from a different perspective in Figure 6.19 where the lookup table is given in Appendix D. From both figures it can be seen that correct fitting cannot be obtained using the MATLAB Curve Fitting tool, especially for certain regions of ϕ where the operation of the MMC enters into overmodulation or is limited by the lower limit according to (6.39). The operation in these regions should be prohibited. Moreover, the MATLAB Curve Fitting tool cannot be used in these regions. This is because the MMC does not operate in this region illustrated by the solid surface (red, blue and yellow without dotted black), due to the attempted increased of the modulation index and therefore this region should be prohibited. Furthermore, non-linear behaviour occurs as $|\phi|$ increases from $|0^\circ|$ to $|180^\circ|$ where the θ_{v3} needs to be increased and changed from positive to negative even when considering the region where the operation can be allowed as shown by the red cuboid in Figure 6.19.e. The dotted points on the surface illustrate steady state operation for positive and negative phase shift 3rd harmonic voltage injection (yellow and red surfaces, respectively). The blue surface illustrates the operation when θ_{v3} is zero. Therefore, four lookup tables are considered for each quadrant of operation. For instance, the red rectangular line and cuboid in Figure 6.17.a and Figure 6.19 illustrate the second quadrant (Q2) on PQ plane.

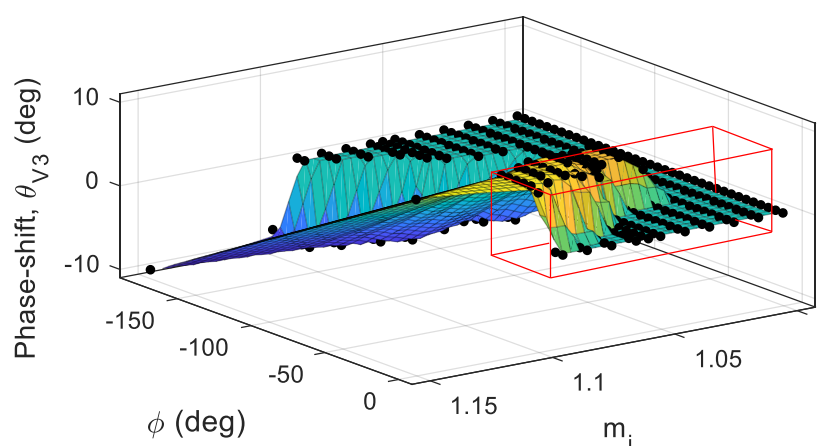


Figure 6.19: The required phase angle of the 3rd harmonic with variation of m_i and $-180^\circ \leq \phi \leq 0^\circ$ at rated $|S|$.

Therefore, a similar solution to that given by (6.41) cannot be achieved using the same MATLAB Curve Fitting approach. The difficulty is to develop a mathematical solution expressing the nonlinear relationship as shown in Figure 6.19. Considering quadrant Q_2 in Figure 6.14 and Figure 6.19, the selected region illustrated by the red cuboid in Figure 6.16 ensures operation and also meets the limit requirements and the operating conditions. The relationship between m_i , ϕ and θ_{v3} can be re-plotted and represented by Figure 6.20.

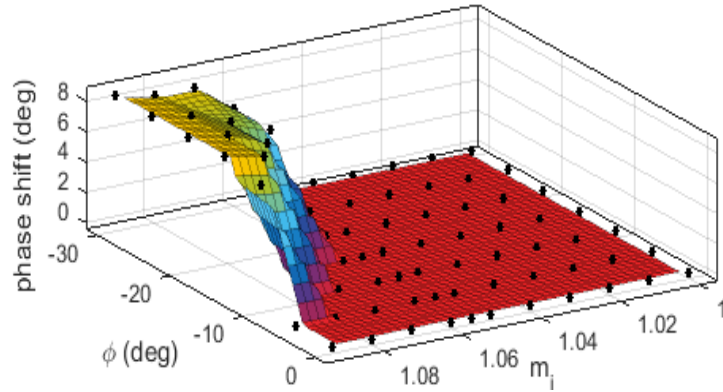


Figure 6.20: The required phase of the 3rd harmonic with variation of m_i and $-30^\circ \leq \phi \leq 0^\circ$ at rated $|S|$.

Considering Figure 6.20, it is noted that because of the nonlinearity, the MATLAB Curve Fitting tools cannot develop a satisfactory mathematical solution expressing the relationship illustrated in Figure 6.20. Therefore, a more convenient method that can be implemented is to use the latter as a lookup table. In fact, four lookup tables are required for each quadrant of operation as illustrated by Figure 6.17 and Figure 6.19 where leading and lagging 3rd harmonic phase-shift are considered for the four quadrants of operation as shown by Figure 6.14. The input to the look up table is the modulation index required by the controller. Figure 6.21 illustrates the block diagram of the look up table with a control signal and the measured power factor expressed by P and Q . The system is also provided with a quadrant selector. The quadrant selection unit determines the quadrant of the operating point of the MMC while the lookup table is used to determine the required phase angle of the 3rd harmonic component. It is also noted that lookup tables are obtained for the variation of the modulation index and are based on the rated power $|S|$. It is noted that when the MMC operates at less than the rated $|S|$ not only is the AC current reduced, but also the modulation index for the same ϕ .

Therefore, when MMC operates with power less than the rated value, the modulation index will be less than that for operating at rated $|S|$ and is covered by the lookup table as illustrated by Figure 6.17 for small values of modulation index. It can be seen that when the modulation index reduces to less than 1.0, the phase-shift of the 3rd harmonic injection will not applied for any value of ϕ especially when operating at only $Q=-0.33$. In this case the generated modulation index to meet this operating point is the smallest while the available voltage is also small due to the operation with 33% of the rated power $|S|$ so that the energy variation is reduced compared to the operation with the rated $|S|$.

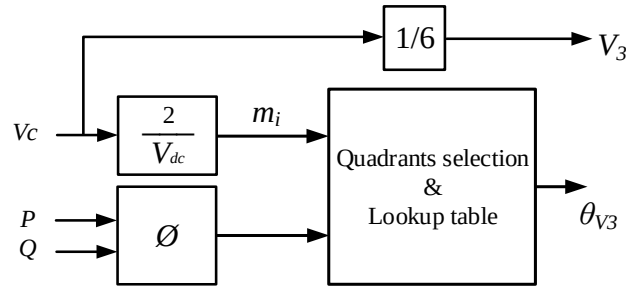


Figure 6.21: Phase-shift look up table and quadrants selector.

6.5 Benefits of Using Third Harmonic Injection

From the previous analysis and discussion it can be seen that 3rd harmonic voltage injection can reduce the peak synthesised voltage for the same generated fundamental component where the optimal reduction occurs when 16.6% of 3rd harmonic is used. The reduced peak synthesised voltage can be then used to increase the fundamental component of the generated voltage using zero phase shifted 3rd harmonic injection. This can be further improved by using phase-shifted 3rd harmonic injection. As an alternative, the advantage of the reduced peak synthesised voltage can be also used to resize the SM capacitor size by allowing larger ripple for the same generated fundamental voltage. Therefore, the two proposed methods to improve the operation of the MMC are discussed in this section.

6.5.1 Benefits on Increasing the MMC Terminal Voltage

From the previous analysis it can be seen that using optimal 3rd harmonic injection of 1/6 of the fundamental can lead to a reduction of the peak synthesised voltage by 0.134 as given in (6.20). The advantage of this reduction is that it can be used to improve the performance of the MMC and improve its design. First, reducing the peak synthesised voltage increases the available voltage margin. For a given AC grid voltage and a given

transformer turns ratio, this can allow operating the MMC with large Q as shown by Figure 6.17 where the MMC operation can be maintained with a greater modulation index than the case without 3rd harmonic injection.

However, if the operation of the MMC needs to be restricted to a specific PQ region, then the use of optimal 3rd harmonic injection using the phase-shift concept can allow operation with a higher modulation index as illustrated by the lookup table in Figure 6.20. For an MMC designed with big SM capacitor sizes or operating with low rated power, this increase can reach 1.155 as given by (6.23). Operating with high modulation index leads to an increase in the output voltage at the MMC terminals. This means that for a given AC system, the transformer turns ratio can be reduced leading to lower MMC current as shown by Figure 6.22 by increasing the converter terminal voltage. However, the operation with high modulation index in some operating points is limited so that the 1.155 figure is not achievable even with the use of the phase-shifted 3rd harmonic as discussed earlier and expressed by Figure 6.17. The reason for this limitation is related to the fact that any further increase of the peak limb synthesised voltage is limited either by the lower limit (6.39) or by overmodulation. In fact the line impedance is a key factor as the generated modulation index and also the limb available voltage are imposed by P and Q considering the line impedance. This prevents further increase in the fundamental component of the MMC output voltage at the PCC point where the latter is given by (6.42).

$$V_s = V_c - jX_L I_c \quad (6.42)$$

where V_s and I_c are the measured voltage and current at the PCC point of the three phases, A, B and C, V_c is the output voltage at the MMC terminals which are connected to the AC system via the converter transformer with an equivalent reactance of X_L . In case of the positive sequence, the voltage of the three phases, V_{cab} , are equal with 120° displacement so that $|V_{ca}| = |V_{cb}| = |V_{cc}| = |V_c|$ which can be applied to the grid voltage and the line current. The single line diagram of an HVDC transmission system and its equivalent circuit are illustrated by Figure 6.23 and Figure 6.24 respectively.

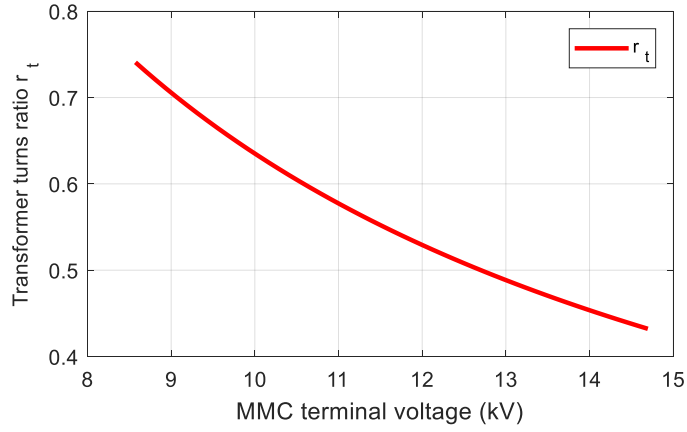


Figure 6.22: The MMC converter transformer ratio and modulation index.

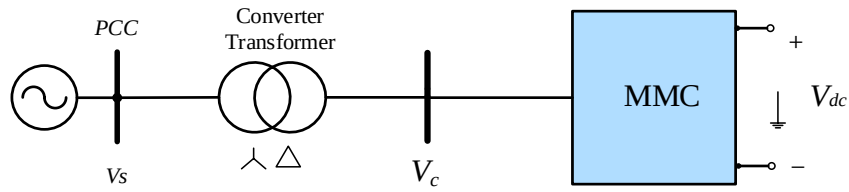


Figure 6.23: Single line diagram of HVDC transmission system based on MMC.

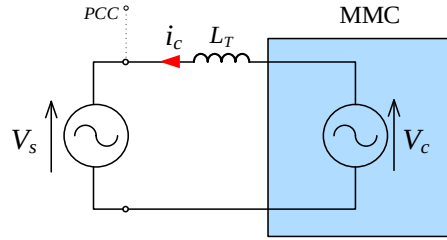


Figure 6.24: The equivalent circuit of HVDC transmission system based on MMC.

Considering Figure 6.23, the AC and DC systems in the single line diagram are connected using an MMC via a YD transformer configuration as discussed in Chapter 4. Using the simplified single line diagram, the transformer ratio can be given in (6.43).

$$r_t = \frac{n_1}{n_2} = \frac{V_{ph.pri}}{V_{ph.sec}} \quad (6.43)$$

where n_1 , n_2 , $V_{ph.pri}$ and $V_{ph.sec}$ are numbers of turns and phase voltages of primary and secondary windings, respectively. From the single line diagram, it can be seen that the transformer configuration is Y/Delta and therefore, the transformer ratio can be given in (6.44). Moreover, the primary side is connected to the point of common

coupling, PCC, with its voltage of V_s while the secondary side of the transformer is connected to MMC output terminals V_c .

$$r_t = \frac{V_{pri} (L)}{\sqrt{3}V_{sec} (ph)} \quad (6.44)$$

Using line-to-line voltages and the RMS value of the MMC output terminal voltage given in (6.45), the approximation of r_t can be expressed by (6.46) given by the AC system voltage at the PCC, V_s and the DC link voltage V_{dc} .

$$V_c = \sqrt{3}m_i \frac{V_{dc}}{2\sqrt{2}} \quad (6.45)$$

$$r_t = \frac{2\sqrt{2}V_s}{3m_i V_{dc}} \quad (6.46)$$

where m_i is the rated modulation index of the MMC terminal AC voltage.

According to Figure 6.20 and Figure 6.21, using the advantages of phase-shifted 3rd harmonic voltage injection can allow operation with higher modulation index. It can be seen that as the modulation index increases, the transformer ratio decreases. Moreover, as the output of the fundamental component increases, the line currents in the secondary side are reduced for the same rated power. This can lead to reduce the power loss. Simulation results will be presented in the next chapter examining this.

6.5.2 Effect of Optimal Third Harmonic Injection on SM

Capacitor Size

The impact of 3rd harmonic voltage injection on energy deviation and therefore capacitor size has been presented in Chapter 4 and 5 and given by Figure 3.38. Based on the energy deviation, it has been shown that for a desired voltage ripple in the SM capacitors the advantages of using the 3rd harmonic voltage injection in the capacitance reduction are negligible.

However, another option to reduce SM capacitors is to take advantage of the large voltage margin provided by the 3rd harmonic voltage injection to allow larger voltage ripple in the SM capacitors.

6.5.2.1 SM Capacitor Size Analysis

The size of capacitor is given as a function of the numbers of SM, maximum energy deviation, capacitor voltage ripple and the DC link voltage. Considering the MMC specifications in Table 3.2 and assuming that the number of SM has already been chosen, then it can be said that the capacitor size is directly proportional to the maximum energy deviation. It is assumed that a given MMC operates with rated power where the capacitor size is designed based on the worst case at the rated operating points of m_i and P and Q. The energy deviation in Equation (5.12) can be considered to be constant as the operating conditions are the same even if the SM capacitor is replaced with a bigger or smaller size. It can be seen that the size of capacitor is inversely proportional to the desired capacitor voltage ripple, δ , as expressed by (6.47) where $k = \frac{N_{SM}\Delta E_{max}}{V_{dc}^2}$ is the constant of the proportionality.

$$C_{SM.min} \propto \frac{k}{\delta} \quad (6.47)$$

The capacitor voltage ripple is given as a percentage of the SM capacitor voltage, as discussed in Chapter 5, and given by Equation (5.5). The available voltage is given by (3.42) as the sum of the SMs capacitors voltages. Therefore, the ripple can be re-expressed by (6.48).

$$\delta = \frac{\Delta \Sigma V_{C_{SM}}}{\Sigma V_{C_{SM}}} \quad (6.48)$$

where $\Sigma V_{C_{SM}}$ and $\Delta \Sigma V_{C_{SM}}$ are the average and ripple of the positive or negative limb available voltage respectively ($\Sigma V_{C_{SM}} = V_{dc}$). As discussed in Section 6.3.1, using 3rd harmonic voltage injection can reduce the peak synthesised voltage as expressed in (6.20). The reduced peak synthesised voltage can allow operation with a larger voltage ripple ($\Delta \Sigma V_{C_{SM}}$). This can be achieved using a smaller capacitor size compared with the nominal design.

The reduced synthesised voltage can be used to increase the available voltage ripple while its average remains constant. A suitable margin between the synthesised and available voltages can still be provided to maintain the converter operation. Therefore,

the ripple in (6.48) can be increased. Using the rated stored energy in one limb given by (5.2), the size of capacitor in relation to voltage ripple can be re-expressed by the total rated stored energy capability, W_{rated} , in (6.49) given in $(kJ/MW = ms)$ where the latter is 6 times the rated energy of one limb divided by the rated power S [16], thus,

$$W_{rated} = 3C_{SM.min}N_{SM}\frac{V_{C_{SM}}^2}{|S|} \quad (6.49)$$

where $V_{C_{SM}}$ is the nominal SM capacitor voltage. $C_{SM.min}$ is given by (6.47) while $|S|$ is the rated apparent power thus:

$$W_{rated} \propto \frac{1}{\delta} \quad (6.50)$$

From the previous discussion, it can be seen that the size of capacitor can be reduced due to the reduced peak synthesised voltage. Therefore, the focus in this section is on resizing the SM capacitor based on the optimal classical and phase-shifted 3rd harmonic injection.

The nominal (full) reduction of the limb synthesised voltage is given by Equation (6.20). Therefore, the same amount of reduction can be used to increase the available voltage. To maintain the margin between the synthesised voltage and the available voltage, the increase of the available voltage should be the same amount as the reduction in the synthesised voltage or less so that the conditions in Equation (3.6) always apply. Using the entire total reduction, the new ripple can be given by (6.55) where the amount of reduction is added to both the maximum and minimum peak values of the available voltage so that the new maximum and minimum available voltage of (3.40) now is given by (6.51) and (6.52) respectively.

$$\hat{\Sigma}V_{C_{SM}.new}(t) = \hat{\Sigma}V_{C_{SM}.old}(t) + V_{red} \quad (6.51)$$

$$\check{\Sigma}V_{C_{SM}.new}(t) = \check{\Sigma}V_{C_{SM}.old}(t) - V_{red} \quad (6.52)$$

The implemented method is demonstrated by an example as shown in Figure 6.25. As can be seen, the voltage ripple is not symmetrical in this operating point, therefore when

considering the new size of capacitor the MMC can enter overmodulation due to this assumption. Therefore, phase-shifted 3rd harmonic injection can be used to maintain normal operation.

Therefore, the ripple of the new available voltage can be obtained by subtracting (6.51) and (6.52). This yields;

$$\Delta\Sigma V_{C_{SM}.new.V_{red}} = \Delta\Sigma V_{C_{SM}.old} + 2V_{red} \quad (6.53)$$

where $\Delta\Sigma V_{C_{SM}.new.V_{red}}$ and $\Delta\Sigma V_{C_{SM}.old}$ are the new and the previous available voltage ripple, respectively. However, a smaller value of the total reduction is used in order to increase the safety margin between the new available voltage and the limb synthesised voltage.

To consider the asymmetry of the voltage ripple with respect to the average as shown in Figure 6.25 and Appendix C and also to include a safety margin, only half the reduction in (6.54) can be used with a zero phase-shifted 3rd harmonic voltage injection. In the following analysis it will be shown that the phase-shifted 3rd harmonic can solve the problem allowing full reduction. The ripple of the new available voltage due to using half the reduction can be obtained by subtracting (6.51) and (6.52) where half of the reduced voltage is added to the maximum and subtracted from the minimum available voltage. This yields;

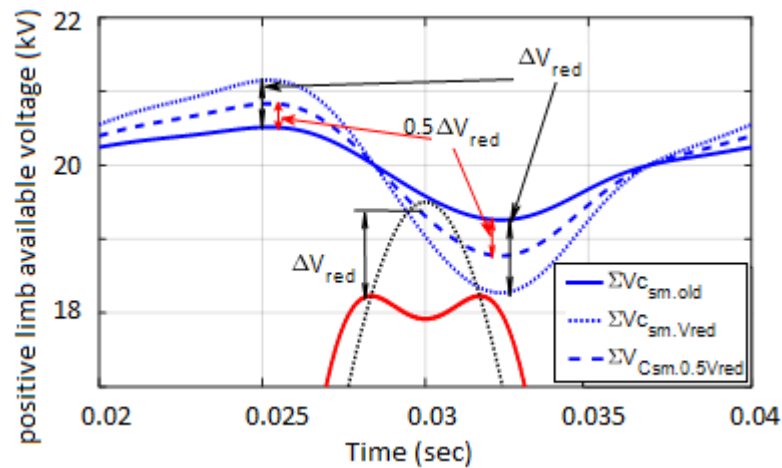


Figure 6.25: Increasing the available voltage by full and half amount of the reduced synthesised voltage.

$$\Delta\Sigma V_{C_{SM}.new.0.5V_{red}} = \Delta\Sigma V_{C_{SM}.old} + V_{red} \quad (6.54)$$

Using (6.53) and normalising by V_{dc} , the new ripple for the full and half voltage reduction can be expressed by (6.55) and (6.56), respectively.

$$\delta_{new.V_{red}} = \delta_{old} + 2V_{red,p.u} \quad (6.55)$$

$$\delta_{new.0.5V_{red}} = \delta_{old} + V_{red,p.u} \quad (6.56)$$

where $V_{red,p.u}$ is the nominal reduced synthesised voltage (in per-unit) based in (6.20) where the base is V_{dc} which is expressed in (6.57). $\delta_{new.V_{red}}$ and $\delta_{new.0.5V_{red}}$ are the new available voltage ripple levels based on using the full nominal and half amounts of the reduced synthesised voltage respectively.

$$V_{red,p.u} = \frac{1}{2} \left(1 - \frac{\sqrt{3}}{2} \right) m_i \quad (6.57)$$

In order to evaluate the benefits of phase shifted 3rd harmonic in comparison with zero phase 3rd harmonic voltage injection, in the following analysis the design of the C_{SM} will be presented for both cases. Substituting (6.57) in (6.55) and (6.56), the new available voltage ripple based on full and half the reduced synthesised voltage are expressed below in (6.58) and (6.59). m_i is the nominal modulation index of the fundamental.

$$\delta_{new.V_{red}} = \delta_{old} + 0.134m_i \quad (6.58)$$

$$\delta_{new.0.5V_{red}} = \delta_{old} + 0.067m_i \quad (6.59)$$

Using (6.47), for the old and also for the new capacitor voltage ripple, δ_{old} and δ_{new} (in case of nominal, V_{red} , and $0.5V_{red}$ reduction respectively, the new SM capacitor size is given by (6.60).

$$C_{new} = \frac{\delta_{old}}{\delta_{new}} C_{old} \quad (6.60)$$

Substituting (6.58) and (6.59) into (6.60), the new size of capacitor considering nominal reduced synthesised voltage ($C_{new.V_{red}}$) is given by (6.61), whilst using half reduction ($C_{new.0.5V_{red}}$) is expressed by (6.62).

$$C_{new.V_{red}} = \frac{\delta_{old}}{\delta_{old} + 0.134m_i} C_{old} \quad (6.61)$$

$$C_{new.0.5V_{red}} = \frac{\delta_{old}}{\delta_{old} + 0.067m_i} C_{old} \quad (6.62)$$

Therefore, the new size of SM capacitor can be given by (6.63) and (6.64) considering the maximum energy deviation using Equation (5.12).

$$C_{new.V_{red}} = \frac{\Delta E_{max}}{(\delta_{old} + 0.134m_i)N_{SM}V_{CSM}^2} \quad (6.63)$$

Or by;

$$C_{new.0.5V_{red}} = \frac{\Delta E_{max}}{(\delta_{old} + 0.067m_i)N_{SM}V_{CSM}^2} \quad (6.64)$$

where ΔE_{max} is the maximum energy deviation and m_i is the modulation index of the nominal AC voltage at the MMC terminals.

The previous analysis and discussion considered the possibility of reducing the size of capacitor by increasing the available voltage ripple using full and half reduction of the synthesised voltage, (V_{red}) and ($0.5V_{red}$) respectively. The analysis can be further extended by investigating the benefit of greater voltage reduction. For instance, one and half times ($1.5V_{red}$) greater than the full voltage reduction (V_{red}) is investigated, For the MMC parameters given in Table 6.1, ($1.5V_{red}$) reduction is applied instead of the (V_{red}) reduction expressed by (6.53). Therefore, the ripple of the new available voltage due to the ($1.5V_{red}$) reduction is expressed by (6.65).

$$\Delta \Sigma V_{CSM.new.1.5V_{red}} = \Delta \Sigma V_{CSM.old} + 3V_{red} \quad (6.65)$$

Using (6.65), similarly, the new ripple and the SM capacitor size can be given by (6.66) and (6.68) respectively.

$$\delta_{new.1.5V_{red}} = \delta_{old} + 0.201m_i \quad (6.66)$$

$$C_{new.1.5V_{red}} = \frac{\delta_{old}}{\delta_{old} + 0.201m_i} C_{old} \quad (6.67)$$

$$C_{new.1.5V_{red}} = \frac{\Delta E_{max}}{(\delta_{old} + 0.201m_i)N_{SM}V_{C_{SM}}^2} \quad (6.68)$$

The previous analysis of the new design of the capacitor size can be given by the equations below in a general form based on the percentage of the reduced synthesised voltage applied.

$$\delta_{new} = \delta_{old} + 0.134k_{V_{red}}m_i \quad (6.69)$$

$$C_{new} = \frac{\delta_{old}}{\delta_{old} + 0.134k_{V_{red}}m_i} C_{old} \quad (6.70)$$

$$C_{new} = \frac{\Delta E_{max}}{(\delta_{old} + 0.134k_{V_{red}}m_i)N_{SM}V_{C_{SM}}^2} \quad (6.71)$$

where $k_{V_{red}}$ represents the reduced synthesised voltage of the nominal (V_{red}) reduction is used to increase the available voltage ripple in per-unit. $k_{V_{red}}$ equals 1.0, 0.5 and 1.5 for nominal (V_{red}), $0.5V_{red}$ and $1.5V_{red}$ reduction given by (6.63), (6.64) and (6.68) respectively. The last three equations are the general solution of the new design of the SM capacitor and describing the relationship between the reduced synthesised voltage the capacitor size.

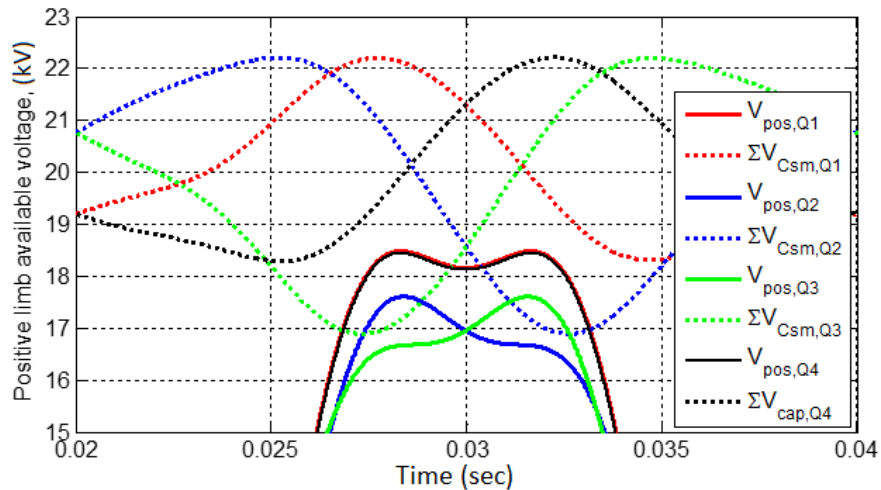


Figure 6.26: Positive and negative phase-shifted 3rd harmonic and capacitor resizing.

It is noted that the phase-shift of the 3rd harmonic must take into consideration P and Q operating point where positive phase shift is required for quadrants Q1 and Q2 shown in Figure 6.14 whilst negative phase shift is required for the other quadrants, Q3 and

Q4. This is further explained by the modulation indices for each operating point ($S_{Q1} = +0.94 + j0.41$), ($S_{Q2} = +0.94 - j0.33$), ($S_{Q3} = -0.94 - j0.33$) and ($S_{Q4} = -0.94 + j0.41$) which are $m_{i,Q1} = 0.974$, $m_{i,Q2} = 0.824$, $m_{i,Q3} = 0.824$ and $m_{i,Q4} = 0.974$ respectively based on Table 6.1.

For better performance, a lookup table is required for each operating point. For more convenience, the phase-shifted 3rd harmonic at the worst case operating points as shown in Figure 6.26 can be used which can always provide correct operation. However, this might not be the optimum value of the phase-shift.

6.5.2.2 Simulation Results

The results obtained below are based on the MMC and AC system parameters given in Table 6.1 where the SM capacitor size is selected based on the PQ envelope given by Figure 6.14. The two possible suggested solutions in the analysis above are considered including Equations (6.63) and (6.64).

The SM capacitor size given in the MMC parameters in Table 6.1 was designed based on capacitor voltage ripple of 0.1. Figure 6.27 illustrates a comparison between three operating cases where three SM capacitor sizes are redesigned based on an initial size as given in Table 6.1. The SM capacitor size reduction was considered using full, V_{red} , half, $0.5V_{red}$, and $1.5V_{red}$ reduction relationships expressed by (6.61), (6.62) and (6.67), respectively. The new ripple for each case can be calculated using (6.58), (6.59) and (6.66), respectively. It is noted that the initial SM capacitor size is designed based on modulation index of 0.824 and capacitor voltage ripple of 0.1.

It can be seen that using the total increase of the available voltage ripple of $0.5V_{red}$, V_{red} and $1.5V_{red}$ can lead to reduce the SM capacitor size by 35.56%, 52.4% and 62.3%, respectively as illustrated by the three points shown by Figure 6.27 and the actual capacitor sizes are given by Figure 6.30.a.

Using the new designs of the SM capacitor given by (6.63), (6.64) and (6.68), the waveforms of the available voltage are illustrated by Figure 6.28.a where the MMC is operating at rated $S = +0.94 - j0.33$ p.u and with m_i of 0.824. It should be noted that the 3rd harmonic phase shift must be applied as shown by solid red line to increase the safety margin in the case of using (6.64) where the selected SM capacitor size in this case is 4.785mF and $\theta_{v3} = 10^\circ$. It can be seen that in the case of using (6.68), the margin

to prevent an overmodulation when $\theta_{v3} = 10^\circ$ is small as shown by the dashed blue line in Figure 6.28.a with the legend of $\Sigma V_{C_{SM}, 1.5V_{red}}$ even though the initial selected SM capacitor size is large. Therefore, bigger θ_{v3} can solve this issue, $\theta_{v3} = 20^\circ$ as shown in Figure 6.28.b.

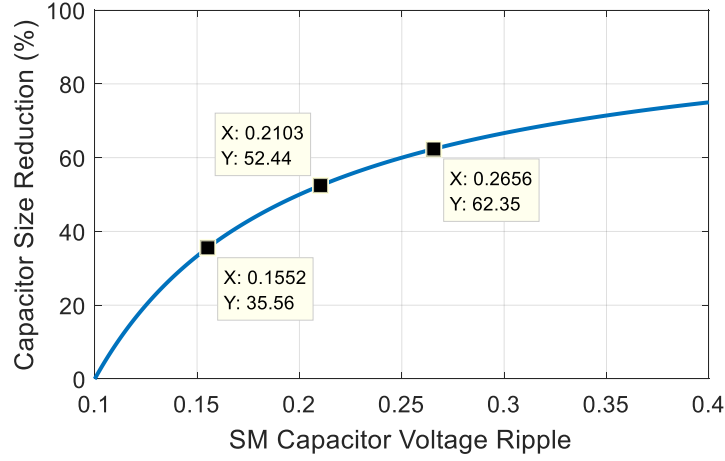
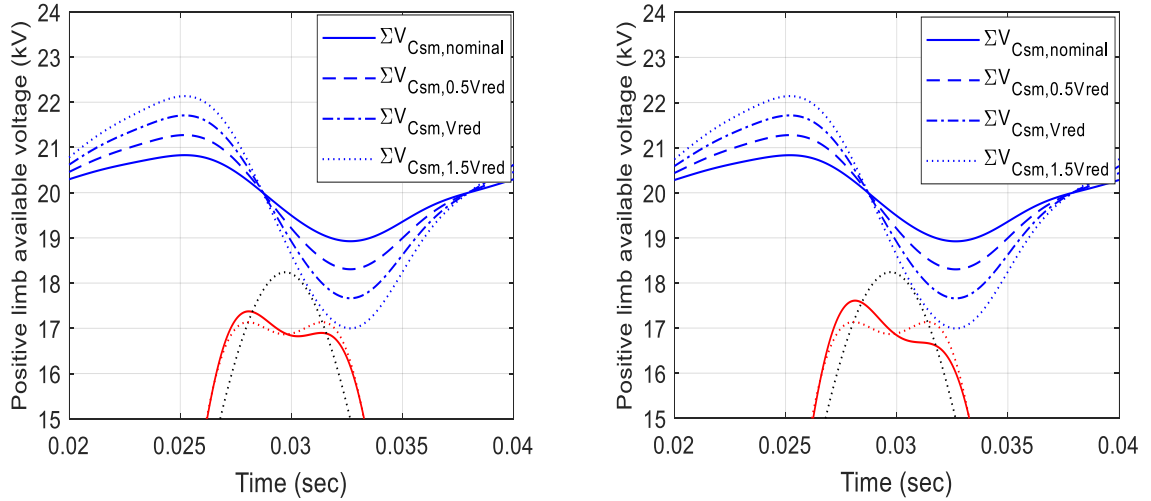


Figure 6.27: Capacitor size reduction due to advantage of peak synthesised voltage reduction, $m_i = 0.824$ and initial peak to peak ripple 0.1.

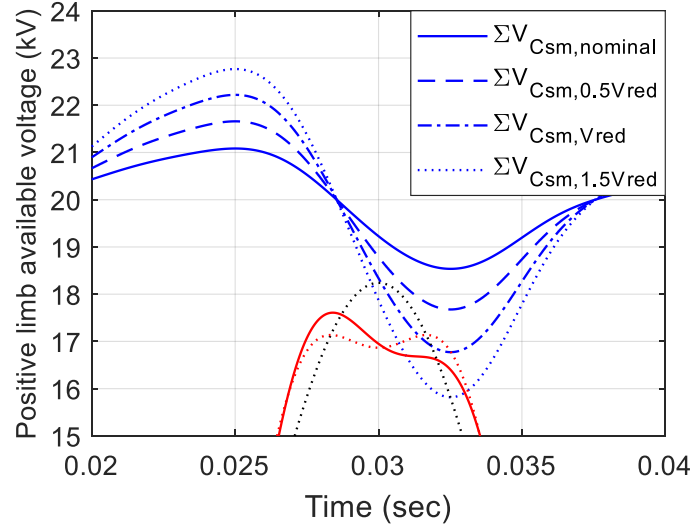


(a) initial $C_{SM.in} = 10.07\text{mF}$ and $\theta_{v3} = 10^\circ$. (b) initial $C_{SM} = C_{SM.in}$ and $\theta_{v3} = 20^\circ$.

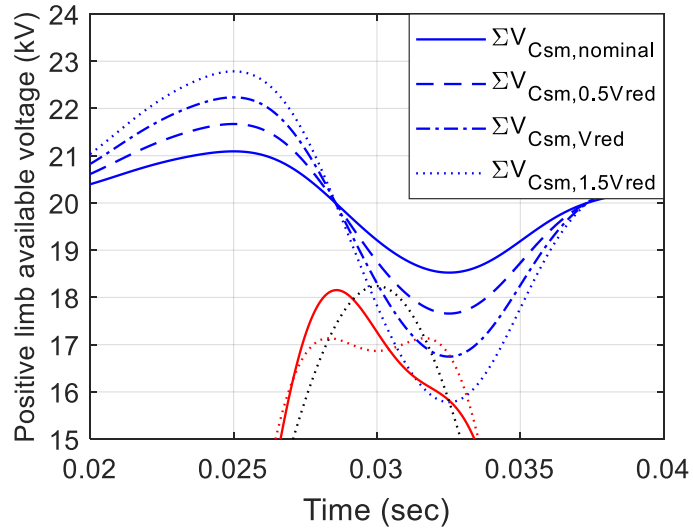
Figure 6.28: Capacitor size re-design for modulation index of 0.824, initial ripple of 0.1.

For further capacitor size reduction, phase shifted 3rd harmonic must be considered where positive and negative phase shift is applied as shown by Figure 6.26. For the MMC parameters given in Table 6.1 the SM capacitor can be further reduced from

$10.07mF$ to $3.791mF$. However, if the initial size of capacitor was smaller than $10.07mF$ by 20% ($C_{SM} = 0.8C_{SM.in} = 8.056mF$), for instance, then even a bigger value of phase shift of the 3rd harmonic cannot provide correct operation due to overmodulation as shown in Figure 6.29.b where in this case $\theta_{v3} = 45^\circ$ is applied compared to $\theta_{v3} = 20^\circ$ in case of Figure 6.29.a.



(a) initial $C_{SM.in}=10.07mF$, $\theta_{v3} = 20^\circ$.



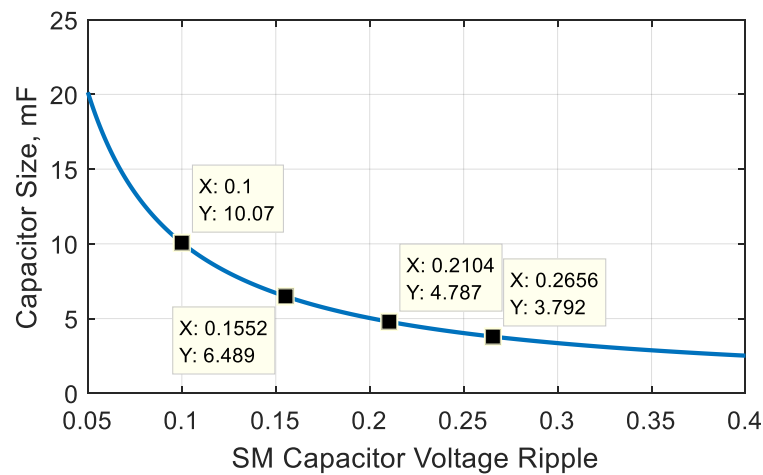
(b) initial $C_{SM}=0.8C_{SM.in}$, $\theta_{v3} = 45^\circ$

Figure 6.29: Capacitor size re-design for modulation index of 0.824, initial ripple of 0.1 two different initial size of capacitors $C_{SM.in}$ and $0.8C_{SM.in}$ and $\theta_{v3} = 20^\circ$ and 45° .

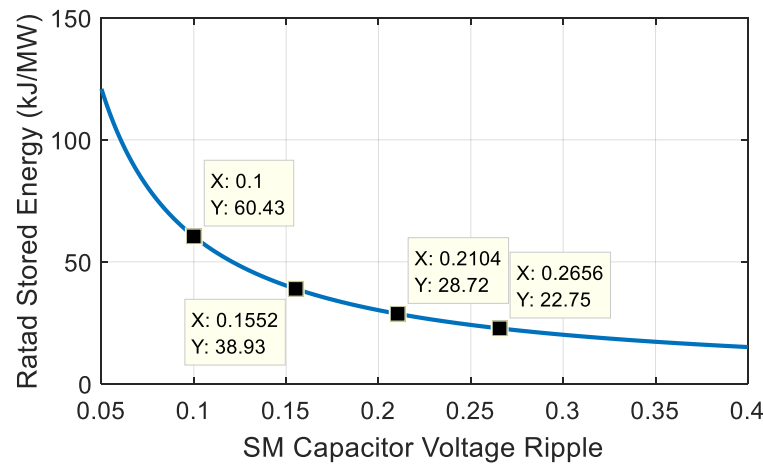
Applying phase shifted 3rd harmonic injection can lead to a reduction in the size of capacitor while using the rated modulation index and by considering the new capacitor voltage ripple. In general, the capacitor size can be given as a function of the capacitor voltage ripple expressed by (6.63) as shown by Figure 6.30.a. The total rated stored

energy of the three SM capacitor sizes is illustrated by Figure 6.30.b where the former can be reduced by increasing the capacitor size.

The rated energy storage of the three SM capacitor sizes compared to the initial SM capacitor size is illustrated by Figure 6.30. The initial SM capacitor size can be reduced from 10.07mF to 6.487mF, 4.485mF and 3.791mF [Y axis] as the SM voltage ripple increases [X axis]. Therefore, the rated stored energy can be reduced from 60.41 kJ/MW (ms) to 38.9 2ms, 28.71ms and 22.74ms, respectively.



(a) Capacitor size in *mF* and voltage ripple.



(b) Total rated stored energy and voltage ripple.

Figure 6.30: The total rated stored energy reduction using the advantage of peak synthesised voltage reduction.

6.6 Conclusion

The previous analysis and simulation results show that 3rd harmonic injection has the advantage of reducing the peak synthesised voltage. It was shown that this reduction

increases the safety margin between the synthesised voltage and the available voltage. Two different possible approaches presented by this thesis discussing the benefits from this reduction.

- Firstly, increase the output voltage at the MMC terminal by increasing the fundamental component of the synthesised voltage by the limbs. In this thesis, an approach was developed in order to increase the operation of the MMC by applying, and for the first time, phase shifted 3rd harmonic voltage injection. Limits of the phase shift was obtained then used to create lookup tables of the optimal phase shift. Other benefits have been discussed including the converter transformer and also the AC current.
- The second possible solution is that increasing the safety margin between the synthesised voltage and the available voltage, therefore, has the option to increase the ripple of the latter. Hence the SM capacitor size can be reduced. This reduction is further improved by applying phase shifted 3rd harmonic. A general solution was obtained describing the relationship between the synthesised voltage reduction and SM capacitor size.

The optimal 3rd harmonic voltage injection for the MMC has been presented considering both classical and phase-shifted 3rd harmonic injection. Considering the peak synthesised voltage limits, a look up table was obtained allowing operation of the MMC with a higher modulation index. The benefit of the optimal 3rd harmonic voltage injection on reducing the peak of the synthesised voltage has been discussed. The increase of the AC voltage at the MMC terminal due to the reduced synthesised voltage also has been discussed for zero and phase-shifted 3rd harmonic voltage injection.

In Chapter 7, the control schemes implemented are discussed and the controls are designed. The MMC modelling is presented including the switching model.

Chapter 7

7 Control Design and MMC Modelling

7.1 Introduction

In Chapter 6, the classical and phase-shifted optimal 3rd harmonic voltage injections were discussed. The effects of the optimal 3rd harmonic on the peak synthesised voltage was presented. The advantage of the reduced peak synthesised voltage were used in two different ways to improved MMC operation and design. First, the reduced peak synthesised voltage was used to increase the output voltage at the MMC terminals. The second method was considering resizing the SM capacitor due to the increase of the available voltage ripple. All the analysis for the 3rd harmonic injection has been done considering steady state waveforms. However, validation of the method in simulation and experimentally, requires a fully functional MMC with closed loop control which is presented in this chapter.

This Chapter considers the control design and MMC modelling with regards to the assumptions presented in this analysis. The control design is firstly presented including AC and DC controllers to ensure a proper operation. Based on the assumptions made and the control scheme used, MMC modelling is secondly presented including a mathematical model using MATLAB, an average modelling and finally the switching model using PLECS Simulation package.

7.2 MMC Control objectives and Scheme

The principle of operation of the MMC was presented in Chapter 3 as shown by Figure 3.1. The applied control scheme presented in this chapter is based on the assumptions made in Chapter 3 as shown in Figure 7.1. The objective of the control scheme is to ensure an appropriate operation with regards to the assumptions made earlier in the analysis.

There are many control schemes and approaches which have been reported in the literature. However, in general, the MMC control can be divided into two main categories [47, 50]:

- **AC Power Flow Control:** this is to control the power flow (AC voltage and current) from the MMC to the grid and vice versa.
- **Voltage Control:** the objective of the voltage control is to balance the voltage of the N_{SM} floating DC capacitors. This category can also be divided into categories:
 - **Averaging control:** this is to force the limb voltages to remain at a desired average level which is normally the V_{dc} .
 - **Balancing control:** the objective of this control is to force SM capacitors not to charge or discharge beyond a set value.

There many control schemes to control a VSC (i.e. MMC) connected to an AC system. Control in the abc-frame is possible but due to complexity of the control problem for a 3-phase converter system the $\alpha\beta$ -frame is presented to reduce the complexity. Compared to the abc-frame control, the $\alpha\beta$ -frame control reduces the number of plants to be controlled from three to two. Moreover, instantaneous decoupled control of the real and reactive power, i.e. exchanged power between the AC and the VSC systems, is easily possible in the $\alpha\beta$ -frame. However, the control variables are sinusoidal functions of time, i.e. feedback signals, feed-forward signals, and control signals which can lead to steady-state error. However, “zero” steady-state error can achieved in the $\alpha\beta$ -frame control with appropriate bandwidth of the closed-loop control system. The bandwidth must be selected to be adequately larger than the AC system frequency [28]. An alternative solution that can be used is to include complex-conjugate pairs of poles in the compensators at the AC system frequency and other frequencies of interest to increase the loop gain at those frequencies. This approach is known as a Proportional plus Resonant (PR) control [79].

Compared to the $\alpha\beta$ -frame control, dq-frame control variables are DC quantities in the steady state; this has the feature that in the dq-frame control zero steady-state error is readily achieved by including integral terms in the compensators. However, the dq-frame control requires a synchronization mechanism which usually achieved through

the phase-locked loop (PLL). This requirement can be considered a disadvantage of the dq-frame control [28].

Considering an MMC connected to an three-phase AC grid and for the reasons above, the dq frame control is implemented to control the MMC terminal voltage and current [49]. In order to inject the desired optimal 3rd harmonic component and its phase shift, the control signal from the dq controller can be used to determine the 3rd harmonic component. Meanwhile the latter phase-shift can be determined using PQ quadrant selection and a lookup table to select the desired 3rd harmonic phase shift.

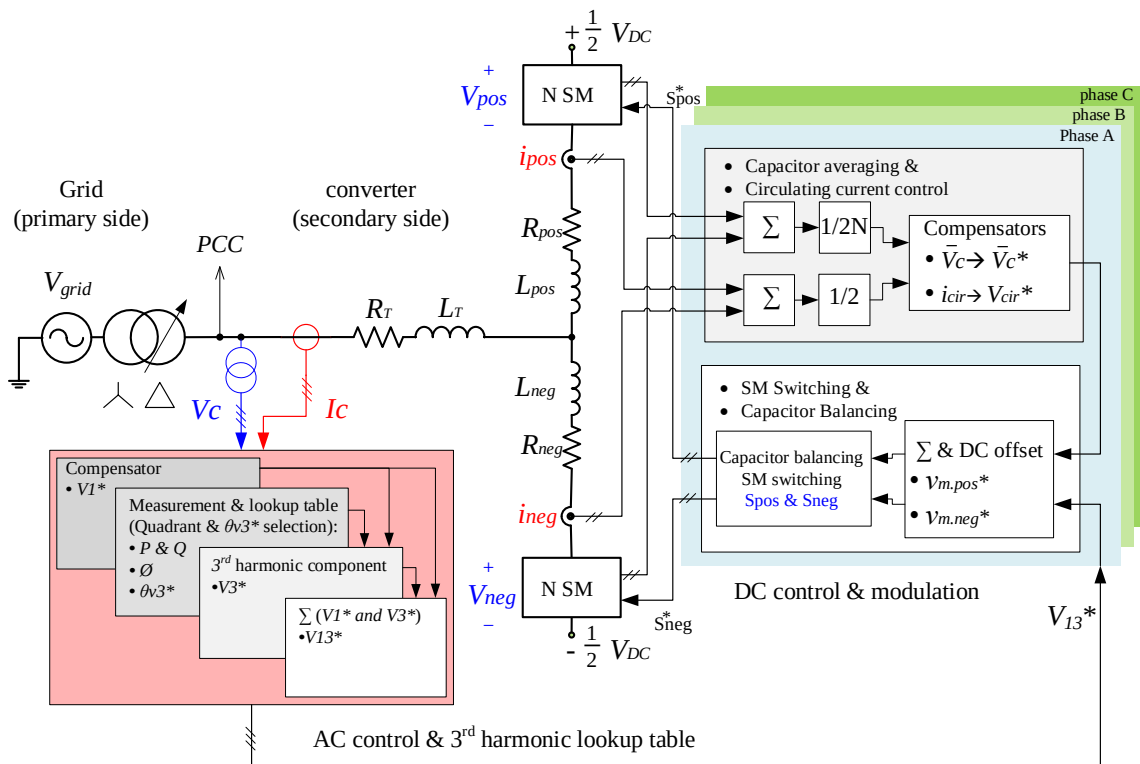


Figure 7.1: A control scheme used for the MMC applying 3rd harmonic voltage injection.

To ensure that the stored energy in the SM capacitors is kept at a given set point, an average capacitor voltage control is used [47]. The authors in [47] proposed a DC control including averaging and a balance controller in order to maintain the energy and the capacitor voltage in each SM at the desired level. Their method is implemented in this thesis, however, instead of controlling each SM capacitor individually, the SM capacitors' voltages are kept balanced using a sorting algorithm [80] which will be presented later. Moreover, the circulating current is controlled to be zero as discussed in Chapter 3 using a proportional resonant controller, PR controller [64].

In this thesis, closed loop AC voltage and power flow control are considered for the fundamental component using dq transformation. However, for the 3rd harmonic component, a feedforward lookup table is used to determine the desired 3rd harmonic voltage injection as will be presented in more details in the next section. Moreover, circulating current suppression and also SM capacitor voltage and energy balancing controllers are also considered.

The formed control scheme shown in Figure 7.1 is based on the control proposed by [47] in addition to DC voltage control and modulation. A combination of dq transformation and optimal 3rd harmonic voltage lookup table are implemented in one block. Therefore, based on the main categories listed in the ballpoints above, the control scheme in this thesis can be divided into two main parts:

- AC controller and 3rd harmonic lookup table (AC voltage and current).
- DC voltage control and modulation.

The objective of the AC control is to control the AC voltage at the MMC terminal to control power flow from/to the grid. First, the three-phase AC voltages and currents, v_{abc} , and i_{abc} , are measured at point of common coupling (PCC) to control the output voltage at the MMC terminals, V_C , by obtaining the fundamental component of the converter demand, V_{C1}^* .

The grid voltage is controlled by the utility (kept fixed with a tolerance $\pm\Delta V$), the MMC terminal voltage, V_C , can be calculated using Equation (6.42) where X_L is the equivalent impedance due to the equivalent reactance seen from the AC grid which will be discussed later and illustrated by Figure 7.3 and Equation (7.1). The MMC terminal voltage, V_C , is controlled in order to meet the active and reactive power demand, exchanged between the VSC system and the AC system, which will be discussed in details in this chapter.

V_{C1}^* varies based on the operating point, i.e. the active and reactive power exchanged between the MMC and the grid. This control signal, V_{C1}^* , will be used to obtain the amplitude of the 3rd harmonic component, $\hat{V}_{C3}^*$, using the optimal value of 1/6 of the fundamental as discussed in Chapter 6, $\hat{V}_{C3}^* = \frac{1}{6} \hat{V}_{C1}^*$. Thus, the amplitude of the 3rd harmonic component varies along with the fundamental component. The phase-shift, θ_{v3} , is determined using a lookup table. The inputs of the lookup table are the

fundamental modulation index, m_{i1}^* , and the power factor, ϕ . m_{i1}^* is determined from V_{C1}^* while ϕ is measured using the MMC output voltage and current. The power factor is first used to determine the operating quadrant of the MMC and then to determine the desired 3rd harmonic phase-shift using a lookup table. The output of this block, V_{C3}^* , is added to the fundamental demand, V_{C1}^* , to obtain the final AC signal, V_{C13}^* . The block diagram of the AC controller is illustrated by Figure 7.2 which is detailed in Section 7.3.

The capacitor voltages of the SMs, $v_{Ci.pos}$ and $v_{Ci.neg}$ are measured for the positive and negative limbs respectively and are used to obtain the average capacitor voltage, $\bar{v}_c^*$. The circulating current, i_{cir} , compensator is fed with the measured limbs currents, i_{pos} and i_{neg} , respectively. The output of the two compensators, $\bar{v}_c^*$ and v_{cir}^* , together with the output of the control signal from the AC compensator, v_{13}^* are fed to a modulation generator block to obtain the modulation demands of the positive and negative limbs $v_{m.pos}^*$ and $v_{m.neg}^*$, respectively. The generated modulation signals are fed to the balance and switching block generator to generate the gate signal for the IGBTs, S_{pos}^* and S_{neg}^* as shown by Figure 7.1.

Table 7.1: Circuit parameters used in the simulation and the experimental results.

Rated output power	P	11.0 kW
Nominal AC system voltage	V_{ac}	170.948 V
Nominal converter AC voltage	V_c	476.944 V
Nominal converter transformer turn ratio	r_t	0.208
DC voltage	V_{DC}	1420 V
Number of SM	N_{SM}	10
Nominal voltage of SM capacitance	V_{cap}	142 V
SM capacitance	C_{SM}	1.08 mF
Inductance of limb inductor	L_{pos}^{**}	6.7 mH
Limb resistance	R_{pos}^{**}	0.51 Ω
AC system inductance	L_t	9 mH
AC system resistance	R_t	2.6 Ω

**The inductance and resistance of positive and negative limb inductors are equal.

It is noted that the analysis presented in the previous chapters are presented based on an HVDC system where the MMC nominal power is 20MW as listed in Table 6.1. In this

chapter, the control design and simulation results are obtained based on scaling down the system in Chapter 6 for the laboratory tests. The scaled down parameters for the experimental rig are listed in Table 7.1. The nominal power of the MMC rig is chosen as 11kW while the SM capacitor voltage is set to 142V where 10 SMs (2×10 SMs = 20 SMs per leg) are utilised for each limb. Therefore, the DC link voltage of the rig is set to 1420V (i.e. $10 \times 142V = 1420V$). The capacitance of the SM capacitor and the limb inductance should be scaled down based on these selections. To ensure realistic validation, the rated stored energy given by (6.49) has to be the same for the two systems, i.e. the original and the scaled down systems. Therefore, the capacitance is calculated to be 1.098mF while the selected capacitance for the experimental work is chosen to be 1.08mF. The calculated the limb inductance is 0.3mH. However, the one available in the laboratory is 6.7mH. This in fact can help reduce the ripple in the limb current.

For a typical HVDC system, the transformer and limb inductances for an MMC are reported in [77] while the equivalent resistance for an HVDC converter is typically small and can be neglected. However, for the scaled down prototype with low voltage specification given in Table 7.1, the measured resistance is 0.51Ω which cannot be neglected. This needs to be considered by the control design. The simulation results obtained by the switching module presented in Section 7.9.2 considered this.

The analysis presented in the previous chapters are achieved for a typical HVDC system with MMC parameters given in Table 6.1. The rated stored energy (kJ/MW) is the same for the analysis and the experimental work. Therefore, this analysis is validated using a switching model which will be presented in Section 7.9.1. A switched model for the prototype presented in Chapter 8 is presented in Section 7.9.2.

The AC control and 3rd harmonic lookup table block is detailed in Section 7.3. The capacitor average voltage and circulating current blocks are considered in Section 0. The capacitor voltage balance and switching generator are given in Section 7.6.

7.3 AC Control (voltage and Power Flow)

In this section, the typical decoupled dq -frame transformation approach is used with the MMC considered in this thesis. DQ transformation approach shows the ability and the ease of implementation with converters based on 3-phase system. Moreover, the main

signal of the dq signal can be used directly to determine the 3rd harmonic signal after considering the required 3rd harmonic phase-shift.

The control signal of the fundamental component, $\hat{V}_{C1}^*$, is used to calculate the 3rd harmonic component signal, $\hat{V}_{C3}^*$. The 3rd harmonic voltage injection normally synchronised with fundamental component and can be achieved by adding the desired 3rd harmonic component signal with the fundamental signal. However, once the phase-shift of the 3rd harmonic is required as in the case of this research, a lookup table is used to provide the controller with the required 3rd harmonic phase angle. Both fundamental and 3rd harmonic components are then added to generate the modulation signal for the MMC. Therefore, the AC control in this thesis is divided into two controllers:

- Fundamental component control.
- 3rd harmonic lookup table.

The fundamental component control is aimed to provide the required fundamental component of the AC voltage at the MMC terminal. The output of this block is fed to the 3rd harmonic component block to generate the required 3rd harmonic component. Both signals then are used to generate the modulation signal for the MMC.

Considering the fundamental component, it is important first to discuss the MMC model which is presented in Section 7.3.1. The block diagram of the combination of the fundamental and 3rd harmonic is illustrated by Figure 7.2.

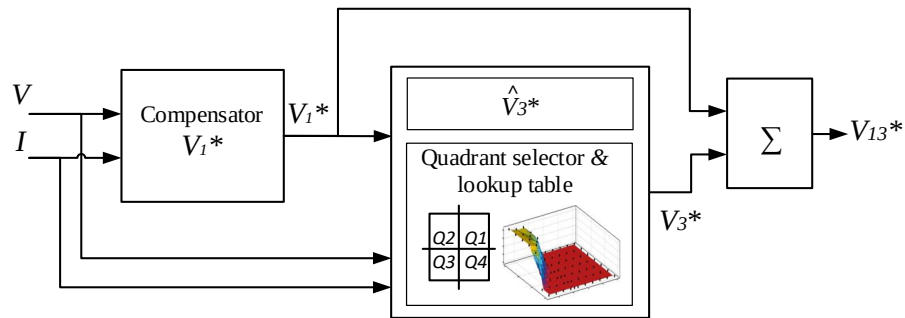
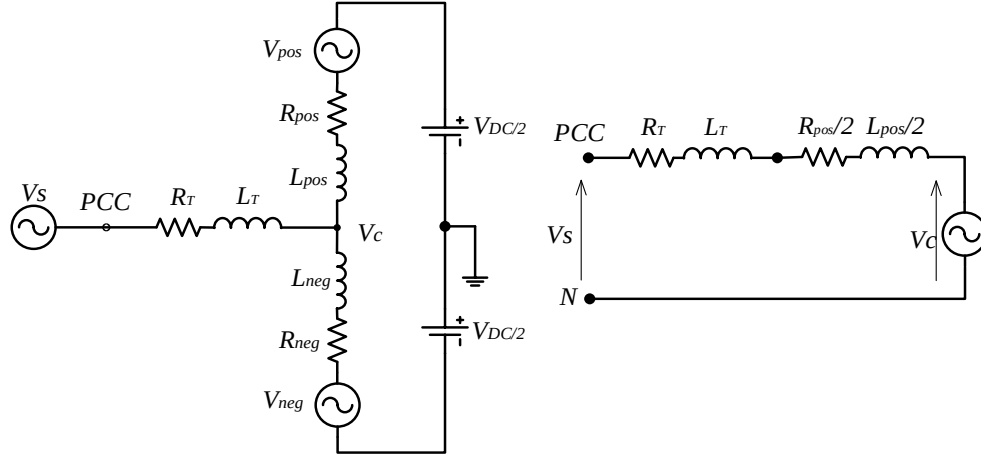


Figure 7.2: Block diagram of AC fundamental and 3rd component control.

7.3.1 The AC system and MMC model

In order to design the controller parameters, the transfer function from i_{abc} to V_{C1}^* is first deduced. The MMC model and its equivalent are presented by Figure 7.3. It can be noted that the inductor of positive and negative limbs are equal and are connected in

parallel. Therefore, their equivalent inductance is presented by the half of positive or negative inductance as shown by Figure 7.3.b. The detailed derivation of the dynamic equations is presented in Appendix E.



(a) AC system and MMC model

(b) AC MMC equivalent model

Figure 7.3: The AC MMC model and its equivalent single line diagram.

Based on the AC MMC equivalent model illustrated by Figure 7.3.b, the output voltage at the MMC terminals can be represented similarly to the conventional VSC by (7.1) [28].

$$v_{s,j} = (Ri_j + L \frac{di_j}{dt}) + v_{c,j} \quad (7.1)$$

where R is the resistance in the AC system considering the MMC as seen from the utility grid. For an HVDC system based on VSC converters, the internal resistance R_{pos} and R_{neg} of the MMC is small enough and it can be neglected. However, this will be considered to give a general analysis. L is the system equivalent inductance including line, converter transformer and half of the limb inductance. j is the three phases a , b and c . where;

$$L = L_T + \frac{L_{pos}}{2} \quad (7.2)$$

And;

$$R = R_T + \frac{R_{pos}}{2} \quad (7.3)$$

Using a dq transformation, (7.1) can be represented in the frequency domain by the dq transformation frame as expressed by (7.4) and (7.5) as shown in the Appendix E.

$$v_{s,d}(s) = (R + Ls) i_d(s) - L\omega_s i_q(s) + v_{c,d}(s) \quad (7.4)$$

$$v_{s,q}(s) = (R + Ls) i_q(s) + L\omega_s i_d(s) + v_{c,q}(s) \quad (7.5)$$

Based on (7.4) and (7.5), the transfer function can be obtained as:

$$G_{idq}(s) = \frac{i_{dq}(s)}{v_{c,dq}(s)} = \frac{1}{R + Ls} \quad (7.6)$$

where R and L are the equivalent resistance and inductance given by (7.2) and (7.3) respectively. The dq transformation can be obtained using abc/dq transformation as illustrated in Figure 7.4 where θ is the angular position in radians [39]. θ can also be calculated directly using the stationary frame, $V_{\alpha\beta}$, where $\theta = \tan^{-1} \left(\frac{V_\beta}{V_\alpha} \right)$.

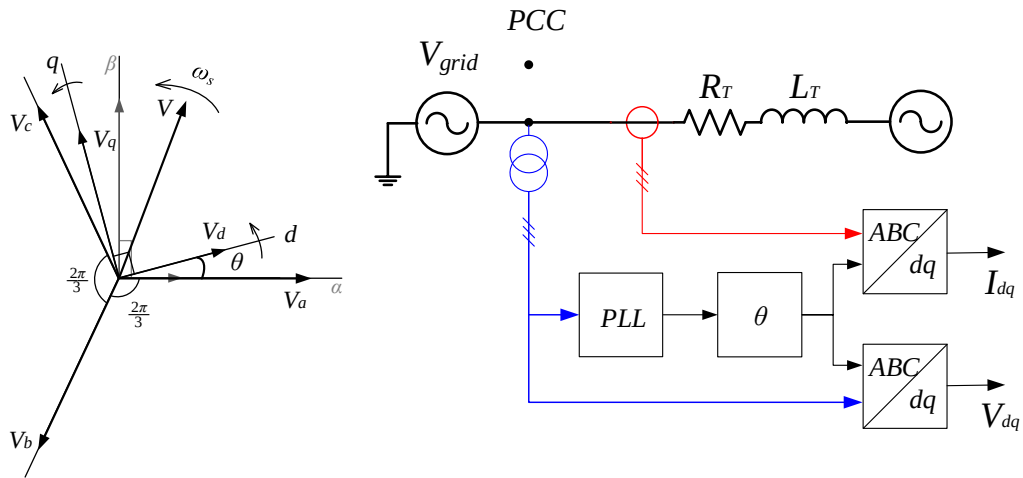


Figure 7.4: Block diagram of the transformation from stationary abc to the dq rotating frame.

7.3.2 Fundamental component AC controller Design

Based on the transfer function given by (7.6) and the control scheme presented in Figure 7.1, the control design is discussed in this subsection.

The control block diagram of the fundamental component is presented by Figure 7.5. The classical root locus controller design is adopted. Typically, an integral term must

be included in the compensator to eliminate the steady-state error of the AC current in the controller design. Therefore, the proportional-integral (PI) controller chosen for $K_i(s)$ is given by (7.7) while the proportional and integral gains are given by (7.8) and (7.9), respectively.

$$K_i(s) = k_p + \frac{k_i}{s} \quad (7.7)$$

$$K_p = 2L\omega_0\zeta - R \quad (7.8)$$

$$K_i = \omega_0^2 L \quad (7.9)$$

The block diagram of the PI controllers for the rotating dq currents is shown by Figure 7.6 where the decoupling feedforward has been taken into account.

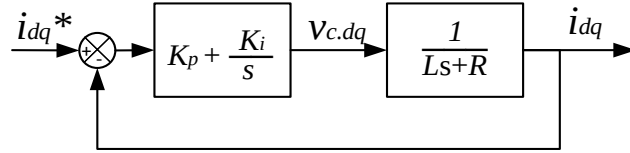


Figure 7.5: Control block diagram of the AC voltage and current controller.

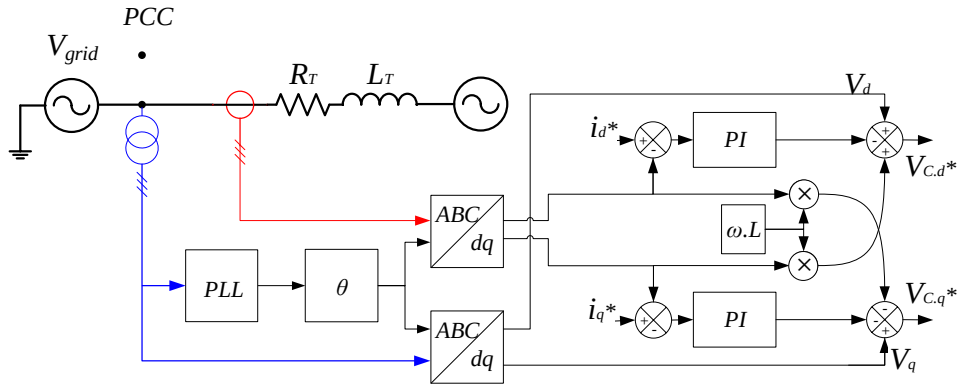


Figure 7.6: The control block diagram including the decoupling feed-forward.

The controller parameters implemented in the simulation model are given by Table 7.2.

Table 7.2: AC power flow control parameters

Control Parameter	
ω_o	$200\pi \text{ rad/s}$
ζ	0.707
K_p	16.9
K_i	7520

7.3.3 3rd Harmonic Lookup Table

In this section, the 3rd harmonic component of the synthesised voltage will be discussed in detail.

As discussed in Sections 7.2 and 7.3 and shown by Figure 7.1 and Figure 7.2, the detailed block diagram of the 3rd harmonic component is presented by Figure 7.7. The modulation index, m_{i1}^* , of the control signal of the fundamental component is first calculated. This signal has two objectives including the calculation of the optimal 3rd harmonic modulation index, m_{i3}^* expressed by (7.10), and also to select the 3rd harmonic phase shift, θ_{v3}^* . Where, the optimal m_{i3}^* is obtained using a lookup table.

$$m_{i3}^* = \frac{1}{6} m_{i1}^* \quad (7.10)$$

It is noted that the fundamental component is synchronised with the grid using a PLL at the PCC point where the reference angle is set to zero, $V_s \angle 0^\circ$. However, the 3rd harmonic component needs to be synchronized with the fundamental component control signal to ensure that it has the same angle reference, $V_{c1}^* \angle \vartheta^\circ$, where ϑ° is the angle between the grid voltage and the MMC terminal voltages. Otherwise, the 3rd harmonic will always be shifted by ϑ° . The total dq control scheme is given by Figure 7.8.

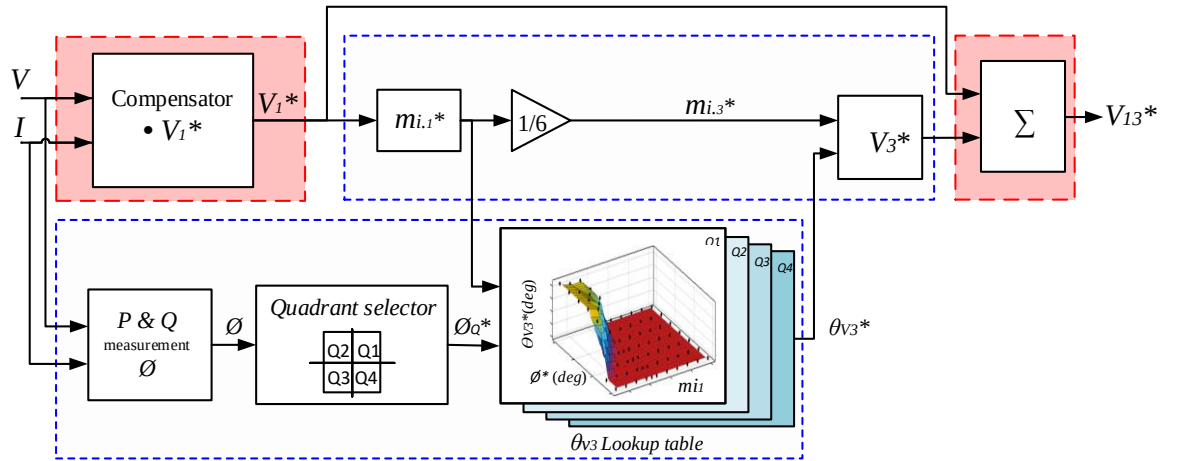


Figure 7.7: The block diagram of AC fundamental and 3rd harmonic component control.

correct order of the quadrants where j is 1,2,..4. This is because each of the four quadrants has its own lookup table. This has been discussed in detail in Chapter 6. The flow chart in Figure 7.9 illustrates the process of determining the operation point used.

The output of the phase-shift selector, θ_{v3}^* , is used with the output of the 3rd harmonic modulation index block, m_{i3}^* . Both signals are used to generate the 3rd harmonic voltage component demand signal, v_3^* where the latter is given by (7.11).

$$v_3(t) = m_{i3}^* \cos(3\omega_1 t - \theta_{v3}^*) = \frac{1}{6} m_{i1}^* \cos(3\omega_1 t - \theta_{v3}^*) \quad (7.11)$$

The output signal from the AC fundamental and 3rd harmonic controller can then be added to generate the final AC control signal, V_{13}^* . This signal is then be ready to be used with the DC control signals to generate the modulation signal for each limb, positive and negative limbs, $m_{pos}(t)$ and $m_{neg}(t)$, respectively. The DC control is detailed in the next section. DC Voltage Control and Modulation

Based on the control scheme shown by Figure 7.1, the objective of this control block is to generate the modulation and switching signals of the SMs in order to synthesise the required output voltage and meet the power flow demand. In the meantime, the SM capacitor average voltage of each limb is maintained to a desired level while providing a voltage balance to each individual SM capacitor. The control block diagram of the DC voltage control and modulation is illustrated by Figure 7.10.

In order to generate the modulation signal, the AC control signal is first obtained by the AC controller as discussed in the previous section. The DC control signal, including average and circulating current suppression, then needs to be generated. This is discussed in the next sections, Sections 7.4 and 7.5. Once the control signals are obtained, the modulation signal then can be achieved by taking into account the positive and negative limb voltages as expressed by (4.12) and (4.14). The modulation signals are given by (7.12) and (7.13).

$$v_{m.pos} = \frac{1}{2} - \frac{V_{13}^* + \bar{V}_{cir}^*}{V_{DC}} \quad (7.12)$$

$$v_{m.neg} = \frac{1}{2} + \frac{V_{13}^* - \bar{V}_{cir}^*}{V_{DC}} \quad (7.13)$$

where V_{13}^* and $\bar{V}_{cir}^*$ are the AC and DC control signal respectively.

$$\bar{V}_{cir}^* = V_{cir}^* + \bar{V}_c^* \quad (7.14)$$

It is noted that the DC control signal is a combination of the average and 100Hz circulating current controllers, $\bar{V}_c$ and V_{cir} , respectively. Both controllers are detailed in Sections 7.4 and 7.5 [15].

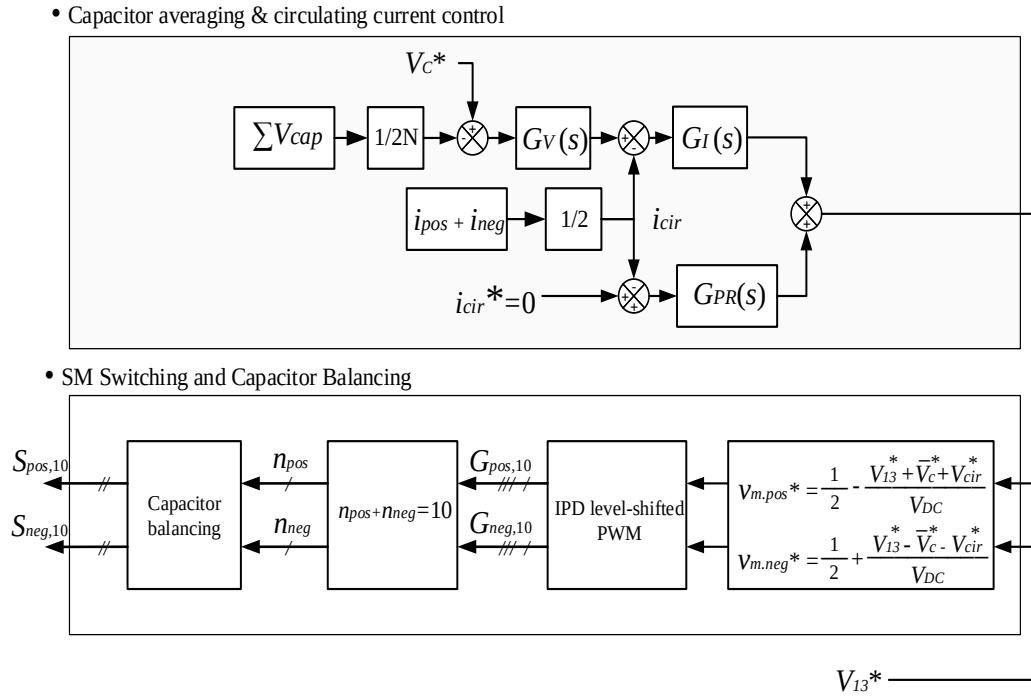


Figure 7.10: Control block diagram of the DC control and modulation.

The modulating signals given by (7.12) and (7.13) are used to generate the gate signals using a suitable PWM scheme. A good harmonic performance (THD) can be achieved using the In-phase disposition (IPD) PWM especially with multilevel converters [39, 81]. However, voltages on the individual SMs cannot be balanced using IPD, so a rotation algorithm is required as detailed later.

The IPD PWM scheme is implemented where a number of N_{SM} carriers is required to generate a $(N_{SM} + 1)$ output voltage level at the MMC terminals. The IPD PWM modulating signal and ten carriers for each limb of the MMC in this thesis are illustrated by Figure 7.11 [82]. In this thesis, the applied switching frequency of the power semiconductor devices is 6kHz and therefore the equivalent switching frequency is 360kHz [15].

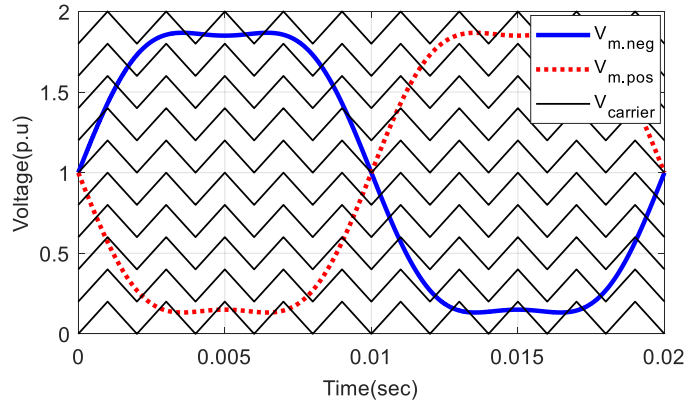


Figure 7.11: In-phase desposition (IPD) level-shifted PWM.

The ten output gate signals, $G_{pos,10}(t)$ and $G_{neg,10}(t)$, from the IPD level-shifted PWM unit will be used to determine the required number of SM, $n_{pos}(t)$ and $n_{neg}(t)$, that needs to be switched ON/OFF for each limb. It is noted that, the ideal operation should ensure that the total number of SMs switched ON or OFF for the one arm is equal to $N_{SM} = 10$. The signal of the required number of the SMs for the positive and negative limb, $n_{pos}(t)$ and $n_{neg}(t)$, are generated to be used with a suitable rotating algorithm to ensure balanced capacitor voltages. The switching signals, $S_{pos,10}(t)$ and $S_{neg,10}(t)$, for the positive and negative limbs are then generated accordingly. In the next sections, the control design is detailed considering the circuit parameters in Table 7.1

7.4 Averaging Control

The averaging control design is presented in this section. The averaging control is used to force the voltage of each limb to remain at a desired set point. In order to control the average voltage, the circulating current needs to be controlled as illustrated by Figure 7.12. Overall, this control monitors the energy stored in the SMs individually and for all of the limbs and legs.

The implemented control considers the scheme introduced by [47]. However, unlike [47], for each phase, one control signal of the average control is generated to control the entire arm, positive and negative limbs, while capacitor voltages were controlled individually in [47]. This is to reduce the complexity. The used balancing control uses a rotating algorithm so that the capacitors are balanced and their average voltage is equally averaged.

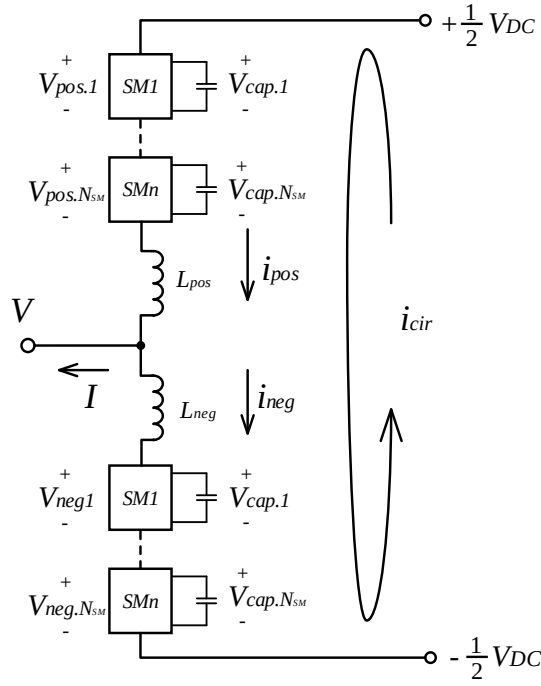


Figure 7.12: The circuit diagram of one arm of MMC and its circulation current.

In order to control the average voltage of the limbs, current and voltage controllers are required as shown by the block diagram in Figure 7.13. The demand signal of the inner loop of the current controller is fed from the voltage controller. The detailed control design is considered below.

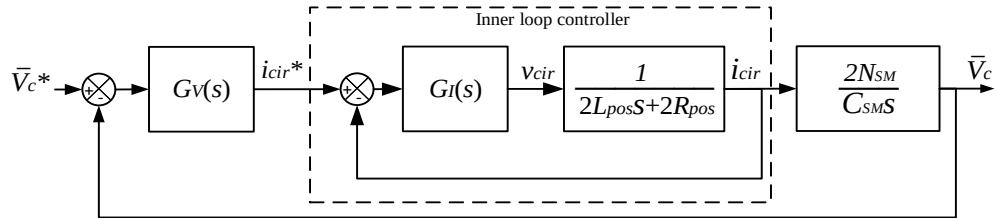


Figure 7.13: Control block diagram of average voltage control.

7.5.1 Inner loop Current Controller

To control the average voltage, the circulating current needs to be forced to follow its reference value. This can be firstly achieved by obtaining the transfer function after considering the equation below (7.15).

$$V_{DC} = \sum_{j=1}^{2N_{SM}} V_{SM,j} + (R_{pos} + R_{neg})i_{cir} + (L_{pos} + L_{neg}) \frac{di_{cir}}{dt} \quad (7.15)$$

Considering the symmetry of the positive and negative limbs, the above equation can be re-expressed by (7.16).

$$V_{DC} = V_{SM,j}^{\Sigma} + 2R_{pos}i_{cir} + 2L_{pos} \frac{di_{cir}}{dt} \quad (7.16)$$

Where $V_{SM,j}^{\Sigma}$ is the sum of the inserted voltages by number $2N_{SM}$ of SMs, $\sum_{j=1}^{2N_{SM}} V_{SM,j}$ of one arm, both positive and negative limbs. $V_{SM,j}^{\Sigma}$ can be replaced by the average voltage of the limb expressed by $\bar{V}_C$. It can be noted that the number of SMs switched ON/OFF for one arm at any instant time, $n_{SM}(t)$, is N_{SM} , where;

$$n_{SM}(t) = N_{SM} \quad (7.17)$$

The transfer function of the capacitor voltage average to the circulating current can be expressed by (7.18).

$$G_I(s) = \frac{i_{circ}(s)}{\bar{V}_C(s)} = -\frac{1}{2R_{pos} + 2L_{pos}s} \quad (7.18)$$

Considering Figure 7.12, the current i_{cir} , cannot be measured directly as it is loop current. However, this can be calculated by (7.19).

$$i_{cir} = \left(\frac{I_{pos} + I_{neg}}{2} \right) - \frac{I_{DC}}{3} + i_{cir.100Hz} \quad (7.19)$$

where $i_{cir.100Hz}$ is the circulating current with double the frequency of the fundamental component which is controlled to zero using a PR controller as will be discussed later.

7.5.2 Outer Loop Voltage Controller

By considering the DC loop, the average DC loop voltage equation can be written by (7.20) and thus (7.21).

$$\bar{V}_C = \frac{1}{2N} \sum_{j=1}^{2N} V_{cap,j} \quad (7.20)$$

Thus;

$$\frac{d\bar{V}_C(t)}{dt} = \frac{1}{C_{eq}} i_{cir}(t) \quad (7.21)$$

where $C_{eq} = \frac{C_{SM}}{2N_{SM}}$.

The transfer function can be obtained by (7.22):

$$G_V(s) = \frac{\bar{V}_C(s)}{i_{cir}(s)} = \frac{2N_{SM}}{C_{SM}s} \quad (7.22)$$

where $\bar{V}_C$ in the equation represents the average voltage of the SMs for one arm and similarly for phases B and C.

7.5.3 Control design

From the control block diagram shown in Figure 7.13, the inner controller can be represented by Figure 7.14. The classical root locus controller design is used where the implemented transfer function of the inner controller is given by (7.23).

$$G_I(s) = K_{p,I} + \frac{K_{i,I}}{s} \quad (7.23)$$

where, the proportional and integral gains of the current controller are represented by $K_{p,I}$ and $K_{i,I}$, respectively. These are given by (7.24) and (7.25).

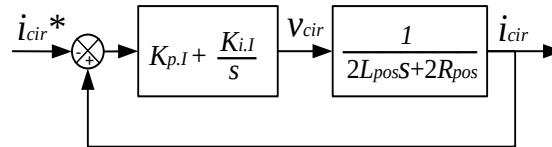


Figure 7.14: The block diagram of the inner control loop.

$$K_{p,I} = 4\zeta\omega_o L_{pos} - 2R_{pos} \quad (7.24)$$

$$K_{i,I} = 2\omega_o^2 L_{pos} \quad (7.25)$$

Considering Figure 7.13, it should be noted that the controller of the inner loop should be faster than the outer loop so that it reaches the steady state faster than the outer loop. This allows the inner controller enough time to compensate for inner loop disturbances before they affect the outer loop. Otherwise the inner loop will keep constantly correcting the disturbances and will be unable to apply a consistence signal. This means

that the bandwidth of the inner loop should be greater than the outer loop. Therefore, the gain of the closed inner loop is $G_I(s) = \frac{1}{1+\tau s} \cong 1$ as illustrated by the block diagram of outer control loop represented by Figure 7.15.

The implemented transfer function of the outer controller is given by (7.26).

$$G_V(s) = K_{p.V} + \frac{K_{i.V}}{s} \quad (7.26)$$

where, $K_{p.V}$ and $K_{i.V}$ are the proportional and integral gains of the voltage controller and are represented by Figure 7.15. The controller gains are given by (7.27) and (7.28), respectively.

$$K_{p.V} = \zeta \omega_o \frac{C_{SM}}{N_{SM}} \quad (7.27)$$

$$K_{i.V} = \omega_o^2 \frac{C_{SM}}{2N_{SM}} \quad (7.28)$$

The controller parameters and gains of the inner and outer controllers are given by Table 7.3 below.

Table 7.3: DC control parameters.

Inner controller		Outer controller	
ω_{oI}	20Hz	ω_{oV}	5Hz
ζ_I	0.7	ζ_V	0.7
$K_{p.I}$	7.07	$K_{p.V}$	0.0024
$K_{i.I}$	634.8	$K_{i.V}$	0.053

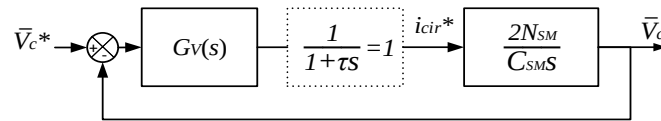


Figure 7.15: The block diagram of the outer control loop.

7.5 100Hz Circulating Current Suppression

In this section, the 2nd harmonic component which appears in the circulating current expressed by (7.19) is controlled. The advantages of using a PR controller over the PI controls has already been described especially with single phase systems including the steady state error and the need for dq transformation [83]. In this thesis, therefore, the

alternative controller used instead of PI is the PR controller. The objective of a PR controller is to eliminate the 2nd harmonic component of the circulating current [64].

One of the important features of the PR controller is that it creates small gain at zero frequency but a large gain at the resonant frequency. Ideally, the PR controller can create an infinite gain at the resonant frequency and zero at zero frequency. However, the infinite gain can cause stability issues. Therefore, no-ideal PR controller can be used where a finite gain can be obtained. The ideal and no-ideal PR controllers are expressed by (7.29) and (7.30) respectively.

$$G_{PR}(s) = K_{p.PR} + \frac{2K_{i.PR}s}{s^2 + \omega_o^2} \quad (7.29)$$

$$G_{PR}(s) = K_{p.PR} + \frac{2\omega_c^2 K_{i.PR}s}{s^2 + 2\omega_c s + \omega_o^2} \quad (7.30)$$

where, K_p and K_i are the proportional and integral gains respectively. ω_o is the resonant frequency where the bandwidth can be widened using the non-ideal PR controller by ω_c frequency.

The block diagram of the DC control and modulation scheme is given by Figure 7.10 while a block diagram of the PR controller is shown by Figure 7.16. It is noted that the 2nd harmonic of the circulating current is suppressed and therefore its reference is set to zero. However, 2nd harmonic current can be used to reduce the energy deviation. Therefore, the reference set point is set to the desired one [69]. In this control design, the reference is set to zero.

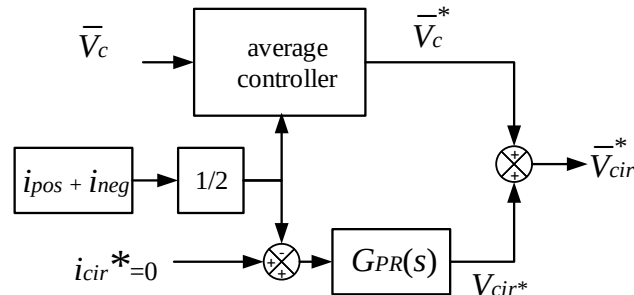


Figure 7.16: Control block diagram of the PR and the average controllers.

The control design is included with the experimental MMC parameters considered in Chapter 7. The MMC parameters are given in Table 7.1. At the resonant frequency of

$200\pi \text{ rad/s}$, the gain of $G_{PR}(s)$ needs to be large. The gain at zero frequency on the other hand needs to be small to minimise the influence on the other controllers. Therefore, the proportional and integral gains of the controller are chosen to be $K_{p.PR} = 0.1$, whilst $K_{i.PR} = 500$.

7.6 Balancing Control of the SMs Capacitors

The average voltage of the MMC has already been discussed where the average of the voltage among each limb and therefore among each arm is averaged. To ensure that this average is also controlled within the individual SMs, the modulation scheme needs to be chosen accordingly. The modulation scheme used in this thesis was discussed earlier and shown in Figure 7.10 using the IPD level-shifted PWM presented by Figure 7.11. Therefore, unlike phase-shifted PWM, IPD approach can cause a balancing voltage issue due to unequally switched SMs. Therefore, a rotating or sorting algorithm is needed to overcome the balancing issue.

Traditional sorting algorithms have disadvantages with regards to long time for processing and also high switching frequency [10]. A dual sorting algorithm is used [84, 85] which overcomes the drawbacks of the traditional sorting algorithm. The dual sorting offers better performance by mitigating the issues associated with time consumption and also the high switching frequency [80]. The approach used in this thesis is based on a dual sorting algorithm. However, unlike [15], the dual sorting control level used is the last control and modulation stage whilst in [15] it is implemented during PWM. The dual sorting approach used in this thesis is illustrated by Figure 7.17, whilst the flow chart is illustrated by Figure 7.18. As discussed in section 7.4 and illustrated by Figure 7.10, n_{pos} and n_{neg} are the required number of SMs required to switch ON/OFF.

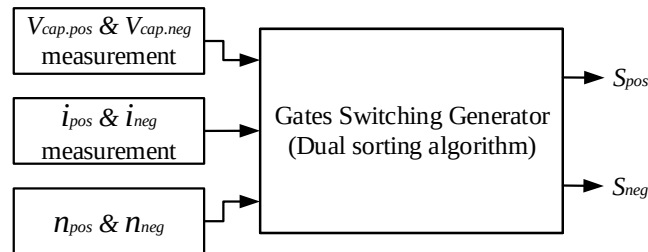
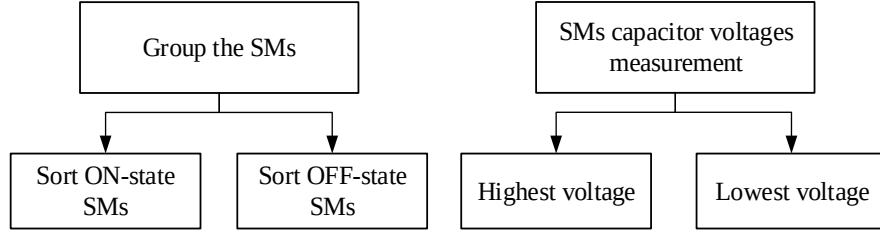


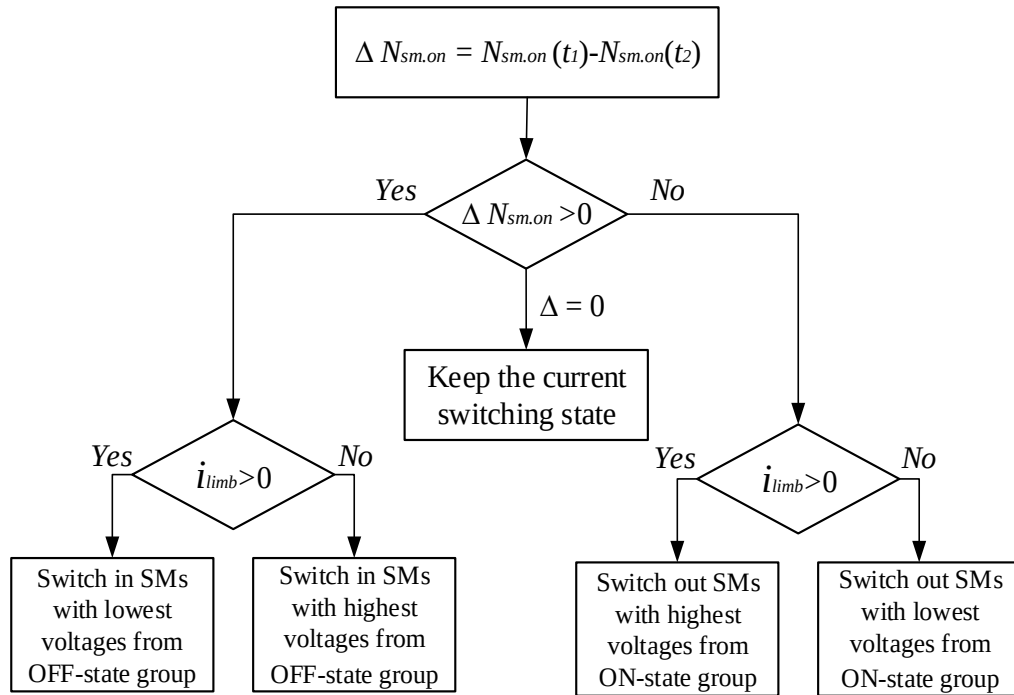
Figure 7.17: Block diagram of the dual sorting algorithm scheme and the required inputs.

The measured SM capacitor voltages are sorted into two groups, the highest voltages group and the lowest voltages group. These groups will be used to select which capacitors need to be inserted during the charging, or bypassed in case of discharging. Charging and discharging states are determined by the direction of the measured limbs currents.



(a) Grouping the SMs ON/OFF-state.

(b) Grouping the SMs capacitor voltages.



(c) Dual sorting algorithm.

Figure 7.18: Flow chart of dual sorting algorithm [80, 85].

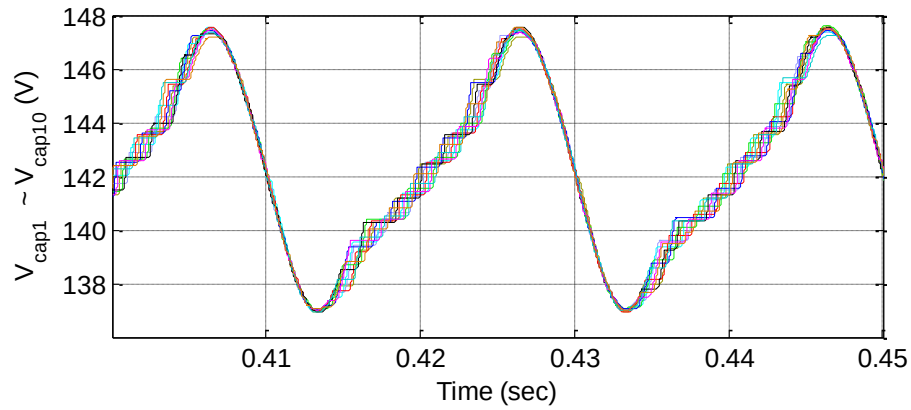


Figure 7.19: Voltage of the SMs capacitors balance due to dual sorting algorithm.

Figure 7.19 shows the waveforms of the capacitors voltages of 10 SMs for the MMC converter presented in Table 7.1. The operation point of the MMC was chosen so that the MMC delivers 1.0 p.u of positive real power, P (+ve), whilst $Q=0$. It can be seen that a good balancing voltage can be achieved due to a dual sorting algorithm. The simulation result is obtained using the PLECS package. However, for better plotting MATLAB was used.

7.7 MMC Modelling and Simulation Results

The simulation results presented in the previous chapters were obtained using a MATLAB simulation package based on the mathematical equations presented in the analysis. In order to validate the analytical study and its mathematical model, an average model based in the PLECS simulation package was used. It is noted that to achieve realistic results, the impact of the AC system line impedance should be considered. The impact of the AC impedance will, of course, affect the phase angle and amplitude of the target voltage depending on PQ. Therefore, the modulation index is affected as discussed earlier where the output voltage if the MMC terminal can be given by Equation (6.42).

In this section, the focus will be on validating the analytical study presented in this thesis using a switching model based on the MMC parameters given by Table 7.1. The table presents the circuit's parameters used for an experimental rig. The latter has already been built to study the modular multilevel concept with two modes of operation. These are the traditional MMC and also the ACC circuit.

The benefit of using 3rd harmonic injection on increasing the fundamental component of the output voltage at the MMC terminals is further validated using the experimental

rig as presented in Chapter 6. However, using the 3rd harmonic to re-design the SM capacitors as presented by Section 6.5.2 in Chapter 6 is only validated using the switching model presented in this chapter. In addition, the modelling and the control scheme presented in this chapter will also be validated. Experimental results based on the laboratory scale will be considered in Chapter 8.

In this Section, a detailed MMC switching model based on PLECS simulation package is presented. The presented simulation results are aimed at validating the analytical research and the average models presented in the thesis.

7.8.1 AC System

The AC system including the converter transformer is considered here. A three phase AC voltage source is used. The converter transformer is modelled using a lossless three phase transformer with the possibility of changing its turn ratio as discussed in Chapter 5. This in order to force the MMC to operate with higher modulation index as discussed in Chapter 5. The primary side of the transformer however is kept fixed as given in Table 7.1.

The transformer impedance and losses are modelled using line impedance where the primary side impedance is transferred to the secondary side as shown in Figure 7.1. Therefore, the point of common coupling, PCC, now can be considered as the secondary side for a lossless transformer.

7.8.2 MMC Modelling

The configuration of the generalised MMC converter presented in this thesis is based on the half-bridge submodule as shown by Figure 3.1.

The MMC is modelled with three identical legs. Each arm has two identical limbs considering the impact of the buffer inductor and internal resistance as given by Table 7.1. Each limb has ten SMs where the individual SM is modelled as presented here for more convenience by Figure 7.1. The DC-link is modelled with two identical and ideal DC voltage sources where no DC inductance or resistance is used.

The inserted voltage of the SM is measured between the 10 SMs of the limb while the available voltage is the sum of the measured capacitor voltages. Using the measured

limb voltage and current it is possible to calculate the instantaneous power and energy for each limb.

The optimal 3rd harmonic is synchronised with the fundamental. This is in order to ensure that at classical 3rd harmonic operation, the zero crossing is equal for both fundamental and 3rd harmonic component as shown by Figure 4.3. The phase-shifted 3rd harmonic is simulated based on the discussion on the control where a lookup table was used. The quadrant selector is also considered to provide the correct signal for the lookup table to select the required phase shift, Section 7.3.3.

7.8 Simulation Results

In this section simulation results based on the detailed switching model are presented for both the 20MW and the 11kW MMC using the MMC circuit parameters in Table 6.1 and Table 7.1.

The simulation results are organised into two main categories including;

- Waveforms based on the rated design specification for 20MW MMC given by Table 6.1. The waveforms of the limbs voltages and currents are presented. The available voltage and also the capacitor balancing and circulating currents are presented. It is noted that the operating point of the MMC is $S=0.94-j0.33$ p.u. The two methods presented in Chapter 6 that could help to improve the MMC operation or design are validated by simulation results.
- Simulation results will be also presented based on the ratings of the experimental rig given by Table 7.1. The focus of the simulation results is on the two methods used to benefit from the reduced peak synthesised voltage presented in Chapter 6 including increase the voltage at the MMC terminals or reduce the SM capacitor size.

7.9.1 Simulation results for the 20MW MMC

It is noted that the previous control parameters are given for the 11kW MMC. Therefore the control parameters of the 20MW MMC parameters given by Table 6.1 are given in the tables below. The switching frequency is 6kHz which the equivalent switching frequency is 360kHz.

The waveforms presented in the figures below are based on the switching model of the MMC parameters given in Table 6.1. A step response at time of 0.25 sec of the controllers from normal operation to traditional 3rd harmonic injection has been presented. Figure 7.20 shows the waveforms of the positive and negative limb synthesised voltages and currents.

Table 7.4: AC control parameters of 20MW MMC.

AC current controller	
ω_o	$200\pi \text{ rad/s}$
ζ	0.707
K_p	4.1
K_i	1818

Table 7.5: DC control parameters of 20MW MMC.

Inner controller		Outer controller	
ω_{oI}	20Hz	ω_{oV}	5Hz
ζ_I	0.707	ζ_V	0.707
$K_{p.I}$	1.92	$K_{p.V}$	0.0085
$K_{i.I}$	342.3	$K_{i.V}$	0.189

Table 7.6: Circulating current control parameters of 20MW MMC.

Control Parameter	
$\omega_{o.PR}$	$200\pi \text{ rad/s}$
$K_{p.PR}$	0.1
$K_{i.PR}$	500

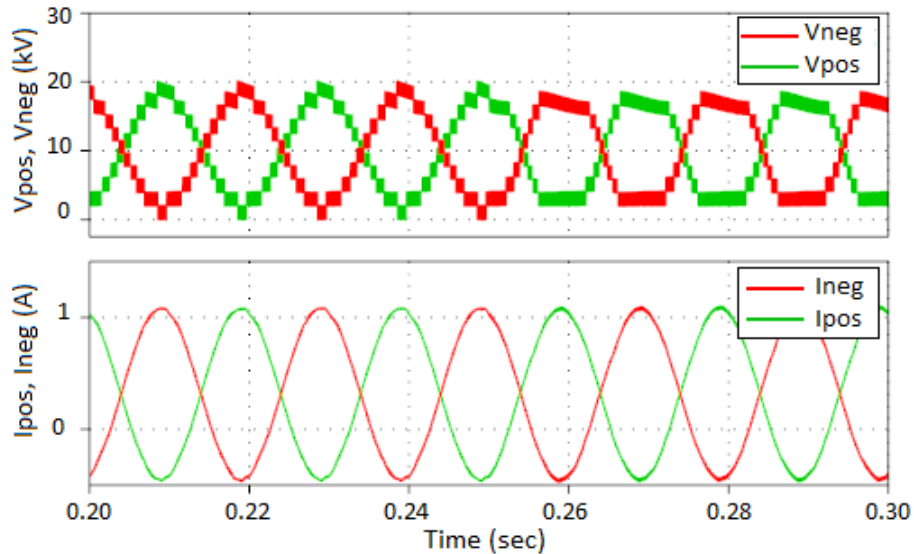


Figure 7.20: Positive and negative limb voltages and currents for phase A.

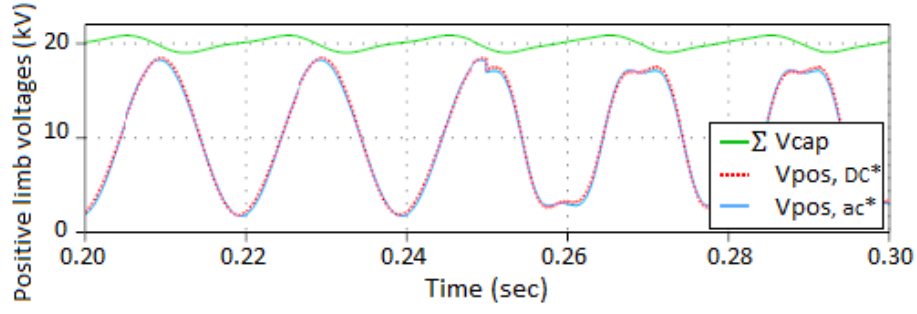


Figure 7.21: Positive limb available and synthesised voltages with and without the DC control.

Positive limb voltages including the available voltage and the generated modulators before and after the DC controller is shown by Figure 7.21 where a slight difference can be noted due to the DC controller. The phase voltage at the MMC terminals and the line currents are presented in Figure 7.22. Figure 7.23 shows the feedback of the capacitor voltage balancing and the circulating current.

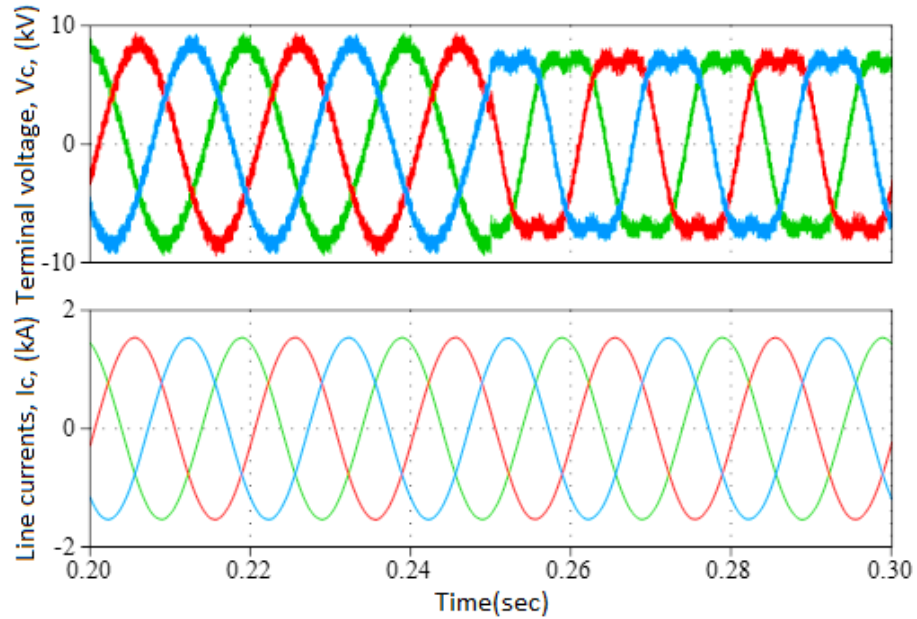


Figure 7.22: MMC terminal AC voltage and current.

Considering Figure 7.1 and for $R_T \ll X_{LT}$, the single line and its phasor diagram is illustrated by Figure 7.24. Figure 7.25 shows the waveforms of the available and synthesised voltages due to the increase of the fundamental component as discussed in Section 6.5.1 where two operating points are considered, point A ($P=0.94$ and $Q=-0.33$) and D ($P=0.94$ and $Q=+0.41$), ($\phi = -19.26^\circ$ and $+22.3^\circ$) respectively as illustrated by Figure 7.24.

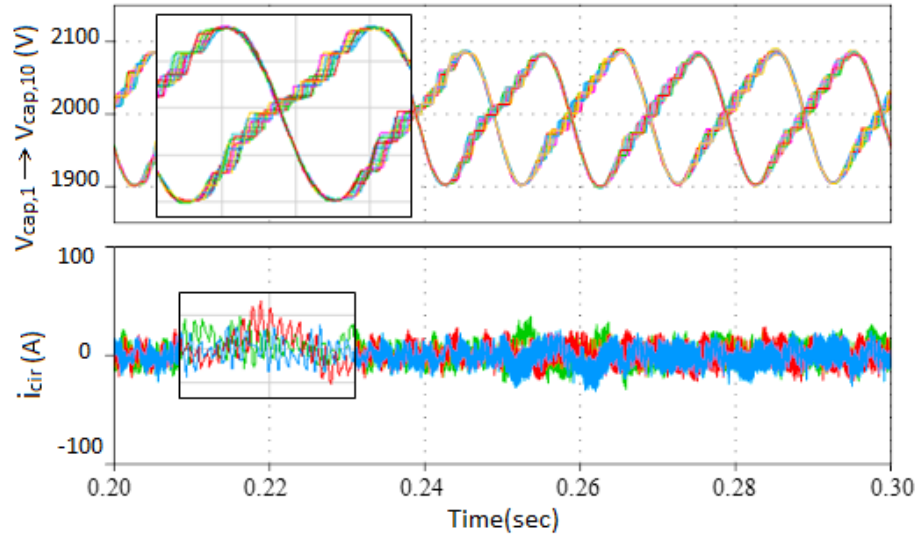


Figure 7.23: SM capacitors voltages balancing and legs circulating currents.

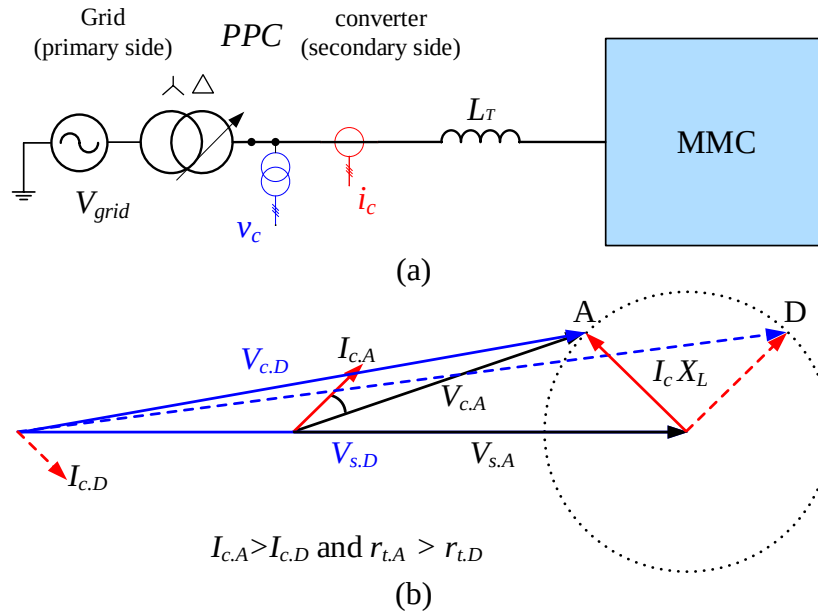


Figure 7.24: (a) The single line diagram of MMC and (b) its phasor diagram of the inverting and inductive and capacitive mode of operation.

The waveforms below represent the simulation results of the switching model where a step response at $t = 0.26\text{s}$ is presented by increasing the source voltage (grid) by 1.15% or by changing the transformer turn ratio (in case of using a tap changer transformer) to the corresponding increase of 1.15%. In the first case, the grid voltage is changed (increased). However, in the case of using tap changer transformer to increase the secondary voltage as shown by Figure 7.24, the grid voltage is kept fixed which are implemented with the switching model. The first method are used with the experimental

rig using a controllable AC source. The results however is the same as the idea is to force the MMC to increase its synthesised voltage above 1.0 p.u.

It is noted that the waveforms in Figure 7.25 represents the MMC operating with the nominal voltage and rated power as given in Table 6.1 with classical 3rd harmonic voltage injection ($t < 0.26$ s). However, when ($t \geq 0.26$ s), the tap changer transformer is used to increase the secondary voltage to 1.15 p.u whilst the primary voltage is kept fixed. It can be seen that the controller forces the MMC to increase its synthesised voltage using the phase-shifted 3rd harmonic voltage injection. Figure 7.26 shows a zoomed view of Figure 7.25 where both upper and lower limits are considered. The solid and dashed lines represent the waveforms for $Q = -0.33$ p.u ($\phi = -19.2^\circ$) and $Q = +0.41$ p.u ($\phi = +22.3^\circ$) whilst $P = +0.94$ p.u in both cases.

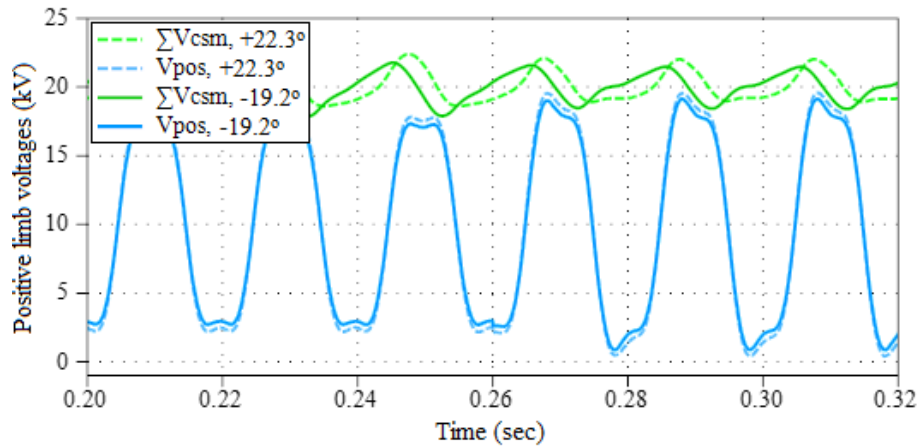


Figure 7.25: Synthesised and available voltages when increasing the fundamental by 1.15% for $S = 1.0$ p.u and $\phi = -19.26^\circ$ and 22.3° .

The operating points A and D represents two of the four worst operating points of the MMC, when the MMC operates as an inverter. The first point, A ($P > 0$ and $Q < 0$), is considered to be the worst case with regards to the minimum available voltage when the MMC operates as an inductor whilst the second, D ($P > 0$ and $Q > 0$), the maximum synthesised voltage when the MMC injects reactive power to the grid. Considering the upper limit and lower limit as shown by Figure 6.8, the terminal voltage of the MMC parameters given in Table 6.1 is increased by 1.15% as given by Equation (6.24) for both worst operating points, A and D using the advantage of phase shifted 3rd harmonic voltage injection.

It can be seen that for the given MMC parameters in Table 6.1, even though 1.15% is used the terminal voltage can be further increased as the margin between the upper and lower limits is still big. Considering the tap changer transformer, the grid voltage is still fixed whilst the MMC terminal voltage is increased. For a fixed operating point of P and Q, the AC current is reduced in the secondary side (converter side) while is fixed at the grid side.

Figure 7.26.b shows that by increasing the output voltage at the MMC terminals by increasing the fundamental component, the peak-to-peak available voltage is reduced due to the reduction of energy deviation as discussed in Chapter 3 and 4. The energy deviation is reduced due to the reduction of the AC current component (i.e. increase of the fundamental component of the synthesised voltage). For a fixed grid voltage, the turns ratio of the transformer is reduced where the new turns ratio should be calculated based on Point D where $V_{c,D} > V_{c,A}$, therefore, $r_{t,D} < r_{t,A}$. Based on the same transformer and limb inductance 18% and 15% given by [77], the MMC parameters due to new the transformer turns ratio are given in Table 7.7.

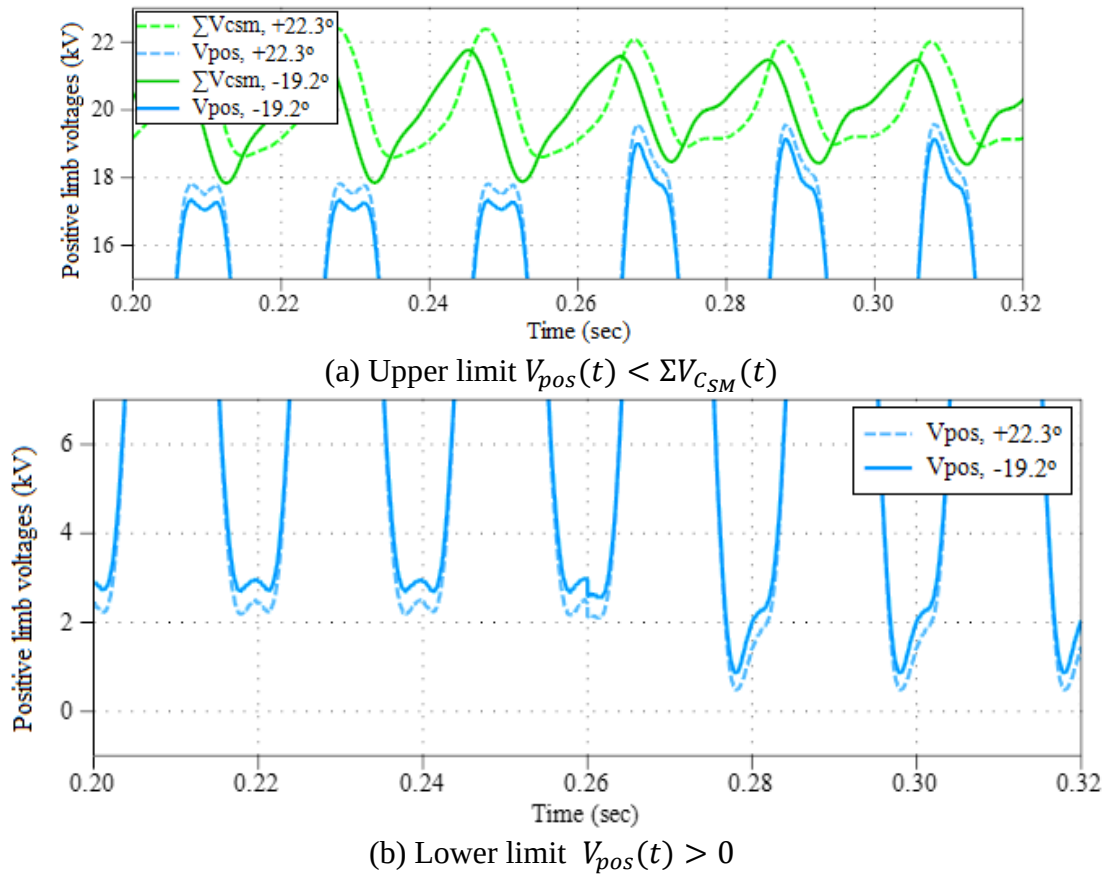


Figure 7.26: Zoomed in the synthesised and available voltages when increasing the fundamental by 1.15% for $S=1.0$ p.u and $\phi = -19.26^\circ$ and 22.3° .

Table 7.7: Circuit parameters MMC due to the new transformer turns ratio design.

Rated output power	P	20MW
Nominal AC system voltage	V_{ac}	11kV
Nominal converter AC voltage	V_c	10.65kV
Converter transformer turn ratio	r_t	0.5183
DC voltage	V_{dc}	20kV
Number of SM	N_{SM}	10
Nominal voltage of SM capacitance	V_{cap}	2000V
SM capacitance	C_{SM}	10.07mF
Inductance of limb inductor	L_{pos}^{**}	2.71mH
Limb resistance	R_{pos}^{**}	0.0 Ω
AC system inductance	L_T	4.29mH
AC system resistance	R_T	3.3m Ω

**The inductance and resistance of positive and negative limb inductors are equal [77].

The terminal voltages and line currents are shown in Figure 7.27. It can be seen that the phase voltage at the MMC terminals is increased while the line current is reduced. Considering the capacitor resizing, the waveforms of the simulation results in Figure 7.28 shows the benefit of using the reduced peak synthesised voltage due to the optimal value of the 3rd harmonic injection (1/6) to redesign the size of the SM capacitor. It is noted that in this case of the operation and design, the grid voltage is kept at the nominal voltage.

The waveforms in Figure 7.28 are based on the operating point D shown in Figure 7.24 and with different size of capacitors; 10.07mF, 6.487mF, 4.785mF and 3.791mF. These capacitor sizes are based on nominal design and using the reduced synthesised voltage to redesign the capacitor size given by Equations (6.62), (6.61) and (6.65) respectively as shown by Figures 6.27.b and 6.29. These are illustrated by solid, dotted, dash-dot and dashed lines respectively.

It can be seen that the size of capacitor can be reduced using the benefit of the phase shifted 3rd harmonic injection as discussed in Chapter 6 and illustrated by Figure 6.28.b. Comparing Figure 7.28 to Figure 6.28.b, it can be seen that the synthesised voltage using the mathematical model is the same for the four operating and design conditions. On the

other hand, the synthesised voltages in the switching model are not identical even though the generated fundamental component of the modulator is the same for the same P and Q operating point. This is related to the fact that the generated modulator signals due to the average and the circulating current controllers in the mathematical model is assumed to be neglected while in the switching model these signals are different for different capacitor size values.

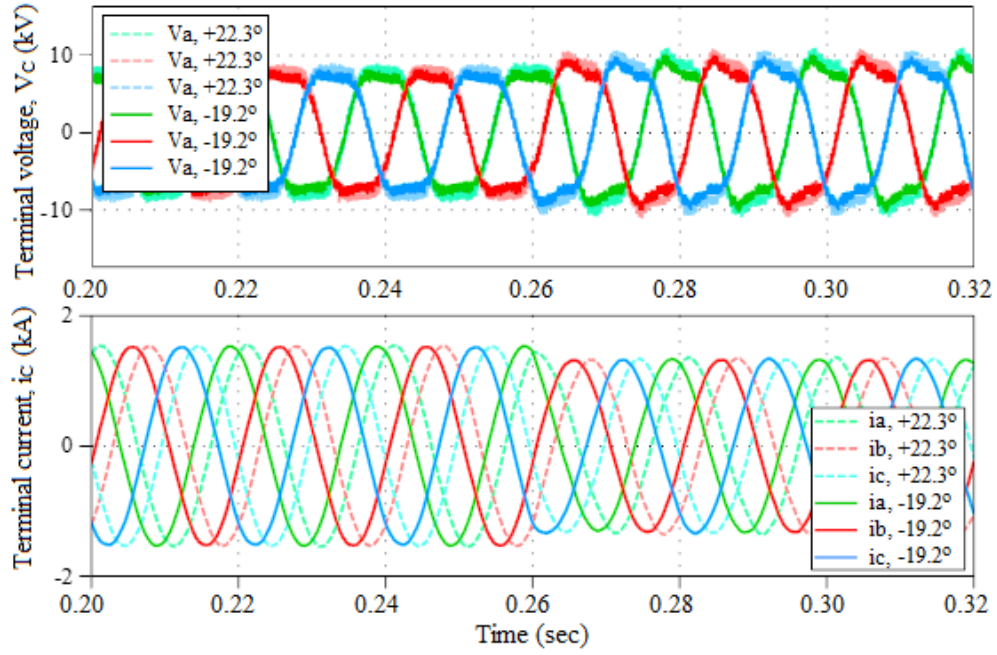


Figure 7.27: Phase and line current at the MMC terminals when increasing the fundamental by 1.15% for $S=1.0$ p.u and $\phi = -19.26^\circ$ and 22.3° .

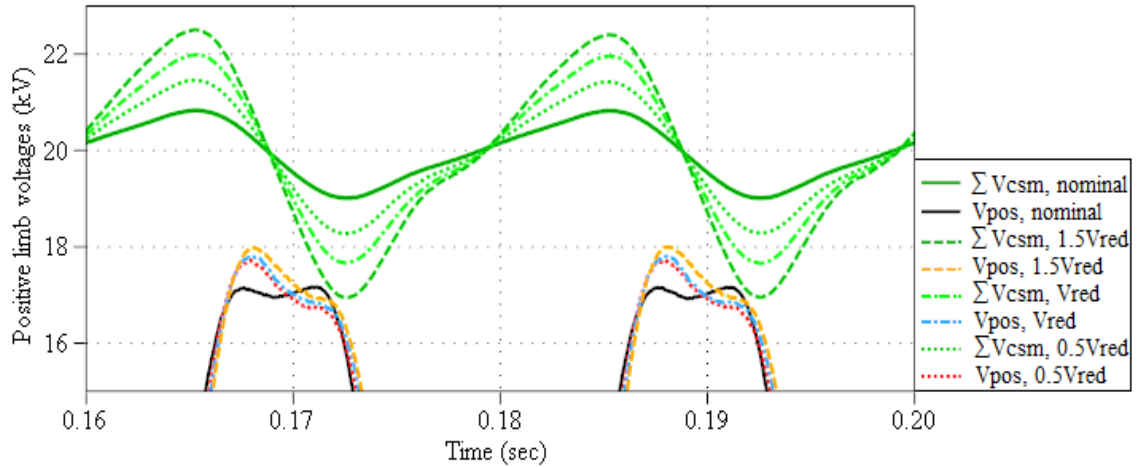


Figure 7.28: Positive limb synthesised and available voltages for different size of capacitors for 20MW, $S=0.94-j0.33$ and $\theta_{v3} = -20^\circ$ and ($\theta_{v3} = 0^\circ$ at nominal design).

The capacitor size influences the control signals. In fact, the DC controller including the circulating current controller is intended to keep the average of the limb voltage at the desired level, $V_C^* = V_{dc}$. Therefore, for larger capacitor size, DC offset in the synthesised voltage, $\bar{V}_C^*$ is smaller which increases by using smaller capacitor sizes. On the other hand, the smaller the capacitor size is, the less stored energy which requires an increase in the DC offset in the synthesised voltage in order to maintain the average at the desired level for the same reference of the average controller. However, the fundamental component of the synthesised voltage will not be affected since it is controlled by the AC controller. In our case, the DC controller is used for the four cases with the same reference of the average controller $V_C^* = V_{dc}$.

Figure 7.29 illustrates the operation with different average voltage references, nominal and increased by 0.5%, and for the same capacitor size due to $1.5V_{red}$. The figure explains the impact of the DC controller not only on the synthesised voltage but also on the available voltage. The synthesised voltage is reduced, from dashed orange to solid blue where the synthesised voltage with legend of $V_{pos, 1.5V_{red}}$ in Figure 7.28 and the synthesised voltage with legend of $V_{pos, V_C^* \text{ nominal}}$ in Figure 7.29 are the same. It is replotted here for comparison purposes only. The average of the available voltage, however, is increased by increasing the capacitor voltage of each SM.

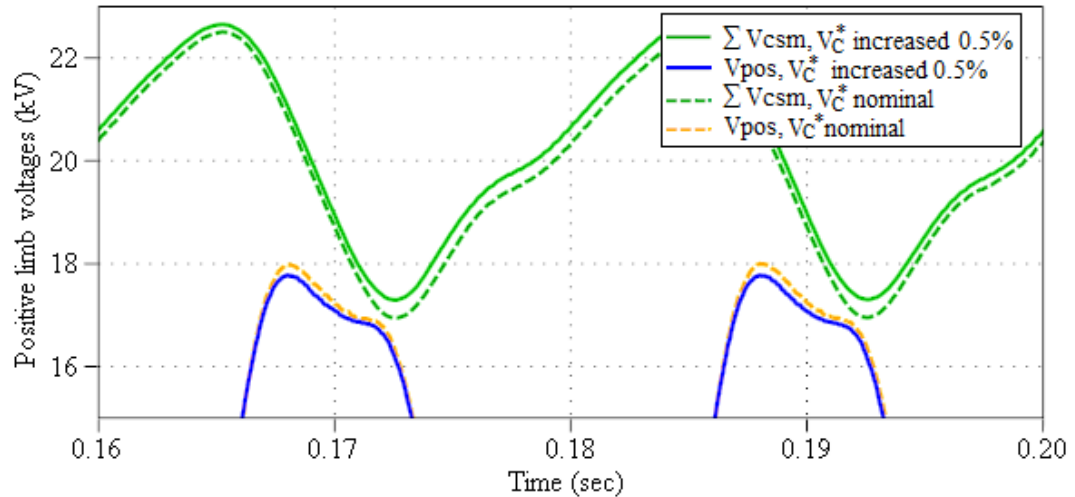


Figure 7.29: Comparison of synthesised and available voltages for different average voltage references, nominal (V_{dc}) and increased by 0.5%.

This explain the reason why the peak of the synthesised voltages are different where the maximum occurs with the smallest capacitor size, a with the legend of $V_{pos, 1.5V_{red}}$. The four sizes of capacitor are used in Figure 7.28 are 10.07mF, 6.487mF, 4.785mF and

3.791mF based on nominal design, i.e. 10.07mF with 0.1 voltage ripple, and using the reduced synthesised voltage to redesign the capacitor size given by Equations (6.64), (6.63) and (6.67) respectively as shown by Figure 6.28.a and 6.30.a.

7.9.2 Simulation results for 11kW MMC

The simulation results of the switching model below are obtained based on the 11kW MMC parameters given in Table 7.1. The waveforms of the positive limb available voltage and synthesised voltages of the switching model shown in Figure 7.30 illustrates the idea of increasing the voltage at the MMC terminal using the advantage of phase-shifted 3rd harmonic voltage injection.

The waveforms represent operation with the nominal voltage and rated power while applying classical 3rd harmonic voltage injection when ($t < 0.26$ s). However, when ($t \geq 0.26$ s), the MMC is forced to increase its synthesised voltage (output voltage) using the advantage of the phase-shifted 3rd harmonic voltage injection. The solid and dashed lines represent the waveforms of the operating points of A and D shown by Figure 7.24 where the operating conditions for each point is; point A, $Q=-0.33$ p.u ($\phi = -19.2^\circ$) and for point D, $Q=+0.41$ p.u ($\phi = +22.3^\circ$) whilst $P=+0.94$ p.u in both cases.

Compared to the 20MW MMC simulation results, the AC voltage at the MMC terminals is only increased by 9% whilst 20MW is increased 15%. This is related to the difference of the initial design of the both circuits. At no load, the modulation index of the 20MW MMC is 0.87 whilst in the 11kW version it is 0.95 since the line impedance of the 20MW MMC is 18% as listed in Table 6.1[77] whilst in the 11kW version it is about 13.5% (based on base values of 11kW and 477V). Therefore, the voltage drop in the second case is smaller as shown by Figure 7.30.

Figure 7.31.a and Figure 7.31.b, are Figure 7.30 zoomed in. considering the operating point A, the margin between the synthesised and the available voltage is still big even by increasing the fundamental by 9%. This allows further increase of the synthesised voltage. Thus, the terminal voltage can be increased since the fundamental component is increased. However, by considering the opposite operating point, D when $P=0.94$ and $Q=+0.41$, the lower limit prevents increase the voltage above 9% as illustrated by the dashed blue. Therefore, the maximum increase of the terminal voltage is limited by the D operating point up to 9%.

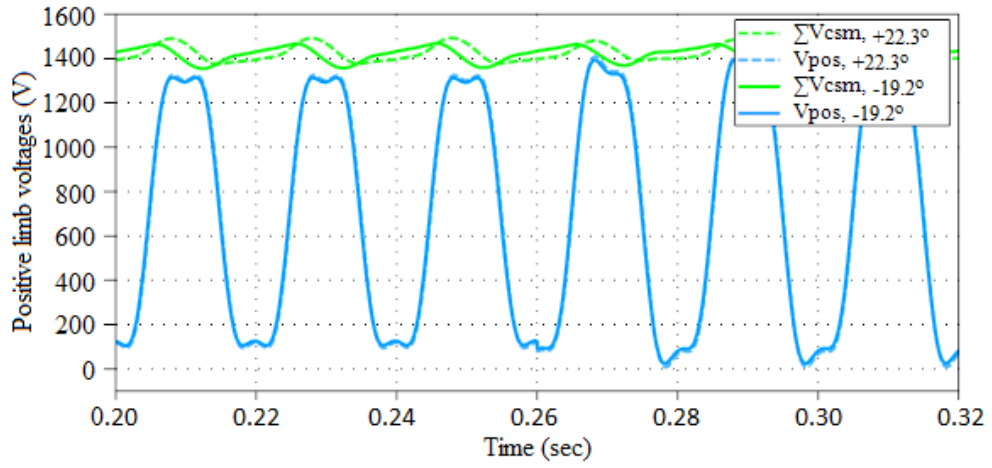
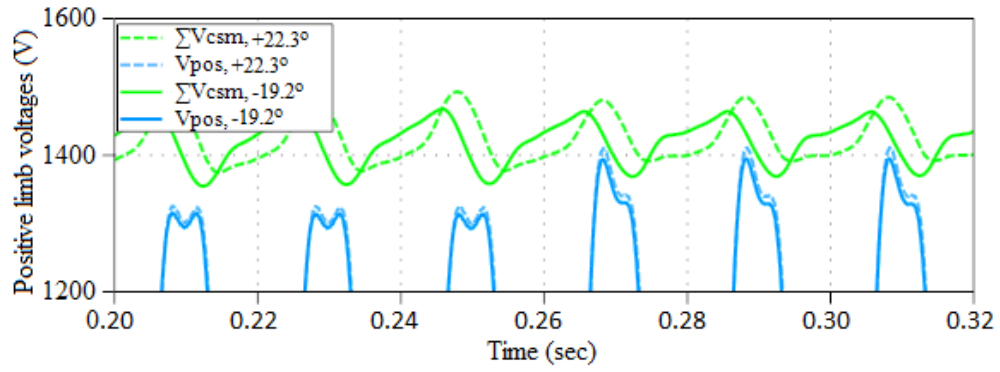
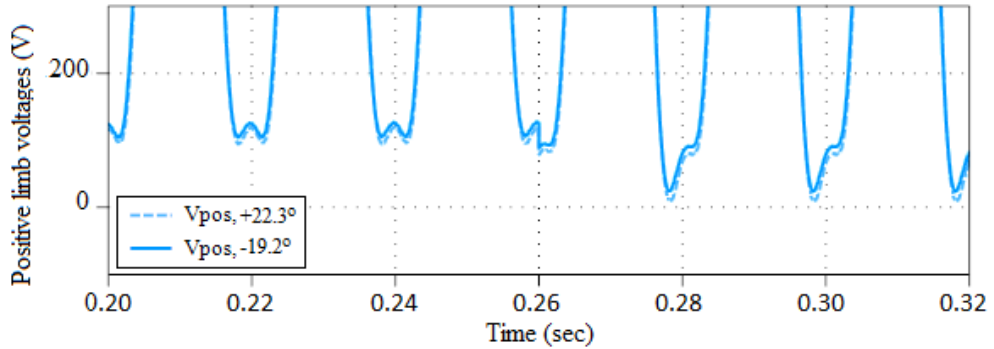


Figure 7.30: The synthesised and available voltages when increasing the fundamental by 1.09% for $S=1.0$ p.u and $\phi = -19.26^\circ$ (point A) and $+22.3^\circ$ (point D).



(b) Upper limit $V_{pos}(t) < \Sigma V_{C_{SM}}(t)$



(c) Lower limit $V_{pos}(t) > 0$

Figure 7.31: Phase voltage and line current at the MMC terminals when increasing the fundamental by 1.09% for $S=1.0$ p.u and $\phi = -19.26^\circ$ (point A) and 22.3° (point D).

With regards to resizing the SM capacitor of the MMC given in Table 7.1, the waveforms of the simulation results shown by Figure 7.32 are obtained so that grid voltage and the terminal voltage of the MMC are kept at the nominal values, 170.5V and 477V as listed in Table 7.1. The SM capacitor size, however, is redesigned. It can be seen that the maximum capacitor size reduction is approximately 20% since the new

size that can be used is 0.864 mF instead of 1.08mF. The results are obtained for the operating point of A. It can be seen that correct operation is achieved by increasing the available voltage ripple due to reduction of the capacitor size by 20%. Thus, the rated stored energy is reduced by approximately 20%.

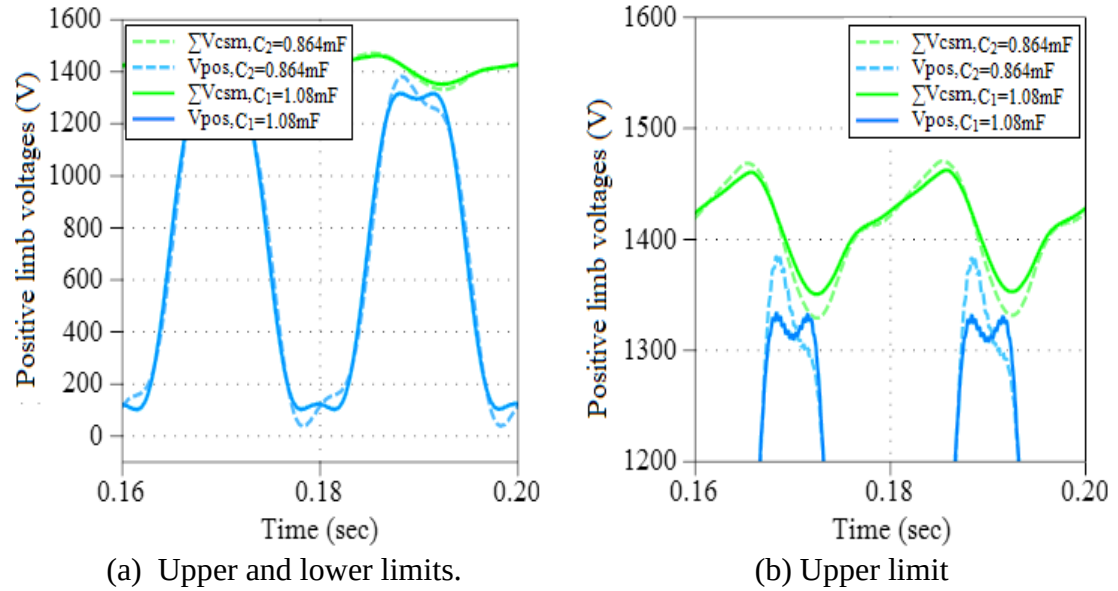


Figure 7.32: Positive limb synthesised and available voltages for different size of capacitors for 11kW, $S=0.94-j0.33$ and $\theta_{v3} = -10^\circ$ and ($\theta_{v3} = 0^\circ$ at nominal design).

7.9 Conclusion

The analysis presented by the previous chapters was achieved for a three phase modular multilevel converter (MMC) for an HVDC system applications with high power and high voltage. A mathematical model was presented based on some assumptions as listed in Chapter 3. Chapter 7 presented the control scheme and modelling the MMC to validate the mathematical model. Therefore, this analysis was validated using a switching model of a 20MW MMC. Another switched model for an 11kW MMC was presented which will be validated by the experimental work of the prototype presented in Chapter 8.

The control and modelling of the MMC were presented in this chapter. The control was first presented including AC and DC controls. The MMC modelling was then presented. Considering the control objectives, an overall control scheme was proposed considering the AC and DC voltage controllers. Based on the deduced transfer functions, the control scheme and design were detailed including AC current, limb DC average voltage and circulating current controllers. Taking into consideration the phase-shift of the 3rd

harmonic, a lookup table was used to achieve the required phase shift of 3rd harmonic component.

MMC modelling was discussed including an averaged MMC model and a detailed switching model. Simulation results were provided to verify the control approach and the modelling. The MMC models were used to verify the proposed methods in Chapter 6 to achieve an improved MMC operation and design. The experimental results will be presented in Chapter 8.

Chapter 8

8 Experimental Results

8.1 Introduction

The control and modelling of a three phase MMC were presented in previous chapter. Simulation results based on switching models of 20 MW MMC and 11kW MMC were presented. The focus of the simulation results was on validating the mathematical model and analysis of the two proposed methods presented in Chapter 6 in order to benefit from the reduced synthesised voltage due to optimal 3rd harmonic injection.

The two different methods presented in Chapter 6, which can be used to improve the operation and design of the MMC, were validated. It has been shown that using the 3rd harmonic voltage injection can be used to reduce the peak synthesised voltage which can be exploited in two different ways as discussed in Chapter 6:

- Increasing the output voltage at the MMC terminals. Or,
- Reducing the size of the SM capacitor by allowing higher voltage ripple.

The first approach requires a dedicated design of the conversion system, and its applicability is therefore limited to new installations where the effected component is the converter transformer. However, this can be validated easily either in the lab or by the switching model when considering variable AC source, e.g. a programmable AC supply. On the other hand, the second option considers a given design of the MMC and only revisits the size of the SM capacitors by exploiting the increased available voltage margin after 3rd harmonic injection. This can be easily validated using the switching model. However, in the lab this would require installation of the new capacitors based on the new improved size.

It has been shown that the terminal voltage increase by about 1.15 and 1.09 of the nominal for the MMC based on 20MW and 11kW respectively. The operating point of the MMC of the active and the reactive power P and Q are based on the power convention presented in Appendix A, $P=+0.94$ and $Q=+0.41$.

As shown in Figure 7.32.b, it is noted that increasing of the AC voltage at the MMC terminals is limited by the peak of the synthesised voltage. In fact, this limitation is presented by the lower limit ($V_{pos}(t) > 0$) when operating in the first quadrant (Q1) with rated S ($P > 0$ and $Q > 0$) shown in Figure 6.14 as represented by point D in Figure 7.24. The power convention implemented in this chapter was discussed before and presented in the Appendix A. Therefore, this operating point considers the worst case scenario, if the increase of the terminal voltage is mentioned. However, if the size of the capacitor is considered, then the worst case of the operating points of the MMC is when the MMC operates with $Q < 0$ as shown by point A in Figure 7.24, for instance. At this point, it has been seen that the proposed capacitor size reduction analysis can be implemented where the phase-shifted 3rd harmonic injection is required as a key factor to maximise the margin between the synthesised and available voltage. Therefore, the capacitor size reduction can be maximised.

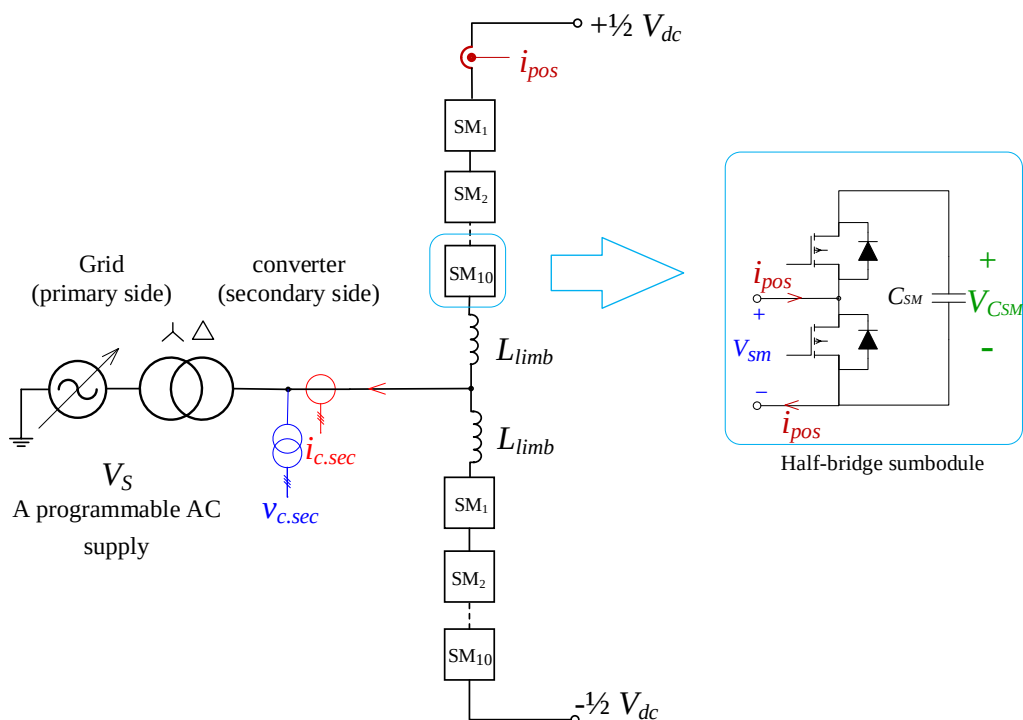


Figure 8.1: Single line diagram of the experimental rig with MMC mode.

In order to experimentally test the proposed method for increasing the terminal voltage, as presented in Chapter 6, and to validate the modelling and control scheme presented in Chapter 7, an existing MMC experimental rig has been used. As mentioned in Chapter 7, the primary mode of the experimental rig is AAC converter. However, it is built so that it can be re-configured as an MMC converter. Therefore, the experiments were

carried out on the hardware rig detailed in [86] which was re-configured as an MMC for the purposes of these tests. The single line diagram of the MMC mode is shown in Figure 8.1. The parameters of the experimental MMC rig after the re-configuration are listed in Table 7.1. However, for the more convenience it is relisted in Table 8.1. The second proposed method requires the installation of sixty capacitors at 0.864mF instead of 1.08mF. Unfortunately, due to time constraints, validation of the second method is complex and considered out of scope of the thesis and will be considered in future work.



Figure 8.2: The small scale prototype hardware: SM and IGBT module interface PCBs (left), and backplane PCBs (right)[86].

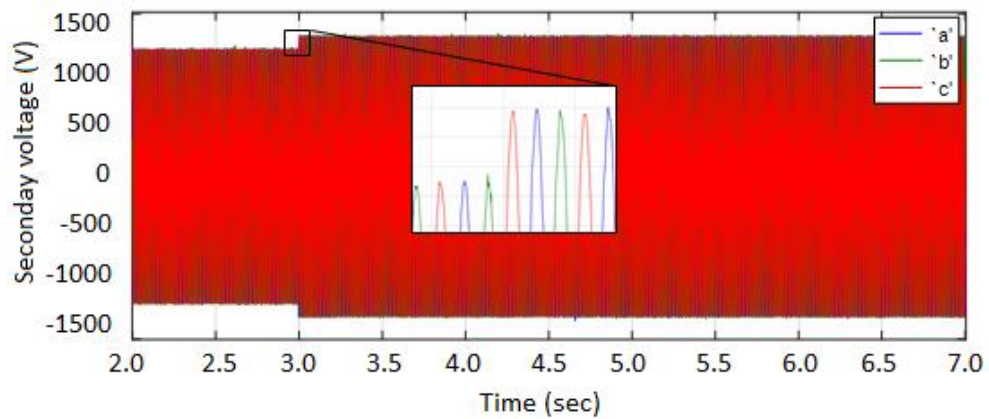
Table 8.1: Circuit parameters used in the simulation and the experimental results.

Rated output power	P	11.0 kW
Nominal AC system voltage	V_{ac}	170.948 V
Nominal converter AC voltage	V_c	476.944 V
Nominal converter transformer turn ratio	r_t	0.208
DC voltage	V_{DC}	1420 V
Number of SM	N_{SM}	10
Nominal voltage of SM capacitance	V_{cap}	142 V
SM capacitance	C_{SM}	1.08 mF
Inductance of limb inductor	L_{pos}	6.7 mH
Limb resistance	R_{pos}	0.51 Ω
AC system inductance	L_t	9 mH
AC system resistance	R_t	2.6 Ω

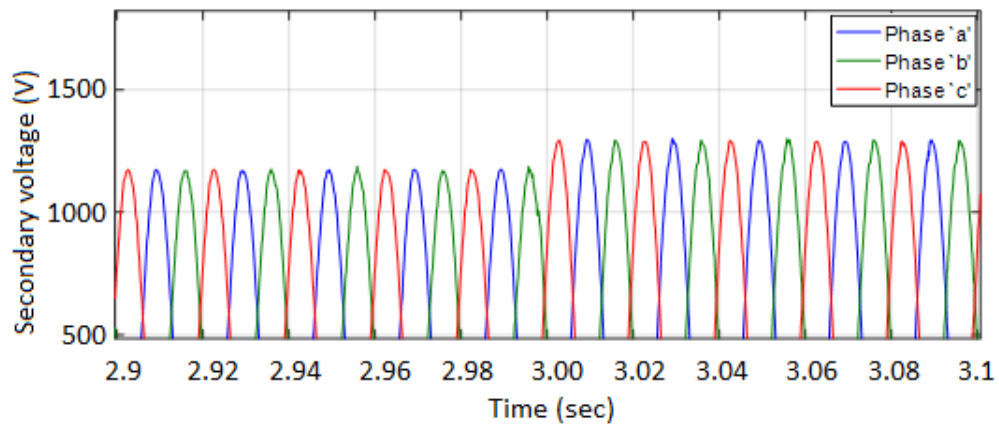
8.2 Experimental waveforms

Experimental waveforms from the existing MMC rig are presented in this section. The experiment is carried out with the same conditions as the simulation. The experimental results are then compared to the simulation results. The operation point of P and Q was at the first quadrant of the P and Q envelope at the point of the maximum synthesised voltage of the MMC.

As shown in Figure 8.1, a controllable AC source is used instead of a tap changer transformer to force the MMC to increase the synthesised voltage by increasing grid voltage, V_{grid} thanks to the phase-shifted 3rd harmonic voltage injection. The controller will force the MMC increasing its synthesised voltage which will be allowed using the lookup table. As a result, the output voltage at the MMC terminals (secondary side of the transformer) will increase from 1.0p.u to 1.09p.u. The AC currents at the secondary side of the transformer will be reduced for the same operating point of P and Q.



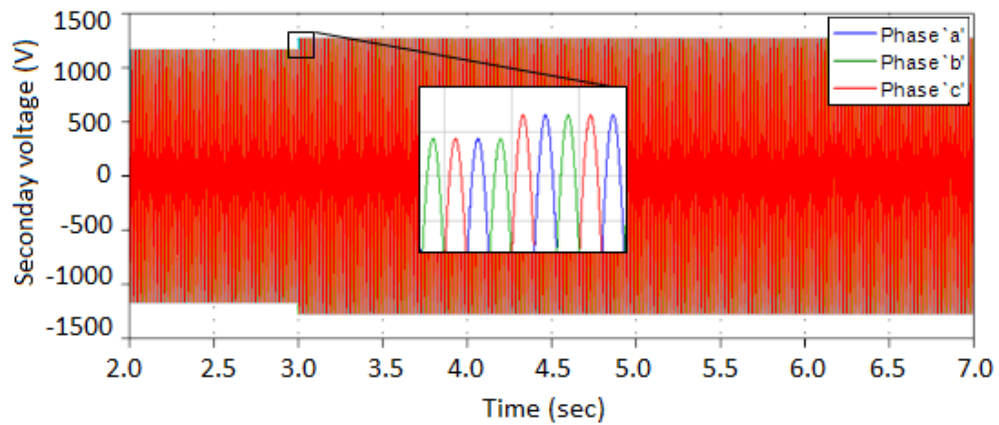
(a) Total view.



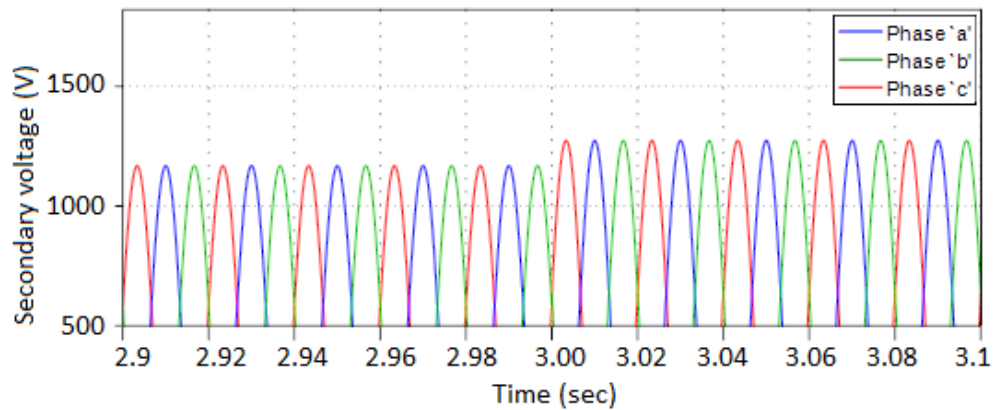
(b) Zoomed view.

Figure 8.3: The experimental results of line voltages at the secondary windings of the converter transformer; (a) total view, (b) zoomed view.

It is noted that the converter is controlled to inject P^* and Q^* of +0.94p.u and +0.41p.u, respectively no matter what happened to the grid voltage within the lookup table margin. It is also noted that the operation will be tested first with conventional 3rd harmonic and then with phase-shifted 3rd harmonic. The grid voltage is increased at ($t = 3.0$ s) in order to force the converter to use phase-shifted 3rd harmonic voltage injection using a programmable AC source. Figures below show the experimental waveforms of the line current and output voltage of the MMC at the secondary winding of the converter transformer as shown by Figure 8.1.



(a) Total view.



(b) Zoomed view.

Figure 8.4: The simulation results of line voltages at the secondary windings of the converter transformer; (a) total view, (b) zoomed view.

Figure 8.3 and Figure 8.4 show the experimental and simulation results of the line voltages at the secondary side of the transformer as indicated by the voltage ($v_{c.sec}$) as shown in Figure 8.1. For more convenient, detailed and zoomed views are shown which presented by (a) total and (b) zoomed views for both experimental and simulation results. Comparing between the experimental and the simulation results of the secondary

line voltages, it can be seen that the MMC has the ability to increase the output voltage above 1.0 p.u ($t < 3.0s$) to 1.09 at ($t \geq 3.0s$) which is the same result achieved by the simulation model. This increase leads to a reduction of the line current by the same percentage as shown by the step response in Figure 8.5 and Figure 8.6.

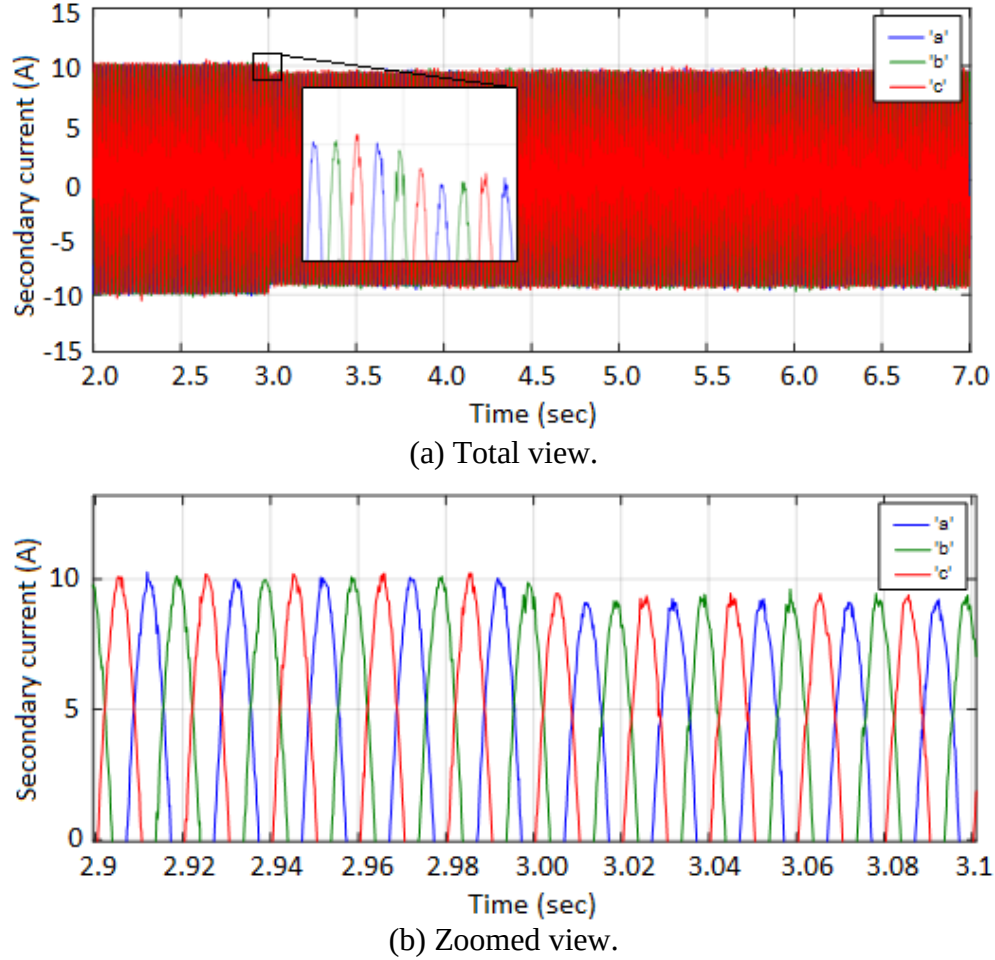


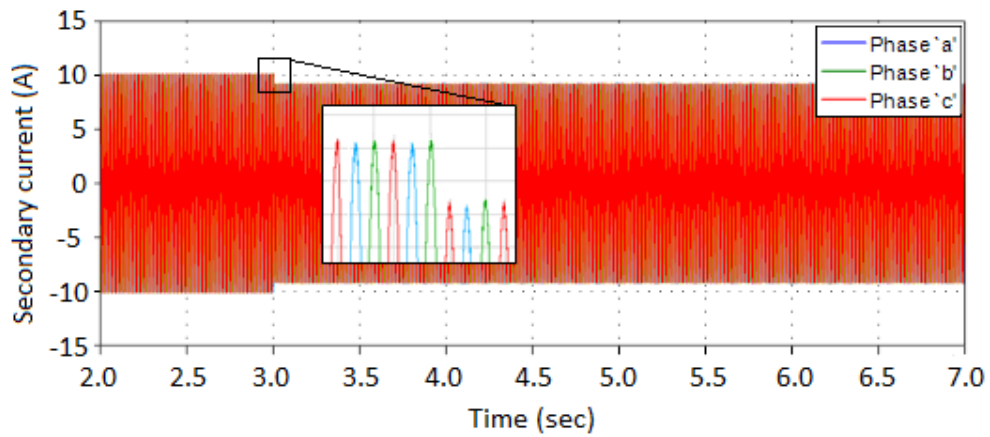
Figure 8.5: The experimental results of line currents; (a) total view, (b) zoomed view.

Figure 8.5 and Figure 8.6 show the line currents measured at the secondary side of the converter transformer using the current transformer represented by ($i_{c.sec}$) as shown in Figure 8.1. It can be seen that by increasing the primary side voltage (grid) from 1.0 p.u to 1.09 p.u, the AC current is reduced for the same operating point of P^* and Q^* (+0.94 p.u and +0.41 p.u), respectively. This is the same response as obtained by the simulation results as seen in Figure 8.6. It is noted that the secondary line currents are reduced due to the increase of the output voltage at the MMC terminals.

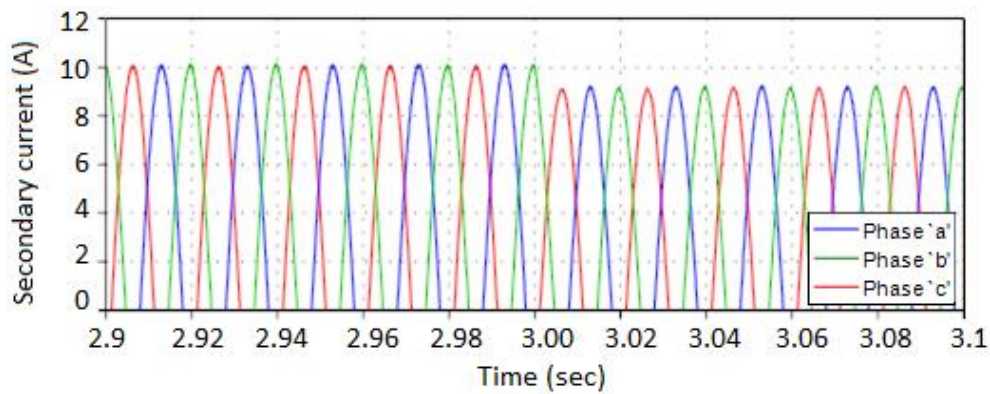
It is noted that, the line current of the experimental results seems to decay with more time constant compared to the simulation results as shown by Figure 8.5 and Figure 8.6

respectively. This can be related to that the controller of switched model is faster than prototype rig since the switched mode was also consider to validate the mathematical models and the simulations results based on the mathematical model covered from Chapter 3 to Chapter 7. Therefore, the bandwidth of the switched model AC controller was chosen to achieve an accuracy so that it can compromise between the average model presented by the mathematical model and the prototype rig.

However, it can be say that the comparison shows that experimental and simulation results are the similar. This means that both the control scheme and also the method presented in Chapter 6 of increasing the output voltage and reducing the AC current can be validated.



(a) Total view.



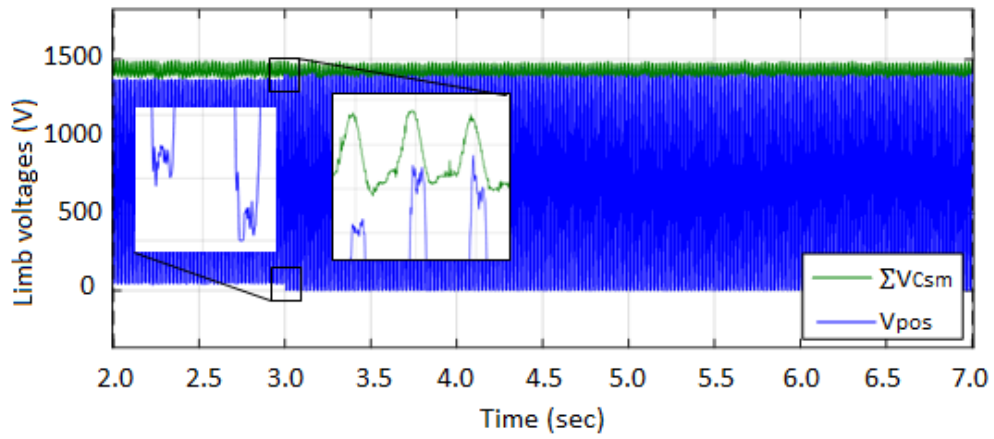
(b) Zoomed view.

Figure 8.6: The simulation results of line currents; (a) total view, (b) zoomed view.

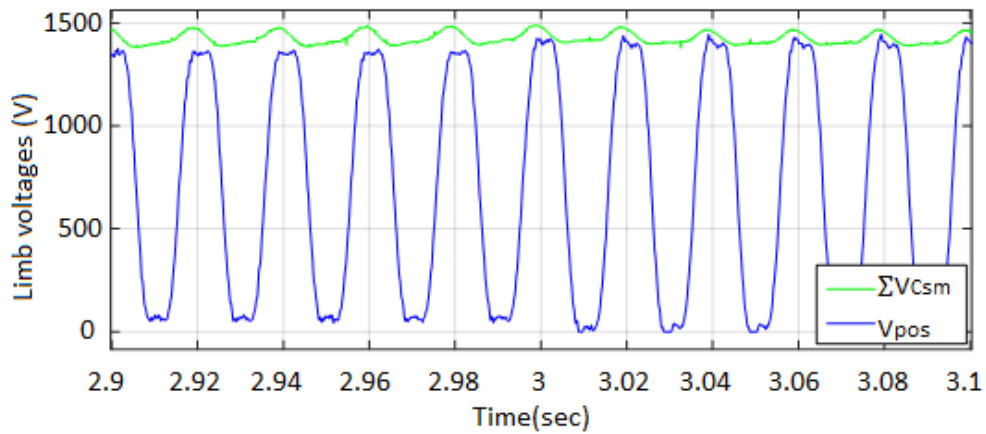
With regards to the available and the synthesised voltages of the limbs, the comparison of the experimental and simulation results are shown by Figure 8.7 and Figure 8.8 respectively where the positive limb voltages are considered.

It can be seen that the experimental waveforms are the same as the simulation waveforms. The peak of the synthesised voltage is increased due to the increase of the fundamental component which can be seen by considering the increase of the output voltage at the MMC terminals shown in Figure 8.3 and Figure 8.4.

It is noted that to increase the output voltage at the MMC terminals the synthesised voltage needs to be increased as discussed in the previous chapters. This can be seen by comparing the experimental and simulation results presented by Figure 8.7 and Figure 8.8. It can be seen that due to the increase of the grid voltage at ($t=3.0$ s) the MMC is forced to increase the synthesised voltage by increasing its fundamental component from 1.0 p.u to 1.09 p.u.



(a) Total view.



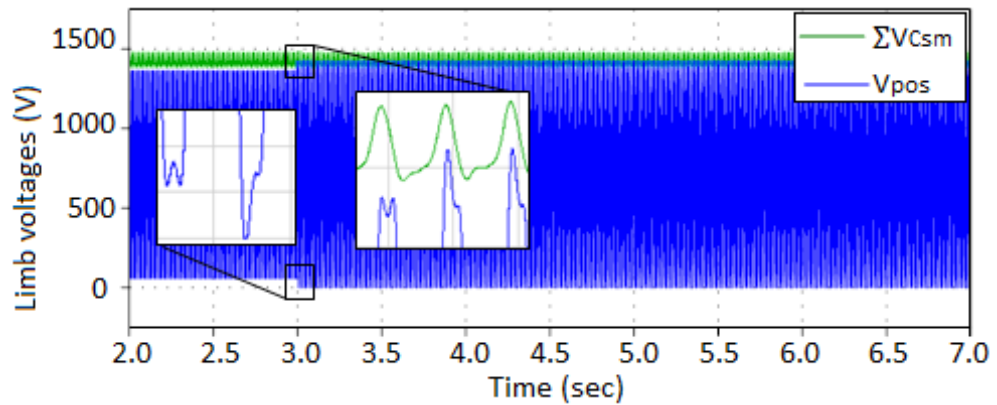
(b) Zoomed view.

Figure 8.7: The experimental results of positive limb synthesised and available voltages; (a) total view, (b) zoomed view.

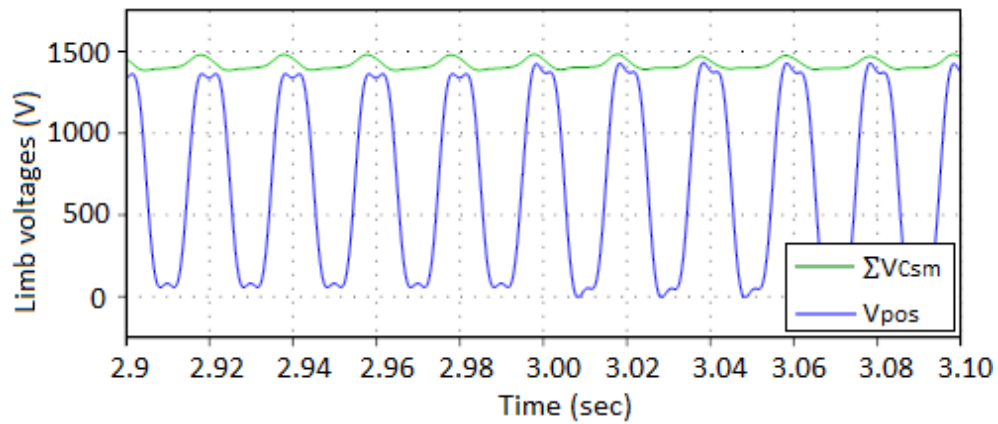
From Figure 8.7 and Figure 8.8 it can also be seen that using 3rd harmonic injection increase the margin between the available and synthesised voltage. As the fundamental

of the synthesised voltage is increased from 1.0 p.u to 1.09 p.u, the output voltage at the MMC terminals increases and therefore the line currents are reduced. The detailed simulation results given by Figure 8.9 give a better view.

It can be seen that the 3rd harmonic voltage component is forced to be shifted at ($t \geq 3.0$ s) allowing the maximum increase of the synthesised voltage compared to the previous operation, ($t < 3.0$ s) of classical 3rd harmonic voltage injection. Considering the comparison, the experimental and simulation results are the same.



(a) Total view.



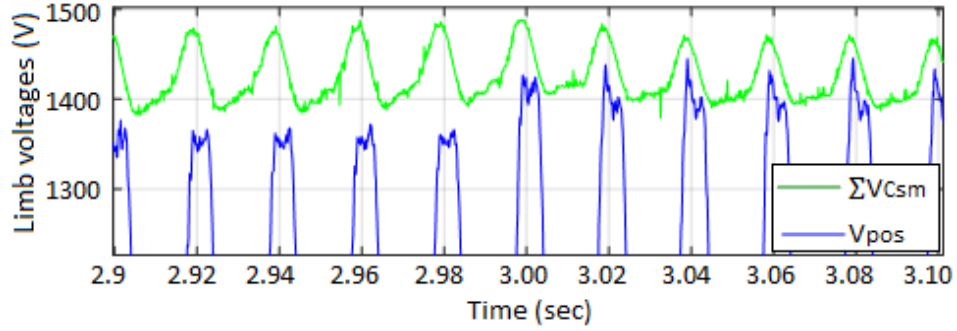
(b) Zoomed view.

Figure 8.8: The simulation results of positive limb synthesised and available voltages;

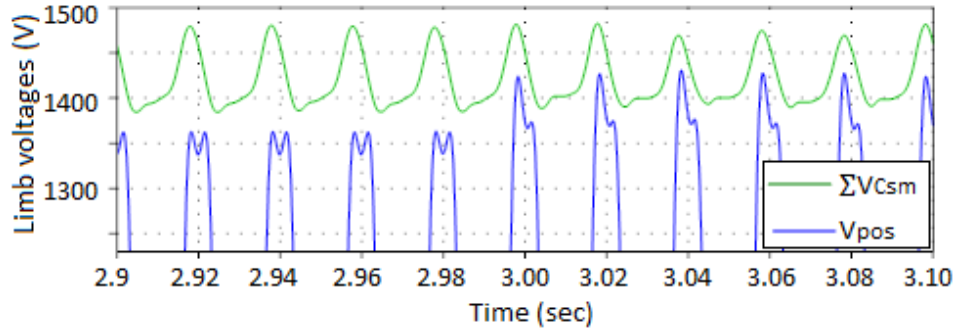
(a) total view, (b) zoomed view.

Considering the available voltage shown indicated by the green trace in Figure 8.9, it can be seen that as discussed in Chapters 4 and 6, increasing the fundamental of the synthesised voltage (modulation index) can lead to a reduction of the peak-to-peak of the available voltage since the reduction of the energy deviation reduced. This can be seen by Figure 3.8 and Figure 4.22 where the energy deviation is reduced by increasing the modulation index (i.e. output voltage) This reduction can be attributed to the reduction of the AC component of the limb's since the AC voltage increased.

Figure 8.10 shows a comparison between the experimental and simulation results of positive limb current of one phase. From the comparison it can be seen that the waveforms of the limb currents are the same for both the experimental and simulation results. The peak currents are reduced due to the reduction of the line current and increasing the output voltage for the same P and Q operating point, $S=+0.94+j0.41$ p.u.



(a) Experimental results.



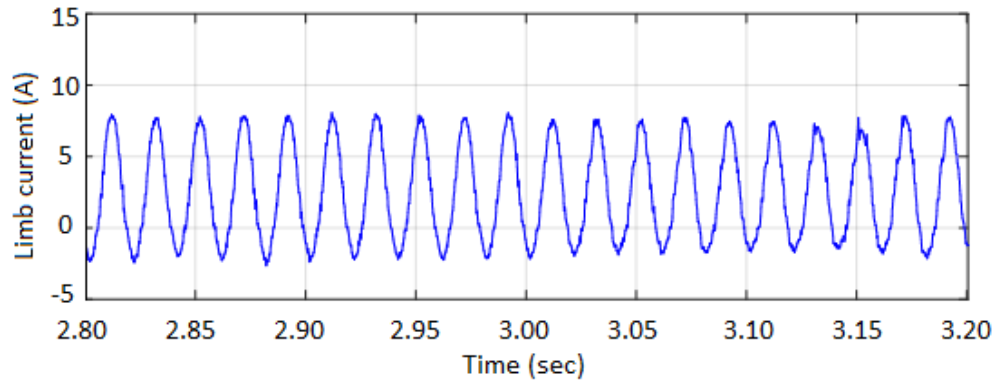
(b) Simulation results.

Figure 8.9: The increased peak of the synthesised voltage with the reduction of the peak-to-peak of the available voltage.

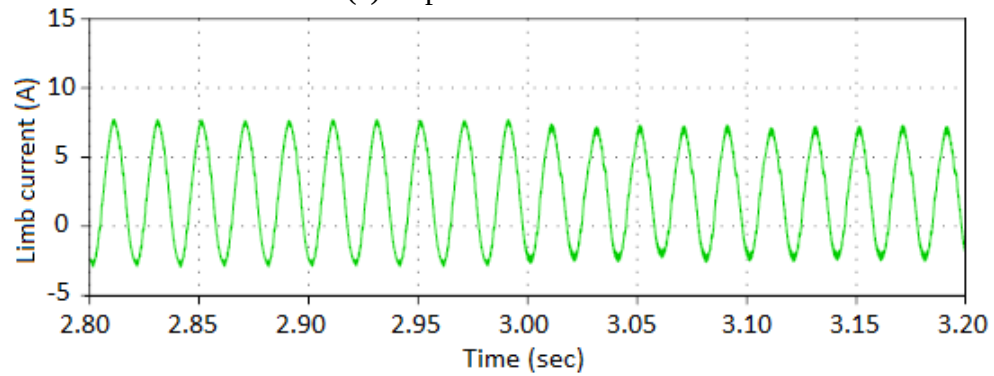
Increasing the operating voltage and for the same rated power leads to reduce the line current with same amount since $I_C = \frac{S(\text{constant})}{\sqrt{3}V_C}$. As discussed in Chapter 3 as presented by Equation (3.10), the current flows in each limb has two components, DC and AC currents. The limb DC current is one third of the converter DC current, $1/3 I_{dc}$, whilst its AC current component is half the converter AC current, $1/2 I_C$. In fact, each limb provides half the power demand. As it can be seen, the reduction of the limb current cannot be seen easily. This can be explained by the fact that the line current, I_C , is reduced approximately 9% due to the increase of the terminal voltage by the same amount, i.e. 9%, from 1.0 p.u to 1.09 p.u. As presented by Equation (3.10) and by considering Figure 8.5 and Figure 8.6, the line current is reduced from 10.2 A to 9.4 A (peak). Therefore, the AC current component of the limb is reduced from 5.1 to 4.6 A (peak) while the DC current has not been affected. Therefore, it can be only seen a slight

reduction, i.e. it reduced $5.1 - 4.6 = 0.5\text{A}$ (peak). For this reason, the reduced limb current cannot be seen clearly from Figure 8.10.

From previous comparisons, it can be concluded that the first proposed method presented in Chapter 6, with regards to the increased the output voltage of the MMC terminals using the advantage of the phase-shifter 3rd harmonic, can be achieved using the advantage of phase-shifted 3rd harmonic voltage injection. In practical application, with the proper transformer turns ratio, the AC voltage at the MMC terminals can be increased and therefore reducing the AC current at the secondary including the MMC legs. The grid voltage, however, is fixed with the permissible tolerance defined by the utilities for the same P and Q envelope. This can lead to a reduction of the power loss and therefore the cooling system requirements and cost.



(a) Experimental results.



(b) Simulation results.

Figure 8.10: Comparison of the experimental and simulation results of the positive limb current.

8.3 Conclusion

Simulation results and the experimental results obtained from an existing experimental rig in the laboratory of the MMC has been presented in this chapter for an MMC with nominal output power of 11kW. The focus was on validating the first method of

improving the operation and design of the MMC as presented in Chapter 6, with regards to increasing the output voltage and reducing the AC current. This was achieved by the proposed method of applying the phase-shifted 3rd harmonic voltage injection. A comparison between the experimental and simulation results has been presented validating the first presented method. The comparison has also validated the modelling and control scheme presented in Chapter 7.

According to the experimental results, phased-shifted 3rd harmonic voltage injection can be successfully used to increase the output voltage of the MMC voltage which leads to a reduction of the line currents and therefore the loss. Due to the increase of the fundamental component of the synthesised voltage (reducing the AC current), the peak-to-peak available voltage is reduced due to the reduction of the energy deviation as discussed in Chapter 4.

Chapter 9

9 Conclusion and Future Work

9.1 Conclusion

The demand for higher voltage and power ratings and better power quality for HVDC transmission systems drives the intense research on modular multilevel converters. Due to the higher power ratings, MMC are subject to an increase in their size and cost due to complex structures and the large quantity of components of the submodules. Therefore, since the MMC has been introduced, research on improving and optimising design, control and operation of the converter has been carried out in the work reported in this thesis. One of the options to optimise design and operation of the MMC is by using harmonic injection including voltage and current harmonics.

This thesis has first investigated, with a generalised analysis, the impact of 3rd harmonic voltage and current and 2nd harmonic current on the operation and design of the MMC. For HVDC applications such as off-shore power transmission, classical and phase shifted 3rd harmonic voltage injections were identified as the most promising solutions, and in particular the phase shifted 3rd harmonic voltage injection has been explored in detail, providing as a result two new methods to achieve an improved operation and design of the MMC, with a focus on reducing the peak of the synthesised voltage.

The thesis outcome can be summarised as follow;

- Firstly, the work filled the gap of the previous studies by presenting an extensive study considering the impact of harmonic injection on energy storage requirements including comparison between different harmonic injections, i.e. 3rd harmonic voltage, 3rd harmonic current and 2nd harmonic current injections.
- Secondly, the impact of harmonic injection, in particular 3rd harmonic voltage injection on the MMC operation and design including the operating voltage and capacitor sizing was studied and identified. The peak of the synthesised voltage is reduced due to the 3rd harmonic voltage injection. In this part, the thesis

proposed two different methods to improve the MMC operation and design to utilise the reduced peak synthesised voltage.

With regards to the first point, the achievements can be listed as follow;

- For the first time, the impact of 3rd harmonic current injection for the MMC is investigated with and without 3rd harmonic voltage injection. The result of the analysis indicated some applications that can benefit from applying 3rd harmonic voltage and/or current injection.
- Proposed the phase shift of the 3rd harmonic injection concept, positive and negative phase shift.
- For the first time a comparison between 3rd harmonic voltage injection and 2nd harmonic current injection on the operation of the MMC was presented. In particular, the energy storage requirements and the rated voltage were studied. This thesis gives further insights by considering both 3rd harmonic voltage and 2nd harmonic current injection.

In the second point, two new different methods using the reduced peak synthesised voltage and using phase-shifted 3rd harmonic voltage injection were proposed for improving the operation or design of the MMC. These methods utilised the advantage of optimal 3rd harmonic voltage injection and the proposed phase-shift to reducing the peak of the synthesised voltage.

- Limits and other aspects due to using phase-shifted 3rd harmonic injection were discussed and lookup tables were implemented giving the optimal classical phase shift based on the reported the optimal reduction of the peak of the synthesised voltage.
- Utilised the optimal solution of the classical 3rd harmonic injection, for the first time, limits of classical and phase-shifted 3rd harmonic voltage injection, positive and negative phase shift, were discussed.
- Discussed the limits and other aspects due to using phase-shifted 3rd harmonic injection on the MMC operation including the synthesised voltage.
- Proposed lookup table method for applying the proposed phase shift in order to develop a method of increasing the output voltage at the MMC terminals.

- Highlighted other aspects including loss reduction and converter transformer design due to using phase-shifted 3rd harmonic injection.
- And finally, proposed two methods, based on using the phase-shifted 3rd harmonic voltage injection and due to the reduced peak synthesised voltage including, developing a method of increasing the output voltage at the MMC terminals and the capacitor sizing. The two methods proposed by the thesis can be summarised as follow;
 - i. Further increasing of the voltage at MMC terminals could be achieved by applying the phase-shifted 3rd harmonic since the fundamental component of the synthesised voltage increased. Power loss can be reduced since line currents are reduced for the same capability power with proper scaling of the transformer turns ratio.
 - ii. Higher voltage ripple of the available voltage can be allowed which can be used to reduce the size of the SM capacitor.

To achieve that, a literature review on the existing converters for HVDC was first covered by Chapter 2. The operating principle of the MMC was presented in Chapter 3. A generalised analysis of the impact of harmonic injection on stored energy requirements was presented in Chapter 4 including 3rd harmonic voltage, current and 2nd harmonic current injection. For the first time, the phase-shifted 3rd harmonic voltage injection was presented in Chapter 4. Moreover, it has been also shown that 3rd harmonic voltage injection with 2nd harmonic current injection has the better operation. The 2nd harmonic currents reduces capacitor voltage ripples whilst the 3rd harmonic reduces the peak of the synthesised voltage. Further analysis to investigate the impact of 3rd harmonic voltage injection on stored energy requirements including the SM capacitor size was presented in Chapter 5. Using the reduced peak synthesised voltage due to 3rd harmonic voltage injection, the focus of Chapter 6 was on presenting two possible and different methods to improve the operation and design of the MMC using the advantages of the phase shifted 3rd harmonic as presented in Chapter 4.

In order to validate the mathematical analysis presented in Chapters 3, 4, 5 and 6, a switching model is presented in Chapter 7 considering the control strategies and schemes for both, AC and DC controllers. Waveforms of simulation results are presented comparing the mathematical and switching model to validate the two different proposed methods presented in Chapter 6.

The two methods that exploit the potential of the phase shifted 3rd harmonic are based on using the reduced peak synthesised voltage in two different ways including increasing the output voltage at the MMC terminals or reducing the size of the SM capacitor by allowing higher voltage ripple.

Using the advantage of phase-shifted 3rd harmonic injection, an increase in the output voltage at the MMC terminals can be achieved due to the increase of the fundamental component of the synthesised voltage. This can lead to lower current through the converter considering proper scaling of the transformer turn ratio. Instead, the second method considers the reduction of the SM capacitor for a given design of the MMC. The analysis and simulation results presented in Chapters 6 and 7 showed that reducing the size of the SM capacitor by allowing higher voltage ripple can be achieved. It is noted that the first approach requires a dedicated design of the conversion system, and its applicability is therefore limited to new installations. The second option, however, considers a given design of the MMC and only revisits the size of the SM capacitors by exploiting the increased available voltage margin after 3rd harmonic injection.

Finally, the presented switching model and control schemes in Chapter 7 were validated in Chapter 8 using an existing experimental rig. A comparison of waveforms of the experimental and switching model were also presented, validating the first method proposed in Chapter 6 that can be used to improve the operation and design of the MMC. Experimental results have shown that the phase-shifted 3rd harmonic can be used to improve operation and design of the MMC by increasing the output voltage at the MMC terminal and therefore reducing the AC current and hence the loss.

The main focus in Chapter 4 and 5 was on the generalised impact of harmonic injection on energy deviation reduction including the 3rd harmonic voltage injection, 3rd harmonic current, 3rd harmonic voltage and current, 2nd harmonic current injection with 3rd harmonic voltage injection. One of the observations in Chapter 4 was that using phase-shifted 3rd harmonic injection has no impact on the energy deviation. Therefore, phase-shifted 3rd harmonic voltage injection was not considered in Chapter 5 where the peak to peak energy variation is discussed for SM capacitor sizing.

However, in Chapter 6, the two different methods presented to improve the operation and design of the MMC, including increasing the MMC output voltage or reducing its SM capacitor sizes, are based on the optimal 3rd harmonic amplitude with phase shift.

In the proposed analysis, the amplitude of the phase shifted 3rd harmonic has been chosen equal to the amplitude of the optimal value for the classical 3rd harmonic injection, for simplicity. However, the selected value might not be the optimal solution in case of phase shifted 3rd harmonic voltage injection. Considering both the amplitude and phase shift, makes the analysis more comprehensive but also more complicated.

Another limitation of the study presented in this thesis is that only steady state operation has been considered for the phase shifted 3rd harmonic injection. This means that the dynamic response to active and reactive power demand changes was not investigated, as well as any transient operation due to changes in the DC and AC grids or system fault. This is due to the fact that most of the focus of the work has been on the detailed analysis of the different injection methods and on the potential benefits of the new concept of phase shifted 3rd harmonic. Operation of the method under transient and abnormal operating conditions can be carried out in future study based on the same method considering all the limitations such as fault ride through capability. It can also be observed that harmonic injection has an impact on switching losses. Due to time constraints, conduction and switching loss were not considered in this study.

Another potential issue is the impact of the 3rd harmonic voltage injection on the 100Hz circulating current control. The circulating current is normally eliminated using a PR controller. Therefore, an investigation on the impact of 3rd harmonic voltage injection on the circulating current control needs to be included considering both classical and phase-shifted 3rd harmonic injection. Even though the experimental results in this thesis do not show any unexpected behaviour, further investigation is needed to prevent undesired effects on the converter.

Finally, only one operating point was validated experimentally. This is due to the limitation caused by synthesised voltage due to the initial design of the experimental rig, which has a small limb inductance and large limb resistance. Therefore, full experimental validation of the method with variable operating points should be conducted.

9.2 Future Work

According to the discussion above, it can be concluded that the following points are opened for further investigation, both theoretical and experimental:

- Future investigation to determine the optimal 3rd harmonic amplitude and phase shift when phase-shift 3rd harmonic voltage injection is considered.
- The impact of both 3rd harmonic voltage and 2nd harmonic current injection presented in Chapter 4 on the two methods proposed in Chapter 6.
- An investigation on the impact of 3rd harmonic voltage injection (classical/phase-shifted) on the MMC fault ride-through capability.
- Investigation on the impact of classical and phased-shifted 3rd harmonic voltage injection on switching loss for the two methods proposed in Chapter 6.
- The impact of 3rd harmonic voltage injection on the circulating current control.
- Full experimental validation of the two proposed methods with variable operating point.

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Appendix A

A Power Convention

In this thesis, the power convention used is the Generator convention as illustrated by Figure A.1.

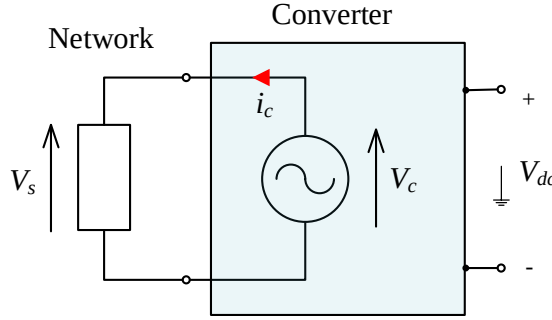


Figure A.1: The VSC converter connected to grid.

In this convention it is assumed that the measured current of the converter is positive when it is out from the converter. The active power is given by;

$$P = V_c I_c \cos \phi \quad (\text{A.1})$$

where ϕ is the phase angle between the current and voltage and is taken to be positive when the current lags the voltage with the labelling conventions specified in Fig 1.

The sign of active power can be given by the ballpoints below:

- Positive P the MMC is inverting, $-90^\circ < \phi < 90^\circ$
- Negative P the MMC is rectifying, $-90^\circ > \phi > 90^\circ$

$$Q = V_c I_c \sin \phi \quad (\text{A.2})$$

The sign of reactive power can be positive (+ve) or negative (-ve) as given by the ballpoints below:

- Positive Q the MMC looks capacitive, $\phi = 90^\circ$
- Negative Q the MMC looks inductive, $\phi = -90^\circ$

Figure A.2 illustrates a PQ curve where the positive and negative P and Q for the convention used. In this case, the converter is considered like an equivalent active component (generator) on the AC side.

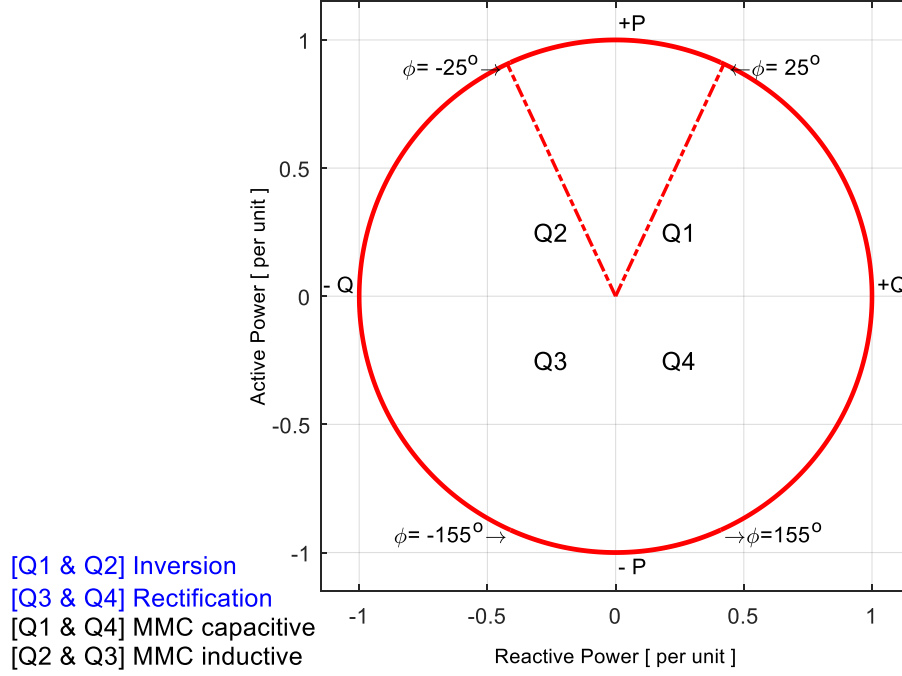


Figure A.2: P and Q curve for the generator power convention.

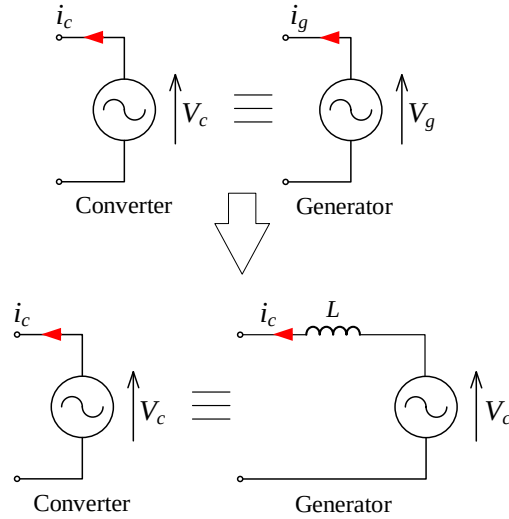


Figure A.3: The equivalent circuit of the VSC converter for the generator convention.

The phasor diagrams of Figure A.4 illustrated below determine the definition of positive (+ve) and negative (-ve) P and Q used in this thesis.

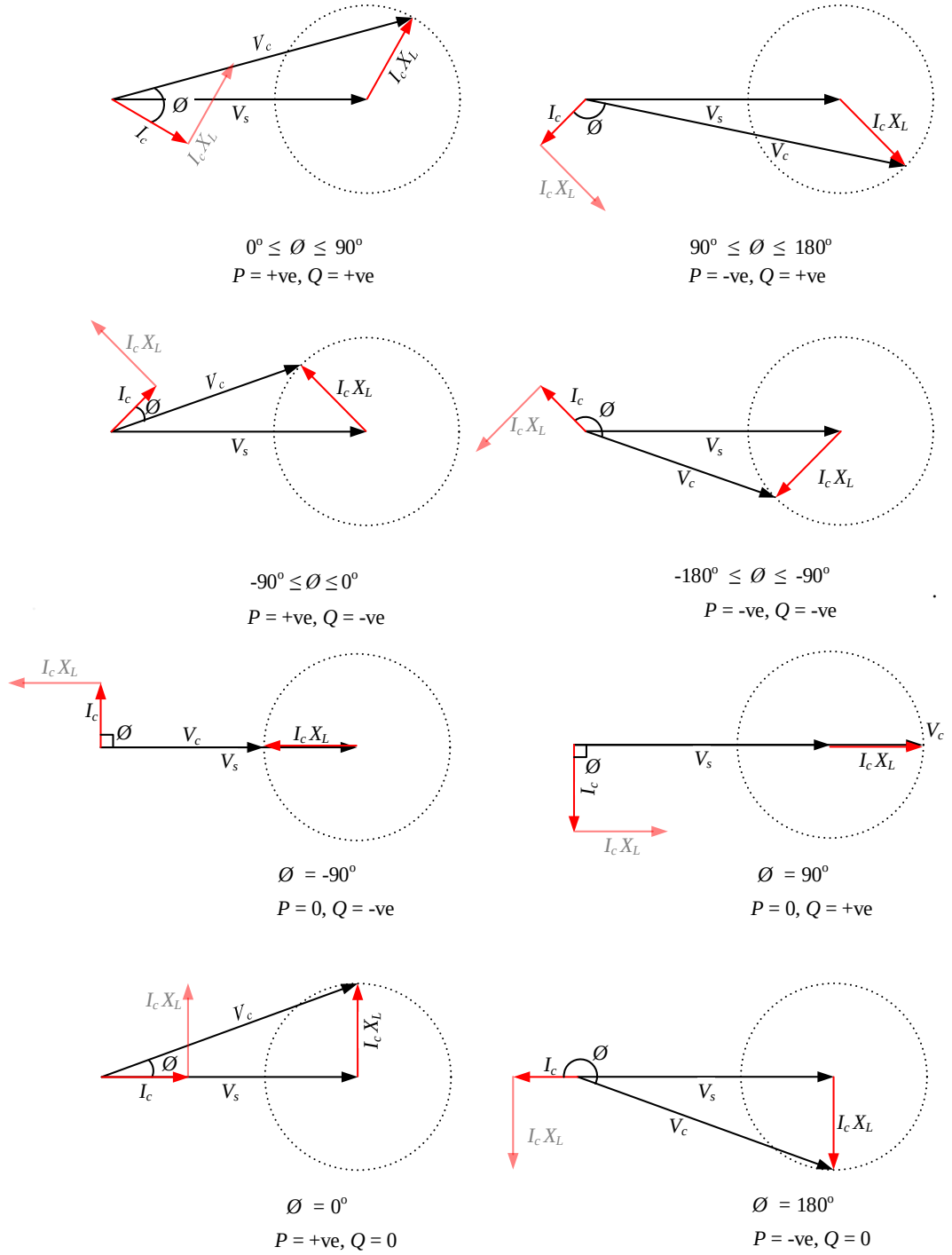


Figure A.4: Phasor diagram of the converter terminal voltage and the AC current.

Appendix B

B Classical and phase-shifted 3rd harmonic injection

The concept of the 3rd harmonic phase shift can be expressed mathematically by (B.1) or (B.2) as illustrated by Figure B.1 or Figure B.2 respectively. The phase-shift can be expressed by how much 3rd harmonic component is advanced or delayed as expressed by (B.3).

$$v(t) = \hat{V}_{C1} \sin(\omega_0 t - \theta_{v1}) + \hat{V}_{C3} \sin(3\omega_0 t - \theta_{v3}) \quad (\text{B.1})$$

$$v(t) = \hat{V}_{C1} \cos(\omega_0 t - \theta_{v1}) - \hat{V}_{C3} \cos(3\omega_0 t - \theta_{v3}) \quad (\text{B.2})$$

$$\theta_{v3} = 3\omega_0 t_{\text{shift}} \quad (\text{B.3})$$

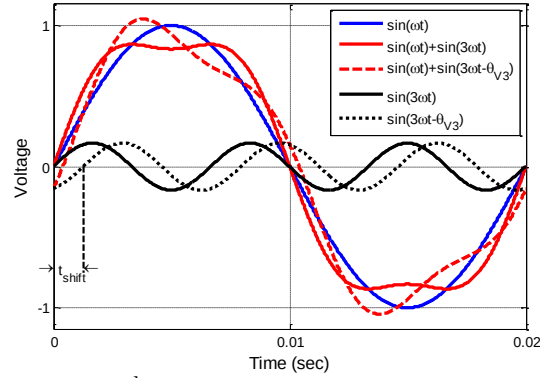


Figure B.1: 3rd harmonic phase shift for sine wave.

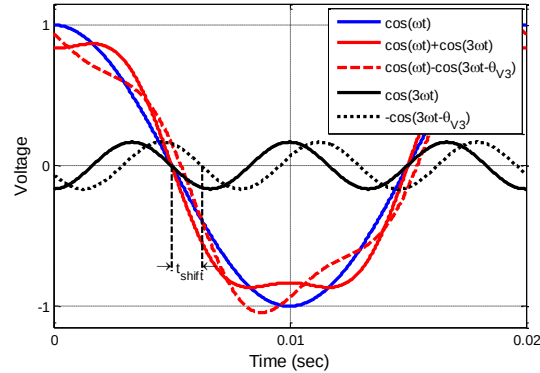


Figure B.2: 3rd harmonic phase shift for cosine wave.

Appendix C

C Capacitor Voltage Ripple

The capacitor average energy can be expressed by the equation below.

$$\bar{E} = \frac{e_{max} + e_{min}}{2} \quad (C.1)$$

Because $e(t)$ contains only harmonics plus DC.

For the MMC, the nominal SM capacitor average voltage is given by (C.2).

$$V_{C_{sm}.n} = \frac{V_{dc}}{N_{SM}} \quad (C.2)$$

This however can be expressed using the maximum and minimum capacitor voltage as given below.

$$V_{C_{sm}.n}^2 = \frac{V_{C_{sm}.max}^2 + V_{C_{sm}.min}^2}{2} \quad (C.3)$$

Put;

$$V_{C_{sm}.max} = V_{C_{sm}.min} + \Delta V \quad (C.4)$$

Substituting (C.4) into (C.3), thus;

$$V_{C_{sm}.n}^2 = \frac{(V_{C_{sm}.min} + \Delta V)^2 + V_{C_{sm}.min}^2}{2} \quad (C.5)$$

$$V_{C_{sm}.n}^2 = \frac{V_{C_{sm}.min}^2 + 2\Delta V V_{C_{sm}.min} + \Delta V^2 + V_{C_{sm}.min}^2}{2} \quad (C.6)$$

$$V_{C_{sm}.n}^2 = \frac{V_{C_{sm}.min}^2 + 2\Delta V V_{C_{sm}.min} + \Delta V^2 + V_{C_{sm}.min}^2}{2} \quad (C.7)$$

$$V_{C_{sm}.n}^2 = V_{C_{sm}.min}^2 + \Delta V V_{C_{sm}.min} + \frac{\Delta V^2}{2} \quad (C.8)$$

After the manipulation, (C.8) becomes;

$$V_{C_{sm}.n}^2 = \left(V_{C_{sm}.min} + \frac{\Delta V}{2}\right)^2 + \frac{\Delta V^2}{4} \quad (C.9)$$

However, if $\frac{\Delta V^2}{4} \ll V_{C_{sm}.max}$, this yields;

$$V_{C_{sm}.n} = V_{C_{sm}.min} + \frac{\Delta V}{2} \quad (C.10)$$

The capacitor average voltage given by (C.10) is only true if $\frac{\Delta V^2}{4} \ll V_{C_{sm}.max}$ otherwise the term $\frac{\Delta V^2}{4}$ in (C.9) should be considered as illustrated by Figure C.1 and Figure C.2.

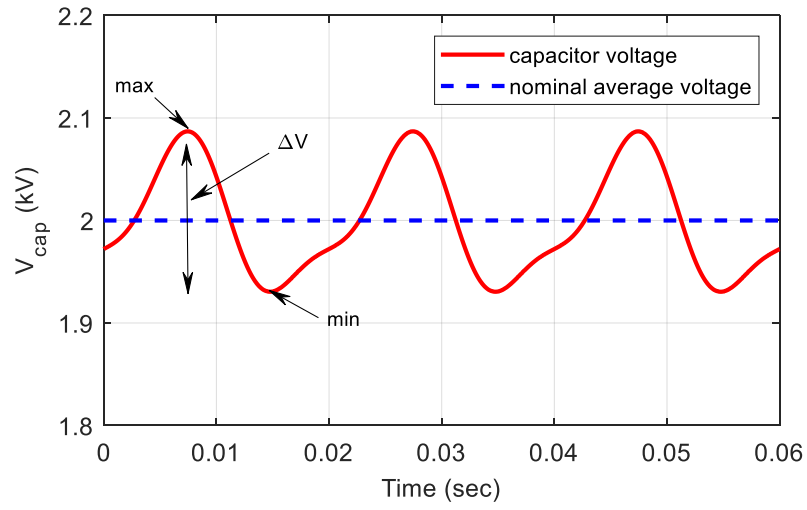


Figure C.1: SM capacitor voltage and its average for $P = 0.94\text{p.u}$ and $Q = 0.41\text{p.u}$.

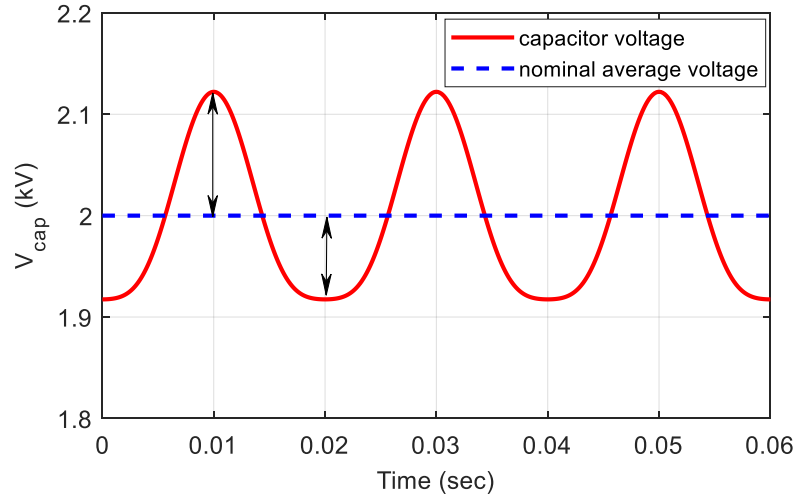


Figure C.2: SM capacitor voltage and its average with only $Q = 1.0\text{p.u}$.

Appendix D

D Optimum phase angle of the 3rd harmonic voltage injection

Considering the PQ curve in Figure A.2, the lookup tables for the optimum phase angle of the 3rd harmonic voltage injection is given by the tables below. The overlap points between quadrants are considered as listed below where Limit means the lower limit is exceeded whilst Over means the upper limit is exceeded (the entry into overmodulation).

Table D.1: Lookup table from the operating point of $-90^\circ \leq \phi \leq 0^\circ$.

Quadrant	ϕ in degrees	m_i													
		1.2	1.15	1.1	1.09	1.08	1.07	1.06	1.055	1.05	1.04	1.03	1.02	1.01	1
Q ₂	θ_{v3} in degrees														
	0	Limit	Over	8	6	0	0	0	0	0	0	0	0	0	0
	-5	Limit	Over	8.5	7	6	4.5	0	0	0	0	0	0	0	0
	-10	Limit	Over	8.5	8	7	6	0	0	0	0	0	0	0	0
	-15	Limit	Over	8.5	8.5	8	6.5	0	0	0	0	0	0	0	0
	-20	Limit	Over	8.5	8.5	8	7	5.5	0	0	0	0	0	0	0
	-25	Limit	Over	8.5	8.5	8	7	6	0	0	0	0	0	0	0
	-30	Limit	Over	Over	8.5	8	7.25	6	5.5	0	0	0	0	0	0
	-35	Limit	Over	Over	Over	8	7.5	7	6	0	0	0	0	0	0
	-40	Limit	Over	Over	Over	8	7.75	7	6	6	0	0	0	0	0
	-45	Limit	Over	Over	Over	Over	8	7	6	6	6	4	0	0	0
	-50	Limit	Over	Over	Over	Over	Over	7	6	6	6	4	0	0	0
	-55	Limit	Over	Over	Over	Over	Over	Over	6	6	6	4	0	0	0
	-60	Limit	Over	Over	Over	Over	Over	Over	Over	6	6	4	0	0	0
	-65	Limit	Over	Over	Over	Over	Over	Over	Over	Over	6	4	0	0	0
	-70	Limit	Over	Over	Over	Over	Over	Over	Over	Over	Over	4	0	0	0
	-75	Limit	Over	Over	Over	Over	Over	Over	Over	Over	Over	4	0	0	0
	-80	Limit	Over	Over	Over	Over	Over	Over	Over	Over	Over	Over	0	0	0
	-85	Limit	Over	Over	Over	Over	Over	Over	Over	Over	Over	Over	0	0	0
	-90	Limit	Over	Over	Over	Over	Over	Over	Over	Over	Over	Over	0	0	0

- (Q_1 and Q_2) when $\emptyset = 0^\circ$, same phase shift as shown in Table D.1 and Table D.3.
- (Q_3 and Q_4), $\emptyset = \pm 180^\circ$, the same phase shift as shown in Table D.2 and Table D.4.
- Similarly, (Q_2 and Q_3) when $\emptyset = 90^\circ$, same phase shift as shown in Table D.1 and Table D.2.
- (Q_1 and Q_4) when $\emptyset = -90^\circ$, same phase shift as shown in Table D.3 and Table D.4.

Table D.2: Lookup table from the operating point of $-180^\circ \leq \emptyset \leq -90^\circ$.

		m_i													
		1.2	1.15	1.1	1.09	1.08	1.07	1.06	1.055	1.05	1.04	1.03	1.02	1.01	1
Quadrant	$\emptyset$ in degrees	θ_{v3} in degrees													
Q ₃	-90	Limit	Over	Over	Over	Over	Over	Over	Over	Over	Over	Over	0	0	0
	-95	Limit	Over	Over	Over	Over	Over	Over	Over	Over	Over	Over	0	0	0
	-100	Limit	Over	Over	Over	Over	Over	Over	Over	Over	Over	Over	0	0	0
	-105	Limit	Over	Over	Over	Over	Over	Over	Over	Over	Over	-4	0	0	0
	-110	Limit	Over	Over	Over	Over	Over	Over	Over	Over	Over	-4	0	0	0
	-115	Limit	Over	Over	Over	Over	Over	Over	Over	Over	-6	-4	0	0	0
	-120	Limit	Over	Over	Over	Over	Over	Over	Over	-6	-6	-4	0	0	0
	-125	Limit	Over	Over	Over	Over	Over	Over	-6	-6	-6	-4	0	0	0
	-130	Limit	Over	Over	Over	Over	Over	-7	-6	-6	-6	-4	0	0	0
	-135	Limit	Over	Over	Over	Over	-8	-7	-6	-6	-6	-4	0	0	0
	-140	Limit	Over	Over	Over	-8	-7.75	-7	-6	-6	0	0	0	0	0
	-145	Limit	Over	Over	Over	-8	-7.5	-7	-6	0	0	0	0	0	0
	-150	Limit	Over	Over	-8.5	-8	-7.25	-6	-5.5	0	0	0	0	0	0
	-155	Limit	Over	-8.5	-8.5	-8	-7	-6	0	0	0	0	0	0	0
	-160	Limit	Over	-8.5	-8.5	-8	-7	-5.5	0	0	0	0	0	0	0
	-165	Limit	Over	-8.5	-8.5	-8	-6.5	0	0	0	0	0	0	0	0
	-170	Limit	Over	-8.5	-8	-7	-6	0	0	0	0	0	0	0	0
	-175	Limit	Over	-8.5	-7	-6	-4.5	0	0	0	0	0	0	0	0
	-180	Limit	Over	-8	-6	0	0	0	0	0	0	0	0	0	0

- Limit: means the lower limit is exceeded.
- Over: means the upper limit is exceeded (the entry into overmodulation).

Table D.3: Lookup table from the operating point of $0^\circ \leq \varnothing \leq 90^\circ$.

Quadrant	$\varnothing$ in degrees	m_i													
		1.2	1.15	1.1	1.09	1.08	1.07	1.06	1.055	1.05	1.04	1.03	1.02	1.01	1
		θ_{v3} in degrees													
Q ₁	0	Limit	Over	8	6	0	0	0	0	0	0	0	0	0	0
	5	Limit	Limit	4	0	0	0	0	0	0	0	0	0	0	0
	10	Limit	Limit	0	0	0	0	0	0	0	0	0	0	0	0
	15	Limit	Limit	0	0	0	0	0	0	0	0	0	0	0	0
	20	Limit	Limit	0	0	0	0	0	0	0	0	0	0	0	0
	25	Limit	Limit	0	0	0	0	0	0	0	0	0	0	0	0
	30	Limit	Limit	0	0	0	0	0	0	0	0	0	0	0	0
	35	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	40	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	45	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	50	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	55	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	60	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	65	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	70	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	75	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	80	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	85	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	90	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0

- Limit: means the lower limit is exceeded.
- Over: means the upper limit is exceeded (the entry into overmodulation).

Table D.4: Lookup table from the operating point of $90^\circ \leq \varnothing \leq 180^\circ$.

		m_i													
		1.2	1.15	1.1	1.09	1.08	1.07	1.06	1.055	1.05	1.04	1.03	1.02	1.01	1
Quadrant	$\varnothing$ in degrees	θ_{v3} in degrees													
Q ₄	90	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	95	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	100	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	105	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	110	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	115	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	120	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	125	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	130	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	135	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	140	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	145	Limit	0	0	0	0	0	0	0	0	0	0	0	0	0
	150	Limit	Limit	0	0	0	0	0	0	0	0	0	0	0	0
	155	Limit	Limit	0	0	0	0	0	0	0	0	0	0	0	0
	160	Limit	Limit	0	0	0	0	0	0	0	0	0	0	0	0
	165	Limit	Limit	0	0	0	0	0	0	0	0	0	0	0	0
	170	Limit	Limit	0	0	0	0	0	0	0	0	0	0	0	0
	175	Limit	Limit	-4	0	0	0	0	0	0	0	0	0	0	0
	180	Limit	Limit	-8	-6	0	0	0	0	0	0	0	0	0	0

- Limit: means the lower limit is exceeded.
- Over: means the upper limit is exceeded (the entry into overmodulation).

Appendix E

E The equivalent circuit of the MMC

Similar to the conventional VSC, the three-phase modular multilevel converter (MMC) can be represented by a single phase equivalent circuit. illustrates a single line diagram of the equivalent circuit of ideal MMC. The synthesised voltage of MMC is represented by the AC source V_C . The grid AC voltage is represented by voltage source V_S and line components R_T and L_T . Therefore, and in order to control the power flow, the converter synthesised voltage magnitude and phase is controlled. To achieve correct performance, the measurements of the parameters should be at the point of common coupling (PCC). This point represents the point in which the MMC should synchronise with the AC system as shown in Figure E.1.

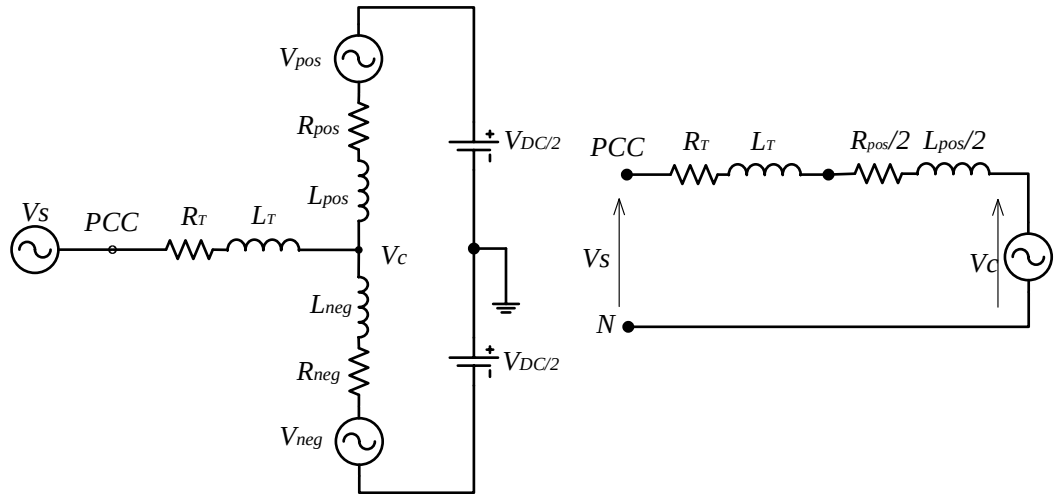


Figure E.1: Single line diagram of MMC converter connector to AC system and the equivalent circuit of MMC.

Based on the AC MMC equivalent model illustrated by Figure E.1, the output voltage at the MMC terminals can be represented similarly to the conventional VSC by (E.1)[28].

$$v_{S,j} = (Ri_j + L \frac{di_j}{dt}) + v_{C,j} \quad (E.1)$$

Where $v_{s,j}$ and i_j are the 3-phase system voltages and currents respectively. $v_{c,j}$ is the fundamental AC voltages synthesised by the converter connected to the grid through impedance of $R_T + j\omega L_T$. where R is the equivalent resistance of the AC system considering and the MMC as seen from the utility grid. For an HVDC system based on VSC converters, the limb resistance R_{pos} and R_{neg} of the MMC is small enough and it can be neglected. However, this will be considered in order to achieve a general analysis. L is the system equivalent inductance including line inductance, converter transformer leakage inductance and half of the limb buffer inductance. j is denoted for the three phases a, b and c . where;

$$L = L_T + \frac{L_{pos}}{2} \quad (E.2)$$

And;

$$R = R_T + \frac{R_{pos}}{2} \quad (E.3)$$

In order to control the power flow, it is required to control the magnitude of the synthesised voltage of the converter and the phase angle. One approach is the $dq0$ technique. This basically based on transforming the ABC system into two stationary frames; $\alpha\beta$, which in return is transformed to the rotating frame $dq0$ as shown by Figure E.2.

E.1 The transformation from ABC to $\alpha\beta$ Frame

The relationship between the stationary $\alpha\beta$ frame and the ABC of the three-phase voltages and currents transformation can be achieved by;

$$v_{s,\alpha\beta} = T \cdot v_{s,j} \quad (E.4)$$

$$i_{\alpha\beta} = T \cdot i_j$$

$$T = \frac{2}{3} \begin{bmatrix} +1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & +\frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \quad (E.5)$$

E.2 The transformation from $\alpha\beta$ Frame to the rotating $dq0$ frame

And similarly, $dq0$;

$$\begin{bmatrix} v_{sd} \\ v_{sq} \\ v_{s0} \end{bmatrix} = \frac{2}{3} \begin{bmatrix} \cos \theta & \cos \theta \\ -\sin \theta & \cos \theta \\ 0 & 0 \end{bmatrix} \cdot \begin{bmatrix} v_{s\alpha} \\ v_{s\beta} \end{bmatrix} \quad (\text{E.6})$$

$$\theta = \tan^{-1} \left(\frac{v_{s\beta}}{v_{s\alpha}} \right) \quad (\text{E.7})$$

Assuming a balanced positive sequence ABC system, the $v_{s0} = 0$ ($Va + Vb + Vc = 0$)

otherwise, it is third the summation $v_{s0} = \frac{1}{3}(Va + Vb + Vc)$.

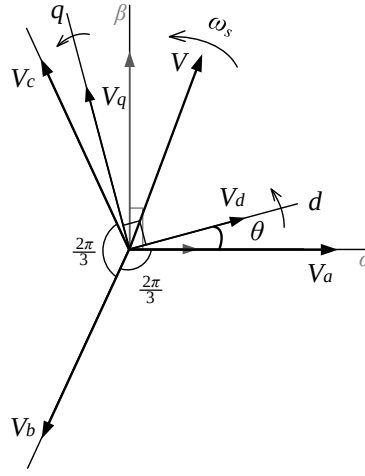


Figure E.2: The transformation from stationary abc to the dq rotating frame.

E.3 Derivation of the Dynamic Equations used in the control

In this section, the dynamic equations of the MMC and the AC system in the dq0 are derived. Considering Figure E.1 and from equation (E.1)

$$v_{s,j} = (Ri_j + L \frac{di_j}{dt}) + v_{c,j} \quad (\text{E.8})$$

Where; $v_{s,j}$ is the grid voltage, $v_{c,j}$ is the converter voltage, i_j is the line current and j denotes the three phases a, b and c .

The equation above represents the ABC system which can be represented in the two stationary frames by:

$$v_{s,\alpha\beta} = (Ri_{\alpha\beta} + L \frac{di_{\alpha\beta}}{dt}) + v_{c,\alpha\beta} \quad (\text{E.9})$$

And the relationship between $dq0$ and $\alpha\beta$ is given by below.

$$v_{S.\alpha\beta} = v_{S.dq} e^{j\theta} \quad (\text{E.10})$$

$$v_{C.\alpha\beta} = v_{C.dq} e^{j\theta} \quad (\text{E.11})$$

$$i_{\alpha\beta} = i_{dq} e^{j\theta} \quad (\text{E.12})$$

Substitute equation (E.10), (E.11) and (E.12) in (E.9). Thus.

$$v_{S.dq} e^{j\theta} = R \cdot i_{dq} e^{j\theta} + L \frac{d(i_{dq} e^{j\theta})}{dt} + v_{C.dq} e^{j\theta} \quad (\text{E.13})$$

Applying the partial integration to (E.13), the yields;

$$v_{S.dq} e^{j\theta} = R \cdot i_{dq} e^{j\theta} + L \left(i_{dq} \frac{d(e^{j\theta})}{dt} + e^{j\theta} \frac{d(i_{dq})}{dt} \right) + v_{C.dq} e^{j\theta} \quad (\text{E.14})$$

$$v_{S.dq} e^{j\theta} = R \cdot i_{dq} e^{j\theta} + L i_{dq} \frac{d(e^{j\theta})}{dt} + L e^{j\theta} \frac{d(i_{dq})}{dt} + v_{C.dq} e^{j\theta} \quad (\text{E.15})$$

$$v_{S.dq} e^{j\theta} = R \cdot i_{dq} e^{j\theta} + j\omega_s L i_{dq} e^{j\theta} + L e^{j\theta} \frac{d(i_{dq})}{dt} + v_{C.dq} e^{j\theta} \quad (\text{E.16})$$

where the voltage vectors (i.e. dq0 frames) rotate at an electrical speed of;

$$\frac{d(\theta)}{dt} = \omega_s$$

Thus;

$$\frac{d(e^{j\theta})}{dt} = j\omega_s e^{j\theta} \quad (\text{E.17})$$

$$v_{S.dq} e^{j\theta} = R \cdot i_{dq} e^{j\theta} + j\omega_s L i_{dq} e^{j\theta} + L e^{j\theta} \frac{d(i_{dq})}{dt} + v_{C.dq} e^{j\theta} \quad (\text{E.18})$$

$$v_{S.dq} = R \cdot i_{dq} + j\omega_s L i_{dq} + L \frac{d(i_{dq})}{dt} + v_{C.dq} \quad (\text{E.19})$$

The vector of the voltages and current are represented by the dq frame;

$$v_{S.dq} = v_{S.d} + j v_{S.q} \quad (\text{E.20})$$

$$v_{C.dq} = v_{C.d} + j v_{C.q} \quad (\text{E.21})$$

$$i_{dq} = i_d + j i_q \quad (\text{E.22})$$

Substituting (E.20), (E.21) and (E.22) into (E.19) this yield;

$$v_{s,d} = R \cdot i_d + L \frac{d(i_d)}{dt} - L \cdot \omega_s \cdot i_q + v_{c,d} \quad (\text{E.23})$$

$$v_{s,q} = R \cdot i_q + L \frac{d(i_q)}{dt} + L \cdot \omega_s \cdot i_d + v_{c,q} \quad (\text{E.24})$$

The Equations (E.23) and (E.24) can be represented by the frequency domain. Thus;

$$v_{s,d}(s) = (R + Ls)i_d(s) - L\omega_s i_q(s) + v_{c,d}(s) \quad (\text{E.25})$$

$$v_{s,q}(s) = (R + Ls)i_q(s) + L\omega_s i_d(s) + v_{c,q}(s) \quad (\text{E.26})$$

From equations (E.25) and (E.26), the ABC system in can be represented in the rotating frame dq as shown in Figure E.3.

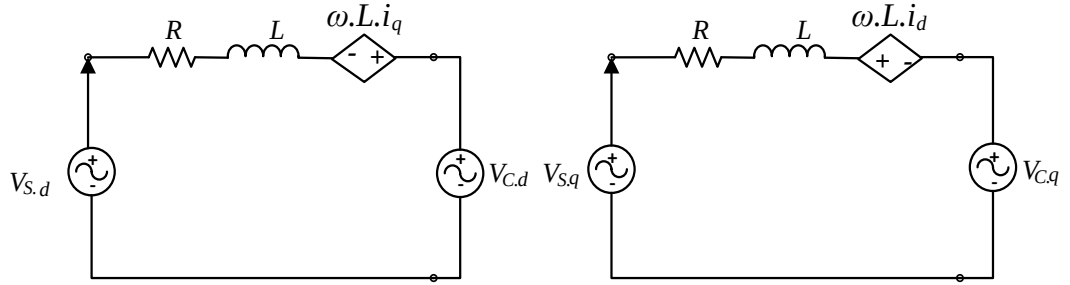


Figure E.3: The dq representation of ABC system in Figure E.1.