

Contrôle et modélisation du convertisseur multi-modulaire AC-AC pour les liaisons offshore à haute tension et basse fréquence

Antony Dasco

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préparée au sein du Laboratoire de Génie Électrique de Grenoble (G2Elab) dans l'École Doctorale Électronique, Électrotechnique, Automatique Et Traitement Du Signal (EEATS)

Contrôle et modélisation du convertisseur multi-modulaire AC-AC pour les liaisons offshore à haute tension et basse fréquence

Modelling and Control of AC-AC Modular Multilevel Converter for High Voltage and Low Frequency Offshore Application

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A mi familia, a la vida y a mi ilustre Universidad Simón Bolívar

Résumé

L'objectif principal de cette thèse consiste à développer des stratégies de modélisation et de contrôle pour la conversion de fréquence en utilisant un convertisseur multimodulaire sans utilisation de lien en continu. Ce convertisseur est intéressant pour l'intégration des énergies renouvelables offshore dans un réseau de transport à basse fréquence. La basse fréquence électrique bénéficie à la fois des avantages des réseaux alternatif et des réseaux continu. L'utilisation de convertisseurs modulaires dans ces réseaux permet, entre autres, de supprimer le transformateur du côté de la basse fréquence électrique. Dans cette optique, l'énergie est convertie par exemple depuis la fréquence nominale à 50 Hz ou 60 Hz (considérée comme la haute fréquence) vers un tiers de 50 Hz. Le convertisseur est aussi capable de transformer d'autres valeurs de fréquence. Dans un premier temps, la méthode de modélisation du convertisseur est reliée à la théorie du transformateur de tension triphasé avec un couplage en YDY. Dans un deuxième temps, l'application du théorème de superposition nous permet d'évaluer séparément la contribution de chaque fréquence. Dans un dernier temps, le contrôle principal du convertisseur utilise des correcteurs résonants, capables de travailler directement avec les courants internes du convertisseur qui sont à deux fréquences nominales. Grâce au découplage entre l'entrée du convertisseur et sa sortie, les défauts et les situations de déséquilibre ne se propagent pas vers la sortie du convertisseur, fonctionnement requis pour l'application en offshore.

Abstract

The purpose of this thesis is to develop strategies to model and control a modular multilevel converter in order to transform frequency without passing through a direct current link. This converter is needed for application in offshore systems where the energy, produced by the offshore renewables, is transmitted using low frequency links. Low frequency presents benefits from alternative and direct current technologies. Moreover, a voltage transformer is not required at the low frequency side when the proposed converter is used as a frequency transformer. In this scenario, frequencies are converted from the highest nominal frequency (50 Hz or 60 Hz) to one third of 50 Hz. Conversion to other frequency values has been tested as well using the same converter. A methodology based on the electrical transformer theory when connected in $Y\Delta Y$ coupling and, the superposition theorem to separate the contribution from both frequencies, are proposed to link these currents with the two three-phase systems at different nominal frequencies. Resonant controllers are introduced to efficiently correct internal current signals of the converter that nominally are constituted of two frequencies: input and output frequencies. The converter is able to decouple the input side from the output. Therefore, faults and unbalanced conditions on the input side are not propagated to the output-side of the converter, which is required for offshore applications.

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Preface

When talking about frequency, distinctions are made to continents and countries, to division and stability, to power and distances. With this idea, the authors propose to start this work with a small passage through history of the electrical frequency [Cowd06].

The idea of changing the electrical frequency has been proposed several times since its first use. However, at the beginning of electricity, a normalized value of frequency was not the real problem because electrical machines were implemented over isolated systems; hence, several frequencies were working at the same time. The selection between Alternating and Direct Current was also an important aspect of the Electrical History, commonly called the War of the Currents. After this moment, AC systems were widely expanded, developing and improving electrical AC devices at one fixed frequency.

Nowadays, with the advances in the power electronic filed, frequency can be changed more easily and DC high power circuits are available. Using the advantage of power electronics, a system working in a lower frequency value is studied in this work.

Defining the Electrical Power System: AC against DC

After 50 years of work trying to invent a successful incandescent lamp, in 1879 Thomas Edison was able to get a carbonized-thread-filament bulb to burn for 13.5 hours. After one year, Edison's lamps were able to keep light for at least 1.000 hours. Later in 1882 he started to create an entire system of electricity generation and distribution. By 1887, Edison's direct current electrical system had more than 120 power station delivering electricity at short distances and 240 V.



Figure a. Edison's Light Bulb, 1879. Photo credits to [Port00]

Low-voltage level are related to more losses of the line and, in order to limit them, power station had to be built close to final users which was a costly effort. To carry the high current value to meet the demand of distributed load; expensive and large diameter copper wire had to be used for electrical cables and lines. Meanwhile in Europe, Serbian inventor Nikola Tesla was trying to eliminate the commutation step of the DC machine and to produce AC directly. In 1884 Tesla travelled to United States to work with Edison. Nevertheless, because of their very different personalities and different visions of their works, their relationship did not last for long.

In 1887, Tesla founded the Tesla Electric Company and began working on AC alternators. During this period, he also formulated the basic mathematical theory still used today in the modern electrical systems. He began with a two identical voltage waveforms, separated by 90 degrees, using four cables. Then, he progressed into the actual three-phase system; using only three wires with 120 electrical degrees out of phase between them. In 1883, Tesla built his first working AC motor. In the other hand, George Westinghouse introduces the transformer who would allow stepping-up the generator voltage to a high value, reducing current level and therefore losses of the line and permitting a longer transmission range. He bought patent rights to the transformer invented by Frenchman Lucien Gaulard and Englishman John Gibbs. In addition, he purchased Tesla's patents to build AC motors. In 1889, Westinghouse had built 870 central lighting stations.

The end of the war was marked by the contest to harness the energy of Niagara Falls; sponsored by the International Niagara Commission and chaired by Lord Kelvin. Kelvin was initially opposed to AC and he changed his mind after visiting the Chicago Fair lighted by Westinghouse, awarding him with the contract on October of 1893.



Kelvin recognized the advantages of the lower cost of an AC system and the ability to transmit energy over long distances. The power plant was successfully inaugurated in 1895.

> Figure b. Tesla's Power Plant from Niagara Falls, producing 15.000 horsepower in 1895. Photo credits in [Kent14]

Designation of the Electrical Frequency Value

Since the demonstration of the principle of electromagnetic induction in 1831 by Faraday, and until the period of System Interconnection (from 1925 to 1990), AC frequency has had different values. The first AC application was made in a lighthouse located at La Hève, France, in 1863. This first commercial machine had a nominal frequency of 53 Hz. but, back in that time, frequency was not considered that important; it was denominated as alternations, full-alternations, turns, and periocity. The decade of 1880 – 90 was considered the Light Period, where frequencies of 133, 125 and 140 Hz were used to transmit electricity. At that time, it was known as the approximately 130 Hz group in North America. Meanwhile, in Europe were implemented lower frequencies: 42, 40 and 30 Hz. In 1891, the company AEG raised their standard frequency to 50 Hz, to avoid any possibility of light flicker (Owen, 1997).

From 1890 to 1925, the biggest changes on frequency matters were made. In 1890, the use of a high frequency at Westinghouse was impeding the development of their induction motors. This is the main reason for their change to 60 Hz. At GE, it was at 1893 when they realized the need of a lower frequency: Steinmetz was able to identify a series resonance in the transmission line, excited by harmonics of the power at 125 Hz. He proposed to reduce frequency at 62.5 Hz and then GE changed to 50 Hz, as the European were using at that moment. Because of the commercial lagging behind Westinghouse, GE decided to change their frequency once more to 60 Hz in 1920. The need to interconnect the different cities and electrical systems start to show up in 1925. The problem was the impossibility to convert frequency at that time, as we know today. In 1925 began the planning to interconnect the electrical system in North America, it was completed in 1948 at 60 Hz. England also began its change to 50 Hz from 1924 until 1938.

The selection of 50 or 60 Hz was suited to the needs of that time. So the outcome was determined by operation conditions in the field. Concepts about power losses, voltage drop problems, distances limits, and any other problem that can be counted in the actual electrical system, were not even supposed to adequate the frequency value at that time. Today's electrical frequency is changing towards a lower value in order to reduce unnecessary losses and to cope with climate changes [RiSa16].

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Notations

Ampere
Celsius Degrees
Faraday
Henry
Hertz
meters
Per unity
second
Volt-Ampere
Volt-Ampere Reactive
Voltage
Watt
Ohm

List of Symbols and Acronyms

$lpha_N$	Negative Firing Angle
α_P	Positive Firing Angle
1φ	One-Phase
3φ	Three-Phase
A	Core Cross-Sectional Area
AC	Alternative Current
ACLF	Alternative Current Low Frequency
В	Susceptance
BP_{BO}	Band in Open Loop
Bx	Flux Density
C	Capacitor
CMC	Conventional Matrix Converters
C_R	Resonant Corrector
d	Distance
DC	Direct Current
D_{max}	Maximal Distance
e	Error
EWEA	European Wind Power Association
f	Frequency in Hertz
G	Conductance
g1, g2, g3, g4	Submodule Gates
GTO	Gate Turn-Off Thyristor
Н	Transfer Function
HV	High Voltage
HVDC	High Voltage Direct Current
Ι	Current

i	Current
$I_{\Sigma H}$	Sigma Current
Idiff	Differential Current
IGBT	Insulated-Gate Bipolar Transistor
IGCT	Integrated Gate-Commutated Thyristor
i_{hn}	Equivalent Arm Current
I_{Hn}	Source Current
IMC	Indirect Matrix Converters
i_{SM}	Submodule Current
i_{xh}	Current Contribution between input and output system sources
$I_{\Delta H}$	Delta Current
Ki	Integrator Gain
kp	Proportional Gain
Lc	Converter Inductance
LF	Low Frequency
Lo	Output Inductance
Ls	Source Inductance
M	Total Submodule Quantity per Arm
m	Modulation Index
MC	Matrix Converters
MMC	Modular Multilevel Converter
MTT	Multi Terminal
N	Turns Number
OPS	Offshore Power Systems
Р	Active Power
PF , $cos(\theta)$	Power Factor
PI	Proportional Integrator
PLL	Phase Lock Loop
PM	Phase-Margin
PMG	Permanent Magnet Generator
PWM	Pulse Width Modulator
Q	Reactive Power
R	Resistance
rms	Root Mean Square
Ro	Output Resistor
Rs	Source Resistor
S	Apparent Power
SM	Submodule
SVM	Space Vector Modulation
<i>t</i>	Time
THD	Total-Harmonic-Distortion
T's	Sample Time
$V_{\sum H}$	Sigma Voltage
V1	Slack Bus Voltage
V2	Generation Bus Voltage
Varm	Arm Voltage
Uc	Capacitor Voltage

v_{hn}	Equivalent Arm Voltage
V_{Hn}	Source Voltage
VSC	Voltage-Sourced Converter
v_{SM}	Submodule Voltage
$V_{\Delta H}$	Delta Voltage
Wn	Energy
X	Reactance
Y	Star Coupling
YDY	Star-Delta-Star Coupling
Z	Impedance
γ	Current Phase-Shift
Δ	Delta Coupling
ΔG	Gain Error
ΔP	Active Power Losses
Δv_{hn}	Voltage Drop In Arms
η	Efficiency
$\mathcal{V}C\Sigma hn$	Aggregated Capacitor Voltage
V_{xh}	Voltage Contribution between input and output system sources
Ψ	Voltage Phase-Shift
ω	Frequency In Radians Per Second

Indexes

8	Source-Side
0	Output-Side
H	H = 1 for phase A, 2, for phase B and 3 for phase C
h	h = 1 for phase a, 2, for phase b and 3 for phase c
n	n = 1 for first converter arm, $n = 2$ for second converter arm
max	Maximum
i	Input
CL	Closed Loop
OL	Open Loop
ii	Other Harmonic Frequency
circ	Circulating

Chapitre i Résumé : Liaisons offshore en basse fréquence

i.1 Introduction

Les travaux effectués lors de cette thèse seront présentés dans quatre chapitres rédigés en langue anglaise. Un résumé de ces quatre chapitres est exposé dans cette première partie du document.

L'objectif principal de cette thèse consiste à développer des stratégies de modélisation et de commande pour la conversion de fréquence en utilisant un Convertisseur Multi-Modulaire (MMC) sans utilisation de lien en continu (DC). Ce convertisseur est nécessaire pour l'intégration des énergies renouvelables offshore dans un réseau de transport à basse fréquence.

La Commission Européenne a promu l'utilisation des réseaux en Haute Tension et Courant Continu (HVDC) pour les développements en offshore [Euro08, Euro14]. En revanche, quelques incertitudes liées aux développements de cette nouvelle technologie ont entraîné des délais sur l'exploitation d'énergie prévue pour les années 2020 et 2030 [MAWK11a, Ren17]. L'HVDC est une technologie qui fonctionne à 0 Hz, et le problème principal dans son développement est lié aux interrupteurs qui, à cette date, ne sont pas commercialisés [BeSm17]. La principale raison est l'absence du passage par zéro des signaux de tensions et courants [LXXZ17]. Cela n'est pas le cas pour les courants alternatifs (AC), qui sont à la fréquence nominale de 50 Hz ou 60 Hz. Nonobstant le passage par zéro des signaux AC, cette technologie est très limitée en distance pour le transport d'énergie par câble sous-marin [FiBE12a].

Dans cette optique, l'utilisation de l'AC à basse fréquence (LF) est introduite dans cette thèse comme une option qui se trouve à mi-chemin entre l'HVDC et l'AC à fréquence nominale [WaWa96a]. La LF permet d'augmenter la distance maximale de transport électrique par câble sous-marin [Bhar96, MROE12]. Les réseaux multiterminaux et les interrupteurs sont très utilisés en AC. La faisabilité de la technologie est donc étudiée et les composants requis pour son développement sont évalués ; les câbles [KPHY00], les interrupteurs [Abb10], les transformateurs [JuKr77] et les convertisseurs AC-AC sont analysés [LiXY11, RuMO16, WrUm09]. Les recherches de ce document se concentrent sur l'étude d'un nouveau convertisseur AC-AC capable de convertir la fréquence pour l'application à haute puissance, haute tension et haute qualité des signaux électriques [CWBW11].

L'utilisation de MMC dans ces réseaux permet, entre autres, de supprimer le transformateur du côté de la basse fréquence électrique, considéré comme le principal inconvénient des réseaux LF. Le convertisseur sera donc utilisé pour passer de la fréquence nominale (considérée comme la haute fréquence) à un tiers de 50 Hz

[Lang10]. Le convertisseur est aussi capable de transformer d'autres valeurs de fréquence.

La modélisation du MMC AC-AC est divisée en deux parties : dans un premier temps, la méthode de modélisation du convertisseur est reliée à la théorie du transformateur de voltage triphasé avec un couplage en star-delta-star (Y Δ Y). Dans notre cas, le convertisseur a une connexion en Δ et les sources des réseaux d'entrée et de sortie sont connectées en couplage Y. Cela permet de créer un lien entre les courants des réseaux avec les courants internes du convertisseur. Ces derniers sont constitués de deux fréquences nominales : fréquences d'entrée et de sortie. En deuxième lieu, l'application du théorème de superposition nous permet d'évaluer séparément la contribution de chaque fréquence.

Le contrôle principal du convertisseur gère les échanges de puissance et d'énergie dans le système entrée-convertisseur-sortie. Il utilise des correcteurs résonants réglés à la fréquence (ou aux multiples fréquences) de travail du convertisseur. Ce correcteur travaille directement avec les signaux sans avoir besoin d'aucune transformation et donc la procédure de contrôle devient plus efficace, surtout dans le MMC AC-AC à deux fréquences. Le contrôleur résonant est aussi utilisé pour le contrôle des courants de circulation. Une nouvelle notion des courants de circulation est introduite pour le MMC AC-AC. Cette notion relie les déséquilibres de courants dans les réseaux avec les courants internes du convertisseur sans fixer une seule valeur de fréquence comme dans d'autres types de MMC.

Grâce au découplage entre l'entrée du convertisseur et sa sortie, les défauts et les situations de déséquilibre ne se propagent pas vers la sortie du convertisseur, fonctionnement requis pour l'application en offshore.

i.2 Augmentation de la distance du transport électrique

L'effet capacitif est considéré comme le principal facteur limitant de la distance du transport d'énergie dans le réseau offshore en AC. Cela est dû à l'utilisation en grande quantité des câbles sous-marins (voir équations 1.1 et 1.2).

En HVDC, les câbles XLPE présentent un problème de charge d'espace dans le diélectrique. Cette accumulation de charge apparaît quand la fréquence utilisée est inférieure à 0.1 Hz. Le problème se résout avec des fréquences supérieures à 1 Hz (voir section 1.2.2).

Pour le calcul de la distance maximale dans une liaison fonctionnant à basse fréquence, un algorithme est proposé. Premièrement, une puissance générée uniquement active est supposée. Compte tenu des effets capacitifs de la ligne, une puissance réactive (Q) est ajoutée au fur et à mesure que la distance augmente. Deuxièmement, pour éviter que la puissance apparente (S) en sortie du câble soit plus grande que sa valeur nominale, une contrainte est ajoutée : cette puissance doit toujours être inférieure ou égale à 1 *p.u.* La quantité de puissance active (P) qui peut être injectée au câble est donc affectée par la distance. Enfin, il est supposé que la distance maximale est atteignable quand le facteur de puissance (PF) en sortie du câble est réduit jusqu'à 0.87. Cette valeur du PF a été choisie de façon à être compétitif par rapport à l'HVDC (voir section 1.1.3). Cela permet d'avoir une disponibilité de l'ordre de 87 % de puissance active prête à être fournie au réseau et donc de limiter la quantité de puissance réactive due au câble. Cette puissance réactive est généralement capacitive ; ce qui entraine des surtensions dans le réseau.

i.2.1 Présentation de l'étude réalisée

Le réseau sous-marin présenté dans ce document a été défini comme un réseau de puissance avec deux nœuds : le nœud bilan (côté d'interconnexion au réseau européen) est modélisé en tant que source de tension à la valeur de 1 p.u. et le nœud génération offshore (côté source), modélisé comme une source de courant avec une puissance uniquement active, celle-ci est une variable contrôlée dont la valeur est fixée à 1 p.u. au départ. Le schéma simplifié du réseau est présenté dans la Figure i suivante.



Figure i. Représentation du réseau modélisé : indice 1 pour le nœud bilan et 2 pour le nœud génération.

La distance maximale autorisée est analysée pour plusieurs types de câbles. Cette étude est faite en utilisant 15 Hz comme fréquence la plus basse. Cette valeur est retenue dans le développement des réseaux de transport ferroviaire de certains pays d'Europe (16 2/3 Hz). Au fur et à mesure que la distance augmente, la puissance générée est progressivement réduite, et le PF du côté nœud bilan est surveillé pour ne pas dépasser la limite imposée de 0.87. La distance maximale, appelée D_{max} , correspond donc à cette valeur du PF.

i.2.2 Modélisation des câbles sous-marins

Les impédances des câbles ont été calculées à partir des références commerciales ABB-XLPE sous-marins [Abb11] (voir Table 1.2.b). Le logiciel EMTP-RV permet de construire le câble à partir de ses dimensions physiques et d'obtenir la valeur des impédances à plusieurs fréquences. Ces valeurs d'impédances ont été implémentées dans le logiciel.

L'étude a été réalisée avec les valeurs de sections suivantes : 500, 630, 800, 1000, 1200, 1400 et 1600 mm². Chaque section du câble est reliée à différents niveaux de tension : 220, 275, 330 et 400 kV.

Les bases de l'algorithme reposent sur un modèle en PI de la ligne, distribué par unité de distance et développé dans le logiciel [Matl15]. Les résultats de l'algorithme ont été testés en simulation avec le logiciel [Emtp16]. Ces résultats ont été comparés avec deux méthodes : le modèle en PI, qui a été reproduit avec vingt éléments distribués et le modèle Wide-Band (WB) du logiciel. Le modèle WB est considéré comme le plus sophistiqué pour les câbles et lignes car il relie les paramètres du câble avec toute la plage de réponse en fréquence. Une comparaison entre les deux méthodes et l'algorithme développé est proposée dans la partie 1.4.3.

i.2.3 Calcul de la distance maximale du transport énergétique

La résolution des flux de puissance du réseau d'étude utilise la méthode de Gauss-Seidel (plus de détails sur la méthode développée se trouvent dans la partie 1.4). Cette méthode s'avère intéressante du fait des itérations de tension dans un réseau à deux nœuds. En effet, seules sont connues la tension du nœud-1 (nœud bilan) et la puissance injectée dans le nœud-2 (nœud génération). Le modèle en PI de la ligne est traité avec la méthode de quadripôles distribuée par unités de distance. Finalement, l'algorithme de résolution est défini dans la Figure ii.



Figure ii. Algorithme de calcul de la distance maximale pour le transport de puissance sous-marin.

À titre d'exemple, un graphique de puissance en fonction de la longueur d'un câble 1400 mm² à 400 kV et 15 Hz est présenté ci-après.



Figure iii. Variation de puissance par rapport à la distance du transport. Limite FP à 204 km. Puissance triphasée = 830 MVA.

Suivant la longueur du câble utilisé, une puissance maximale de génération peut être attribuée. À mesure que la distance augmente, Q1 augmente aussi. Cette augmentation de Q1 entraine une baisse de P2 et, par conséquent, de P1. Ces variations de puissances font diminuer PF1. Dans l'exemple présenté, la distance maximale admissible est de 204 km. En admettant que la longueur du câble est inférieure à 204 km, il est donc possible d'utiliser une fréquence plus élevée pour obtenir le même résultat. Cette idée correspond à l'étude de variation de fréquence présentée dans la partie 1.6.2.

Avec l'utilisation de la basse fréquence AC, il est possible de développer un réseau constitué de câbles sous-marins dont les longueurs seraient supérieures à 200 km. Un avantage important des réseaux à basse fréquence est que ces réseaux sont déjà en utilisation dans les réseaux ferroviaires de quelques pays d'Europe depuis 1905 [Lang10]. Ces réseaux peuvent donc être étudiés pour simplifier l'intégration vers les réseaux offshores.

i.3 Réseaux à basse fréquence électrique

L'Allemagne, l'Autriche, la Suisse, la Norvège et la Suède ont standardisé leur système électrique ferroviaire pour fonctionner à 15 kV et 16 2/3 Hz (voir Figure 2.1). Cependant, en raison d'une volonté de réduire les échauffements dans les conducteurs, la fréquence a été modifiée à 16,7 Hz en 1995 [LiHe02]. Un système ferroviaire est structuré de la même manière que tout autre système électrique. Il a besoin de la source de génération, de la transformation de voltage (et fréquence dans ce cas), du transport, de la protection et de la répartition des charges. Cette répartition est plus difficile à gérer dans les réseaux ferroviaires parce que les charges électriques sont d'une grande capacité, en mouvement et peuvent utiliser du freinage régénératif, ce qui introduit des harmoniques supplémentaires dans le réseau [Wagn89].

Ces réseaux à basse fréquence se sont donc développés pour fonctionner à différents niveaux de tension et tous les composants requis pour leur fonctionnement ont déjà été développés, ce qui offre la possibilité de les utiliser dans un réseau offshore.

i.3.1 Niveaux de tension à basse fréquence

En Allemagne, la puissance totale installée est d'environ 3000 MW [Abb10] dans un réseau maillé (voir Figure 2.2) ; plus de 19.000 km des lignes sont distribuées par 180 postes sources fonctionnant à 110 kV et 16.7 Hz. Ce réseau alimente le réseau ferroviaire qui fait une longueur de 7745 km. Une tension encore plus élevée devrait néanmoins être mise en place pour un réseau offshore afin de réduire les pertes du transport électrique.

Le niveau de voltage atteint jusqu'à 132 kV pour la Suisse. Les applications au niveau offshore auraient besoin de quelques approfondissements car les niveaux de tension attendue est d'environ 400 kV. Ces recherches devraient donc se focaliser sur l'adaptation des composants comme les relais de protection pour pouvoir travailler à une tension plus élevée. Néanmoins, ceci reste en-dehors du cadre de cette thèse.

i.3.2 Relais de protection existants

Il est nécessaire que tout relais de protection soit de petite taille, fiable, rentable, et qu'il puisse garantir des niveaux de sécurité et disponibilité élevés. En outre, il doit minimiser les temps d'arrêt nécessaires à l'installation. Un exemple est proposé par [Abb10] avec les relais *PASS* : ce-sont des disjoncteurs spécialement conçus pour être utilisés dans les réseaux ferroviaires où ils doivent dégager des arcs électriques plus longs à cause d'une fréquence plus lente de 16,7 Hz. Ils sont remplis à gaz SF6 [LSYP85] et sont capables de fonctionner à 110 kV.

Ces interrupteurs sont faciles à transporter et peuvent être remplacés rapidement en cas de défaillance. D'autres versions utilisent un appareillage complètement isolé avec de l'air, réduisant sa taille jusqu'à 60 % par rapport au modèle conventionnel, ce qui présente un grand intérêt pour les réseaux offshores. Ces relais de protection sont certainement moins coûteux et moins volumineux que ceux utilisés dans l'HVDC [Fran11].

i.3.3 Transformateurs à basse fréquence

Un inconvénient important de la basse fréquence est lié à l'augmentation de la taille des transformateurs électriques comme présenté en [WTRY15] :



Figure iv. Transformateur de 250 MW à: 50 Hz (a); 16 2/3 Hz à la même puissance en augmentant la taille du fer (b) ; 16 2/3 Hz à la même puissance en augmentant les enroulements (c).

En comparant avec le transformateur à 50 Hz et en garantissant la même puissance de transformation (voir partie 2.1.4), le volume du transformateur est augmenté de 155.1 %, avec 188.5 % de poids en plus, cela en changeant la taille du fer uniquement. Si par contre, seuls les bobinages sont changés, l'augmentation reste de 75.5 % pour le volume, et 77 % pour le poids. Un équilibre entre variations du fer et des bobinages pourrait être trouvés, mais le résultat final sera toujours un transformateur de plus grande taille. L'utilisation d'un transformateur de voltage à basse fréquence n'est pas souhaitable. Le choix d'un MMC comme variateur de fréquence permet d'éliminer le besoin du transformateur du côté de la basse fréquence.

i.4 Modélisation du convertisseur multi-modulaire en AC-AC

Le convertisseur multi-niveau modulaire (MMC) est devenu la topologie de conversion à multi-niveaux la plus attractive pour les applications à moyenne et haute puissance, en particulier pour l'application dans les réseaux HVDC [DQBS14].

Ces convertisseurs sont généralement développés en connexion AC-DC réversibles [HIFS17, WLGD17, YDLH17], et les applications en basse fréquence [TRWA15] proposent comme solution l'utilisation de deux convertisseurs connectés en configuration *Back-to-Back* (voir partie 2.3.3 et 2.4.3). Cette solution n'est pas intéressante pour l'offshore car elle double la quantité totale de composants requis par rapport à l'HVDC, et devient donc plus coûteuse.

Le MMC peut cependant, être connecté directement en configuration AC-AC [GlMa05, PRFK12, BaMe11]. Il profite alors de la même quantité des composants par rapport au cas AC-DC (voir partie 2.4). Une comparaison avec d'autres convertisseurs de conversion directe de fréquence pour l'application offshore à basse fréquence est présentée dans le tableau ci-dessous.



Tableau i. Comparaison entre différents convertisseurs transformateurs de fréquence pour l'application offshore.

Le MMC en conversion directe AC-AC est donc retenu comme solution pour le changement de fréquence dans les réseaux offshores fonctionnant à basse fréquence. Une nouvelle topologie de modélisation est proposée pour ce convertisseur quand il est connecté entre deux réseaux triphasés

i.4.1 Modèle exact du MMC AC-AC

Le modèle exact est le modèle le moins élaboré du convertisseur [BaMB14]. Il décrit les équations électriques pour chaque configuration de circuit.

Sans lien en DC et en utilisant une configuration interne en Δ , le convertisseur permet de connecter deux systèmes triphasés. La configuration Δ est faite en joignant les bras du convertisseur entre eux et donc en formant un Δ entre les phases. Ceci est nécessaire pour garder les harmoniques produites dans le convertisseur, générant de faibles taux d'harmoniques (THD) en entrée et en sortie du convertisseur.

Toutes les variables utilisées dans ce document appartiennent à deux groupes : dans le premier se trouvent les variables reliées à l'intérieur du convertisseur (*Variables du Convertisseur*) et dans le deuxième, il y a les variables externes (*Variables des Lignes*). Elles sont définies dans la Figure 3.5 du document.

Dans la Figure v ci-dessous, un MMC est représenté en configuration triphasée où le côté d'entrée et de sortie sont en AC et peuvent avoir différentes fréquences. Dans cette figure, six courants (ia1, ia2, ib1, ib2, ic1, ic2) circulant dans chaque demi-bras sont représentés. Ces demis-bras sont composés d'une quantité M de sous-modules. Chaque phase est constituée d'un bras, qui se compose de deux demis-bras.



Figure v. Modèle exact du MMC AC-AC.

Chaque sous-module contient un condensateur où l'énergie rentre dans une fréquence et sort dans une autre fréquence, ce qui rend possible la conversion de fréquence dans le convertisseur. Plus de détails sur les sous-modules sont présentés dans la partie 3.2.2. Lorsque les lois de contrôle doivent être mises en œuvre, il faut que le modèle exact discontinu soit transformé en un modèle invariant continu qui fournisse la meilleure représentation du système pour l'analyse dynamique et pour le réglage du contrôle. Dans ce contexte, un modèle moyen du convertisseur est proposé pour développer les équations de contrôle. La différence avec le modèle exact est l'absence des hautes fréquences de commutations, qui sont indésirables et normalement atténuées. Le modèle moyen du convertisseur est présenté dans la section 3.3.

Une dernière simplification est ajoutée sur le modèle moyen pour pouvoir développer le contrôle principal du convertisseur. Cela consiste à transformer les condensateurs du modèle moyen du MMC en sources de voltages, contrôlées directement avec la forme d'onde requise dans chaque demi-bras. La contribution de chaque réseau à chaque fréquence dans le modèle simplifié du convertisseur est obtenue en appliquant la méthode de superposition comme suit en Figure vi (voir aussi partie 3.3.2). La représentation est faite pour une seule phase et peut-être appliquée de la même façon pour les autres phases.



Figure vi. Modèle moyen simplifié du convertisseur et théorème de superposition appliqué.

Lors de l'application du théorème de superposition, deux nouvelles variables sont définies pour chaque valeur de fréquence : source à la fréquence d'entrée et, charge à la fréquence de sortie. Du côté de la source (s) : la contribution du courant de la source dans la charge (i_{xhs}) et la tension au point de connexion de charge induit par le côté source (v_{xhs}). Et du côté de la charge (o) : la contribution du courant de charge dans la source, déphasé de -2 π /3 de la phase analysée ($i_{x(h+1)o}$) et la tension induite dans le point de connexion de la source ($V_{x(h+1)o}$).

Ces courants sont normalement égaux à zéro afin d'éviter d'introduire la fréquence de la source dans la charge, ou la fréquence de la charge dans la source. Les index H et h ont déjà été définis comme les phases pour les variables des lignes et du convertisseur (voir Figure 3.5). L'indice x représente uniquement le nom de la variable inconnue. Cette tension peut être dans n'importe quelle fréquence du système ou même dans une combinaison des deux fréquences.

Les équations du contrôle principal du convertisseur ont été obtenues avec cette méthode. Deux équations pour chaque fréquence en sont obtenues : les équations Δ , qui définissent la valeur de voltage résultant de la différence entre deux voltages des demisbras et, les équations Σ , résultant de la somme de deux voltages des demi bras. Ces équations sont définies pour la fréquence de la source en 3.33 et 3.34, et pour la fréquence la charge en 3.37 et 3.38.

Les valeurs des courants contenus dans ces équations sont mesurées avec les deux fréquences en même temps, et aucune contribution de fréquence ne peut être mesurée instantanément. La décomposition de Fourier [Arri97, WeWi73] risque d'être un système trop lent, en conséquence une autre méthode doit être utilisée. Une façon de calculer la contribution de chaque fréquence instantanément a été trouvée en appliquant une analogie qui relie le comportement du convertisseur à un transformateur de tension.

i.4.2 Analogie avec le transformateur de voltage

Les mesures de courants sont réalisées à l'entrée et à la sortie du convertisseur, où chaque contribution de fréquence peut être obtenue instantanément. Le lien entre ces courants et les courants des demis-bras du convertisseur dépend d'un facteur de $\sqrt{3}$ et

d'un déphasage de 30° (voir partie 3.3.3). Cela correspond à la relation entre voltages et courants dans un transformateur de tension en connexion YDY (présenté en Figure vii), avec un rapport de transformateur arbitraire et sans un noyau magnétique, qui dans le MMC a été remplacé par les inductances internes.



Figure vii. Analogie avec un transformateur en $Y\Delta Y$.

La relation entre les courants des réseaux et du convertisseur est représentée vectoriellement comme suit.



Figure viii. Diagramme vectorielle des courants entre le convertisseur et les réseaux.

Le courant de chaque demi-bras est défini par rapport à la contribution de chaque fréquence. Les courants des demi-bras relié à la fréquence de la source sont définis avec l'équation 3.55 et les courants reliés à la fréquence de la charge avec l'équation 3.57.

Le convertisseur MMC AC-AC peut être commande grâce aux méthodologies présentées dans cette partie car tous les courants des demis bras sont connus. Un résumé des stratégies appliquées est présenté dans la section suivante.

i.5 Contrôle du changeur de fréquence

Le fonctionnement du MMC peut être décrit comme suit : la sortie du MMC fournit une valeur de puissance de charge. En conséquence, le niveau d'énergie contenu dans les condensateurs des sous-modules est réduit. Le côté entrée du convertisseur va à compenser ces changements d'énergie en maintenant la charge des condensateurs à un certain niveau. Dans les deux topologies, la commande doit permettre de travailler dans l'autre sens de puissance, où la charge fournit l'énergie qui est dissipée du côté de la source. Ces stratégies seront présentées dans cette section.

i.5.1 Commande principale du MMC

Le contrôle principal de ce convertisseur est utilisé pour calculer les courants d'entrée et de sortie, qui ont été corrigés à l'aide des correcteurs résonnants. Un correcteur résonnant permet de contrôler directement les variables sinusoïdales, et aucune transformation est requise. La réponse en boucle ouverte des convertisseurs résonants utilisés est détaillé ci-après.



Figure ix. Réponse en boucle ouverte des correcteurs résonants.

Ces correcteurs ont été particulièrement utiles dans le MMC AC-AC, car ils permettent de commander des signaux qui sont naturellement formés de deux fréquences différentes. Lorsque les paramètres des correcteurs sont reliés au système à utiliser (voir partie 4.2.2), on obtient des performances très dynamiques, même dans les situations de défaut.

i.5.2 Commande secondaire du MMC

La commande secondaire du convertisseur est organisée en deux niveaux différents, la gestion de l'énergie et le contrôle de puissance réactive. Les deux suivent les variations des niveaux de voltage dans les condensateurs et ils sont liés à l'énergie interne du convertisseur (voir Figure x).

Lorsque l'énergie change, une référence de courant d'entrée triphasé est calculée à partir de ce contrôle secondaire afin de maintenir l'énergie dans un certain niveau.

L'idée est de maintenir les voltages du condensateur en-dessous de leurs limites opérationnelles.



Figure x. Énergie dans les condensateurs et procédure de recharge gérée par la commande secondaire.

Un lien entre le niveau de charge des condensateurs et la puissance réactive dans le système *entrée-convertisseur-sortie* est aussi présenté dans la section 4.4.2. Le contrôle de la puissance réactive injectée par la source est essentiel pour éviter quelques problèmes de divergence dans les voltages des condensateurs d'un même bras (un exemple de cette divergence est visible dans la Figure 4.23).

Néanmoins, les transitoires dans le réseau, comme par exemple les variations de puissance, modifient la valeur du voltage. C'est pourquoi une nouvelle référence de puissance réactive doit être définie automatiquement par la commande. Au cours de cette étape, un courant de circulation pourrait être généré dans le système. Ce courant apparaît pendant les transitoires et reste dans le système, une autre étape de contrôle doit donc être appliquée pour éliminer le courant.

i.5.3 Contrôle des courants de circulation

Un contrôle tertiaire permet de maintenir la stabilité dans le convertisseur en minimisant la valeur du courant de circulation, qui a été définie comme nulle dans la partie de modélisation étant donné que le convertisseur a été supposé équilibré et invariant.



Figure xi. Représentation des valeurs de courants Δ à minimiser.

Ce contrôle relie les courants d'entrée et de sortie avec les courants des demis-bras afin de générer une valeur de tension à chaque extrémité de demi-bras. Cela permettra de maintenir n'importe quelle valeur d'harmonique à l'intérieur du convertisseur. Cette valeur, étant défini comme intruse, correspond à toute valeur de fréquence différente de l'harmonique fondamental à l'entrée ou à la sortie du convertisseur. Afin de réduire la valeur du courant de circulation, toutes les tensions des condensateurs des différents demis-bras doivent se maintenir autour d'une même référence.

Les méthodes de commande développées sont aussi appliquées dans le modèle exact du convertisseur. Cependant, une nouvelle stratégie doit aussi être introduite pour équilibrer chaque demi-bras du convertisseur.

i.5.4 Algorithme d'équilibrage des condensateurs

L'insertion des M-condensateurs dans chaque demi-bras du modèle exact du convertisseur nécessite l'utilisation d'un contrôle d'équilibrage par tri (BCA). Ce contrôle permet donc de maintenir les M-voltages des condensateurs superposés au même niveau.



Figure xii. Algorithme d'équilibrage par tri selon le niveau de tension des condensateurs (MMC à 20 niveaux, fréquence maximale de découpage).

Le BCA permet d'acquérir la même forme d'onde dans le groupe de condensateurs d'un demi-bras que celle obtenue avec le modèle moyen (voir Figure xii). Nous avons utilisé un BCA simple pour cette tâche, néanmoins il offre une grande performance lorsque les sous-modules sont physiquement invariants.

Le BCA fait la sélection des sous-modules à utiliser dans chaque cycle qui est défini par la fréquence du générateur de signal modulé en largeur d'impulsion (PWM). Le choix de sous-modules se fait selon trois éléments : la valeur du voltage des condensateurs, le signe du courant de chaque demi-bras et le signe et la valeur de l'indice de modulation (voir section 4.7).

i.6 Conclusions

Une nouvelle méthode pour la modélisation et la commande d'un convertisseur multimodulaire en topologie directe AC-AC est présentée dans ce travail de thèse.

La méthode proposée dans cette thèse vise à intégrer la production d'énergie renouvelable dans des systèmes d'énergie offshore fonctionnant à basse fréquence AC. Le MMC présenté dans ce travail offre l'utilité de pouvoir changer la fréquence et peut
être réglé à n'importe quelle valeur de fréquence. Le MMC peut être conçu pour une application à haute puissance et en haute tension.

Des simulations utilisant le modèle exact du convertisseur sont présentées dans les parties 4.8 et 4.9. Les résultats de la simulation montrent la haute qualité des signaux avec un THD très faible, ce qui est nécessaire pour les réseaux de transport. Les résultats de simulation étudient :

- La variation de la puissance de la charge.
- La réponse phase à la modification des inductances des demis-bras sans modification du contrôle.
- Des changements brusques de tension
- De fortes variations de fréquence et de conversion entre fréquences à 60 Hz, 51 Hz, 50 Hz, 49 Hz, 40 Hz et 30 Hz en entrée et 50/3 Hz en sortie. La conversion de 50 Hz à 60 Hz est aussi possible.

Il est montré dans cette thèse que le MMC peut fonctionner en configuration AC-AC directe et convient aux réseaux offshores en basse fréquence.

Introduction

A 2008 communication from the European Commission (EC) entitled "Offshore Wind Energy: Action needed to deliver on the Energy Policy Objectives for 2020 and beyond" [Euro08] stated that it will be necessary to further develop the use of oceans and seas to promote the EU's energy goals, presented later in this introduction. In order to foster the market deployment, some of the made proposals were:

- To prioritize the offshore deployment over onshore wind, because it is already among the "most competitive technologies", and,
- To boost the modern High Voltage Direct Current (HVDC) cable technology, where *"European Industry has a unique potential"*.

These imposed constraints require an enormous implication from a research point of view from many countries. In France for instance, the EC approved a grant of 86.6 M \in in order to create the *SuperGrid Institute* [Euro14]. Their main implications are the Offshore Wind Energies in HVDC [Supe14]. Nevertheless, HVDC is a relatively new technology [Euro08], and some normal uncertainties in its development might introduce some unwanted delays on the expectations set by the EC; it was expected a growth of the total wind power installed in 2008 of 30-40 times by 2020 and 100 times by 2030. For 2008, the total power was 121 GW. For 2014, the power raised to 370 GW and for 2016 it increased to 487 GM [Ren17]. With this rhythm, it is expected a total power of 750 GW for 2020, which is only 6.2 times more than the total installed power in 2008.

One of the main problems in HVDC is the that until this date no commercially mature products of HVDC circuit breakers exist [BeSm17]; one of the reasons is because of the absence of the natural zero crossing of voltage and current signals when compared to alternative current (AC), increasing the price of the resulting circuit breaker [LXXZ17]. Then, when compared to AC at nominal 50 Hz or 60 Hz, the problem is that this technology is characterized by a very limited submarine transmission distance [FiBE12a].

With this perspective, the use of AC Lower Frequency (LF) is introduced in this thesis as a middle option between nominal AC frequencies and HVDC. It is proved that the use of a lower electrical frequency (LF) enhances the distance of power transmission in AC submarine cable networks. Meshed grids and AC breakers are quite common in AC networks. The idea is therefore to study the feasibility of the technology and the required components to link these LF networks with the existing interconnected systems in AC 50 Hz or 60 Hz. The main components to be studied in the LF system are cables, protection relays, transformers and AC-AC converters. From these components, the priority is given to the study of a newer converter capable of transforming AC-AC for high power, high voltage and high waveforms quality applications. In accordance to the previous perspective, the main objective of this work is to develop a new model and control strategy for a Modular Multilevel Converter (MMC) capable to interconnect two three-phase systems working at different frequencies in a direct and decoupled AC-AC link. Moreover, the converter should be able to work under unbalanced conditions in the electrical system and to recovery autonomously after faults.

This thesis is divided in four chapters. A context based on developments in offshore power system is exposed in *chapter 1*. First, a comparison between the two opposed technologies is presented: AC and DC networks in offshore. Low frequency networks are introduced as a middle option, taking benefits from both technologies that can be advantageous for offshore networks. The choice of the low frequency value has been discussed and a method for calculating the maximum cable length in a link for power transmission has been proposed.

Existing low frequency networks are studied in *chapter 2*, where the main developments of the technology have been introduced thanks to LF railways existing since 1905. LF nevertheless carries some drawbacks like oversize of the electrical transformer when compared to the same power system at 50 Hz or 60 Hz, and longer zero crossing of signal, which is related to longer extinction time of the circuit breakers. For these drawbacks, it has been exposed that circuit breakers that are already in use for the railway system could be adjusted for offshore application, while the elimination of the LF electrical transformer is related to the chosen converter. Three main groups for AC-AC converters have been compared: cycloconverters, matrix converters and back-to-back converters.

None of these structures is suitable for offshore application when compared to the MMC performances. From this context, the MMC is chosen for application in AC low frequency systems and studied in *chapter 3*. The MMC AC-AC has a different topology but the same quantity of components when compared to the reversible AC-DC MMC. Nevertheless, there is only a few amount of researches made on this structure when is working in AC-AC three-phase topology [KaBM15]. A new model methodology is therefore proposed. The main difficulty in the model is that the internal currents of the converter are working at two fundamental frequencies. A link between the two external systems and all internal currents of the converter is proposed. This methodology is followed by the control strategies of the converter.

The input side of the converter regulates the submodules energy using the control strategies exposed in *chapter 4*. This regulation is necessary to compensate changes occasioned by the output side of the converter, where power is supplied from the internal available energy. A resonant controller ensures the principal control of the converter; it has been selected because is capable of tracking waveforms that are a several frequencies at the same time. A new concept of *circulating current* has been introduced for the MMC AC-AC; this current resulted to be different from other studied cases in the literature [BSBV14]. The converter has been analysed using three different models: perfect model, average model and the switching model; all three are tested on simulations

Chapter 1. Offshore Power Systems Links and Impact of the Frequency

Offshore wind turbines started to be installed in 1991, at 2.5 km away from the Danish coast with a rated power of 4.95 MW. For 2011, the total amount of offshore wind power overpassed 1000 MW. And the same year, the European Wind Power Association (EWEA) forecasted a power increment of 40 GW to be operational in 2020 over 18 European Countries, and 150 GW in 2030 [MAWK11b].

All offshore wind turbines have been installed near from the coasts in order to limit the alternative current (AC) transmission distance. AC is a technology established in 1887 [Cowd06] and relatively less expensive to implement than High Voltage Direct Current (HVDC) [QYXA09a]. However, AC transmission distance is limited in distance, and after some length becomes more expensive than DC. A relationship between the cost and the distance can be appreciated in the following Figure 1.1.



Figure 1.1. Distance/Cost tendency within AC and DC submarine networks.

EWEA's expectations are related to the construction of an interconnected oversea network at around 200 km from the coast. On the one hand, such long transmission distances are associated to technological limitations in AC 50/60 Hz, principally imposed by the submarine cables, so 50/60 Hz cannot longer be used. On the other hand, HVDC multi-terminal networks are still in development process, creating many researches subjects. Some of the development areas in HVDC are electrical protections, charge space and harmonics in cables, offshore conversion platforms [PBGM14].

Problems of both technologies could be solved by using a lower frequency than 50/60 Hz. Thus, the resulting admittance value of a cable is directly linked to the frequency value, consequently limiting the transmission distance. By dropping the frequency,

therefore the distance will be increased [FiBE12b], and line protection relays are already existing in an AC system: the low frequency was used for the first time in a rail system in 1902 at 15 Hz, which was the nominal frequency of the induction machines at the time [Duff03]. Low frequency at 16 2/3 Hz is still in use today, mainly in the railway network of some European countries. In addition, these networks have already developed their own low frequency protections, contrarily to point-to-point HVDC, which today uses the AC side to clear failures.

Many advantages for low frequency systems have already been exposed [Bhar96, MROE12], some of these are the increase of the power transmission capability when compared to the 50 Hz equivalent system and also the distance increase. However, a drawback in a low frequency system is the increased size of transformers when compared to 50/60 Hz. However, a transformer could be neglected by using an appropriate AC/AC converter, as it will be presented in Chapter 3.

In this chapter, the feasibility of the low frequency technology is analysed and compared to DC technology in transmission of power using submarine cables. A power flow method is therefore proposed in order to establish the maximum transmission distances using the reactive power level as indicator. Some optimizations of power losses and of the frequency value are presented using several kind of cable sections.

1.1. Offshore Power Systems

One of the advantages in Offshore Power Systems (OPS) is the larger quantity of wind resources as presented in Figure 1.2. The wind profile has a more constant wind speed and less environmental impact because of the largest distance to populated areas. However, OPS have some important disadvantages too: difficult access because of the largest distances, supplementary cost because of the needed of platforms, sea transportation and maintenance. The most advanced technologies in generation, power electronics, system power transmission and security are being introduced to make possible OPS [BlMa17].



wind resources over open sea (more than Tokin offshole) for five standard neights										
	10m		25 m		50m		100m		200 m	
	$\mathrm{ms^{-1}}$	$W m^{-2}$	$\mathrm{ms^{-1}}$	$W m^{-2}$	$\mathrm{ms^{-1}}$	$W m^{-2}$	$m s^{-1}$	$W m^{-2}$	m s ⁻¹	$W m^{-2}$
	>8.0	>600	>8.5	>700	>9.0	>800	>10.0	>1100	>11.0	>1500
	7.0 - 8.0	350-600	7.5-8.5	450-700	8.0-9.0	600-800	8.5-10.0	650-1100	9.5-11.0	900-1500
	6.0 - 7.0	250 - 300	6.5 - 7.5	300-450	7.0-8.0	400-600	7.5 – 8.5	450 - 650	8.0 - 9.5	600 - 900
	4.5 - 6.0	100 - 250	5.0 - 6.5	150 - 300	5.5 - 7.0	200 - 400	6.0 - 7.5	250 - 450	6.5 - 8.0	300 - 600
	<4.5	<100	<5.0	<150	<5.5	<200	<6.0	<250	<6.5	<300

Figure 1.2.Offshore wind power resources [Leit07].

1.1.1 High Voltage Alternative Current OPS

High distances from the coast are related to OPS, requiring transmission of electricity through submarine cables [DaBR16]. When compared to overhead lines, the cable shunt capacitance becomes much more important, limiting the total transmission distance when using AC.



Figure 1.3. (a) Power against distance in an AC offshore power system, (b) Participation of each component in total transmission losses. Reference: [NTAN06].

In Figure 1.3.a, a power losses repartition in a given AC OPS is presented. In that study [NTAN06], the authors estimated that 87 % of the losses are associated to the submarine cables and, in order to reduce that value, the transmission distance must be limited between 50 and 100 km, which is an important disadvantage for classical AC systems.

Three maximum power values were presented in Figure 1.3.a: 900, 500 and 250 MW, related to three different nominal voltages: 400, 220 and 132 kV respectively. The bigger the distance (d) and the voltage (V) are, significantly quicker the power amount drops in a high voltage system. This is because the generated reactive power (Q) by the capacitance (C) is related directly to the voltage at a given frequency (ω):

$$Q_{capacitance} = V \cdot I_{shunt} = (Y \cdot d) \cdot V^{2} = \omega \cdot C \cdot d \cdot V^{2}$$
1.1

$$P_{out} = \sqrt{\left(S_{in}\right)^2 - \left(\omega \cdot C \cdot d \cdot V^2\right)^2}$$
 1.2

For OPS case, where required distances are at least of 150 km, transmission of power in classical AC system is therefore impossible; this is why many authors agree with the use of direct current (DC) power lines [BKHV07, GZGH13, Laza05, PBGM14]. In this case, shunt capacitances are almost inexistent and only the cable resistance limits the distance.

1.1.2 High Voltage Direct Current OPS

High Voltage Direct Current (HVDC) systems are a technological viable option for integration of long distance offshore wind power into the interconnected power system [Raul14]. In the case of point-to-point HVDC connection, protections are needed in the AC-side only. However, the main drawbacks compared to AC system are the needed power converters and, for Multi-Terminal (MTT) implementation, it may be necessary to integrate DC breakers, which will be the case for offshore networks. An HVDC network is presented in Figure 1.4.b. A technical (but not viable) solution to avoid HVDC breakers will require the implementation of more conversion station and therefore using only AC breakers, but this solution would be more difficult and costly to implement in a very mesh network.



Figure 1.4. General comparison between (a) AC OPS and (b) HVDC OPS.

Even if more components are needed in the DC case as presented in Figure 1.4, this is the best possibility to interconnect with offshore power systems when compared to AC 50 Hz. In order to deal with MTT in HVDC, which is the challenge of these kinds of networks [Whit11], many researches have been carried on this technology, principally focused on the power converter [Akag11, WZCZ16, ZBBF16] and to the breaker algorithm area [ARDN16, DeRC13, HoMa14].

As it was said, HVDC permits an increased transmission distance when compared to AC. Nevertheless, as it was presented in Figure 1.1, a distance limit that benefices AC over DC can be found and it depends on the power factor value. A power analysis is then proposed to compare those systems and to fond the required power factor.

1.1.3 Power transmission comparison between AC and DC networks

Multi-phase systems are possible to be built in AC by angular dephasing [GrSt94]. The main advantage of using three phases is to reduce the neutral return in the system. In balanced operation, an equivalent angle of 120° separates the phases. In DC power transmission, where frequency is equivalent to zero, a multi-phase system is inexistent and two poles are used to transmit the power. In some cases is implemented an asymmetrical bipolar network were the ground can be used as the return of current, but in this case, the ground resistance has to be introduced in the model and normally the resistance value is bigger than in a copper cable [Kers12].

A power comparison between both system frequencies will set the distance limit for which DC become a better choice than AC technology. System are first compared at the generation point, for example at the output of the converter from Figure 1.4.b. Power generated in both cases are entirely active and, at this point, no elements introduce reactive power to the system. It will be supposed that the used cable, with a length of d, is the same for both the AC and DC systems. Then, a voltage value is imposed for both cases: it is supposed that a factor ζ is used to multiply the AC voltage in order to compare it with DC. This is because the AC voltage could take many different values as peak or rms, 1 phase or 3 phases. Consequently, all of them will be compared. A one-phase comparison gives:



Figure 1.5. Used cable for comparison between AC and DC.

The nominal current of the cable will be the same in both cases, which correspond to the maximum current value of the used cable. Current and voltages are defined as it follows:

$$I_{DC} = \left| Irms_{AC_{1\phi}} \right|$$
 1.3

$$V_{DC} = \zeta \cdot V_{AC} \tag{1.4}$$

Power calculated in each individual case is:

$$P_{DC} = V_{DC} \cdot I_{DC}$$
 1.5

$$P_{AC_{1\phi}} = Vrms_{AC_{1\phi}} \cdot Irms_{AC_{1\phi}}$$
 1.6

Consequently, a power comparison could be defined for different ζ values:

Peak value
$$\zeta = \sqrt{2}$$
 $P_{AC_{1\phi}} = \frac{1}{\sqrt{2}} \cdot Vrms_{AC_{1\phi}} \cdot Irms_{AC_{1\phi}} = \frac{1}{\sqrt{2}} \cdot P_{DC}$ 1.7

rms value
$$\zeta = 1$$
 $P_{AC_{1\phi}} = Vrms_{AC_{1\phi}} \cdot Irms_{AC_{1\phi}} = P_{DC}$ 1.8

As the actually value that can be used for comparison is not known, it was therefore supposed that the value is in a range between equations 1.7 and 1.8. These power values are in one phase; introducing now a three-phase case for AC and a bipolar DC link, it goes:

$$P_{DC} = 2 \cdot V_{DC} \cdot I_{DC}$$
 1.9

$$P_{AC_{3\phi}} = 3 Vrms_{AC_{1\phi}} \cdot Irms_{AC_{1\phi}}$$
 1.10

$$Vrms_{AC_{3\phi}} = \sqrt{3} \cdot Vrms_{AC_{I\phi}}$$
 1.11



Figure 1.6. Generation point and power transmission between: (a) DC and; (b) AC.

Introducing now the transmission system, it can be defined the Power Factor ($PF = cos(\theta)$), reactive power (Q) and apparent power (S). It has been supposed that power generation is only active, and the reactive power was added by the cable. For a given cable length x, output apparent power would be 1 in per unity of the cable. Active power losses were avoided in both AC and DC cases.



Figure 1.7. Three-phase transmission by cables.

From Figure 1.7 it can be said that:

$$\left(S^{\text{in}}\right)_{AC_{3\phi}} = \sqrt{3} \cdot V_{AC_{3\phi}} \cdot I_{AC_{I\phi}} \cdot FP \qquad \left(S^{out}\right)_{AC_{3\phi}} = \sqrt{3} \cdot V_{AC_{3\phi}} \cdot I_{AC_{I\phi}} \qquad 1.12$$

$$\left(P^{\text{in}}\right)_{AC_{3\phi}} = \sqrt{3} \cdot V_{AC_{3\phi}} \cdot I_{AC_{I\phi}} \cdot FP : \qquad \left(P^{out}\right)_{AC_{3\phi}} = \sqrt{3} \cdot V_{AC_{3\phi}} \cdot I_{AC_{I\phi}} \cdot FP \qquad 1.13$$

$$\begin{pmatrix} Q^{out} \end{pmatrix}_{AC_{3\phi}} = \sqrt{\left(\sqrt{3} \cdot V_{AC_{3\phi}} \cdot I_{AC_{I\phi}}\right)^2 - \left(\sqrt{3} \cdot V_{AC_{3\phi}} \cdot I_{AC_{I\phi}} \cdot \frac{\sqrt{3}}{2}\right)^2} \\ \dots = \sqrt{3} \cdot V_{AC_{3\phi}} \cdot I_{AC_{I\phi}} \cdot \sqrt{1 - FP^2}$$
 1.14

Where S^{in} and P^{in} have the same reduced power value. For the DC case it would be:



Figure 1.8. Bi-polar transmission by cables.

And the power is the same in both sides as no active power losses are taken into account:

$$\left(P^{in}\right)_{DC} = 2 V_{DC} \cdot I_{DC} \qquad \left(P^{out}\right)_{DC} = 2 V_{DC} \cdot I_{DC} \qquad 1.15$$

Both AC and DC can be compared from equations 1.13 and 1.15, using definition from equation 1.4, it goes:

By introducing several values ζ , the required power factor angle that gives the same active power in DC and AC at the output side of the cable is presented in Table 1.1:

Condition	ζ	Resulting Equation	$ \theta \text{ when } \\ \left(P^{out} \right)_{AC_{3\phi}} = P_{DC} \cdot \cos(\theta) $	
1-phase peak	$\zeta = \sqrt{2}$	$\left(P^{out}\right)_{AC_{3\phi}} = \frac{3}{2\sqrt{2}} \cdot P_{DC} \cdot \cos\left(\theta\right)$	19.5°	
1-phase rms	$\zeta = 1$	$\left(P^{out}\right)_{AC_{3\phi}} = \frac{3}{2} \cdot P_{DC} \cdot \cos\left(\theta\right)$	48.1°	
3-phase peak $2 \cdot V_{DC} = \sqrt{2} \cdot \sqrt{3} \cdot V_{AC_{I\phi}}$	$\zeta = \frac{\sqrt{6}}{2}$	$\left(P^{out}\right)_{AC_{3\phi}} = \frac{3}{\sqrt{6}} \cdot P_{DC} \cdot \cos(\theta)$	35.3°	
3-phase rms 2· $V_{DC} = \sqrt{3} \cdot V_{AC_{1\phi}}$	$\zeta = \frac{\sqrt{3}}{2}$	$\left(P^{out}\right)_{AC_{3\phi}} = \frac{3}{\sqrt{3}} \cdot P_{DC} \cdot \cos(\theta)$	54.7°	

Table 1.1. Power factor angle values for comparing AC and DC power transmission.

Cases were the lower transmission distance is obtained for AC, presented in Table 1.1, are related to the use of the peak value of voltages. Therefore, in order to be less optimistic, these values are selected to mark the corresponding limit between DC and AC. Angle values and the corresponding power factor would be defined by:

$$19.5^{\circ} \le \theta \le 35.3^{\circ}$$
 1.17

$$0.94 \ge FP \ge 0.82$$
 1.18

In order to define the limit angle for calculation in this thesis work, it will be selected 30° , because it is located between limits presented in equation 1.17. The maximum distance of a system could be calculated with the procedure presented in this subsection (with equations 1.12, 1.13 and 1.14). Nevertheless, the actually 50 Hz or 60 Hz power system is not capable of long distance transmission using submarine cables. A solution might be the reduction of the frequency value for the required transmission system; this will be explained in the following subsection.

1.1.4 Other technological solutions for OPS

It has been presented that for OPS located between 50 and 100 km from the coast, the best technological solution is the use of AC systems as presented in Figure 1.4.a. This is because is simpler to install and needs a reduced amount of components than DC: only a transformer platform in the worst case and no AC-DC conversion platform is needed, which is related to less changing steps of power, less maintenance and more robustness, less offshore platforms and therefore much lower cost.

If more concentrated and steady wind sources of power are located at approximately 200 km of the coast, therefore for that length, HVDC technology can be used because it presents much less distance limitations. HVDC systems are characterized for presenting a great system quality and efficiency. However, these kinds of systems came with much higher costs and technological challenges that still are in investigation status and the technology is today applied in a point-to-point network only. An alternative solution for OPS is the use of an AC working in a lower frequency than the classical 50 or 60 Hz.



Figure 1.9. OPS representation working at low frequency alternative current (ACLF)

An AC Low Frequency (LF) system is an equivalent from the DC one from Figure 1.4.b and ACLF has the possibility of inserting of Multi-Terminal Networks. Moreover, the offshore side converter could be eliminated if the output frequency from generators were defined directly in low frequency. Power flows in LF will be therefore analysed in the present chapter in order to evaluate its viability. In addition, low frequency is a technology that has being developed since more than a century, and tis technology can therefore be used.

1.2. Low Frequency and DC Systems

Railway electrification was tested and implanted for the first time on the Adriatic Valtellina (Italy) in 1902. In this city, Ganz & Co made a comparative ratio between battery-electric traction, low voltage DC traction using a conductor rail, and 3400 V_{AC} three-phase traction at 15 Hz. After the tests, the low frequency three-phase system was judged a success and was extensively developed in Italy. Its development continues until 1933, and following the widespread wreckage of the system in the 1939-45 war, it was declared obsolete and replaced with 3 kV DC [Duff03].



Figure 1.10. The world first 3-phase-15 Hz locomotive in Valtellina (1898-1902). Designed by Kálmán Kandó in Ganz and Co.

Germany, Sweden and other European countries adopted the 16 2/3 Hz of frequency for their railway system in 1912 [Abb10] and is this frequency value is still in use today. Before power electronic devices became available, power supply in DC was chosen in some other countries.

The advantage of DC is the ease to control the speed of DC motors that depends only on the applied voltage. Other countries chosen alternating current and commutator motors. AC motors speed can be easily controlled, but input frequency must be lower than 50 or 60 Hz to avoid sparking in a commutator and without using a converter [Bhar96]. Frequency transformers were at first rotating machines and the first power supply converter with powerful turn-off semiconductors in the form of Gate Turn-off Thyristors (GTO) went into operation in 1994. Since then, the Integrated Gate-Commutated Thyristor (IGCT) was developed and features a much more advanced switching capability, lower losses and a low-inductance gate as integrated component [Abb10].

The world's largest railway power plant is positioned in Datteln, West Germany. It is 1100 MW hard coal fired power plant completed in 2011 and can provide up of 413 MW in reduced frequency of 16.7 Hz and 110 kV-2 phases to the railway system. Application of this frequency in OPS will reduce the capacitive effect and therefore increase the transmission distance capability in AC.

1.2.1 Low frequency alternative current OPS

Some authors already applied LF and demonstrated its effects over the system distance by testing a range of frequency on a cable. In 1996 is presented one of the first studies in this area [WaWa96b], it was called fractional frequency and it was demonstrated the increased transmission distance and the possibility of using magnetic frequency changer with 95% of efficiency and reversible active power. Cycloconverters are also a viable option to change the frequency [LiXY11], more details about converters will be exposed in chapter 2.

A comparison between 50 Hz and 50/3 Hz system is presented in Figure 1.11.a. A compensation curve is also showed; this compensation is related to the introduction of distributed reactors in the transmission system. The use of reactors will increase costs in the systems, this is why compensation is excluded in analysis proposed in this work. The same frequency value is used as the frequency used in the Germany's train system presented before. It is showed how the transmission distance is largely increased and it can be remarked how the power transmission capability could be raised. An optimization of the transmission distance versus distance is presented in Figure 1.11.b. However, this variable is strongly related with the used cable parameters and the applied voltage level as presented in equations 1.1 and 1.2. A power analysis of the OPS, presented further in this work, is therefore necessary to explore the relationship between frequency, voltage and power of a system and the cable.



Figure 1.11. (a) Submarine transmission power in AC 50 and 16.7 Hz compared at 155 kV [FiBE12a]. (b) Optimization of the offshore distance [MROE12].

1.2.2 Cable characteristics in low frequency

The extruded insulation material for AC power cable cannot be easily applied to DC operation. This is because of the space charge accumulation in the material. It occurs in the dielectrics of normal XLPE [KPHY00] cable when the AC frequency is lower than 0.1 Hz; under these operating conditions, the XLPE electrical insulation properties may become degraded when trapped or low mobility electrically charges rise the space charge, resulting in localized electric stress enhancement [RuTM12]. This can cause further concentration of space charge and lead to premature failure of the material.

The effect is neutralized for bigger frequencies than 1 Hz [FuMa00]. Thus, many researchers have tried to find an effective solution which decreases the space charge accumulation in the materials, for example, addition of conductive inorganic fillers or polarized inorganic fillers to cross-linked polyethylene (XLPE) [PLJK14]. Some of these solutions can raise the DC voltage limit in cables but despite the results, in practice the XLPE cable is chosen by over-dimensioning its limit. This has as result an increased price.

In LF AC instead, it is possible to use the standard XLPE AC cables. Some experiments have been carried out in order to evaluate the space charge face to the applied frequency [TITT00]. Results presented in Figure 1.12 show how the space charge phenomenon variates with the applied frequency value and, from at least 1 Hz, the effect is minimized.



Figure 1.12. Space charge behavior in a cable with: a) 10 kV DC, b) 15 kV DC.

It can be concluded that LF AC does not have the space charge problem as in HVDC systems. If supposing an LF AC system, this grid will only require AC protection relays in LF, is will use at least the same quantity of converters as in the DC case and, it will be presented in chapter 3 that an electrical transformer in the low frequency side can be avoided. It has been presented that when frequency is lower, the transmission distance increases and can be applied in offshore.

As it was presented in section 1.1.3, the output power in a three-phase AC system can be the same than the output power in DC system if cables are also the same and, when the power factor is limited to around 0.86, proving that a three-phase system could be more attractive solution for developments in offshore. Low frequency in AC will minimize the space charge effect in cables increasing lifetime, robustness and reducing implementation costs. Low frequency systems are then retained and studied in this thesis work.

1.3. Analytical Method proposed for Low Frequency Systems

Some of the analyses that will be presented further in this chapter are power analysis and frequency optimization in the system link. These analyses are therefore related to impedances, which are no longer a constant parameter. A cable model to be used in simulations and to estimate the needed parameters is therefore proposed.

First, parameters of a cable are calculated in a defined range of frequencies. All the obtained values will be fit into a line equation that depends on the applied frequency and, each equation is related to a specific cable. This approximation is useful for optimization purposes.

Classical AC Power System techniques of analysis are proposed to study power flow at reduced frequency. The cable was analysed using two-port theory [DaMR15, Welf98]; which is quite useful because of its simplicity to be applied with analysis techniques as Gauss-Seidel [GrSt94]. A three-phase PI-Line model with distributed parameters was used to model the transmission cable system. The model performance was tested and compared in simulation using the Wideband model, which is considered of great precision because it uses a wide range for frequency analysis [Gust04] and the wave propagation theory [WeWi73]. Cable parameters were also calculated at different frequencies using the wideband model [Emtp16]. Mutual impedances were introduced in the model.

1.3.1 Cable parameters

In the classical electrical power system, cable parameters are generally considered with constant impedance values and only temperature has an influence in the cable resistance value [Guer13]. As the frequency value cans variate, cable parameters are no longer constant an all variations must be taken in the model.

ABB XLPE submarine cables parameters were used into the system model [Abb11]. As an example, parameter of a 1400 mm² cables are presented next.

(a)								
Cross-	Diameter	Insulation	Diameter	Lead	Outer	Cable	Cable	
section	of	thickness	over	sheath	diameter	weight	weight	
of	conductor		insulation	thickness	of cable	(Aluminium)	(Copper)	
conductor								
[mm ²]	[mm]	[mm]	[mm]	[mm]	[mm]	[kg/m]	[kg/m]	
1400	44.4	23.0	93.8	3.1	128.6	26.1	46.2	
	(b)							
	I	/oltages [kV]	S	Sections [mm				
		220	500 620 80	0 1000 1900				
		275	900, 630, 8t	10, 1000, 1200)			
		330	620 800	1000 1900 1				
		400	050, 800,	1000, 1200, 1				

Table 1.2. (a) Geometrical specification of a 1400 mm² cable, parameters for a single-core with lead sheath. (b) Used cable sections and voltages.

A three-phase system was build using three single core cables. Were used the default cable distances proposed by the EMTP Wideband model [Emtp16], these are: cables are separated by one-meter each other; they are one-meter depth measured from the centre of the cable to the earth's surface into the sea. Variating these distance will only affects the mutual impedances of cables. More information about submarine cable modelling can be found in [Desc14].



Figure 1.13. Proposed cable positioning for simulations.

Cable characteristics are listed in Table 1.3. As the simulations were supposed at constant temperature of 20 °C ignoring temperature changes, this may introduce an error related to the conductor resistivity. The produced error is only linked to active power losses; however, distance limits are mainly associated to reactive power losses as presented in equation 1.2.

Table 1.3. Physic properties of the used cables at temperature of 20 °C.

Cooper	Aluminium	Conductor	Insulator	XLPE -	Insulator	Earth	Earth
Resistivity	Resistivity	Relative	Relative	Teflon	Loss	Resistivity	relative
of	of tape	Permeability	Permeability	Relative	Factor		permeability
conductor	shield			Permittivity			
$[\Omega \cdot m]$	$[\Omega \cdot m]$					[Ω·m]	
1.72e-8	1.92e-8	1.00	1.00	2.25 - 2.1	0.001	100	1.00

These specifications were introduced in EMTP-RV. Wideband model was selected to calculate the cable parameters in a large range of frequencies [Lean13], it was chosen because it presents the most accurate response in modelling.



Figure 1.14. Representation from 0 to 100 Hz of the self-cable parameters.

A wideband model allows to increase the precision in a system that uses high frequency elements as a switching converter, or to study transients when faults are produced. In this work instead is presented each element against a variable frequency that in steady state will not be bigger than 100 Hz. The electrical parameters presented in Figure 1.14 shown a linear response of resistance and impedances with the frequency changes. These values will therefore be used into power and optimization analysis of the system. The obtained results are 6x6 matrices:

$$[M] = \begin{bmatrix} Ms_{3x3} & Msg_{3x3} \\ Msg_{3x3} & Mg_{3x3} \end{bmatrix}$$
 1.19

From this matrix, four different 3x3 matrices can be obtained as showed in equation 1.19. Being *M* the matrix representation for impedances and admittances parameters; *Ms* is the 3x3 matrix for the conductor self and mutual components; *Msg* also a 3x3 matrix for components between the conductor and the screens; and, *Mg* is the matrix representation for the screen self and mutual components.

As example, an impedance and admittance matrix for 1400 mm² cable is presented next for 50 Hz of frequency. The impedance results are in $\Omega/_{km}$.



And the admittance matrix in $\frac{Siemens}{km}$:



These values will be treated using the symmetrical components theory (see appendix B.1).

1.3.2 Impedance Matrix and Symmetrical Components Transformation

Transforming equation 1.19 into symmetrical components is convenient for simplification of a three-phase system into three one-phase systems: positive, negative and zero sequences. In a balanced system, zero sequence voltages have zero value, and normally positive sequence is used to solve the entirely three phase system, in this particular case the negative sequence will be also zero. However, when the system is not balanced, negative and zero sequence will not be zero. Solving a system using symmetrical components in this case, will require to analyse three individual systems, which increases the difficulty [MeDT00].

In our case presented in the Figure 1.13, cable self-impedances and admittances are the same; this is not the case for the mutual impedances. Mutual of phases 1-2 are the same of phase 2-3, but these two are different from mutual 1-3 as presented in equation 1.20. As the difference between these values is not as important as it will be presented in the following equations, a singular mutual term can therefore be estimated that will introduce a minor error. This error exists because the simulated cable system was no transposed because it was supposed than a real submarine cable will not be transposed either. Therefore, in order to minimize the error between the different phases, a singular mutual term is obtained from the average value of the mutual impedance Zm.

$$Zm = \frac{Z_{1-2} + Z_{2-3} + Z_{1-3}}{3}$$
 1.22

The introduced error for the 50 Hz case is calculated as it follows:

$$e_{1} := \left(1 - \frac{\frac{200 \cdot |0.05 + 0.43| + 100 \cdot |0.05 + 0.39|}{3}}{100 \cdot |0.05 + 0.43|}\right) = 3.057\%$$

$$e_{2} := \left(1 - \frac{\frac{2 \cdot |0.05 + 0.43| + |0.05 + 0.39|}{3}}{|0.05 + 0.39|}\right) = -6.732\%$$

$$Error = \sqrt{\frac{2 \cdot (|e_{1}|)^{2} + (|e_{2}|)^{2}}{3}} = 4.619\%$$

17

As it can be seen in equations 1.20 and 1.21, because of the different distances between conductors, only the imaginary part is different at each case. The real part depends only on the self-dielectric. When compared with Figure 1.14, for lower frequencies a lower error from equation 1.23 is therefore expected. In the one hand, this error only affects the mutual parameters, which will not strongly affect results, and therefore it can be introduced in the main calculations. In the other hand, a system supposed transposed by physically exchanging the phase physical position several times, until the cable ends, will have identical mutual impedances, and this will keep balanced the power transmission.

Supposing steady state and a balanced system, the 6x6 matrix presented in equation 1.19 can be transformed into a 3x3 matrix using only the Ms_{3x3} terms, and while the system is balanced, the others matrices can be ignored. This is possible because the zero sequence voltages in a balanced system have zero values, introducing no currents into the neutral return. This procedure change when faults in the system are studied and depending on the type of fault, a strong value of zero sequence current can be present into the ground.

So, introducing the concept of balanced system into equation 1.19 and using the example presented in the equation 1.20, the final impedance matrix a 50 Hz is:

$$[Zline_{3x3}] = \begin{bmatrix} 0.07 + 0.68i & \dots & \dots \\ 0.05 + 0.43i & 0.07 + 0.68i & \dots \\ 0.05 + 0.39i & 0.05 + 0.43i & 0.07 + 0.68i \end{bmatrix}$$
1.24

Assuming then a transposed system and applying equation 1.22 into equation 1.24 the final line impedance is:

$$[Zl_{3x3}] = \begin{bmatrix} 0.07 + 0.68i & \dots & \dots \\ 0.05 + 0.417i & 0.07 + 0.68i & \dots \\ 0.05 + 0.417i & 0.05 + 0.417i & 0.07 + 0.68i \end{bmatrix}$$
1.25

The same procedure can be applied to the admittance matrix of equation 1.21. The symmetrical components are now introduced to the defined impedance matrix. Fortescue [Fort18] proposed this concept in 1918 to solve a three-phase unbalanced system using three separate balanced systems. Symmetrical components are still in use today for the analysis of modern power systems. The symmetrical components procedure is explained in annex B.1.

Values from the impedance matrix from equation 1.25 can now be converted into symmetrical components using equation B.4 of annexes, as it is presented next:

$$A^{-1} \cdot \begin{bmatrix} 0.07 + 0.68i & \dots & \dots & \dots \\ 0.05 + 0.417i & 0.07 + 0.68i & \dots \\ 0.05 + 0.417i & 0.05 + 0.417i & 0.07 + 0.68i \end{bmatrix} \cdot A$$

$$= \begin{bmatrix} 0.17 + 1.514i \\ 0.02 + 0.263i \\ 0.02 + 0.263i \end{bmatrix}$$

$$1.26$$

Equation 1.26 is an example of zero, positive and negative components resulting from a three-phase submarine cable system from Figure 1.13 using as example a 1400 mm² cable. Only positive sequence value is considered for power flow analysis.

As there are many values which depend on each cable section, voltage and frequency, a fitting tool can therefore be used in order to obtain one equation that define the evolution of parameters face to the frequency changes, for instance at 50 Hz from Figure 1.14. An equation will reduce the calculation time for simulation and will be more adapted to optimization purposes.

1.3.3 Impedance and Admittance values fitting

Instead of using the entire range of each calculated data of cables at each simulation, we propose instead the use of a preliminary step for fitting values into an equation. Fitting cable parameters is useful for optimization purposes. Figure 1.14 shows that the cable parameters depend on the frequency and, for impedances and resistances, a linear response was obtained into a defined range of frequency from 1 Hz to 100 Hz. In order to reduce the quantity of values related to each frequency, cable section and voltage, it was proposed to find a first order polynomial response $g(\varsigma)$ for each cable that depends on the listed parameters. For this purpose, it was used Curve Fitting Toolbox [Matl15].

For
$$0.01 \le \varsigma \le 100 \, Hz$$
, $\epsilon(\varsigma) = a \cdot \varsigma + b$ 1.27

Parameters a and b, are automatically calculated and depends on the used cable: voltage and section. The obtained cable equation will be used in the power calculations for the LF system.

1.4. Power flow analysis for submarine cable application

The system will be analysed using a two-port matrix equivalent and a voltage iteration method of Gauss-Seidel, that will be presented in section 1.4.1 and 1.4.2 respectively. It was considered a two nodes system as the one presented in Figure 1.15, one node is defined as the slack bus and the other as a Generation bus side.



Figure 1.15. Two buses system: BUS 1 as Slack, BUS 2 as Generation. Power factor (PF) obtained at each distance (d).

In the defined power system, three different distances can be identified: d_1 is the first limit found when the inductive losses of the system are the same than the capacitive losses, so only active power is obtained at this point; d_2 represent the maximal distance, limited when the required power factor is reached. Distance d_3 is a theoretical distance point where not active power can be transmitted in the system. Power factor will progressively decrease from d_1 until the cable ends because of the cable capacitance. In this work, minimal PF was selected to be 0.86. This value represents 30° of angle shift and a 50% of reactive power. This PF value was chosen in order to compare with a DC equivalent as demonstrated in 1.1.3. From d_2 and until d_3 , the PF will progressively be reduced until zero value.

In all analyses, it will be considered that the calculating distance d from BUS2 to BUS1 is equal or less than the maximum d_2 .

1.4.1 Two-Port method and π-model

Because of the physical structure of a submarine cable and a core surrounded by a dielectric and a very close neutral point, the cable susceptance is more important than its equivalent value for over-head line. To treat this kind of data, a π -model with distributed parameters of the line is generally used [GrSt94]. Impedance and admittance matrixes from equations 1.20 and 1.21 were presented with distributed parameters.

A power flow of the cable can be performed using several sections and different nominal voltages. The idea is to calculate limits as maximum current, maximum voltage drop and maximum distance at a given frequency of the system. As an example, the current behaviour in the output side of a transmission system against the system frequency value can be represented while keeping constant the voltage and the distance value of the system; this is showed in Figure 1.16. Current and voltage are in per unity (p.u.) of the cable nominal values [Abb11] (which change the power and voltage depending on the used cable), and the current was calculated using different cable sections. It was used a p.u. system of the cable in order to be able to represent the current behaviour while using the same scale.



Figure 1.16. Behavior of the real part of the current against frequency value. 300 km cable case.

For lower frequencies, the cable has lower losses, and current in p.u. does not differ from other cable sections. This is only an illustrative example where an important distance was used to observe the dropping power effect. In addition, the voltage was considered as constant in only one side of the cable-end (the slack bus) and, the voltage drop was not taken into account. Voltage drop will therefore reduce the practical results; this will be introduced later in this document. For this particular case of distance, the frequency of the system should not be bigger than 10 Hz in order to cope with the requirements presented in chapter 1.1.3. To achieve the presented results, a two nodes power system was modelled as illustrated in Figure 1.17. The impedance is defined as Z_{Cable} and the admittance as Y_{Cable} , resistance (R), reactance (X), conductance (G) and susceptance (B) are then presented.



Figure 1.17. Model of the power system and representation of the cable parameters.

The π -model per kilometre is introduced into a two-port matrix where voltages and currents are calculated at each kilometre. The proposed method is fully described in [Welf98]; equations for π -model would be:

$$R + jX = Z$$
, and $G + jB = Y$ 1.28

The voltages and current relationship in used system is:

$$\begin{bmatrix} \bar{V}_1\\ \bar{I}_1 \end{bmatrix} = \begin{bmatrix} K \end{bmatrix} \cdot \begin{bmatrix} \bar{V}_2\\ \bar{I}_2 \end{bmatrix} = \begin{bmatrix} K_{11} & K_{12}\\ K_{21} & K_{22} \end{bmatrix} \cdot \begin{bmatrix} \bar{V}_2\\ \bar{I}_2 \end{bmatrix}$$
(1.29)

And the two-ports K matrix from equation 1.29 is:

$$[K] = \begin{bmatrix} 1 + Z \cdot Y & -Z \\ -Y \cdot (1 + Z \cdot Y/_4) & 1 + Z \cdot Y/_2 \end{bmatrix}$$
 1.30

Equation 1.30 is related to a cable section with a length defined by the parameters unit, which in this case is per kilometres creating as consequence a distributed model. Multiplying several identical section matrices, will give as result the same section matrix with an exponent number. This number corresponds to the total length of the studied case. This length can be used in a more precise scale by scaling the parameter values too. The final equation would be:

$$\begin{bmatrix} \bar{V}_1\\ \bar{I}_1 \end{bmatrix} = \begin{bmatrix} K \end{bmatrix}^{length} \cdot \begin{bmatrix} \bar{V}_2\\ \bar{I}_2 \end{bmatrix}$$
 1.31

These equations are used to solve the Gauss-Seidel iteration method presented in the following section. The final PI-model of the cable for tests can be represented as following.



Figure 1.18. Model of the electrical cable and system: 1 for slack bus, and 2 for generation bus.

1.4.2 Resolution method of Gauss-Seidel

In the modelled power system, there are two known variables: the voltage of the Slack bus 1 represented in the Figure 1.17; and the available power from bus 2. This power depends instantly on the wind energy but only a constant value was taken in account.

Power variable is related to three other variables at the same time; voltage, current and angle of bus 2; the only way to calculate these variable is by performing iterations techniques with the bus 1. Gauss Seidel method makes it possible to find the required variables. By introducing the two-port matrix equation 1.29, the equations are:

$$\bar{V}_1 = K_{11} \cdot \bar{V}_2 + K_{12} \cdot \bar{I}_2 \tag{1.32}$$

$$\bar{I}_1 = K_{21} \cdot \bar{V}_2 + K_{22} \cdot \bar{I}_2 \tag{1.33}$$

$$S_2 = \bar{V}_2 \cdot \bar{I}_2^* \tag{1.34}$$

$$\bar{I}_2 = \frac{P_2 - jQ_2}{\bar{V}_2^*}$$
 1.35

Using equations 1.35 and 1.33 into the equation 1.32, the iteration Gauss-Seidel voltage equation is:

$$\bar{V}_2 = \frac{1}{K_{11}} \left(\bar{V}_1 - K_{12} \cdot \frac{P_2 - jQ_2}{\bar{V}_2^*} \right)$$
 1.36

Voltage V_2 is the newest calculated value and V_2^* is the precedent one.

$$\bar{V}_2^* - \bar{V}_2 = e$$
 1.37

Where e is the error between newest calculated voltage and its precedent chosen by the user to be a value around 1e-4 and indicates the iterations to stop.

1.4.3 Model validation

Two simulation methods were chosen in order to compare the results between them and the proposed algorithm. The first method consists of the physical distribution of the cable parameters; in a given number of PI-elements that, for practical simulation reasons was choose twenty distributed elements. A PI element is recommended to be applicable each 300 km in an over-head line [KuBL94]. It is supposed that in the worst supposed case system scenario of 500 km, 20 elements will be conformed of 25 km each, which is twelve times better than the recommended value and may be a good choice for submarine cables.

The second method is the Wide-Band model. The algorithm was built in Matlab. An exact PI-model representation per distance unit of the circuit from Figure 1.23 was used. This corresponds theoretically to the first used simulation method. Known variables are V1 and P2. The value of I2 is related to P2 and was calculated with the algorithm and, used in simulation for each method in order to fix the power generation. V1 is the nominal voltage of the cable. Values of V2 and I1 were measured and compared with those obtained with the algorithm.

The error is calculated in two steps; first, the error between amplitudes (X), where X symbolizes studied variables (V2 or I1). Second, the error between phase angles:

$$\Delta X = \frac{X_{algorithm} - X_{model}}{X_{algorithm}} \cdot 100\%$$
 1.38

$$\Delta Angle = \frac{Angle_{algorithm} - Angle_{model}}{360^{\circ}} \cdot 100$$
 1.39

All error results were reduced in a single table to compare the obtained results between simulations: exact-PI with the Wide-Band method. Then, the variation of the error is represented in Figure 1.19.

Table 1.4. Generalized errors of the used methods compared with the algorithm results.

Algorithm error when compared to models:	ΔV2 [%]	∆angV [%]	ΔI1 [%]	∆angI [%]
PI	e < 0.09	e < 0.35	e < 0.1	e < 0.35
WB	e < 17.8	e < 7.87	e < 25.21	e < 1.22

The presented errors in Table 1.4 correspond to the error of the proposed algorithm when it is compared to the PI model and to the WB model. A very small error when compared to the PI model indicates that the algorithm was correctly developed and it is like a PI model. However, when it is compared to the WB model, it indicates that some limitations must be made before using it.



Figure 1.19. Comparison error within simulation methods and developed algorithm: (a) $\Delta V2$ Wide-Band method, (b) Currents waveforms.

From Figure 1.19.a, the error of the algorithm increases with the distance value when compared to the wideband model. It is therefore important to set a maximum error of around 10 % in order to use the proposed PI model based method. The maximum distance is around 250 km, and all lower distances will have a lower error. For instance, around 200 km then the error would be of 5 %.

The error rises when compared to the PI model because the wideband method evaluates, with hyperbolic functions, the propagation constant of the line. This propagation effect became more important with the distance, so for longer distances another method must be used. All calculations made with the method are only estimations of the final system; therefore, it should be first verified in simulation by taking into account the estimated error. For selecting frequency, voltage or current in the system, an optimization approach is proposed, and in all cases, the resolution method stands on Gauss-Seidel technique presented before.

1.5. Optimization in the Low Frequency Power System

By using the proposed methodology, an optimization of the system can be applied. All estimated results contain an error that increases with the transmission distance as presented in Figure 1.19. As many cable sizes, voltages, distances and frequencies can be applied to a low frequency power system, an optimization toolbox to analyse them [CaPá10, Matl15] is therefore proposed.

For all cases, the objective is to find the minimum of a constrained nonlinear multivariable function. The vector \mathbf{x} is composed of three variables, x(1) is the frequency of the system in Hz, x(2) is the generation power of BUS2 in pu and x(3) is the system length in km. At each optimization, it is found a variable that will be tested in a defined range that will depend on the studied case.

1.5.1 Distance Optimization

In order to find the maximal allowed distance in a system, the minimal active power at BUS1 was limited to 0.86. Additionally, only active power is generated at BUS2 coming from wind resources. The optimization was repeated using different frequency values (x(2) which is defined as the changing variable). From equation 1.31, the function can be defined as:

$$\epsilon(x) \ge 0 \quad \forall x \in \mathbb{R}: \quad \begin{cases} 100 \ge x(1) \ge 0.1 \, Hz \\ 1 \ge x(2) \ge 0.1 \, p.u. \quad \rightarrow \quad changing \, variable \\ 1 \le x(3) < \infty \, km \\ 1 \ge P_1 \ge \sqrt{3}/2 \, p.u. \end{cases}$$

$$1.40$$

In a 1400 mm² system and 400 kV of nominal voltage, obtained results can be illustrated as it follows:



Figure 1.20. Results of the distance optimization in a two nodes system using a 1400 mm² cable.

The system length decreases exponentially when the frequency raises, when this value is less than 5 Hz, distances between 500 km and 2000 km can be reached. This result proves also that a DC system, which is at 0 Hz, has the higher reachable distance. From around 1 Hz, PF is at is maximal value, allowing 100% of active power. While increasing the frequency, the distance will be reduced, and at around 1.5 Hz approximatively, the PF decreases in order to cope with the maximal value of distance. The minimal constrained power factor value is found at around 5 Hz.

In the actually AC system with the lowest frequency $(15\frac{2}{3} Hz)$, the system length is around 200 km for the used voltage level. The distance can be improved by reducing the transmission voltage. For 50 or 60 Hz systems, the allowed distances are between 50 km and 100 km, as it was discussed before in chapter 1.1.1.

1.5.2 Frequency Optimization: Maximal Frequency of the System

The selection of the minimized or maximized parameters to find the best frequency value in a power system will depend on the used constraint: the first presented study will maximize the frequency over different distance values. Then, in subsection 1.5.3, the voltage drop is minimized at different frequencies. It has been supposed that only active power is generated from BUS2, thus there are no reactive power losses at the generation side ($S_2 = P_2$).

The first studied function is:

0

0

120

60

180

240

$$\max[\epsilon(x)] = \max[x(1)]$$
1.41
$$\begin{cases} 60 \ge x(1) \ge 0.01 \, Hz \\ 1 \ge x(2) \ge 0.1 \, p.u. \\ 1 \le x(3) < \infty \, km \quad \rightarrow \ changing \ variable \\ 1 \ge P_1 \ge \sqrt{3}/2 \, p.u. \end{cases}$$
1.41
$$\begin{cases} 0 \ge x(1) \ge 0.01 \, Hz \\ 1 \ge x(2) \ge 0.1 \, p.u. \\ 1 \le x(3) < \infty \, km \quad \rightarrow \ changing \ variable \\ 1 \ge P_1 \ge \sqrt{3}/2 \, p.u. \end{cases}$$

Figure 1.21. Frequency maximization at each distance value by limiting received power.

Length [km]

300

400

500

600

700

800

900

1000

This study identifies the maximum frequency value that can be used in a particular system length. As it can be seen, the frequency was limited between 60 and 0 Hz, and in this particular case the maximum frequency value scope the higher frequency for around 50 km. The frequency decreases and it reaches 204 km for $15\frac{2}{3}$ Hz, and 250 km for 10 Hz.

In order to cope with the required constraints, the maximum allowed frequency value is being reduced from the same moment that the generated power reaches its minimum. From this test, it can be appreciated that depending on the total distance of a given cable, a maximum frequency value can be set for the system (for instance at 60 km, the maximum frequency would be around 50 Hz). This is when supposing a variable frequency system; nevertheless, this is out of the sight of this work.

1.5.3 Voltage drop minimization

In this study, voltage drop is defined from Figure 1.17 as the difference between the slack bus voltage (V1) with the generation bus voltage (V2). A lower voltage drop is required for keeping operations in a system, where normally there are some upper and lower voltage drop limits. Additionally, because of the capacitive effect of the line, voltages tend to increase in the evaluated system, therefore this voltage drop tend to increase and be positive, as it will be presented next. It is evaluated with the following equation 1.42.

$$\Delta V = \frac{\left|\bar{V}_{1}\right| - \left|\bar{V}_{2}\right|}{\left|\bar{V}_{1}\right|} \cdot 100\%$$
 1.42

From the optimization point of view, the idea is to minimize voltage drop as it is presented in equation 1.43 below. The optimization function is evaluated at different values of distances.

$$\min[\epsilon(x)] = \min[|\Delta V|]$$
 1.43

$$\epsilon(x) \ge 0 \quad \forall x \in \mathbb{R}:$$

$$\begin{cases}
60 \ge x(1) \ge 15 \\
1 \ge P_1 \ge \sqrt{3}/2 \\
V_1 = 1 \\
1 \ge x(2) \ge \sqrt{3}/2 \\
1 \le x(3) < \infty \quad \rightarrow \quad changing \ variable
\end{cases}$$

Results can now be illustrated as following.

٢



Figure 1.22. Voltage drop optimization at different distances values

A bigger voltage drop is appreciated with a bigger distance length as it can be appreciated in Figure 1.22. It was found that the minimal voltage drop is obtained also with the lower frequency value that in this case was constrained to be 15 Hz. Finally, the output apparent power S1 reaches 1 p.u. at 204 km, and the voltage drop is 4 %, which is acceptable for OPS.

As it was presented in this section, a lower frequency value will result in a more optimized system and, in order to propose realistic results, it will be considered 15 Hz as the lowest limit of frequency for calculations in an OPS.

1.6. Calculations in a OPS

Optimization results were used as criteria to build a power flow algorithm and to calculate unknown variables as distances, power or frequencies in a given system. First, the maximum distance is calculated in a system with two buses Depending on the made test, different cable sections and voltages could be used at constant frequency of 15 Hz. The maximum distance is calculated by using the same method described in subsection 1.4.1. Then, the resolution algorithm is presented next:



Figure 1.23. Algorithm for calculation of the maximal transmission distance.

1.6.1 Maximal Distance Calculation

In this section is presented a power analysis were maximal distances are calculated using the proposed algorithm of Figure 1.23. A second part of analysis will be followed by calculations of the allowed frequency range in a given system. The first result will consider only one cable. It was used as example a 400 kV system with 1400 mm² of cross-sectional cable surface. The nominal power of this system is 830 MVA. In the following figure it is therefore traced the power evolution versus the distance increase.



Figure 1.24. Power behavior against distance variations. Minimal PF at 204 km.

From Figure 1.24 it can be said that the nominal power of the system will be defined by the total length of the system at one specific PF. This power is different from one cable section to another and the PF variations will directly affect the total allowed power. In the presented example, the found maximal distance was 204 km for a FP1 of 0.87 at 15 Hz.

Supposing now that the actually length of the system is lower than the indicated limit, this means that the system frequency could be higher in order to keep the same FP conditions. The advantages of rising frequency are related to a shorter time for fault clearing in a system. Variation of frequency is related to control the reactive power amount of the system and this is related to control the voltage drop levels. This idea will be explained in the next section.

In order to make a choice of voltage, distance and frequency, multiple cable sections and nominal voltages have to be tested and then choose the convenient one: for long distance applications the idea is to rise voltage at its maximum possible and minimize frequency, this will increase the transmission distance and minimize losses. Voltages drop and power losses comparison, are presented in Figure 1.25 from all cable sections and nominal voltages of Table 1.2.b.



Figure 1.25.Algorithm results of: voltage drop (a), and active power losses (b).

The maximum distance of 409 km is related to the minimum allowed PF; nevertheless, it was obtained at the lower nominal voltage that carries the higher voltage drops of 25 %. Supposing 10 % of voltage drop as the maximum accepted then the maximal distance will be around 250 km, which is an expected distance value for offshore application. Voltages at this point are between 275 and 330 kV, and active power losses around 6 % for a rated value between 362 and 716 MVA.

The power behaviour can be applied using a large range of frequency in order to observe the evolution in distance and voltage drop effect. First, variations in the active power of the load is presented:



Figure 1.26. Active power representation at constant generated power.

Then, the voltage drop for each case can be observed:



Figure 1.27. Voltage Drop representation at constant generated power.

Depending on the system length, the use of a given value of frequency could optimize active or reactive power and can change the voltage drop of the system. Being able to change the frequency on an electrical system will increase the control possibilities that could improve the networks of the future. Variations of frequency were tested with a submarine cable; this is presented in the following section.

1.6.2 Frequency variation range

Presented results were calculated using a nominal frequency of 15 Hz. Nevertheless, higher frequency can be useful for many reasons. One of them is the more frequent zero crossing, and this allow a quicker fault clearing than in the 15 Hz case. In addition, with a higher frequency, a lower size transformer can be used (which will be presented in chapter 2). It is therefore recommended to study the maximal frequency limit in a system where the distance of the cable is known. In this way, the range of frequency in a distance will depend in its total length; for low lengths higher frequencies could be applied as it was showed in Figure 1.21.

The frequency range is calculated using the previous obtained value of maximal distance (D_{max}). The total length of D_{max} , which depends on each used cable, is divided using the same proportion for all cable cases as presented in Table 1.5.

Then, frequency is progressively raised at each element until PF1 reaches the 0.87 limit. The proposed algorithm is presented below.



Figure 1.28. Algorithm for calculation of the maximal frequency.

As in the previous section, an example for a 400 kV cable case is presented next, here are showed all frequency values that can actually be used for each proportion of D_{max} .



Figure 1.29. Allowed frequency range in submarine cables of different lengths.

The cable capacitive effect limits the maximal frequency that can be used in the system. For a 122 km of cable length, the highest applicable frequency is 28 Hz. Nevertheless, higher frequencies values than 60 Hz are allowed in a 41 km cable length. As depicted in Figure 1.29, this is actually the average distance used today for AC offshore interconnections as presented in section 1.1.1. With the developed algorithm, we obtained the same frequency range, no matters the cable section or the applied voltage; this is because all cables present a linear response as suggested in subsection 1.3.3. Results are presented in the following table. D_{max} is the obtained maximal distance value for each case of voltage and section and represented in Figure 1.25.a.

Table 1.5. Frequency range for all cable cases.

>60
43
28
20

From Table 1.5, the maximal allowed value of frequency depends only on D_{max} , no matters the applied voltage. This is an expected result because D_{max} was calculated by taking into account the maximal allowed reactive power of the system, therefore it is proportional to minimal PF. These results introduce the following hypothesis: it is possible to variate the value of frequency in a system in order to optimize the power flow and reduce losses.

1.7. Conclusions in LFAC OPS

In this chapter, it was suggested the use of the low frequency for power transmission in OPS. To examine the feasibility of LF, an algorithm for maximal transmission distance calculations was developed using [Matl15]. Several cable sections and voltages were used for the precision tests and its performance was tested in simulations when compared with two models, the Wideband and the PI model of the line or cable of EMTP.

The use of the low frequency will allow a transmission system with distances that can achieve 250 km and even 300 km, which are adapted to OPS. The maximal reachable distance is related to the cable parameters and the applied voltage level. A tool for sensitivity analysis is proposed. A frequency range was also identified for a given cable length, which makes it possible to adapt its value in order to manage the power flows in the system. With this, an optimization method can be used in submarine networks to adapt the system frequency and increase the power exchanges.

The retained LF value will be between 15 Hz and 16 2/3 Hz, because some railway networks are still operative at this frequency, proving the feasibility of the technology. Nevertheless, its application to a wind generator will depend on its construction as it is presented in appendix A.1.1; it will be supposed that permanent magnet generators were used and that LF will not impose any problem for those generators.

Only a point-to-point system was tested in this section, nevertheless the use of the LF will allow the insertion of multi-terminal networks as in the AC case. A LF system will therefore require equipment adapted at this lower frequency, this will be discussed in the following chapter.

Chapter 2. Low Frequency in Offshore Power Systems

The standard frequency of 16 2/3 Hz corresponds to the first three-phase catenary supply interconnecting two cities in 1905. It related the cities of Brig (Switzerland) and Iselle (Italy) [Lang10]. From 1910 until nowadays, Germany, Austria and Sweden also adopted this frequency for their railway electrical system.

The application of LF to an overhead line for power transmission issues started between 1994 and 1996, proposed by Wang (et al.) [WaWa96a]. Articles have focused on its feasibility from many points of view as generation, conversion and transmission. Studies in low frequency prove that this kind of systems can be applied without technological challenges, which is an advantage faced to HVDC systems [FuMa00].

One of the main drawbacks in low frequency systems is the oversizing of the LF transformer when compared to 50 Hz one. The design of the transformer must be optimized for LF application in order to minimize its size [JuKr77]. A bigger transformer will raise its implementation price and maintenance costs, which are no attractive for OPS.

Converters are needed if supposing power generation at the conventional 50 or 60 Hz at the offshore side, as it has been supposed for researches in HVDC [Berg15, BKHV07, Sark12]. This means that a transformer would be necessary in order to step-up the line voltage for long-distance transmission of power.

In HVDC, some of the research proposals are based on the introduction of a DC bus at the offshore side [RiSa16, ZFSG16]. Generators would be connected in series to raise the voltage without needing a transformer. This solution does not require an offshore transformer and can be applied to LF too if supposing LF generation and series connection of generators.

In any possible case that might be applied, a converter capable to manage frequencies between 50 Hz or 60 Hz and LF will be needed in order to interconnect with the main electrical system of any country. Many converters are suitable for this task, and references have being published with cycloconverters [LiXY11], matrix converters [RuMO16] and back-to-back converters [WrUm09]. These converters shave their own drawbacks as poor quality of the signal, which is related to high losses and the need of big filter devices. This increases the volume and size of the converter, maintenance and implementation costs that are not wanted for offshore application. The supplementary conversion stages are also taken into account, which would not be necessary when conversion through DC stage is avoided.

All converters are compared with the retained solution for HVDC, that is the modular multilevel converter [CWBW11], proving that this technology will offer the best results for waveforms quality, efficiency and robustness making it the best solution for OPS.

2.1 Low Frequency Systems: Advantages and drawbacks

Main improvements in the LF systems have been performed within the railway electrification developments in different European countries. Generation in low frequency is currently made using two technologies: motor-generators, which convert energy from one frequency to another using two rotating machines; and converters, which are based on power electronics. Motor generators [StOm96] are not suitable for OPS because of the larger size and cost and losses when compared to power electronics. Moreover, modern power plants for railway systems working in LF are all designed using static converters.

2.1.1 Impact of the railway as designer on the LF systems and components

Germany, Austria, Switzerland, Norway and Sweden standardized their railway electrical system in one-phase working at 15 kV and 16 2/3 Hz as showed in Figure 2.1. However, Germany, Austria and Switzerland passed to 16.7 Hz of frequency in 1995, which is still close from 16 2/3 Hz. This solved overheating problems with their motorgenerators [LiHe02]. A railway system is structured the same way as any other electric system; they need generation, transformation, transmission, protection and distribution of loads. The last one is more difficult to manage in railways because the electric loads are of great capacity. These loads are in constant movement with changing speed and can even regenerate power during braking [Wagn89]. This creates important voltages drop in the system and supplementary harmonics.



Figure 2.1. European railway electrification [CDSS14].

The combination of frequency converters and asynchronous motors proved to be particularly advantageous. It allowed a largely drive concept that was independent of the type of power supplied on the contact wire. This introduced some standardization and made easier to build locomotives, which were able to work under different voltages and frequencies for international trains. Some of the advantages of using low frequency in railway systems presented in [Bhar96] are listed next:
a. Less substation installations

Because of the reduction of the line inductance, thus increasing the transmission distance while keeping the same voltage drop when compared to 50 Hz or 60 Hz systems.

b. Parallel operation of catenary from adjacent substation

Paralleling the seconders will enable the power to be drawn from two or more substations, thus decreasing the voltage drop further in the catenary and also, distributing the load in a multi-terminal system. With modern power electronics and an advanced adjustable-speed-drive control system, the voltage at a substation can be controlled for optimal operation.

c. Reduced harmonics into the system

The frequency conversion equipment filters on the system would further reduce the harmonics generated from the locomotives and reduce the harmonics entering into the utility system. The German electric system is a meshed network as it was presented in Figure 2.2.



Figure 2.2. Catenaries in the railway electric system of Germany [Deut10].

The railway working in LF was therefore an important development area for converters working to adjust frequency in locomotives [BOHS10]. In the past, locomotives were equipped with mercury-arc rectifiers [Prin26] in order to work with the reduced frequency. Then in 1990 were implemented the first GTO-converters in locomotives. This proves also the technological advances made in the low frequency system faced to the needed improvement in their service.

This kind of systems requires the development of the electrical protection system capable to work in a LF system. Some examples will be introduced further in this work.

2.1.2 Existing voltage and power levels in the AC LF systems

An AC OPS could be built as the railway network: in Germany, over 19,000 km of the railway network are electrified at 110 kV for a traction network that has a total length of 7745 km. The importance of high-voltage is to ensure an uninterrupted supply and high network availability. A higher voltage nevertheless should be implemented for a power transmission system in order to reduce losses.

It is also proved that this LF frequency network could bet in a multi-terminal configuration as the 110 kV, 16.7 Hz grid, which is connected using 180 substations with the catenary line of the railway system. Some examples of substation are presented in Table 2.1. The installed power is approximatively 3000 MW.

Location	Total/unit AC apparent power at 16.7 Hz [MVA]	Total/unit AC apparent power at 50 Hz [MVA]	Total/unit active power [MW]	Three phase system 50 Hz [3AC kV]	Railway voltage 16,7 Hz [2AC kV]	Number of units
Düsseldorf	18.9 / 18.9	17.7 / 17.7	15 / 15	20	110	1
Karlsfeld	132 / 66	112 / 56	100 / 50	110	110	2
Limburg	151.2 / 18.9	141.6 / 17.7	120 / 15	20	110	8
Wolkramshausen	37.8 / 18.9	35.6 / 17.8	30 / 15	110	15	2
Doberlug-Kirchhain	37.8 / 18.9	35.6 / 17.8	30 / 15	110	15	2
Genin	37.8 / 18.9	35.6 / 17.8	30 / 15	110	15	2
Neckarwestheim	186 / 93	172 / 86	150 / 75	400	110	2

Table 2.1. Examples of ABB converter stations in operation in Germany [Abb10].

In sparse traffic regions in the northeast of Germany, it is not employed a transmission grid, but uses power converters to supply the 15 kV catenary directly. Modern static converters feature operation modes in both, the frequency of the public grid, or synchronized with the central 110 kV 16.7 Hz grid. The forced-outage availability of the current generation of static power converters exceeds 98%, and overall efficiency is higher than 97%. Low losses even during partial load operations allow a high flexibility using the converters in load sharing mode and for and balancing the grid.

The voltage level of 132 kV is the maximum operative value for 16.7 Hz electrical systems. However, offshore applications are expected with voltage levels around 400 kV. Therefore, some researches and tests must be still performed in order to adapt the LF system components into OPS: protection relays and converters must be adapted to a higher voltage. Nevertheless, this is out the sight of this work.

2.1.3 Existing protection relays working in LF

In LF systems, the needed extinction time is longer than in a 50 Hz system, requiring three times more in 50/3 Hz than in 50 Hz as showed in the following Figure 2.3:



Figure 2.3. Comparison between 50 Hz and 50/3 Hz waveforms.

However, compared to the electrical network, the catenary has to frequently withstand short circuits caused by vegetation and the mechanical stress from pantographs, in consequence some commercial protection relays has been developed for this use.

Switchgear and circuit breakers plays an important role in substations by ensuring safe and flexible operation [WeFa07]. Ideally, utilities or operators look for switchgear that is reliable during its lifetime, flexible in configuration, easy to operate and requires minimum maintenance. Suitable solutions in any protection system are related to a small footprint, which is reliable and cost effective, and guarantee high safety and availability. In addition, they minimize the downtime needed for installation. One example is the ABB's plug and switch system *PASS* family [Abb10]. Breakers are specially designed to be used for railways where they have to extinguish longer burning arcs in 16.7 Hz networks. It is filled with SF6 self-blast interrupting chamber [LSYP85]. It is able to insulate the high-voltage live parts of the contact assembly from the housing. Bushings connect the interrupter chamber with the high-voltage terminals.

The 110 kV PASS circuit breaker went into operation at the Datteln [WrUm09] converter station. An example of this kind of protection relays is presented in Figure 2.4. It is easy to transport and can be simply replaced in case of failure. Other versions uses a complete air insulated switchgear, and each module requires 60 % less space than a conventional one, which is attractive for OPS. These protection relays are therefore operational for 110 kV. In the case of higher voltages as in the offshore case, the circuit breaker would be not so different and these protection relays are certainly less expensive and less voluminous than the ones used in HVDC OPS [Fran11].



Figure 2.4. ABB 110 kV PASS M0-16.7 Hz switchgear for railways [Abb10].

2.1.4 The electrical transformer and the applied frequency

The relationship between size and frequency of a transformer is given by its universal EMF equation [Pans99]:

$$\mathbf{V} = 4.44 \cdot f \cdot B\mathbf{x} \cdot N \cdot A \tag{2.1}$$

Where the magnetic flux density (Bx) is almost not affected by variations of the frequency (f). Only the turns number (N) and the core cross-sectional area (A) can be modified. When the nominal f is lower in a system, a bigger A or more N must be used in other to keep the same nominal voltage (V). A higher dimension transformer is therefore not suitable for offshore application where it requires the transport into the sea and construction of a bigger and costly platform.

A study design of a classical transformers using mathematical optimization for a given set of constraints is explored in [JuKr77]. In this article, it is also made a comparative study when using a range of frequencies from 25 Hz to 100 kHz, which will give an idea of the transformer behaviour when a low frequency is applied.

In the following Figure 2.5 is presented the compromise between the transformer efficiency with a given power capability and the magnetic flux and the current density. The power capability can be defined as the nominal power of the transformer at a given frequency value, generally 50 Hz or 60 Hz.



Figure 2.5. Maximum power capability in a range of operational frequencies in a transformer when comparing different core structures [JuKr77].

The required efficiency of a transformer may introduce some constraints on its design that may be in opposition to some wanted optimums. For instance, when the efficiency must be larger than its nominal value, some compromise must be made. For instance, in order to keep the same power capability the transformer design is physically larger. If the design cannot be changed, then the power capability is less than its nominal value. A transformer can be designed for minimum losses at a given load power, which is a function of the operating frequency for any laminated core structure.

An optimization of power capability in a transformer where the design parameters were kept the same for each studied case is presented in Figure 2.5. It is showed how the power capability of a transformer is being reduced with the applied frequency. For all studied cases, the difference between 50 Hz and 25 Hz has an exponential behaviour. Nevertheless, for low frequency, when the frequency is reduced in two, the power is also approximatively reduced by a half. An approximation of transformer efficiency is presented in Figure 2.6. It is showed how the construction design can improve the transformer efficiency. However, it is still low when compared with the 50 or 60 Hz case.



Figure 2.6. Efficiency approximation of a transformer faced to variations in the frequency [JuKr77].

2.1.5 Transformer comparison with LF 16 2/3 Hz case

When directly comparing a 50 Hz transformer with is 16 2/3 Hz equivalent, the resulting transformer will be either larger or wider. As example, it is presented the following comparison:



Figure 2.7. A 250 MW transformer dimensions compared at: (a) 50 Hz, (b) at 16 2/3 Hz wide design, (c) at 16 2/3 Hz tall design [WTRY15].

By following equation 2.1, in Figure 2.7.b only the magnetic core cross sectional area was changed, leaving the turns number equal to the 50 Hz design. It produces an augmentation of 188.5 % of weight and 155.1 % on the volume of the transformer. In Figure 2.7.c it was only modified the turns number of the transformer, the result was an augmentation of 77 % in the weight and 75.5 % in the volume. Only these two extreme cases were tested in [WTRY15], it is evident that an optimization can be applied in order to find the best combination of the parameters. This subject is outside the sight of this thesis work.

It has been proved that a transformer, working in a lower frequency must have a bigger design in order to keep the same efficiency, in addition to work at the same power level than its equivalent at 50 Hz. These are unwanted characteristics for application in an offshore power network and a solution is presented with the developed converter in section 3.

2.2. Proposed System Configurations for Low Frequency Systems

In order to propose a solution to interconnect with far location of hydropower electric resources in China, with a range of transmission distances coming from 1000 to 2500 km, [WaWa96a] proposed the implementation of a low frequency transmission system. It is mainly because DC transmission system was too expensive for widely implementation and the highest voltage level of AC 50 Hz transmission was 550 kV, which does not satisfy power transmission over 1000 km.

The Low Frequency Transmission System (LFTS) can provide an efficient approach to solve the distance limitation of AC 50 Hz. LF (for instance 50/3) could reduce the reactance value of the AC transmission system, hence can multiply the transmission capacity and remarkably improve the operating performances with the only drawback of having a bigger transformer.

2.2.1 Overhead lines in LF

In 1996, [WaWa96a] proposed the following configuration for overhead line in a transmission system interconnecting two 50 Hz systems through a 50/3 Hz system:



Figure 2.8. 50 Hz systems connected on LF [WaWa96a].

The simulated system is of 1200 km length at 550 kV with a nominal power of 1700 MW. Nevertheless, the designed converters at that time were magnetic frequency changers [BiLa76]. These converters are not introduced in this work because of their low efficiency (5% of losses in the proposed overhead line system in Figure 2.8), especially when compared with newer technologies as it will be presented in section 2.3.

The same system from Figure 2.8 was introduced again in 2005 [XiCZ05], but this time it was chosen a cycloconverter as a frequency changer. A particular configuration is presented in Figure 2.9; it was chosen a generation directly at 50/3 Hz.



Figure 2.9. LF overhead line used for experiments [XiCZ05].

Their result probed that the transmission capability can be increased by 2.5 times when compared to AC 50 Hz. Efficiency of the proposed converter was 96.5 %, which still presents high losses. However, cycloconverters are still in use today for application in power machines as a frequency changer; this is why it will be described using more details in section 2.3.1.

Generating directly in LF has the advantage of elimination of the high power converter located at the offshore side; this of course will have some problems in the transformer size as presented in section 2.1.4. In order to compare the best solution between choosing a converter and choosing an electrical transformer, a technicaleconomic analysis should be performed; however, this is out of sight of this work.

A technical analysis supposing generation in LF in the system is presented in the following subsection.

2.2.2 Direct generation in LF

A direct LF generation of power would have the following topology:



Figure 2.10. AC OPS with direct LF generation.

As presented in appendix A.1.2, the most suitable solution in generators for offshore application is the Permanent Magnet Generator (PMG), which is connected to a full power converter in order to ensure a constant frequency at the output side. This frequency could be set directly to a low value, and therefore the offshore system would be in LF, avoiding the needed converter in offshore platforms.

Nevertheless, a lower frequency will require a bigger transformer and motors [Bhar96], increasing the weight and volume of locomotives. This will partially be the case for offshore wind turbines, because only LF transformers will be larger, and electric generators instead will remain the same.

A way to eliminate the needed transformer in the offshore system may be applied too, this solution was proposed for the HVDC case as explained next [GaMo12, HBMM13]. Instead of using generation at 50 Hz at the wind turbines output, the idea was to generate in high frequency and to use a transformer that would be placed inside each wind turbine.

The high frequency will allow the use of smaller transformers as it was explained in subsection 2.1.5, which are easier to build and to transport into offshore. At the transformer output, another AC-DC converter is used in order to create a DC link directly from the generation side. By interconnecting generators in series, the individual voltages will then be added and a large offshore transformer will be avoided. An example is presented in Figure 2.11.

By replacing the AC-DC converter for an AC-ACLF topology, this solution could therefore be applied to the LF system as well and therefore eliminating both the large offshore converter and transformer.

This is a not so attractive solution tough; the series-connection of wind turbines will reduce the system robustness because, losing one of them for faulty reasons for instance, will therefore create an open circuit or will step-down the total voltage needed for transmission in an interconnected system. This is the main reason why this configuration was not held for the presented thesis work.



Figure 2.11. Series connection of wind turbines in (a), and proposed configuration for each turbine in (b) [GaMo12].

They proposed a matrix converter because it can convert AC to AC directly reducing the size and the conversion stages without passing through a capacitor link. This link normally requires large components and produces additional losses. Matrix converters are also quite efficient for a long range of frequencies (even 400 Hz for aircraft operations [Lope10]). More details about matrix converters will be presented in subsection 2.3.2.

2.2.3 Generation and LF conversion

The LF system held in this work should be at least an equivalent from the supposed HVDC solution presented in Figure 1.9.b. This means that the generation is made at 50 Hz, the use of a step-up transformer at the offshore side, a conversion station and a transmission system without a transformer. The transmission distance is also limited in order to transmit the same power amount in both cases; this was presented in subsection 1.1.3.

An example of a multi-terminal offshore low frequency power system is presented in Figure 2.12. This solution was proposed the first time for offshore application in reference [FuMa00], they proposed the use of cycloconverters as a frequency changer, and they proposed some benefits in LF such as enhancing of the transmission distance and reducing the charge space in XLPE cable, which was presented in subsection 1.4.2.



Figure 2.12. Generation and LF transmission system allowing a multi-terminal connection.

In the last decade, LF system were highly studied presenting researches over several subjects as optimization of the transmission distance, experiments with several cable types, optimization of the power control, increase of the line voltage, protections devices, etc. In the selected references [CKAO15, FiBE12b, LiXY11, QYXA09a, RuMO15, WaWM15, ZhWW10], authors agree that the AC-AC converter is the main limitation for offshore LF developments. Solutions in the LF are normally proposed using known topologies of converters as cycloconverters, matrix converters and back-to-back converters. They will be described in the following section.

2.3. Converter Proposals for LF Systems

The offshore wind farm connection with a LF system requires a frequency converter or a variable frequency transformer that will be chosen depending of three main axes: use, low-cost and efficiency [QYXA09b]. Two main kinds of converters can be compared, this depend on the implementation or not of a middle DC link (used as a middle field to pass from one frequency to another). The use of a DC link has some advantages and drawbacks, presented as it follows [Chat97]:

DC link converter advantages:

- Output frequency can be varied from zero to rated value. Frequency variation time is quite low due to the use of transistors.
- Simpler control implementation.
- It still has a high input power factor when a diode rectifier is used in the first stage.

DC link converter drawbacks

- Conversion is made in three stages: rectifier storage inverter.
- Forced commutation required for the inverter.
- Regeneration feature is difficult because must include a battery for example, in order to recharge the DC link of the converter.

A converter without a DC-link is considered an AC-direct converter. An example of this kind of converter is the Cycle converter, presented in the following subsection.

2.3.1 Cycloconverters

Cycloconverters were conceived and patented by Hazeltine in 1926. The principle is to use switching arrangement to construct an alternating lower frequency voltage signal from a higher frequency multiphase ac supply. Cycloconverters are today a practical proposition for large power application [ChJA13, QYXA09a]. They can be used for instance in gearless mill drive, centrifugal pump, compressors, electric traction, rolling mills, variable-speed constant frequency systems, static Scherbius Drives, mine winders, ship propellers.

Cycloconverters can be connected in multiple configurations as presented next. An increase of the converter size is more or less proportional to the quantity of phases of its design. Step-up frequency in a cycloconverter can be achieved by implementing high frequency switching devices as Insulated-Gate Bipolar Transistor (IGBT) instead of thyristors for instance.

One-phase to one-phase cycloconverters

This converter consists of a back-to-back connection of two full-wave rectifier circuits. In Figure 2.13, the operating waveforms are shown when a resistive load is connected at the output. For this kind of converter, \propto_P is the positive firing angle and \propto_N is the negative one. Considering as example, to obtain 1/4 of the input frequency at the output side, the first two input cycles must supply positive current to the load; therefore, the load sees four positive half cycles and then four negative half cycles.



Figure 2.13. One-phase to one-phase cycle converter waveform output: a) input signal; b) ¼ rectified signal at firing angle of 60°; d) ¼ rectified signal at varying firing angle.

A square \propto operation gives a discontinuous waveform as showed in Figure 2.13.b. The average of this resulting waveform is of the square type. The \propto modulation could be improved in order to reproduce a more sinusoidal like output that will therefore reduce the harmonic content.

Three-phase to one-phase cycloconverters

This kind of conversion can be achieved using either the half-wave dual converter or the dual-bridge converter.



Figure 2.14. Three-phase to one-phase cycloconverters in dual-bridge (a), and half-wave configuration (b).

The difference between both configurations presented in Figure 2.14 is the resulting harmonics, where for the half-wave only six devices per converters are needed, reducing implementation costs but raising harmonic content because it is controlled with only six pulses. For the dual bridge instead, the number of pulses raises to twelve and there is a supplementary circulating current control, which can improve the output signal quality.

Three-phase to three-phase cycloconverters

Normally, this is the configuration used for low frequency application; the converter topology is presented in the following figure:



Figure 2.15. Three-phase bridge cycloconverter for offshore LF system application [NaFM02].

A three-phase bridge cycloconverter is also called a 6-pulse or a 36-thyristor cycloconverter. A cycloconverter can supply lagging, leading or unity power factor loads while its input is always lagging. The main problem of these converters is the poor signal quality that requires big filtering devices in order to reduce the harmonics quantity.

In Figure 2.16 is presented an example of results from reference [NaFM02], where the output voltages and currents without filtering and the generated power are showed. It can be appreciated the distortion in waveforms.



Figure 2.16. Obtained results using a three-phase cycloconverter [NaFM02].

There is also a half-wave cycloconverter, but generated waveforms will have even worst quality form. In addition, as it can be appreciated in Figure 2.16, three threephase transformers are needed instead of only one, rising implementation cost for offshore application. For all the presented reasons, cycloconverters are discarded for analysis in this work.

2.3.2 Matrix Converters

Earliest publications of the Matrix converters (MC) begin at the year 1959 describing the first MC topology [FrKo02]: a half-bridge MC using bipolar junction transistors. The actual development of the modern high-frequency modulated MC topology started with the work of Venturini and Alesina in 1980 [AlVe81].

Two main MC topologies are divided in two: Conventional or direct Matrix Converter (CMC) and Indirect MC (IMC). The difference between both technologies lies on the fact that IMC creates a fictitious DC link [RoJ83] without any capacitive storage but decouples the input current control and the output voltage control. A complete comparison between CMC and IMC has been made by Kolar in [FrKo10] and an important contribution using fast switching IGBT is presented in [WeLL10]. They explained that matrix converters have as advantages the reduction of size, lightweight, durability and high efficiency power supplies, compared to the use of a Pulse Width Modulator (PWM) rectifier and an inverter system.

The difference between both topologies remain in the needed quantity of switching devices, whereas in the ICM, components are reduced from almost a half of the number of the CMC. Both topologies schematics can be compared using simplified and a fully topological representation as it follows:



Figure 2.17. Full topological representation for CMC in (a) and IMC in (b). CF represents the filter capacitors.



Figure 2.18. Simplified representation for CMC in (a) and IMC in (b).

To avoid short-circuiting and current interruption in the input voltage path, a multistep scheme must be applied. The most effective method to prevent commutation failures is to detect both the polarity of the input voltages and the direction of the output currents [KaIt07]. The same commutation schemes known from CMC can be applied to IMC. Modulation methods were mainly triggered by Venturini's work in 1981 [AlVe81]in which are presented PWM and Space Vector Modulation (SVM) [MTBR13]. Even if many improvements for MC appears every year [ChEn03], [ZWCB08]. Still, this converter presents some drawbacks that exclude it from offshore applications. These are listed below:

- Due to the absence of intermediate energy storage, the operation faced to unbalanced input voltages is more critical when compared to a dc-link converter systems;
- MC operational safety risk in case of failure;
- MC are developed for low and medium voltage applications from 3.3 kV and 6 kV with maximal rated power in industrial application of 5 MW;
- A voltage transfer ratio of 86% (for sinusoidal modulation);
- Complex commutation and restricted reactive power compensation capability.

2.3.3 Back-to-Back Converters

The back-to-back converter consists of force-commutated rectifier and a forcecommutated inverter connected with a common DC-link [Carl98]. An important property of the back-to-back converter is the possibility of fast control of the power flow. By controlling the power flow to the grid, the DC-link voltage can be held constant. The properties of this combination are:

- The line-side converter normally operates at a given sinusoidal waveform;
- With this operation, the DC-link voltage must be higher than the peak of the main voltage;
- The regulation of the DC-link voltage is made by controlling the power flow to the AC grid and;

The topology model of this converter is presented in the following figure:



Figure 2.19. Topological model for a back-to-back converter.

As it can be appreciated in Figure 2.19, six switching devices define each side of the converter: the input and output side, both of them considered as inverter devices. It is consequently expected a signal quality with high harmonic content and therefore the need of big filtering devices. Nevertheless, AC-DC-AC converters are the main choice for converting frequency in the low frequency railway systems. The world's largest power supply converter is located in Datteln, Germany [WrUm09]. It has a nominal power of 413 MW distributed in four units (a unit is presented in Figure 2.20).

Some of the advantages of back-to-back converters are:

- Power is bidirectional.
- The DC-link voltage can be boosted to a level higher than the amplitude of the grid line-to-line voltage in order to achieve full control of the grid current.
- The capacitor between the inverter and rectifier makes it possible to decouple the control of the two inverters, allowing the compensation of system asymmetry (as unbalanced phases) on both the input side and the output side.
- The component costs are low.



Figure 2.20. A 103 MW back-to-back converter unit [Abb10].

Some of the disadvantages of the back-to-back converters are:

- The presence of the heavy and bulky DC-link capacitor increases the costs and reduces the overall lifetime of the system
- Every commutation in both the input and output between the upper and lower DC-link branch is associated with a hard switching and a natural commutation that increases the conduction losses.
- The high switching speed to the grid may also require extra filters.

The listed disadvantages are also unwanted characteristics for offshore application, whose reduction of components size and high waveforms quality are required. Moreover, when compared to HVDC case, using a back-to-back converter will require at least two AC-DC-AC conversion stations, which increases in two the conversion stages. This is way this converter is not held in this work.

In the HVDC case, none of the already existing solutions in AC-DC conversion was retained, and a newer converter recently developed seems to be the best choice. The converter is the Modular Multilevel that will be introduced in the following section.

2.4. Modular Multilevel Converters used in HVDC

The modular multilevel converter (MMC) has become the most attractive multilevel converter topology for medium and high-power applications, especially for voltage-sourced converter (VSC) in HVDC transmission systems [DQBS14]. In comparison with other multilevel converter topologies, the salient features of the MMC include:

- Modularity and scalability to meet any voltage level requirements,
- High efficiency, which is of significant importance for high-power applications,
- Superior performance in low harmonic content, specifically in high-voltage applications where a large number of identical submodules with low-voltage ratings are stacked up, thereby the size of passive filters can be reduced.

Therefore it has the best equilibrium between high quality signals and size design, its application is therefore convenient for OPS [CWBW11, NNAK13]. The MMC is a new and a potential candidate for HVDC applications [GDRS08]. The MMC does not have the same drawbacks of the multilevel and multi-module converters. However, the MMC-HVDC system still requires comprehensively analysis in the control strategies and very specifics operational characteristics in order to normalize it [KuDa12, XuDD13, YYJJ09].

2.4.1 MMC Modelling and Control

The schematics for a one-phase MMC in DC-AC connection is showed in the following Figure 2.21:



Figure 2.21. One phase MMC DC-AC schematics.

Variables can be defined as it follows: upper arm current (i1) and lower arm current (i2), arm inductances (Lc), DC-link voltage (V_{DC}) and AC source voltage (V_{AC}) . More information about submodules will be presented in chapter 3.

The MMC consists of two arms per phase-leg, where each arm contains M seriesconnected, nominally identical submodules (SM), and a series inductor Lc. The SM in each arm is controlled to generate the required AC phase voltage.

Main control of the converter is based on direct modulation technique, where the two arms per phase-leg are modelled and controlled by two complementary sinusoidal reference waveforms. The control of the converter is made using signal values from the input and output, currents and voltages measures, and capacitor voltages and arm currents as well. The major drawback of the direct modulation technique is the presence of circulating currents, which increases the converter power losses and rating values of the components. From the one phase case presented in Figure 2.21, the model equations are:

$$V_{DC}(t) = Lc \cdot \frac{d}{dt} \left[i_1(t) \right] + V_{AC}(t) + V_{Arm1}$$
2.2

$$-V_{DC}(t) = -Lc \cdot \frac{d}{dt} \left[i_2(t) \right] + V_{AC}(t) - V_{Arm2}$$

$$2.3$$

And two complementary equations can be built using the upper arm voltage (V_{arml}) and the lower arm voltage (V_{arml}) as following:

$$\frac{V_{Arm1} - V_{Arm2}}{2} = -V_{DC}(t) - Lc \cdot \frac{d}{dt} \left[\frac{i_1(t) - i_2(t)}{2} \right] - V_{AC}(t)$$
 2.4

$$\frac{V_{Arm1} + V_{Arm2}}{2} = V_{DC}(t) - Lc \cdot \frac{d}{dt} \left[\frac{i_1(t) + i_2(t)}{2} \right]$$
 2.5

In equation 2.4 the differences of currents i₁ and i₂, which are the arm currents, gives the AC source-side current. And in equation 2.5, the addition of currents gives the called circulating current of the MMC [LiJW13].

2.4.2 Basic control method for MMC used in DC-AC or AC-DC topology

A simplified representation of the MMC control is proposed in [BBEL13]. In Figure 2.22 are represented the main controlled variables in order to ensure a good operation of the MMC. These are related to energies calculation (Wn) and energy variations (Wn_{Δ}) contained in the MMC capacitors that satisfy the load demanded power. This power is related to the load current (I_{diff}) defined from equation 2.4:

$$I_{diff} = \frac{i_1(t) - i_2(t)}{2}$$
 2.6

In order to compare the proposed control for the MMC in AC-AC (presented later in chapter 4), in the following Figure 2.22 it is therefore presented a simple control diagram used for the MMC in AC-DC or DC-AC operation. In the figure it is used the twice of the fundamental frequency of the load, which is needed to control the circulating current of this converter. It will be showed in chapter 4 that the twice of the frequency does not exists in AC-AC configuration. As it can be appreciated, the main controlled variables are the capacitor energies and the load current defined in equation 2.6.



Figure 2.22. A simplified control representation for the MMC in DC-AC operation.

This control technique is characterized for using a Park and Clarke transformations [OvHR16], which are applied in order to set a continuous reference and being able of using PI correctors. This kind of corrector is going to be introduced further in this work. The main objective of this control is to keep a balance between energies and to supply the demanded load energy.

Energy changes are mainly related to the demanded output power in the MMC, while the energy level in capacitors have to be kept constant, and this is related to the circulating current control.

Many authors propose therefore newer control techniques in order to be able to work with this converter [HIFS17, WLGD17, YDLH17]. This is presented for instance in the reduction of the voltage ripple from Figure 2.23, where are showed the improvement results in the control techniques of a newer publication compared to an older one.



Figure 2.23. (a) Voltage ripple reduction in [BBEL13], face to results in (b) presented in [AnAN09].

2.4.3 AC LF system and the MMC

Some authors show the advantages of using a MMC in a LF system, the problem with these proposals are that two modules in AC-DC and DC-AC are used in their experiments in order to pass from one frequency value to another [TRWA15]. This doubles the needed components when compared to HVDC application. In this reference, the proposed scheme topology is presented in the following figure.



Figure 2.24. Two MMC in back-to-back operation in order to change frequency [TRWA15].

The overall efficiency for the presented converter in Figure 2.24 is about 98.8%. The example design was a ± 120 kVdc 250MW Back-to-Back MMC system and computer simulations showed good performance in both normal and fault conditions as presented in Figure 2.25. It was supposed a voltage drop to 0.8 p.u. of only one phase during 100 ms. While this fault occurred at the 50 Hz side, it could be observed how the LF-side is not affected by the fault, and only both DC voltages react to this particular condition.



Figure 2.25. Back-to-back MMC – voltages and currents behaviour when fault occurred at the 50 Hz side [TRWA15].

2.5. Conclusions

Many advantages are related to the use of LF systems were presented in this chapter. The technology exists and had been developed for over a century in the railway systems. The use of LF allows also the increase of the transmission distance in any power system because it reduces parasitic losses in overhead lines and cables. LF can therefore be implemented offshore instead of HVDC, like this, a multi-terminal system, as the one existing today in the German's railway, can be implemented without almost any physical and technological limit.

The reduction of frequency changes certain system operations when compared to a 50 Hz system and this affects some of the components of the system. As cables and overhead lines impedances are affected by the change of the frequency, also does the electrical transformers: when the frequency is lower, the power capability of a given the electrical transformer will be reduced too, and therefore its elements have to be bigger in order to keep the same efficiency. Another problem when frequency is reduced is the longer required time for zero crossing of voltages and currents waveforms as presented in the example of Figure 2.3. The crossing time in a 50/3 Hz system is 30 ms, which is three times longer when compared to a 50 Hz system. Nevertheless, protections for 50/3 Hz systems already exists and are applied to LF railways at nominal voltages going up to 132 kV.

A LF protection relay is much smaller than its equivalent in HVDC and can handle longer extinction times thanks to the natural zero crossing of alternative voltages and currents. In addition, even if the extinction time is three times longer than in AC 50 Hz, protection relays are already operational.

Power electronics are todays the most efficient ways to convert frequency, and after comparison presented in section 2.3, any AC-AC existing converter in direct operation is not suitable for offshore application because of their big size, need of large filtering devices, or the required power to manage, the comparison is synthetized in Table 2.2. Only back-to-back converters can that high power quantity expected for offshore applications, but they also need filtering devices. All presented converters need a stepup transformer to be placed in the LF side, which is unwanted for the expected application. Only MMC in back to back operation are suitable for the task.

	Cycloconverters	Matrix Converters	Back-to-Back	MMC
Small size	\mathbf{x}	\mathbf{x}	\mathbf{x}	$\mathbf{\otimes}$
Good signal quality	\bigotimes			
Without transformer	\bigotimes	\bigotimes	\bigotimes	
Without filter	\bigotimes		\bigotimes	
Large power management	\bigotimes	\bigotimes		

Table 2.2. Comparison between converters for frequency changing application.

MMC are capable to transform frequency in high voltages level systems when two modules are connected in back-to-back topology. However, the main drawback presented in this chapter is the needed components that doubles the quantity when compared to HVDC systems. A newer converter could be developed in AC-AC direct conversion using only one MMC device, this will require the adaption of its topology and the methodology for modelling and building a control as presented further in this work.

Chapter 3. AC-AC Modular Multilevel Converter

A Modular Multilevel Converter (MMC) was first proposed in 2003 [LeMa03]. The structure facilitates the renewable energies integration in medium and high voltage grids, especially for remote locations [HoMa14]. This is the main reason why a MMC was considered for studies in Offshore Power Systems (OPSs) [ZBBF16]. Although, MMC requires multiple isolated capacitors, modelled as a direct current (DC) sources that have to be balanced at a given energy value, introducing an extra difficulty for large scale OPSs which also are in High Voltage (HV).

MMC has been highly investigated for offshore applications [KMCS16], offering the starting bases in any subject as: application in multi-terminal grids, control techniques needed for high efficiency operation and, energy balancing [DQBS14] which is the main challenge for this kind of converters. All researches are mostly based on conversion between the alternative current (AC) and DC and therefore making a back-to-back AC-DC-AC topology, which in HVDC is characterized by a long DC link. However, this back-to-back configuration corresponds to a particular configuration of the MMC.

It is also possible to create AC-AC in direct conversion using a MMC and therefore to use a low-frequency transmission system instead of HVDC, which is the interest of this work. Moreover, the configuration requires the same quantity of elements as for the AC-DC conversion; in consequence the implementation effort of an MMC AC-AC would be the same than for the MMC AC-DC or DC-AC. Because of the alternative voltage in AC-AC configuration, only full-bridge for each submodule can be implemented. A comparison between full-bridge and half-bridge submodules will be presented in this chapter.

A MMC in direct AC-AC topology is the best choice for applications in OPS that are developed in AC Low Frequency (LF). This converter works as a frequency changer without DC passage in a transformer-less system, which is essential when working at 50/3 Hz as it was presented in chapter 2. The use of two different frequencies for only one system adds a difficulty in the control techniques that will be developed in this chapter.

It is therefore proposed a new modelling strategy for a MMC in direct AC-AC threephase structure developed using the same number of components that the AC-DC one. The main idea lies in the reorganization of its elements in order to connect independent AC systems at the input and output, and therefore be able of converting both voltage and frequency in order to keep the same power in the system.

A state of the art of earlier MMC working in AC-AC configuration will be presented in the first part of this chapter. It includes the proposed topologies and model. The proposed switching model of the converter in three-phase configuration will be introduced in the second part; this switching model represents the real configuration of the converter.

The full-bridge submodules topology will be described too and, for modelling purposes, it will be proposed the use of an averaged model that neglects the unwanted high frequencies produced by the submodule switches. The third part of this chapter evaluates the feasibility of the converter and it will be proposed a modelling strategy that consists in analysing each frequency-side separately. In addition, it will be presented that the arm currents are working at two frequencies at the same time and a method will be proposed in order to link these internal variables with the line currents that work at one frequency.

The final part of this chapter will evaluate the maximum voltage limits inside the converter, which are necessary in order to set the limits for the converter control that will be presented in the following chapter.

3.1. MMC AC-AC

MMC connected in AC-AC can be considered as a new technology because only a few publications as being found. Even if a three-phase topology already existed when this thesis work started, the proposed MMC in AC-AC connection was developed from the one-phase model detailed in [PRFK12] and from the three phase models in AC-DC configurations. This was useful because a newer modelling methodology could be developed, as it will be exposed in this chapter.

Some of the existing converters in AC-AC configuration from other previous works are listed in the following subsections. For this state of the art, some control techniques will be mentioned but not developed. This will help to compare the different approaches used between one-phase and three-phase topologies.

3.1.1 MMC direct-AC in one phase structure

One-phase AC MMC topologies have been used since 2004 [GlMa05]. Perez (et al) presented an improvement in the converter modelling in [PRFK12], they were able to introduce the circulating current parameter, which is controllable and helped to produce better results in the waveforms quality. In their studies, energy was controlled in two steps, defined by the authors as *balancing control* and *average control*.

The one-phase AC MMC proposal requires two legs and four arms as shown in Figure 3.1, these two legs are required to create a free circulation of the load frequency and to keep it inside the converter to avoid intrudes frequencies in the source and in the load. Three one-phase structures could be used in a three-phase system; nevertheless, this solution would require too many components. A single three-phase structure nevertheless reduces in a half the required elements per phase compared to three one-phase topology.



 $\label{eq:Figure 3.1.} \begin{tabular}{ll} MMC one-phase AC topology (a) and, Source, Load and Circulating current distributions (i_{s}, i_{o}, i_{c}) in (b). \begin{tabular}{ll} (PRFK12] \end{tabular}$

3.1.2 MMC direct-AC in three phase structure

Baruschka (et al) [BaMe11] presented in their article in 2011 the first three-phase MMC AC-AC. In this proposal, the converter was modelled using the consumed and generated power from each arm. Therefore, the entire development of frequencies in

this system is required in order to control the device (see Table 3.1). This may result in an interested approach when two frequencies values have to coexist in the same system.

Their proposed approach is only valid when the system is in balanced conditions and in steady state. Some examples of the system response in dynamic-state will reveal the presence of more harmonic frequencies (that normally are unpredictable as they are originated from fault conditions); this will be presented in chapter 4.

The converter structure is showed in Figure 3.2.



Figure 3.2. Hexverter topology proposed in reference [BaMe13a].

Even if the topology from Figure 3.2 was not used for developments in this thesis work, it presents the same converter topology as showed further in section 3.2. Nevertheless, a difference is found as it can be appreciated in the figure above; both AC systems have to be near one from the other because their model and control require the ground measures between AC system neutrals. They developed the entire coupled model of the system as presented next.

Frequency	Resulting component in branch power (amplitude)
f_1	$\hat{p}_{\rm bm}^{f_1} = \left[I_{\rm cir}^2 \hat{v}_{\rm g1}^2 + \frac{1}{3} V_{\rm st}^2 \hat{i}_{\rm g1}^2 + \frac{2(-1)^m}{3} I_{\rm cir} V_{\rm st} P_{\rm g1} - \frac{2\sqrt{3}}{9} I_{\rm cir} V_{\rm st} Q_{\rm g1} \right]^{\frac{1}{2}}$
f_2	$\hat{p}_{\mathrm{b}m}^{f_2} = \left[I_{\mathrm{cir}}^2 \hat{v}_{\mathrm{g2}}^2 + \frac{1}{3} V_{\mathrm{st}}^2 \hat{i}_{\mathrm{g2}}^2 + \frac{2(-1)^m}{3} I_{\mathrm{cir}} V_{\mathrm{st}} P_{\mathrm{g2}} + \frac{2\sqrt{3}}{9} I_{\mathrm{cir}} V_{\mathrm{st}} Q_{\mathrm{g2}} \right]^{\frac{1}{2}}$
$f_1 + f_2$	$\hat{p}_{\rm bm}^{f_1+f_2} = \frac{1}{6} \left[3\hat{v}_{\rm g2}^2 \hat{i}_{\rm g1}^2 + 3\hat{v}_{\rm g1}^2 \hat{i}_{\rm g2}^2 + \frac{4}{3} P_{\rm g1} P_{\rm g2} + \frac{4}{3} Q_{\rm g1} Q_{\rm g2} + \frac{4(-1)^m}{\sqrt{3}} \left(P_{\rm g1} Q_{\rm g2} - P_{\rm g2} Q_{\rm g1} \right) \right]^{\frac{1}{2}}$
$f_1 - f_2$	$\hat{p}_{\rm bm}^{f_1-f_2} = \frac{1}{6} \left[3\hat{v}_{\rm g2}^2 \hat{i}_{\rm g1}^2 + 3\hat{v}_{\rm g1}^2 \hat{i}_{\rm g2}^2 + \frac{8}{3} P_{\rm g1} P_{\rm g2} - \frac{8}{3} Q_{\rm g1} Q_{\rm g2} \right]^{\frac{1}{2}}$
$2f_1$	$\hat{p}_{ m bm}^{2f_1} = rac{\sqrt{3}}{6} \hat{v}_{ m g1} \hat{i}_{ m g1}$
$2f_2$	$\hat{p}_{ m bm}^{2f_2} = rac{\sqrt{3}}{6} \hat{v}_{ m g2} \hat{i}_{ m g2}$
const	$\hat{p}_{\rm bm}^{\rm const} = -\frac{1}{6}P_{\rm g1} - \frac{1}{6}P_{\rm g2} + (-1)^m \left(\frac{\sqrt{3}}{18}(Q_{\rm g1} - Q_{\rm g2}) - I_{\rm cir}V_{\rm st}\right)$

Table 3.1. Power calculations for the MMC proposal in reference [BaMe13a].

The variables proposed for the MMC model from [BaMe13b], are presented in Table 3.1, it is defined:

- frequency (*f*);
- indexes for both systems (1 and 2), for used arm (b) and sub-module index with (m);
- generated active and reactive power (*Pg* and *Qg*);
- a constant value of power measured in the neutral wired (*const*);
- and individual power calculations at each submodule (*P*_{bm});
- voltages from generators (Vg) and their respective currents (Ig), and;
- a value of circulating current (I_{cir}) and voltage between neutrals (V_{st}) .

Nevertheless, the spectrum of power used to balance the internal energy may not be so simple to be extended to general cases. This development requires the knowledge of the voltage between neutrals of both systems (AC₁ and AC₂) in order to calculate a defined "*adjacent power*" (P_{bm}^{const}). This power is calculated using the neutral-wire information between the two systems. This approach may not result in an attractive solution for far neutral distances as in the case of OPS.

For this kind of converter, the model and control implementation seems to be difficult. For instance, a calculation of the instantaneous power inside any arm requires a product between two waveforms at two different frequencies. In addition, because of the phase-shift between waveforms, analysis using sinusoidal expressions became quickly unmanageable. Fourier could be implemented, but this tool is not so a dynamic because it requires the knowledge of at least a half of cycle of the longest waveform to make the calculations and it is only precise in steady state. The same problem is related to Park and Clarke transforms. In lower frequencies systems this waiting time becomes bigger and therefore unattractive.

A Euler's transformation [GrSt94] can be used to make the product of these signals in steady state as presented in equations 3.1 and 3.2. It is only supposed that each arm voltage and currents are composed of two different frequencies and with their respective phase shift, the precise value of these variables are *a priori* unknown. A method to calculate these values will be proposed in section 3.3.1. The source side is defined with (s) index and the load side with (o) index. The arms instantaneous Power ($P_{\text{instantaneous}}$) is calculated with the arm voltages (V_{arm}) and currents (i_{arm}). The total power is the sum of four elements that depend on the product of voltages (V) and currents (i) using only two frequencies and also a combination of their respective phase shifts (Ψ and γ) presented in equation 3.3.

With the product of two waveforms, there are at least three different frequencies in the system: the source frequency (f_s) , the output frequency (f_o) and the sum of both of them.

$$v_{arm} = V_s \cdot \cos(\omega_s \cdot t + \Psi_s) + V_o \cdot \cos(\omega_o \cdot t + \Psi_o) = V_s \cdot e^{j \cdot (\omega_s \cdot t + \Psi_s)} + V_o \cdot e^{j \cdot (\omega_o \cdot t + \Psi_o)}$$
3.1

$$i_{arm} = i_s \cdot \cos(\omega_s \cdot t + \gamma_s) + i_o \cdot \cos(\omega_o \cdot t + \gamma_o) = i_s \cdot e^{j \cdot (\omega_s \cdot t + \gamma_s)} + i_o \cdot e^{j \cdot (\omega_o \cdot t + \gamma_o)}$$

$$3.2$$

 $P_{instantaneous} = v_{arm} \cdot i_{arm} = \dots$

$$Re\left[V_{s}\cdot i_{s}\cdot e^{j\cdot\left(2\omega_{s}\cdot t+\Psi_{s}+\gamma_{s}\right)}+V_{o}\cdot i_{o}\cdot e^{j\cdot\left(2\omega_{o}\cdot t+\Psi_{o}+\gamma_{o}\right)}+\ldots\right]$$
3.3

$$V_{o} \cdot i_{s} \cdot e^{j \cdot \left(\omega_{o} \cdot t + \omega_{s} \cdot t + \Psi_{o} + \gamma_{s} \right)} + V_{s} \cdot i_{o} \cdot e^{j \cdot \left(\omega_{o} \cdot t + \omega_{s} \cdot t + \Psi_{s} + \gamma_{o} \right)} \right]$$

$$\omega_s = 2 \cdot \pi \cdot f_s \qquad \qquad 3.4$$

$$\omega_o = 2 \cdot \pi \cdot f_o \tag{3.5}$$

In this thesis work, the direct-AC MMC is analysed from a different perspective than the one exposed in this section using reference [BaMe13b]. Any power analysis is required in the proposed model as the new perspective allows to decoupling the system in two and simplifies the modelling process. Therefore, currents and voltages can be modelled individually; this will be presented in section 3.3. Moreover, the developed strategies increases the MMC adaptability face to faults or irregularities in the system as it will be explained in this work.

3.2. Direct AC-AC Structure

Most of the researches about the MMC are based on the AC-DC reversible topology presented in Figure 3.3. There are only a few researches made on the AC-AC topology as it was exposed in section 3.1. The reason might be the extra difficulty in modelling and control from the existing proposals in the literature, and mostly when compared with AC-DC structures.



Figure 3.3. MMC DC-AC three-phase.

Instead of using a back-to-back MMC as it was suggested by some authors [ESWF00, SWNL16, TRWA15], the proposed topology configuration of this work consists in using only one MMC, eliminating the DC link and reorganizing all its elements in order to connect two AC systems directly.

A MMC is characterized for having a large number of submodules that allows high voltage operations. However, this large number introduces modelling challenges [ZBBF16]. For instance, in the electromagnetic transient (EMT) simulation programs, the switching operation is modelled by admittance matrix; the dimension of this matrix is given by the converters state variables number. This matrix must be inverted at each switching operation. Therefore, regarding the high dimension of the system and without an appropriate model according to this type of study, it is difficult to find a balance between simulation time and accuracy in systems containing MMC converters on EMT-type simulation programs. To solve this issue, different models have been developed in [ZBBF16] for studying the normal operation of MMC: Detailed model, Equivalent SM model and Averaged model.

The switching model of the converter represents the real model that should be used for OPS applications as it is explained next.

3.2.1 MMC Switching Model

The switching model is the less elaborated model of the converter [BaMB14]. It describes the electrical equations for each circuit configuration. It is also named exact model because it describes the converter behaviour exactly.

Without a DC link and by using an internal Delta (Δ) configuration, the converter makes it possible to connect a three-phase system. The output of the converter remains

with the same connection as for the DC topology. The Δ configuration is made by joining the converter legs between them and therefore creating a Δ interconnection between phases. This is necessary because it helps to keep the produced harmonics inside the converter, generating low THD waveforms. The converter structure is presented in the following Figure 3.4.



Figure 3.4. Switching model of the MMC in AC-AC topology.

In Figure 3.4, it is represented a MMC in three-phase configuration where the source and the output side are in AC and can be at different frequencies. There are six arm currents (i_{a1} , i_{a2} , i_{b1} , i_{b2} , i_{c1} , i_{c2}) circulating in each arm, which are equally composed of a total M-quantity of submodules. Each phase is constituted of one leg, which is composed of two arms from the same defined source (a, b or c).

All the variables presented inside this delta are identified as the *Converter Variables*. Between two arms from the same phase leg is connected each output phase. The sourceside and the output side are connected in star configuration, and their variables are defined as the *Line Variables*. In order to identify each used variable in the two frequencies system, it is proposed the following identification process.



Figure 3.5. Definition of variables and indexes.

The sources variables at input and output side are represented in upper caps, and in Figure 3.5 is generically denoted with (*K*) which can be, for either voltages (*V*) or currents variables (*I*). In addition, indexes (*s*) for the source-side and (*o*) for output load are added in order to identify the used frequency. Indexes *s* and *o* can also be used inside the converter variables in order to define a specific frequency. In this case, variables are represented with lower caps. Arms inside a leg are defined with (*n*) index, where $n = \{1, 2\}$.

Phases are defined with upper caps H or lower caps h index, which represent a number between 1 and 3, 1 representing phase a or A, 2 phase b or B, and 3 phase c or C. Normally, all developed equations are referenced to phase A or a. This means that H will have a zero value. Supposing as example the line variables, all other phases are therefore referenced as presented in the following equations:

For phases A,	K_H	3.6	
В,	$K_{(H+1)}$	3.7	

C,
$$K_{(H+2)}$$
 3.8

The switches of the converter can be found inside of each submodule, which in this thesis are all identical. In order to complete the topologic switching model from Figure 3.4, the submodules are therefore represented as it follows:



Figure 3.6. Full-Bridge submodule representation for an MMC.

Submodules contain also a capacitor (C_{SM}), this is where energy is stored in one frequency and produced in another frequency, making possible the frequency conversion in the converter. More details about submodules are presented in the following subsection.

3.2.2 Equivalent submodule model

MMC in AC-DC configuration can use submodules in half-bridge, which is the simplest submodule topology. In order to compare it with the full-bridge topology, it is represented in Figure 3.7. First, it can be appreciated that the switches quantity is the half when compared it with Figure 3.6; and therefore produces less conduction losses. Nevertheless, it cannot be used because it allows only two states of conduction that produce only positive and zero voltage. Current instead can go in both directions. One of the drawbacks of the half-bridge is that there is no DC-current limitation and the modulation factor is limited to one.



Figure 3.7 .Half-Bridge topology of submodules.

As presented in Figure 3.4, a MMC have an M-quantity of submodules (SM). The different voltages possibilities at the submodules output are showed in Figure 3.8. These submodules allow an increased modulation factor as presented in reference [IBHN00] (up to 1.4) and it has a better performance faced to faults (four blocking switches per submodule) [SHNN00].

The three main general states of the topology are showed in Figure 3.8: $+v_{SM}$, $-v_{SM}$ and zero, the last one was represented using either both upper switches, or both lower switches. All these states can be represented with a modulation index *m*. Each one of the presented states are in charging mode (left side) and discharging mode (right side).



Figure 3.8. Switching states of a submodule in Full-Bridge configuration.

In Table 3.2, are represented five possibilities for blocking state, which offer more robustness from security aspects than the Half-Bridge topology.

State	Gate 1	Gate 2	Gate 3	Gate 4	Result
Blocked 0	0	0	0	0	$V_{C} = cte. i_{arm} = 0$
Blocked 1	1	0	0	0	$V_{C}\!=\!V_{SM}$ if $i_{arm}\!>0,V_{SM}\!=\!0$ and $V_{C}\!=\!cte$ if $i_{arm}\!<0$
Blocked 2	0	1	0	0	-V_C = V_{SM} if $i_{arm} < 0$, $V_{SM} = 0$ and V_C = cte if $i_{arm} > 0$
Blocked 3	0	0	1	0	-V_C = V_{SM} if $i_{arm} < 0$, V _{SM} = 0 and V _C = cte if $i_{arm} > 0$
Blocked 4	0	0	0	1	$V_{\rm C}{=}V_{\rm SM}ifi_{arm}{>}0,V_{\rm SM}{=}0$ and $V_{\rm C}{=}cteifi_{arm}{<}0$

Table 3.2. Blocking state for a Full-Bridge submodule. [SJLJ16]

Exchanges of energy in the MMC take place inside the submodules: in the presented MMC, which is working in AC-AC coupling, switches act with combination of at least both frequencies of the system in steady state, and also compensating the other submodules energies when needed [YDLH17]. In dynamic state, there may occur some

not predictable cases, as fault or rapid changes of frequencies in the line for instance, it can therefore exist more than two frequencies in the switching that will be introduced into the energy contained in capacitors.

Based on the definitions presented with Figure 3.8, it is possible to describe the dynamics of an individual MMC SM; it is presented the submodule capacitor (C_{SM}), the capacitor voltage (v_c), the submodule current (i_{SM}) and the submodule voltage (v_{SM}):

$$m = (-1, 0, 1) \tag{3.9}$$

$$C_{SM} \cdot \frac{d}{dt} \left[v_c(t) \right] = m \cdot i_{SM}(t)$$
3.10

$$v_{SM}(t) = m \cdot v_c(t) \tag{3.11}$$

In order to generalize equations 3.10 and 3.11, it is applied the equation representation from Figure 3.5 (where hn index will specify the phase and the arm), and a sm super index used to identify the precise submodule where calculations are taking place, then equations are:

$$sm = (1, 2, \dots, M)$$
 3.12

$$\left(C_{SM}\right)^{(sm)} \cdot \left[\frac{d}{dt} \left[v_{c_{hn}}(t)\right]^{(sm)}\right] = m_{hn}^{(sm)} \cdot \left[i_{SM}(t)\right]$$

$$3.13$$

$$\begin{bmatrix} v_{SM_{hn}}(t) \end{bmatrix}^{(sm)} = m_{hn}^{(sm)} \cdot \begin{bmatrix} v_{c_{hn}}(t) \end{bmatrix}^{(sm)}$$

$$3.14$$

If for instance is supposed a converter with M = 200, then the total amount of submodules in the three-phase MMC would be M, times 2 arms, times 3 phases, which is 1200 submodules. In order to use a reduced order model, an average operation is suitable. By making some basic simplifying hypothesis presented in the following subsection 3.2.3, each arm can be represented by an equivalent continuous voltage source.

3.2.3 Submodule average model

The difference between a switching model and an averaged one is the absence of the high frequency switching phenomena that are unwanted and must be attenuated and even neglected. When control laws need to be implemented, the original discontinuous model must be transformed in a continuous invariant model that provides the best representation of the system for the dynamic analysis and control tuning. The simulation issues are interested as well to use such averaged because their rapidity [BaMB14].

In order to find the average model, all submodules must be averaged and reduced to only one equivalent source. First, the total arm voltage $(\nu_{hn}(t))$ is averaged from equation 3.14:

$$v_{hn}(t) = \sum_{sm = 1}^{M} \left[v_{SM_{hn}}(t) \right]^{(sm)} = \sum_{SM = 1}^{M} \left[m_{hn}^{(sm)} \cdot \left[v_{c_{hn}}(t) \right]^{(sm)} \right]$$
3.15

Every submodule arm voltage is considered for having the same value. Then equation 3.15 becomes:

$$v_{hn}(t) = \sum_{SM = 1}^{M} \left[m_{hn}^{(sm)} \cdot \left[v_{c_{hn}}(t) \right]^{(sm)} \right] = M \cdot m_{hn} \cdot v_{c_{hn}}(t) = m_{hn} \cdot v_{c\Sigma_{hn}}(t)$$
3.16

$$m_{hn} = \frac{\sum_{SM = 1}^{M} \left(m_{hn}^{sm} \right)}{M}$$

$$3.17$$

Where $\nu_{C\sum hn}(t)$ from equation 3.16 is the aggregated capacitor voltage. The same procedure is applied to the equation 3.13. In addition, in this case the equivalent capacitor for each arm must be established by taking into account its series-connection. By supposing that all submodules capacitors have the same value, it comes:

$$\sum_{sm = 1}^{M} (C_{SM})^{(sm)} = \frac{C_{SM}}{M}$$
 3.18

$$C = \frac{C_{SM}}{M}$$
 3.19

$$\left[i_{SM}(t)\right]^{(sm)} = i_{hn}(t) \tag{3.20}$$

$$\frac{C_{SM}}{M} \cdot \left[\frac{d}{dt} \sum_{sm = 1}^{M} \left[v_{c_{hn}}(t) \right]^{(sm)} \right] = \sum_{sm = 1}^{M} \left[m_{hn}^{(sm)} \right] \cdot i_{hn}(t)$$

$$3.21$$

$$\frac{C_{SM}}{M} \cdot \frac{d}{dt} \left[v_{c\Sigma_{hn}}(t) \right] = m_{hn} \cdot i_{hn}(t)$$
3.22

The average model can now be represented as following.


Figure 3.9. Equivalent topology from the switching model of one full-bridge submodule to average model containing m submodules.

The average model of one MMC submodule is presented in Figure 3.9, the resistor R represents some losses due to some imperfections of the considered capacitor.

The relationship between the energy transfer and the submodules components in the converter is described in the following subsection.

3.2.4 Submodule energy calculations

Changes in the energy contained in capacitors are linked the input and output energies of the converter. This energies are coming from the power exchanges which, in a constant voltage system, these exchanges are related to the currents values. The energy contained in capacitors is therefore used to adjust the input current value and to supply the output power demanded by the load. This process can be modelled as it follows:

$$Energy_{hn}(t) = \frac{1}{2} \cdot \left[v_{c\Sigma_{hn}}(t) \right]^2 \cdot C$$
3.23

$$m_{hn} \cdot i_{hn}(t) = C \cdot \frac{d}{dt} \left[v_{c\Sigma_{hn}}(t) \right] + \frac{v_{c\Sigma_{hn}}(t)}{R}$$

$$3.24$$

$$m_{hn} \cdot \left[v_{c\Sigma_{hn}}(t) \right] \cdot i_{hn}(t) = v_{hn}(t) \cdot i_{hn}(t) = \frac{C}{2} \cdot \frac{d}{dt} \left[v_{c\Sigma_{hn}}(t) \right]^2 + \frac{\left[v_{c\Sigma_{hn}}(t) \right]^2}{R}$$

$$3.25$$

Introducing equation 3.23 into equation 3.25 then:

$$i_{hn}(t) = \frac{2}{v_{hn}(t)} \cdot \left[\frac{d}{dt} \frac{Energy_{hn}(t)}{2} + \frac{Energy_{hn}(t)}{C \cdot R} \right]$$
3.26

The link between the arm current and the energy variations inside the converter is represented with equation 3.26. The entire average model of the converter can now be obtained.

3.3. Converter average model

Introducing the submodules average model presented in Figure 3.9 in the converterswitching model from Figure 3.4 and reorganizing the converter to create a delta topology, then the converter average model can be represented as following.



Figure 3.10. Average model of the MMC in YDY configuration.

This particular structure has to deal with two different frequencies at the same time. Equations from a MMC DC-AC [Cong15], which were presented in section 2.4, do not model the effect of a second system in AC, and therefore they cannot be directly implemented. All impedances presented in Figure 3.10 are mainly reactive, and a small series resistor can symbolise some active power losses in the arm inductances. The voltage drop for each arm inductance is given by the following equation:

$$v_{hn_{Lc}}(t) = Lc \cdot \left[\frac{d}{dt} \left[i_{hn}(t) \right] \right] + Rc \cdot i_{hn}(t)$$

$$3.27$$

Also for each source and output reactance, voltage equations are:

$$V_{Hs_{Ls}}(t) = Ls \cdot \left[\frac{d}{dt} \left[I_{Hs}(t) \right] \right] + Rs \cdot I_{Hs}(t)$$
3.28

$$V_{Ho_{Lo}}(t) = Lo \left[\frac{d}{dt} \left[I_{Ho}(t) \right] \right] + Ro I_{Ho}(t)$$
3.29

Where *Lc*, *Ls and Lo* are respectively the arm, source and load inductances. Normally equations 3.28 and 3.29 are at their own frequency while two frequencies values are contained in equation 3.27, and therefore the voltage measures cannot be decomposed instantly; more details about these mixed frequencies will be discussed in the next section. Some of the problems encountered in the modelling task are the presence of two frequencies in arm waveforms, the link between input and output and arm currents, which are related to use two three-phase systems at two different frequencies in only one link [DaBR17a]. A solution for each one of these problems will be addressed with the developed in the following subsections.

3.3.1 Feasibility of the MMC AC-AC

In order to develop the main model of the converter, it will be use a perfect model, which is a simplification from Figure 3.10. The simplification consists in transforming the capacitors submodules into arm voltage sources as it is presented in Figure 3.11. In this case, there is not a modulation index; instead, there will be the same waveform than the wanted voltage waveform linked to each arm. These sources will produce exactly the required voltage, but without the energy limitation from capacitors, which is why it is called a perfect model. Sources at input and output sides are connected in star coupling (Y) and the converter instead have a coupling in Δ .



Figure 3.11. Perfect average model for the MMC in YDY configuration.

When looking to equations 3.1 and 3.2, only two frequencies are taken into account, source and load side. In addition, what is more important is that voltages and currents are represented by only these two frequencies independently, allowing a separate analysis of each one of them. This is the first step proposed for analysis.

A second step stands for the identification process of the current values inside the converter, which is composed of two frequencies. These currents cannot be controlled directly, and a control can be applied on the input and output currents. The link between these currents in magnitude and phase is the same that the one found in an electrical transformer. The theory is explained in section 3.3.4.

Given the numerous variables that have been defined for this converter, we propose a list for identifying each frequency variable used in the model. Table 3.3 presents these variables and those from application of the superposition theorem, presented in the following subsection.

		Input Source (s)	Output Load (o)	Converter	
	Voltages	V_{As} V_{Bs}	V _{Ao} V _{Bo}	v_{a1}, v_{a2} v_{b1}, v_{b2}	e ation
	Currents	$V C_8$ I_{As} I_{Bs} I_{Cs}	$V C_{o}$ I_{Ao} I_{Bo} I_{Co}	$egin{array}{c} v_{c1}, v_{c2} \ i_{a1}, i_{a2} \ i_{b1}, i_{b2} \ i_{c1}, i_{c2} \end{array}$	Simpl represent
Source-side superposition	Voltages, phase A referenced Currents, phase A referenced	$egin{aligned} V_{Hs} \ V_{(H+1)s} \ V_{(H+2)s} \ I_{Hs} \ I_{(H+1)s} \ I_{(H+2)s} \end{aligned}$	- - - - -	v_{hns} $v_{(h+1)ns}$ $v_{(h+2)ns}$ i_{hns} $i_{(h+1)ns}$ $i_{(h+2)ns}$	Equation representation proposed in Figure 3.5
Output-side superposition	Voltages, phase A referenced Currents, phase A referenced	- - - - -	V_{Ho} $V_{(H+1)o}$ $V_{(H+2)o}$ I_{Ho} $I_{(H+1)o}$ $I_{(H+2)o}$	v_{hno} $v_{(h+1)no}$ $v_{(h+2)no}$ i_{hno} $i_{(h+1)no}$ $i_{(h+2)no}$	Equation representation proposed in Figure 3.5
		_	т (<i>п</i> +2)0	v(n+2)no	

Table 3.3. Model variables definition.

3.3.2 Superposition theorem

In order to correctly model the proposed converter from Figure 3.11, the contribution at each frequency needs to be studied individually; the superposition theorem is therefore proposed [RaKu13].

The converter is separated in two subsystems by taking into account the concerning elements whose frequency form is expected to exist: source side and load side elements. It can be remarked that the converter arms are a common space for these two frequencies. The model will also suppose the presence of unwanted frequencies that normally are zero when supposing a balanced system. Presented results are based in only one-phase analysis for each side; this is possible when supposing a balanced system where the sum of all phase currents gives zero as in equation 3.30.

$$I_{AS} + I_{BS} + I_{CS} = 0 = I_{Ao} + I_{Bo} + I_{Co}$$
3.30

Superposition theorem is applied first to the source side as presented in Figure 3.12; the input frequency is defined as (*fs*). While applying this theorem, two new variables are defined: the source current contribution to the load side (i_{xhs}), and the voltage at the load connection point induced by the source side (v_{xhs}).

These variables are normally equal to zero in order to avoid intruded source frequencies at the load side. H and h indexes were already defined as the phase values for line and converter variables respectively, and x index represents only the unknown voltage variable. This voltage can be in any system frequency or even in a combination of both system frequencies.



Figure 3.12. Superposition theorem applied to the source side.

Kirchhoff's voltage law is applied to the scheme of Figure 3.12. Resulting equations are:

$$-V_{As}(t) + Ls \cdot \frac{d}{dt} \left[I_{As}(t) \right] + Lc \cdot \frac{d}{dt} \left[i_{a1s}(t) \right] + v_{a1s}(t) = v_{xas}$$

$$3.31$$

$$V_{Bs}(t) - Ls \cdot \frac{d}{dt} \left[I_{Bs}(t) \right] + Lc \cdot \frac{d}{dt} \left[i_{a2s}(t) \right] + v_{a2s}(t) = -v_{xas}(t)$$

$$3.32$$

Applying the superposition theorem from Figure 3.12, two equations can be formulated. These equations are defined in this thesis work as the sigma voltages from the source ($V_{\Sigma H_s}$) and delta voltages from the source ($V_{\Delta H_s}$) [DaBR17b]. They correspond to an analogy from the equations 2.4 and 2.5 used for controlling the MMC in AC-DC topology. Generalizing to all phases:

$$V_{\Sigma_{Hs}}(t) = \frac{v_{h1s}(t) + v_{h2s}(t)}{2} = \frac{V_{Hs}(t) - V(H+1)s(t)}{2} - \dots$$

$$Ls \cdot \frac{d}{dt} \left[\frac{I_{Hs}(t) - I(H+1)s(t)}{2} \right] - Lc \cdot \frac{d}{dt} \left[\frac{i_{h1s}(t) + i_{h2s}(t)}{2} \right]$$

$$V_{\Delta_{Hs}}(t) = \frac{v_{h1s}(t) - v_{h2s}(t)}{2} = \frac{V_{Hs}(t) + V(H+1)s(t)}{2} - \dots$$

$$Ls \cdot \frac{d}{dt} \left[\frac{I_{Hs}(t) + I(H+1)s(t)}{2} \right] - Lc \cdot \frac{d}{dt} \left[\frac{i_{h1s}(t) - i_{h2s}(t)}{2} \right] - V_{xhs}(t)$$
3.33

The same procedure is repeated at the load side from Figure 3.13



Figure 3.13. Superposition theorem applied to the load side.

In this case, currents are at the output frequency (f_o) and this frequency should not appear at the source side. As in the case before, two new variables are defined, the load current contribution from the other source side with a phase shit of $-2\pi/3$ from the analysed phase ($i_{x(h+1)o}$), and the voltage at that source connection point induced by the load side ($V_{x(h+1)o}$). Equations can therefore be expressed as:

$$-V_{Ao}(t) - Lo \cdot \frac{d}{dt} \left[I_{Ao}(t) \right] + Lc \cdot \frac{d}{dt} \left[i_{a2o}(t) \right] + v_{a2o}(t) = v_{xbo}$$

$$3.35$$

$$V_{Bo}(t) + Lo \cdot \frac{d}{dt} \Big[I_{Bo}(t) \Big] + Lc \cdot \frac{d}{dt} \Big[i_{b1o}(t) \Big] + v_{b1o}(t) = v_{xbo}$$
3.36

Adding and subtracting equations 3.35 and 3.36 and then generalizing will also give two equations, sigma-voltages of the load ($V_{\Sigma ho}$) and delta-voltages of the load ($V_{\Delta ho}$). It can be appreciated that arm currents used for the input side model are different from those used for the output side model (see Figure 3.12 and Figure 3.13); this is why equations are different in both cases.

$$V_{\Sigma}_{Ho}(t) = \frac{v_{h2o}(t) + v(h+1)_{Io}(t)}{2} = \frac{V_{Ho}(t) - V_{(H+1)o}(t)}{2} + \dots$$
3.37
$$Lo \cdot \frac{d}{dt} \left[\frac{I_{Ho}(t) - I_{(H+1)o}(t)}{2} \right] - Lc \cdot \frac{d}{dt} \left[\frac{i_{h2o}(t) + i(h+1)_{Io}(t)}{2} \right] - v_{x(h+1)o} + \dots$$

$$V_{\Delta}_{Ho}(t) = \frac{v_{h2o}(t) - v(h+1)_{Io}(t)}{2} = \frac{V_{Ho}(t) + V_{(H+1)o}(t)}{2} + \dots$$

$$Lo \cdot \frac{d}{dt} \left[\frac{I_{Ho}(t) + I_{(H+1)o}(t)}{2} \right] - Lc \cdot \frac{d}{dt} \left[\frac{i_{h2o}(t) - i(h+1)_{Io}(t)}{2} \right]$$
3.38

Each individual response is added in order to obtain the sigma and delta equations. These equations are used to calculate the needed modulation index that will allow the control of the converter. Control details will be presented in chapter 4.

$$V_{\Sigma_{H}}(t) = V_{\Sigma_{Hs}}(t) + V_{\Sigma_{Ho}}(t)$$
3.39

$$V_{\Delta_H}(t) = V_{\Delta_{Hs}}(t) + V_{\Delta_{Ho}}(t)$$
3.40

Finally, the arm voltages that are related to the modulation index are calculated as presented in the following equations:

$$v_{h1}(t) = V_{\Sigma_H}(t) + V_{\Delta_H}(t)$$
 3.41

$$v_{h2}(t) = V_{\Sigma_H}(t) - V_{\Delta_H}(t)$$

$$3.42$$

$$m_{h1} = \frac{v_{hI}(t)}{V_{DC}} \qquad * \tag{3.43}$$

$$m_{h2} = \frac{v_{h2}(t)}{V_{DC}}$$
 * 3.44

Where m_{h1} and m_{h2} are the submodule modulation indexes. They are needed to control the voltages sources from the converter average model with its equivalent capacitors. Developed equations are also valid for the switching model. In that case, V_{DC}^* is the continuous voltage reference from the submodule capacitors. This reference value must be chosen correctly, this will define the contained energy in capacitors and its capability to transfer power as explained next in section 3.4.

As the converter currents are measured with both frequencies at the same time, and each frequency contribution cannot be measured instantaneously; and Fourier decomposition [Arri97, WeWi73] risk the system to be too slow, then another method must be used. A way to calculate the instantaneous frequency contribution will be proposed in the following subsections.

3.3.3 Transformer analogy

Three-phase transformers can have two rating voltages depending on their internal windings connection in delta (Δ) or in star (Y) and all configurations between both of them [GrSt94]. What is relevant to analyse in this theory is the analogy with the phase shift introduced by a transformer, has an example a Y- Δ transformer is analysed next.



Figure 3.14. Y- Δ wiring diagram.

Capital letters define the high voltage side in this example. Winding AN is magnetically linked to the winding ab and, V_{AN} is in phase with voltage V_{ab} , producing phase-shift between V_A and V_a . The phase-shift and amplitude adjustment when passing from one configuration to another is defined using the following equations. Only are studied the phase A and phases a, and b because the process is equivalent for the other phases.

This leads to a line-to-neutral voltage composed of line-to-line data:

$$\overrightarrow{V}_{a} = \frac{|V_{ab}|}{\sqrt{3}} \cdot e^{j \cdot \left(\omega t + \Psi_{ab} - \frac{\pi}{6}\right)}$$

$$3.45$$

$$\overrightarrow{V}_{b} = \frac{|V_{ac}|}{\sqrt{3}} \cdot e^{j \cdot \left(\omega t + \Psi_{bc} - \frac{\pi}{6}\right)}$$

$$3.46$$

$$\overrightarrow{V}_{c} = \frac{|V_{ca}|}{\sqrt{3}} \cdot e^{j \cdot \left(\omega t + \Psi_{ca} - \frac{\pi}{6}\right)}$$

$$3.47$$

Where Ψ_{ab} , Ψ_{bc} and Ψ_{ca} are the phase-shift angles of the corresponding voltages. These equations can be used in any direction. The process is equivalent for currents and the result is expressed as it follows:

$$\vec{i}_{ab} = -\left(\frac{|ia|}{\sqrt{3}}\right) \cdot e^{j \cdot \left(\omega t + \gamma_a + \frac{\pi}{6}\right)}$$

$$3.48$$

$$\overrightarrow{ibc} = -\left(\frac{|ib|}{\sqrt{3}}\right) \cdot e^{j \cdot \left(\omega t + \gamma_b + \frac{\pi}{6}\right)}$$

$$3.49$$

$$\vec{i}_{ca} = -\left(\frac{|ic|}{\sqrt{3}}\right) \cdot e^{j \cdot \left(\omega t + \gamma_c + \frac{\pi}{6}\right)}$$

$$3.50$$

Where γ_a, γ_b and γ_c are the corresponding phase-shift current angles.

3.3.4 Link between arm and references currents

Measurements are made at the input and output of the converter where each corresponding frequency contribution can be obtained instantaneously. The link between these currents and the arm currents of the converter is the same that the one presented in the previous subsection. This introduces the following analogy: the converter variable waveforms presents the same behaviour as an electrical transformer with an arbitrary turns ratio and without a magnetic core, which in the MMC has been replaced by the internal inductances.

In the converter scheme from Figure 3.12 and Figure 3.13 and from their delta equations 3.34 and 3.38, delta currents can be respectively defined as:

$$\frac{i_{h1s}(t) - i_{h2s}(t)}{2} = i_{\Delta_{hs}}(t)$$
3.51

$$\frac{i_{h2o}(t) - i_{(h+1) \cdot Io}(t)}{2} = i_{\Delta_{ho}}(t) : \qquad 3.52$$

Using the same analysis and from sigma voltages in equations 3.33 and 3.37, the respectively sigma currents are found:

$$\frac{i_{h1s}(t) + i_{h2s}(t)}{2} = i_{\Sigma_{hs}}(t)$$
3.53

$$\frac{i_{h2o}(t) + i_{(h+1) \cdot Io}(t)}{2} = i_{\Sigma_{ho}}(t)$$
 3.54

Current values from equations 3.51 and 3.52 must be zero in order to avoid intrudes frequencies at the converter source and output sides. Then, sigma currents from equations 3.53 and 3.54 can be linked to the energy transfer between the source-side, the converter and the output-side. This can be illustrated in Figure 3.15 by supposing a start-delta-star coupling in the system.



Figure 3.15. Transformer analogy

In order to facilitate the representation between the line and the converter variables discomposed from the superposition theorem (which was introduced in subsection 3.3.2), and to reveal the induced phase shift developed in section 3.3.3, a graphic support is used. Phase vector were developed and presented in the following figure.



Figure 3.16. Phasor diagram and application of the superposition theorem.

By supposing a balanced system, arm currents are related with source and load currents from sigma equations 3.53 and 3.54 and by changing amplitude and phase as presented on the diagram from Figure 3.16.

It can be remarked that the source and load currents are represented differently. First, load currents have a phase-shift of π because they were defined like this the diagram from Figure 3.11 as they supply the load side. Second, because of their connection point in the delta, the resulting contribution is not the same in the converter currents. This precise configuration is introduced into the control equations.

Finally, when adding the phase shifts to the source and load currents (γ_{Hs} and γ_{H0}) and from equations 3.48, 3.49 and 3.50, the final current equations are:

$$i_{h1s}(t) = i_{h2s}(t) = \frac{|I_{Hs}|}{\sqrt{3}} \cdot e^{j \cdot \left(\omega_s \cdot t + \gamma_{Hs} + \frac{\pi}{6}\right)}$$
3.55

$$\omega_s = 2 \cdot \pi \cdot fs \tag{3.56}$$

$$i_{h2o}(t) = i_{(h+1) \cdot Io}(t) = \frac{|I_{Ho}|}{\sqrt{3}} \cdot e^{j \cdot \left(\omega_0 \cdot t + \gamma_{Ho} + \frac{\pi}{6} + \pi\right)}$$
3.57

$$\omega_o = 2 \cdot \pi \cdot fo \qquad \qquad 3.58$$

Equations 3.55 and 3.57 give the relationship between the arm currents and the source-side and load-side currents.

An average model of the system using only perfect voltage sources, which equations were developed in this section, was built and simulated in order to present some basic simulation results.

In Figure 3.17 are presented the preliminary results of an MMC working as a frequency and voltage changer. The used parameters for simulation are presented in Table 3.4, as the control is not developed at this stage yet, nominal power was fixed to 1 kW at both sides and the current reference was manually calculated. A control to automatically adjust the needed input power will be implemented in chapter 4.

Table 3.4. Converter parameters for	or simulation.
-------------------------------------	----------------

	Input	Output
Voltage Amplitude [V]	500	250
Frequency [Hz]	50	50/3
Source Inductance [mH]	0.2	0.2
Arm Inductances [mH]	0.2	
Active and Reactive Power	1000 W and 0 VAr	



Figure 3.17. Input and output of the converter three-phase voltages and currents

It can be appreciated both frequency separately and a high quality of signals, of course these are results coming from a perfect average model and in a more exact model is expected to present some higher frequencies in waveforms. Currents inside the converter delta were simulated individually in order to analyse each contribution; these are presented in Figure 3.18.



Figure 3.18. Arm current compared with input current (a). Arm current compared with output current (b). Arms currents (1 and 2) when both sides activated.

In Figure 3.18.a and b are showed the expected results from the described transformer theory presented with equations 3.55 and 3.57

- In (a), it is used only a source-side current. Arm currents are at 50 Hz and it can be appreciated a phase-shift that correspond to the transformer analogy presented before in subsection 3.3.3. It is of $\pi/6$ with the source and the input amplitude is $\sqrt{3}$ times of the arm current.
- In (b), using only load currents there is only 50/3 Hz of frequency, the expected phase-shift of $\pi/6$ plus a negative direction (+ π) as defined in Figure 3.16, and it has the same $\sqrt{3}$ times amplitude of the arm current.
- Finally, in (c) there is not only the sum between the currents from different frequencies but also there is a phase shift, built with the system equations in order to maintain harmonics inside the converter only.

3.4. MMC Limits

MMC is designed using one-inductor and M-capacitors per arm, which are related to the equations developed in this chapter. Inductors are necessary in order to integrate two frequencies in only one system while limiting the harmonic content produced by the switching devices [QZHJ10]. These inductances will reduce the raising rate of transient faults currents, inside or outside the converter [ZBBF16]. In addition, capacitors are needed to produce the required voltages at each arm, which is composed of both frequencies.

These inductances and capacitors have to tolerate some voltages and currents limits that are related to normal operation in steady state. These limits, which are associated to the required operation point of power and nominal voltage, are going to be calculated in this section.

Each input and output sources have their own inductances, which are not related to the converter control but they add stability to the converter and act as a current filter face to the switching devices of the converter. The value of these inductors could be related to the internal impedance from generators, the impedance of the lines and even to filter designs if inductors are added to the input and output side of the converter [LiBH05, SHNN00, ZyGN13], nevertheless this is out the sight of this thesis work.

The voltages and currents limits for a given range of inductance values are evaluated. It will be showed how nominal power and nominal voltages of the converter are related with its internal components by analysing in steady state and balanced conditions.

3.4.1 Arm inductors, steady state analysis

The steady-state analysis is made in order calculate a particular operation point of the converter. This operation point is generally chosen in order to obtain the maximum value of a given variable. For instance, calculation of the maximum voltage that any inductor might face during its normal operation. This method will help to define the nominal voltage of components for a given operation point. Only one phase is evaluated and the results are generalized and can be applied to all phases. However, the arm currents given by equations 3.55 and 3.57, which are related to each frequency side, are supposed as known.

All currents are defined with a sinusoidal behaviour at their respective frequencies. Source and load currents are:

$$I_{Hs}(t) = I_{Hs'} \cos(\omega_{s'} t + \gamma_{Hs})$$

$$3.59$$

$$I_{Ho}(t) = I_{Ho} \cos(\omega_o \cdot t + \gamma_{Ho})$$
3.60

Arm currents represented with each frequency contribution are therefore:

$$i_{h1s}(t) = i_{h2s}(t) = \frac{I_{Hs}}{\sqrt{3}} \cdot \cos\left(\omega_s \cdot t + \gamma_{Hs} + \frac{\pi}{6}\right)$$

$$3.61$$

$$i_{h2o}(t) = i_{(h+1) \cdot Io}(t) = \frac{I_{Ho}}{\sqrt{3}} \cdot \cos\left(\omega_o \cdot t + \gamma_{Ho} + \frac{\pi}{6} + \pi\right)$$
 3.62

In order to simplify the following results, a change of variable is proposed:

$$I_s = \frac{I_{Hs}}{\sqrt{3}} \tag{3.63}$$

$$I_o = \frac{I_{Ho}}{\sqrt{3}}$$
 3.64

From equation 3.61, arm currents 1 and 2 from the source frequency side are from the same phase h. This is not the case for the load frequency side. In this case, arm currents 1 and 2 are from different phases h and (h+1). This was represented in Figure 3.16. In order to calculate the corresponding arm currents and phases from the two frequencies, then the phase shift between currents have to be taken into account as it is detailed in the following equations.

$$i_{a1}(t) = i_{a1s}(t) + i_{a1o}(t) = i_{a2s}(t) + i_{c2o}(t)$$
3.65

$$i_{a2}(t) = i_{a2s}(t) + i_{a2o}(t) = i_{a1s}(t) + i_{b1o}(t)$$
3.66

$$i_{b1}(t) = i_{b1s}(t) + i_{b1o}(t) = i_{b2s}(t) + i_{a2o}(t)$$
3.67

$$i_{b2}(t) = i_{b2s}(t) + i_{b2o}(t) = i_{b1s}(t) + i_{c1o}(t)$$
3.68

$$i_{c1}(t) = i_{c1s}(t) + i_{c1o}(t) = i_{c2s}(t) + i_{b2o}(t)$$
3.69

$$i_{c2}(t) = i_{c2s}(t) + i_{c2o}(t) = i_{c1s}(t) + i_{a1o}(t)$$
3.70

Arm current equations can therefore be generalized as following,

$$i_{h\,Io}(t) = i_{(h+2)2o}(t) = I_o \cdot \cos\left(\omega_o \cdot t + \gamma_{Ho} + \frac{\pi}{6} + \pi + \frac{2 \cdot \pi}{3}\right)$$
3.71

Therefore each individual arm current from the load is,

$$i_{h\,Io}(t) = -I_o \cdot \cos\left(\omega_o \cdot t + \gamma_{Ho} + \frac{5 \cdot \pi}{6}\right)$$
3.72

$$i_{h\,2o}(t) = -I_o \cdot \cos\left(\omega_o \cdot t + \gamma_{Ho} + \frac{\pi}{6}\right)$$

$$3.73$$

And for the source side both arm currents are the same,

$$i_{hns}(t) = I_s \cdot \cos\left(\omega_s \cdot t + \gamma_{Hs} + \frac{\pi}{6}\right)$$

$$3.74$$

Then, integrating source and load side,

$$i_{h\,I}(t) = i_{h\,Is}(t) + i_{h\,Io}(t) = I_s \cdot \cos\left(\omega_s \cdot t + \gamma_{Hs} + \frac{\pi}{6}\right) - I_o \cdot \cos\left(\omega_o \cdot t + \gamma_{Ho} + \frac{5 \cdot \pi}{6}\right)$$

$$3.75$$

$$i_{h2}(t) = i_{h2s}(t) + i_{h2o}(t) = I_{s} \cos\left(\omega_{s} \cdot t + \gamma_{Hs} + \frac{\pi}{6}\right) - I_{o} \cdot \cos\left(\omega_{o} \cdot t + \gamma_{Ho} + \frac{\pi}{6}\right)$$

$$3.76$$

Voltage drops on inductances are calculated by introducing source current equations 3.75 and 3.76, and arms current equations 3.72, 3.73 and 3.74 into equations 3.27, 3.28 and 3.29. Source and output sides voltage drops are:

$$\begin{split} V_{Hs_{Ls}}(t) &= Ls \cdot \left[\frac{d}{dt} \Big(I_{Hs} \cdot \cos \Big(\omega_s \cdot t + \gamma_{Ps} \Big) \Big) \right] + Rs \cdot \Big(I_{Hs} \cdot \cos \Big(\omega_s \cdot t + \gamma_{Ps} \Big) \Big) = \dots \\ &- Ls \cdot \omega_s \cdot I_{Hs} \cdot \sin \Big(\omega_s \cdot t + \gamma_{Ps} \Big) + Rs \cdot I_{Hs} \cdot \cos \Big(\omega_s \cdot t + \gamma_{Ps} \Big) \end{split}$$

$$(3.77)$$

$$V_{Ho_{Lo}}(t) = Lo \cdot \left[\frac{d}{dt} \left(I_{Ho} \cdot \cos\left(\omega_{o} \cdot t + \gamma_{Ho}\right) \right) \right] + Ro \cdot \left(I_{Ho} \cdot \cos\left(\omega_{o} \cdot t + \gamma_{Ho}\right) \right) = \dots$$

$$-Lo \cdot \omega_{o} \cdot I_{Ho} \cdot \sin\left(\omega_{o} \cdot t + \gamma_{Ho}\right) + Ro \cdot I_{Ho} \cdot \cos\left(\omega_{o} \cdot t + \gamma_{Ho}\right)$$

$$3.78$$

And voltage drops of arm inductances are:

$$v_{hns_{Lc}}(t) = v_{h\,Is_{Lc}}(t) = v_{h\,2s_{Lc}}(t) = Lc \cdot \left[\frac{d}{dt} \left(I_s \cdot \cos\left(\omega_s \cdot t + \gamma_{Hs} + \frac{\pi}{6}\right) \right) \right] + Rc \cdot \left(I_s \cdot \cos\left(\omega_s \cdot t + \gamma_{Hs} + \frac{\pi}{6}\right) \right) - Lc \cdot \omega_s \cdot I_s \cdot \sin\left(\omega_s \cdot t + \gamma_{Hs} + \frac{\pi}{6}\right) + Rc \cdot I_s \cdot \cos\left(\omega_s \cdot t + \gamma_{Hs} + \frac{\pi}{6}\right)$$

$$3.79$$

$$v_{h Io_{Lc}}(t) = Lc \cdot \left[\frac{d}{dt} \left(-I_o \cdot \cos\left(\omega_o \cdot t + \gamma_{Ho} + \frac{5\pi}{6}\right) \right) \right] + Rc \cdot \left(-I_o \cdot \cos\left(\omega_o \cdot t + \gamma_{Ho} + \frac{5\pi}{6}\right) \right) = \dots$$

$$Lc \cdot \omega_o \cdot I_o \cdot \sin\left(\omega_o \cdot t + \gamma_{Ho} + \frac{5 \cdot \pi}{6}\right) - Rc \cdot I_o \cdot \cos\left(\omega_o \cdot t + \gamma_{Ho} + \frac{5 \cdot \pi}{6}\right)$$

$$3.80$$

$$v_{h_{2o_{Lc}}}(t) = Lc \left[\frac{d}{dt} \left(-I_o \cdot \cos\left(\omega_o \cdot t + \gamma_{Ho} + \frac{\pi}{6}\right) \right) \right] + Rc \left(-I_o \cdot \cos\left(\omega_o \cdot t + \gamma_{Ho} + \frac{\pi}{6}\right) \right) = \dots$$

$$Lc \cdot \omega_o \cdot I_o \cdot \sin\left(\omega_o \cdot t + \gamma_{Ho} + \frac{\pi}{6}\right) - Rc \cdot I_o \cdot \cos\left(\omega_o \cdot t + \gamma_{Ho} + \frac{\pi}{6}\right)$$

$$3.81$$

The input and output side voltages can be defined by integrating their internal voltage drops as following:

$$V_{s}(t) = V_{Hs}(t) - V_{Hs}_{Ls}(t) = ..$$

$$V_{Hs} \cdot \cos(\omega_{s} \cdot t + \psi_{Hs}) - (-Ls \cdot \omega_{s} \cdot I_{Hs} \cdot \sin(\omega_{s} \cdot t + \gamma_{Ps}) + Rs \cdot I_{Hs} \cdot \cos(\omega_{s} \cdot t + \gamma_{Ps}))$$

$$3.82$$

$$V_{o}(t) = V_{Ho}(t) + V_{Ho_{Lo}}(t) = ..$$
$$V_{Ho} \cdot \cos(\omega_{o} \cdot t + \psi_{Ho}) + (-Lo \cdot \omega_{o} \cdot I_{Ho} \cdot \sin(\omega_{o} \cdot t + \gamma_{Ho}) + Ro \cdot I_{Ho} \cdot \cos(\omega_{o} \cdot t + \gamma_{Ho}))$$
3.83

Voltages from equations 3.82 and 3.83 are dependent on known variables as the source side voltages and currents, respectively V_H and I_H . Using voltage equations 3.79, 3.80, 3.81, 3.82 and 3.83 it is possible to calculate the maximum voltage value that each arm has to endure. Therefore, voltage design for each component has to be set. Moreover, this procedure will reveal the minimal voltage needed for each submodule and therefore for each capacitor in order to ensure the converter performances.

The individual voltage contribution for each arm are:

$$\begin{split} \upsilon_{h\,I}(t) &= -V_{s}(t) - \left[\upsilon_{h\,Is_{Lc}}(t) + \upsilon_{h\,Io_{Lc}}\right] + V_{o}(t) = \dots \\ &\cos(\psi_{Ho} + t \cdot \omega_{o}) \cdot V_{Ho} - \cos(\psi_{Hs} + t \cdot \omega_{s}) \cdot V_{Hs} + \dots \\ &I_{s} \cdot \left(\sqrt{3} \cdot Rs \cdot \cos(\gamma_{Ps} + t \cdot \omega_{s}) - Rc \cdot \cos\left(\frac{\pi}{6} + \gamma_{Hs} + t \cdot \omega_{s}\right) + Lc \cdot \omega_{s} \cdot \sin\left(\frac{\pi}{6} + \gamma_{Hs} + t \cdot \omega_{s}\right) - \sqrt{3} \cdot Ls \cdot \omega_{s} \cdot \sin(\gamma_{Ps} + t \cdot \omega_{s})\right) + \\ &I_{o} \cdot \left(Rc \cdot \cos\left(\frac{5 \cdot \pi}{6} + \gamma_{Ho} + t \cdot \omega_{o}\right) + \sqrt{3} \cdot Ro \cdot \cos(\gamma_{Ho} + t \cdot \omega_{o}) - Lc \cdot \omega_{o} \cdot \sin\left(\frac{5 \cdot \pi}{6} + \gamma_{Ho} + t \cdot \omega_{o}\right) - \sqrt{3} \cdot Lo \cdot \omega_{o} \cdot \sin(\gamma_{Ho} + t \cdot \omega_{o})\right) \end{split}$$

$$\begin{split} v_{h2}(t) &= V_{s}(t) + \left[v_{h2s_{Lc}}(t) + v_{h2o_{Lc}}(t) \right] - V_{o}(t) = \dots \\ & \cos\left(\psi_{Hs} + t \cdot \omega_{s}\right) \cdot V_{Hs} - \cos\left(\psi_{Ho} + t \cdot \omega_{o}\right) \cdot V_{Ho} + \dots \\ & I_{s} \cdot \left[Rc \cdot \cos\left(\frac{\pi}{6} + \gamma_{Hs} + t \cdot \omega_{s}\right) - \sqrt{3} \cdot Rs \cdot \cos\left(\gamma_{Ps} + t \cdot \omega_{s}\right) - Lc \cdot \omega_{s} \cdot \sin\left(\frac{\pi}{6} + \gamma_{Hs} + t \cdot \omega_{s}\right) + \sqrt{3} \cdot Ls \cdot \omega_{s} \cdot \sin\left(\gamma_{Ps} + t \cdot \omega_{s}\right) \right) + \\ & I_{o} \cdot \left[Lc \cdot \omega_{o} \cdot \sin\left(\frac{\pi}{6} + \gamma_{Ho} + t \cdot \omega_{o}\right) - \sqrt{3} \cdot Ro \cdot \cos\left(\gamma_{Ho} + t \cdot \omega_{o}\right) - Rc \cdot \cos\left(\frac{\pi}{6} + \gamma_{Ho} + t \cdot \omega_{o}\right) + \sqrt{3} \cdot Lo \cdot \omega_{o} \cdot \sin\left(\gamma_{Ho} + t \cdot \omega_{o}\right) \right) \end{split}$$

The essence of the converter remains on voltage equations 3.84 and 3.85. As it can be appreciated, it was supposed that the converter is connected between two power systems working at different frequencies. Therefore, system voltages are known. Supposing that the output source is replaced by an output impedance, then it must be considered only a change of the variable V_{Ho} in equations, where the new variable would be related to the output impedance, but this is out the sight of this work. The maximum tolerated voltages of this converter can be calculated with these equations, this is examined in the following subsection.

3.4.2 Voltage limitation design

Some limits have to be retained in order to guaranteeing the converter operation. Those limits depend on the operation of the converter, and the internal elements as inductances and capacitors have to tolerate those limits. They depend therefore on operation from each arm per phase.

An extreme operation case can be supposed in order to set a security margin for the converter. This operation is evaluated when all amplitudes are at their maximum values, with every sinusoidal component equal to one, and all terms are positives. This is a required but not decisive condition for guaranteeing the converter function [BaMB14].

Then, the maximum value will be the same for both arm voltages from equation 3.84 and 3.85, and can be expressed with the following equation:

$$\begin{bmatrix} v_{h}(t) \end{bmatrix}_{max} = \begin{bmatrix} v_{hIs}(t) \end{bmatrix}_{max} + \begin{bmatrix} v_{hIo}(t) \end{bmatrix}_{max} = \begin{bmatrix} v_{h2s}(t) \end{bmatrix}_{max} + \begin{bmatrix} v_{h2o}(t) \end{bmatrix}_{max}$$

$$\begin{bmatrix} v_{h}(t) \end{bmatrix}_{max} = \sqrt{\left(V_{Hs}\right)^{2} + \left(Rc \cdot I_{s}\right)^{2} + \left(Lc \cdot \omega_{s} \cdot I_{s}\right)^{2} + \left(\sqrt{3} \cdot Rs \cdot I_{s}\right)^{2} + \left(\sqrt{3} \cdot Ls \cdot \omega_{s} \cdot I_{s}\right)^{2} + \dots$$

$$3.86$$

$$\sqrt{\left(V_{Ho}\right)^{2} + \left(Lc \cdot \omega_{o} \cdot I_{o}\right)^{2} + \left(Rc \cdot I_{o}\right)^{2} + \left(\sqrt{3} \cdot Ro \cdot I_{o}\right)^{2} + \left(\sqrt{3} \cdot Lo \cdot \omega_{o} \cdot I_{o}\right)^{2}}$$

Equation 3.86 defines the maximum voltage that each arm has to tolerate in a given moment for a given operation point, in steady and balanced state. Moreover, equation 3.16 could be introduced too in order to calculate the maximum allowed voltage for each capacitor. Following the same procedure, it can be calculated the maximum voltage of the arm inductances. These voltages will depend on the nominal current value of the converter as expressed in the following equation:

$$\begin{bmatrix} v_{h_{Lc}}(t) \end{bmatrix}_{max} = \begin{bmatrix} v_{h_{Is_{Lc}}}(t) \end{bmatrix}_{max} + \begin{bmatrix} v_{h_{Io_{Lc}}}(t) \end{bmatrix}_{max} = \begin{bmatrix} v_{h_{2s_{Lc}}}(t) \end{bmatrix}_{max} + \begin{bmatrix} v_{h_{2o_{Lc}}}(t) \end{bmatrix}_{max} = \dots$$

$$\sqrt{\left(Lc \cdot \omega_{s} \cdot I_{s}\right)^{2} + \left(Rc \cdot I_{s}\right)^{2}} + \sqrt{\left(Lc \cdot \omega_{o} \cdot I_{o}\right)^{2} + \left(Rc \cdot I_{o}\right)^{2}}$$

$$3.87$$

This equation gives the relationship between voltage and currents, which are related to the maximum value operation point. These limits have to be taken in account for physical design of the MMC, an example will represent these limits when a high voltage system is used (see Table 3.5).

The maximum voltage values of inductances and capacitors will be tested using equations 3.86 and 3.87. The effect on each variable can be appreciated with graphical support as presented next. It is supposed an OPS working at 250 kV and distributing 100 MW. A same power level of a MMC was evaluated in [SABC10] and values are presented next.

Defined Variables	Evaluated Variables
$V_{Hs} = \frac{\sqrt{3}}{\sqrt{2}} \cdot 250kV$	V_{Ho}
P = 3.100 MW	I_o
$I_s = \frac{1}{\sqrt{2}} \cdot 800A$	Lc
$\omega_s = 2 \cdot \pi \cdot 50$	-
$\omega_o = \frac{\omega_s}{3}$	-
Ls = Lo = 5 mH	-

Table 3.5. Used variable for maximum voltage evaluation in high voltage and high power.

First, the same voltage level at both sides is supposed, then, application of equation 3.86 using variables presented in Table 3.5 gives:

$$\left[\upsilon_h(t)\right]_{max} = 622kV \tag{3.88}$$

For regular operations, equation 3.88 represents the maximum voltage value that the arm of the presented converter would have to generate (in steady and balanced state) at nominal power. It is called maximum voltage because at lower power, the required voltage would be lower too. Therefore, if the converter arm is capable to generate more voltage than the one presented in 3.88, then more kinds of operations, as the one exposed next, would be allowed.

The output voltage can vary supposing two cases:

- 1- With constant output, power no matters the needed current, this evaluates the behaviour when the converter has to keep the same output power while the output voltage changes its value.
- 2- Using a constant current of the line no matters the needed output voltage (V_o) nor the power; only for evaluation purposes as a current source, normally not related to a power system.

The obtained results are presented in the following figures.



Figure 3.19. Output voltage variations evaluated when: (a) constant power and constant current; (b) constant power using different values of Lc in (H).

As it can be appreciated in Figure 3.19.a, depending on the tested case, the maximum allowed arm voltage would have a different behaviour when the output voltage decreases. If the current is kept constant, then converter can vary its output voltage from zero to a defined upper limit. In this case, it was supposed that the upper limit is the one from equation 3.88, like this, the maximum allowed power is 100 MW.

A second behaviour can be observed at constant power operation. In this case, the output voltage should be limited with a lower and an upper limit. This represents the normal behaviour of the converter when supposing over voltages or voltages drops: if the limits are crossed, then the converter might not be able of continuing its operations. Therefore, these conditions should be avoided in order to keep stable the energy contained in capacitors.

A variation of the internal converter inductances will reveal a voltage drop effects, calculated with equation 3.87. This is presented in Figure 3.19.b, where for inductances lower than 0.01 H almost does not change the voltage behaviour, but when raising the inductance values, the allowed range of voltage became smaller. Therefore, the maximum voltage limit will not be affected by the inductance values that are commercially used for the MMC, as they are around some small inductance values [Opal17a, SABC10]. A small difference in the voltages is actually appreciated for inductance values bigger than 0.1 H.

The same experiment is repeated but this time the input and the output voltages were kept at the same value while the output current value was variated.



Figure 3.20. Output current variations evaluated with constant output voltage using different values of Lc.

The maximum current value of the converter become smaller when a higher inductance value is used, this will therefore reduce the maximal power capability of the converter.

The best voltage capability of the converter is when the lowest inductance value is used in the converter. An inductance is required for this converter because two threephase systems have to work together in the same link. An inductance value until 0.01 H seems to be satisfactory for the voltage capability of the converter.

The presented system in Table 3.5 will be used too for testing the control in chapter 4 and limits presented in this section will be verified with simulations.

3.5. Conclusions

In this section was presented the structure and the model of a MMC working as a frequency changer in a direct AC-AC topology. Other authors suggested the use of the MMC in a back-to-back configuration in order to apply it in offshore (as presented in section 2.4.3), but this solution was not suitable because it doubles the required components when compared to HVDC. This is not the case of the proposed solution in the present chapter, where a single MMC is used to convert frequency with the same number of components that in HVDC.

The only difference between the components of a MMC HVDC and a MMC AC-AC is the needed full-bridge components in submodules. It is considered as a drawback because it rises the conduction losses in the converter arms, nevertheless this losses are not so high when compared to the total efficiency of the converter and can be up to 0.5% more losses than the half-bridge case [SJLJ16]. However, multiple advantages are found in a full-bridge component as fault ride-through capability, which is desired for OPS application.

The model of the converter was defined from its average model, which neglects the high frequencies resulting from the switching model. This is not a problem because those frequencies are unwanted and are normally eliminated from the system. The elimination can be done thanks to the arm inductances of the converter.

The proposed modelling stands in the superposition theorem plus a three-phase transformer analogy that allow the study at each frequency individually. Some preliminary results from the perfect model showed the feasibility of the converter. The converter presents relative high difficulty in its model. Nevertheless, the implementation of some analogies simplified the link between the internal current values with the measured parameters, which is different from other AC-AC converters.

A balanced three-phase system was considered for the proposed model and this is certainly not true for all kind of operations that the converter may face. It is therefore essential to discuss about an automatic control for normal operations of the converter and for compensate instabilities that will be introduced in the following chapter.

Chapter 4. Frequency Changer Control

In chapter 3, it was evaluated the feasibility of the MMC when it is configured in AC-AC topology, and in this section it will be presented the applied control and its operations on simulation. Studied controls for the one-phase topology [GlMa05] and the one working in AC-DC, or DC-AC operation [Akag11, BBEL13, GnGJ11], generally they present the same structure: the MMC output-side supplies a required load power. In consequence, the energy level contained in the submodules capacitors will be reduced. The input-side of the converter compensates these energy changes and charges the capacitors. In both topologies, the control must allow to work in the other direction, where the load supplies the energy, which is dissipated by the source-side. These strategies will be applied in this work for the proposed converter.

In this chapter, the term of circulating current will be introduced. The circulating current value was supposed to be zero at the modelling stage. However, this current exists because of some irregularities in the system (physical and control one) and its behaviour cannot to predicted, but it can be minimized. Some authors propose the suppression of this current by applying control strategies that are related to the second harmonic of the fundamental frequency [AASI11, QiZL11]. Some other authors propose the circulating current control in order to improve the balance of the system [LZXY16, YaSa17]; they use this value in order to compensate the energy charges and discharges in capacitor due to transients, they are also based in the second harmonic of the fundamental frequency.

However, not only the circulating current value should be suppressed in the arm of the converter, but also the highest harmonics caused by the switching frequency that can be eliminated with the arm inductances [LiJW17]. In the DC case, the capacitor voltages must be equally distributed in order to reduce harmonic distortions at the AC side [DQBS14]. A balancing control algorithm will be used to keep all submodules energies at the same level.

The circulating current value is an unwanted and however needed current for the MMC operations. For the MMC in AC-DC reversible topology, the circulating current value is normally not delivered out of the converter and therefore it raises the converter power losses.

A direct AC-AC MMC with YDY coupling can tolerate inequalities in the capacitor voltages distribution because any possible generated harmonic will remain inside the delta. In addition, the proposed strategy works directly with a resonant controller, which simplifies the implementation task and presents excellent waveform quality. It will be presented that, because of the frequency conversion, arm voltages are naturally different and therefore capacitor voltages will not have the same waveforms and a new strategy for keeping voltages at the same level will be proposed.

The bases of the control and its general application in the MMC in AC-AC configuration will be introduced in this chapter. Then, it will be introduced the model for calculating the average energy of the converter and the generated circulating current value. For these values, an energy control strategy will be proposed and evaluated; it avoids energy divergence from the use of reactive power, and a circulating current suppression strategy will be presented. Finally, the balancing control algorithm applied to the switching model will be presented at the end of this section with some simulation results.

4.1. Averaged Model - General Control Schemes

Since the studied MMC AC-AC in direct configuration is represented by a relatively new topology, every used control strategy will be prudently defined. These control strategies were developed for application of the proposed converter that, in most of the cases, are different from the AC-DC topology as it will be specified and compared further in every specific case.

From the developed model presented in chapter 3, all arm currents can be considered as known and therefore they are controllable. In consequence, the converter representation can be simplified as following:



Figure 4.1. Simplification of the MMC model representation in YDY configuration.

The main variables related to the control of the MMC are presented in Figure 4.1:

- Input and output side currents and voltages (*I*_{*Hs*}, *I*_{*Ho*}, *V*_{*Hs*}, *V*_{*Ho*} respectively), measurements used to calculate power;
- Capacitors voltages ($\nu_{e\Sigma hn}$) that are used for measure the energy of each submodule;
- The resulting modulation index (m_{hn}) , which value remains between -1 and +1 and will control the energy of each submodule;
- Finally arm currents (*i*_{hn}) that have many functions as the link between sourceside currents and contained energies in capacitors; the link between these energies and the output current of the converter; and calculation and control of the circulating currents.

4.1.1 Control Application

As it is explained in appendix A.1.1, it is supposed that the energy being processed by the MMC in AC-AC configuration is coming from offshore wind resources. The energy available at the input side minus some losses is used as a reference for the output side of the converter. This process is resumed by giving a reference of the output apparent power (S_0^*).

The output power is evaluated from the measures of the line currents and voltages at the output side. This power is generated from the stored energy in the submodules capacitors. The available energy is calculated from the voltage measured at each submodule capacitor and the compensation is made by controlling the current at the input-side. As the relationship between input and output currents was showed to be no linear in chapter 3, the proposed model decouples the input from the output side, allowing an individual control implementation.

The objectives of the proposed converter can be resumed as it follows:

- Energy from AC 50 Hz using a power reference at the output side will be converted to AC 50/3 Hz from a given stocking energy reference.
- Energy going in the other direction is tolerated as well.
- To avoid any divergence of the capacitor energy levels using reactive power.
- To suppress the harmonic content in the input and output signals by keeping any generated harmonic inside the delta.

In order to achieve those objectives some control strategies must be applied. They are explained in the following subsections.

4.1.2 Output current reference calculations

With the given output power reference and a required reactive power value, the converter can therefore start its operation by calculating the equivalent output current reference as showed in the following Figure 4.2. Input variables are related to the energy of the converter, which will be presented further in section 4.4.



Figure 4.2. Output current reference calculation from power reference.

In Figure 4.2 is represented a current calculation using complex theory with the apparent power (S) and power factor $(cos(\varphi))$ as presented in equation 4.1. The output reactive power is zero by default, and its value can be adjusted depending on the required uses of the MMC: for instance, the reactive power is useful to maintain the voltage level of the line at any required level. In this case, the output power angle (φ^*_{o}) is automatically adjusted from voltage measurements. Results will show how the converter is able to follow the imposed reference of angle.

$$\overline{I^{*}_{o}} = \frac{S^{*}_{o} \cdot \cos(\varphi^{*}_{o})}{\sqrt{3} \cdot |V_{o}|} + j \frac{S^{*}_{o} \cdot \cos(\varphi^{*}_{o})}{\sqrt{3} \cdot |V_{o}|}$$

$$4.1$$

The amplitude of the current with the given angle is converted in a sinusoidal reference using the output frequency (ω_0), measured from the line voltage with the aid of a phase lock loop (PLL).

A PLL is based in an oscillator that can provide a variable frequency compared to a source reference coming from a quartz or from another PLL. There are several PLL architectures divided in three groups: analogical, digital and mixed [Kief14]. More information can be found in appendix D.1.

This calculated output current reference is used in the first level control of the MMC: first, compared to the output current and then, corrected using a resonant controller. This value is required for the control equations 3.37 and 3.38, which will be used to generate the modulation indexes for controlling submodules.

4.2. Principal control: resonant controller

Using the output voltage measurements, an output current value is therefore calculated and used as a reference to the converter. This current is in alternative form, and should be at the output frequency only. In this current however, it is possible to find some other frequencies as the source frequency, this is because of the possible irregularities in the physical model. For instance, they exist because some small differences between all inductances or all capacitances, switching transients or even dynamics states. The used controller has to be able to correct signals containing many harmonics and, at the same time, to efficiently track a signal that is at only one frequency. These unwanted frequencies should be at first neglected and then suppressed

It was found that the most suitable corrector for this kind of situations is the multiresonant corrector [DBRG17, LGEB06]. Using general words, it is capable to correct the error signal in order to produce the required reference waveform at the required frequency. Multi-frequency reference signals can also be used with this corrector.

The resonant corrector avoid the use of the Park's transformation, which is usually needed for the MMC control when is connected in AC-DC topology as presented in Figure 2.22. This requires more computational resources [PiME09], it is less dynamic because the signal should be filtered and can use only one frequency value. Resonant controllers instead are capable of tracking sinusoidal references, of arbitrary frequencies, with multiple frequency vales at the same time, of both positive and negative sequence with zero steady-state error.

For this converter, where at least two frequencies exist in one link, it is therefore convenient to use resonant corrector. A classical multi-resonant structure consist of the addition of single resonant controllers in parallel configuration [GHEB05, LGEB06].

4.2.1 Resonant corrector

A resonant corrector is a mathematical equivalent of the PI corrector working in a frame reference [LGEB06], but without the use of a transformation. The corrector moves the frequency reference, ideally at the same frequency value of the alternative input signal (ω_i) and therefore the PI corrector can be transformed in a complex resonant corrector (C_R) as it follows:

$$C_R(S) = kp + \frac{ki}{S + j\omega_i} + \frac{ki}{S - j\omega_i} = kp + \frac{2 \cdot ki \cdot S}{S^2 + \omega_i^2}$$

$$4.2$$

The C_R frequency response in a stationary frame is not symmetrical when compared to the origin because of its complex nature. It is modelled by a pronounced gain and a null phase at the input frequency (see Figure 4.3). Sinusoidal current regulation is commonly used to achieve a high level of performance in many applications as AC-AC converters [GDJG11].



Figure 4.3. Frequency response for a CR controller [Etxe09].

The maximum gain value is obtained at the wanted frequency ω_i , and therefore, it is possible to reduce the error of alternative signal turning at ω_i as presented in Figure 4.3. For parameters adjustments, the maximum peak can be placed at a positive or at a negative gain, by consequence the corrector is capable of withdraw the positive of negative sequence of a given frequency.

In order to calculate the closed loop of the system when a resonant corrector is applied, it is presented the transfer function from each arm; it was calculated from Figure 4.1:

$$H_{am(S)} = \frac{i_{hn}(S)}{v_{hn}(S)} = -\frac{1}{Lc} \cdot \frac{1}{S + \frac{Rc}{Lc}}$$
4.3

The closed loop response of the system when using a C_R is:

$$H_{(arm_{CL})} = \frac{C_{CR}(S) \cdot H_{arm}(S)}{1 + C_{CR}(S) \cdot H_{arm}(S)} = \frac{kp \cdot S^2 + 2 \cdot ki \cdot S + kp \cdot \omega_i^2}{Lc \cdot S^3 + Rc \cdot S^2 + S^2 \cdot kp + Rc \cdot \omega_i^2 + kp \cdot \omega_i^2 + 2 \cdot S \cdot ki + Lc \cdot S \cdot \omega_i^2} \quad 4.4$$

The maximum gain from equation 4.4 is obtained at the frequency ω_i , and therefore the static error is zero [Etxe09]. The corrector parameters are therefore calculated at two frequencies; the input and output system frequencies, using the arm inductance and resistor values.

4.2.2 Calculation of the resonant controller parameters

In order to avoid the integration of unwanted harmonics into the corrector response, as for instance those harmonics coming from system perturbations, the corrector must be correctly adjusted. It is therefore used a method based on the open-loop frequency response of the system in order to guarantee the stability and to minimize the action from other harmonics. The complete methodology is presented by ETXEBERRIA-OTADUI in [Etxe09].

As presented in equation 4.2, two parameters must be adjusted: kp and ki. The proportional one cannot be specified to only one harmonics because its influence is the same for the entire frequency range. By consequence, the system is adjusted with one proportional parameter and at least one integrator placed in parallel. This integrator parameter is adjusted at the required frequency and, multiple integrators can be added in parallel, as frequencies are required to control or to eliminate. This way, a multi-resonant controller could be built.

First, it is defined the allowed band in open loop (BP_{BO}) , which is imposed by the proportional parameter:

$$kp = L \cdot \sqrt{\left(BP_{BO}\right)^2 - \left(\frac{Rc}{Lc}\right)^2}$$

$$4.5$$

$$PM = 180^{\circ} - atan \left(\frac{Lc}{Rc} \cdot BP_{BO}\right) - \frac{3}{2} \cdot BP_{BO} Ts \cdot \frac{180^{\circ}}{\pi}$$

$$4.6$$

In equations 4.5 and 4.6, the term Ts is the sample time of the system, which is defined as the discrete time intervals used to sample a continuous-time signal, and subsequently converted to a sequence of digital values for processing [Carw03]. It is used to calculate the Phase Margin (PM) of equation 4.6: the allowed BP_{BO} is chosen by taking into account the system stability related to a positive PM. A recommended range for PM stability is between 30° and 60° [Ogat10]. Then Ts is chosen from the sample time used in simulation [Etxe09] or the sampling frequency of the switching devices that will be introduced later in section 4.7. As example, two Ts are therefore compared: 100 µs and 200 µs.

Considering the following internal parameters of the arm converter, then the stability band can be calculated with equation 4.5 and represented in Figure 4.4.

Table 4.1. Considered parameters for phase-margin and allowed band calculations

	Arm
Inductance [mH]	7.5
Resistance $[\Omega]$	0.1



Figure 4.4. Calculation of the stability band for the MMC.

As it can be appreciated, a simple time of 200 μ s would be critically stable with the minimum allowed phase margin. This means that the system would be too much sensible to parameters mismatches. A 100 μ s system instead has an allowed band going from 3500 rad/s to 7000 rad/s, offering more robustness.

In order to avoid some problems, as unwanted frequencies during transient states, ki parameters are calculated by taking into account other harmonic correctors used in the same system: as 50 Hz and 50/3 Hz are the main frequencies used, their interharmonic distance is $\Delta \omega = 210 \text{ rad/s}$. Then, the gain from one harmonic corrector must limit the gain from the other harmonic frequency. Therefore, ki can be calculated as it follows [Etxe09]:

$$ki = \Delta \omega \cdot \sqrt{\left(\Delta G + kp\right)^2 - kp^2}$$

$$4.7$$

$$\Delta G = \left| C_{CR}(\omega_{ii}) \right| - kp \tag{4.8}$$

Where ω_{ii} is the other used harmonic frequency: for instance, when calculating the corrector for the 50 Hz-side, then the other harmonic frequency would be the 50/3 Hz-one. ΔG must be chosen low enough in order limit the gain of the other harmonic frequency. It was therefore selected a value of $\Delta G = 5 \cdot 10^{-3}$, which does not affect the performances of the given corrector faced to other frequencies. The open-loop response for each corrector can now be obtained and it is presented in the following figure.



Figure 4.5. Open-Loop response for the resonant corrector.

Both resonant correctors were chosen for each system frequency in order to follow the frequency of the current reference and reject any other contained harmonic in the measure. When comparing to the closed-loop response from Figure 4.6, the gain will be one only at the selected frequency and the steady state error would be zero.



Figure 4.6. System closed loop response using resonant correctors for each frequency.

As it can be appreciated in Figure 4.6, the 50 Hz corrector rejects every system frequency until the selected one (314.159 rad/s), where also the phase shift becomes zero; the output signal will not be phase-shifted from input signal, allowing an instantaneously response having consequently a very dynamic controller. The same response is obtained for the 50/3 Hz corrector, in this case all frequency are rejected except for 104.72 rad/s where the phase-shift is also zero. Both responses are combined and the final response would be the sum of both of them. It can be noted that the phase at 50/3 Hz is around -360°, which is also 0°, and then at 50 Hz it is zero too.



Figure 4.7. System closed loop response using resonant correctors that combine both system frequencies.

The system can now be controlled using a perfect model; this model represents the ideal response of the system and this is presented in the following section.

4.3. Simulation results I: evaluation using the perfect model

In this section, it will be studied the first simulations results made with the perfect model of the MMC. This model does not have the submodules, which have been replaced by voltage sources controlled directly by the modulation voltage. This model is useful to make the main adjustments of the converter: output current calculation, performances of the resonant controller and verification of equations presented in chapter 3.

Simulation results from the perfect model will be presented in this section using only a proposed prototype converter at 300 kV. More information about models and a comparison between them can be found in [BaMB14, Berg15, ZBBF16]

4.3.1 Perfect model and basic performances

The control schematics of the converter at this stage are represented in Figure 4.8. It can be appreciated both, the input and the output side of the converter, with their respective variables and the controlled voltage sources connected to each arm of the converter.

The input current value is related to changes of energy, as it will be presented further in section 4.4, consequently for these simulation results, the input current value was calculated directly from the output current value, and therefore results give only a general idea of its behaviour where its value may not be accurate. Nevertheless, this is not a drawback for this first stage of simulations.



Figure 4.8. Control application to the MMC perfect model.

For the offshore prototype, it is proposed to use an adapted data for that end:

Table 4.2.	MMC	AC-AC for	offshore	application	[SABC10]
1 4010 1.2.	111110	110 110 101	011011010	application	

Nominal AC-AC Voltage	250 kV peak per phase
Output power	750 MW
Inductance value	5 mH
Arm resistance	$2 \text{ m}\Omega$
Equivalent arm capacitance	$130 \ \mu F$

For the presented simulation using a perfect model, capacitor sub-modules are not needed at this stage; it will be used for presenting the average model in section 4.6. Simulation results will reveal the operation of the resonant correctors and the modulation waveforms resulted of the proposed converter control.

4.3.2 Perfect model: steady-state waveforms

First, it is presented the obtained arm voltage, which is directly related to the modulation index as explained in equation 3.16.



Figure 4.9. One cycle of modulation waveforms in steady state.

These waveforms are the results of applying equations 3.41 and 3.42. Each waveform controls each arm voltage of the converter, they are showed in the following Figure 4.9, and the order was chosen in order to represent together arm voltages connected to each phase of the source side (see Figure 3.11: phase As with a1 and c2; phase Bs with b1 and a2; and phase Cs with c1 and b2).

One of the most interesting behaviours of the MMC operated in direct AC-AC is the modulation presented Figure 4.9 that is different from the conventional AC-DC one (see Figure 4.10). This is because the resulting arm voltages are all different between them. This behaviour can be described mathematically and it is presented in subsection 4.3.4. Moreover, because the output frequency is 1/3 of the input one, these cycles will always be repeated after 60 ms and each voltage will keep its own waveform. Other frequency value will modify the cycle of repetition. For instance, using 16.7 Hz will repeat the cycle every 10 seconds, and voltage will be better distributed in all six phases during this cycle. An example of transforming 60 Hz will be presented further in this section.



Figure 4.10. Modulation index for a MMC in AC-DC reversible operation [Berg15].

In AC-DC configuration, an example of arm voltages related to the modulation index is presented in Figure 4.10, where all voltages are the same and only some distortions are introduced to the waveforms because of the use of a DC link. This allows an easier balancing of capacitors between arms.

In steady state, the resonant correctors work with almost no error as presented in Figure 4.11. The input error is at 2 mA peak, while the output error is 0.1 μ A, this difference is related to the parameters of the resonant controller that were calculated separately for each frequency side, also it could include some mathematical errors from the software around 10⁻⁶ order (related to the used time step). The presented correctors generate the waveforms corresponding to the arm inductance voltages.



Figure 4.11. Waveforms of the error and the resonant corrector output in steady state.

Presented waveforms can have more than one frequency as presented in Figure 4.12.



Figure 4.12. Harmonic content in the corrector waveforms.

The corrector introduces both system frequencies in waveforms, which is normal because voltage of arm inductances are composed of both frequencies of the system. Moreover, the corrector is able to generate the required harmonics in order to keep the input and the output at their own frequency. This behaviour can be compared to the transient state response, where the corrector must generate the same output signal even when changes are produced in the system.

4.3.3 Perfect model: transient-state waveforms

Supposing that the system is placed at steady state and maximum power, and the power is decreased at 37 seconds, in Figure 4.13 it can be observed how the controller follows very fast the changes in the reference. It was used a load generation for the input side of the converter (that consumes power) and generation convention for the

output side (that supplies power), both values will therefore have the same value in normal operations.



Figure 4.13. Variations of power in the system.

In Figure 4.14, it can be seen how the error in the corrector raises when the output current value is different from the reference, therefore the output signal from the corrector must be reduced, this can be noticed over many cycles but this is not the objective of this analysis.



Figure 4.14. Resonant corrector response faced to transients.

The idea is to take only one cycle of the low frequency value, from each signal from Figure 4.14. Then, to calculate the harmonic content during the transient. In the resulting Figure 4.15, it can be appreciated an increase of the inter-harmonic content which are related to the changes of reference in the systems. Nevertheless, when observing the correctors output at one cycle of the transient-state, and comparing them in steady state from Figure 4.12 before, it is observed the same harmonic content. This means that the controller is rejecting all intrudes frequency harmonics coming from both systems, which is required to keep stable the system.



Figure 4.15. Harmonic content in the corrector waveforms with transients.

The obtained results proved the good dynamics of the used resonant controllers. In addition, the converter was able to transfer power from the one side to the other. The following step is therefore the introduction the sub-module capacitors into the model and definition of the associated control strategies.

4.3.4 Modulation voltage calculations

Usually, voltages and current waveforms are at only one frequency. A voltage turning at two different frequencies requires a special attention as the voltage distribution in the three phases system might not be the same; an example was presented in the Figure 4.9.

This can be proved with the application of Kirchhoff's voltage laws. It is shown that every voltage waveform will depend on two parameters. The first one will be the phase shift reference of the entire system from where voltages at different frequencies are referenced: this could be seen as a fictitious slack bus at the 0° reference of voltage, and every voltage side at the input and output of the converter can be at the same 0° reference.

For convenience, it is supposed that every inductance voltage drop is condensed into one variable called Δv_{hn} . Then, it is supposed that the voltage drop due to each source is taken into account for each voltage source, and therefore the voltage variable is measured directly from the connection point between the sources and the formed delta of the converter MMC in YDY topology, this is presented in the following figure.


Figure 4.16. MMC AC-AC: design for arm voltage calculations.

Variables where already defined with Figure 3.5. Then, all equations can be expressed as it follows:

$$V_{\text{Hs}}(t) = V_{\text{s}} \cdot \cos(\Psi_{\text{s}} + t \cdot \omega_{\text{s}})$$

$$4.9$$

$$V_{\text{Ho}}(t) = V_{\text{S}} \cdot \cos(\Psi_{\text{O}} + t \cdot \omega_{\text{O}})$$

$$4.10$$

The voltage of each arm is therefore:

$$V_{As}(t) - V_{Ao}(t) = v_{a1}(t) - \Delta v_{a1}(t)$$
 4.11

$$V_{Bs}(t) - V_{Bo}(t) = v_{b1}(t) - \Delta V_{b1}(t)$$
 4.12

$$V_{Cs}(t) - V_{Co}(t) = v_{c1}(t) - \Delta V_{c1}(t)$$
4.13

$$V_{Ao}(t) - V_{Bs}(t) = v_{a2}(t) - \Delta V_{a2}(t)$$
 4.14

$$V_{Bo}(t) - V_{Cs}(t) = v_{b2}(t) - \Delta V_{b2}(t)$$
4.15

$$V_{Co}(t) - V_{As}(t) = v_{c2}(t) - \Delta V_{c2}(t)$$
4.16

Then, introducing equations 3.41 and 3.42 it goes:

$$v_{a1}(t) - \Delta v_{a1}(t) = V_{s} \cdot \left(\cos\left(\Psi_{s} + t \cdot \omega_{s}\right) - \cos\left(\Psi_{o} + t \cdot \omega_{o}\right) \right)$$

$$4.17$$

$$v_{b1}(t) - \Delta v_{b1}(t) = V_s \cdot \left(\cos\left(\Psi_s + t \cdot \omega_s - \frac{2\pi}{3}\right) - \cos\left(\Psi_o + t \cdot \omega_o - \frac{2\pi}{3}\right) \right)$$

$$4.18$$

$$v_{c1}(t) - \Delta v_{c1}(t) = V_{s} \cdot \left(\cos\left(\Psi_{s} + t \cdot \omega_{s} + \frac{2\pi}{3}\right) - \cos\left(\Psi_{o} + t \cdot \omega_{o} + \frac{2\pi}{3}\right) \right)$$

$$4.19$$

$$v_{a2}(t) - \Delta v_{a2}(t) = V_{s} \cdot \left(\cos\left(\Psi_{o} + t \cdot \omega_{o}\right) - \cos\left(\Psi_{s} + t \cdot \omega_{s} - \frac{2\pi}{3}\right) \right)$$

$$4.20$$

$$v_{b2}(t) - \Delta v_{b2}(t) = V_{s} \cdot \left(\cos\left(\Psi_{o} + t \cdot \omega_{o} - \frac{2\pi}{3}\right) - \cos\left(\Psi_{s} + t \cdot \omega_{s} + \frac{2\pi}{3}\right) \right)$$

$$4.21$$

$$v_{c2}(t) - \Delta v_{c2}(t) = V_{s} \cdot \left(\cos \left(\Psi_{o} + t \cdot \omega_{o} - \frac{2\pi}{3} \right) - \cos \left(\Psi_{s} + t \cdot \omega_{s} \right) \right)$$

$$4.22$$

Finally, it is applied a cosine trigonometric law and equations were expanded:

$$v_{a1}(t) - \Delta v_{a1}(t) = Vs \cdot \cos(t \cdot \omega_{s}) \cdot \cos(t \cdot \omega_{o}) \cdot \cos(t \cdot \omega_{o}) - Vs \cdot \sin(t \cdot \omega_{s}) \cdot \sin(t \cdot \omega_{s}) + Vs \cdot \sin(t \cdot \omega_{o}) \cdot (\pi \cdot \omega_{o}) \cdot \sin(t \cdot \omega_{o}) \cdot \sin(t \cdot \omega_{o}) \cdot (\pi \cdot \omega_{o$$

$$\frac{v_{b1}(t) - \Delta v_{b1}(t) =}{\frac{Vs \cdot \cos(t \cdot \omega_{0}) \cdot \cos(\Psi_{0})}{2} - \frac{Vs \cdot \cos(t \cdot \omega_{s}) \cdot \cos(\Psi_{s})}{2} + \frac{Vs \cdot \sin(t \cdot \omega_{s}) \cdot \sin(\Psi_{s})}{2} - \frac{Vs \cdot \sin(t \cdot \omega_{0}) \cdot \sin(\Psi_{0})}{2} + \frac{\sqrt{3} \cdot Vs \cdot \cos(t \cdot \omega_{s}) \cdot \sin(\Psi_{0})}{2} - \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_{0}) \cdot \sin(\Psi_{0})}{2} - \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_{0}) \cdot \cos(\Psi_{0})}{2} + 4.24$$

$$v_{b2}(t) - \Delta v_{b2}(t) = \frac{Vs \cdot \cos(t \cdot \omega_s) \cdot \cos(\Psi_s)}{2} - \frac{Vs \cdot \cos(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} - \frac{Vs \cdot \sin(t \cdot \omega_s) \cdot \sin(\Psi_s)}{2} + \frac{Vs \cdot \sin(t \cdot \omega_o) \cdot \sin(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \cos(t \cdot \omega_s) \cdot \sin(\Psi_s)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_s)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(\psi_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \cos(\psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \cos(\psi_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \sin(\psi_o) \cdot \cos(\Psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \cos(\psi_o) \cdot \cos(\psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \cos(\psi_o) \cdot \cos(\psi_o)}{2} + \frac{\sqrt{3} \cdot Vs \cdot \cos(\psi_o)}{2} + \frac$$

$$v_{a2}(t) - \Delta v_{a2}(t) = \frac{Vs \cdot \cos(t \cdot \omega_s) \cdot \cos(\Psi_s)}{2} + Vs \cdot \cos(t \cdot \omega_o) \cdot \cos(\Psi_o) - \frac{Vs \cdot \sin(t \cdot \omega_s) \cdot \sin(\Psi_s)}{2} - Vs \cdot \sin(t \cdot \omega_o) \cdot \sin(\Psi_o) - \frac{\sqrt{3} \cdot Vs \cdot \cos(t \cdot \omega_s) \cdot \sin(\Psi_s)}{2} - \frac{\sqrt{3} \cdot Vs \cdot \sin(t \cdot \omega_s) \cdot \cos(\Psi_s)}{2} - \frac{4.26}{2}$$

4.25

$$v_{b2}(t) - \Delta v_{b2}(t) = \frac{Vs \cdot cos(t \cdot \omega_{s}) \cdot cos(\Psi_{s})}{2} - \frac{Vs \cdot cos(t \cdot \omega_{o}) \cdot cos(\Psi_{o})}{2} - \frac{Vs \cdot sin(t \cdot \omega_{s}) \cdot sin(\Psi_{s})}{2} + \frac{Vs \cdot sin(t \cdot \omega_{o}) \cdot sin(\Psi_{o})}{2} + \frac{\sqrt{3} \cdot Vs \cdot cos(t \cdot \omega_{s}) \cdot sin(\Psi_{s})}{2} + \frac{\sqrt{3} \cdot Vs \cdot sin(t \cdot \omega_{o}) \cdot sin(\Psi_{o})}{2} + \frac{\sqrt{3} \cdot Vs \cdot sin(t \cdot \omega_{o}) \cdot cos(\Psi_{o})}{2} + \frac{\sqrt{3} \cdot Vs \cdot sin(t \cdot \omega_{o}) \cdot cos(\Psi_{o})}{2} + \frac{\sqrt{3} \cdot Vs \cdot sin(t \cdot \omega_{o}) \cdot cos(\Psi_{o})}{2} + \frac{\sqrt{3} \cdot Vs \cdot sin(t \cdot \omega_{o}) \cdot sin(\Psi_{o})}{2} + \frac{\sqrt{3} \cdot Vs \cdot sin(t \cdot \omega_{o}) \cdot cos(\Psi_{o})}{2} + \frac{\sqrt{3} \cdot Vs \cdot sin(t \cdot \omega_{o}) \cdot cos(\Psi_{o})}{2} + \frac{Vs \cdot sin(t \cdot \omega_{o}) \cdot sin(\Psi_{o})}{2} + \frac{Vs \cdot$$

$$v_{c2}(t) - \Delta v_{c2}(t) = Vs \cdot sin(t \cdot \omega_{s}) \cdot sin(\Psi_{s}) - \frac{Vs \cdot cos(t \cdot \omega_{o}) \cdot cos(\Psi_{o})}{2} - Vs \cdot cos(t \cdot \omega_{s}) \cdot cos(\Psi_{s}) + \frac{Vs \cdot sin(t \cdot \omega_{o}) \cdot sin(\Psi_{o})}{2} - \frac{\sqrt{3} \cdot Vs \cdot cos(t \cdot \omega_{o}) \cdot sin(\Psi_{o})}{2} - \frac{\sqrt{3} \cdot Vs \cdot sin(t \cdot \omega_{o}) \cdot cos(\Psi_{o})}{2} + 4.28$$

It was therefore proved that each arm voltage has its own voltage equation and therefore they are all different, even when supposing zeros in their voltage angle sides, Ψ_s and Ψ_o . These are the angle references for the source-side group of voltages and the load-side group of voltages.

4.4. Energy Management of the MMC

The energy-control links the changes in the energy contained in the capacitors submodules with the input power. The energy is also influenced by the total output power. Therefore, when the energy is supplied to the load, the energy control will raise the input current amplitude in order to compensate the energy changes; this idea is represented in Figure 4.17. In the case when the output power is reduced, the energy control will therefore reduce the amplitude of the input current in order to maintain the energy at a constant value.



Figure 4.17. Energy changes behaviour between input and output power.

The contained energy in capacitors of a given arm was expressed in equation 3.23. It depends on the square of the aggregate capacitor voltage $(\nu_{c \sum hn}(t)^2)$. The link between the input current and the energy contained in the arm submodules was presented in equation 3.26. With this equation, it is therefore possible to calculate the power at each arm $(p_{hn}(t))$ and to express the transfer function of each submodule as presented in equation 4.30. In Laplace Transform [Schi00], it was used S (in upper caps) as the Laplace operator.

$$p_{hn}(t) = v_{hn}(t) \cdot i_{hn}(t) = \frac{d}{dt} \left[Energy_{hn}(t) \right] + 2 \frac{Energy_{hn}(t)}{C \cdot R}$$

$$4.29$$

$$H_{Energy}(S) = \frac{Energy_{hn}(S)}{p_{hn}(S)} = \frac{1}{S + \frac{2}{C \cdot R}}$$

$$4.30$$

Energy can be controlled using equation 4.30, which is linear-time invariant. The dynamics can be evaluated in open loop by applying a step response to the function, and this is represented in Figure 4.18.



Figure 4.18. Step response of the OL energy transfer function.

As it can be appreciated in the open-loop response, the settling time of the function is too long (about 23 minutes). This response risk the system instability because it cannot follow quick changes in the output power. A faster response can be obtained in different ways. First, R could be sized with a lower value (by introducing a resistor in parallel to the capacitor for instance): this resistance is in shut connection as presented in Figure 4.1, which is related to higher losses and therefore an unwanted condition. Second, if the C value becomes lower instead of R: then less energy could be stocked in the converter system; the nominal power will therefore be reduced. This condition should be avoided in order to have a more stable system.

A controller should therefore be implemented in order to improve the signal dynamics without varying the system parameters. In this case, it was chosen a PI controller because is normally suggested for many MMC applications [HIFS17, PRFK12, YLWC17]. The PI controller is presented in appendixes E.1.

The required PI must to be fast in order to efficiently respond to the load requirements and overshot must be avoided in order to protect the converter from overcurrent and overheating. For this purposes, a response time of 50 ms is quickly enough for the presented system as it represents less of one cycle of the lowest used frequency in the system. For control purposes, a recommended overshoot is of about 25 % [Ogat10] but, this value will be traduced in 25 % of overcurrent, which for high power systems is not convenient. The overshoot will therefore be limited to around 10.7

%. In consequence of this lower overshoot, a settle time risks to be longer. In this system, it was obtained a settle time of 0.35 seconds, and this value represented no problem for simulated energy exchanges, as it will be showed in section 4.6.

The closed loop (CL) of the system is stable and is much more efficient when equipped with a PI controller.



Figure 4.19. Closed-loop step response of the submodule plant with a PI controller.

This is the main energy control applied to the converter. It gives a direct relationship with the input current as it was expressed in equation 4.29. This current can now be built in the converter.

4.4.1 Input current reference

As the input current value is linked to the energy contained in capacitors, it is therefore important to define some values that allow the current calculation. First it is defined the capacitor voltages; it can be set to any value inside the defined constraint given by equation 3.86. It was selected using the most adaptable value for the system, which consider the YDY coupling of the converter, and by following the same principle applied to electrical transformers (which was explained in subsection 3.3.2), it can be calculated the phase to phase voltage applied to the ensemble of submodules per arm; this is represented in the following Figure 4.20.



Figure 4.20. Calculation of the maximum voltage applicable to each arm submodule.



Figure 4.21 represents then the applicable value at each converter stage.

Figure 4.21. Voltage distribution in the MMC AC-AC system.

In order to define the system; it is supposed that both system sides have the same voltage value, then a lower voltage value can be define for any side, while the other remains unchanged. A study was made using, as example, a range of suitable values for the output side voltages; this was explained in section 3.4.2.

Some simulation results will present how the converter can operate with the same input and output voltage, and also as a voltage transformer. Finally, the amplitude of the input current reference can be calculated and it is represented in Figure 4.22. The current angle is related to the reactive power of the system and will be explained in subsection 4.4.2



Figure 4.22. Input current reference calculation from capacitors energy.

The error is calculated with the following equation:

$$\Delta v_c * \Sigma = v_c * \Sigma - \left[\frac{Energy_{h1}(S) + Energy_{h2}(S)}{2} \right] \cdot \frac{1}{C \cdot v_c * \Sigma}$$

$$4.31$$

The PI controller in Figure 4.22 is the one defined in subsection 4.4. A PLL (see appendix D.1) is implemented too, in this case for measures of the input frequency of the system. It is used to adjust the frequency value of the input current reference. The current amplitude is calculated using the error coming from the capacitors energy.

4.4.2 Input angle reference

One-phase input current amplitude was built from the average of upper energy and lower energy sources, which values normally are close to the selected reference. Nevertheless, if the reactive power of the system is not correctly supplied, these values could diverge in opposite directions causing instability. The input angle reference is therefore referenced in this document as the reactive power control (RPC). One example of this singularity is showed in Figure 4.23, where a random system without a RPC was used as example. The showed voltage divergence can be corrected thanks to the angleshift of the input current, as it will be demonstrated next.



Figure 4.23. Divergence of the capacitor-voltages.

Charge and discharge of the submodules energy values are directly related to their voltage level, which are represented by the six arm sources (v_{hn}) from Figure 4.1. Each source energy level will depend on the two arms that share a connection point with each phase of the input source and with those of the output source. For instance, a given current shift-angle of the phase-As (input-source), will affect the charge of the arm capacitors described by v_{a1} while discharge v_{c2} . The effect can be changed when the current angle-shift value from phase as is modified; this process is repeated in order to keep a balanced system.

A definition of upper energies from all voltage sources v_{h1} and lower energies from all v_{h2} is therefore convenient in this case. A charge factor (*Chargehn*) can be used only to demonstrate the effect of the input and output side sources at each arm. This will give one index of the charging factor and its sign. It can be proven using Kirchhoff's current laws with Euler's representation.

$$I_{As}(t) + I_{Bs}(t) + I_{Cs}(t) = 0$$
 4.32

$$I_{As} \cdot \begin{bmatrix} j \cdot \left(\omega_{s} \cdot t + \gamma_{As}\right) & j \cdot \left(\omega_{s} \cdot t + \gamma_{As} - \frac{2\pi}{3}\right) \\ e & e \end{bmatrix} + e^{j \cdot \left(\omega_{s} \cdot t + \gamma_{As} + \frac{2\pi}{3}\right)} = 0$$

$$4.33$$

Introducing the phase shift represented in Figure 3.16 and generalizing to all phases using the definition of variables from Figure 3.5, and using equations 3.75 and 3.76, then the source side current (I_{Hs}) can be written as it follows:

$$\begin{split} I_{Hs}(t) &= i_{hI}(t) - i_{(h+2)2}(t) = I_{Hs} \cdot e^{j \cdot \left(\omega_{s} \cdot t + \gamma_{Hs}\right)} = \dots \\ \frac{I_{Hs}}{\sqrt{3}} \cdot \left[e^{j \cdot \left(\omega_{s} \cdot t + \gamma_{Hs} + \frac{\pi}{6}\right)} - e^{j \cdot \left(\omega_{s} \cdot t + \gamma_{Hs} + \frac{5\pi}{6}\right)} \right] - \frac{I_{Ho}}{\sqrt{3}} \cdot \left[e^{j \cdot \left(\omega_{o} \cdot t + \gamma_{Ho} + \frac{5\pi}{6}\right)} - e^{j \cdot \left(\omega_{o} \cdot t + \gamma_{Ho} + \frac{5\pi}{6}\right)} \right] = \dots \\ \frac{I_{Hs}}{\sqrt{3}} \cdot \left[e^{j \cdot \left(\omega_{s} \cdot t + \gamma_{Hs} + \frac{\pi}{6}\right)} - e^{j \cdot \left(\omega_{s} \cdot t + \gamma_{Hs} + \frac{5\pi}{6}\right)} \right] = \dots \\ 4.34 \end{split}$$

It can be seen that the term relating the output current is zero, which is normal because it is defined the source-side current only.

A power value for each arm can be calculated from equation 4.34, nevertheless this power calculation will introduce too many terms in equations that are not entirely necessary to demonstrate the flux of energies.

Some simplifications can be made at this stage:

- First, it is supposed that each three-phase voltage from the source-side and the load-side, form two voltages groups.
- Each voltages group can be referenced to an *angle side* as it was presented in subsection 4.3.4.
- By supposing now that both angle-side are zero, then the active and reactive power of the system will depend only on the angle of the current; γ_{Hs} .
- A real part of the source current is then related to the input active power, and the imaginary part will be related to the input reactive power.

As the actually power used and dissipated in any system is only the active power [GrSt94], then it is the active power that will charge and discharge the capacitors in the MMC. Finally, the input active power, as it was said before, is related directly to the real part of the input current. Therefore, the real part of equation 4.34 define the charging factor as presented next, it is introduced the phasor form of the arm current $(\overline{t_{hs}})$:

$$Charge_{h\,I} - Charge_{(h+1)2} = \Delta \ Charge_{hs} = \dots$$
$$\Re\left[\frac{I_{Hs}}{\sqrt{3}} \cdot \left(e^{j \cdot (\omega_s t)} \cdot e^{j \cdot (\gamma_{Hs} + \frac{\pi}{6})} - e^{j \cdot (\omega_s t)} \cdot e^{j \cdot (\gamma_{Hs} + 5\frac{\pi}{6})}\right)\right] = \Re\left[\frac{I_{hs}}{I_{hs}} \cdot \left(e^{j \cdot (\gamma_{Hs} + \frac{\pi}{6})} - e^{j \cdot (\gamma_{Hs} + 5\frac{\pi}{6})}\right)\right]$$
4.35

From equation 4.35, if h = 1, the term $e^{j(\gamma_{Hs} + \frac{\pi}{6})}$ is related to the arm current i_{a1} while the term $e^{j(\gamma_{Hs} + \frac{5\pi}{6})}$ is related to the current i_{c2} , both terms multiply the phasor value of the arm current and they depend only on the angle from the source current. For a given angle value, the resulting value from each term will give the charging factor for each arm, related only to the input source current. For instance, if an angle $\gamma_{Hs} = \pi/2$, then equation 4.35 will result in: $\Delta Charge_{hs} =$

$$\Re\left[\overline{\iota_{hs}}\cdot\left(e^{j\cdot\left(\frac{\pi}{2}+\frac{\pi}{6}\right)}-e^{j\cdot\left(\frac{\pi}{2}+\frac{5\pi}{6}\right)}\right)\right]=\Re\left[\overline{\iota_{hs}}\cdot\left(\left(-\frac{1}{2}+j\frac{1}{2}\sqrt{3}\right)-\left(-\frac{1}{2}-j\frac{1}{2}\sqrt{3}\right)\right)\right]=0$$

$$4.36$$

Equation 4.36 proves that an angle of $\gamma_{Hs} = \pm \pi/2$ will charge and discharge both arms using the same factor; therefore there is not any energy transfer in the system, which is expected because at those angles the resulting power would be entirely reactive.

The same procedure can now be repeated to the load side, and it is get:

$$I_{Ho}(t) = i_{hI}(t) - i_{h2}(t) =$$

$$\frac{I_{Hs}}{\sqrt{3}} \cdot \left[e^{j \cdot \left(\omega_{s} \cdot t + \gamma_{Hs} + \frac{\pi}{6} \right)} - e^{j \cdot \left(\omega_{s} \cdot t + \gamma_{Hs} + \frac{\pi}{6} \right)} \right] - \frac{I_{Ho}}{\sqrt{3}} \cdot \left[e^{j \cdot \left(\omega_{0} \cdot t + \gamma_{Ho} + \frac{5\pi}{6} \right)} - e^{j \cdot \left(\omega_{0} \cdot t + \gamma_{Ho} + \frac{\pi}{6} \right)} \right] =$$

$$(\overline{i_{ho}}) \cdot \left[-e^{j \cdot \left(\gamma_{Ho} + \frac{5\pi}{6} \right)} - e^{j \cdot \left(\gamma_{Ho} + \frac{\pi}{6} \right)} \right]$$

$$(4.37)$$

$$\begin{aligned} Charge_{h1} - Charge_{h2} &= \Delta \ Charge_{ho} = \dots \\ \Re \left[\left(\frac{j}{i_{ho}} \right) \cdot \left[-e^{j \cdot \left(\gamma_{Ho} + \frac{5\pi}{6} \right)} + e^{j \cdot \left(\gamma_{Ho} + \frac{\pi}{6} \right)} \right] \right] \end{aligned} \tag{4.38}$$

When comparing equation 4.38 with equation 4.35, it can be seen that it is obtained the same effect but, this time, the factor relating h1 is negative and the one relating h2is positive. This only means that the input side must compensate the angle changes at the load with the same negative angle (plus some reactive losses, which are automatically taken into account). As example, it is chosen an angle $\gamma_{Ho} = \pi/6$. The charging factor of the load would be $\Delta Charge_{ho} = \overline{\iota_{ho}} \cdot (1 + 0.5)$. In order to have the same charging factor at the load side, the angle must be $\gamma_{Hs} = -\pi/6$, therefore $\Delta Charge_{hs} = \overline{\iota_{hs}} \cdot (1 + 0.5)$. In consequence, the reactive power of the load is supplied.

The reactive power control will have the following expression, supposing that all submodules have the same capacitance value, the energy correction angle is given by the average of the difference:

$$\Delta Energy_{arm} = \frac{C}{2} \cdot \sum_{h} \left[\left(v_{c\Sigma_{h1}} \right)^2 - \left(v_{c\Sigma_{h2}} \right)^2 \right] \cdot \frac{1}{3}$$

$$4.39$$

The implemented control can therefore be represented as:



Figure 4.24. Input angle reference calculation from capacitors energy.

Some elements must be defined from Figure 4.24: *PFo* is the output power factor calculated using equation 4.40 below, where P_o and Q_o are the active and reactive output power respectively. Coefficients k_2 and k_1 are constant used to adapt the value unities; k_1 is used to reduce the error reference in a per unity value ($\pi/180^\circ$), while k_2 is used to adjust the sign of the output angle with -1 or +1 when the output *PF* is positive or negative (respectively). A low pass filter is implemented to the *PFo* signal because it is used directly.

In addition, a slope limiter is implemented in order avoid the error coming from highspeed variations in the energy link.

$$PF_{o} = \frac{P_{o}}{\sqrt{\left(P_{o}\right)^{2} + \left(Q_{o}\right)^{2}}}$$
4.40

The dynamic of the proposed controller is supposed to be at least three times slower than the energy controller presented in subsection 4.4. Two reasons justify this choice: on the one hand, to avoid unnecessary fast changes of the reactive power in the system and prioritize active power variations. On the other hand, the converter can handle some momentary variation in the internal energy without affecting the input or the output signals of the converter.



Figure 4.25. Closed-loop step response of the angle plant with a PI controller and comparison with the energy response.

The control of the reactive power is used to keep the upper and lower capacitance voltages at a given reference value. Nevertheless, transients in the system, as power variations, could change their original voltage level state. A new reactive power value must be set and the control will provide a new angle reference. During this process, a circulating current could be generated in the system. This current appears during transitory and remains in the system.

Minimization of the circulating current will avoid unbalances in the system and overcurrent in the converter arms.

4.5. Circulating Current Control

For MMC in AC-DC reversible topology, this current is also known as the differential current [BBEL13], which is the resulting current from delta equations introduced in chapter 3. Differential current is defined in DC case as the AC arm currents at twice the fundamental grid frequency value [Raul14]. It will be showed that this second harmonic is not present the AC-AC topology under balanced operations.

4.5.1 Circulating Current for the MMC AC-AC

The circulating current term was introduced for the MMC AC-AC one phase topology in [PRFK12]; it was defined as a current that appears naturally caused by parameter mismatch, switching transients and capacitor voltage fluctuations. It was concluded that the circulating current value needs to be minimized in order to reduce the branch losses and ensure a proper control of the capacitor voltages. In the three-phase topology [BaMe11], the circulating current was defined as the average of all arm currents. It also proposed the minimization of this value in order to reduce the energy variances and the required switching power.

In this section, a new circulating current concept and control strategy is proposed. Two objectives are related to this control; on the one hand, it has to minimize the current value in order to reduce losses in the system and on the second hand, it has to keep the balance between capacitor voltages in the system.

Even if a reactive power control is already used to control the voltage balance of arm capacitors, this control is related to the average of a three-phase system, which was assumed balanced too. A balance based on circulating current control offers more flexibility because it can individually balance voltages with a self-produced current of the converter.

4.5.2 Circulating current calculations

When looking to equations 3.34, 3.37, 3.51 and 3.52, there were some defined terms that were assumed zero for modelling purposes. These terms are the delta currents ($i_{\Delta hs}$ and $i_{\Delta ho}$) and the x voltages (V_{xhs} and $v_{x(h+1)o}$). Nevertheless, for multiple reasons related to operations of the converter, these terms are not zero and therefore theirs values must be minimized. Some of these reasons are; the *natural unbalance* in arm voltages as it was presented subsection 4.3.2; also because normal transients in the system. Delta current values are represented in the following figure.



Figure 4.26. Representation of delta current values in the converter.

Normal operations that might induce any circulating current value should be allowed for the converter, and therefore it must be implemented a control that minimizes the values of delta variables presented in Figure 4.26. However, there is still a limitation in the control implementation that should be considered: as presented in subsection 3.3.2, there are only four equations that can be used to control the converter; these are delta and sigma equations from the source frequency and; delta and sigma equations from the output frequency.

Sigma equations have already been used to control the energy exchanges in the converter (see equations 3.53 and 3.54), this leaves only two controllable equations and four variables each phase: two delta currents and two x voltages. It will be assumed that the x voltages are *already been controlled*; this assumption can be made because the delta current control searches precisely to minimize those x voltages. In order to calculate the circulating current value for this converter, the following definition will be studied:

It is supposed that all sources from the source-side and the loadside of the system contribute to the creation of a current value called the circulating current. Moreover, this current exists only inside the converter.

When looking to Figure 4.26 and introducing currents from the source and the load as they were defined for this system, it is obtained:

$$i_{a1} + i_{b1} + i_{c1} = I_{As} + i_{c2} + I_{Bs} + i_{a2} + I_{Cs} + i_{b2} = i_{a2} + i_{b2} + i_{c2}$$

$$4.41$$

$$i_{a2} + i_{b2} + i_{c2} = -I_{Ao} + i_{a1} - I_{Bo} + i_{b1} - I_{Co} + i_{c1} = i_{a1} + i_{b1} + i_{c1}$$

$$4.42$$

From equation 4.41 and 4.42 it can be seen that the sum of current is not zero as in a three-phase system. Therefore, in order to introduce the circulating current defined

before, it is convenient to represent a new equivalent model of the converter; variables are the same as it is presented in Table 4.3 below, but it is only evaluated the contribution to the circulating current value:



Figure 4.27. Schematics for circulating current control of the MMC in AC-AC topology.

In order to calculate the required circulating current value, in Figure 4.27 it has being added a *circ* upper index, which represents the link with the circulating current. Moreover, a new representation becomes convenient for representing the control. This modifies only the sign of certain variables as presented next. The idea is to simplify the resulting equations.

	•	
	Circulating Current	Equivalent from original
	Variables	variables
e c	$I_{As}{}^{circ}$	I_{As}
ouro Side	$I_{Bs}{}^{circ}$	I_{Bs}
N N	$I_{{\mathcal C} s}{}^{circ}$	I_{Cs}
	I_{Ao}^{circ}	-IAo
oad ide	$I_{Bo}circ$	$-I_{Bo}$
L_{c}	$I_{Co}{}^{circ}$	-ICo
S	i_{a1} circ	i_{a1}
rent	i_{a2} circ	$-i_{a2}$
. cur	i_{b1} circ	i_{b1}
ertei	$ib2^{circ}$	-ib2
on ve	i_{c1} circ	i_{c1}
Ŭ	i_{c2} circ	$-i_{c2}$

Table 4.3. Variables equivalent for circulating current calculations.

The circulating current value for each phase can be defined as:

$$\dot{i}_{h1}^{circ} - \dot{i}_{(h+2)2}^{circ} = \dot{i}_{circulating}^{ho}$$

$$4.43$$

$$i_{h2} circ - i_{h1} circ = i_{circulating} hs$$

$$4.44$$

Finally, the general form of the current values that needs to be minimized are presented in the following equations:

$$I_{Hs} \, circ - (i_{h1} \, circ - i_{(h+2)2} \, circ) = i_{\Delta ho} \tag{4.45}$$

$$I_{Ho}^{circ} - (i_{h2}^{circ} - i_{h1}^{circ}) = i_{\Delta hs}$$

$$4.46$$

Consequently, equations 4.45 and 4.46 correspond to the delta current values introduced into equations 3.34 and 3.38. The total circulating current value can be obtained with the application of the superposition theorem in Figure 4.27 as it follows. From the input-side frequency, all arm currents from each arm should have the same amplitude and frequency value in order to avoid this frequency at the output-side of the converter. Therefore,

$$i_{a_{1s}} circ + i_{a_{2s}} circ = i_{circulating} a_{s}$$

$$i_{b_{1s}} circ + i_{b_{2s}} circ = i_{circulating} b_{s}$$

$$i_{c_{1s}} circ + i_{c_{2s}} circ = i_{circulating} c_{s}$$

$$4.47$$

Using the same principle for the load variables,

$$i_{a2o} \operatorname{circ} + i_{b1o} \operatorname{circ} = i_{circulating} \operatorname{ao}$$

$$i_{b2o} \operatorname{circ} + i_{c1o} \operatorname{circ} = i_{circulating} \operatorname{bo}$$

$$i_{c2o} \operatorname{circ} + i_{a1o} \operatorname{circ} = i_{circulating} \operatorname{co}$$

$$4.48$$

And, as presented in equations 3.65 to 3.70, the total circulating current value can be defined as:

$$(i_{a1}^{circ} + i_{b1}^{circ} + i_{c1}^{circ}) + (i_{a2}^{circ} + i_{b2}^{circ} + i_{c2}^{circ}) = i_{circulating}$$

$$4.49$$

Introducing equation 4.49 into equations 4.41 and 4.42, it can be defined the circulating current term as it goes:

$$i_{circulating} = \left(I_{As} + i_{c2} + I_{Bs} + i_{a2} + I_{Cs} + i_{b2}\right) - \left(-I_{Ao} + i_{a1} - I_{Bo} + i_{b1} - I_{Co} + i_{c1}\right)$$

$$4.50$$

The sum of currents in a balanced three-phase system is zero. This concept is applied to the sum of the source-side and the load-side currents from equation 4.50. Therefore, the difference of the arm currents from the same equation should be zero too; this defines the circulating current value of the converter.

Because of the use of resonant correctors, the control of the circulating current does not requires Park's transformation. Therefore, references are introduced directly in AC creating a more efficient process. The general circulating current control diagram is represented in the following figure.



Figure 4.28. Schematics for delta current control.

Control strategies presented at this point where developed with the average model of the MMC when working in direct AC-AC topology. The presented energy control from section 4.4 and the effects of the circulating current control from section 4.5 will be tested in simulations in the following section.

4.6. Simulation results II: evaluation using the average model

The new simulated system ads the equivalent submodule capacitors and the new control schemes that produce the input current reference and the delta current errors related to the circulating current.



Figure 4.29. Average model schematics of the MMC AC-AC.

The implementation of an energy management control is essential to operate the energy transfer of the converter. This control will be tested in this section using an average model, where all submodules per arm are condensed in one equivalent as it was explained in chapter 3. This equivalent capacitor per arm, which is still accurate, allows a much faster simulation when compared to the switching model, where their only difference should be the lack of high frequency harmonics product of the switching devices that in all cases must be neglected [BaMB14]. The accuracy of the average model gets lower when faults are studied in the system [ZBBF16], and it is useless when individual submodules must be studied.

Circulating current control is also required to increase the balance between arm capacitors, as it will be exposed next. Two model using two different voltages levels will be studied, the first one using the model presented in Table 4.2, which is related to an application in offshore. The second one will use a benchmark parameters in a scaled model (see Table 4.5); this model is proposed in order to be set the bases for a future Real-Time simulation, accessible in the laboratory where this thesis was prepared. Tests of the circulating current control and with a bigger arm inductance are presented in appendix F.1.

4.6.1 Average model: high voltage - high power simulation results

The used simulation data are the following ones:

Table 4.4. MMC AC-AC averaged parameters for offshore application.					
Nominal AC-AC Voltage	250 kV peak per phase				
Output Power	750 MW				
Arm Inductance value	5 mH				
Equivalent Capacitance Value	$137 \ \mu F$				

A steady state simulation of the input and output voltages and currents waveforms gives:



Figure 4.30. Voltage and currents waveforms in a high power system converter.

It can be appreciated the used of the both system frequencies and a peak of voltages and currents of 250 kV and 2 kA respectively, which are almost in phase. A slight reactive power value was found and it is related to the inductive losses of the system, which are also required to the operation of the converter. At the represented period in Figure 4.30, input apparent power was of $S_{in} = 250.3 MW - 10.3 MVA$, which gives a $PF_{in} = 0.9992$. And an output power of $S_{out} = 249.9 MW - 46.1 kVA$, with a $PF_{out} =$ 1.0000. Power losses (ΔP) and the efficiency (η) of the presented system are calculated as following:

$$\Delta P = 1 - \frac{P_{out}}{P_{in}} \tag{4.51}$$

$$\eta = 100\% - \Delta P \tag{4.52}$$

In this case $\Delta P = 0.160$ % and therefore $\eta = 99.840$ %. A Fourier analysis of the presented current waveforms results in:



Figure 4.31. Harmonics of the current

A small presence of the lowest frequency component is observed at the input current side from Figure 4.31. This is a problem related to the resonant controller adjustment; it is presented as example to show the unwanted harmonic at the input current side. The output-side current showed that unwanted harmonic could be complete eliminated when the controller is correctly adjusted.

The Total-Harmonic-Distortion (THD) was calculated using equation 4.53 [Knes99]

$$THD = \frac{\sqrt{\sum_{harmonic} RMS(harmonic)^{2}}}{RMS(1)}$$

$$4.53$$

For this simulation, the circulating current control was also tested. In the following figure, the behaviour of this current during the entire simulation can be appreciated. During the first 3 seconds, the current control was inactive; it can be seen how the circulating current raises its value until the circulating current control is manually activated. It can be appreciated how the minimal obtained circulating current value is almost four times lower than the maximum one at 2.5 seconds. Disturbance in the circulating current value are also presents at the same moment in the capacitor voltages, which is expected because this current helps to keep balanced the capacitor voltages.



Figure 4.32. Circulating current and capacitor voltages waveform behaviour.

At 10 seconds, the circulating current control was manually deactivated and it can be appreciated the distortion in the capacitor voltages from Figure 4.32, which is normal because it can be considered as an internal fault of the converter. The voltages remain in oscillation state until 11.5 seconds, and then at this moment the circulating current control was reactivated. From that moment, the oscillations ended and the converter starts its balancing process with the aid of the circulating current control.

These results show not only the converter operations and its capacity to be used in a high voltage and high power link, but also the robustness of the proposed circulating current control. This control reduces the circulating current value, keeps balance of the capacitor link and helps to prevent instabilities in the converter. It is proposed to continue the following simulations using a lower scale of voltage. This will offer results that could be compared with a laboratory benchmark.

4.6.2 Average model: scaled low voltage model

Until this date, no MMC in direct AC-AC configuration has been commercially built; moreover, the dimensioning of components is out of the sight of this work. It is therefore supposed that components are the same from the MMC AC-DC in reversible configurations and the MMC parameters are given for a reduced model:

Number of Levels	20
Cell Voltage / All submodule voltages each arm Output power Current (AC RMS) AC voltage per phase 1 / 3-phases AC RMS Cell Capacitor Single arm inductor	40V / 800 V 10.5 kW (or 6 kW) 16.7 A 208 V / 360 V 20 mF (or 12 mF) 2.5 mH (or more in series) < 50 kHz
Switching nequency	_ 00 MIL

Γable 4.5. MMC OP1200 parameters from	OPAL-RT [Opal17b].
---------------------------------------	--------------------

The model is the one of 10 kW presented in Table 4.5. In simulations, results presented in Figure 4.33 it is showed the converter operation face to power references changes and even changes of the power direction. It can be appreciated that power reaches 3.5 kW per phase (between 14 and 15 seconds), which is the maximal tolerated

power of the converter. A more dynamic behaviour can be observed in the reactive power, which is normal as the control is trying to keep all equivalent capacitor voltages to around the same vale.



Figure 4.33. Power when using 2.5 mH in arm inductances.

It can also be appreciated the behaviour in the capacitor voltages and the circulating current waveforms in Figure 4.34. All equivalent capacitor voltages are kept together thanks to the reactive power and circulating current control of the system; even in difficult situations as changing the power direction, where the capacitor voltages seems to diverge at 18 seconds, but they get balanced even when there is no input nor output power in the system (between 18 and 22 seconds).



Figure 4.34. Capacitor voltages and circulating current when using 2.5 mH in arm inductances.

It is proved that the control can be adapted to a lower voltage system with a different inductance value and other power parameters. The same converter working with another inductance value of 10 mH is presented in appendix F.1, where only the required input reactive power is increased of four times as the inductance value, and the converter behaviour faced to power changes is the same.

These results were presented with an average model that uses an equivalent submodule model; a switching model that integrate all submodules will be presented next. For this task, an internal capacitor balancing is required. More robust results will therefore be presented in section 4.8 integrating all submodules capacitors.

4.7. M-Capacitors Balancing

In the average model, results shown that all arm equivalent capacitors voltages were kept together after each power variation made to the system. For the switching model used in this section, the same results are expected even that each arm is composed now of M submodules per arm. This is because all arm capacitors must be kept at the same value in order to guarantee the good operation of the converter.

In order to reach this task, it is required a new low-level control scheme, called the *Balancing Control Algorithm* (BCA). It will follow all voltages from capacitors and compensates those who get far from a given reference, which was fixed to one voltage value in these simulations.

Generated pulses from the BCA are then introduced in a phase shifted carrier pulse width modulator (PSC-PWM) in order to generate, at a maximum of 50 kHz of frequency (see Table 4.5), the corresponding pulses to control the converter switches.

A sorting algorithm is used to analyse the capacitor voltages and select them according to their voltage level. As it was presented in the full bridge configurations from Figure 3.8, selecting a submodule state (out of 6) will depend on the arm current sign and the modulation index sign. The objective of this algorithm is to insert submodules at a precise moment to charge or discharge them and therefore to keep all of capacitor voltages at almost the same level.

The selected BCA (presented in Table 4.6) guarantees a voltage balancing under all MMC normal operating conditions [DQBS14], but it produces unnecessary switching transitions among the submodules; improved BCA methods could reduce the switching losses, but they offer the same results from a general point of view, which is accepted for this work. It is therefore not proposed a new sorting algorithm, and the used one is simple to apply and offers good results [BBEL13, Berg15, BPBA15]. The voltage dynamics of individual capacitors are analysed using equation 3.22, which is recalled in the following Table 4.6:

Table 4.6. Implemented BCA from current and modulation index signs.

	Recalled equation:	$\frac{C_{SM}}{M} \cdot \frac{d}{dt} \left[v_{c\Sigma_{hn}}(t) \right] = m_{hn}$	$\cdot i_{hn}(t)$
mhi	n < 0	mhn	> 0
i_{hn} < 0	$i_{hn} > 0$	$i_{hn} < 0$	$i_{hn} > 0$
Charge	Discharge	Discharge	Charge

In order to sort the capacitor depending on their voltage levels, the function *sortrows* from [Matl15] was used. Their level will depend on the chosen state from Table 4.6 as it is explained next:

Four states could be used to carry out the capacitor voltage balancing task. The other two states will keep the capacitor voltage without variations. The BCA requires the measure of each voltage, which is measured and sorted.

If the arm current and the modulation index have the same sign, out of M submodules in the corresponding arm, those with the lowest voltages are identified, inserted and therefore charged. Voltage value will increase in consequence.

Supposing that arm current and modulation index have now different signs, out of M submodules in the corresponding arm, those with the highest voltages are identified, inserted and therefore discharged.

Regardless of the direction of the arm current, when the modulation index is zero, the corresponding capacitor voltage remains unchanged.

In conjunction with the BCA that previously selected the required switching function for each submodule, a PSC-PWM is then used to control the high frequency component of the MMC arm currents. This particular PWM generates one carrier signal for every power module, and carriers are phase-shifted by an angle that is obtained by dividing II by the number of submodules per arm. This is useful to achieve better performance results in waveforms, this technique is presented in [QiZL11].

As example, in the following Figure 4.35, it is presented a period from one arm capacitor voltages in which all submodules voltages can be appreciated.



Figure 4.35. Sorting algorithm applied to the switching model of the MMC (switching frequency of 50 kHz).

A very slight difference between all voltages can be appreciated, which confirms the good performance of the selected BCA.

4.8. Simulation results III: evaluation using the switching model

The switching model is considered as a high-fidelity model, which requires an enormous amount of computational time [SDMD14]. This is because the program creates an admittance matrix, which has a size equal to the total number of nodes in the network subsystem. This nodal matrix must be inverted to calculate the new voltages (as in the case of the nodal admittance matrix in power systems [GrSt94]), and

this might occur several hundred times in a cycle. The presented converter in Table 4.5 has a huge number of submodules; simulation time is therefore expected to be very large and dynamics tests would be difficult to execute in this converter using the exact switching model. A change in the number of submodules can reduce the required simulation time.

4.8.1 Proposed parameters

In order to keep the same switching model, it is proposed to reduce the simulation time of results by supposing only 10 submodules per arm instead of 20 submodules. This will not affect the implemented control. The only effect is that it doubles the required voltage per submodule. Therefore, in order to keep the same power of the converter, it is supposed that each capacitor can manage 80 V instead 40 V.

All tests were performed using the following data, with modified number of levels from the fabricant parameters presented in Table 4.5.

_	
Number of Levels	10
Cell Voltage	80V
Output power (3 phases / 1 phase)	10.5 kW / 3.5 kW
AC RMS Current / peak value	$16.7{ m A}$ / $23.6{ m A}$
AC voltage per phase 1 / 3-phases AC RMS	208 V / 360 V
Cell Capacitor	12 mF
Single arm inductor	2.5 mH
Single arm resistor	1 mΩ
Switching frequency	$50 \mathrm{~kHz}$

Table 4.7. MMC AC-AC parameters for simulations using the switching model.

The first part of results will present some power variations and frequency analysis. Frequency conversion between 60 Hz and 50/3 Hz will be also presented and compared with the studied one of 50 Hz to 50/3 Hz. The idea is to observe the impact kon the capacitor voltages, which are related to equations showed in subsection 4.3.4.

The second part of simulations present a robustness analysis. Results will present the converter behaviour faced to one of the perturbations listed below:

- First, some internal perturbations related to changes in the system parameters as arm inductances is evaluated;
- Second, some external perturbations as frequencies changes and voltage drops in the generation system are tested;
- A final part of results will present the MMC behaviour when it has to face faults in the generation system.

Loss of submodules and variations of the inductance values of the converter are not tested. This is because the used BCA is not programmed to identify a faulty submodule and therefore, an improved and more adapted BCA for this MMC should be developed.

4.8.2 Power variations in the MMC AC-AC switching model

Active power was variated during the first 6 seconds of simulations as it can be observed in Figure 4.36. From 6 s the system remained at constant power. The

simulation was stopped at 10 seconds, considered in this case as steady state. The same steps of power were used for both studied cases of the frequency converter: 60 Hz to 50/3 Hz and 50 Hz to 50/3 Hz. In both cases, the converter was capable to follow any power reference, and changes of the power direction. Reactive power, as it was not demanded by the output system, is low when compared to the active power.



Figure 4.36. Power variations in the switching model of the MMC AC-AC

Between 2.5 and 3 seconds, it has been reached a power of 3.5 kW at the output side. At 2.8 seconds, for both cases input and output power values are:

- For the 50 Hz case, $Sin = 3.569 \ kW 68.3 \ VA$ and $Sout = 3.483 \ kW 11.9 \ VA$. Power losses are therefore, $\Delta P = 2.421 \ \%$ and $\eta = 97.58 \ \%$.
- And the 60 Hz case, $Sin = 3.570 \ kW 119.5 \ VA$ and $Sout = 3.483 \ kW 11.5 \ VA$. Power losses are, $\Delta P = 2.437 \ \%$ and $\eta = 97.563 \ \%$.

Supposing for instance that is used an unwanted bigger arm resistor, it should result in bigger active power losses. To verify this, the simulation was repeated for a resistor value of 0.1 Ω and, power losses increase to 4.130 %. Therefore, the loss value will always depend on the quality of the inductance component used for the MMC, and this value might change from one application to another.

For both systems, it was obtained the same active power losses, which is normal because the system parameters were not modified. Nevertheless, it can be appreciated that the input reactive power increased from 68.3 VA at 50 Hz to 119.5 VA at 60 Hz. This is related to the raise of frequency and the reactive impedance from arm inductances.

The capacitor voltages using both structures for frequency conversion are presented in Figure 4.37. It can be seen that the entire simulation voltages remained below 80 V, which is the highest allowed capacitor voltage for the used converter.



Figure 4.37. Capacitor voltages when using an exact model MMC in AC-AC

One of the most relevant results from Figure 4.37 is the difference between the capacitor voltages of both examples. First, they can be compared in steady state: at almost 10 seconds of simulation where only 1 kW of power is going into the source from the load. The difference between them is related to the distribution of voltages in all six arms of the converter, as it was explained in subsection 4.3.4.

- When the conversion is made from the 50 Hz to 50/3 Hz, arm voltages present the same distribution of voltage every 60 ms. This is because there is a conversion factor of 3 (a natural number) between frequencies. This voltage distribution was presented in Figure 4.9, and it might be undesirable because the voltage is not equally distributed in all arms permanently. This increases power losses and heating in arms carrying the highest voltages. This could be interpreted as an unbalanced situation; supposing the presented converter as a six phases system (six arms), a normal operation of the converter implies that 2/6 of phases will permanently have a lower voltage than the other 4/6 phases (see Figure 4.9). Nevertheless, the control was developed in order to work under this situation and to balance the system.
- This effect is not the same when the conversion is made from 60 Hz to 50/3 Hz; the frequency factor is now of 3.6. In this case, it is required at least 5 cycles of the low frequency value in order to repeat the cycle (3.6 times 5 equal 18, which again results in a natural number). This means that, during 5 cycles of the lowest frequency, one arm voltage of the converter will result in a different amplitude every cycle of the highest frequency. Doing this, for every arm of the converter that are all phase shifted, then the resulting arm voltage will have a better distribution when taking into account all arm voltages. An example is presented in Figure 4.38 for the 60 Hz case.

This is the same reason why the presented capacitor amplitudes are different at each frequency case. When comparing both cases at the maximum power of the system, at around 2.3 seconds, there is a 32.66 V in the 50 Hz case and 8 V in the 60 Hz case, which represent better performances in dynamic situations.

Every arm voltage distribution was developed with equations presented in section 4.3.4, and an example for the 60 Hz case, of one arm capacitor voltage during five cycles of the lowest frequency is presented in the following Figure 4.38. It can be appreciated that the waveform at 9.2228 seconds is repeated again at 9.5228 seconds, which correspond to 18 cycles of the high frequency value into five cycles of the low frequency value; it can be appreciated that the maximum voltage value change every 20 ms, and this is the same behaviour for all other five arms.



Figure 4.38. Capacitor voltage during five cycles when converting 60 Hz to 50/3 Hz.

Arm currents instead do not have a continuous component and, besides the normal presence of both system frequencies, it will be showed that they present the highest THD when compared to other currents of the converter, as the control was designed to keep all harmonics inside the converter. Results are presented for one cycle of the low frequency and using only upper phases $(i_{a1}, i_{b1} \text{ and } i_{c1})$. Lower phases $(i_{a2}, i_{b2} \text{ and } i_{c2})$ present the same waveforms but phase shifted.



Figure 4.39. Arm currents and harmonic content.

In Figure 4.39 waveforms present the two main frequencies used in the system (which values are 100 %). This is normal as it was exposed in section 4.5. The input and output currents of the converter present much less harmonic content as observed next.



Figure 4.40. Input and output current waveforms and harmonic content in steady state.

In Figure 4.40 are presented the input and output current of the converter at both frequencies. The harmonic effects of the switching devices and noises in waveforms can be appreciated in the resulting frequency spectrum. In this spectrum, high frequency harmonics are lower than 0.25%, producing a THD lower than 0.4%, which is an excellent result compared to the IEEE standard [Ieee14], where it is recommended a THD lower than 4% at this current level. It can also be appreciated that the load and the source frequency has been almost entirely suppressed from source and the load side respectively. The obtained circulating current value is presented in appendix F.5.

It was presented that the converter is able to operate normally under steps of power references of the system. Variations of the power factor are also possible as presented next.

4.8.3 Power Factor changes

The modelled system can handle changes in the output power factor as it has been tested in this section. When a new output current angle reference is given, the output current change immediately and the input current follows it with a time lag has it has been defined in section 4.4.2.

As it can be appreciated in Figure 4.41, the converter was able to follow the imposed angle references and when sharp variations were made (from -30° to $+30^{\circ}$ for instance), a described divergence effect can be appreciated in the capacitor voltages. However, this divergence is transitory only, and small angle variations do not impose any kind of problem to the converter. The same example also probes that variations from capacitive to inductive angle can be made in the converter and this effect is observed on the power responses.



Figure 4.41. Power factor changes and influence on capacitor voltages, active and reactive powers.

These power responses shown the variations between the active and reactive power at each variation of the angle, the apparent power normally remains at 1 in p.u. When angle variations stop, the capacitor voltages return to their equilibrium point as it can be appreciated after 15 seconds of simulation.

Activity under normal operations has been tested in this section. The response is evaluated when is faced to fault conditions in the following subsections.

4.9. Simulation results IV: assessment under disturbance conditions

This section presents the behaviour of the converter when it has to face some faulty situations. All tests are performed at a fixed output power of 1.5 kW for each phase, which corresponds to the half of the nominal power of the converter. A balanced steady-state simulation is therefore made to compare with the disturbance cases. These results are showed in Figure 4.42.

It can be noted that the steady state of reactive power and capacitor voltages is reached at 4 seconds. At this point, some values can be used as comparison:

- Active and reactive power are at the constant values presented on the figure.
- Peak value of the arm current are around 10 A.
- Maximum amplitude between all capacitors and of each capacitor are presented. The upper (76.83 V) and lower amplitudes (72.73 V) are 2.33% far from the reference value.



Figure 4.42. MMC at 1.5 kW, steady state and balanced conditions.

Figure F.10 in appendix presents the obtained circulating current in steady state; it was analysed using Fourier to obtain the frequency response. The amplitude of this current was around 0.3 A, and it could be compared to the other presented cases in the same appendix.

Fault simulations are divided into internal faults; where the parameter mismatches are evaluated: external voltages drop and over voltages, external frequency variations and input source-side faults.

4.9.1 Parameter mismatches

Internal perturbations are related to some parameter mismatch at the modelling stage, or to variation of the internal parameters of the converter as the inductances or capacitances [Hern11, HRSD10]. In some cases, the value of the components of the converter might not be precisely the same and, to evaluate the response of the proposed

control faced to these situations. Up to 10% of difference with the original inductance values where tested as it is presented in Table 4.8.

Arm impedances modifi	cation
Changed Variable	New value
L_{a1}	95%
L_{a2}	100%
L_{b1}	98%
L_{b2}	110%
L_{c1}	90%
Lc_2	100%

Table 4.8. Parameters mismatch as internal variations.

In the following Figure 4.43, it is observed a normal operation of the converter when the system parameters are different from the used to design the control.



Figure 4.43. Results with parameters mismatch.

When compared to Figure 4.42, active power, reactive power and arm currents did not present noticeable variations. The capacitor voltages instead presented some variations; the total amplitude raised of 2 V and each individual arm amplitude raised about 0.1 V. The highest and lowest amplitude values are at +4.3 % and -3.9 % from the reference. There is therefore a new point of operation in the converter, which is still supplying the required power without any disturbance to the source or to the load.

Small mismatches are tolerated as presented before; and some more important variations are analysed next.

4.9.2 Voltage Variations

Disturbances as load steps and input voltage variations are quite common in any system [HRSD10]. In the particular case of OPS, frequency changes from the source could occur because of instantly and unpredictable variations of wind. This is why some frequency drops are tested at the input side of the converter. In addition, some voltage

variations can also occur because of faults in the system, or because of sharp changes of power.

The impact of voltage variations (VV) in all three-phases coming from the generation system is analysed on the converter. Those VV affect directly the arm voltages as it was presented in section 4.3.4. Arm voltages are liked to capacitor voltage, which are the most affected part of the converter.

In Figure 4.44, it can be observed the Voltage Variations (VV) used as a reference to the input source-side of the converter. The study was performed with over voltages of +10 %, +20 %, +30 %, +40 % and +50 %, where each one of the variation was applied during 200 ms (arbitrary chosen when compared to the selectivity time for relays in [JeSy13]). And with voltages drop of -10 %, -20 %, -30 %, and -50 %, where the last two variations remained for more time.



Figure 4.44. Overvoltage and voltages drop references.

Results from overvoltage variations are presented first. 50 % of variation is applied to the waveforms from Figure 4.45. On the one hand, a sudden increase of the input voltage can be observed (which correspond to the made test), while the current decreases progressively, it required 40 ms to decrease the current amplitude and 110 ms to stabilize it at its new value. On the other hand, output voltage and currents seem to be unchanged, this can be observed when comparing input and output power measurements.



Figure 4.45. Voltage and current waveforms when 50 % of overvoltage is applied.

Some power peaks related to these voltage variations are observed in Figure 4.46. At the instant where an over voltage occurs, the system will suddenly see an increase of

power, which is absorbed by the capacitors. The energy control will therefore react in order to reduce this exceeded energy.



Figure 4.46. Active and reactive power, capacitor voltages and arm currents faced to overvoltages.

From Figure 4.46, it is important to remark that the output power was never affected by voltage variations that occurred at the input side, this effect can also be observed in back-to-back converters (presented in subsection 2.3.3). Therefore, the MMC in YDY configuration is capable of decoupling the input from the output side. Peak of voltages for each test are 7.1 %, 13.0 %, 18.2 %, 22.6 % and 26.4 % bigger when compared to the steady-state response presented in Figure 4.42, it can be seen that the impact becomes bigger when the input disturbance does it too. Every peak lasted no more than 60 ms, which correspond to one cycle of the lowest frequency of the system and, it is related to the duration of the stabilization of current presented before.

Capacitor voltages instead changed suddenly their value, it can be seen that their maximum amplitude was reduced of 2.5 times when compared to steady state, and therefore there are not risk of destabilization. The only risk related to over voltages would be to overpass the voltage or current limits of internal components; nevertheless, in this case the arm current amplitude did not passed 12.5 A of amplitude, only 10% more when compared to steady-state, which is normal because no overcurrent was in the converter.

Voltages drop will offer the same responses when compared to overvoltages. When comparing the maximum applied variation, the converter was able to work with -50 % of voltage drop, but it risked to destabilize the system as it approaches to the lower limit of the converter (discussed in subsection 3.4.2). Results are presented next:



Figure 4.47. . Capacitor voltages, active and reactive power and arm currents faced to voltages drop.

Negatives peaks of power are related to voltage drops, while positive peaks represent the restore of the voltage value and it is seen as an overvoltage. The amplitude of the peak will depend on the step intensity used for variations.

At 3.1 seconds, the maximum amplitude between capacitor was 1.3 more when compared to steady state and the bigger voltage limit was not overpassed. The lower limit was also reduced as the input voltage of the system was 50 % lower, and therefore no problems were in the capacitor voltages, which stabilized after 3 seconds during the disturbance. Nevertheless, at this lower voltage point, current increased in order to keep the same power (as in an electrical transformer). It can be seen than the arm currents almost reached its limits. By consequence, this is not so attractive because at the presented operation point, the converter is working at maximum of its current capacity and only a half of the designed power.

4.9.3 Frequency Variations

Changes of the nominal frequency are not allowed in a Conventional Power System (CPS) working at 50 or 60 Hz. However, frequency can variate as a consequence of faults in the system [Ewar78], generating what is called violent frequency swings. Those are the effects of spontaneous power outages and governors overreactions in the

systems. In OPS this frequency effects are expected more often than in a CPS because of the low inertia of the offshore wind generators [JuPG15].

In order to test the robustness of the converter, sharp variations of frequency (VF) at the input side of the system were made. Variations of +1, -1, and -10 Hz were tested. Additionally, the frequency was decreased to 30 Hz, this may not be real in a power system, but it proves the response of the converter faced to lower frequencies. Variations are presented in Figure 4.48. The system continued its operations at each tested frequency. Changes in the system where monitored by a Phase-Lock-Loops (PLL) [Best07], presented in appendix D.1, it is used to update the frequency value of the system in order to build the input currents references.



Figure 4.48. Variations of frequency and response of the PLL.

As it can be appreciated, the PLL was able to follow the applied references and its dynamic was enough for the presented simulation as the converter was able to follow every change without destabilizing the system.

As in the case of voltage variations, the output power was not affected by variations of frequency made at the input side. Moreover, peaks are also observed in the input power at the same moment that the frequency changed as the input side is being influenced. For changes of 1 Hz, these peaks are maximum of 0.5% when compared to the steady state from Figure 4.42. For changes of 10 Hz, peaks raised to 3 % the input power. These variations are therefore accepted. The results in current waveforms, capacitor voltages and powers are presented next.



Figure 4.49.Active and reactive power and capacitor voltages faced to frequency variations.

Changes in the frequency of the system are observed at the capacitor voltages from Figure 4.49. Some variations are observed after the frequency changed, but it did not destabilized the system. This variations are normal as those values depend on arm voltage waveforms and therefore, on the source voltages. An individual increase of voltage amplitude can be observed for the lowest tested frequency passed from 1.1 V in steady state to 1.86 V, this was still accepted for the converter.

After each variation, the control was able to restore the normal operation of the converter and to keep working at the new frequency, proving its robustness faced to variations of frequency. Figure F.11 in appendix shows the obtained circulating current related to parameters mismatches, no particular variations are observed when compared to the steady state response.

Figure 4.50 shows an example of the good quality in current waveforms when changing big values of frequencies in two cases: when the frequency was converter from 49 Hz to 40 Hz and from 40 Hz to 30 Hz.



Figure 4.50. Input current waveforms faced to big frequency variations.

The converter can also work by stepping-up the frequency value; transforming 50 Hz to 60 Hz, results are presented in appendix F.3. Some fault tests will probe that the system input and the output system are decoupled and faults from one side are not propagated to the other.

4.9.4 Response of the MMC AC-AC under voltage dips at the source side

External disturbances presented before are related to balanced faults. In this part, one-phase and two-phase unbalances are tested. The applied disturbances at the inputside source are listed in Table 4.9. Faults to the output side were also applied, and results are presented in appendix F.4.

				1				
Time [s]	4	5	6	8	10	12	14	
V _{As} factor	0.9	0.8	0.9	0.8	0.5	0.5	0	
V_{Bs} factor	-	-	0.9	0.8	-	0.5	-	
V_{Cs} factor	-	-	-	-	-	-	-	

Table / C) Faults	to ind	leuhivi	nhasee
1 abre 4.8). rauns	10 110	inviuuai	phases

Each fault was applied during 100 ms, input current and voltages during all 16 seconds of simulations are presented next.



Figure 4.51. One-phase and two-phase faults in the input voltage of the converter.

In Figure 4.51, currents returned to its steady state value in less than 500 ms after each fault and the converter was able to continue its regular operations as expected. Even if asymmetrical faults were applied, currents presented a symmetrical behaviour for all kind of faults. This is because the energy control builds a perfect three-phase current as a reference, and the resonant corrector is capable to follow this reference even in fault condition. Measured active and reactive power, capacitor voltages and arm currents are presented next.



Figure 4.52. Active and reactive power, capacitor voltages and arm currents faced to faults at the input-side of the converter.

The output power was unchanged for each one of the applied faults as it can be appreciated in Figure 4.52. Compared to steady state, input power raised in about 14 % for the first three applied faults; in 18% for faults applied at 8 and 10 seconds and in 26% for the last two faults. The converter was able to continue its operations for all applied faults. Nevertheless, faults applied at 12 and 14 seconds cannot stand 100 ms in the converter because the capacitor voltages over-passed their maximum limit. For these two cases, arm currents almost doubled their initial amplitudes, which proves the high intensity resulting from these faults that normally are tripped in a much shorter time. Figure F.11 in appendix shows the obtained circulating current related to parameters mismatches, amplitude of this current raises at each case; the maximum obtained amplitude was 0.6 A which is twice the value in steady state presented in appendix F.5, this current returned to its normal value after the fault was cleared.

A three-phase fault response is presented next. It was applied during 50 ms, and the converter continues its normal operation after the fault as presented in Figure 4.53. Even if an over-current is in the system, the loss of voltage at the input side discharges the capacitors in a symmetrical way, and therefore a risk of overvoltage of capacitors is inexistent when a three-phase fault begins. The real risk resides when the voltage is restored because the current increased while trying to supply the same power: it was obtained an overcurrent in the system that should be limited. In the simulated case, the current reached 30 A peak at the input and, of about 20 A peak inside the converter arms. In this case, the current in the arm almost reached the nominal value defined in Table 4.5.
In case of maximum power, a higher current level would have been obtained, and then a protection algorithm should have triggered the relay in order to protect the converter. An internal protection algorithm could also be useful to limit the output power if the application of the converter allows it. Finally, blocking states of submodules can also be used to limit the fault current, yet, this is out of the sight of this work.



Figure 4.53. Response faced to a three-phase fault.

The tested fault from Figure 4.53 represent the good quality of converter as its good coordination between the control device parameters, which allows a very dynamic recovery from the fault [StBH06]. This is thanks to the developed control and applied resonant correctors that created a very manageable, adaptable and robust converter.

4.10. Conclusions

In this chapter, it was presented a new proposal for controlling the MMC when it is working in AC-AC in YDY topology. This proposal showed excellent results when some external instabilities were applied.

The primary control of this converter is used to calculate the input and output currents, which were corrected by using a resonant corrector. A resonant corrector allows controlling variables directly in a sinusoidal form, without needing a frame reference transformation. This was especially useful in the MMC AC-AC, because it allowed controlling signals that are naturally constituted of two different frequencies. In addition, when its parameters are correctly calculated, this corrector showed good and very dynamic performances, even during faulty situations.

The secondary control of the converter is organized in two different levels: the energy management and the reactive power control. Both of them follow changes in the capacitor voltages that are linked to the internal energy of the converter. When energy changes, a three-phase input current reference is calculated from this secondary control in order to maintain the energy at a given level. The idea is to keep the capacitor voltages below their operational limits.

A tertiary control allows maintaining the stability in the converter by minimizing the circulating current value, which is naturally zero in a balanced converter. This control links the input, output and arm currents in order to generate a voltage value at each arm; this will maintain any foreign harmonic inside the converter. A foreign harmonic is, for instance, any other frequency, different from the source-frequency, at the source-side of the converter. In order to reduce the circulating current value, all capacitor voltages must be as equal as possible; this is why the corrector also maintains the stability in the converter.

Finally, a low-level control was implemented to balance the M-quantity of submodule capacitors. This control called BCA is required for the switching model of the converter; it allowed acquiring a voltage response in the capacitor like the one obtained with the averaged model, and therefore every strategy developed with this simpler model was integrated into the switching model. It was chosen the most basic BCA for this task, and it has been proved to be of great performance when submodules are invariant.

Moreover, the control showed an outstanding performance when it is faced to external disturbances. It was tested: sharp frequency changes (even 10 Hz of change), sharp voltage variations (even $\pm 50 \%$), angle changes (from $\pm 30^{\circ}$ to $\pm 30^{\circ}$) and faulty situations of one-phase, two-phase and three-phase voltages drop. After each tested situation the control was able to recovery its normal operation. Moreover, these fault conditions only affected one side of the converter, keeping the other side unchanged, and demonstrating robustness of the developed control and capability of decoupling the input from the output side. In addition, a high quality of the converted waveforms was ensured with a THD lower than 0.4 % in steady state.

This converter has therefore proved to be capable of transferring power while converting frequency efficiently. It showed every needed performance to be installed offshore, for transmitting high quantity of power, over long distances in low frequency and without needing a low frequency transformer. Some of the proved performances required for offshore application were capacity of changing frequency, self-recovering after transient faults, high voltage and high power application and high quality of produced waveforms.

The following section will present some conclusions and further works.

Conclusions and Further Works

Summary

A new method to model and control a Modular Multilevel Converter in AC-AC direct topology was presented in this thesis work.

Three main features distinguish the proposed method from previous work on the MMC:

- 1- The introduction of a voltage transformer analogy with $Y\Delta Y$ coupling plus the superposition theorem to evaluate each frequency contribution, used to separate the input-side from the output-side frequencies.
- 2- The development of a method for calculating the required reactive power of the converter, which is essential to keep the balance of the internal energy of the converter.
- 3- The introduction of a new circulating current concept that does not have a defining frequency value as for other MMC cases. In this case, the current is calculated from unbalances of the system currents and compared to the arm currents of the converter.

The proposed method in this work aims at integrating renewable energy generation into an offshore power systems working at AC low frequency. Results on a developed iteration algorithm calculated the possible transmission distance for ACLF in a submarine cable. The obtained distance was proved to be between 200 km and 250 km without any kind of external compensation, with normal losses and acceptable voltage drop, which is suitable for offshore links.

The MMC developed in this work could be useful for changing frequency between many different system frequencies. It can be designed for high power and high voltage application. In addition, a transformer at the lowest frequency-side is not required. Simulations using the switching model of the converter were applied for testing the developed methods. Simulation results showed the high quality of signals with very low THD, which is required for high power systems. In addition, some robustness results can be highlighted:

- As it was expected, the converter was capable of following several power references.
- Some parameters mismatches going up to ± 10 % of variations of the arm inductances were endured by the proposed control without affecting the response of the system when compared to its balanced response.

- The converter was not robust against internal faults. This is because the balancing algorithm was developed to work in balanced operations and was not capable of identifying the faulty submodule.
- Sharp voltages-drop and over-voltages were tolerated by the control of the converter. Nevertheless, these situations might overpass the current or voltage limits of the converter; this depends on the intensity of the phenomenon.
- Sharp variations of frequency were also tested in the converter, and it had almost no effect at the output power that remained unchanged. Small variations of about 0.5 % and 3 % were observed at the input power.
- The converter was able to work as a frequency transformer, switching frequencies from 60Hz, 51 Hz, 50 Hz, 49 Hz, 40 Hz, and 30 Hz to 50/3 Hz. In addition, it was tested the case of transforming frequency from 50 Hz to 60 Hz, proving that: the output can also be at a higher frequency than the input and; it can be used for interconnecting power systems between 50 Hz and 60 Hz.

It was proved that the MMC can work in direct AC-AC configuration and is suitable for OPS. The development of this topology, its model and basic control, required a high amount of time to understand it. In consequence, some points from this vast subject have not been studied.

Perspectives

After dealing with the MMC in AC-AC topology and thanks to the results obtained in this work, some interesting research perspectives can be proposed.

Short-term perspectives are:

- As the proposed methodology was based on a power and control point of view, the study of this converter from its internal components point of view is also interesting. It is related to standardization of the nominal values of components that is made, until this date, only for AC-DC topology.
- An improvement of a balancing algorithm more adapted to this converter is required in order to improve the converter when has to face a loss of submodule for example.
- The development of a tripping algorithm related to the blocking states of submodules and an internal voltage protection is needed: it was found that in some particular faulty cases, the maximum voltage of capacitors risk to be overpassed.
- Real-time simulations of this converter in a benchmark. It is therefore required to change the internal configuration between the arms of the AC-DC converter and to use full-bridge submodules.
- The converter could be capable of working as a voltage transformer. Stepping-up of two times the input voltage was tested. Nevertheless, this kind of configuration might require higher limits of its components and more comprehensive developments.

Some long-term perspectives can also be proposed:

- This converter proved its efficiency to work at different frequencies; working as speed changer could be studied too. This will also require an adaptable multi-frequency resonant-controller in order to generate optimal signals quality.

- Simulations in a multi-buses offshore power system interconnected by low-frequency links and using MMC converter as converter stations.
- In this work, it was proved that not only the low frequency value (including DC) is always the best choice for transmitting power. The best choice will depend on the length of the system. In this scenario, it could be assumed an electrical system with several nominal operating frequencies. Therefore, for lower system dimensions, higher frequency values will allow reducing the size of components (transformer, circuit breakers, generators), while keeping the same power characteristics of the system.

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Appendix A. Wind Energy

A.1.1 Wind Energy

Energy is produced through the effect of wind on rotatory blades generating kinetic energy, which is then converted into electricity using power machines. When increasing the swept area (A_r), more power can be obtained from the wind, which is related to raising the radio size of blades (r) [Gipe95]:

$$A_r = \pi \cdot r^2 \tag{A.1}$$

$$P_{wind} = \frac{1}{2} \cdot \rho_{wind} \cdot A_r \cdot \left(\nu_{wind}\right)^3$$
 A.2

Where P_{wind} is the power of the wind measured in Watts, ρ is the density of dry air equal to 1.225 kg/m³ at average atmospheric pressure at the sea and temperature of 15° C and ν_{wind} is the velocity of the wind measured in m/s.

For a given turbine size, the total amount of power is mainly determined by the wind speed. Nevertheless, Betz's law [RaRa11] impose that the possible maximum extracted power from the wind is less than 16/27 of equation A.2, introducing a power coefficient (C_p) to the equation:

$$P_{wind} = \frac{1}{2} \cdot \rho_{wind} \cdot A_r \cdot \left(\nu_{wind}\right)^3 \cdot C_p$$
 A.3

When power is generated using the wind, some part of the power get loss in the process, the different steps are represented in Figure A.1.



Figure A.1. Wind generation and different steps of power losses.

For a given wind power value, almost 60% is lost from Betz's coefficient. The remaining power can be transformed, but still some losses are present over three different areas: aerodynamics, mechanics and electrical. All three depends on the turbine construction quality.

A.1.1 Wind Turbines

A general representation of the components inside a wind turbine can be made as follow:



Figure A.2. General parts of wind turbines.

The generation process start when the wind impacts the blades of the turbine; generating a torque in a low rotating speed. Depending on the used generator, a gearbox can be needed in order to increase the speed and to agree with the nominal speed of the generator. Finally a converter is used to maintain a constant voltage and frequency at the generators output.

A.1.2 Electrical wind generators

Two main kinds of generator technologies are suitable for wind power generation, the first one being the Permanent Magnet Generators (PMG) and the second one the Double-Fed Induction Generators (DFIG). A comparison between both technologies is made next:

	Table A.1. Comparison between DFIG and PMG [Swit14].			
Performance	Machine Type			
	DFIG	PMG		
Stator	Same Botor Coil	Same Permanent Magnet, multiple poles Not needed		
Carbon Brush	Available			
Maintenance	Heavy maintenance work	No need for rotor maintenance		
Converter	25-30% output power converted	100% output power converted		
Reactive Power Adjustment	Varies according to the generator speed	100% across the entire speed range		
Ability to connect and support Power Grid	Poor	Very good		
Advantages	Lower initial investment	No need for rotor maintenance High, medium and low speed High efficiency Generator is lighter and smaller Higher annual power output		
Disadvantages	Not applicable to medium or low speed generator Low efficiency	The initial investment is higher Requires professional design software for complicated calculation Requires more sophisticated process		

Supposing an Offshore network far from the coast where lower maintenance, robustness and able to connect independently to the power grid are the most important subjects. For that, PMG are the suitable solution.

Appendix B. Symmetrical Components

B.1 Method description

The method starts by defining an operator for referencing all phasors to only one of them. The operator $a = \angle 120^{\circ}$ will refer all phasors to phase A as it is presented in the following voltage equation:

$$\begin{bmatrix} V_a \\ V_b \\ V_c \end{bmatrix} = \begin{bmatrix} 1 & 1 & 1 \\ 1 & a^2 & a \\ 1 & a & a^2 \end{bmatrix} \cdot \begin{bmatrix} V_a^{2ero} \\ V_a^{positive} \\ V_a^{negative} \end{bmatrix} = \mathbf{A} \cdot \begin{bmatrix} V_a^{2ero} \\ V_a^{positive} \\ V_a^{negative} \end{bmatrix}$$
B.1

Currents are obtained using the same procedure. The equation of the line is then represented by the following equation, were the self-impedance of the line is Zs and the mutual one is Zm, and the voltage between cable ends for each phase are defined by $V_{aa'}$, $V_{bb'}$ and $V_{cc'}$.

$$\begin{bmatrix} V_{aa'} \\ V_{bb'} \\ V_{cc'} \end{bmatrix} = \begin{bmatrix} Zs & Zm & Zm \\ Zm & Zs & Zm \\ Zm & Zm & Zs \end{bmatrix} \cdot \begin{bmatrix} I_a \\ I_b \\ I_c \end{bmatrix}$$
B.2

Introducing equation B.1 into equation B.2, the voltage equation of the line yields:

$$\boldsymbol{A} \cdot \begin{bmatrix} V_{aa'} & Zrro \\ V_{aa'} & Positive \\ V_{aa'} & Pogative \end{bmatrix} = \begin{bmatrix} Zs & Zm & Zm \\ Zm & Zs & Zm \\ Zm & Zm & Zs \end{bmatrix} \cdot \boldsymbol{A} \cdot \begin{bmatrix} I_a & Zrro \\ I_a & Positive \\ I_a & Positive \\ I_a & Positive \end{bmatrix}$$
B.3

Then, the transformation of the line impedance matrix into symmetrical components can now be defined from equation B.3 as:

$$\begin{bmatrix} Z_0 \\ Z_+ \\ Z_- \end{bmatrix} = \mathbf{A}^{-1} \cdot \begin{bmatrix} Zs & Zm & Zm \\ Zm & Zs & Zm \\ Zm & Zm & Zs \end{bmatrix} \cdot \mathbf{A} = \begin{bmatrix} Zs + 2Zm & 0 & 0 \\ 0 & Zs - Zm & 0 \\ 0 & 0 & Zs - Zm \end{bmatrix}$$
B.4

The zero sequence impedance is represented by Z_0 , the positive sequence by Z_+ and the negative sequence by Z_- .

Appendix C. Possible configurations of the MMC AC-AC

C.1 MMC AC-direct, delta coupling

The transformer analogy made in section 3.3.13.3.3 may allow other coupling connections for the MMC AC. In this thesis work, developments and tests were only based on the YDY coupling. However, delta connection of the source and the load and a combination of both are also possible as presented in Figure C.1. Changing the source connection will not make any difference to the control if the currents and voltages are measured using the same equivalent.

Supposing the entire model of a DDD coupling, this will lead to decoupling currents between phases, as for instance, i_{ABs} will be equal to i_{abs} of the converter, and i_{abo} of the converter will be equal to i_{abo} of the load without changing neither angle nor amplitude. Nevertheless, these are only conjectures and no practical experiments were performed.



Figure C.1. Different source and load connections for the MMC AC in $\Delta.$

A configuration in Y of the converter may be possible too as presented next.

C.2 MMC AC-direct, star coupling

The star-coupling converter does not have a clear path for circulating currents as the delta one presented before. Therefore, the control might be more complicated to implement. Nevertheless, it is possible to create a delta path of component of any kind as inductances or capacitances between phases that will be controlled by the converter. One of the advantages in this topology is the middle neutral point that interconnects all arms in one point; this neutral path could be used in order to keep a better balancing

in the system. Moreover, a three-phase balanced circulating current will be zero in that point.



Figure C.2. Proposal of the converter in star coupling.

Appendix D. Frequency measurements

D.1 PLL definition

A phase lock loop (PLL) is based in an oscillator that can provide a variable frequency compared to a source reference coming from a quartz or from another PLL. There are several PLL architectures divided in three groups: analogical, digital and mixed [Kief14].

As the MMC AC-AC is connected to a two system frequencies, two different PLL are therefore used; one for the high frequency side, another for the low frequency side. The architecture of the used PLL is presented in the following figure.



Figure D.1. Applied PLL for simulations with the MMC [Matl15].

The PLL tracks the frequency and phase of a sinusoidal signal by using an internal frequency oscillator. The input signal is compared with the internal oscillator signal. Any DC component of the compared signal, which is proportional to the phase difference between these two signals, is extracted and rejected with a variable frequency mean value calculation. A proportional-integral-derivative (PID) controller with an automatic gain control keeps the phase difference to zero by acting on the controlled oscillator. The PID output, corresponding to the angular velocity, is filtered and converted in hertz for frequency measures.

Appendix E. PI Controller

E.1 The PI parameters

A Proportional-Integral (PI) is the classic corrector used for controlling continuous systems. It is composed of a proportional (kp) and an integrator (ki) constants. The representation of this corrector using Laplace operator (S) is of the form of:

$$C_{PI}(S) = \frac{k_P \cdot S + ki}{S}$$
E.1
$$\frac{k_p}{k_p} + \frac{k_i}{s}$$

Figure E.1. PI corrector.

The PI corrector will be used for controlling continuous signal coming from the submodules capacitors. The principal drawback of this converter is that is not capable of eliminating the steady state error when it works with an alternative signal. This can be demonstrated with a Bode diagram in a close-loop of a first order system, as the case of the MMC arms.



Figure E.2. Frequency response of a PI corrector. [Etxe09]

The corrector as a pole in the origin and a zero placed at ki/kp, measured in rad/s. Therefore, the gain is theoretical infinite at the origin and it decreases until the value 20 log(kp), measured in dB.

The used controller was built using PID tuner application from Matlab Toolbox [Matl15]. This tool allows choosing the required values for the needed controller in a plant.

The open loop transfer function of this corrector is:

$$H_{BO}(S) = C_{PI}(S) \cdot H(S)$$
 E.2

With H(S)s the converter model presented in chapter 3. Then the closed loop function is:

$$H_{CL}(S) = \frac{C_{PI}(S) \cdot H(S)}{1 + C_{PI}(S) \cdot H(S)}$$
E.3

For the case of an alternative signal at the input side, the static error will not be zero and cannot be suppressed. Moreover, this error will quickly increase with the frequency value at the controller input. In this case it is needed a frame referenced PI corrector, as for instance with the use of Park's transformation. By turning at the same frequency than the alternative signal, a continuous signal can be recreated and this allows the use of a PI controller. This is presented in the following figure.



Figure E.3. Using PI in a frame referenced with Park's transformation.

This example represents only one frequency transformation. The use of other frequencies will require as multiple frame references as frequencies with their respective PI controller. Moreover, this representation requires the knowledge, or the measure with a PLL of any frequency that could have the alternative signal at the measurement point. Nevertheless, as it was presented in Chapter 4, many of these frequencies cannot be measured directly nor predicted because they exist at the same time in an alternative link. In consequence, the PI corrector is not suitable for this kind of application.

Appendix F. Complementary Results

F.1 Average Model using 10 mH as arm inductance

These results complement those presented in section 4.6. Used parameters where presented in Table 4.5, for a 10 kW converter.

In the first part of simulations, the converter operation when is faced to power references changes will be presented: the energy control keeps the balance between the input and the output power, even when the power changes of direction as it is presented in the following figure:



Figure F.1. Power variations in the MMC averaged model in low voltage.

The circulating current control is also tested at the second part of simulation, where it is disabled in order to compared it with normal operations of the converter.

The maximum reached power of the converter was 3.2 kW per phase. It was not able to reach 3.5 kW because of the used arm inductance values of 10 mH (when the converter is originally designed with a 2.5 mH inductance); this is because the voltage drop in each arm becomes more important. Therefore, the required submodule voltage reaches its limit before being able to transmit the maximum power as it was explained in section 3.4. The simulation made with a 2.5 mH inductance showed that the converter reached the maximum power limit without any problem (see subsection 4.6.2). Figure F.1a, shows that the converter can supply positive, negative or zero power value without any difficulty. Finally, the input reactive power has the main function of keeping stable the capacitor voltages, and therefore a lot more variations of its values when it is compared with the output reactive power are appreciated. As a result, the capacitor voltages are all around the given reference and changes when the power does as it can be appreciated in Figure F.2.



Figure F.2. Capacitor voltages in the MMC averaged model in low voltage.

In Figure F.2 it is first presented the upper and lower voltages waveforms in steady state (after 90 seconds), which are expected in that way as it was explained in subsection 4.3.2. The maximum voltage ripple at maximum power is about 50V, which is 7% of the converter nominal voltage and therefore provide very good results when compared to other AC-DC MMC [BBEL13, Berg15, TRWA15, ZBBF16]. It can be also appreciated that all capacitor voltages are lower than 800 V for the entire simulation, as the converter requires it (see Table 4.5).

Power transmission in the other sense is evaluated from 28 seconds to 48 seconds and it does not affect the normal operations of the converter.

At around 60 seconds, the circulating current control of the system was deactivated and many oscillations in the capacitor voltages can be appreciated, as it was the case for the high voltage system. The circulating current control is reactivated again at 82 seconds and it stabilizes the system. Low ripple amplitudes are related to very low (or even zero) power in the system.

The circulating current response can be appreciated in Figure F.3. In the entire simulation, the control seems to keep the circulating current value at around 0.02 A; this value is reduced when the power in the system does also. When the corrector is turned off at 63 seconds, the current seems to raise its value progressively, and even if it is represented by a very small value of current, it has a big effect in the voltage of capacitors. From 74 seconds, the converter was working at reduced power and it raises to 3 kW per phase until 77 seconds. From this simulation time, there are many oscillations in the system that tends to unbalance the capacitor voltages.



Figure F.3. Circulating current of the MMC averaged model in low voltage.

The presented results showed the utility of the circulating current control to reduce unbalances in the capacitor voltages system. Nevertheless, this control is not enough to ensure a durable stability; this could be appreciated when the circulating current control was turned off and the capacitor voltages started to oscillate. In the results, it was appreciated that, the circulating current amplitude is not the primordial problem of this converter; the real problem would be the destabilization of the capacitor voltages that are related to the amplitude of the circulating current.

The arms currents in the converter and a comparison with the circulating current in steady state are exposed in the following figure.



Figure F.4. Arms currents ia1 and ia2 and total circulating current in steady state.

In Figure F.4 it can be appreciated that for both arm currents, the 50 Hz current value is bigger than the 50/3 Hz current value. This is because the source side supplies the power losses of the converter. Odd harmonics are however detected in the circulating current value. These harmonics, corrected with the circulating current control, are of low amplitude. Nevertheless, they could generate some heating problems in the converter if they are not properly corrected.

F.2 Switching Model – Reactive Power Variations

Some reactive power variations of the load were performed to the converter as a complement of results from section 4.8.

Figure F.5 shows that the capacitor voltages were affected after raising the reactive power at 5 seconds, nevertheless the capacitor voltages stabilizes after one second. An extreme variation is made at 6 seconds, going from +30° to -30°: this exchanged the value of the capacitor voltages, which is a normal effect of the reactive power as it was explained in subsection 4.4.2. However, the proposed control is capable to follow the reference and to stabilize the capacitor voltages to continue its normal operations. The circulating current value remained at the same level the entire simulation.



Figure F.5. Results with parameters mismatch and angle variations.

It can be appreciated that reactive power variations are accepted in the converter. Sharp changes were tested as example between 5.9 s and 6.5 s, and even if voltages are far from the reference (almost 7 %), the converter was still able of keeping its operations. It can be seen that the input reactive power has almost the same value than the output reactive power but with an opposite sign, this was explained in subsection 4.4.2.

Some changes can be evaluated regarding the voltage reference at 4.5 seconds, which was 73 V in this case, therefore:

- From 4.5 s to 5.1 s, upper and lower voltages passed from ± 2.62 % from a given voltage reference to ± 4.513 %, ± 3.82 % respectively. This means that the reactive power tends to separate upper from lower voltages. This is observed also in the following change of the reactive power.
- From 5.9 s to 6.5 s, upper and lower voltages passed from +3.75 % and -2.21 % to
 -3.0 % and +6.58 % respectively. An exchange between upper and lower voltage of capacitors is also showed and it is related to the change of sign in the reactive power.
- The individual voltages amplitudes remained the same from 4.5 s to 6.5s as the same apparent power is in the system. Nevertheless the maximum amplitude from the lowest capacitor voltage to the biggest raised to 215 %, passing from 3.39 V to 7.29 V
- Finally, active and reactive power at 4.5 s, 5.5 s and 6.9 s are:
 - t = 4.5 s, Sin = 1.51 kW 39.2 VA and Sout = 1.50 kW 0 VA.
 - t = 5.5 s, Sin = 1.32 kW 793.9 VA and Sout = 1.31 kW + 746.1 VA.
 - t = 6.9 s, Sin = 1.31 kW + 715.7 VA and Sout = 1.30 kW 738.3 VA.

F.3 Switching Model – 50 Hz to 60 Hz conversion mode

These results complement simulation results from subsection 4.9.3, where different values of frequencies are studied. In this case, results prove that the converter could be used to interconnect 50 Hz with 60 Hz systems, moreover the output is at the highest frequency in this case, and the converter operates normally as the input and the output power are kept constant the entire simulation.



Figure F.6. Power and capacitor voltages of the MMC converting AC 50 Hz to AC 60 Hz.

It was observed that new waveforms from the capacitor link were obtained. As it was described before in this work, that depends on the arm voltages (calculated in subsection 4.3.4), which this time are a fast frequencies at both sides. The converter circulating current also presents a different waveform when compared to results in section 4.8 and 4.9.



Figure F.7. Circulating current waveforms when working in a 50 Hz-60 Hz system.

F.4 Switching Model – Evaluation of output-side disturbances

These results complement simulation results from subsection 4.9.4, where different voltage dips were evaluated to the input-side source. In this case, three fault tests have been made at the output side: one-phase and two-phase during 100 ms and a three-phase disturbances during 30 ms. Results are presented next:

Time [s]	6	7	8
V _{Ao} factor	0.8	0.8	0
V_{Bo} factor	-	0.8	0
V_{Co} factor	-	-	0

Table F.1. Voltage dips to individual phases - output side

In this case, the disturbance is propagated to the input side current as it can be appreciated in the following Figure F.8. This is expected as the energy control explained in section 4.4 follows changes at the output side of the converter. These results can be compared to the steady state representation from Figure 4.42



Figure F.8. Current waveforms for each applied disturbances at the output side.

The converter was able of continuing its operations after each fault. Active and reactive power, capacitor voltages and arm current behaviours are exposed in Figure F.9. For this kind of faults, both sides of the converter were affected. Nevertheless, when analysing the obtained responses, it can be seen that peak of power are 12.5% bigger at the input side than at the output side for all cases. This proves that the converter control is reacting with the same intensity for each case.

For the tests made, neither capacitor voltages nor arm currents overpassed their maximum limits.



Figure F.9. Active and reactive power, capacitor voltages and arm currents faced to dips at the output side.

F.5 Switching Model – Circulating current from simulation results

These results complement simulation results from sections 4.8 and 4.9, where the circulating current from each result is presented next. First, it is presented the circulating current from the steady state analysis presented at the beginning of section 4.9:



Figure F.10. Circulating current from steady-state analysis.

The obtained circulating current during the entire range of simulation plus a harmonic analysis made during 100 cycles of the lowest frequency of the system is
presented in Figure F.10. During all evaluated cycles, the system was considered as balanced and unchanged. The maximum evaluated harmonic was the 600th, which is the 10.000 Hz frequency equivalent. As the circulating current present a very low value, highest harmonics coming from the switching devices can be appreciated. Only odds harmonics are observed, and what is interesting is that, during the 100 cycles, the circulating current managed to eliminate the first and the third harmonic of this current. In addition, this is exactly the performance expected from this controller; supposing the presence of 1st (50/3 Hz) or 3rd (50 Hz) harmonics. This means that a fragment of one of these currents is not well distributed to the input nor the output side of the converter, creating a circulation of unnecessary currents that will heat the converter arms.

As the response is the same, harmonics will not be analysed for circulating currents applied during the following tests: power variations is related to section 4.8.2; mismatched parameter to section 4.9.1; frequency variations to section 4.9.3, and; faults to voltages dips from section 4.9.4.



Figure F.11. Circulating current for various tests.

It can be appreciated that for all made tests the circulating current remained not bigger than 0.3 A. The only special case is related to faults in the system; especially when 0 V one phase fault was tested and the circulating current amplitude raised to 0.6 A during the fault.