

Investigating Converter Control Interactions in the Transmission Grid

Von der Fakultät für Elektrotechnik und Informationstechnik
der Rheinisch-Westfälischen Technischen Hochschule Aachen
zur Erlangung des akademischen Grades eines
Doktors der Ingenieurwissenschaften
genehmigte Dissertation

vorgelegt von

Matthias Andreas Quester, M.Sc.
aus Aachen

Berichter: Universitätsprofessor Dr.-Ing. Albert Moser
 Universitätsprofessor Antonello Monti, Ph.D.

Tag der mündlichen Prüfung: 11. Juni 2021

Diese Dissertation ist auf den Internetseiten der Universitätsbibliothek online verfügbar.

Acknowledgment

This thesis was written during my time as a research associate at the Institute for High Voltage Technology, and the successor institute, the Institute of High Voltage Equipment and Grids, Digitalization and Energy Economics at RWTH Aachen University.

My special thanks go to Univ.-Prof. Dr.-Ing. Armin Schnettler for giving me the opportunity to pursue a Ph.D. and support my thesis. I would like to express my gratitude to Univ.-Prof. Dr.-Ing. Albert Moser for supervising my thesis. The fruitful discussions and the feedback have contributed significantly to the success of my thesis. Moreover, I would like to thank Univ.-Prof. Antonello Monti, Ph.D. for his interest in my research and co-supervising my thesis.

I would like to thank my colleagues at the institute. In particular, the discussions and support of the research team System Stability and the PROMOTioN project team have helped me a lot to advance my research. I would also like to thank Andreas Roehder who believed in my skills and abilities and encouraged me to work as a research associate. And thanks to Moritz Mittelstaedt for making me aware that there is a problem worth investigating. My thanks to Carola Meier, Fisnik Loku, Markus Knittel, Philipp Ruffing, and Sandor Simon for reading and commenting on the first draft of my thesis. I would also like to show my appreciation to all the students who supported my research.

Lastly, I want to thank my parents who supported my education and made it possible to study abroad for two semesters. And special acknowledgments go to my spouse, Lyn Zeichner, and my father-in-law, Sam Zeichner, for their assistance in proofreading and editing.

Aachen, July 2021

Matthias Quester

Abstract

Motivation, Goal and Task of the Dissertation

The progressive integration of HVDC systems into the transmission grid has led to an increase in HVDC converters, particularly MMCs, that impact system stability. It has been seen that the converters and their control system can interact with the connected AC system, causing high oscillations in the grid current and voltage. To prevent outages resulting from the oscillations, the interactions need to be identified in the planning stage of HVDC systems. A promising method is to investigate the interactions in the frequency domain using impedance models. For this purpose, several models have been proposed to represent the converters. However, relying on full-system knowledge, these models cannot represent manufacturer converters, particularly their control system. In addition, the impact of simplified MMC models on the analysis is not addressed. Thus, this work aims to analyze the suitability of simplified converter models and develop a method to investigate converter control interactions, which is also applicable for black-boxed systems.

Major Scientific Contributions

This work develops a numerical method applicable for deriving frequency-dependent impedances of time-domain converter models, physical laboratory converters, and converter control replica systems without full system knowledge. Based on an online perturbation approach, the method directly determines the impedance during the measurements, ensuring high accuracy while keeping data requirements to a minimum. As a result, numerous impedances can be derived for different operating points and control settings, describing the converters' frequency behavior. Thus, the method facilitates studies employing the actual frequency behavior of converters to investigate converter control interactions.

Comparing the impedance of a small-scale laboratory MMC to the impedance obtained from time-domain average value models shows that the simplified model can be used to represent the electrical structure of the MMC when interactions are assessed that cause oscillations between 30 Hz and 3 kHz. The comparison further reveals that the converter transformer causes a deviating frequency behavior at sub-synchronous frequencies that cannot be seen in the time-domain model. The investigation indicates when simplified MMC models are adequate for assessing converter control interactions with the transmission grid.

Using the derived impedance models of the MMCs, several test cases reveal in which grid situations interactions might cause unstable systems. The results show that interactions

can cause an unstable system for moderate short-circuit power ratios depending on the converter's control system and the power flow. Operating in grid-forming control mode, the MMC can sustain a stable operation point for lower short-circuit ratios compared to grid-following operating mode.

Major Publications

1. M. Quester, V. Yelliseti, F. Loku, and R. Puffer, "Assessing the Impact of Off-shore Wind Farm Grid Configuration on Harmonic Stability," in *2019 IEEE Milan PowerTech*, Milan, Italy, Jun. 2019 - Jun. 2019, pp. 1–6.
2. M. Quester, F. Loku, V. Yelliseti, and R. Puffer, "Online Impedance Measurement of a Modular Multilevel Converter," in *2019 IEEE PES Innovative Smart Grid Technologies Europe (ISGT-Europe)*, Bucharest, Romania, Sep. 2019 - Oct. 2019, pp. 1–5.
3. M. Quester, F. Loku, V. Yelliseti, and A. Moser, "Frequency Behavior of an MMC Test Bench System," in *2020 6th IEEE International Energy Conference (ENERGYCon)*, Gammarth, Tunis, Tunisia, Sep. 2020 - Oct. 2020, pp. 74–79.
4. M. Quester, F. Loku, O. El Azzati, L. Noris, Y. Yang, and A. Moser, "Investigating the Converter-Driven Stability of an Offshore HVDC System," *Energies*, vol. 14, no. 8, p. 2341, 2021, doi: 10.3390/en14082341.

Contents

Acknowledgment	iii
Abstract	v
List of Abbreviations	xi
Nomenclature	xiii
1 Introduction	1
1.1 Introduction	1
1.2 Converter Control Interaction Incidents	2
1.3 Representation of Modular Multilevel Converters in Grid Studies	5
1.4 Literature Review	7
1.5 Research Gap and Research Questions	7
1.6 Objective and Structure	9
2 Analysis	11
2.1 Classification of Interaction Phenomenon	11
2.1.1 Power System Dynamics	11
2.1.2 Power System Stability	12
2.1.3 Power Quality	16
2.1.4 Terminology	17
2.1.5 Conclusion	18
2.2 Discussion of Methods and Approaches	18
2.2.1 Time-Domain Simulation	19
2.2.2 Laboratory Test Bench	20
2.2.3 Eigenvalue Analysis	21
2.2.4 Impedance-based Stability Analysis	23
2.2.5 Conclusion	29
2.3 Impedance Modeling	31
2.3.1 Frequency Coupling	32
2.3.2 dq-Domain	32
2.3.3 Sequence Domain	34
2.3.4 Analytical Derivation	37
2.3.5 Numerical Derivation	43
2.4 Method Approach	47
2.4.1 Choice of Domain	47
2.4.2 Choice of Impedance Derivation Approach	48
2.4.3 Method Requirement	48
2.5 Investigation Approach	49

3	MMC Test Bench and EMT Models	53
3.1	MMC Test Bench Laboratory	53
3.1.1	Power Amplifier	54
3.1.2	MMC Rack	55
3.2	Electromagnetic Transient Time-Domain Models	56
3.2.1	MMC Electrical Model	56
3.2.2	AC Grid Model	57
3.3	MMC Control System	59
4	Impedance Model Derivation	63
4.1	Online Impedance Measurement Method	63
4.1.1	Frequency Generator	65
4.1.2	Perturbation Source	65
4.1.3	Signal Processor	69
4.1.4	Impedance Calculator	73
4.1.5	Impedance Measuring: Laboratory	74
4.1.6	Impedance Measuring: Simulation	76
4.1.7	Method Validation and Evaluation	76
4.2	Impedance Modeling	79
4.2.1	Frequency Response Analysis	79
4.2.2	Frequency Coupling	83
4.2.3	AC Grid Impedance Model	86
5	Results	89
5.1	Identification of Model Limitations	89
5.1.1	MMC Test Bench and Laboratory-Scaled Model Comparison	89
5.1.2	Influence of Transformer	91
5.1.3	Conclusion	92
5.2	Impedance-based Stability Assessment Validation	93
5.2.1	Frequency-Domain Stability Assessment	93
5.2.2	Time-Domain Validation	97
5.3	Stability Assessment	99
5.3.1	Test Case Analysis	99
5.3.2	Conclusion	106
6	Summary and Outlook	109
6.1	Summary	109
6.2	Outlook	111
	References	113
	List of Publications	129
	List of Supervised Theses	131

List of Figures	133
List of Tables	137
A Appendix	139
A.1 Power Amplifier Bandwidth Verification	139
A.2 Scaling Up Laboratory-Scaled Impedances	140
A.3 MMC Test Bench Frequency Response	143
A.4 MMC TB and LS Model Negative-Sequence Impedances	144
A.5 Stability Assessment Nyquist Plots	145

List of Abbreviations

AC	Alternating Current
CCM	Component Connection Method
CORDIC	Coordinate Rotation Digital Computer
CPU	Central Processing Unit
D/A	Digital-to-Analog
DAE	Differential-Algebraic Equations
DC	Direct Current
DDS	Direct Digital Synthesis
DFIG	Double-Fed Induction Generator
DFT	Discrete Fourier Transform
DUT	Device Under Test
EMT	Electromagnetic Transient
FFT	Fast Fourier Transform
FPGA	Field Programmable Gate Array
FS	Full-Scale
GB	Gigabyte
GDP	Grid Development Plan
GUI	Graphical User Interface
HVDC	High-Voltage Direct Current
IbSC	Impedance-based Stability Criterion
IGBT	Insulated-Gate Bipolar Transistor
IPR	Intellectual Property Right
LS	Laboratory-Scaled
LTI	Linear Time Invariant
MB	Megabyte
MIMO	Multiple Input and Multiple Output

MMC	Modular Multilevel Converter
MOSFET	Metal–Oxide–Semiconductor Field-Effect Transistor
NLM	Nearest-Level Modulation
PA	Power Amplifier
PC	Personal Computer
PHIL	Power-Hardware-in-the-Loop
PI	Proportional Integral
PLL	Phase-Locked loop
PM	Phase Margin
pu	per unit
PWM	Pulse-Width Modulation
RES	Renewable Energy Sources
RMS	Root Mean Square
RTS	Real-Time Simulator
SCR	Short-Circuit Ratio
SISO	Single Input and Single Output
SM	Submodule
STATCOM	Static Synchronous Compensator
TB	Test Bench
THD	Total Harmonic Distortion
TRL	Technology Readiness Level
TSO	Transmission System Operator
TYNDP	Ten-Year Network Development Plan
VSC	Voltage Source Converter

Nomenclature

Symbols

χ	Signal that can be either a current or a voltage
Δt	Time step
ω	Angular frequency
θ	Grid angle
$\underline{a}$	Complex phasor
$\underline{k}$	Complex scaling factor
φ	Phase angle
a_n, b_n	Fourier coefficients
C	Capacitance
e	SM string voltage of MMC arm
F	Frequency of one period
f	Frequency
F_{DFT}	Fourier fundamental of the signal processor
f_{eval}	Frequencies for which the measured signal is evaluated
f_{pert+}	Positive coupling frequency
f_{pert-}	Negative coupling frequency
f_{pert}	Frequency for which the system is perturbed
f_{series}	Vector of a series of perturbation frequencies
I, i	Current
i_c	Coupling current at $f = 2f_{grid} - f_{pert}$
$i_{c,c}$	Coupling current at f_{pert} resulting from grid response
$i_{c,p}$	Coupling current at $f = 2f_{grid} - f_{pert}$ resulting from grid response
k_φ	Phase scaling factor
L	Inductance
m	Number of discrete time steps
MP	Measurement parameters
N	Number of samples within one integration period
n	Number
n_{onl}	Number of recorded values online method
n_{ppr}	Number of recorded values post-processing method

P	Active power
Q	Reactive power
R	Resistance
S	Apparent power
s	Complex variable equal to $j\omega$
S_{SC}	Short-circuit power
SCR	Short-Circuit Ratio with respect to $P_{n,MMC}$
T	Time of one period
t	Ttime
t_{delay}	Time a signal is delayed
T_{valid}	Time span outputting valid results
ts	Sampling time
U, u	Voltage
u_c	Coupling voltage at $f = 2f_{grid} - f_{pert}$
W	Window of signal processor
X	Reactance
Y	Admittance
$Y_{c,c}$	Additional coupling admittance
Y_{pert}	DUT impedance without coupling considered
Z	Impedance
Z_{DUT}	Impedance of connected Device Under Test

Subscripts

abc	Three-phase system phases a , b and c .
AC	AC signal
$AC,1$	Primary side of transformer
$AC,2$	Secondary side of transformer
Arm	MMC arm
C	Converter
c	Coupling
CC	Circulating current signal
CPU	Part of the method that is implemented on the CPU
$ctrl$	Control signal for power amplifier
d	d-axis
DC	DC signal
f	Frequency

FPGA Part of the method that is implemented on the FPGA

grid AC grid

l Lower arm of MMC

max Maximum

meas Measured or numerically derived signal

min Minimum

n Nominal

pert Perturbation

q q-axis

s Samples

SC Short circuit

SM Submodule

Tr Transformer

u Upper arm of MMC

Superscripts

*

Reference value

+

Positive DC pole

−

Negative DC pole

n

Negative Sequence

p

Positive Sequence

SM

Submodule

1 Introduction

1.1 Introduction

Both the European Union and Germany have set climate goals to reduce greenhouse gas emissions and become almost climate-neutral by 2050. This requires a transformation of the energy sector from fossil fuels to nearly climate-neutral Renewable Energy Sources (RES) [1]. Starting with the energy transition, Germany monitors its climate goals by annually evaluating the status quo and assessing measures taken [2]. Consequently, Germany is significantly expanding offshore wind capacities. Moreover, Germany decided to gradually decommission its coal-fired power plants by 2038 [3].

As of 2019, the offshore wind capacity in operation in Germany has reached 7.5 GW, having the second-largest capacity worldwide [4]. Due to the long distances and the resulting reactive power demand, many offshore wind farms are connected to the shore via High-Voltage Direct Current (HVDC) transmission systems. This requires converters, both on- and offshore, to transform the transmitted power from Alternating Current (AC) to Direct Current (DC) and vice versa [5]. By February 2020, the Transmission System Operator (TSO) TenneT connected 96 % of the 7.1 GW offshore wind power capacity using HVDC technology and nine offshore as well as onshore converters [6]. Additional 3 GW are expected to be connected by 2027, further increasing the number of HVDC converters offshore and onshore [6]. Due to the increasing power transfers, the TSO partnered with five HVDC suppliers to develop a new 2 GW offshore grid connection system [7]. Related projects are expected to operate in 2028 and 2029 in the Netherlands, and 2030 in Germany [8].

At the same time, the shift to more RES introduced new challenges to the transmission system. Because RES rely on intermittent wind conditions and solar irradiation, they are more volatile and less predictable. Additionally, the decommissioning of many coal-fueled and nuclear power plants, along with the increase of offshore wind power capacities in the North Sea, shifts the power generation to the North while the industrial load centers remain in the South and West of Germany. As a result, power flows from North to South have increased significantly, requiring network expansions and other measures to ensure a reliable and stable transmission system operation.

To overcome these challenges, the ENTSO-E publishes the Ten-Year Network Development Plan (TYNDP). By developing future scenarios for the European power system, the TYNDP assesses potential projects reinforcing the European transmission grid. In Germany, the TSOs identify and publish required grid expansions and reinforcement measures in the Grid Development Plan (GDP) [9]. Both TYNDP and GDP have identified HVDC systems as an integral part for the grid expansions [3, 10]. HVDC systems

allow for long-distance power transmission without reactive power losses. Moreover, modern converter technology, such as Voltage Source Converters (VSCs), can independently control the active and reactive power. Thus, they can provide ancillary services to the grid. For instance, by providing additional reactive power, the voltage profile can be maintained in case of under- or over-voltages. Hence, several HVDC projects have been proposed and are currently under construction [11]. In Germany, SuedLink, SuedOstLink and Ultranet are proposed to link the northern and southern parts of the transmission system [12] to facilitate the increase of offshore wind power. Additionally, the 2020-commissioned ALEGrO project connects the German and Belgian transmission system [13], facilitating an increase of trade in electricity in Europe [10].

Because HVDC technology is a major part of the energy transition, enabling the integration of RES into the transmission grids, more and more HVDC links are integrated into the transmission system. Consequently, the number of HVDC converters increases, which impacts the stability of the transmission system. State-of-the-art HVDC converters have response times of 30 ms to 50 ms [14]. Therefore, they react significantly faster than traditional power system components, requiring electromagnetic phenomena to be considered when assessing the transmission system's stability. With the large-scale integration of converters, a new stability problem was observed in the power system. TSOs experienced that HVDC converters interacted with the AC system, causing high oscillations in the grid current and requiring a shutdown of the converter in order not to damage the components of the system [15]. Investigations have shown that this stability problem is caused by the converter's control system, interacting with the connected AC system.

1.2 Converter Control Interaction Incidents

This section gives an overview of the reported interaction phenomenon related to converters and their control systems and the connected AC grid. The methods to assess the interaction phenomenon can be analyzed in the following chapter by reviewing the reported incidents.

Interactions between converters and the AC system were first reported for locomotives using high-frequency converters. Starting in April 1995, on several occasions, regional trains in the Swiss railway were forced to stop as protection systems shut down the locomotives due to extraordinary high harmonic currents. Investigations showed that the experienced shutdowns correlated with the integration of new locomotives using high-switching frequency converters. In certain grid situations, e.g., low load scenarios on the weekend, the converters interacted with each other, and the supplying AC network [16].

After commissioning of the first HVDC-connected wind farm Bard Offshore 1, the TSO TenneT experienced a previously-unknown phenomenon, requiring a temporary shut down of the HVDC system, BorWin 1, that connects Bard Offshore 1 [15, 17]. According to TenneT, the connection of a cable changed the resonance frequency of the AC

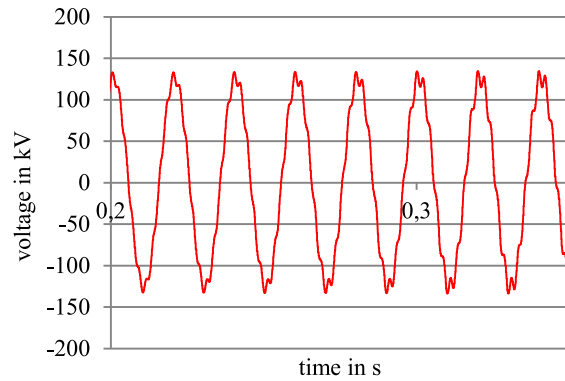


Figure 1.1: Voltage measurements at TenneT wind farm [18].

offshore grid to which the converter was connected. This resulted in the unstable operation of the converter. Given that this problem was unknown to the power supply industry during the project's planning phase, potential converter interactions with the AC grid were not considered in the project. Measurements of distorted currents and voltages led to the conclusion that oscillations can occur at nearly every frequency [15]. A subsequent publication pointed out that an observed frequency of 451 Hz in the recorded voltages of a wind farm shown in Figure 1.1 could not be attributed to steady-state harmonics fed in by the converters. Even under worst-case conditions, present steady-state harmonics such as the 9th harmonic could not explain the oscillations due to the high magnitude of the current, and the deviating frequency of the measured oscillations [18]. Therefore, unstable controllers interacting with grid resonances are thought to have caused the problem. Specifically, the current controllers and the Phase-Locked loop (PLL) are considered likely causes, in addition to the unique characteristics of the AC offshore grid. Because the offshore grid comprises cables with no connection to rotating masses resulting in low short circuit power, the grid's resonance frequency is shifted in regions of the controllers' bandwidth [15, 18]. Further investigations confirmed the theory that the converter and the AC grid formed a marginally-stable system due to a marginally damped series resonance-inducing oscillations at the series resonance frequency. This explains the non-characteristic frequency, and the high magnitude of the distortion [19].

Similar incidents have also been reported for onshore HVDC systems. The 2015 commissioned HVDC INELFE project connecting Spain and France by 2 GW HVDC-VSC link experienced resonances between the HVDC system and the AC grid, as shown in Figure 1.2 [20]. The system operator RTE found that the oscillations, referred to as harmonic interactions, resulted from network configurations that lead to harmonic interactions between the AC grid and the HVDC controls, which could be reproduced by real-time simulators and control replicas [21]. The investigations showed that the AC-line parameters and grid configuration and the short circuit power influence the interactions and that for reproducing the phenomenon, the exact behavior of the controls given by the actual control system is required [20, 22].

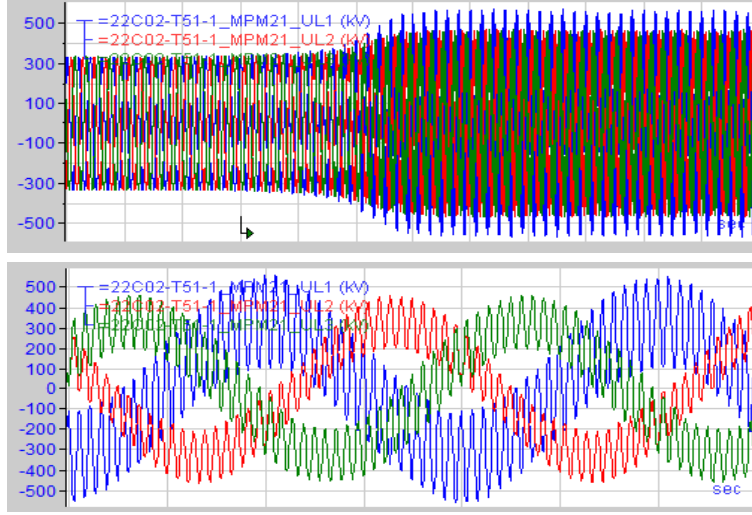


Figure 1.2: On-site voltage measurements INELFE project [20].

In April 2017, China Southern Power Grid reported a 1270 Hz resonance between the 525 kV AC grid and the connected ± 350 kV and 1 GW Luxi back-to-back VSC-HVDC system. After being subjected to the 1270 Hz oscillations, reaching a magnitude of 68.9 kV and 60.1 A (as shown in Figure 1.3), the tap changer of the converter transformer was out of order due to frequent operation [23, 24]. Using AC system impedance scans, the controllers were modified so that the AC system could not get into resonance with the converter, considering every potential operation mode of the HVDC system. It was further revealed that the high frequency of the oscillations is influenced to a great extent by the long delay time of the Modular Multilevel Converter (MMC)-VSC, negatively affecting system stability [24].

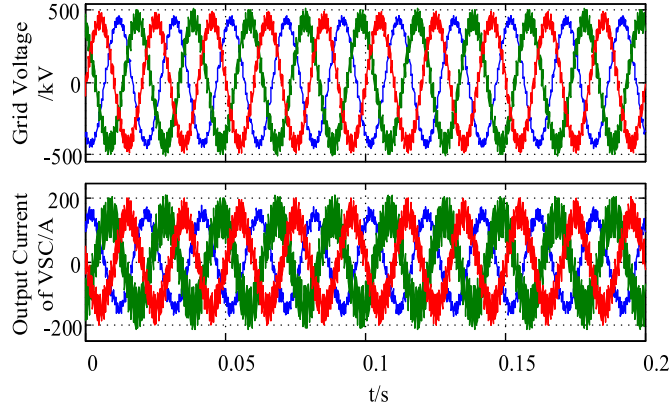


Figure 1.3: 1270 Hz resonance at the Luxi back-to-back VSC-HVDC system [23].

Another oscillation phenomenon was observed in a weak part of the China Southern Grid's transmission system: In specific grid situations, after a Static Synchronous Compensator (STATCOM) was put into operation, voltage and current measurements contained 2.5 Hz and 97.5 Hz components, as it can be seen in Figure 1.4 [25]. It was determined that interactions between STATCOMs and a weak grid caused the interactions with a low grid strength, negatively impacting the severity of the oscillations. At

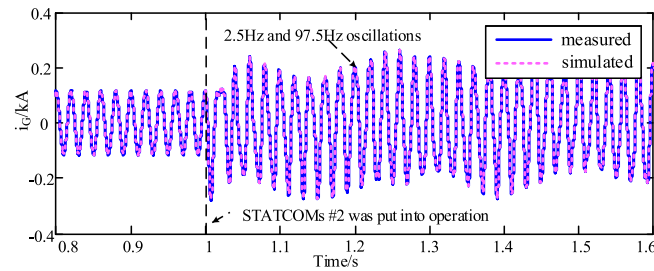


Figure 1.4: 2.5 Hz and 97.5 Hz oscillations in a transmission system of China Southern Grid [25].

higher grid strengths, unstable situations could be avoided completely. However, series compensation and the loading level of the HVDC system only had a small effect on the oscillation magnitude and frequency [25].

In summary, the reported incidents have the following characteristics in common:

- The incidents are connected to power system converters and their control systems.
- They are triggered by small disturbances changing the grid impedance, such as switching connected lines or operation of particular components.
- The currents and voltages show excessive distortions with oscillations magnitudes exceeding normal harmonic emission levels.
- The oscillations have unusual, non-characteristic frequencies, not being an integer multiple of the fundamental or switching frequencies.

1.3 Representation of Modular Multilevel Converters in Grid Studies

Stability studies dealing with converter-dominated systems are challenging because converter manufacturers do not provide insight into controllers and other converter parts due to Intellectual Property Right (IPR) concerns. Although predominantly influencing the system behavior, the converter itself is a black box to TSOs.

A way to overcome the challenge of unknown system components, particularly controllers, is to use black-boxed replica systems of the converter controls and couple them with Real-Time Simulators (RTSs) that model the system's electrical structure. Figure 1.5 depicts an MMC, divided into the control system and the electrical structure. While not mimicking the whole converter station, replica systems are an identical copy of the control and protection system installed on-site at an HVDC station. They can be supplied by manufacturers as control cubicles [26]. By simulating the electrical structure of the MMC and the connected AC system in the time domain on RTS, manufacturer controllers can be used in stability studies without on-site testing [21]. Converter replica systems are provided as control cubicles that consist of the converter manufacturer control system.

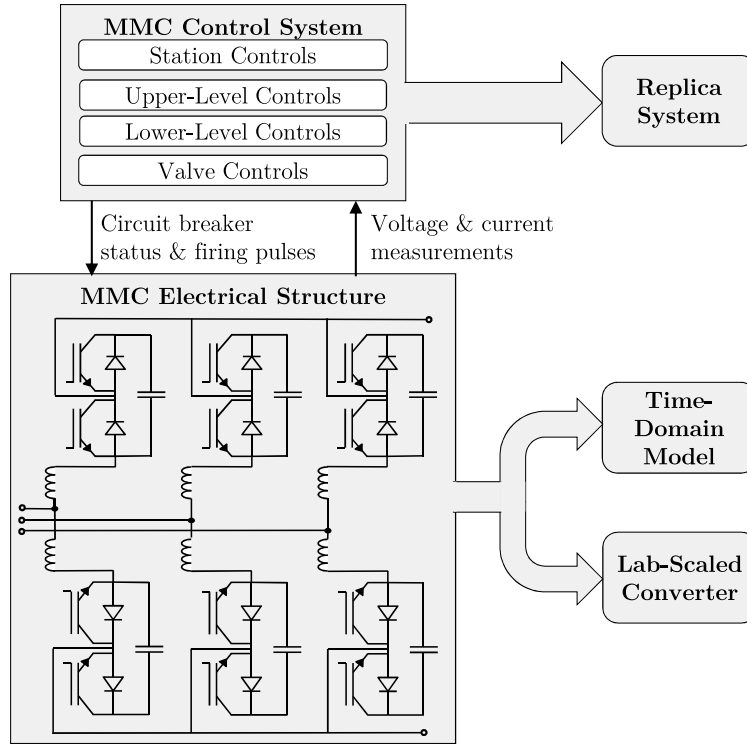


Figure 1.5: MMC electrical structure and control system.

As a result, replica systems enable an analysis of a converter's actual control system and its impact on system stability. However, replica systems are expensive and require extensive space because they can have sizes as large as several server racks. Moreover, while using real controllers, the physical parts, such as the converter's electrical structure, remain a model, and model assumptions and simplifications come with deviations from the real converter behavior. For instance, typically-used average value models (type 5 and 6 Electromagnetic Transient (EMT) models) use controlled voltage and current sources to represent the electrical structure of an MMC [27]. While also representing the harmonic content and switching functions that influence the AC waveforms [27], the converter arms, comprising the Submodules (SMs) that contain Insulated-Gate Bipolar Transistors (IGBTs) and capacitances, are modeled solely as a voltage source. More detailed models are intended for AC or SM fault studies and are not applicable for grid studies because they can reach a relative computing time that is 500 times larger than that of average value models [27].

Using lab-scaled converters for representing the electrical structure of an MMC makes it possible to avoid potential deviations due to model assumptions and simplifications. While consisting of significantly fewer SMs than full-scale MMCs, the electrical structure of a lab-scaled MMC matches that of a full-scale MMC, so that potential resonances can be identified. However, when used in system studies, interfaces or scaling approaches need to be considered to couple lab-scaled converters with full-scale system components.

1.4 Literature Review

In academia, research publications and Ph.d. theses address the aforementioned incidents by developing converter models and applying suitable investigation methods.

Particularly, frequency-domain analyses using frequency-dependent impedance models have been used as a measure to analyze converter control interactions. Promoting the impedance-based approach [28], VSC models have been developed that describe the frequency behavior of the converter by analytically derived transfer functions [29]. The model was later generalized to also represent the characteristics of MMCs [30]. Similar models have been also published by other authors [31]. While these models have been developed in the positive and negative sequence, additional models in the dq-domain have been proposed for VSCs [32, 33] and also MMCs [34]. Investigating the difference in modeling converters in positive and negative sequence or dq-domain, research has also focused on developing appropriate transformation methods, and principles between the domains [35]. Due to the complexity, other models have been developed in the phase domain only [36, 37] with the latter using a laboratory-scaled MMC to validate the developed analytical model. So far, the models are derived on the basis of control systems that are also developed by the same authors and intended to be used for frequency-domain investigations. Thus, models are simplified by neglecting control loops that are assumed to be too slow (e.g. outer control system) or too fast (inner control system) for causing interactions.

Investigations focus mainly on offshore systems such as HVDC connected offshore wind farms [32, 38] or AC connected wind farms and specifics of cable models [39]. While some authors look into the stability of power systems dominated by a high share of RES [40] or consisting of multiple VSC HVDC converters [36], investigations are limited to the converter models developed by the authors. The results indicate that specific converter controllers, such as the PLL, are responsible for instabilities [29], particularly when converters are connected to weak grids [41].

1.5 Research Gap and Research Questions

The literature review shows that several theses investigated the observed interaction phenomenon [29, 30, 36, 39]. Many analytical models of converters and their control systems have been developed to investigate the stability of converter-dominated grids in the frequency domain [30, 31, 34, 37]. While these models can investigate the interaction phenomenon and examine causes and mitigation methods, the models require detailed knowledge of the grid components, particularly the converter control system. As a result, the developed methods and models cannot assess complex and manufacturer-based converters' interactions. Therefore, an extensive investigation of converters' interactions

considering various control modes and operating points is missing in the contemporaneous literature. Moreover, the interdependency of outer control modes and different AC grid situations, such as weak or strong grid conditions, are not considered.

Additionally, grid code specifications are not considered, preventing the use of developed models in a commercial context [42]. Also, stability investigations are performed under the assumption that both converter and grid models are developed by the engineer, investigating the stability [39, 43, 44]. However, it is not realistic that, for instance, a transmission system operator shares the grid model with a third party or a converter manufacturer publishes a detailed description of the converter control system. Instead, TSOs use replica systems that, while mimicking the real converter control behavior, are a black-box to the operator. However, published methods and approaches do not address how stability analyses can be conducted based on unknown system components such as replica systems. Moreover, they do not address that system components from different parties need to be coupled and the implications for the analysis method's requirements.

In addition to the control system, the converter's electrical structure and components also impact the frequency behavior. While replica systems allow for an exact representation of the control system, they need to be applied to a converter model that simulates the converter's electrical structure and components. However, only simplified electrical converter models are often applied, and the impact of model assumptions and simplifications is not investigated.

Consequently, analysis tools are required to assess the stability of converter-dominated grids and the observed interaction phenomenon by methods capable of dealing with models provided by different parties along with black-boxed (time-domain) models of system components, such as replica systems. In addition, the impact of model simplifications in the electrical structure of the model needs to be addressed.

Thus, this thesis develops methods that can be applied independent of the converter's electrical representation and the control system, employing preexisting, realistic control system models. It addresses the question of how converter control interactions can be assessed and what approaches allow for the integration of system components and models provided by different entities. Moreover, the thesis investigates in what grid situations interactions might cause unstable systems. Particularly, it investigates the short-circuit power because a TSO is required to provide a network equivalent with minimum and maximum short-circuit power at the connection point so that the HVDC system owner can ensure a stable operation of the system [45].

Thus, this thesis addresses the following research questions:

1. With what methods can a TSO investigate potential converter control interactions that threaten the system stability when manufacturers provide converter controllers only as black-boxed systems, such as control replicas?

2. Are simplified models of the electrical structure of a converter adequate to investigate converter control interactions?
3. In what grid situations do converter control interactions threaten the system stability of the transmission grid?

1.6 Objective and Structure

This thesis aims to identify adequate approaches and develop methods to investigate converter control interactions when the converter control system is unknown and, therefore, a black box to the system operator. Assuming that converter control systems are only provided as black-boxed replica systems, methods need to be able to be applied on RTSs that couple replica systems with a simulated model of the electrical structure of a converter. By addressing the aforementioned research questions, suitable methods are developed to assess converter control interactions for different grid scenarios. In addition, the applicability of simulated converter electrical models to assess interaction is investigated. Figure 1.6 illustrates the approach and procedure to achieve these objectives and answer the research questions of this thesis.

First, the observed interaction incidents are analyzed to classify the phenomenon and determine suitable methods that can be used to assess the interactions. Based on this, the methods are reviewed, and suitable approaches are discussed to simulate and assess the interaction phenomenon, with respect to the underlying assumption of this thesis that TSOs are only given access to black-box control models by converter manufacturers. The conclusions from this analysis will then be used to narrow down and define requirements for the subsequently developed methods. By investigating the frequency behavior of a simulated electrical MMC model and physical laboratory-scaled MMC, both having identical control systems, differences and potential model limitations are identified. Finally, the modeled methods are used to evaluate defined test cases to determine what grid situations threaten a transmission grid's system stability due to converter control interactions.

This thesis will contribute to the planning stage of HVDC systems to be integrated into the transmission system by providing a tool that can predict instabilities in non-fault situations of systems connected to potential HVDC converters. Therefore, TSOs can analyze if HVDC converters and their control systems will threaten the system stability in certain grid situations. This will reduce the risk of unexpected failures and project delays in the commissioning phase of HVDC systems.

To demonstrate the applicability, the method will be used on a test bench with physical lab-scaled converters as well as simulated converter models. The knowledge gained will help to understand how HVDC converters, in particular, MMCs interact with the system in a way that leads to the observed oscillations. Furthermore, such a method can be used

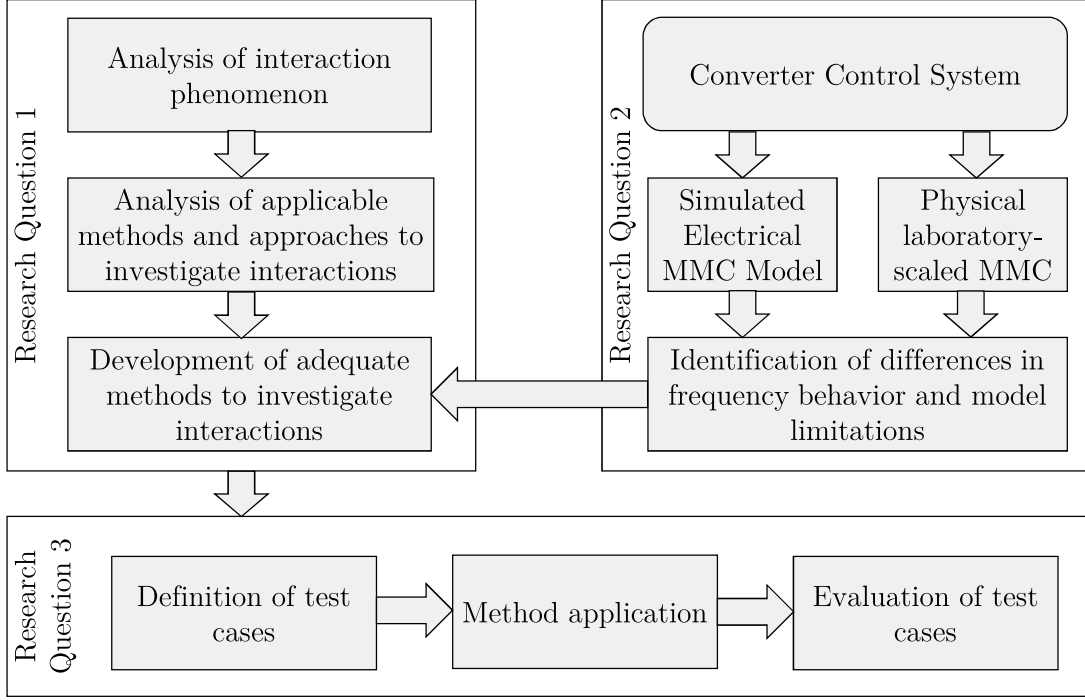


Figure 1.6: Research questions and thesis approach.

to verify manufacturer black-box models by comparing the models' frequency behavior with that of converter control replica to increase the confidence about the validity of the provided models.

The thesis is structured into six chapters. Having introduced and motivated the topic and objective in this chapter, the second chapter analyzes the research problem and questions and investigation methods. Initially, the interaction phenomenon is classified in terms of power system dynamics, stability, and quality. On this basis, methods and approaches are discussed to identify those to be appropriate for investigating converter control interactions concerning the thesis' objective. The chapter concludes with a detailed approach to how the research questions are investigated using the methods and approaches that are previously concluded from the analysis. Chapter 3 introduces the underlying models and test benches used to derive the impedance models used for identifying critical grid situations. Afterward, chapter 4 presents a numerical impedance derivation method that is developed in this thesis and is capable of deriving impedance models for black-box systems, such as converter control replicas, while minimizing data-handling requirements. The chapter ends with the derived impedance models of the MMC and AC grid. Chapter 5 addresses the second and third research question and presents at first the results of the comparison of real laboratory converters with simplified EMT converter models. Afterward, the impedance-based stability assessment approach is validated by comparing the frequency-domain results with those obtained from EMT time-domain simulations. Lastly, the impedance-based approach is used to identify the stability of several test cases. Chapter 6 then summarizes the thesis and presents an outlook of potential further work based on this thesis.

2 Analysis

This chapter breaks down the interaction phenomenon and analyzes the task definition of this thesis. Moreover, power system phenomena and classifications are discussed in the context of power system dynamics, stability, and quality. On this basis, adequate methods are reviewed to assess converter interaction incidents. Finally, the approach to accomplish the objective of this thesis will be defined.

2.1 Classification of Interaction Phenomenon

The electrical power system can be described as a complex dynamic system of interacting components and devices with the common goal to transform primary energy sources first into electricity, which can be transformed again and used by loads and consumers. Employing different voltage levels and AC technology, electricity is traditionally transmitted and distributed from the generation to the loads using transformers, overhead lines, and cables [46]. As described, modern power systems also rely on HVDC systems for transmitting electricity due to lower losses and the inherent capability to control the active and reactive power flow [47]. However, HVDC systems impose new risks to system stability, as seen in the aforementioned interaction incidents. This section gives an overview of power system dynamics and stability and power quality to classify the aforementioned interaction phenomenon.

2.1.1 Power System Dynamics

According to their time frame and physical characteristics, power system dynamics can be classified into four main categories: wave, electromagnetic, electromechanical, and thermodynamic phenomena [48]. Whereas wave phenomena, e.g., traveling waves in overhead lines caused by lightning and switching, occur in a time scale spanning from microseconds to milliseconds, electromagnetic phenomena are much slower. They can range from milliseconds up to a second. Electromagnetic phenomena are typically related to the interaction of electrical machines and the grid, as well as to the operation of protection systems [48]. Electromechanical phenomena correspond to the response of synchronous and asynchronous machines when subjected to disturbances and the effect on the speed of the rotor and the primary control system. Electromechanical phenomena can continue for several seconds. Lastly, dynamics due to thermodynamic phenomena caused by the boiler and combustion controls can last a few hours. Figure 2.1 shows classified dynamics with respect to time scale. The time frame of the converter interaction phenomenon is indicated in gray. According to the time scale, the reported interaction

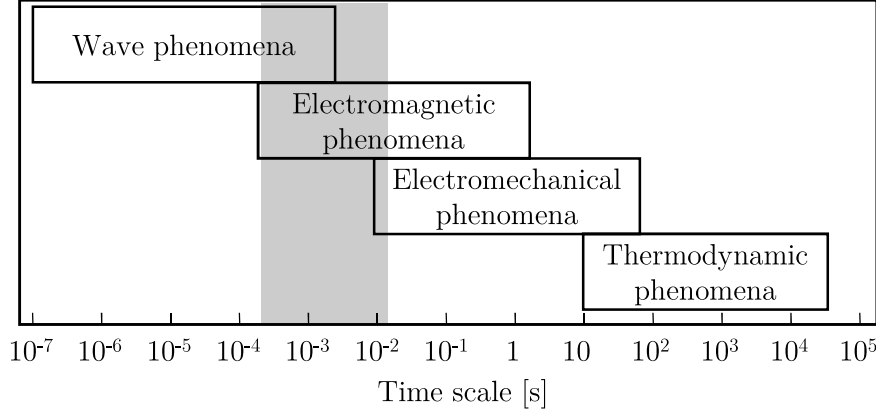


Figure 2.1: Power system dynamic phenomena classification [48].

incidents can be understood as electromagnetic phenomena; they can be attributed to exchange from energy in electric and magnetic fields.

In power systems, electromagnetic phenomena are traditionally understood as interactions within synchronous generators following disturbances such as short circuits with dynamics having a time scale of milliseconds [48, 49]. According to the law of constant flux linkages and energy conservation, the energy within a magnetic field cannot change instantaneously because it is proportional to the magnetic flux. When a short circuit occurs, this implies that the flux is the same both directly before and immediately after the short circuit. Thus, for example, in a DC source linked to a coil, the magnetic energy either induces a constant current when it is purely inductive or when considering the coil's resistance exponentially decaying current, dissipating the magnetic energy. Similarly, a generator responds with an alternating short circuit, having a DC offset which is decaying according to its time constant [48]. Because the reported incidents show sustained and not decaying oscillations, it can be concluded that the incidents result from an energy exchange caused by interactions between the network impedance and an equal negative resistance [36].

2.1.2 Power System Stability

The effects and severity of instabilities do not necessarily depend on the classification. Nevertheless, classifying stability into distinguishable categories helps the engineer studying and analyzing power system stability. Narrowing down stability phenomena reduces complexity and facilitates the identification of factors of influence as well as the development of methods mitigating stability problems [49].

The IEEE/CIGRE Joint Task Force on Stability Terms and Definitions has defined power system stability as follows [50]:

"Power system stability is the ability of an electric power system, for a given initial operating condition, to regain a state of operating equilibrium after

being subjected to a physical disturbance, with most system variables bounded so that practically the entire system remains intact."[50]

Accordingly, stability depends on the initial operating condition and the disturbance to which it is subjected. Stability is assessed according to the subsequent state of the system that is required to achieve equilibrium. Disturbances are categorized as small disturbances, e.g., load changes, or large disturbances, e.g., short circuits. As shown in Figure 2.2 System stability is categorized into the main categories, considering the physical nature of instability, the extent of the disturbance, and the time span required to assess the instability: rotor angle, voltage, and frequency stability. These categories are further divided into small and large disturbances, and short and long-term phenomena [50].

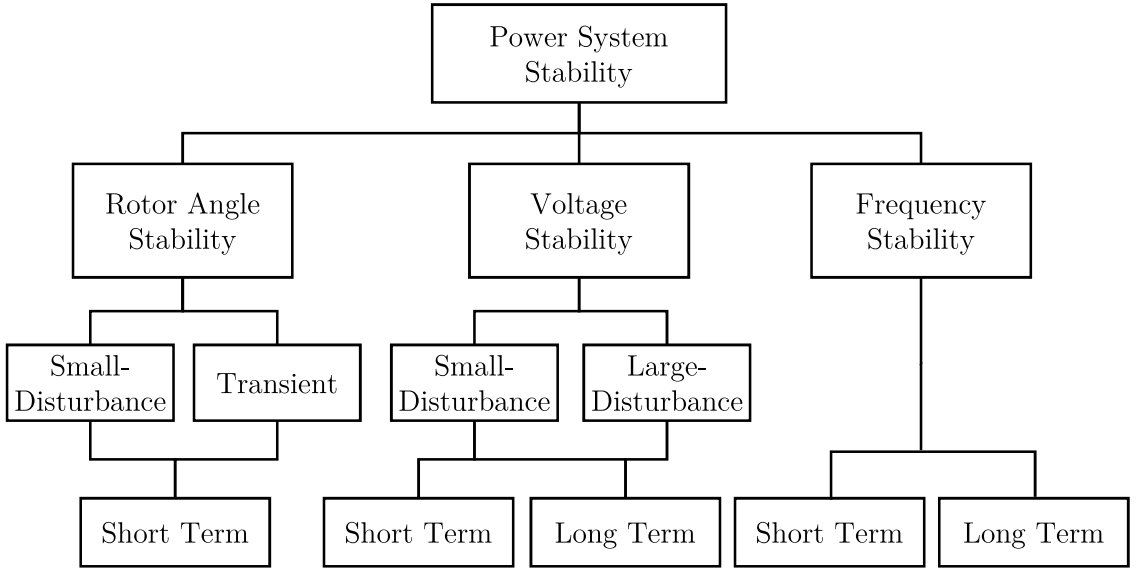


Figure 2.2: Power system stability classification [50].

Rotor angle stability is an electromechanical phenomenon. It refers to the power system's synchronous machines' ability to remain in synchronism and operate at the same speed when subjected to disturbances [49]. For small disturbances, the synchronous machine is required to return to its original state to maintain stability. With large disturbances, the synchronous machine can return to a new equilibrium point, and stable system state [51]. Following a large disturbance, synchronous machines may decelerate or accelerate and change their rotor angle to account for changes in active power infeed. If the rotor angle deviates too much compared to the interconnected synchronous machines, then the synchronous machine slips, loses synchronism. After a small disturbance, synchronous machines can excite other closely-coupled or weakly-connected machines, resulting in so-called inter-area-oscillations and machines swinging against each other. However, both small-disturbance and transient stability are short-term phenomena, short-term being defined as a range of seconds [49].

Frequency stability refers to the power system's ability to maintain a steady grid fre-

quency following major disturbances resulting in a significant loss of power generation or load demand. In the instance of such power imbalances, the power system is required to reach a new equilibrium point without tripping more generation units or loads due to frequency oscillations. Furthermore, frequency stability is of concern when power reserves are insufficient, resulting in unintentional load shedding or loss of additional power generation units [50]. Short-term frequency instabilities refer to power outage cascades and blackouts when the frequency decreases rapidly due to insufficient under-frequency load shedding [52]. Whereas short-term frequency instabilities occur within seconds, long-term instabilities can last for several minutes and are often caused by poorly functioning primary controls of the generation units [53].

Voltage stability is defined as the power system's ability to maintain steady and acceptable voltages at all buses following a disturbance, causing the system to deviate from its initial operation point [50]. Voltage instabilities are recognized by a sudden voltage drop and result from the attempt to operate the power system beyond the power transfer limits of transmission lines [54]. Whereas both frequency and voltage stability depend on the power system's ability to sustain or return to an equilibrium between load demand and generation, the power system's structure, particularly the electrical distance, is an essential factor of influence in the case of voltage instabilities. With increasing inductance between load and generation, the reactive power demand increases and makes the power system vulnerable to voltage instabilities. Hence, a sudden increase in load or a large disturbance that increases the inductance and decreases the voltage threatens the voltage stability. If the active power flow exceeds the maximum capabilities, the voltage progressively drops, leading to a voltage collapse [55]. Whereas small-disturbance voltage stability refers to the power system's ability to maintain stable voltages following perturbations, such as gradual load increases or decreases, large-disturbance instability is caused by loss of generation or system faults. Short-term and long-term phenomena further distinguish voltage stability: short-term stability refers to relatively fast components involving dynamics in the range of several seconds. Long-term stability requires studying the system for several minutes due to slower components, such as tap-changing transformers [50].

Rotor angle stability cannot be considered for explaining interactions between converters and the AC grid because it is an electromechanical phenomenon related to interactions of synchronous machines and the power system. Although the reported interactions show a non-steady grid frequency, the grid frequency itself does not significantly drop or increase. In fact, the superimposed signals at frequencies other than the grid frequency, as seen in the measured signals' oscillations, cause the problems. Furthermore, because the interactions are triggered by events considered normal grid operations, power imbalances and large disturbances are not causing the incidents. Similarly, while manifesting itself in voltage distortions, no large voltage drop of the Root Mean Square (RMS) voltage is seen in the measurements, and the converter maintains its initial operating point. In addition, the definition of short-term instabilities refers to interactions within a time

frame of seconds, being slower than the reported interactions showing frequencies as high as 1.2 kHz.

Thus, the reported converter interaction incidents do not coincide with the classic stability categories rotor angle, voltage, and frequency stability. Although the interactions do not always lead to diverging system states and the system remains at an operating condition around an equilibrium, the reported incidents constitute a stability problem. The interactions do not always lead to an unstable system. However, a damped and marginally stable system often results from the converter interactions. Recognizing that the existing stability definition does not account for phenomena caused by the increasing share of power electronic components, the IEEE Power System Stability Subcommittee has recently extended the stability classification. Therefore, the terms *resonance stability* and *converter-driven stability* are added to consider the "dynamic behavior in systems with high penetration of power electronic interfaced technologies" [56]. Figure 2.3 depicts the new stability categories and their subcategories.

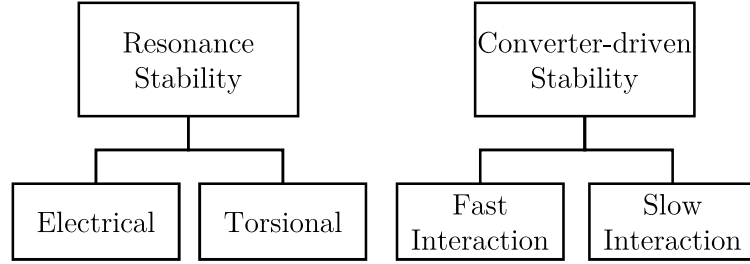


Figure 2.3: Additional stability categories due to the increase of power electronic interfaced technologies [56].

Resonance stability describes sub-synchronous resonances that lead to sustained oscillations at a frequency below the nominal grid frequency [57]. Resonances between series compensation and the generator's turbine-shaft, which excite mechanical modes and cause torsional oscillations, are classified as *torsional*. These oscillations can compromise the generator shaft's mechanical stability if they are insufficiently or negatively damped [58]. They are mainly caused by components with fast-acting control systems such as HVDC systems, static VAR compensators, STATCOMs, and power system stabilizers that interact with the turbine shaft modes [56, 58]. *Electrical resonance* refers to interactions of series compensation with the electrical characteristics of a generator [56]. Whereas these interactions were not observed for synchronous generators, Double-Fed Induction Generators (DFIGs) of wind turbines are known to be susceptible to these kinds of resonances [56]. Effectively acting as an induction generator that is directly connected to the AC grid, a series capacitor can resonate with the effective inductance of the DFIG at sub-synchronous frequencies due to an overall negative resistance of the resonant circuit [59].

Converter-driven stability refers to interaction phenomena of converter-connected systems, in particular VSCs, with the grid that can occur at a wide range of frequencies due

to the wide-ranging time-scales of the employed controllers. To distinguish different phenomena, it is further divided into *fast interaction*, referring to oscillations at tens to hundreds and up to several kHz, and *slow interaction*, with frequencies that are usually below 10 Hz [56]. *Slow-interaction converter-driven stability* describes power-electronic systems that interact with slow power system components such as electromechanical dynamics of synchronous machines and their controllers [56]. Like voltage instability, it can be caused by weak grids. However, not loads but the converter control system, particularly the PLL, are responsible for the instability [60]. Slow-interaction converter-driven stability is further classified into low-frequency oscillations. It refers to interactions of the outer control loops, PLL that are influenced by the grid strength at the converter connection point and weak system stability. The latter describes the failure of the PLL in weak systems to track the grid angle in case of close faults [56]. *Fast-interaction converter-driven stability* refers to an instability phenomenon that results in high-frequency oscillations with up to several kilohertz [61]. However, both sub- and super-synchronous interactions of STATCOMs and weak grids, having oscillations at 2.5 Hz and 97.5 Hz, are classified as fast interactions [56, 62]. High-frequency switching and LCL filters, and interacting converter controllers are considered causes for fast-interaction converter-driven stability [56].

Although both are related to interactions of components with fast controllers, the observed incidents do not meet the criteria of resonance stability because the oscillations happen at frequencies above the grid frequency and are not related to synchronous generators or induction machines. Contrarily, converter-driven stability refers to interactions of VSC control systems with the grid. Thus, the observed interactions can be assigned to this category. Ranging from 2.5 Hz to more than 1 kHz, the interactions can be considered both slow and fast according to the definition. However, slow interactions refer to interactions between the converter and power system components such as synchronous generators and their controllers that have a slow response [56]. Therefore, the observed interaction incidents are best classified as fast-interaction converter-driven stability according to the extended stability classification.

2.1.3 Power Quality

Power quality in traditional power systems is understood as an issue distinct from power system dynamics and stability. Although referred to as power quality, mostly the voltage is being studied for assessing the quality. Operating at a sinusoidal voltage of typically 50 Hz or 60 Hz, any deviations in terms of waveform magnitude or frequency are treated as a power quality problem [63]. Recognized as consumer-driven issues, power quality problems are linked to the failure or maloperation of customer equipment [63, 64]. Because power quality problems address a wide range of different phenomena, Table 2.1 shows the categorization of the phenomena used to evaluate power quality problems[63].

Table 2.1: Categories of power quality phenomena according to [63].

Category	Description
Voltage Imbalance	Relative maximum deviation of the voltages or currents from the average three-phase voltages or currents.
Voltage Sags/ Interruptions	Short-duration voltage variations that are instantaneous, momentary or temporary.
Voltage Fluctuations	Cyclical variations of the voltage envelope or a series of random voltage changes which does not exceed specified limits.
Transients	Momentary event that is described by the transition of a variable from one steady-state operating condition to another.
Harmonic Distortion	Steady-state deviations of the harmonic spectrum of the voltages and currents with respect to the fundamental frequency.

As indicated in gray, the converter interaction phenomenon can be attributed to harmonic distortions according to the description in Table 2.1, because the frequency spectrum of the measured signals contains frequencies other than the fundamental frequency. However, harmonic emissions of power system equipment usually are considered the causes of harmonic distortions [65]. In fact, harmonic distortion and power system harmonics have been studied since the beginning of the AC power system when mostly third harmonics, due to iron saturation in transformers and machines, were of concern. In today's power system, power electronic devices such as nonlinear loads are primarily responsible for harmonic distortions [66]. Traditionally, in power systems, these harmonics are known for potentially causing high distortions when close to the resonance frequency of the grid and being amplified as a result [67, 68]. That said, these amplifications of pre-existing harmonics cannot cause the reported incidents because non-characteristic harmonics were observed, and the distortions reached a severity that cannot be explained even under worst-case assumptions of converters emitting harmonic currents [18, 19, 22, 69].

2.1.4 Terminology

Because the stability classification has only recently¹ considered the impact of power electronic coupled technologies, the term converter-driven stability does not commonly appear in the literature. Moreover, different terms are used when referring to the stability problems caused by converter interactions. Wang established the term *harmonic stability* [70], also used by Buchhagen, along with *harmonic resonance* to describe the stability

¹The extended stability classification was published in April 2020 by the IEEE PES Power System Dynamic Performance Committee Task Force on Stability definitions and characterization of dynamic behavior in systems with high penetration of power electronic interfaced technologies.

incident experienced at TenneT’s offshore wind farms [18]. This term dates back to 1967, but was originally used to describe stability problems associated with harmonics emitted by converters [71]. The term has also been used and translated to *Oberschwingungsstabilität* in the German standard VDE-AR-N 4131 [42]. Sun refers to a *harmonic and resonance problem* while stating the inadequacy of the term as not only harmonics but also frequencies of any value need to be considered [72]. Later, Sun suggested the term *harmonics created by resonance*, with the emphasis that the harmonic frequencies are not restricted by integer multiples of a fundamental frequency and can be above and below the fundamental frequency [73]. This definition also stresses that preexisting harmonics do not cause observed oscillations – instead, the harmonics are a consequence of resonances. Whereas other authors simply refer to *stability of interconnected systems* [37] or *small-signal stability* [44], Bayo Salas proposes the definition of *electromagnetic instability*. Referring to the resonance nature of the interactions, the definition emphasizes that natural frequencies are excited by components adversely controlling the energy at specific frequencies [36].

2.1.5 Conclusion

As pointed out in the previous sections, the reported converter interaction incidents can be understood as electromagnetic phenomena. Although the distortions caused by the interactions suggest a power quality problem, the severity and frequency spectrum of the measured voltage and currents signals indicate a stability problem. With the extension of the power system stability classification diagram, the increasing share of power electronic coupled technologies impacting the transmission system’s dynamic behavior has been recognized. Regulating internal currents and voltages of the power electronics and controlling the power flow of the AC grid, the time scale of converter dynamics is significantly wider than traditional power system components such as synchronous machines [74]. As a result, interactions of converters with the AC grid can cause oscillations at a wide range of frequencies [70].

Because no consistent definition has been established, in this thesis, the reported incidents will be referred to as converter control interactions, emphasizing the systems responsible for the stability problem. However, converter control interactions per se do not inevitably comprise the system stability. For example, converter control systems can purposefully interact with the AC grid, and damp inter-area oscillations [75, 76] or provide voltage support in case of contingencies [14]. Nevertheless, this thesis refers to converter control interactions as interactions causing a deterioration of the system stability.

2.2 Discussion of Methods and Approaches

Several existing methods and approaches can be used to investigate converter control interactions. Consequently, in the next sections, they are analyzed with respect to the

thesis's research questions to derive the appropriate investigation approach for the thesis.

2.2.1 Time-Domain Simulation

Time-domain simulation tools have been widely used to assess the stability of power systems [49]. Because power systems were influenced mainly by the dynamics of synchronous machines and the respective controllers, transient stability tools have been used traditionally to study dynamics in the time-scale of milliseconds to seconds. Transient stability tools, often referred to as RMS or phasor simulation tools, rely on the assumption that power system components can be modeled as a set of nonlinear differential-algebraic equations and that the investigated system is balanced, operating at fundamental frequency [46]. The power system is simplified to a positive-sequence equivalent circuit diagram. Sinusoidal quantities such as voltages and currents are represented by phasors, rotating at the fundamental frequency and becoming DC quantities in a steady state. Thus, efficient variable time step solvers can be employed, enabling the simulation of large-scale power systems with several thousand buses [77]. However, using phasors at fundamental frequency to represent system dynamics is contrary to the research question to investigate dynamics at frequencies other than the fundamental. Therefore, transient stability tools are not suitable for investigating converter control interactions.

Time-domain simulations targeted at simulating EMT phenomena are likewise referred to as EMT-type tools. EMT tools are designed to precisely simulate the time-domain dynamics of power systems at very high precision, without inherent assumptions limiting the scope of the investigations. The actual waveforms of system states are reproduced while including the representation of all wires and relevant system components as three-phase quantities if required [78].

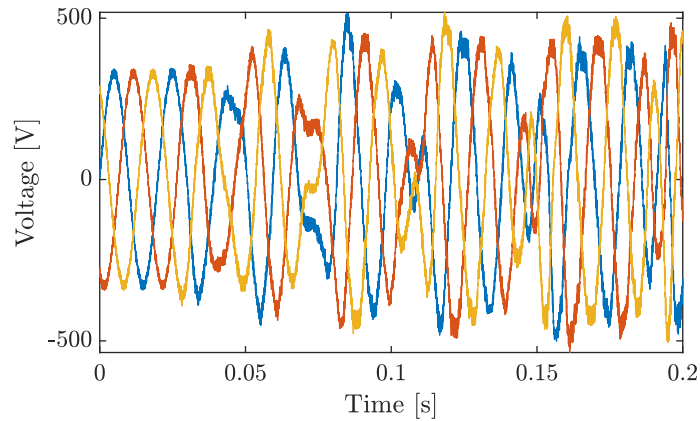


Figure 2.4: EMT simulation showing unstable system.

EMT simulation tools determine the waveforms of the system states by solving the nodal admittance matrix of the systems at each time step Δt . Using the trapezoidal rule, the

system is converted into a set of current sources and conductance matrices forming the nodal admittance matrix [79]. As a result, the accuracy of results obtained with EMT-simulations are only limited by the degree of the depth of the models, the parameter data available and the computing power. Thus, they are often used as a benchmark for verifying developed methods. Whereas EMT simulations can verify the existence and quantify the magnitude of experienced or predicted stability problems, as shown in Figure 2.4, time-domain simulations are ineffective for identifying new, unknown problems. In fact, using EMT-simulations for stability studies comes with the following drawbacks:

- The size of the system is limited due to computational restrictions.
- Precise models require data often not available or not provided due to intellectual property concerns.
- The duration of simulations can be excessive due to the small simulation time steps required by fast controllers.
- EMT simulations require studying the waveforms first following the simulations for any conclusions. They do not provide direct indications of causes for instabilities, making it challenging to automate analyses and investigate numerous cases.

EMT simulations can be executed offline or in real-time. Whereas the duration of a simulation depends on the computational time of solving each time step, real-time simulation tools ensure that each time step equals the actual time. This means that, for instance, simulating a time series of 10 s also lasts 10 s. This enables the coupling of physical components with real-time simulators when appropriate interfaces are available. It should be emphasized that when not coupled with physical components, there is no inherent advantage of simulating models in real-time.

2.2.2 Laboratory Test Bench

Laboratory test benches employ real, physical converters scaled-down to a power and voltage level, enabling use in laboratory setting. As such, laboratory test benches capture the physical behavior of a converter that might not be seen in models due to simplification or underlying assumptions. For instance, switching operations, potential voltage imbalances in the capacitances of MMC SMs and non-ideal characteristics such as parasitic elements of circuit components or asymmetries, are considered inherently when using laboratory converters [37].

Laboratory converters typically have a power-rating of 5 kW to 10 kW and facilitate a DC voltage of 100 V to 400 V [30, 80, 81]. Although down-scaled, MMC test benches employ several SMs consisting of semi-conductor devices that can be switched by applied control systems so that different controllers, including encrypted manufacturer controls, can be tested.

When used for interaction studies, laboratory converters need to be coupled with the potentially interacting systems, e.g. the AC grid. For simple grid representations, this can be done by connecting a passive component with an impedance calculated according to a certain grid strength. However, this can become a complex task when larger grids, including the topology, are investigated because each investigated grid needs to be physically constructed in a laboratory. Alternatively, laboratory converters can be coupled with RTSs that simulate the AC grid and therefore also enable the study of the impact of large-scale grids. Coupling physical components with a simulated AC grid not only requires that the RTS provides simulated voltage and current signals to the converter. Also, the converter needs to feed back measured voltage and current signals to the RTS. This means that a Power-Hardware-in-the-Loop (PHIL) system is created that requires additional interface algorithms for the signal exchange [82]. However, interface methods such as the ideal transformer or the impedance damping method can easily cause unstable systems due to the time delay when exchanging signals between the RTSs and the connected components [83]. This can lead to false conclusions regarding the identification of unstable cases. Thus, the following conclusions can be made for using laboratory test benches to investigate converter control interactions:

- Laboratory test benches facilitate a more realistic representation of a converter's physical structure due to the use of real physical components.
- Using a laboratory environment instead of relying on simulation only makes it possible to increase the Technology Readiness Level (TRL) of investigated equipment or control systems.
- The converters need to be coupled with RTSs through interface methods for interaction studies, which can also be a cause of instability.
- Laboratory test benches only represent a down-scaled system, so that appropriate scaling methods need to be investigated.

2.2.3 Eigenvalue Analysis

Eigenvalue analyses conclude the state-space matrix of Linear Time Invariant (LTI) dynamic systems whose behavior can be described in the time domain by a set of Differential-Algebraic Equations (DAE). Traditionally used to analyze electromechanical oscillations in power systems, system modes can be related to state variables by computing the system states' participation factors. The information can be used to identify the causes of instability and analyzing the sensitivity of control system parameters [46]. However, taking into account electromagnetic dynamics, the great extent of the time scales of converter controllers significantly increases the system order [84].

Following a small-signal linearization of the system around an equilibrium point, the linearized state vector

$$\Delta \dot{\mathbf{x}} = \mathbf{A} \Delta \mathbf{x} + \mathbf{B} \Delta \mathbf{u} \quad (2.1)$$

and the linearized output variable vector

$$\Delta \mathbf{y} = \mathbf{C}\Delta \mathbf{x} + \mathbf{D}\Delta u, \quad (2.2)$$

where $\mathbf{A}, \mathbf{B}, \mathbf{C}, \mathbf{D}$ are the state, control, output and feed-forward matrices, representing the state-space representation of a power system [50]. The eigenvalues are derived from the characteristics equation of the LTI system, given by equation (2.3) with

$$\det(s\mathbf{I} - \mathbf{A}) = 0. \quad (2.3)$$

However, the state-space representation of large-scale power systems, including HVDC converters, requires modeling a high-order system state matrix. In order to ease the computational burden, the Component Connection Method (CCM) is proposed for deriving the state-space LTI of converter-based power systems [85]. The CCM decomposes each component in the power system and linearizes the components locally. The components are interconnected by linear algebraic equations [70] defining interfaces. The resulting state-space component model is given by

$$\begin{aligned} \Delta \dot{\mathbf{x}} &= \mathbf{F}\Delta \mathbf{x} + \mathbf{H}\Delta \mathbf{a} \\ \Delta \mathbf{b} &= \mathbf{J}\Delta \mathbf{x} + \mathbf{K}\Delta \mathbf{a} \end{aligned} \quad (2.4)$$

where the state and system matrices $\mathbf{F}, \mathbf{H}, \mathbf{J}, \mathbf{K}$ are diagonal matrices and $\mathbf{a}$ and $\mathbf{b}$ are the vectors of input and output variables of the components. The interconnections of the components are given by

$$\begin{aligned} \mathbf{a} &= \mathbf{L}_{11}\mathbf{b} + \mathbf{L}_{12}\mathbf{u} \\ \mathbf{y} &= \mathbf{L}_{21}\mathbf{b} + \mathbf{L}_{22}\mathbf{u} \end{aligned} \quad (2.5)$$

where $\mathbf{u}$ and $\mathbf{y}$ are the input and output vectors of the system. The matrices L_{ij} are linear algebraic, and depend on input and output $\mathbf{a}, \mathbf{b}$ of the components.

Computing the eigenvalues of the state-space model permits investigation of the stability of the power system. The eigenvalues in a state-space representation are synonymous with the roots of a characteristic equation of the system transfer function. Thus, the system's small-signal stability can be determined by the computed eigenvalues' real parts, where negative parts denote unstable and positive parts denote a stable system. Figure 2.5 shows the eigenvalues of the state-space representation of an exemplary system. As can be seen, a pair of eigenvalues exist in the right-half plane, indicating a small-signal unstable system. However, conclusions drawn from eigenvalue analyses are only as reliable as the models used for state-space modeling. In addition, every component of the power system needs to be represented and every state considered in state-space models. As a result, eigenvalue analyses mostly rely on time-domain RMS models developed for transient stability simulations that neglect the influence of states related to inductor currents and capacitor voltages, leading to false stability predictions [32]. Moreover,

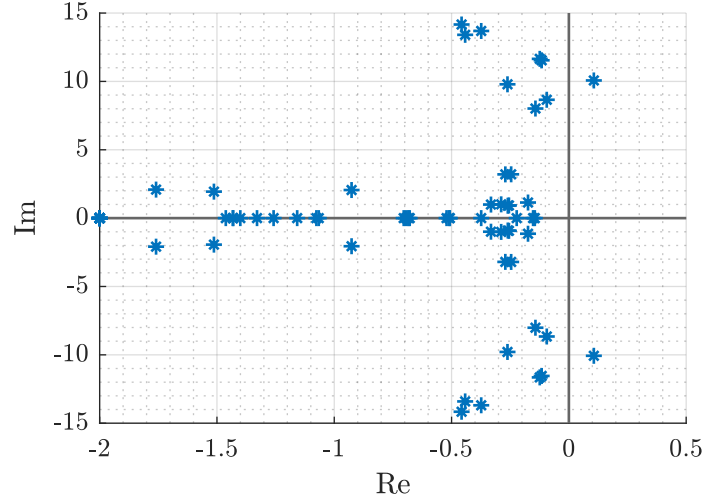


Figure 2.5: State-space of a small-signal unstable system.

converters are represented by average-value models neglecting fast dynamics of Pulse-Width Modulation (PWM) or delays in the control system [44]. To summarize, eigenvalue analysis has the following advantages and disadvantages:

- + Eigenvalue analysis can detect stability globally for the entire system at a given condition.
- + Participation factors of eigenvalues provide insight about controllers affecting system stability.
- State-space models of the entire system are required.
- State-space time-domain models neglect electromagnetic dynamics and the influence of fast control loops of converters.

2.2.4 Impedance-based Stability Analysis

Originally developed and used for the design of input filters for regulated converters [86, 87], the impedance-based approach transforms the system of interest in the frequency domain. This makes it possible to apply classical control theory tools such as the Nyquist criterion and Bode analysis for assessing the stability and investigating the root causes of potential interactions. Before applying the method, a system needs to be divided into two subsystems of those power system components that are assumed to interact with each other. Representing a system by an equivalent circuit diagram, the power system components are modeled as Thévenin or Norton equivalent with impedances representing the components' frequency behavior. Taking a voltage source converter and the connected AC grid as an example, as seen in Figure 2.6 (a), Figure 2.6 (b) demonstrates the principle of the impedance-based method. The inverter subsystem is modeled by its Norton equivalent circuit consisting of an ideal current source in parallel with its output

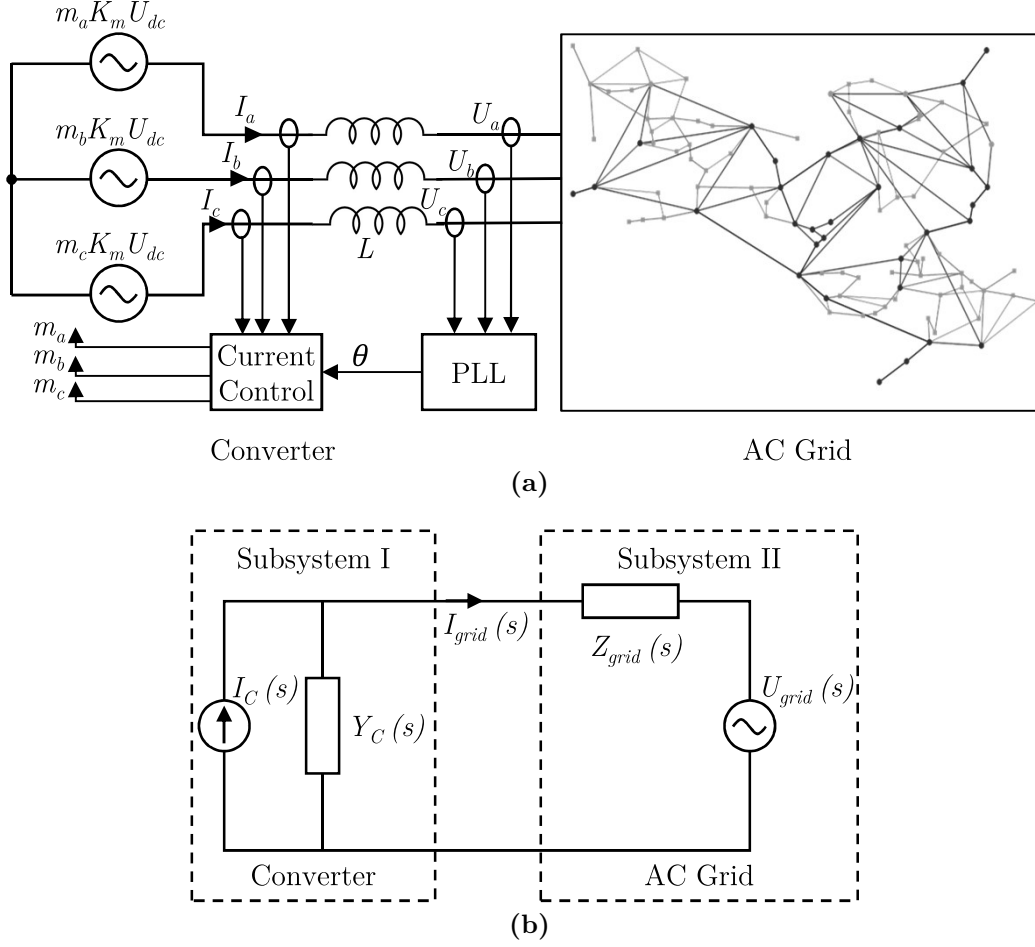


Figure 2.6: Converter connected to AC grid (a) and equivalent circuit diagram (b).

admittance. The grid subsystem is modeled by its Thévenin equivalent consisting of an ideal voltage source in series with the grid impedance [28]. The frequency behavior of the two subsystems is modeled by their admittance and impedance, respectively, where the converter admittance includes the control system's influence. The grid impedance represents the frequency behavior of the grid seen from the connection point of the converter. Because representing the subsystems as admittance and current source and impedance and voltage source implies a linear system, the impedance models require appropriate linearization of the represented components and are only valid around the operation point. Thus, the impedance-based approach only can be used as a small-signal analysis method.

Taking a closer look at the small-signal equivalent circuit, the system can be solved for the grid current $I_{grid}(s)$ and the following expression is obtained:

$$I_{grid}(s) = I_C(s) \frac{1}{1 + Y_C(s)Z_{grid}(s)} - U_g(s) \frac{Y_C(s)}{1 + Y_C(s)Z_{grid}(s)}, \quad (2.6)$$

which can be rearranged as

$$I_{grid}(s) = [I_C(s) - U_g(s) \cdot Y_C(s)] \cdot \frac{1}{1 + Y_C(s) \cdot Z_{grid}(s)}, \quad (2.7)$$

with $I_{grid}(s)$ being the current flowing into the grid, $I_C(s)$ the current from the current source, $Y_C(s)$ the parallel admittance of the current source, $Z_{grid}(s)$ the grid impedance and $U_g(s)$ the grid voltage.

From a control system point of view, equation (2.6) and (2.7) indicate a negative feedback loop system, as seen in Figure 2.7. Moreover, the product of grid impedance to converter admittance $Y_C(s)Z_{grid}(s)$ resembles the negative feedback loop gain of the system. Assuming the converter current $I_C(s)$ as an input variable and the grid current $I_{grid}(s)$ as output variable, the converter and grid subsystem can be understood as a negative feedback loop with the forward gain

$$G(s) = 1, \quad (2.8)$$

and the feedback loop gain

$$H(s) = Y_C(s)Z_{grid}(s). \quad (2.9)$$

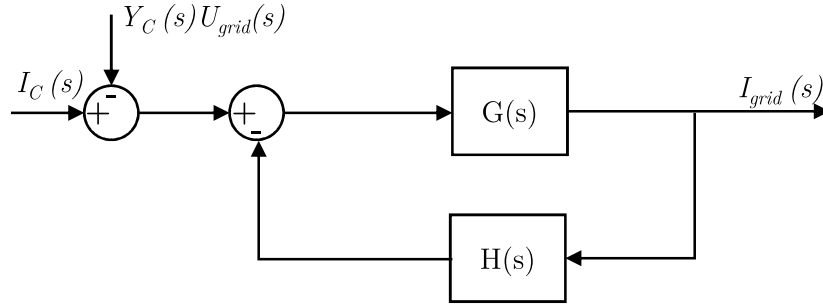


Figure 2.7: Equivalent circuit seen as feedback loop.

As a result, the feedback loop gain can be used to assess the stability of the overall system using the Nyquist stability criterion. The Nyquist stability criterion states that the closed-loop stability of a system depends on the number of unstable right-half plane poles of $G(s)H(s)$ and the number of counter-clockwise encirclements of the point $(-1, j0)$ by the Nyquist plot of $G(s)H(s)$ [88]. However, identifying the number of unstable poles can be challenging, in particular when numerically derived data is used for determining the loop gains of the system. As a consequence, the simplified Nyquist criterion typically is applied along with the impedance-based approach when assessing the stability of the system. The simplified Nyquist criterion is defined as follows [88]:

If the open-loop system is stable, then the closed-loop system is stable if and only if the Nyquist plot of the open-loop system neither encircles nor goes through the critical point $(-1, j0)$.

Assuming an open-loop stable system requires that both the impedance $Z_{grid}(s)$ and admittance $Y_C(s)$ are individually stable, i.e. do not contain any right-half plane poles. Consequently, the impedance-based stability analysis requires the following conditions[28]:

- The current source subsystem with the admittance $Y_C(s)$ is stable if the subsystem can operate in ideal grid conditions (infinite short-circuit power).
- The grid subsystem with the impedance $Z_{grid}(s)$ is stable in no-load conditions, e.g., when the converter subsystem is disconnected.

If the conditions above are satisfied, the subsystems can be assumed to be stable, and the feedback loop gain $H(s)$ does not have any right-half plane poles. Hence, the calculation of poles can be avoided, and the stability of the overall system can be determined through the Nyquist plot of $Y_C(s)Z_{grid}(s)$. Figure 2.8 depicts the Nyquist plot of the equivalent loop gain of an unstable system as seen by the encirclement of $(-1, j0)$.

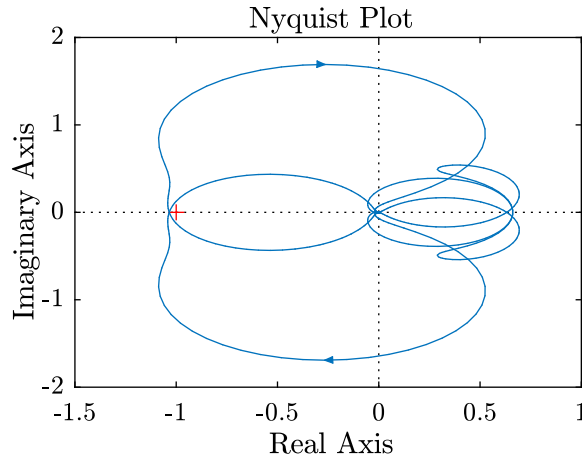


Figure 2.8: Nyquist plot of unstable system.

In addition to the Nyquist plot, the Bode plot offers a more intuitive visualization and makes it possible to demonstrate a system's proximity to unstable regions. Assuming a sinusoidal steady-state with all transients having decayed, σ of the complex variable, $s = \sigma + j\omega$ can be ignored and the transfer function $H(s)$ corresponds to the frequency response $H(j\omega)$ [88]. While the transfer function is a mathematical description of a linear system, the frequency response can be directly physically interpreted and measured [88]. Separating $H(j\omega)$ into magnitude and phase with

$$H(j\omega) = |H(\omega)|e^{j\varphi_H(\omega)}, \quad (2.10)$$

$H(j\omega)$ can be analyzed at the gain crossover angular frequency when $|H(\omega)| = 1$ or in other words at zero dB gain. Hence, a stable system is given when

$$\varphi_H(\omega) < 180^\circ \Big|_{|H(\omega)|=1}. \quad (2.11)$$

The Phase Margin (PM), indicating the degree of stability, is then calculated to

$$PM = 180^\circ - \varphi_H(\omega) \Big|_{|H(\omega)|=1}. \quad (2.12)$$

Considering the equivalent circuit depicted in Figure 2.6b and 2.7, and dropping the angular frequency argument for clarity,

$$PM = 180^\circ - \arg\left(\frac{Z_{grid}}{Z_C}\right) \Big|_{\left|\frac{Z_{grid}}{Z_C}\right|=1} = 180^\circ - (\arg(Z_{grid}) - \arg(Z_C)) \Big|_{|Z_{grid}|=|Z_C|}, \quad (2.13)$$

where $Z_C = 1/Y_C$. Figure 2.9 shows a Bode plot of an unstable system. Having a phase difference of more than 180° at the intersection frequency, and hence, a negative phase margin, an unstable system can be determined by the Bode plot. However, the Bode plots can lead to erroneous stability predictions in some cases when the derivative of the impedance magnitude and phase angle is not considered [89]. When solely evaluating the phase difference at the intersection frequency, it is possible that the corresponding Nyquist plot does not actually encircle the $(-1, j0)$ point [90]. To have a sufficient condition for stability predictions, additionally, magnitude and phase conditions need to be ensured [90]. Alternatively, a Nyquist plot can verify the stability prediction.

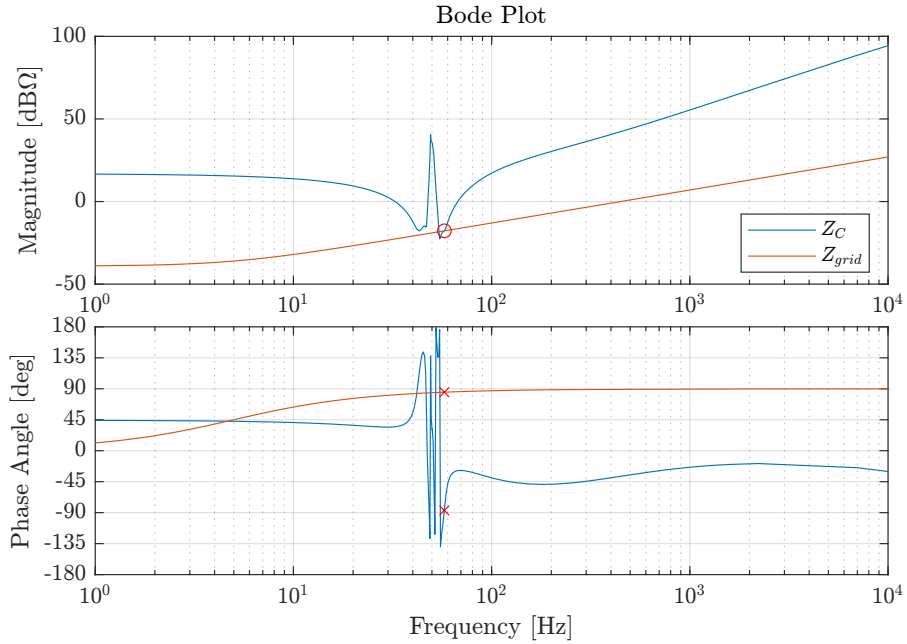


Figure 2.9: Bode plot of an unstable system.

Using the impedance-based approach to investigate the stability of converter-dominated systems in the frequency domain, has the significant advantage that the stability of the system is directly assessed, revealing the responsible system. Applying the classical control theory, the so-called Impedance-based Stability Criterion (IbSC) allows for predicting stability margins and the frequencies at which oscillations might occur.

As a result, the stability assessment only requires simple multiplications when impedance models are given and thus, results in negligible computation times.

An inherent assumption of the impedance-based approach is that a system can be divided into two subsystems potentially interacting with each other. An underlying implication of this assumption is that interfaces need to be identified, dividing the systems into the subsystems S I and S II. Figure 2.10 demonstrates various potential interfaces when considering the transformer and line impedances separately. As can be seen, disaggregating the converter and grid impedance results in several interfaces and potentially different subsystems. Consequently, interface definitions are required before the stability assessment. However, defining different interfaces only changes the calculation of the open-loop gain used to assess the stability. Although the IbSC assess the stability only locally, applying the criterion numerous times successively allows for investigating the interactions of various interfaces and subsystems due to negligible computation times [91].

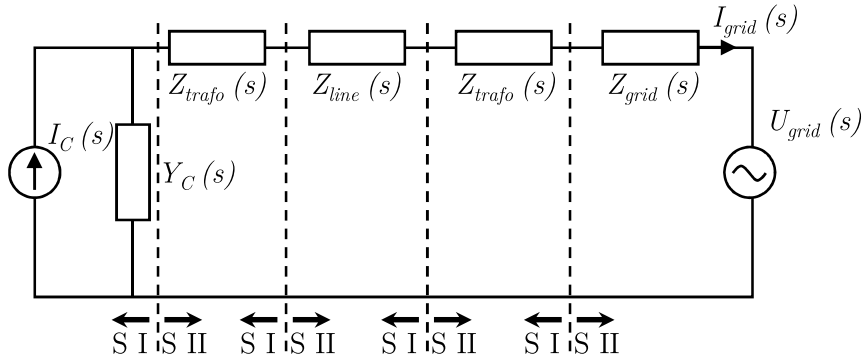


Figure 2.10: Potential interfaces defining subsystems for impedance-based approach.

However, one has to keep in mind that the stability assessment's accuracy depends on the validity of the underlying models, which – when using the impedance-based approach – assume a linear behavior of the components. As a result, stability predictions are only valid for given operating points and non-fault situations and should be verified by time-domain simulations or experimental results.

Nonetheless, the impedance-based approach offers the following advantages, making it an effective tool for studying converter control interactions:

- The impedance approach is intuitive as interactions are related to resonance conditions of two subsystems.
- Impedance models can be derived analytically using small-signal linearization techniques [92–94], as well as numerically, deriving the impedances using frequency sweep techniques on the basis of EMT simulations or measurements [95, 96].
- The stability assessment is direct and straightforward because the impedance-based analysis predicts instability as well as stability margins for stable systems.

- Factors influencing stability, such as control parameters, can be determined by investigating the impact on the frequency behavior.
- The approach can be easily scaled to larger systems by applying the criterion numerous times consecutively and defining different interfaces.

The impedance-based approach is considered also in the grid code "*VDE-AR-N 4131 Technical requirements for grid connection of high voltage direct current systems and direct current-connected power park modules (TAR HVDC)*", allowing impedance models of HVDC systems to be derived analytically as well as numerically with the following requirements [42]:

- The influence of active control, measuring, and protection systems needs to be included in the frequency-dependent impedance.
- Different operation points, including power flow directions, need to be considered.
- Coupling between frequencies or component systems can only be neglected if justified in the model's documentation.
- The impedance model needs to have a minimum resolution of 1 Hz and be valid from 10 Hz up to 2.5 kHz or 9 kHz if requested by a TSO.
- The impedance model needs to be derived in the positive and negative sequence.

2.2.5 Conclusion

Different methods can be used to assess the converter control interactions reported by several transmission system operators. However, they differ in the ability and accuracy of assessing instabilities and applicability for using black-boxed models. Table 2.2 shows a comparison of the discussed methods for analyzing the converter control interactions. A disadvantageous or not applicable method for a given criterion is indicated with -, while + and ++ indicate good or very good suitability, respectively.

Table 2.2: Comparison of analysis methods.

Criterion	EMT Simulation	Laboratory Test Bench	Eigenvalue Analysis	Impedance-based Analysis
Modeling ease	-	-	-	+
Block-box models	+	+	-	++
Numerical models	+	+	-	++
Large-scale systems	-	-	+	+
Computational time	-	-	+	++
Accuracy of results	++	++	-	+
Stability margins	-	-	+	+
Insight in phenomenon	-	-	++	+

EMT simulations and eigenvalue analysis necessitate models of every power system component. A laboratory test bench needs interface algorithms. In contrast, the impedance-based analysis requires only modeling of the components assumed to interact with each other. Thus, it can be applied in case of limited information of the power system. Additionally, black-box models only containing information on converters' frequency behavior can be applied, which protect the manufacturer's IPRs for the control systems. Although EMT simulation tools and laboratory converters can utilize compiled black-box models in which sensitive information such as converter control systems is encrypted, there are concerns among manufactures that compiled black-box models may be reverse-engineered due to the level of detail provided in EMT models [97]. Eigenvalue analyses, however, rely on a state-space representation where dependencies and derivatives of the states and variables of the models need to be fully disclosed, thus, preventing the use of black-box models. Furthermore, the requirement that interdependencies between the components need to be modeled also does not permit the use of numerical models obtained by voltage and current signals for the eigenvalue analysis. Although it is possible to use numerically-derived models in time-domain simulations, the process is not straightforward and requires the definition of an equivalent model and the use of system identification techniques to estimate the respective parameters [98]. Additionally, the procedure, model extraction, and accuracy validations of such models can be challenging, particularly for EMT simulations and converters where time steps are small, and the required level of detail is high [99]. In contrast, impedance models only require high-resolution data of voltages and currents at different frequencies [95]. Due to little computation times, the IbSC can be applied numerous times for investigating large-scale systems. EMT simulations rely on co-simulations, simulating large-scale grids with less detail using RMS-simulation for parts of the system [100] and thus, losing accuracy [72]. The same is valid for laboratory test benches when large-scale AC grids are simulated on RTSs. However, when simulating size-limited systems, EMT simulations provide very high accuracy, as no inherent assumptions and modeling simplifications are made. However, laboratory test benches capture real converters' behavior without the risk that potentially important phenomena are neglected. Eigenvalue analysis depends on state-space models, relying on simplified average value models of the components. Although the impedance-based approach employs linearized models, the IbSC can predict a system's stability with high confidence when underlying assumptions are considered. In addition, phase or gain margins can indicate how close a system is to instability. Stability margins can also be derived by eigenvalue analysis from the proximity of left-half plane poles to the imaginary axis. However, EMT simulations and laboratory test benches provide no additional information, requiring post-processing and analyses of the waveforms to study interactions. In contrast, eigenvalue and impedance-based analyses provide insight into the phenomenon through participation factors and parameter sensitivity analyses.

The analysis shows that the impedance-based approach is adequate to answer the first and third research questions, providing a means to investigate converter control interactions with encrypted black-boxed controllers and identify critical grid situations. However, the

approach should be verified with EMT simulations due to the inherent simplification and component linearization. While laboratory test benches are not well-suited for investigating converter-control interactions because it cannot be easily distinguished whether identified instabilities result from converter-control interactions or the PHIL interface method, they can be used for answering the second research question. Comparing the results of simulated models to the real-component behavior of a test bench converter makes it possible to reveal the impact of model simplifications with respect to the converter's electrical structure.

2.3 Impedance Modeling

The previous section pointed out the impedance-based approach as a preferred method for investigating converter control interactions. Investigating the transmission system, however, requires approaches to consider the three-phase nature of the system. Furthermore, an adequate approach to derive impedance models needs to be chosen, considering the non-linearity of the system components and the thesis's objective.

A widely applied method in power systems to analyze three-phase systems is the sequence-domain transformation which has been used, for example, to investigate symmetrical rotating machinery, transformers, and transposed transmission lines [101–103]. The system is characterized by a positive-sequence, negative-sequence, and where necessary zero-sequence impedance circuit that are decoupled from each other. Another method, often applied when dealing with power electronics and control systems [74], is to apply the Park transformation and investigate the impedance circuit in a dq-domain system, rotating synchronously at the grid frequency and being aligned with the grid angle θ . In addition to the sequence and dq-domain, the modified sequence-domain method was proposed to bridge the gap between dq-domain and sequence-domain impedance modeling [104]. To determine the suitable domain in this thesis, sections 2.3.2 and 2.3.3 review advantages and disadvantages and the implications for impedance-based analyses resulting from the selected domain.

Impedance models can be obtained in two fundamentally different ways: a model is determined analytically by linearizing the components and deriving transfer functions. Deriving the frequency response numerically by voltage and current signals permits a stability analysis employing the impedance-based approach as well. The signals can be obtained from converter models simulated in EMT time-domain simulations or from the measurements of the frequency response of physical components. Both approaches come with different advantages and disadvantages. Thus, sections 2.3.4 and 2.3.5 will provide an overview of both approaches in addition to the implications by choice of domain with respect to the objective and the research questions.

2.3.1 Frequency Coupling

Frequency coupling describes an effect observed for power system converters resulting from the converter's non-linear behavior and its control system and the inter-modulation of the grid frequency f_{grid} and a perturbation frequency f_{pert} . The effect can be demonstrated when perturbing a converter with a frequency f_{pert} that is superimposed on the grid frequency. When perturbing a system at a frequency f_{pert} , the converter responds at frequencies other than f_{pert} in addition to f_{pert} . The literature often refers to this effect as *frequency coupling* because it can be understood as couplings between positive and negative sequence or cross-couplings of the dq-domain impedance matrix's non-diagonal elements [105, 106]. Referring to the specific frequency component $2f_{grid} - f_{pert}$ that is observed in the current response of a three-phase VSC, the term *coupling over frequency* is also used [73]. Other authors point out that the relationship of the resulting coupling frequencies f_{pert-} and f_{pert+} and the grid frequency of the dq-transformation follows a fixed relationship that can be described by the following equation [107]:

$$f_{pert-} + f_{pert+} = 2 \cdot f_{grid}. \quad (2.14)$$

Emphasizing the fact that f_{pert-} and f_{pert+} can be seen as frequency components that are equally distanced from f_{grid} , the term *mirror frequency coupling* is used to describe the same phenomenon [35]. In addition, current controllers that control signals differently in the d- and q-axis are identified for causing the coupling frequencies [35]. Coupling effects could also be seen in measurements of a STATCOM that interacted with a weak part of the China Southern Grid [25]. It is mathematically shown that the dq-transformation of the PLL transforms a component at the frequency f into two simultaneous components in the stationary frame, one at $f_{grid} - f$ and the other at $f_{grid} + f$ [25]. However, also PWM can contribute to frequency coupling [108].

2.3.2 dq-Domain

Traditionally, the dq-domain has been used for analyzing the transient behavior of synchronous machines [49]. Three-phase voltages can be transformed into a new dq0 coordinate system by applying the Park transformation, which contains only real matrix elements. Applying the transformation to the dq-domain voltages

$$\begin{bmatrix} U_0 \\ U_d \\ U_q \end{bmatrix} = \sqrt{\frac{2}{3}} \underbrace{\begin{bmatrix} \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \\ \cos \theta & \cos(\theta - \frac{2\pi}{3}) & \cos(\theta + \frac{2\pi}{3}) \\ -\sin \theta & -\sin(\theta - \frac{2\pi}{3}) & -\sin(\theta + \frac{2\pi}{3}) \end{bmatrix}}_{T_{0dq}} \begin{bmatrix} U_a \\ U_b \\ U_c \end{bmatrix}, \quad (2.15)$$

where θ is the transformation angle. Usually, the transformation angle is chosen in a way that the d-axis is aligned with the peak of phase a [35] with

$$\theta = 2\pi f_{grid} + \varphi_a, \quad (2.16)$$

where f_{grid} is the grid frequency and φ_a is the phase angle of phase a at $t = 0$. For converter systems, θ should be set to the voltage angle at the converter terminal, which is also used to synchronize the converter output voltages with the connected AC grid. A scaling factor of $\sqrt{2/3}$ is used so that $T_{0dq}^{-1} = T_{0dq}^T$ whereas the original Park transformation applies a scaling factor of $2/3$ [109]. The 0-axis can be omitted for balanced three-phase systems or systems without neutral connections when no 0-axis current can flow [84] so that

$$\begin{bmatrix} U_d \\ U_q \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos \theta & \cos(\theta - \frac{2\pi}{3}) & \cos(\theta + \frac{2\pi}{3}) \\ -\sin \theta & -\sin(\theta - \frac{2\pi}{3}) & -\sin(\theta + \frac{2\pi}{3}) \end{bmatrix} \begin{bmatrix} U_a \\ U_b \\ U_c \end{bmatrix}. \quad (2.17)$$

The resulting voltages U_d and U_q are constant in steady-state with respect to the rotation frequency. As a result, linear system theory can be applied, e.g. simple PI-controllers can be used for controlling phase currents transformed by means of the Park transformation. As such, the dq-domain has been used for control loop and filter design of three-phase PWM rectifiers [110, 111].

In the dq-domain, the relationship between voltages and currents is given by an impedance matrix $\mathbf{Z}_{dq}$ where

$$\begin{bmatrix} U_d \\ U_q \end{bmatrix} = \mathbf{Z}_{dq} \begin{bmatrix} I_d \\ I_q \end{bmatrix} \quad (2.18)$$

with

$$\mathbf{Z}_{dq} = \begin{bmatrix} Z_{dd} & Z_{dq} \\ Z_{qd} & Z_{qq} \end{bmatrix}. \quad (2.19)$$

Equation (2.19) illustrates that the q-axis and d-axis impedances, Z_{dd} and Z_{qq} are coupled due to the existence of the off-diagonal elements, Z_{dq} and Z_{qd} in $\mathbf{Z}_{dq}$. As a result, the d-axis and q-axis cannot be evaluated separately and the equivalent impedance circuit of the impedance-based analysis is a Multiple Input and Multiple Output (MIMO) system as seen in Figure 2.11.

Due to multiple inputs and outputs, the loop-gain assessment is not as straightforward when applying the IbSC. Because a 2x2 matrix defines a subsystem's impedance, the loop gain assessed by the IbSC is a 2x2 matrix as well, so that the simplified Nyquist criterion cannot be applied. Therefore, the generalized Nyquist criterion needs to be applied to determine the stability of a MIMO system such as the dq-domain equivalent circuit [112]. The generalized Nyquist criterion plots the graphs of the eigenvalues, $\lambda_i(s)$,

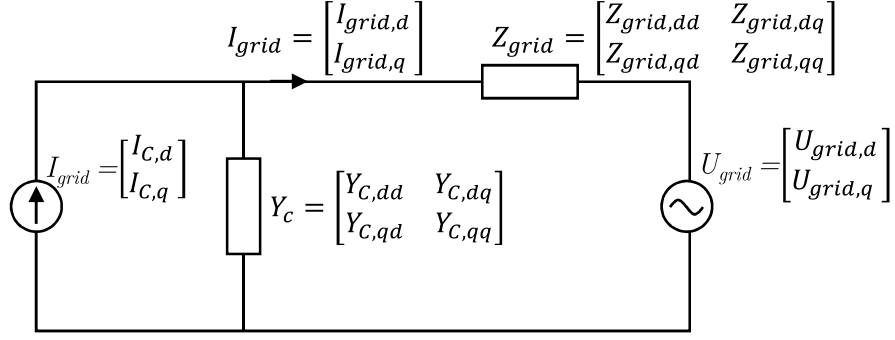


Figure 2.11: dq-domain circuit.

of the loop gain matrix to determine the stability. As a result, the Nyquist plots of both eigenvalues of $Z_{grid}(s)Y_C(s)$ determine the system's stability. Contrary, for Single Input and Single Output (SISO) systems, the Nyquist plot is generated by the magnitude and phase of the loop gain only [113].

To summarize, assessing the stability in the dq-domain using the impedance-based analysis has the following implications:

- A common rotation angle, θ , is required for both impedance models Y_C and Z_{grid} so that the individual models have the same reference frame. For converter-based systems, θ can be chosen as the grid angle at the converter terminal.
- System dynamics in the q-axis and d-axis are closely coupled, resulting in a 2x2 matrix loop gain. Thus, the generalized Nyquist Criterion needs to be applied. Consequently, potential couplings of the impedances are considered inherently already.
- Unbalanced systems result in time-variant 0-axis variables, making the linearization of models impossible [84].

2.3.3 Sequence Domain

Impedance-based stability analyses can be performed for three-phase systems using the sequence transformations to decompose an unbalanced system into single-phase decoupled balanced subsystems in the sequence domain. In the sequence domain, the three-phase voltages and currents are transformed into a set of balanced voltages and currents called symmetrical components, which are defined as follows:

- Zero-sequence voltage and current phasors U^0 and I^0
- Positive-sequence voltage and current phasors U^p and I^p
- Negative-sequence voltage and current phasors U^n and I^n

In a balanced system, the three-phase voltage phasors U_a, U_b, U_c are equal in magnitude, having each a phase shift of 120° . The phasor U_b lags U_a by 120° and phasor U_c lags U_b by 120° . The set of voltage phasors in such sequence have a *positive sequence abc* [114]. Defining U_a as the reference phasor, the three positive-sequence phasors

$$\begin{bmatrix} U_a \\ U_b \\ U_c \end{bmatrix} = \begin{bmatrix} 1 \\ \underline{a}^2 \\ \underline{a} \end{bmatrix} U_a, \quad (2.20)$$

with the complex phasor

$$\underline{a} = e^{j\frac{2\pi}{3}}. \quad (2.21)$$

If the phase sequence of the voltage phasors is *acb*, then it is said to be in negative sequence with

$$\begin{bmatrix} U_a \\ U_b \\ U_c \end{bmatrix} = \begin{bmatrix} 1 \\ \underline{a} \\ \underline{a}^2 \end{bmatrix} U_a. \quad (2.22)$$

If the voltage phasors have the same magnitude and phase angle, the sequence is referred to as zero sequence with

$$\begin{bmatrix} U_a \\ U_b \\ U_c \end{bmatrix} = \begin{bmatrix} 1 \\ 1 \\ 1 \end{bmatrix} U_a. \quad (2.23)$$

These sets of positive-, negative-, and zero-sequence phasors are referred to as the symmetrical components of the three-phase system phasors U_a, U_b, U_c .

Applying the symmetrical component transformation by Fortescue [115], the original phasors

$$\begin{bmatrix} U_a \\ U_b \\ U_c \end{bmatrix} = \begin{bmatrix} 1 & 1 & 1 \\ 1 & \underline{a}^2 & \underline{a} \\ 1 & \underline{a} & \underline{a}^2 \end{bmatrix} \begin{bmatrix} U^0 \\ U^p \\ U^n \end{bmatrix} \quad (2.24)$$

and the inverse matrix determines the sequence phasors where

$$\begin{bmatrix} U^0 \\ U^p \\ U^n \end{bmatrix} = \frac{1}{3} \begin{bmatrix} 1 & 1 & 1 \\ 1 & \underline{a} & \underline{a}^2 \\ 1 & \underline{a}^2 & \underline{a} \end{bmatrix} \begin{bmatrix} U_a \\ U_b \\ U_c \end{bmatrix}. \quad (2.25)$$

The sequence transformation used in (2.25) also can be applied to currents and is not only valid for phasors at the fundamental frequency but also phasors of any frequency. In contrast to the dq-transformation, the matrix elements are complex. The zero sequence is usually omitted because a zero-sequence current cannot flow in a three-leg converter

[92, 116]. However, in low-voltage systems with a neutral wire, the zero sequence should be considered [117]. Disregarding the zero-sequence component, a 2x1 vector results from the sequence transformation. In addition, assuming balanced power systems, impedances of each phase are identical, which has the following implications:

- A positive-sequence current at a frequency, f , causes only a positive- and not a negative-sequence voltage at frequency f , and a positive-sequence voltage causes only a positive-sequence current at that frequency, respectively.
- A negative-sequence current at a frequency, f , causes only a negative and not a positive-sequence voltage at frequency f , and a negative-sequence voltage causes only a negative-sequence current at that frequency, respectively.

Assuming a balanced power system decouples both sequences. Consequently, the sequence transformation can be used to analyze a balanced system separately in the positive and negative sequences. Regarding the impedance-based analysis, this means that the three-phase power system's stability can be assessed by applying the impedance-based stability criterion separately, both for a positive and negative-sequence system. Thus, the required frequency-dependent impedance models of the systems that are assumed to interact with each other can be derived by the positive and negative-sequence voltages and currents with

$$Z^p(s) = \frac{U^p(s)}{I^p(s)} \quad (2.26)$$

and

$$Z^n(s) = \frac{U^n(s)}{I^n(s)}, \quad (2.27)$$

where $s = j2\pi f$ and f can be of any frequency. Figure 2.12 depicts the positive and negative-sequence circuits where the grid impedance Z_{grid}^p and converter admittance Y_C^p represent the positive-sequence system, and Z_{grid}^n and Y_C^n represent the negative-sequence impedance models of the negative-sequence system, respectively. Thus, applying the impedance-based stability criterion for both systems, as explained in section 2.2.4, allows for a stability assessment of the equivalent three-phase system.

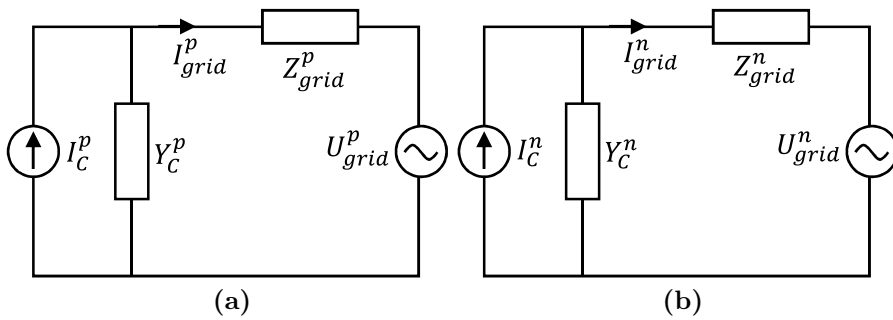


Figure 2.12: Positive-sequence circuit (a) and negative-sequence circuit (b).

To conclude, the following implications should be considered when applying impedance-

based analyses in the sequence-domain:

- The sequence-domain allows for individual impedance modeling in the positive and negative sequence. This enables the application of the simplified Nyquist criterion, and the stability of the systems can be assessed separately.
- No coupling between sequences or frequencies is included inherently. It should be verified that no couplings exist, or techniques to include couplings should be deployed [118, 119].
- The sequence transformation does not rely on a common input variable, such as the grid angle, for aligning positive and negative sequences. This makes it possible to perform stability analyses using models of different parties (e.g., grid impedance model from an TSO and converter model from a manufacturer) when the grid angle is unknown or only known for one party.

2.3.4 Analytical Derivation

Several impedance models have been derived analytically to represent power system converters and their control systems. The derivation depends heavily on the considered domain so that different models of converters have been proposed for dq-domain, sequence domain, or in the abc-phase domain. Analytical models employ linearization techniques to develop an impedance model around an operating point as a transfer function. Superimposing small-signal perturbations on the grid voltage in dq- or sequence domain allows for the linearization of non-linear controllers such as the PLL.

An impedance model of a 2-level VSC in the dq-domain can be found in [33]; a sequence-domain model is given by [120]. The 2-level voltage source converter will be referred to later on as VSC while the modular multilevel voltage source converter will be referred to as MMC. Taking the average-value representation of a VSC shown in Figure 2.13 as an example, the following sections illustrate the analytical impedance modeling approach in the sequence as well as in the dq-domain.

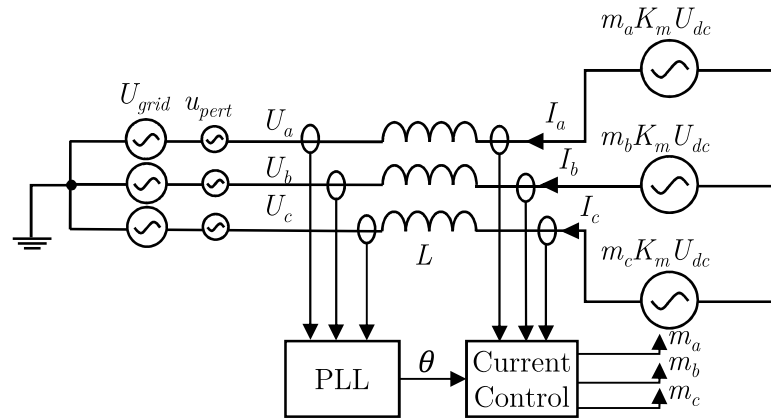


Figure 2.13: Average value model of a grid-connected 2-level VSC.

Analytical Modeling in Sequence Domain

The harmonic linearization technique is used to linearize the converter at an operating point of VSC to derive a small-signal linear impedance model. Again, it should be noted that the term harmonic does not refer to the grid frequency only; the harmonic can be also a multiple-integer of a perturbation frequency. The technique superimposes harmonic perturbations on the system and subsequently determines the induced response at the perturbation frequency to derive the overall impedance model. For instance, a small-signal perturbation voltage is superimposed on the AC grid voltage at the converter terminal, such as that for positive-sequence perturbations, the phase voltages

$$\begin{aligned} U_a(t) &= U_{grid} \cos(2\pi f_{grid}t + \varphi_{grid}) + u_{pert} \cos(2\pi f_{pert}t + \varphi_{pert}) \\ U_b(t) &= U_{grid} \cos(2\pi f_{grid}t - \frac{2\pi}{3} + \varphi_{grid}) + u_{pert} \cos(2\pi f_{pert}t - \frac{2\pi}{3} + \varphi_{pert}) \\ U_c(t) &= U_{grid} \cos(2\pi f_{grid}t + \frac{2\pi}{3} + \varphi_{grid}) + u_{pert} \cos(2\pi f_{pert}t + \frac{2\pi}{3} + \varphi_{pert}) \end{aligned} \quad (2.28)$$

and that for negative-sequence perturbations

$$\begin{aligned} U_a(t) &= U_{grid} \cos(2\pi f_{grid}t + \varphi_{grid}) + u_{pert} \cos(2\pi f_{pert}t + \varphi_{pert}) \\ U_b(t) &= U_{grid} \cos(2\pi f_{grid}t - \frac{2\pi}{3} + \varphi_{grid}) + u_{pert} \cos(2\pi f_{pert}t + \frac{2\pi}{3} + \varphi_{pert}) \\ U_c(t) &= U_{grid} \cos(2\pi f_{grid}t + \frac{2\pi}{3} + \varphi_{grid}) + u_{pert} \cos(2\pi f_{pert}t - \frac{2\pi}{3} + \varphi_{pert}) \end{aligned} \quad (2.29)$$

where U_{grid} is the amplitude of the grid phase voltages, f_{grid} is the grid frequency, u_{pert} is the amplitude of the small-signal perturbation voltage and f_{pert} is the frequency of the perturbation voltage, φ_{grid} is the grid phase angle and φ_{pert} is the perturbation voltage phase angle.

The response of the converter to the small-signal voltage perturbation is given by the phase currents $I_{a,b,c}$. Equation (2.30) describes the electrical model of the converter represented in Figure 2.13 and gives the relationship between the phase currents and voltages as follows:

$$L \frac{d}{dt} \begin{bmatrix} I_a \\ I_b \\ I_c \end{bmatrix} = \begin{bmatrix} m_a \\ m_b \\ m_c \end{bmatrix} K_m U_{dc} - \begin{bmatrix} U_a \\ U_b \\ U_c \end{bmatrix} \quad (2.30)$$

where L is the converter inductance and $m_{a,b,c}$ with gain K_m are the modulating signals required for the control of the PWM that the current controller gives. By solving (2.30) in the frequency domain, the positive-sequence impedance $Z^p(f)$ and negative-sequence impedance Z^n can be obtained by the ratio of the voltage U^p and current $-I^p$ and U^n and $-I^n(s)$, respectively. To include the dynamics and influence of the converter control system on the currents, the modulating signals need to be derived as a function of the voltage, and current perturbations using the block diagrams of the respective control systems [120]. Modeling the influence of the dq-domain current controller and assuming a perfectly tracked grid angle θ results in the following frequency-dependent positive-

and negative-sequence impedances of the converter with

$$Z^p(s) = K_m U_{dc} [H_i(s - j2\pi f_{grid}) - jk_d] + sL \quad (2.31)$$

and

$$Z^n(s) = K_m U_{dc} [H_i(s + j2\pi f_{grid}) + jk_d] + sL \quad (2.32)$$

where

$$s = j\omega = j2\pi f_{pert}. \quad (2.33)$$

In (2.31) and (2.32) H_i is the Proportional Integral (PI) controller of the current controller in the dq-domain and k_d the decoupling gain of the q-axis and d-axis currents.

Additionally considering the PLL that tracks the grid angle θ requires the linearization of the Park transformation which is done by approximating the transformation around the grid angle $\theta + \Delta\theta$. The effect of the PLL on the current control is modeled through the influence of $\Delta\theta$ on the d-axis and q-axis current. The resulting sequence impedance models of the converter are given by

$$Z^p(s) = \frac{k_m U_{dc} [H_i(s - j2\pi f_{grid}) - jk_d] + sL}{1 - k_m U_{dc} [(H_i(s - j2\pi f_{grid}) - jk_d) \frac{I_{grid}}{2} + \frac{c_1}{2} e^{j\phi_c}] T_{PLL}(s - j2\pi f_{grid})} \quad (2.34)$$

and

$$Z^n(s) = \frac{k_m U_{dc} [H_i(s + j2\pi f_{grid}) + jk_d] + sL}{1 - k_m U_{dc} [(H_i(s + j2\pi f_{grid}) + jk_d) \frac{I_{grid}}{2} + \frac{c_1}{2} e^{-j\phi_c}] T_{PLL}(s + j2\pi f_{grid})}. \quad (2.35)$$

The influence of the PLL is given by the closed-loop transfer function T_{PLL} . The variables c_1 and ϕ_C describe the relationship between the AC and DC voltage.

As it can be seen by (2.34) and (2.35), the analytical modeling results in two separate transfer functions of the converter, revealing the influence of controllers and system parameters on the converter impedance. However, the modeling of the controllers requires the linearization of the Park transformation because most controllers regulate the currents in the dq-domain.

Analytical Modeling in dq-Domain

Based on the same average value model shown in 2.13, the converter impedance model can be modeled in the dq-domain [33] by describing the electrical model in the $\alpha\beta$ frame at first. The $\alpha\beta$ is conceptually similar to the dq-domain and also uses a real transformation matrix. However, contrary to the dq-domain, the reference frame is not rotating and the variables remain sinusoidal. As a starting point the complex space voltage and current vectors

$$\begin{aligned} U_{\alpha\beta} &= U_\alpha + jU_\beta \\ I_{\alpha\beta} &= I_\alpha + jI_\beta \end{aligned} \quad (2.36)$$

are used. The dq-frame signals are given by

$$\begin{aligned} I_{dq} &= I_d + jI_q \\ U_{dq} &= U_d + jU_q \\ C_{dq} &= C_d + jC_q \end{aligned} \quad (2.37)$$

where I_{dq} is the current, U_{dq} is the voltage and C_{dq} is the modulating signal in dq-frame. The electrical model is described in the frequency domain with

$$\frac{C}{sL} k_m U_{dc} - \frac{U_{\alpha\beta}}{sL} = I_{\alpha\beta}. \quad (2.38)$$

The signals can be transformed from $\alpha\beta$ to dq-frame by the transformation matrix $T_{\alpha\beta}$ with the real space vectors

$$\begin{bmatrix} U_d \\ U_q \end{bmatrix} = \underbrace{\begin{bmatrix} \cos\theta & \sin\theta \\ -\sin\theta & \cos\theta \end{bmatrix}}_{T_{\alpha\beta}} \begin{bmatrix} U_\alpha \\ U_\beta \end{bmatrix}. \quad (2.39)$$

Using Euler's formula with

$$e^{j\theta} = \cos\theta + j\sin\theta \quad (2.40)$$

$U_{\alpha\beta}$ and accordingly $I_{\alpha\beta}$ can be simplified to

$$\begin{aligned} U_{dq} &= U_{\alpha\beta} e^{-j\theta} \text{ and} \\ I_{dq} &= I_{\alpha\beta} e^{-j\theta}. \end{aligned} \quad (2.41)$$

To transform signals in dq to $\alpha\beta$ frame, the reverse transformation matrix

$$\underbrace{\begin{bmatrix} \cos\theta & -\sin\theta \\ \sin\theta & \cos\theta \end{bmatrix}}_{T_{\alpha\beta}^{-1}} \quad (2.42)$$

can be employed and using (2.40)

$$\begin{aligned} U_{\alpha\beta} &= U_{dq} e^{j\theta} \text{ and} \\ I_{\alpha\beta} &= I_{dq} e^{j\theta}. \end{aligned} \quad (2.43)$$

Modeling the converter impedance in dq-frame also requires linearizing the PLL. In steady-state, the PLL tracks the grid angle with $\theta = 2\pi f_{grid}t$. Aligning the d-axis with the grid voltage by the tracked grid angle of the PLL sets the q-axis to zero resulting in

$$U_{\alpha\beta} = U_{d,grid} e^{j\theta_{grid}}. \quad (2.44)$$

Introducing a voltage perturbation ΔU_{dq} on the aligned dq voltages gives

$$U_{\alpha\beta} = (U_{d,grid} + \Delta U_{dq})e^{j\theta_{grid}}. \quad (2.45)$$

The PLL responds to the voltage perturbation with a small deviation in the tracked grid angle so that

$$\theta = \theta_{grid} + \Delta\theta \quad (2.46)$$

and

$$U_{dq} = (U_{d,grid} + \Delta U_{dq})e^{-j\Delta\theta}. \quad (2.47)$$

Equation (2.47) can be linearized by the first-order Taylor series and used to derive the relationship between the controllers and the voltage perturbation, which results in the final impedance model with

$$\begin{bmatrix} I_d \\ I_q \end{bmatrix} = \underbrace{\left[G_{cl,dq}(s) Y_{PLL}^m(s) - \frac{Y_{m_{dq}}(s)}{1 + T_{dq}^m(s)} \right]}_{Y_{dq}(s)} \begin{bmatrix} U_d \\ U_q \end{bmatrix} \quad (2.48)$$

The term $G_{cl,dq}(s)$ is the closed-loop response of the current controller, $Y_{PLL}^m(s)$ describes the influence of the voltage signals on the current signals through the PLL. $Y_{m_{dq}}(s)$ and T_{dq}^m are part of the current controller as well and model the relationship of the voltage signals on the modulating signal as well as the dependency of the DC voltage on the control signals.

The modeling in the dq-domain requires fewer linearizations of the Park transformation because most controllers are already implemented in the dq-domain. Still, voltage perturbations superimposed on the dq-voltages are needed to linearize the PLL and model its influence on the converter impedance. While the analytical model gives insight into how the controllers shape the converter's admittance, the model is given as a 2x2 matrix so that the generalized Nyquist criterion is required for assessing the stability. As a result, the matrix elements themselves do not provide meaningful insight; conclusions can be drawn only from the impedance or admittance matrix's eigenvalues. While this complicates the analysis compared to the decoupled sequence domain impedances, couplings between frequencies are included. In particular, couplings through non-linear controllers such as the PLL are considered in the model contrary to decoupled sequence domain models [39, 105, 121].

MMC Impedance Models

The electrical structure of an MMC is shown in Figure 2.14. The MMC consists of one upper arm and one lower arm per phase. The arms comprise a string of SMs that can be modeled by controllable voltage sources. In addition to 2-level VSCs, analytical impedance models of MMCs are also proposed in the literature. However, MMCs require

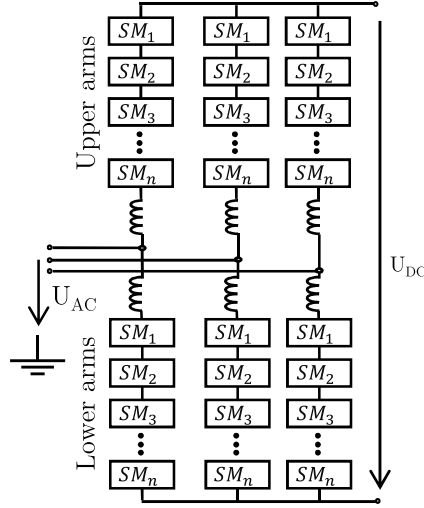


Figure 2.14: Electrical structure of an MMC.

more advanced modeling techniques due to internal dynamics such as harmonic currents within the converter and the more complex electrical structure. Moreover, the impedance model of an MMC needs to include additional controllers such as the circulating current control.

An MMC model by Sun develops MMC impedances in the sequence domain employing a matrix-based approach. It uses multi-harmonic linearization, considering any potential internal harmonics within the converter [92, 116]. However, the model involves numerous matrix operations and needs to obtain either steady-state voltages and currents or the insertion index via simulations or to solve additional non-linear equations. In [31], the impedance model of an MMC is derived based on a single-phase equivalent circuit. Using the harmonic state-space modeling method, which is, as the authors state, equivalent to the multi-harmonic linearization method, the model also requires both a steady-state and small-signal representation of the MMC. Also derived utilizing a single-phase representation, an AC-side admittance model of a MMC is developed in [80] to achieve high accuracy yet avoids complex modeling. As a result, the authors present the influence of different control systems on the converter admittance [122]. While internal control loops such as circulating current control and PLL are included, outer control loops such as AC active or reactive power and DC voltage controllers are not considered. The MMC model recently presented in [34] derives the MMC impedance as SISO matrix using complex vectors and harmonic transfer-function matrices to include also the effect of zero-sequence currents circulating inside the converter. Still, due to the model's complex matrix structure, modeling different types of control systems is hardly possible.

Published MMC models deal with the additional challenges imposed by the complex structure of the MMC by modeling one phase arm of the MMC and taking advantage of symmetries to build the overall converter model. Although this simplifies the models, this approach is only practical when modeling the MMC in the sequence or phase domain. Consequently, MMCs have not been modeled in the dq-domain thus far.

Conclusion

Analytically deriving impedance models can be done both in the sequence and the dq domain with the following advantages and limitations:

- Both dq and sequence domain models reveal the influence of control loops and converter parameters on the impedance model. However, decoupled sequence domain models show a direct influence on the impedances.
- Models in the dq domain show couplings due to non-linear controllers directly through the off-diagonal elements in the impedance matrix.
- A challenge when deriving analytical converter models in both domains is the linearization of the PLL and Park transformation. However, modeling in the dq-domain has the advantage that for VSC most controllers already operate in the dq-domain, simplifying the modeling.
- For MMCs, modeling in the sequence or phase domain is preferable because MMCs can be derived based on a single-phase one-arm model of the converter.
- Analytical models rely on model simplifications and neglect control loops that are assumed not to influence the impedance. While this is often valid, it needs to be verified. In particular, for MMCs, not every control loop can be modeled easily. Modeling different control systems can be challenging and time-consuming. In particular, discrete controllers such as SM sorting algorithms cannot be easily included in analytical impedance models that rely on continuous-state representations.
- Analytical modeling is possible only with full knowledge of the converter parameters and control system.

2.3.5 Numerical Derivation

An advantage of the impedance-based analysis is that the components' impedance models can be derived from voltage and current data, even when no information about the system is given. The current and voltage data can be obtained by EMT simulations or measurements of laboratory converters. In theory, current and voltage signals of HVDC stations can be used as well. However, as it will be explained in the following paragraphs, a numerical impedance derivation relies on perturbations to excite a converter response at different frequencies. While this is unproblematic when using EMT simulations or laboratory converters, perturbations can potentially cause undesired effects such as high distortions or damage of components, and therefore are not practical for in-field measurements.

The principle of numerical impedance derivations is to perturb a system or Device Under Test (DUT) at a given frequency by a small-signal voltage or current superimposed on grid voltage or current. By recording the terminal voltage and current at the perturbed

frequency, the response of the DUT is obtained, and the impedance at that frequency is calculated by applying Ohm's law. Perturbing the system with a series of frequencies results in a model valid between the lowest and highest injected perturbation frequency. The accuracy depends on the number of applied frequencies and the resulting resolution.

For injecting the perturbation frequencies, two main approaches exist. Sequentially applying perturbation frequencies over time, one after another, is the most common approach, referred to as frequency scan or frequency sweep. Contrarily, combining all frequencies in one injection signal applies all perturbations simultaneously. Simultaneously frequency techniques are referred to as wide- or broad-band or multi-tone signal injections [123–125]. Subsequently, the recorded voltage and current signals can be evaluated for the frequency range of interest.

Sequential impedance approaches provide the highest accuracy. The energy of the injection signals can be small because only one injection frequency is applied at a time so that the system is disturbed as little as possible. In addition, the approach allows for analyzing the response frequencies and their origin. For instance, a current response of a converter other than the perturbation frequency can be recognized easily. However, the approach is time-consuming, requiring up to several minutes to obtain the required current and voltage data.

A challenge when using wide-band approaches is to ensure that the injected frequencies' energy is high enough at all frequencies to obtain a meaningful converter response. However, merely increasing the signal's energy is not feasible because high disturbances will change the operating point of the converter and violate the small-signal condition of the injection. Several techniques exist to create an injection signal. In [123], current impulses were injected to create the signal. For estimating the grid impedance of systems with a low-noise ratio, the maximum-length binary sequence injection method is proposed, while for strong grids with a high-noise condition, the discrete-interval binary sequence is recommended by the authors because the power spectrum of the signal can be defined by the user [124, 126].

Choosing between sequential and simultaneous injections is a trade-off between time and accuracy. As a result, the appropriate injection approach depends on the objective and the intended application of the derived impedances.

When accuracy is a priority, and different, potentially unknown effects are of interest, sequential injections are beneficial [96, 107, 127]. Unambiguously mapping the frequency response to a unique perturbation frequency makes it possible to investigate when converters also react to perturbations with frequencies other than the perturbation frequency. Moreover, choosing an appropriate magnitude of the injected perturbation signal is straightforward, and it can be easily verified that the converter operation is not altered.

Lasting less than one second, simultaneous injections are applied if short impedance derivation times are the priority. This is the case when the derived impedances are used for real-time applications requiring a direct frequency response for a range of frequencies. For instance, simultaneous injections are used for adaptive controllers [124, 128, 129] and online stability assessments [125, 130].

Analog to analytical impedance models, numerical impedance models can be derived both in the dq- and sequence domain. However, perturbing a DUT in the positive and negative sequence is more intuitive than the dq-domain, where perturbations have to be aligned with the d-axis and q-axis. The approaches for deriving the impedance models based on voltage and current data in the dq-domain and sequence domain are presented in the following paragraphs. The signal χ can be given by either current or voltage signals.

Sequence Domain

In the sequence domain, the perturbations signals are injected either as currents or voltages in positive and negative sequences in two subsequent simulations or measurements. For each positive- and negative-sequence perturbation, the following steps need to be performed.

1. Transforming the time-domain data in the frequency domain by Fourier analysis gives the magnitude $|\chi(f)|$ and phase angle $\varphi_\chi(f)$ of the voltage and current data for each phase a, b and c and for the injected perturbation frequency f .
2. The sequence transformation in (2.25) is applied on the signal $|\chi(f)|e^{j\varphi_\chi(f)}$.
3. For a positive-sequence perturbation, the positive-sequence component is extracted. For a negative sequence-perturbation, the negative-sequence component is extracted.
4. The positive-sequence impedance $Z^p(f)$ and negative-sequence impedance $Z^n(f)$ is calculated according to (2.26) and (2.27), respectively.

Dq domain

Deriving the impedance model in the dq-domain requires a more complicated approach. While also two different perturbation series need to be conducted: one for d-axis perturbation, one for q-axis perturbation, the injections need to be synchronized with the grid voltage to ensure that the d-axis and q-axis perturbation are orthogonal to each other [95]. This requires tracking of the grid angle by a PLL. Subsequently, the following steps need to be performed for both d-axis and q-axis perturbation:

1. The transformation angle θ needs to be determined for the Park transformation. Choosing θ as the grid angle of the voltage at the converter terminal is beneficial because it provides a common reference for all perturbation series. The angle θ

can be obtained by solving the power flow or evaluating the voltage at the grid frequency using Fourier analysis.

2. Having only real and no complex elements, contrarily to the sequence transformation, the Park transformation can be applied directly to the recorded signals $\chi_{abc}(t)$. to obtain $\chi_d(t)$ for d-axis perturbation and $\chi_q(t)$ for q-axis perturbation.
3. Transforming the dq time-domain signals into the frequency domain gives the magnitude and phase angle of the signals for the perturbation frequency f . For a d-axis perturbation, $\chi_d(f)$, for q-axis perturbation $\chi_q(f)$ is obtained.
4. The impedance matrix as in (2.18) and (2.19) has four unknown elements. As a result, a second set of linear independent perturbation series is required to solve the system. This can be done, for instance, by rotating the perturbation vector of the d-axis and q-axis injections and solving the resulting equation system as described in [131].

Conclusion

Numerically determining the converter response by injecting perturbations at different frequencies makes it possible to derive the impedance of DUT such as power system converters. Numerically-derived impedances can be used in the same way as analytically-derived impedances to investigate converter control interactions in the frequency domain by impedance-based analyses. Likewise, the numerically-derived impedances can be modeled in dq-domain or sequence domain with analog implications compared to the analytical approach:

- Both dq domain and sequence domain impedances show the response of a converter at specific frequencies. Through the off-diagonal elements in the impedance matrix in the dq domain, couplings between frequencies are included. Still, the single matrix elements do not have any direct physical meanings. In contrast, numerically-derived positive and negative sequence impedances allow for a physical interpretation of the converter response. However, couplings at low frequencies might affect the positive and negative sequence impedance and introduce a dependency of the recorded current and voltage data on the connected grid when the grid's short circuit power is low, and the equivalent grid impedance is high.
- Injecting perturbation signals in the sequence domain requires only changing the phase difference of the three-phase injection signal. Contrarily, in the dq-domain, the injected perturbation needs to be aligned with the grid angle so that d-axis and q-axis perturbations are orthogonal. Because the grid angle needs to be tracked by a PLL, the setup in dq-domain is subjected to an additional source of potential errors for the obtained data.
- When applying the Park transformation to determine the current and voltage data in dq-domain, a common transformation angle needs to be applied for both d-axis

and q-axis injections. In addition, more perturbation series need to be performed to solve for the dq-impedance matrix elements. In return, the positive- and negative-sequence impedances can be obtained by only applying the sequence transformation on the obtained current, and voltage data of the positive- and negative-sequence perturbed system after the signals are transformed in the frequency domain.

2.4 Method Approach

As discussed in section 2.2.5 and shown in Table 2.2, the impedance-based analysis is beneficial and offers several analysis possibilities to investigate converter control interactions. Therefore, the impedance-based approach is used in this thesis to investigate the stability of converter-dominated AC systems in the frequency domain. Relying on modeling the system as impedances, different approaches to derive the impedance models were pointed out in the previous sections. The chosen derivation approach, the selected domain, and the implications are discussed with respect to the following thesis's objectives:

- An analysis method needs to be able to investigate the system when different parties provide different parts. For instance, a manufacturer provides a converter model, and a transmission system operator knows the transmission system's frequency behavior at the connection point.
- The impedance modeling method should be able to determine the impedance when converter models are provided as black-boxed time-domain models to protect the intellectual property rights, and no knowledge of the control systems is given.
- The method needs to be able to facilitate the investigation of different control systems and parameter variations.
- The stability of the system should be assessed with high confidence.

2.4.1 Choice of Domain

Impedance models can be derived in dq-domain or sequence domain to represent the three-phase system. Whereas in dq-domain coupling effects are represented, the sequence domain models offer a higher simplicity in modeling as well as analyzing the system. In particular, modeling in the dq-domain requires a common reference angle, making it difficult to utilize models provided by two different parties. In practice, choosing a common reference angle is hardly possible, and also aligning both models on the terminal voltage angle might result in inaccurate models [132]. Hence, in this thesis, positive- and negative-sequence impedance models are used. This allows for assessing the stability of systems where two different parties provide the subsystems' impedance models, i.e., the converter and grid model. However, the sequence domain's simplicity comes with the drawback that coupling effects need to be carefully monitored and possibly included in the models separately for an accurate stability assessment. [73, 118, 133].

2.4.2 Choice of Impedance Derivation Approach

The impedance models can be obtained either analytically or numerically. Analytical models derive the frequency response based on the converter control system. Contrary, numerical models inject a small-signal voltage or current perturbation to analyze the frequency response. While analytical models require the converter control system's exact knowledge, a numerical approach can also be conducted for black-boxed systems. In addition, analytical models cannot include the control system of converters as a whole, which might diminish the model accuracy when not investigated thoroughly. Consequently, in this thesis, a numerical approach is chosen to obtain the impedance models. Thus, impedance models of black-boxed systems can be obtained, and the frequency response always represents one of the actual converters. Because the system does not need to be simplified in a modeling process, numerically derived models guarantee high confidence in the stability assessment. Additionally, the numerical approach can be conducted for simulated time-domain models of converters and also for physical, laboratory scaled MMCs. Deriving converter models through a laboratory test bench allows to include phenomena in the impedance models that might otherwise be neglected in time-domain models. Nevertheless, numerical models are only valid for a given system, so that a numerical approach necessitates determining impedances for various converter settings and operation modes.

2.4.3 Method Requirement

With respect to the thesis' objective, a numerical approach to model the converter as positive- and negative-sequence impedances is chosen. Yet, the question of how the perturbation series are conducted and how the current and voltage data are evaluated remains open. In this section, the requirements for a method to derive the impedance models numerically in the sequence domain are discussed.

With respect to the thesis objectives, the accuracy of the derived impedances should be prioritized over speed. Consequently, a sequential approach is beneficial, ensuring the highest confidence for the impedance models. However, as pointed out, numerous perturbation series need to be conducted to represent the converter behavior accurately. An accurate converter model requires high-resolution frequency data, meaning that perturbation series needs to be conducted for a large number of perturbation frequencies. While simply recording voltage and current signals for further post-processing such as Fast Fourier Transform (FFT) will result in accurate frequency-dependent impedance models, this approach is not feasible when numerous perturbation series are conducted. Transforming the voltage and current time-domain data into the frequency domain in post-processing is hardly possible because the required high sampling rate of the data results in file size, which cannot be handled in practice. In particular, when using RTS and laboratory converters, the data transfer in real-time is critical and constitutes a bottleneck for the number of signals and sampling time. As a result, a numerical approach

is required that guarantees high data accuracy while also permitting a large number of perturbation series.

Because signals having frequencies with $f = 9\text{ kHz}$ need to be evaluated [42], the time step of the method needs to be sufficiently small so that the inherent information can be sampled at discrete times. While the Nyquist-Shannon sampling theorem demands a sampling rate that is equal or greater than $2f$, this applies only for perfectly band-limited signals, which is not the case for actual systems [134]. This requirement is not problematic when using offline simulations because the time step can be decreased sufficiently since only the simulation time is affected by a lower time step. However, to answer the second research question, the MMC Test Bench (TB) is used, and the method needs to run in real-time to be coupled with the laboratory equipment. Preliminary tests have shown that models run on the Central Processing Unit (CPU) of the RTS with time steps between $10\text{ }\mu\text{s}$ and $40\text{ }\mu\text{s}$ without overruns, meaning that each time step can be executed in real-time [135]. Because this might not be small enough, in particular when measurements containing noise are evaluated, the method needs to be implemented on the Field Programmable Gate Array (FPGA) board of the RTS to facilitate sufficiently small time steps [135].

2.5 Investigation Approach

Based on the analysis and the methodical approach concluded to investigate converter control interactions, the following approach is taken to achieve the objectives defined in chapter 1.6. Figure 2.15 illustrates the steps that need to be taken to answer the research questions posed in chapter 1.5.

The basis of the investigations are the methods that are developed in this work to derive and measure the MMC impedances. As pointed out previously, a CPU-based method can be used for deriving the impedance models based on the MMC time-domain models. In contrast, the MMC TB impedance models are derived using the FPGA implementation of the method. Subsequently, the impedance models are used to predict stable and unstable scenarios in the frequency domain. For this purpose, scenarios are defined for the MMC operation, such as defining varying operating points for the active and reactive power flow and different control settings. These scenarios are then used to operate the laboratory-scaled MMC, the MMC TB and simulate the time-domain models of the MMC using EMT simulations. Additionally, varying the short-circuit power defines the AC grid scenarios. The MMC scenarios and the AC grid scenarios define the grid situations investigated in this thesis.

Perturbing a converter and evaluating the current and voltage data is the basis for the time-domain models' and MMC TB's impedance representation for each scenario. The derived impedances include frequency coupling effects that are modeled dependent on the AC grid scenario. The AC grid impedance is modeled according to the previously defined AC grid scenarios. Moreover, the laboratory-scaled impedances of the MMC TB are

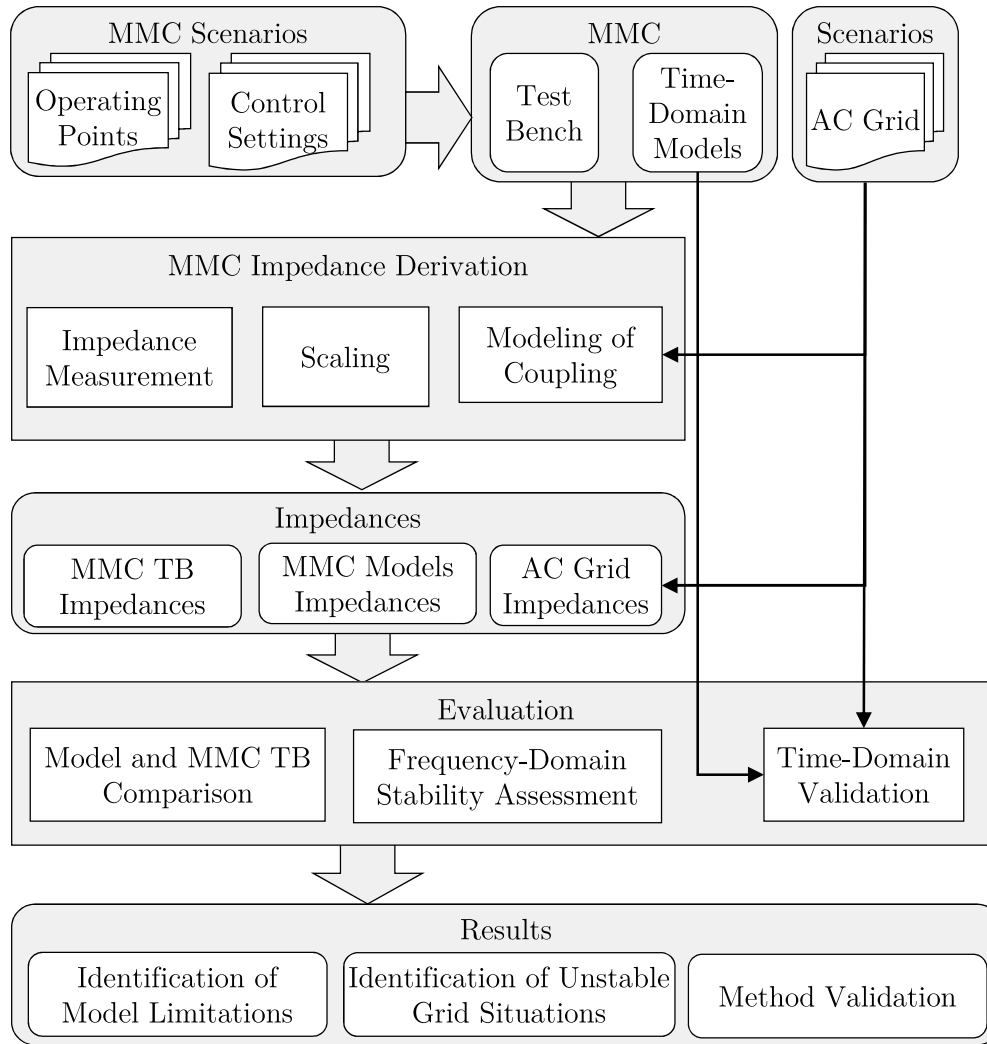


Figure 2.15: Investigation approach.

scaled up so that they can be applied to full-scale system studies. The developed methods demonstrate how converter control interactions that threaten the system stability can be investigated when manufacturers provide converter controllers only as encrypted black-boxed systems such as control replicas.

Evaluating the resulting MMC TB, MMC model and AC grid impedances is the basis for answering the second and the third research question that are posed in this thesis. The MMC model and MMC TB impedances are compared to identify potential limitations of the applied MMC models. Hence, it can be concluded if simplified models of a converter's electrical structure are adequate to investigate converter control interactions.

The impedances of the MMC models for each scenario and those of the AC grid form the basis for the frequency-domain stability prediction. By applying the IbSC, the unstable cases within the defined scenarios are identified, and it can be concluded in what grid situations converter control interactions threaten the system stability. By simulating particular scenarios through EMT tools in the time domain and comparing the results

to the stability predictions in the frequency-domain, the stability assessment approach is further validated.

The impedance derivation method is validated based on the deviations of the derived impedance magnitudes with those of a standard FFT algorithm. It is evaluated according to the required data file sizes, which impact the applicability of the method. The criterion for assessing the stability is the encirclement of the $(-1, j0)$ point by the Nyquist plots. An indication of how close a system is to instability is defined by the PMs that are derived using Bode plots. The criteria for evaluating the model and MMC TB is the proximity of the impedances in different frequency regions. The impedance-based stability analysis is validated if a stable and unstable case can both be replicated with time-domain simulations.

3 MMC Test Bench and EMT Models

This chapter introduces the MMC TB and the EMT models that are the basis for the subsequently-derived impedance models.

3.1 MMC Test Bench Laboratory

The MMC TB laboratory consists of laboratory-scaled, physical MMC racks that are connected through Power Amplifiers (PAs) with RTSs on the AC side of the converter. The RTSs contain a CPU and an FPGA board to simulate models and grids in real-time. Whereas on the CPU Simulink models can be applied with minor modifications, modeling on the FPGA requires more machine-oriented implementations. However, FPGAs facilitate a high degree of parallelization and very low time steps in the range of nanoseconds. While simulating the AC grid signals on the RTSs, the DC grid can be built up using inductive, resistive, and capacitive components emulating DC cables. Figure 3.1 illustrates the laboratory setup of the MMC TB laboratory.

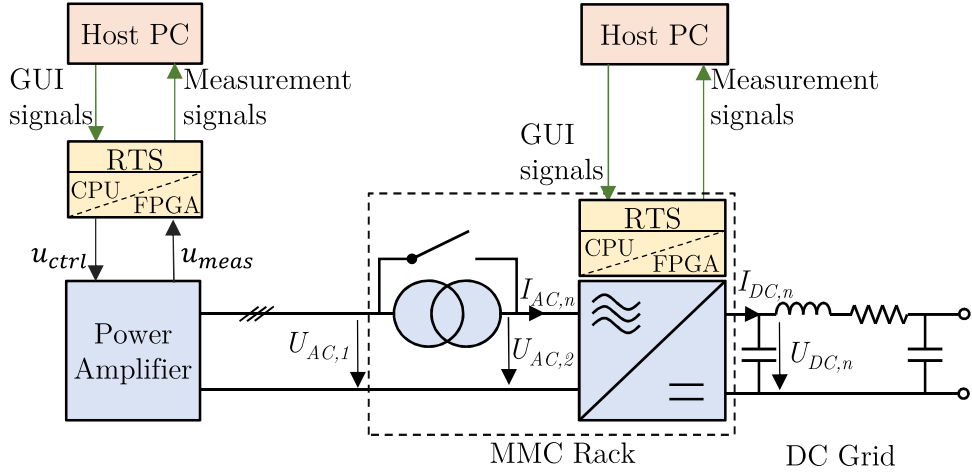


Figure 3.1: MMC TB setup.

Throughout this and the following chapter, as in Figure 3.1, the parts of the models and methods running on the host Personal Computer (PC) are indicated by orange, the parts running on the RTS by yellow and physical components or, if simulated, electrical models are indicated by blue. Two host PCs communicate via Ethernet with the RTSs so that Graphical User Interface (GUI) signals, such as simulation settings, can be sent to the simulators, and measured signals can be received and viewed on the host PC. Exchanging data asynchronously with the RTSs, no major calculations are performed on the host PC. The AC side RTS communicates with the PA by means of analogue voltage signals between -10 V and 10 V . The control voltage signal u_{ctrl} is used to either set a

specified AC voltage U_{AC} or AC current I_{AC} at the terminal of the PA. The RTS of the MMC rack directly connects with the converter through its FPGA board and contains the converter's control system. Figure 3.2 depicts the PA (a) and the MMC rack (b) of the MMC TB laboratory. It should be noted that, in this thesis, no loop is closed within the AC side RTS, meaning that the signals obtained through u_{meas} are not used as inputs for determining signals that are sent back to the PA via u_{ctrl} . Otherwise, a PHIL system is created that requires additional interface algorithms [81], affecting the system's stability and, therefore, possibly distorts stability predictions.

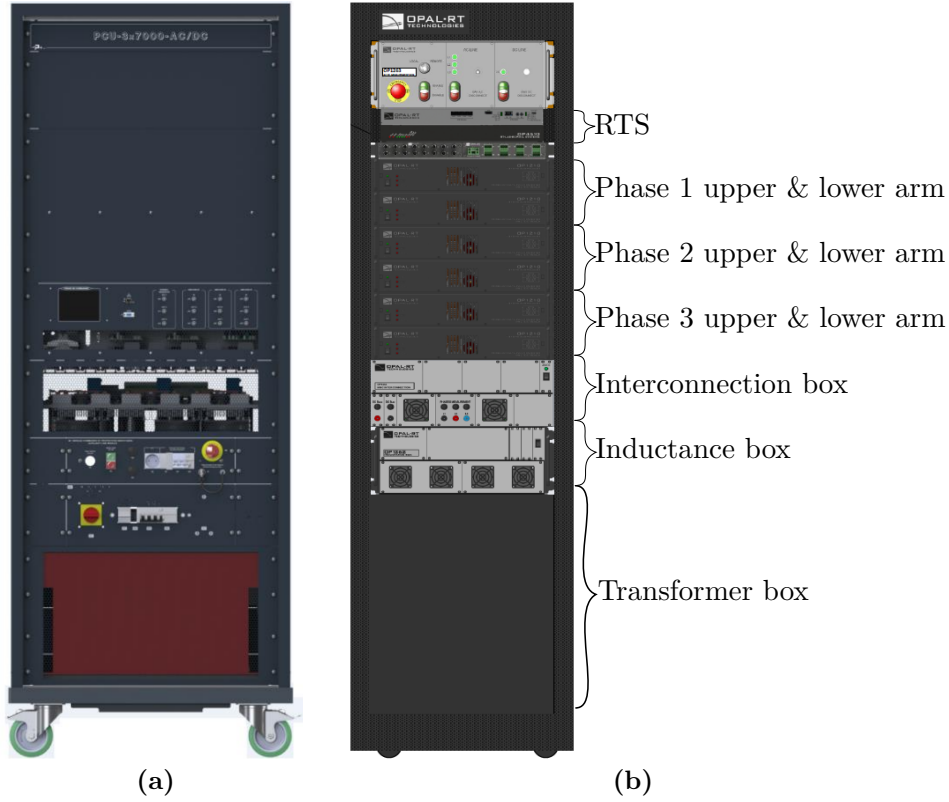


Figure 3.2: Power Amplifier (a) [136] and MMC rack (b) [137].

3.1.1 Power Amplifier

Based on linear technology, the 4-quadrant power amplifier utilized in the MMC TB is able to regulate the voltage, current, and frequency while operating in all four quadrants by absorbing or generating power. The electrical parameters are given in Table 3.1.

The specified bandwidth is verified by additional measurements shown in A.1 and Figure A.1.

Table 3.1: Power amplifier parameters.

Model	PCU-3x7000-AB
Power	21 kVA (3ph) - 3x 7 kVA
Maximum Peak Voltage	400 V or 270 V (RMS)
Maximum Peak Current	18 A or 28 A (RMS)
Voltage distortion at full output power	Typical: < 0.03 % / Max.: < 0.07 %
Bandwidth: Full scale	DC – 5 kHz
Bandwidth: Small signals at -3 dB	25 kHz
Noise – max. RMS	0.02 % of rated voltage
Noise – max. peak to peak	0.3 % of rated voltage
Time response	< 20 μ s
Voltage and current measurement	Included
100 % absorption	Included

3.1.2 MMC Rack

As indicated in Figure 3.1, the MMC rack is comprised of a lab-scaled MMC, an RTS controlling the converter and a transformer that can be bypassed. While Table 3.2 shows the specified parameters, Figure 3.2 (b) illustrates how the converter rack is built up. Consisting of 60 SMs in total, the converter utilizes one SM-box per phase and per arm, having ten SMs each and a nominal cell voltage of 40 V. The converter uses Metal–Oxide–Semiconductor Field-Effect Transistors (MOSFETs) for switching instead of IGBTs due to their lower resistance, ensuring that the overall converter resistance

Table 3.2: MMC Test Bench parameters.

Parameter	Variable	Value
Nominal output power	$P_{n,MMC}$	6 kW
Nominal DC voltage	$U_{DC,n}$	400 V
Nominal DC current	$I_{DC,n}$	15 A
Nominal frequency	f_n	50 Hz
Nominal AC primary voltage (3-phase)	$U_{AC,1}$	400 V (Line-to-line RMS)
Nominal AC secondary voltage (3-phase)	$U_{AC,2}$	208 V (Line-to-line RMS)
Nominal AC RMS current at f_n	$I_{AC,n}$	16.7 A
MOSFET switching frequency	f_{sw}	0 -10 kHz
Number of cells (submodules)	n_{SM}	10
Nominal cell voltage	$U_{cell,n}$	40 V
Cell capacitor	C_{SM}	4.92 mF
Arm inductor	L_{Arm}	2.5 mH
Transformer rated power	S_{Tr}	8 kVA
Transformer vector group		Y/ Δ (Ynd11)

matches the system's rated power and voltage level. The arm inductances are placed in the inductance box above the transformer, connecting the converter with the PA system. Because the AC voltage can be set to 208 V instead of 400 V by the RTS, the transformer can also be bypassed if the MMC rack is connected to the PA and not the household power grid.

The control system of the MMC rack is divided into the lower-level controls that are implemented on the FPGA board of the RTS and the upper-level controls that run on the CPU of the RTS using the modulation index as the interface. Whereas the lower-level controls generate the firing signals for the SM units and are specific to the converter, the upper-level controls are implemented in the OPAL RT-LAB environment on the basis of Matlab Simulink models. Thus, different upper-level controls, as well as black-box control systems, can be applied to the MMC TB.

3.2 Electromagnetic Transient Time-Domain Models

3.2.1 MMC Electrical Model

The electrical models of the MMC are developed on the basis of a Type 5 average value model. Type 5 models replicate the average response of the MMC's SMs and switching devices by controlled current or voltages sources [27]. Figure 3.3 illustrates the time-domain model used in this thesis. The upper and lower arms are represented by controlled voltage sources, and the lumped arm inductances L_{Arm} and resistances R_{Arm} that represent the ohmic losses of the arm inductances and the MMC's SMs. Each controlled voltage source models a string of SMs as it can be seen in Figure 3.3 for the upper arm and phase c . The SM represent the semiconductors that are switched to operate the converter. On the AC side, the converter model is comprised of a transformer that decouples the MMC from the AC grid and transforms the AC voltage to the required amplitude. While the model can also be used to build a bipolar converter station, it is used to analyze a monopolar representation of the converter in this thesis. Thus, only a single converter is used. The electrical model can be parameterized to represent two different systems, namely a digital twin of the MMC TB, the Laboratory-Scaled (LS) model, and a representation of an HVDC MMC, the Full-Scale (FS) model.

The MMC control system described in section 3.3 can be applied to simulate both MMC models and the MMC TB because it is independent of the system's power and voltage ratings. The simulated LS and FS models are implemented in MATLAB/Simulink® with the addition of the OPAL-RT block set for real-time simulation [138]. The RTSs are also used for simulating the MMC models, although they are not coupled with physical components. This is due to consistent and comparable modeling and potentially shorter simulation times. The parameters of the simulated models are given in Table 3.3. The FS parameters are set to model an HVDC system, whereas the MMC TB specifications

represent a down-scaled HVDC system. The LS model's parameters are chosen according to the MMC TB's specifications to match simulations and laboratory tests.

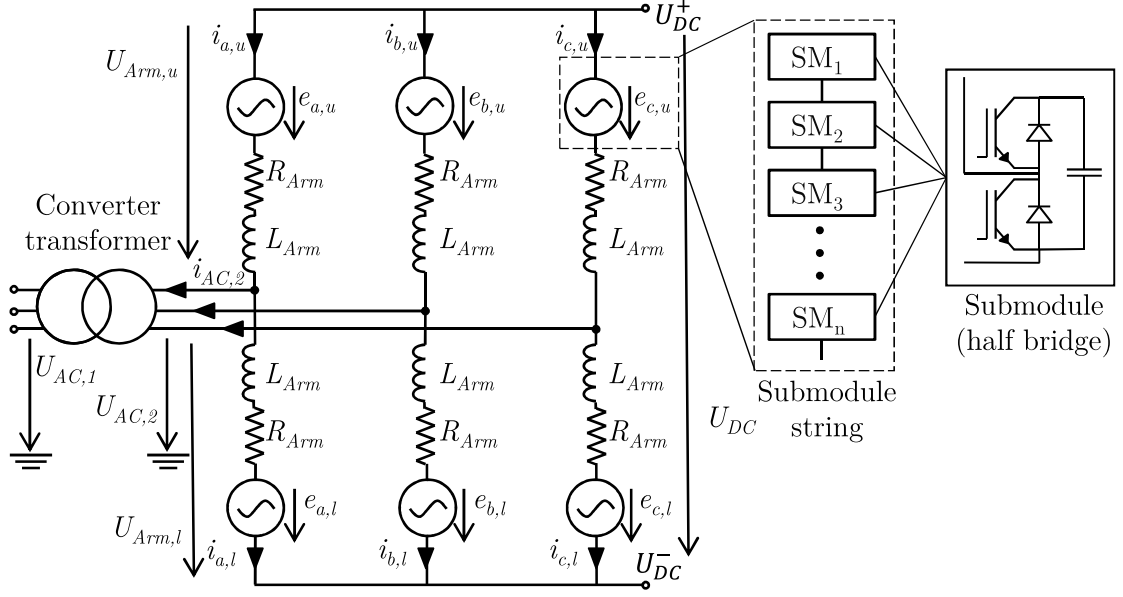


Figure 3.3: Type 5 model representation of the electrical structure of an MMC.

Table 3.3: Full-scale and lab-scale model parameters.

Parameter	Variable	Lab-scale model	Full-scale model
Converter power	$P_{n,MMC}$	6 kW	1200 MW
DC voltage	$U_{DC,n}$	400 V	640 kV
Arm inductance	L_{Arm}	2.5 mH	46.2 mH
Resistance arm inductance	R_{Arm}	0.013 Ω	0.08 Ω
On resistance	$R_{SM,on}$	0.925 m Ω	0.08 Ω
Number of SMs	n_{SM}	10	350
SM capacity	C_{SM}	4.92 mF	8.8 mF
AC primary voltage	$U_{AC,1}$	400 V	400 kV
AC secondary voltage	$U_{AC,2}$	208 V	350 kV
Transformer reactance	$R_{AC,1}$	0.04 Ω	0.316 21 Ω
Transformer inductance	$L_{AC,1}$	0.219 mH	0.403 mH
Transformer reactance	$R_{AC,2}$	0.032 Ω	0.726 Ω
Transformer inductance	$L_{AC,2}$	0.177 65 mH	0.924 73 mH

3.2.2 AC Grid Model

The AC grid is modeled according to Thévenin's theorem as a two-pole with a voltage source in series to an internal impedance, as shown in Figure 3.4. Two independent

measurements of the two-pole could determine the internal impedance. However, in this thesis, the impedance is derived deterministically using the grid voltage U_{grid} and an assumed short-circuit power S_{SC} as input. The parameters of the internal impedance with

$$Z_{grid} = R_{grid} + j2\pi f_{grid}L_{grid} \quad (3.1)$$

are determined to

$$R_{grid} = \frac{U_{grid}^2}{S_{SC} \cdot \sqrt{1 + (X/R)^2}} \quad (3.2)$$

and

$$L_{grid} = \frac{R_{grid} \cdot X/R}{2\pi f_{grid}}, \quad (3.3)$$

where the grid voltage is given as line-to-line RMS voltage and a constant $X/R = 10$ is assumed. The model resembles an electrical-short overhead line in a transmission system, allowing for the assumption that the leakage current and the line capacitance can be neglected [139].

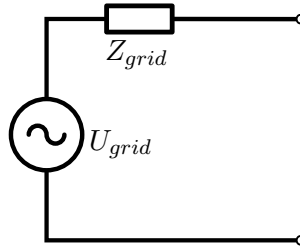


Figure 3.4: Equivalent circuit diagram of AC grid model.

The model allows for the study of interactions between converters and the AC grid when the impact of the grid strength is investigated [41]. In this context, the Thévenin impedance results from a defined Short-Circuit Ratio (SCR) that expresses the AC grid strength relative to the nominal converter power $P_{n,MMC}$ as follows [47, 140]:

$$SCR = \frac{S_{SC}}{P_{n,MMC}} = \frac{\frac{U_{grid}^2}{|Z_{grid}|}}{P_{n,MMC}} \quad (3.4)$$

A qualitative classification of the SCR is given in Table 3.4 [47].

Table 3.4: SCR classification.

Range	Classification
$SCR \geq 5$	High
$3 < SCR < 5$	Moderate
$SCR \leq 3$	Low
$SCR \leq 2$	Very low

3.3 MMC Control System

A cascaded vector control scheme is used to represent the converter control system, controlling both the simulated MMC time-domain models and the MMC TB. As stated in chapter 1, in this thesis, the objective is to be able to investigate interactions independent of specific control systems that are only developed for the purpose of stability investigations. Consequently, a preexisting control system, not developed solely for interaction studies, is employed because converter control replica systems are not available for the thesis. The control scheme is a proven concept for VSCs, in particular MMCs, and guarantees a stable and dynamic operation of the converter [27, 141–143]. It provides a hierarchy of nested controllers that translate the reference values from the dispatch control to switching states that control the semiconductor switches' firing angles.

The control system has been developed and tested in previously-published literature [144, 145]. Furthermore, it has been tested in several industry-related projects and further developed to provide fault ride-through capability of the converter and in AC and DC fault situations [145, 146]. It is comprised of several slow controllers (e.g., outer control loops for active and reactive power and voltage references), as well as fast controllers (e.g., SM sorting algorithms for Nearest-Level Modulation (NLM)).

Figure 3.5 provides a high-level overview of the vector control scheme that forms the basis for the applied control system. In this thesis, the control system is assumed to be a black box because it is unknown in reality to the converter operator. Hence, it is not tuned or enhanced to mitigate potential interactions. However, the converter operator

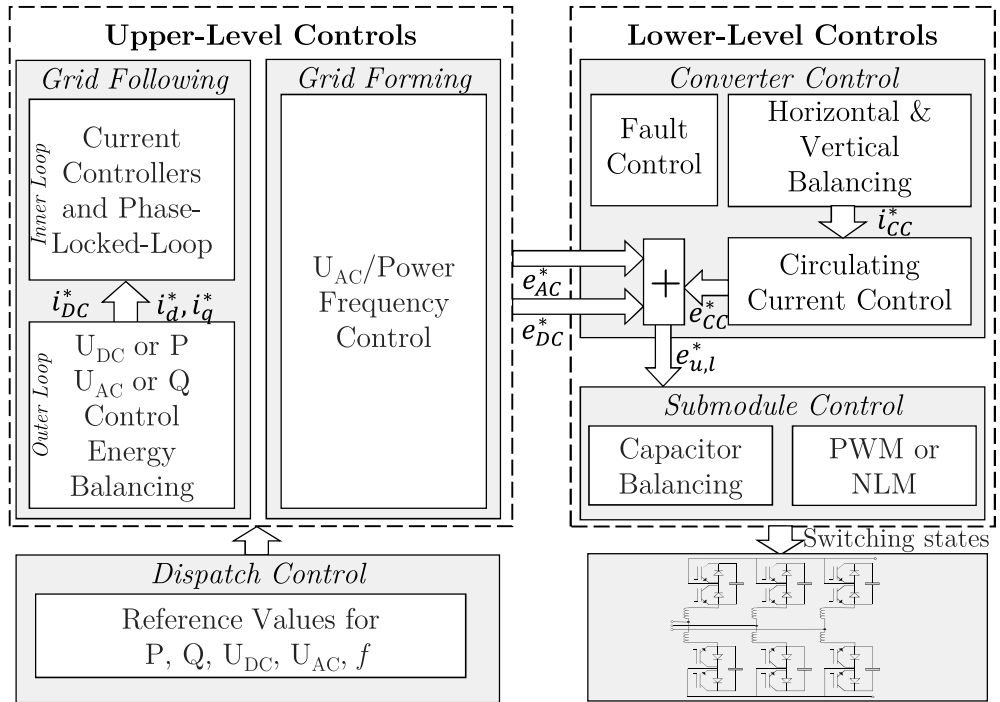


Figure 3.5: High-level overview of the MMC control system.

can switch between two upper-level controllers, serving different purposes that this thesis investigates. When operating in grid-following control mode, the converter can be chosen to either control the DC voltage or active power and the AC voltage or reactive power. In grid-forming control mode, the converter directly defines the AC voltage amplitude and frequency. While the grid-following control mode requires additional current controllers and a PLL to synchronize the converter to the AC voltage, the grid-forming control directly defines grid frequency and angle.

In grid-following control mode, the upper-level controls' outer loop provides the reference values i_d^* and i_q^* that can be independently controlled. As a result, a converter can regulate two different signals. Depending on the role of the converter, i_d^* is defined by the U_{DC} or P controller and i_q^* by the U_{AC} or Q controller. In a point-to-point connection of two grid-following converters, one converter usually controls the DC voltage, whereas the other converter controls the active power flow. Both converters can either control the AC voltage or reactive power flow at the same time. The energy balancing ensures a stable DC voltage by controlling the average capacitor voltage from the DC or AC side. It provides the reference current i_{DC}^* to the current controller. The current controls of the inner loop then regulate the provided reference currents according to the measured AC currents. Typically, PI controllers are used to control the AC and DC grid measurements according to reference values [27, 147]. Because PI controllers are not able to directly control oscillating quantities, the three-phase signals need to be transformed to stationary vectors in a dq-frame rotating at an angular frequency ω_{grid} [147, 148]. Thus, the Park transformation is applied, resulting in additional non-linear behavior of the controllers [25]. A PLL tracks the grid frequency f_{grid} , respectively ω_{grid} , that is used as input to determine the angular frequency ω required for the Park transformation.

Contrarily, the grid-forming controllers employed in this thesis directly define the grid frequency in addition to the AC voltage magnitude and do not require tracking of the grid frequency. Instead, an oscillator at fixed angular frequency replaces the PLL and generates the grid frequency [149]. Converters operating in grid-forming control mode are typically deployed in weak AC grids, such as offshore collector grids where the grid frequency does not result from the rotating masses of synchronous machines.

Both grid-following and forming controllers provide the reference arm voltage e_{AC}^* and e_{DC}^* to the converter control of the lower-level controls. While the upper-level controls guarantee that the converter regulates a desired power flow and couple the converter with the AC grid, the lower-level controls ensure a dynamic, but also stable, operation of the converter [27].

The horizontal and vertical balancing control system controls the average energy stored in the MMC and the SMs. Therefore, it ensures that the average AC power output is equal to the DC power output over each period and is symmetrically distributed among the MMC arms [142, 150]. Referred to as horizontal balancing, the control systems ensure that each converter phase arm is subjected to an equal amount of energy and that the average capacitor voltage is within its limits. Using PI controllers, the energy is balanced

using either the AC or DC power as input [145]. Based on the reference current i_{CC}^* , the circulating current controller suppresses undesired currents flowing between the converter phase legs due to voltage imbalances in the converter arms of the MMC. The spectrum of the circulating currents also consists of second and third harmonics in addition to the DC portion and the fundamental frequency. Therefore, additional resonant controllers are used to suppress these currents and reduce the resulting losses [27, 141]. Moreover, the converter control also comprises additional functionalities for controlling fault currents [145, 151]. The sum of the reference voltages e_{AC}^* , e_{DC}^* and e_{CC}^* is then used to provide the reference voltage $e_{u,l}^*$ for the upper and lower arm of the MMC.

Finally, the control cascade consists of the submodule control that determines the semiconductor switches' switching states. Depending on the reference voltage $e_{u,l}^*$ for the upper and lower arm voltages, the control system decides which SM is inserted or bypassed [152]. Two modulation techniques, PWM and NLM can be applied for this purpose. However, with respect to MMCs, NLM is preferable and implemented in the applied control system due to the large number of SMs and the algorithm's higher efficiency [153]. The capacitor balancing algorithm determines which SMs are switched so that the capacitors of the SMs with the highest voltage are discharged, and those with the lowest voltage are charged. Thus, the energy of the capacitors is evenly distributed.

4 Impedance Model Derivation

This chapter shows the methods developed in this thesis and used for answering the research questions in chapter 5. A numerical impedance derivation method is presented to derive impedance models on the basis of time-domain converter models and the physical laboratory converters of the MMCs. The latter are interfaced with RTSs and host PCs and rely on PAs to provide the current and voltage test signals. Nevertheless, the impedance derivation method itself does not differ when using signals obtained from simulated EMT converter models or physical laboratory converters. Thus, for simplicity, the term impedance measurement will be used both for impedance models derived numerically using EMT simulations and impedance models measured in a laboratory environment based on physical components. Apart from that are impedance measurements of actual, physical HVDC MMC stations, which are not in the scope of this thesis because on-field tests require fundamentally different approaches. Secondly, the derived impedance models are presented.

4.1 Online Impedance Measurement Method

In this section, the method to obtain the frequency-dependent impedance models for stability analyses in the frequency domain is described. As discussed in chapter 1, a measurement-based approach is chosen to fulfill the requirements that a variety of different control systems – including black-boxed time-domain systems – can be investigated. However, a measurement-based approach necessitates new measurements each time another operation point or control mode is investigated. Consequently, many measurement series are required to describe the overall frequency behavior of an MMC. At the same time, computing the impedance in post-processing using FFT algorithms necessitates recording high-resolution voltage and current signals. Because a method needs to be able to derive impedances of laboratory converters and converter control replicas, the limitations of RTSs need to be also considered. Therefore, data sizes are limited further by the data transfer buffer of the RTS. Consequently, in this thesis, an online method is developed that directly determines the impedance during the measurement process, avoiding the need to record high-sampling voltage and current data for subsequent Fourier analyses. The method injects perturbation signals with varying frequency f_{pert} and applies an online Fourier series at the frequencies of interest using the measured voltage and current signals at high sampling for calculating the frequency-dependent impedance. Because the impedance is calculated at the same time the measurements are taken, the output data consist only of the previously calculated impedance, which can be recorded with a lower sampling rate. Furthermore, the method allows the user to investigate the impedance during the measurements directly. The method is implemented in Matlab/Simulink and adapted for real-time simulations with OPAL RT-Lab so that either

simulated time-domain models or physical laboratory components can be investigated [154].

Figure 4.1 gives an overview of the method and shows the implemented building blocks discussed in detail in the following sections. The method is divided into the test circuit and the input and output signal processing parts. The test circuit comprises the grid, an additional perturbation source, and the DUT whose impedance is determined. DUTs can be any physical components and simulated models for which a current flow and voltage can be measured at their terminals when connected to a current or voltage source. With respect to the thesis' objectives, the method is tuned for active components such as MMCs. Depending on the investigated DUT and the grid circuit, the perturbation source consists of a current or voltage source. The signal processing starts with the frequency generator defining a series of perturbation frequencies, $f_{series} = [f_{start}, \dots, f_{end}]$, which are consecutively outputted as perturbation frequencies $f_{pert}(t)$, over time based on the user-defined measurement parameters MP_f , e.g. the frequency range and resolution of interest. The frequency signal f_{pert} and f_{eval} then are fed into the perturbation source and the signal processor, respectively. The perturbation source superimposes a small-signal voltage, $U_{pert}(t)$, at the current perturbation frequency, f_{pert} , on the grid voltage, $U_{grid}(t)$. In case the grid voltage is defined by the DUT, e.g., by a converter operating in grid-forming control, the grid circuit consists of a current source instead, and an additional small-signal current is inserted in parallel to perturb the system. The signal processor evaluates the measurements, $U_{meas}(t)$ and $I_{meas}(t)$, taken at the terminals of the DUT with respect to the frequencies, f_{eval} comprising the current perturbation frequency, $f_{pert}(t)$, and other frequencies of interest. The resulting time-dependent frequency-domain measurement data, $U(f(t))$ and $I(f(t))$, subsequently are used by the impedance calculator to obtain the frequency-dependent impedance, $Z_{DUT}(f)$, that is independent of the time [155].

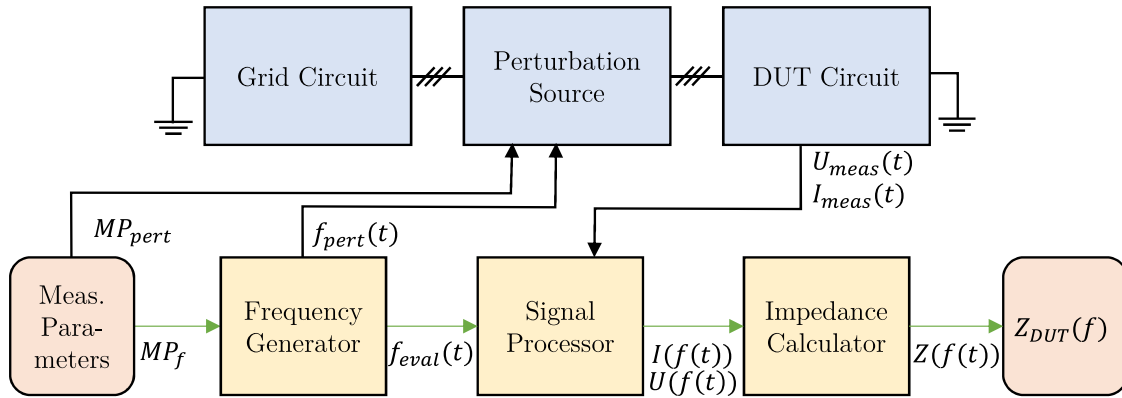


Figure 4.1: Method Overview.

When the frequency response of laboratory converters, such as the MMC TB, is investigated, the method runs on a RTS coupled with the laboratory equipment. In that case, the RTS' computation power limits the minimum time step. Thus, the frequency

generator and the signal processor are implemented on the RTS' FPGA board to allow for a small-enough time step. An FPGA is an integrated circuit containing an array of configurable logic blocks whose interconnects can be programmed after manufacturing for custom designs. It is capable of running with time steps in the range of nanoseconds [156]. However, running the method on the FPGA requires significantly different implementation approaches. Therefore, the following supplemental sections include the developed and implemented FPGA logic for parts of the method.

4.1.1 Frequency Generator

The frequency generator's output is essential to the impedance calculation and serves as input for the perturbation source and signal processor. Initially, the frequency generator creates a frequency series, f_{series} , according to the user-defined minimum and maximum frequencies, f_{min} and f_{max} , the number of frequency steps, n_{fstep} , and the choice of logarithmic or equal distances. The defined series, along with additionally defined parameters such as the measurement duration T_{meas} , is then taken as input for generating the frequency over time signal $f_{pert}(t)$. Figure 4.2 shows the output signal of the frequency generator for an typical frequency series with $f_{series} = [f_1, f_2, f_3]$. For the time, T_{f_1} , T_{f_2} and T_{f_3} , the perturbation source applies a small-signal voltage or current with the respective frequency. Because the frequency signal, f_{pert} , serves as input for every other building block of the method, it also is used as a communication signal. For instance, the change of f_{pert} determines the start of a measurement series and the change of the perturbation signal [155].

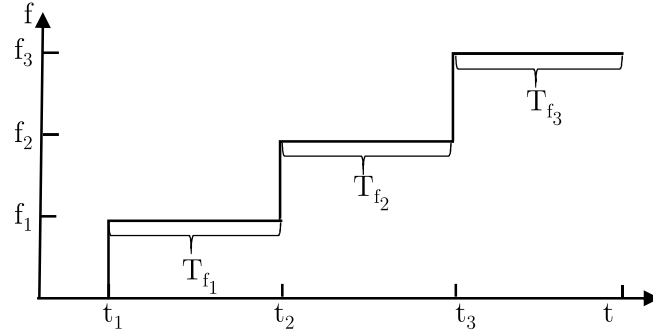


Figure 4.2: Impedance measurement perturbation frequency series.

4.1.2 Perturbation Source

The perturbation source consists of a small-signal voltage or current source with amplitude $|\chi_{pert}|$ whose frequency varies according to the perturbation frequency, f_{pert} , given by the frequency generator. The grid circuit establishes the operating point for the DUT and defines either a voltage or current signal, χ_{grid} , with magnitude, $|\chi_{grid}|$, and

frequency, f_{grid} . The output of the perturbation source is given by

$$\chi(t) = \chi_{grid} + \chi_{pert} = |\chi_{grid}| \cdot \sin(2\pi f_{grid}t + \varphi) + |\chi_{pert}| \cdot \sin(2\pi f_{pert}t + \varphi). \quad (4.1)$$

The perturbation source generates either a positive or negative-sequence signal. The choice of voltage or current source depends on the system under investigation. Figure 4.3 shows two systems with either voltage or current perturbation.

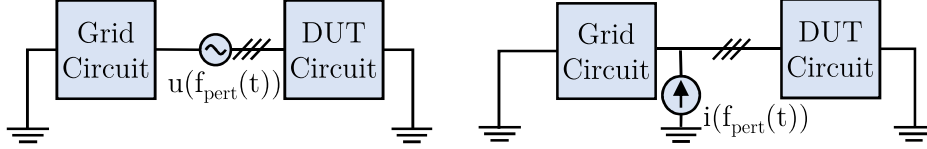


Figure 4.3: Voltage and current perturbation circuit.

For converters, the choice of perturbation source also depends on the converter's control mode. When operating in grid-following control, the adjacent AC grid provides the voltage at the AC terminal of the converter, and a PLL synchronizes the converter with the grid. As a result, the converter can be perturbed with a voltage source that is superimposed on the grid voltage. In that case, the perturbation generator provides both the phase angle φ of the grid voltage and perturbation source. Contrarily, when operating in grid-forming control mode, the converter's control system provides the AC terminal voltage. A perturbation voltage source would influence the converter behavior heavily. Therefore, a current source needs to perturb the converter. In that case, the perturbation generator measures the grid voltage angle φ and applies it to the current perturbation source to have a defined operating point. Compared to the current perturbation, the voltage perturbation is advantageous because the perturbation signal is superimposed on the grid voltage and directly applied at the terminal of the DUT. When injecting a perturbation current, it needs to be ensured that a large enough portion of the perturbation current flows into the DUT and not the grid. However, one should keep in mind that the voltage source approach might not be applicable for higher power ratings, and a current injection needs to be chosen [157, 158]. The amplitude of the perturbation source needs to be small not to affect the operation of the DUT significantly. However, it should be large enough that the measured voltages, $U_{meas}(t)$, and currents, $I_{meas}(t)$, contain a significant frequency spectrum other than the grid frequency of 50 Hz and the signal processor can correctly evaluate the spectrum. Measurements have shown that 5 % of the grid voltage or current is an acceptable trade-off for the perturbation source's magnitude. In that case, the magnitude is high enough that the signal processor can determine the frequency components while not disturbing the converter's operation.

FPGA Design

When the impedance of a physical device is measured, the RTS's FPGA board generates the terminal signal $\chi(t)$ to facilitate smaller time steps and higher resolutions of the perturbation signals at high frequencies. In contrast, the grid signal, χ_{grid} , which employs a lookup table of a sine wave stepped through at constant frequency, the variable frequency nature of the perturbation source, χ_{pert} , necessitates a more elaborate design. The implemented design follows a Direct Digital Synthesis (DDS) using a phase accumulator, phase-to-amplitude conversion, and a Digital-to-Analog (D/A) converter to create a variable frequency sine wave [159]. A standard procedure to control the sine wave's frequency is to adjust the value of the increment the phase accumulator uses to repeatedly step through a lookup table of one period of a sine wave [159]. If a large increment is chosen, the phase accumulator passes many lookup table values, requiring fewer steps to step through and thus, creates a high-frequency sine wave signal. Consequently, stepping through every value in the lookup table determines the lowest frequency signal possible. However, using this approach, the maximum frequency is limited by the available memory required for the lookup table. Because the available memory for one word is restricted on the FPGA, the lookup table approach does not allow for applying very low and very high frequencies.

In this thesis, the phase accumulator is designed to create a sine wave with a fixed number of samples per period independent of the frequency, which can be used in conjunction with the Coordinate Rotation Digital Computer (CORDIC) algorithm. Figure 4.4 demonstrates the components of the implemented DDS sine wave.

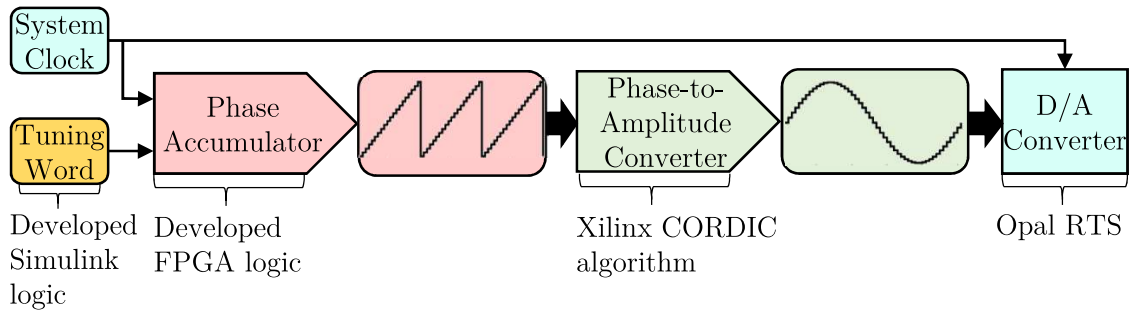


Figure 4.4: Components of the implemented DDS sine wave. Adapted from [159].

The applied hardware sets the system clock and the D/A converter. The system clock of the FPGA board determines how fast it can run and process data. The D/A converter is part of the Opal RTS and outputs the sine wave as analog voltage signal between -10 V and 10 V which is used as a communication channel to the power amplifiers. The tuning word provides the required parameters to set the speed or frequency of the phase accumulator. Using the supplied parameters, the phase accumulator determines the phase argument of the sine wave. The phase argument is a periodic sawtooth signal ranging from $-\pi$ to π , and it is used as input for the phase-to-amplitude converter, which

f_{pert} and $f = 2f_{pert}$. The counter is incremented each time step of the system clock. The output of the counter inc_1 is multiplied subsequently by f_{pert} to yield in inc_2 , which is compared to the reset value $n_{s,rst}$. Whenever inc_1 reaches $n_{s,rst}$, the counter is reset to zero. Consequently, when f_{pert} is doubled, inc_2 reaches the reset value in half the time because inc_2 is doubled as well. Thus, it results in a periodical saw-tooth signal with double the frequency. Multiplying inc_2 by a gain of $\frac{2\pi}{n_{s,max}}$ and subtracting π yields inc_3 , which is scaled as required for the CORDIC algorithm.

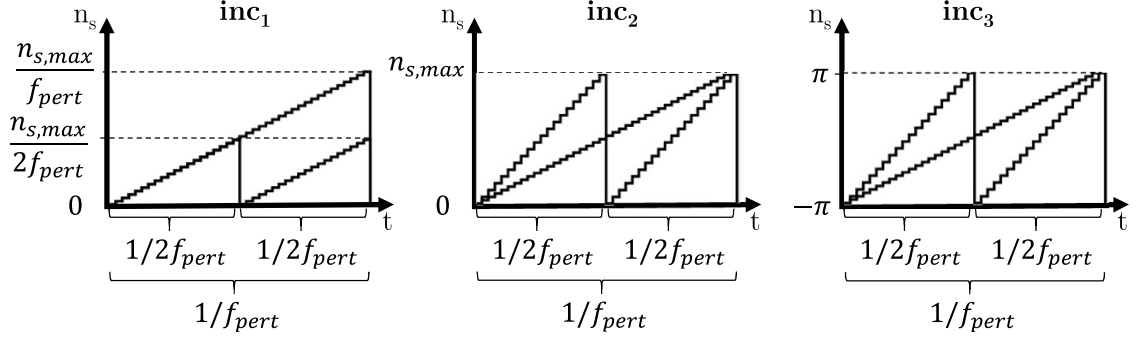


Figure 4.6: Phase accumulator increments for f_{pert} and $2f_{pert}$.

Figure 4.7 depicts a 1 kHz sine wave created by Simulink (green) running on the CPU with $\Delta t_{CPU} = 40 \mu s$ and by the implemented FPGA design (yellow) running at $\Delta t_{FPGA} = 5 ns$.

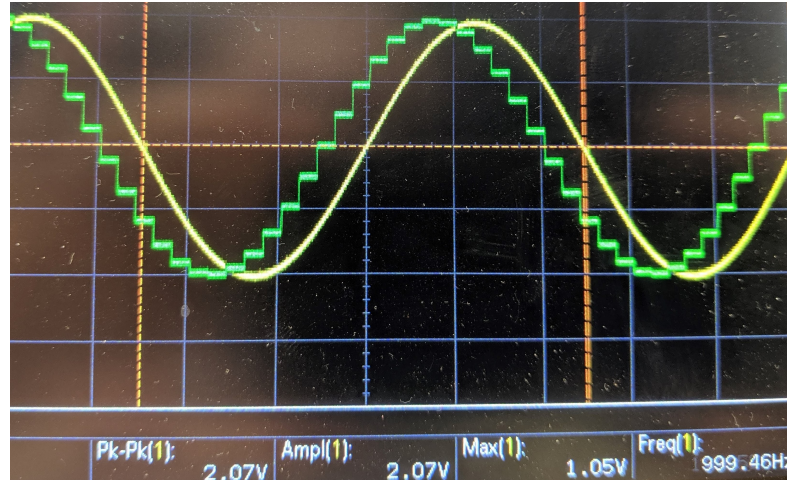


Figure 4.7: 1 kHz sine wave created by CPU and FPGA.

4.1.3 Signal Processor

The signal processor's task is to evaluate the measured voltages and currents for the currently injected perturbation with $f = f_{pert}(t)$ and other frequencies of interest. Using the Discrete Fourier Transform (DFT), a signal can be represented as the sum of a series

of sines and cosines by splitting the signal into separate frequency components [156]. Accordingly, a periodic time-domain signal $\chi(t)$ can be expressed by a Fourier series where

$$\chi(t) = \frac{1}{2} a_0 + \sum_{n=1}^{\infty} a_n \cos(n \omega t) + b_n \sin(n \omega t) \quad (4.6)$$

with

$$a_n = \frac{2}{T} \int_{t-T}^t s(t) \cos(n \omega t) dt \quad (4.7)$$

and

$$b_n = \frac{2}{T} \int_{t-T}^t s(t) \sin(n \omega t) dt. \quad (4.8)$$

The magnitude $|A_n|$ and angle φ_n of each harmonic n is given by the expression:

$$|A_n| = \sqrt{a_n^2 + b_n^2} \text{ and } \varphi_n = \arctan\left(\frac{a_n}{b_n}\right). \quad (4.9)$$

Hence, determining the magnitude and phase of the signal requires the calculation of a_n and b_n with respect to the measured signal, $\chi(t)$, and the frequencies, f_{eval} , to be evaluated by the signal processor. Accordingly, the signal processor evaluates a_n and b_n for

$$n \omega t = 2\pi f_{eval} t \quad (4.10)$$

with the Fourier fundamental

$$F_{DFT} = \frac{1}{T} \quad (4.11)$$

and

$$f_{eval} = n \cdot F_{DFT}, \text{ where } n \in \mathbb{N}. \quad (4.12)$$

Choosing $F_{DFT} = 1 \text{ Hz}$ allows for evaluating every integer frequency.

The DFT is implemented by means of a running window with width $T = 1/F_{DFT}$ evaluating the signals over time, meaning that at a time, t_{W1} , the output is generated based on the signals at $t = t_{W1}$ and $t = t_{W1} - 1/F_{DFT}$. Figure 4.8 depicts the principle of the implemented DFT, showing the running window W at three different times t_{W1} , t_{W2} and t_{W3} . The overall time during which each perturbation frequency is applied and

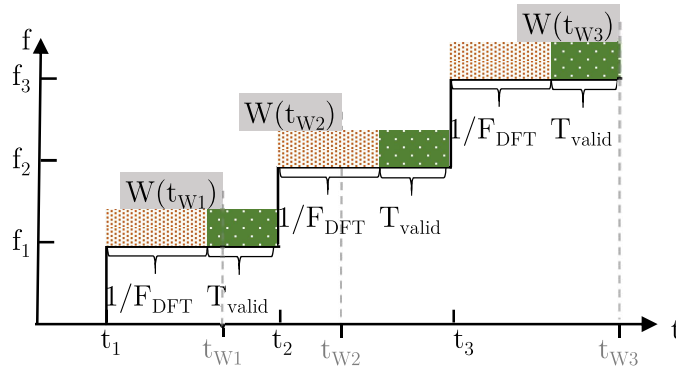


Figure 4.8: DFT window.

the Fourier fundamental, F_{DFT} , determine the valid measuring time, T_{valid} . As a result, the minimum measuring time needs to be larger than the $1/F_{DFT}$ to allow for valid measuring time windows. As can be seen for $W(t_{W1})$ and $W(t_{W3})$, the resulting value of the DFT at $t = t_{W1}$ and $t = t_{W3}$ is calculated based on a window which solely comprises the respective frequencies, f_1 and f_2 . This means that according to (4.7) and (4.8), the product of $\chi(t)$ and $\sin(2\pi f_{1,2}t)$ or $\cos(2\pi f_{1,2}t)$ can be integrated for one period, resulting in valid results. In contrast, at $t = t_{W1}$, the depicted window does not yet contain enough samples of the signal at $f = f_2$ and therefor, no valid results can be obtained.

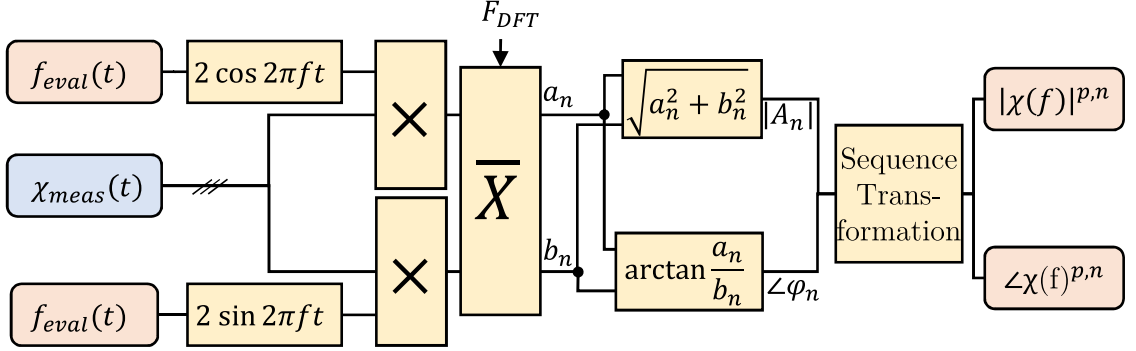


Figure 4.9: Fourier implementation.

Figure 4.9 shows how the signal processor determines the magnitude and angle of the measured voltages, $U_{meas}(t)$, and current, $I_{meas}(t)$, when the DUT is a model and perturbation series are applied in a EMT time-domain simulation. The inputs of the signal processor are the frequencies f_{eval} given by the frequency generator and the measured three-phase voltages and currents, $\chi_{meas}(t)$, at DUT's terminal. Multiplying the sine and cosine of (4.10) with $\chi_{meas}(t)$ gives the inputs for the Fourier integral. The integral is realized by calculating the mean for $t = 1/F_{DFT}$ of the input signals using numerical integration methods [164]. Subsequently, the three-phase voltages' and currents' magnitude and angle are calculated according to (4.9). By applying the sequence transformation matrix

$$T = \frac{1}{3} \begin{bmatrix} 1 & 1 & 1 \\ 1 & \underline{a} & \underline{a}^2 \\ 1 & \underline{a}^2 & \underline{a} \end{bmatrix} \quad (4.13)$$

with complex phasor

$$\underline{a} = e^{j\frac{2}{3}\pi} \quad (4.14)$$

to the three-phase voltage and current phasors, the positive- and negative-sequence components, $|\chi(f)^p| \angle \chi(f)^p$ and $|\chi(f)^n| \angle \chi(f)^n$, along with the zero component of the signals are determined. If the system is perturbed in positive sequence, the positive-sequence components are used for the impedance calculating; for negative-sequence perturbations, the negative-sequence components are used.

FPGA Design

Only the parts of the signal processor that benefit from a low time step are implemented on the FPGA to save its resources. For instance, operations such as using the signals a_n and b_n to calculate the magnitude, $|A_n|$, and angle, φ_n , can be executed on the CPU. In contrast, the integration of the product of the measured voltage and current signals with the perturbation frequency should be calculated on the FPGA to ensure a high sampling rate for the discrete Fourier transformation. Figure 4.10 illustrates the approach to how the signal processor is implemented on the FPGA.

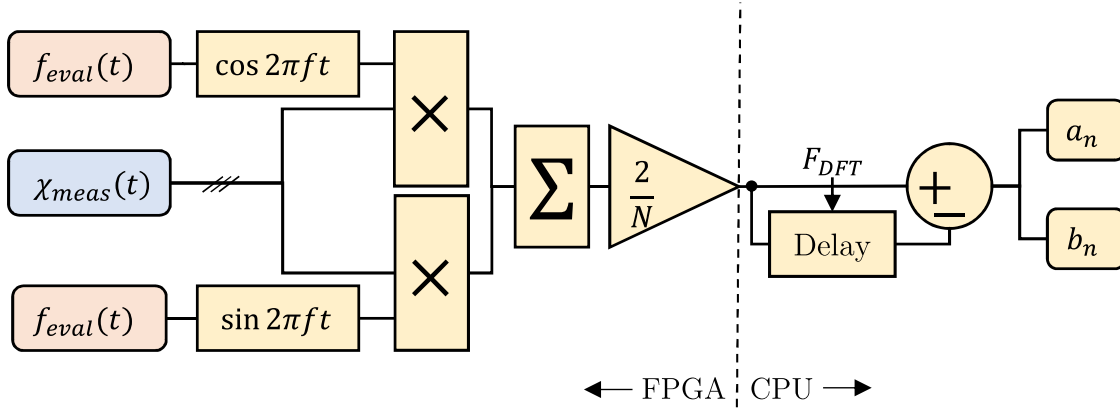


Figure 4.10: FPGA design of signal processor.

The main difference to the CPU implementation is how equations (4.7) and (4.8) are adapted for a discrete system such as the FPGA. Using (4.7) as example, the definite integral can be transformed into Riemann sums so that

$$a_n = \frac{2}{T} \lim_{\Delta t \rightarrow 0} \sum_{m=N}^m \Delta t \cdot \chi(t_m) \cdot \cos(n\omega t_m), \quad (4.15)$$

where parameter, m , is a sequence of integer numbers that represent the number of discrete time steps, Δt , and N is the number of samples within one integration period, T , with $N \cdot \Delta t = T$. Treating Δt as a constant factor, (4.15) can be rewritten to

$$a_n = \frac{2}{N} \lim_{\Delta t \rightarrow 0} \sum_{m=N}^m \chi(t_m) \cdot \cos(n\omega t_m). \quad (4.16)$$

Finally, assuming that $\Delta t = \Delta t_{FPGA} = 5 \text{ ns}$ fulfills $\lim_{\Delta t \rightarrow 0}$

$$a_n = \frac{2}{N} \sum_{m=N}^m s(t_m) \cdot \cos(n\omega t_m). \quad (4.17)$$

Equation (4.17) and its counterpart for obtaining b_n then are used for the implementation of the signal processor. While the calculation of the Riemann sum is realized on the FPGA, the difference resulting from the integration limits is calculated on the CPU using a delay block with $t_{delay} = 1/F_{DFT}$. Thus, a moving window is implemented that

calculates the impedance at each CPU time step for any given time, t , based on the high-sampling calculations on the FPGA at a time, t and $t - t_{delay}$. The moving window makes it possible to monitor the impedance in real-time during the measurements for an applied perturbation frequency series.

On the FPGA, the sine and cosine that use the output of the high-resolution frequency generator are multiplied with the voltage and current measurements, $S_{meas}(t)$, at every clock cycle of the FPGA. Then the products are integrated using an adder-based accumulator that sums up its input at every clock cycle [165]. After scaling the output of the accumulator with $2/N = 2 \cdot 5 \text{ ns}/1 \text{ s} = 10^{-8}$, the output signals are outputted to the CPU of the RTS.

On the CPU, these signals are additionally delayed for $1/F_{DFT} = 1 \text{ s}$ and stored to give the signal at $m - N$ and accordingly at time, $t - T$. The delayed signals then are subtracted from the actual signals at m and accordingly at time, t , to give the coefficients $a_n(t)$ and $b_n(t)$. The coefficients are then processed as written before to determine the impedance, $Z_{DUT}(f)$.

4.1.4 Impedance Calculator

Taking the frequency domain data of the measured voltages U_{meas} and currents I_{meas} , transformed into the frequency domain by the signal processor for the frequencies $f = f_{eval}$, the impedance calculator determines the impedance of the DUT. Evaluating the injected perturbation frequency series gives

$$|Z_{DUT}(f_{pert}(t))| = \frac{|U_{DUT}(f_{pert}(t))|}{|I_{DUT}(f_{pert}(t))|} \quad (4.18)$$

and

$$\angle Z_{DUT}(f_{pert}(t)) = \angle U_{DUT} - \angle I_{DUT}. \quad (4.19)$$

Figure 4.11 shows the measured impedance of an inductive load with $H = 100 \text{ mH}$ for $f_{series} = [1 \text{ Hz}, 10 \text{ Hz}, 100 \text{ Hz}, 1000 \text{ Hz}]$. Because measurements are taken for 3 s for each frequency, the valid measuring range, T_{valid} are the last 2 s of each frequency. Averaging the measured impedance and phase angle over T_{valid} , gives the respective values for each frequency.

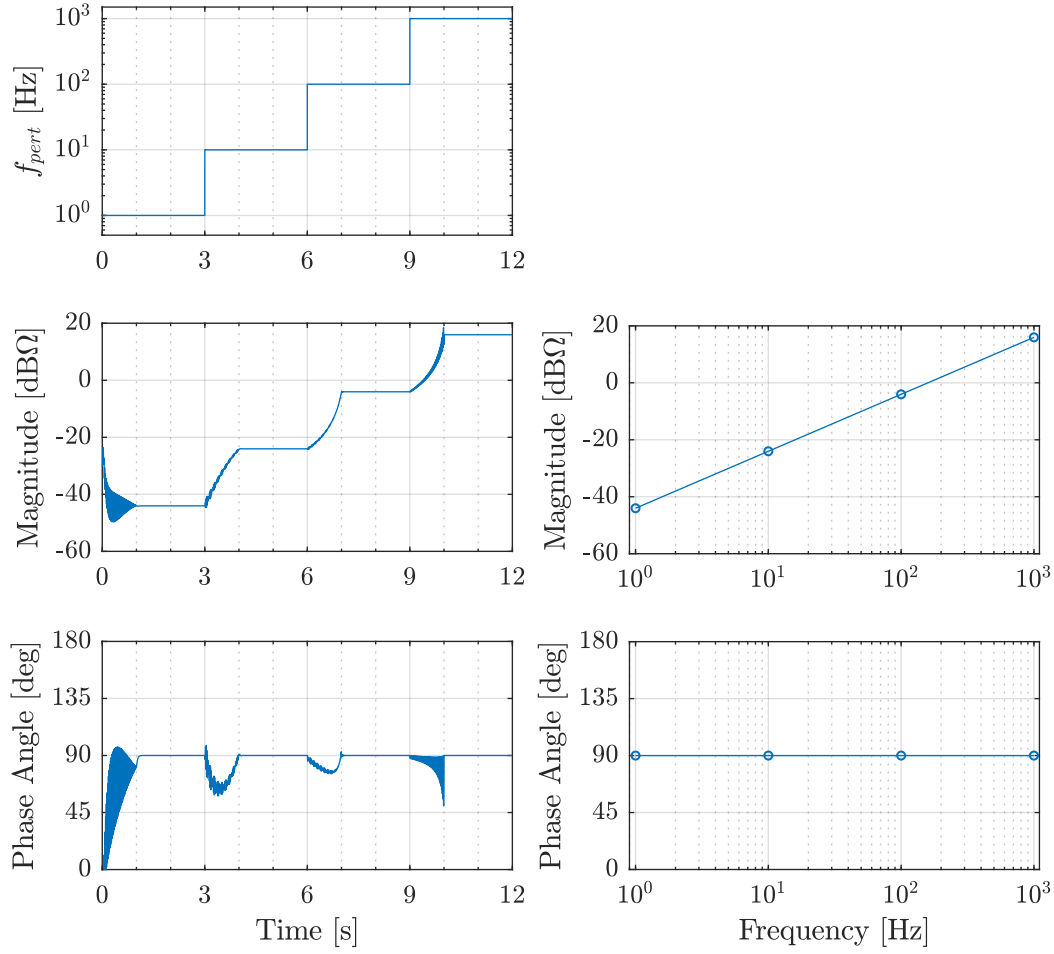


Figure 4.11: Impedance calculation of an exemplary load.

4.1.5 Impedance Measuring: Laboratory

The impedance of the laboratory MMCs is measured using the MMC TB system shown in Figure 4.12. The system is modified using two of the four converter stations and employing the developed measuring method. The impedance measurement setup displayed in Figure 4.13 can be divided into two parts, the test bench and the real-time simulation. The test bench consists of the two MMCs connected physically by inductances and capacitances resembling Π -line sections on the DC side. On the AC side of the converters, the grid voltage is supplied by PAs connected to transformers shifting the grid voltage from 208 V to 400 V as required by the MMCs. In case the transformers are bypassed, the PAs set the grid voltage directly to 400 V. The PAs are coupled with the real-time simulation via analog input and output ports. The ports transmit the control signals, u_{ctrl} to, and the measurement signals, U_{meas} and I_{meas} , from the PAs as voltages between -10 V and 10 V. The control signals comprise the three-phase grid voltages or three-phase currents set at the PAs. While the control signal for PAs 2 consists only of the grid voltage, U_{Grid} , the control signals for PA 1 additionally include the perturbation signal. In case the MMC under investigation operates as a grid-forming converter, PA 1 operates as a

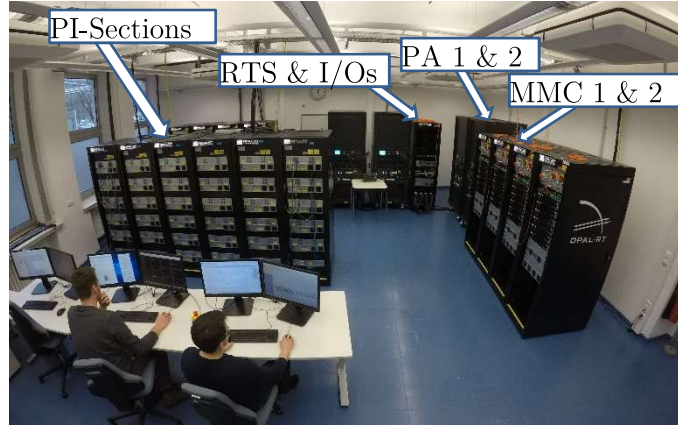


Figure 4.12: MMC Test Bench at RWTH Aachen University.

current source. In both cases, by simulating the signals on the AC side on the RTS, the impedance measuring method can be applied to the test bench. The frequency generator and the signal processor of the impedance measuring method run on the FPGA of the RTS to ensure a sufficiently small time step when high frequencies are evaluated.

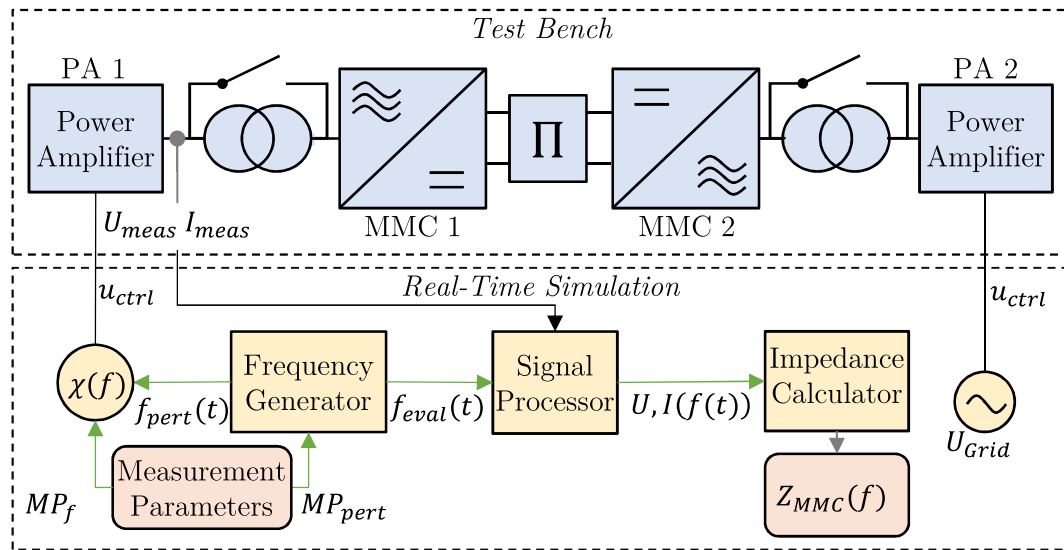


Figure 4.13: MMC Test Bench impedance measurement setup.

Although only the impedance of MMC 1 is measured, the impedance measurement setup takes advantage of the possibility to use an additional MMC. Using two MMCs, the converters' operation mode can be changed while the system is running. For instance, changing the converters' outer control loops from active power control to DC voltage control does not require a shutdown and restart of the overall system. The PAs' can remain an AC voltage source regardless of the connected MMC controlling the active power flow or DC voltage. On the other hand, a system with only one converter would require the second PA to act as a DC voltage source or a DC load, dependent on the converter's operation mode for defining an operation point on the DC side.

4.1.6 Impedance Measuring: Simulation

An impedance setup equivalent to the MMC TB impedance measurement setup is implemented in the EMT time-domain simulation toolbox Matlab/Simulink to achieve high comparability. Figure 4.14 depicts the setup using the simulated MMC models instead of the MMC TB. The MMCs can be either represented by the LS or FS model, dependent on the objective. However, in all cases, the same control system is applied. For MMC 1, the AC grid and perturbation sources can be switched from a voltage source to a current source in case the MMC is operating in grid-forming control mode. Not connected to any physical components, the simulations do not need to run in real-time so that the time step of the CPU can be chosen freely, and the measuring method can be simulated on the CPU. Apart from that, the measuring method is identical to that applied in the MMC TB setup.

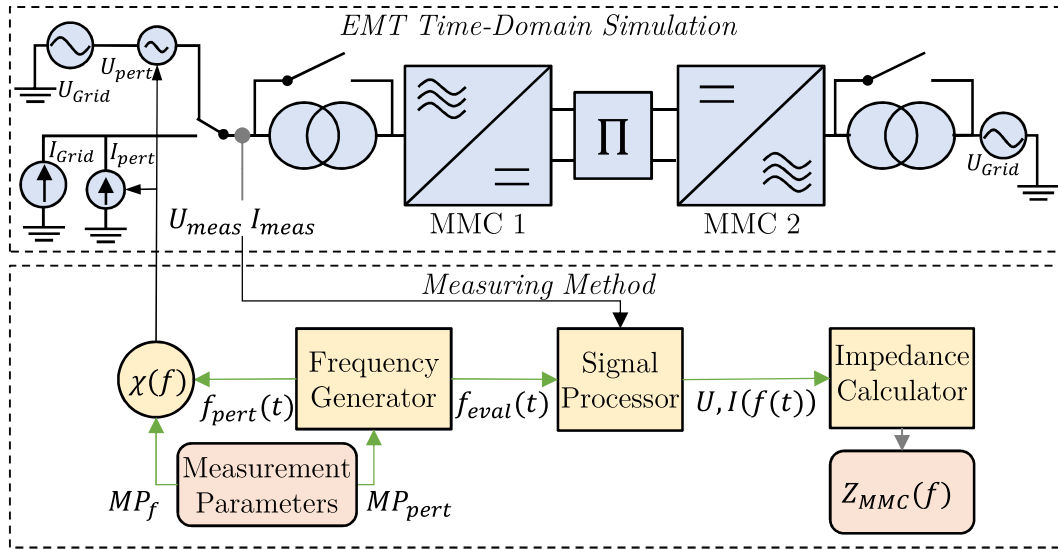


Figure 4.14: Simulated MMC impedance measurement setup.

4.1.7 Method Validation and Evaluation

The developed impedance measurement method is validated by comparing the determined MMC impedance to the impedance obtained from the voltage and current spectrum obtained by post-processing the measured voltage and current signals using an FFT algorithm [166–168].

Figure 4.15 shows the magnitude of the MMC impedance determined by the developed method in addition to the impedance spectrum from 10 Hz to 100 Hz determined by the FFT algorithm serving as benchmark for the validation. Furthermore, Table 4.1 gives the relative difference of the impedance magnitudes of both approaches for frequencies from 10 Hz to 100 Hz. The underlying time-domain simulation is performed using a time step $\Delta t_{CPU} = 10 \mu s$ and changing f_{pert} of the perturbation voltage source every three seconds.

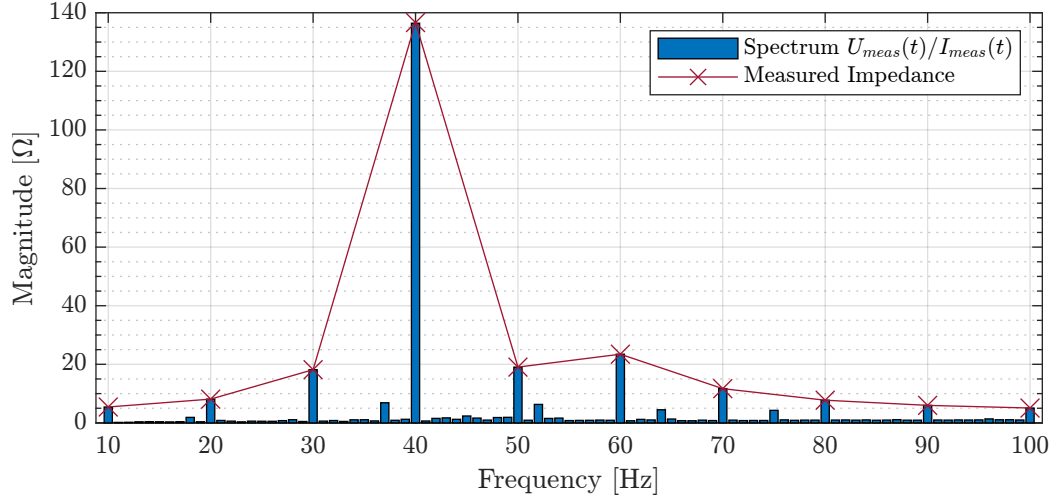


Figure 4.15: MMC impedance magnitude determined by the developed measuring method and by the voltage and current spectrum obtained from a FFT algorithm in post-processing.

Table 4.1: Relative difference of impedance magnitude of measuring method and FFT algorithm.

f_{pert} [Hz]	Magnitude [Ω]	Rel. difference [%]
10	5.43	0.06
20	8.12	-0.04
30	18.16	0.14
40	136.64	0.16
50	19.05	0
60	23.45	0.01
70	11.69	-0.15
80	7.75	-0.05
90	5.99	0
100	5.08	0.23

The impedance spectrum is calculated by applying an FFT algorithm to $U_{meas}(t)$ and $I_{meas}(t)$ and subsequently dividing the voltage by the current spectrum. The results show that the measuring method can match the benchmark data. Differences are less than 0.5% for both the simulated and the laboratory MMC, thus validating the developed measuring method.

Because the impedance calculation is performed online during the measurement period, the results do not become less accurate when the impedance is saved with fewer samples than the time step used during the simulation. Reducing the sampling rate can significantly reduce the size of the measurement data and guarantee a stable data transfer between the RTS and the host PC. As a result, the online impedance measuring method

requires for a measurement period T_{meas}

$$n_{onl} = \frac{n_{\chi} \cdot T_{meas} \cdot n_{f_{pert}}}{ts} \cdot n_{f_{eval}} \quad (4.20)$$

values to be stored, where the sampling time, $ts = 1$ ms of the impedance signal is fixed and independent of the simulation time step, Δt , and therefore, scales only with the number of perturbation frequencies $n_{f_{pert}}$ and evaluated frequencies series, $n_{f_{eval}}$. If only f_{pert} is evaluated, $n_{f_{eval}} = 1$. If an additional frequency series is evaluated, e.g., $f = x \cdot f_{pert}$, then $n_{f_{eval}} = 2$ and so on. The number of signals, n_{χ} , that need to be recorded comprises time and frequency as well as the magnitude and phase of the evaluated signal, χ .

In contrast, the post-processing algorithm depends on the time step, Δt , but does not scale with the number frequencies that are evaluated because the original time-domain data is recorded. Consequently,

$$n_{ppr} = \frac{n_{\chi} \cdot T_{meas} \cdot n_{f_{pert}}}{\Delta t} \quad (4.21)$$

values need to be recorded where the number of signals, n_{χ} , consist of the three-phase values of the evaluated signal in addition to time and frequency. As n_{ppr} is inversely proportional to the time step, Δt , small time steps heavily impact the number of values to be recorded as shown in the following example.

Table 4.2 shows the resulting output data file sizes for both approaches for a measurement series with $n_{f_{pert}} = 10$ and $T_{meas} = 3$ s. The output data are saved as binary MATLAB® files using OPAL RT-Lab for executing the simulation. The example data shows that post-processing data requirements can reach the size of gigabytes for very low time steps. Although the measuring time can be reduced, data requirements remain high even for a low number of frequencies due to the dependency on the time step, Δt . In contrast, the developed online measuring method results in data file sizes less than 1 MB so that even if more than one frequency is evaluated, the data output file sizes remain low.

Table 4.2: Output data file size for a 30 s measuring series.

Δt	Post-processing	Online impedance measurement
10 μ s	175 MB	< 1 MB
5 μ s	356 MB	< 1 MB
1 μ s	1.78 GB	< 1 MB

Comparing the results of the developed online impedance measuring method to those of a post-processing approach validates the developed method. It demonstrates its ability to ensure data file sizes that allow for numerous impedance measurement series, modeling a converter's frequency behavior.

4.2 Impedance Modeling

This section describes how the impedance measuring method's output data is used to build the impedance model of an MMC. Comprising of non-linear elements such as dq-transformations, the control system of MMCs is known to respond to a frequency, f_{pert} , with frequency components other than f_{pert} as pointed out in the analysis in section 2.3.1. Thus, in such cases, the converter system cannot be regarded as a single input and signal output system anymore. As a result, this chapter first evaluates the frequency spectrum of a MMCs for different upper-level control modes to identify significant frequency components when perturbed at f_{pert} . Secondly, it is shown how the response at frequencies other than the perturbation frequency is incorporated in the impedance model [169]. Subsequently, the AC grid impedance model is shown. Additionally a scaling approach for low-voltage impedances is developed and presented in appendix A.2 so that the physical laboratory MMCs can be used to represent the frequency behavior of full-scale MMCs.

4.2.1 Frequency Response Analysis

As stated in section 4.1.2, the choice of perturbation source depends on the converter control mode. As a result, in the case of grid-following control, when the converter is perturbed with a voltage source, the frequency response needs to be analyzed by the converter's current response. In the case of grid-forming control, the converter is perturbed by a current source, and therefore, examining the frequency response requires analyzing the voltage response.

MMC in Grid-Following Control Mode

Figure 4.16 shows the frequency spectrum of the current response when the FS MMC model is subjected to positive as well as negative-sequence perturbation at 40 Hz, 70 Hz, 80 Hz, and 90 Hz. Because the MMC is connected to an AC grid with $f_{grid} = 50$ Hz, the spectrum also shows a large component at that frequency. Figure 4.16 (b) shows that the MMC responds to a negative-sequence voltage perturbation only with a current at the same frequency. In contrast, the current spectrum depicted in Figure 4.16 (a) demonstrates that the converter responds to a positive-sequence voltage perturbation also with currents at frequencies other than the perturbation frequency. For instance, the positive-sequence spectrum shows that, when subjected to a 40 Hz perturbation, the MMC also responds with a current at 40 Hz at 60 Hz and with lower magnitude at 140 Hz. It can be seen that for a 40 Hz perturbation, the current at 60 Hz is 2.8 times larger than that at 40 Hz, indicating a high degree of coupling. When evaluating the frequency spectrum for every perturbation frequency, the results show that the converter responds with a current at $100\text{ Hz} - 2f_{pert}$ and with lower magnitude at $100\text{ Hz} + 2f_{pert}$ in addition to the current at perturbation frequency f_{pert} . These response currents, in the literature, referred to as (mirror) frequency coupling [35, 39, 73], are specific to VSCs and can

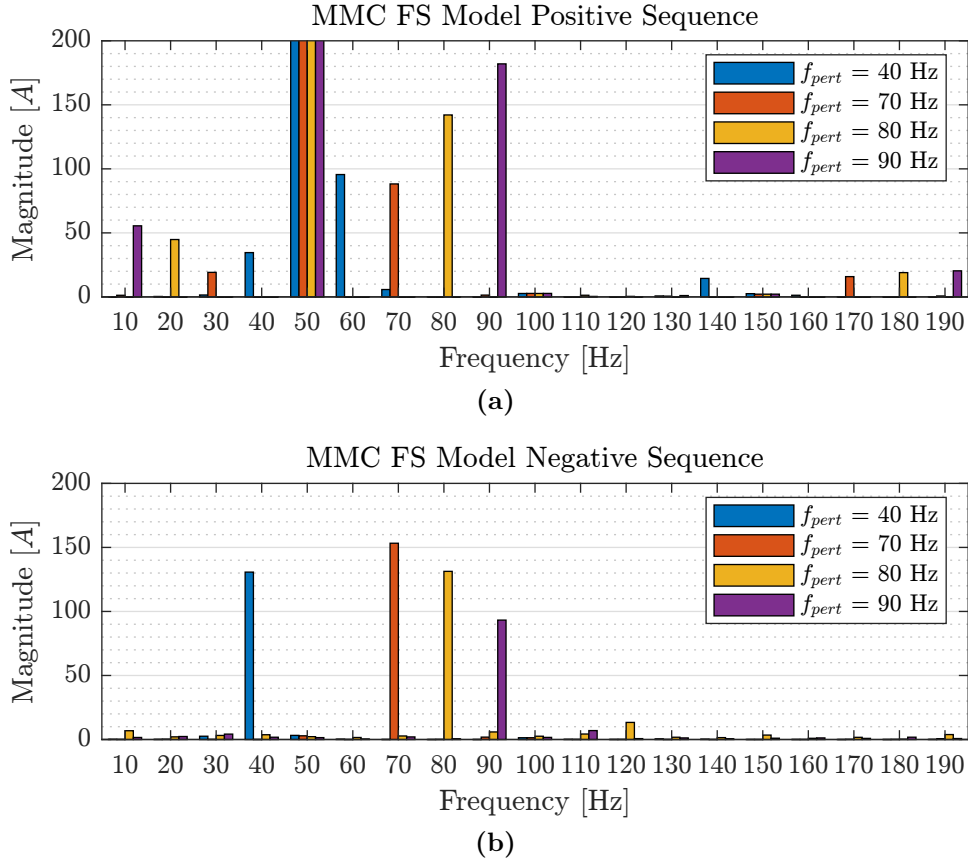


Figure 4.16: MMC current response spectrum with positive-sequence (a) and negative-sequence perturbation (b).

be determined as a negative and positive coupling current relative to the perturbation frequency with

$$f_{pert-} = 2f_{grid} - f_{pert} \quad (4.22)$$

and

$$f_{pert+} = 2f_{grid} + f_{pert}. \quad (4.23)$$

Because the grid frequency is a positive-sequence signal only, no negative-sequence frequency couplings between the grid and perturbation frequency exist as seen in Figure 4.16 (b). The coupling effects occur both for the FS MMC model and the MMC TB whose results are shown in appendix A.3.

The developed measuring method is used to determine the frequency-coupling effect for a wide range of perturbation frequencies. Figure 4.17 shows the current response in absolute values when the converter is perturbed at frequencies ranging from 1 Hz to 10 kHz. For every perturbation frequency, the current at perturbation frequency f_{pert} , at the coupling frequency f_{pert-} and at the coupling frequency, f_{pert+} is shown. That is,

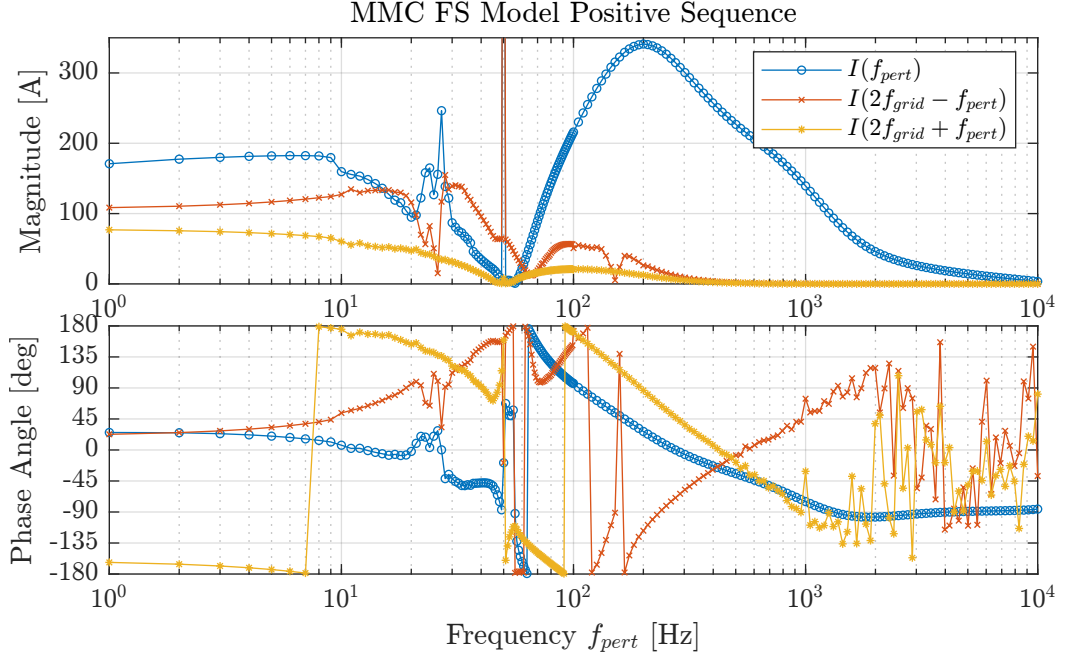


Figure 4.17: MMC current response to a voltage perturbation with frequency f_{pert} .

at $f_{pert} = 40$ Hz the response at 40 Hz as well as 60 Hz and 140 Hz is shown. The curves show that the positive coupling current f_{pert+} is significantly smaller than the negative coupling current f_{pert-} when both coupling currents are compared to the current response at f_{pert} . It can be seen that the frequency coupling effect is relevant mainly at frequencies close to the grid frequency, exceeding the response at the perturbation frequency between 30 Hz and 50 Hz for $f_{pert} = 2f_{grid} - f_{pert}$. With increasing frequency, the response at the coupling frequencies converges to zero. Thus, the frequency response is dominated by the response at f_{pert} only.

MMC in Grid-Forming Control Mode

Figure 4.18 depicts the frequency spectrum of the FS MMC model when operating in grid-forming control and being subjected to a positive- and negative-sequence current perturbation at 40 Hz, 70 Hz, 80 Hz and 90 Hz. The figures show that the MMC reacts to the current perturbation only with a voltage at perturbation frequency, demonstrating that the control system at these frequencies causes no coupling effect contrary to the grid-following control mode. Looking at the frequency response for a wider frequency spectrum as depicted in Figure 4.19 shows that between 1 Hz and 10 kHz no significant coupling effect can be observed. Thus, the coupling effect does not need to be considered in the impedance model when the converter operates in grid-forming control. This shows that the coupling effect is specific to the control system of the converter. Not relying on comparably non-linear controllers such as the PLL, the grid-forming control responds only with voltages at the perturbation frequencies.

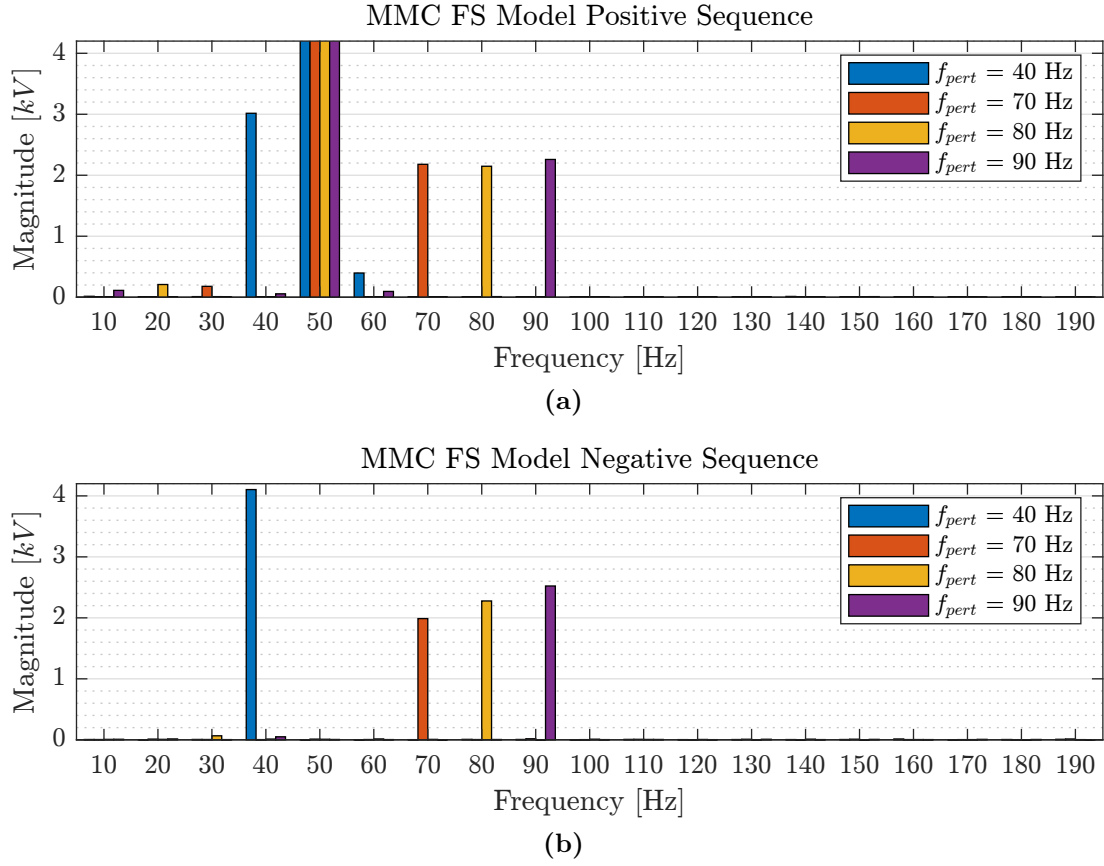


Figure 4.18: MMC voltage response spectrum with positive-sequence (a) and negative-sequence perturbation (b).

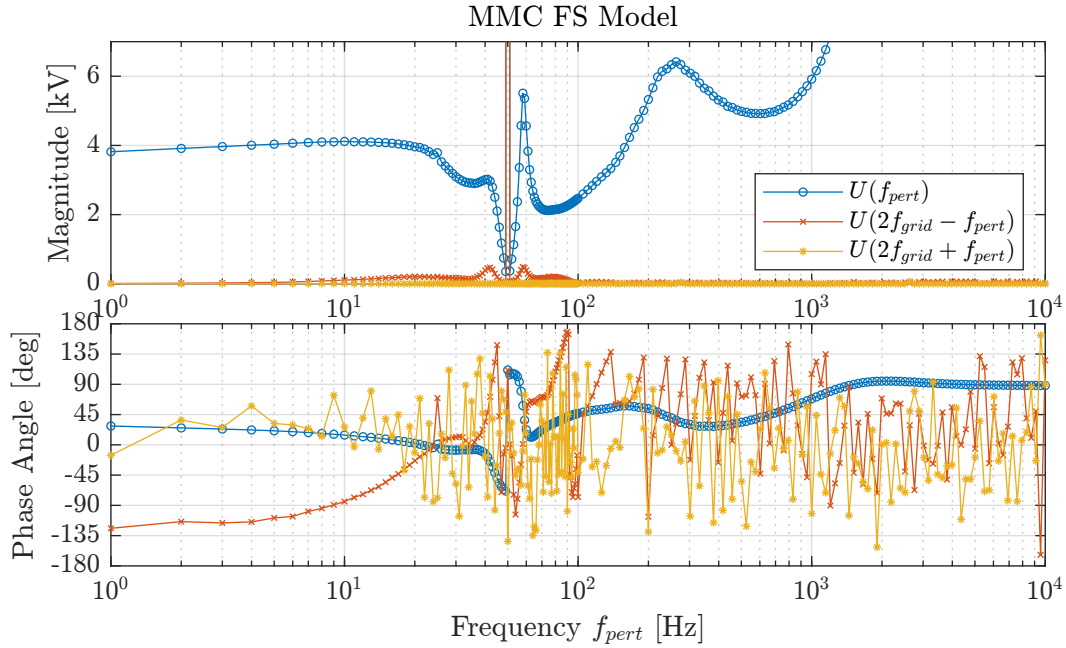


Figure 4.19: MMC voltage response to a current perturbation with frequency f_{pert} .

4.2.2 Frequency Coupling

Section 4.2.1 showed that for converters in grid-following control mode a positive-sequence coupling current $f = 2f_{grid} - f_{pert}$ exists which will be subsequently referred to as i_c . This coupling current influences the output perturbation current $I(f_{pert})$ and depends on the AC grid impedance, to which the converter is connected when the impedance is measured. Figure 4.20 demonstrates how the measured impedance varies for measurement setup with different grid impedances [169].

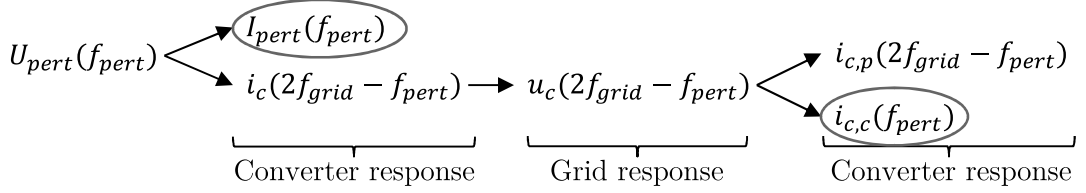


Figure 4.20: Influence of coupling current on perturbation current [169].

Perturbing the converter with a small-signal voltage at frequency f_{pert} not only leads to a response current at the same frequency but also a response current i_c at $f = 2f_{grid} - f_{pert}$ as shown in the previous analysis. This current, however, induces a voltage u_c at the same frequency due to the grid impedance. The voltage, u_c , can be understood as an additional perturbation source to which the converter responds again with a current, $i_{c,p}$, at the same frequency that is $f = 2f_{grid} - f_{pert}$. Moreover, due to the coupling effect, the converter responds to the grid-induced perturbation, u_c also with a current, $i_{c,c}$, at frequency

$$f = 2f_{grid} - (2f_{grid} - f_{pert}) = f_{pert}. \quad (4.24)$$

As a result, the measured current response, $I_{pert}(f_{pert})$, also includes the current, $i_{c,c}(f_{pert})$, that is dependent on the grid impedance. This effect is undesirable because a dependency of the converter impedance on the grid impedance means that the grid impedance needs to be considered for accurate results. While it is possible to include the grid impedance in the measurement setup, this approach would contradict the idea of the IbSC that the impedances of the two subsystems can be derived independently. Hence, the impedance model is extended to incorporate the additional coupling current using ideal measurements. For this purpose, the converter impedance is measured when connected to an ideal grid, meaning a grid impedance close to zero. The effect of the additional coupling current subsequently is included according to the grid impedance of interest.

Impedance Modeling of Frequency Coupling

To include the additional current, $i_{c,c}$, and the resulting grid dependency into the impedance model, an additional impedance, $Z_{c,c}(f)$, is added in parallel to the converter

impedance, Z_{MMC} [73]. For simplification, the additional impedance will be derived as admittance, $Y_{c,c}$, so that the overall converter admittance

$$Y_{MMC}(f_{pert}) = \frac{I_{pert}(f_{pert}) + i_{c,c}(f_{pert})}{U_{pert}(f_{pert})} = \underbrace{\frac{I_{pert}(f_{pert})}{U_{pert}(f_{pert})}}_{Y_{pert}} + \underbrace{\frac{i_{c,c}(f_{pert})}{U_{pert}(f_{pert})}}_{Y_{c,c}(f_{pert})}. \quad (4.25)$$

Because Y_{pert} is a direct output of the impedance measurement method described in section 4.1, only $Y_{c,c}$ is required in addition to include the coupling effect. Based on the method presented in [118] that is developed for analytically impedance models, in the following paragraphs, $Y_{c,c}$ is derived and used for measurement-obtained signals [169].

The coupling admittance Y_c models the relationship of the perturbation voltage and coupling current with

$$Y_c(f_{pert}) = \frac{i_c(2f_{grid} - f_{pert})}{U_{pert}(f_{pert})}. \quad (4.26)$$

As shown in Figure 4.20, the current $i_c(2f_{grid} - f_{pert})$ induces a voltage $u_c(2f_{grid} - f_{pert})$ that depends on the grid and converter admittance and can be calculated to

$$u_c(2f_{grid} - f_{pert}) = \frac{i_c(2f_{grid} - f_{pert})}{Y_{pert}(2f_{grid} - f_{pert}) + Y_{grid}(2f_{grid} - f_{pert})}. \quad (4.27)$$

This coupling voltage can be considered as an additional perturbation voltage causing the converter to respond with the current $i_{c,p}$ at the same frequency along with a second coupling current, $i_{c,c}$, at frequency $f = f_{pert}$ as derived in (4.24) with

$$i_{c,c}(f_{pert}) = Y_c(2f_{grid} - f_{pert}) \cdot u_c(2f_{grid} - f_{pert}). \quad (4.28)$$

Inserting (4.28) in the right summand of (4.25) for $i_{c,c}(f_{pert})$ results in

$$Y_{c,c} = \frac{Y_c(2f_{grid} - f_{pert}) \cdot u_c(2f_{grid} - f_{pert})}{U_{pert}(f_{pert})}. \quad (4.29)$$

Now representing $u_c(2f_{grid} - f_{pert})$ in (4.29) by (4.27) gives

$$Y_{c,c} = \frac{Y_c(2f_{grid} - f_{pert})}{U_{pert}(f_{pert})} \cdot \frac{i_c(2f_{grid} - f_{pert})}{Y_{pert}(f_{pert}) + Y_{grid}(2f_{grid} - f_{pert})}. \quad (4.30)$$

The fraction $\frac{i_c(2f_{grid} - f_{pert})}{U_{pert}(f_{pert})}$ in (4.30) can be replaced by Y_c as in (4.26) so that

$$Y_{c,c}(f_{pert}) = \frac{Y_c(2f_{grid} - f_{pert}) \cdot Y_c(f_{pert})}{Y_{pert}(2f_{grid} - f_{pert}) + Y_{grid}(2f_{grid} - f_{pert})}. \quad (4.31)$$

As a result, the derived admittance, $Y_{c,c}(f_{pert})$, allows for the integration of the current, $i_{c,c}(f_{pert})$, into the overall impedance model, $Z_{MMC}(f_{pert})$, by considering the influence of the grid impedance. Equations (4.26) and (4.31) show that to derive $Y_{c,c}$, the

signals $i_c(2f_{grid} - f_{pert})$ and $U_{pert}(f_{pert})$ need to be measured in addition to the admittance $Y_{pert}(f_{pert})$ that is already measured as previously described. Consequently, the measurement method is extended to determine also the voltage U_{pert} at f_{pert} and the coupling current i_c at $2f_{grid} - f_{pert}$ as presented by the red line in Figure 4.17. The grid impedance model determining $Y_{grid}(f_{pert})$ is described hereinafter in section 4.2.3. Furthermore, (4.31) shows that the required admittances need to be obtained also for the coupling frequency $2f_{grid} - f_{pert}$. Contrarily to the coupling current, i_c , that is measured at $2f_{grid} - f_{pert}$ when perturbing the system at f_{pert} , the admittances at $2f_{grid} - f_{pert}$ are obtained by shifting the measured signals at f_{pert} by $2f_{grid} - f_{pert}$.

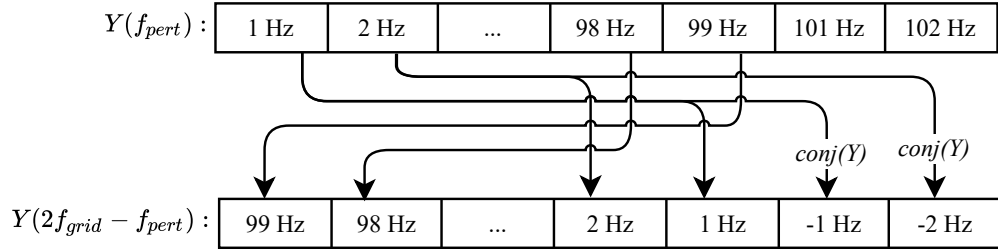


Figure 4.21: Shifting of admittances as in (4.31).

Figure 4.21 illustrates how signals are shifted. While the first row shows the elements of the admittance when sorted according to f_{pert} , the second row shows the same admittance in the order according to $f = 2f_{grid} - f_{pert}$ using f_{pert} of the first row as input, meaning that for $f_{pert} > 100$ Hz a negative admittance is needed. Utilizing the fact that the underlying signals used for calculating Y are purely real signals in the time domain, the complex conjugate of Y at the corresponding positive frequency value is taken [170]. It should be noted that the existence of measurement data at $2f_{grid} - f_{pert}$ depends on the method's parameterization. However, the impedance measuring method is parameterized such that at frequencies below 100 Hz, impedance values at every integer frequency are obtained. Therefore, sufficient accuracy in this frequency region is achieved while ensuring that the values at the required frequencies exist. As a result, the shifted admittances can be obtained for frequencies until 200 Hz. As it can be seen in Figure 4.17, the coupling current is significantly smaller at frequencies above 200 Hz so that the frequency-coupling effect does not need to be included in the impedance model anymore.

Equations (4.26) - (4.31) show that when perturbing the converter at f_{pert} the voltage $U_{pert}(f_{pert})$ and the current, $I(2f_{grid} - f_{pert})$, is needed to obtain the coupling admittance, $Y_{c,c}(f_{pert})$. Consequently, the measuring method is extended to obtain the required signals along with the converter impedance. Contrary to standard, commercial off-the-shelf measuring tools that require complex workarounds to measure frequencies other than the perturbation frequency [171], the developed measuring method can be adjusted easily for the additional signals. Figure 4.22 illustrates how modeling of frequency coupling affects the MMC impedance. Compared to an impedance without modeling

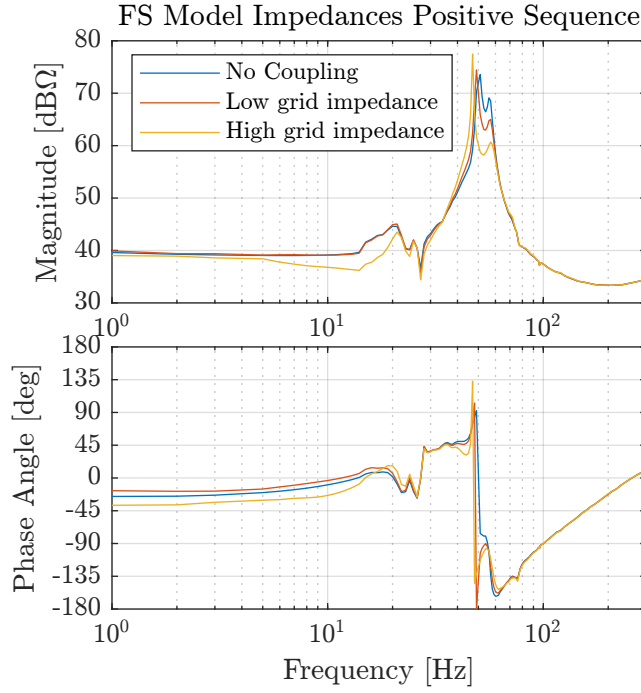


Figure 4.22: MMC impedance model with and without integration of coupling effect.

of the frequency coupling effect, a high grid impedance significantly alters the MMC impedance at frequencies close to 50 Hz. While a low grid impedance that is three times smaller still impacts the MMC impedance, the effect is significantly reduced. It can also be seen that at frequencies above 70 Hz, no effect can be seen anymore, verifying the previous assumptions that it is sufficient to model the coupling effect only for frequencies up to 200 Hz.

4.2.3 AC Grid Impedance Model

The impedance model of the AC grid is an exact representation of that of the time-domain presented in section 3.2.2. Consisting only of passive components, the AC grid model meets the assumption of the IbSC which requires the AC grid to be stable in open-circuit when being unloaded. Figure 4.23 illustrates an example of an AC grid impedance with a short-circuit power of 5 GV A with $X/R = 10$ because the inductance is typically significantly smaller compared to the resistance in transmission grids [139].

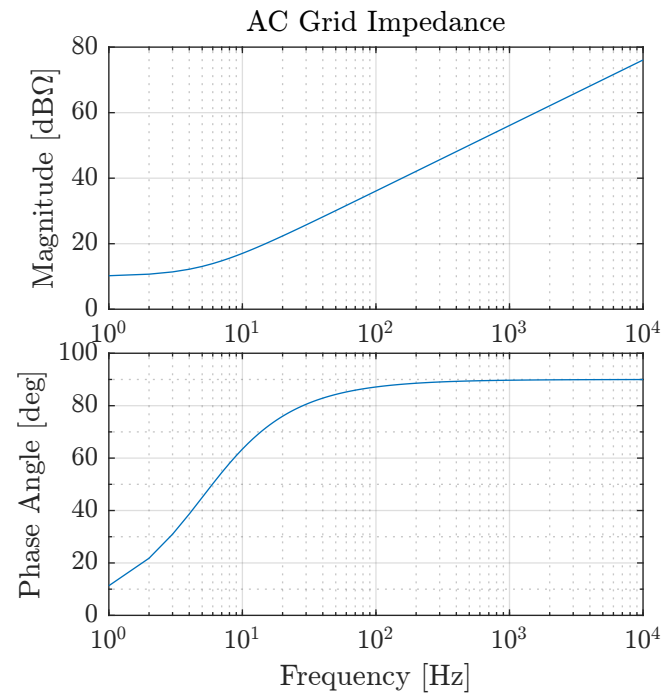


Figure 4.23: Exemplary AC grid impedance.

5 Results

This chapter addresses the research questions posed in chapter 1.5. Chapter 5.1 analyzes whether a Type 5 average MMC model that utilizes controllable voltage sources to represent the respective phase-arms can be used for investigation converter control interactions. Chapter 5.2 validates the chosen stability assessment approach and shows how the impedance derivation method developed in this thesis can be used to investigate converter control interactions even when the converter is a black-box to the person conducting the stability studies. Lastly, chapter 5.3 presents several test cases that are used to identify unstable grid situations caused by converter control interactions with the AC grid.

5.1 Identification of Model Limitations

The limitations of the MMC model are investigated by comparing the measured impedances of a physical laboratory MMC, the MMC TB, to those of the simulated MMC model. Because the MMC TB represents a down-scaled model, the equivalent simulated LS MMC model is used for the comparison. Bypassing the converter transformer in a subsequent analysis, the influence of the converter transformer on the impedance trajectories can be investigated.

5.1.1 MMC Test Bench and Laboratory-Scaled Model Comparison

This section compares the derived AC-side impedances of the MMC TB and the LS MMC model to examine their frequency behavior and identify frequency areas where the model is not able to determine the frequency behavior accurately. Although having fewer SMs compared to HVDC MMCs, the MMC TB is able to capture the internal dynamics of an MMC because it consists of individual SMs compared to controlled voltage sources that are used in the simulated MMC model. Figure 5.1 shows the positive-sequence impedances of the MMC TB (solid line) and the LS model (dashed line) for the converter operating in grid-following control mode (blue), as well as grid-forming control mode (red).

Investigating the influence of the different control modes at first by comparing the blue and red lines with each other, makes it possible to separate the impedances in frequency ranges that are dominated by the control system or by the electrical structure of the MMC. Analyzing the MMC TB trajectories (solid lines) in Figure 5.1 shows that the magnitudes of the MMC TB impedances in grid-following and grid-forming control mode deviate at frequencies below 20 Hz but show a similar trend. However, between 20 Hz and

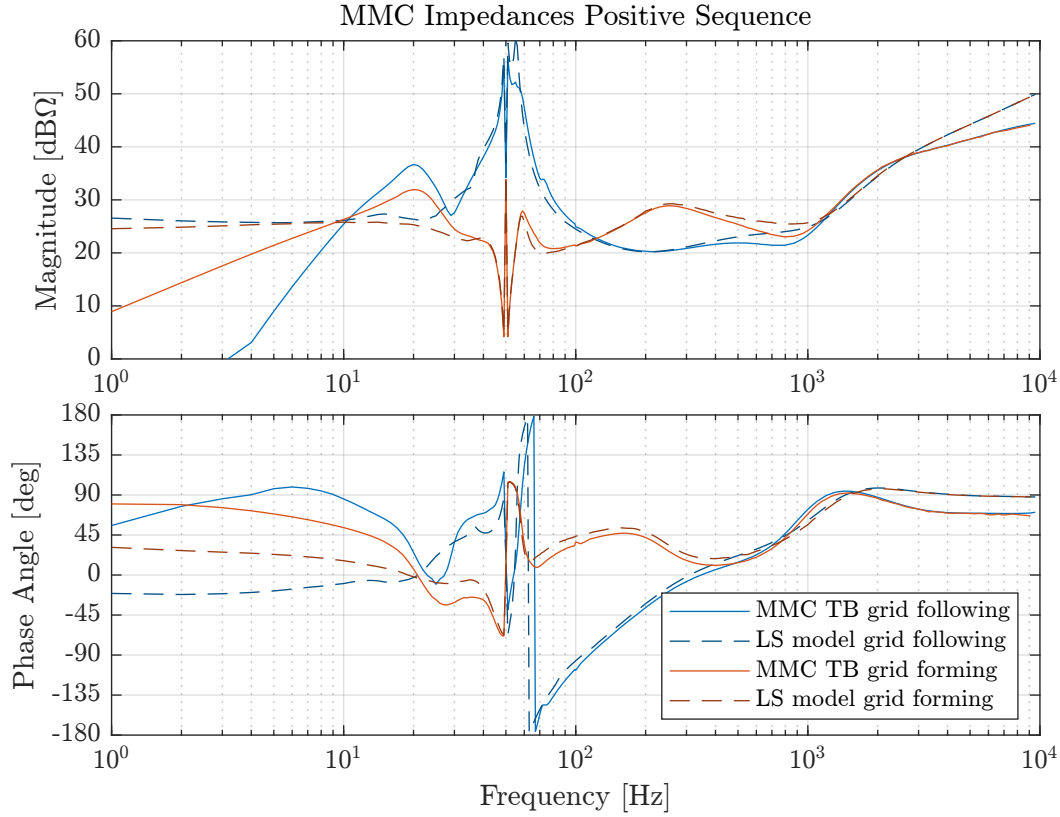


Figure 5.1: MMC TB and LS model impedance for grid-following and grid-forming control.

1 kHz the converters show entirely different frequency behaviors in grid-forming and grid-following control mode. While in grid-following control mode, the magnitude increases initially and then decreases after 50 Hz, in grid-forming control mode, a local minimum at 50 Hz followed by two maxima can be observed. However, above 1 kHz magnitude and phase align, and no differences between the control modes can be recognized. At 50 Hz there is an outlier that can be explained by the fact that at 50 Hz the applied perturbations are indistinguishable from the grid current and voltage. With an equal operation point of 0.5 per unit (pu) active power transmission, the impedance at 50 Hz is the same for grid-following and grid-forming control modes. In comparison, the impedance trajectories of the simulated LS model (dashed lines) show the same differences for grid-following and grid-forming control modes with the exception that at lower frequencies, the impedance not only follows a similar trend, but also shows no significant differences in magnitude. Because the different control modes do not result in different trajectories of the impedances above 1 kHz, it can be concluded that the frequency behavior is mainly influenced by the electrical structure, in particular the arm inductances of the MMC TB and LS model which shape the impedances at high frequencies. In contrast, below 1 kHz, the control system dominates the frequency behavior as grid-forming or following control modes cause varying impedance trajectories. Additionally, the results indicate that below 20 Hz, the control system plays a less significant role compared to the converters' electrical structure.

Investigating the difference between the MMC TB and the LS model shows that between 40 Hz and 1 kHz, where the control system dominates the frequency behavior, the MMC TB impedances in grid-following, as well as grid-forming control mode, match those of the LS model. Above 1 kHz, however, the LS model impedances begin to deviate from those of the MMC TB with the LS impedance showing purely inductive behavior, and thereby converging to a phase angle of 90° . In comparison, the MMC TB phase angle converges to 70° . Also, below 40 Hz the MMC TB and the LS model impedances start to vary as well, showing a completely different frequency behavior below 20 Hz. Analogue observations can be made for the negative-sequence impedances that are presented in Figure A.6 in appendix A.4. During the measurements, it can be seen that the MMC TB reacts to perturbations below 30 Hz with extremely high currents at the same frequency as it can be seen in Figure A.5. Unapparent in the LS model, the cause of these over-currents is narrowed down by subsequently investigating the influence of the converter transformer.

5.1.2 Influence of Transformer

The influence of the transformer on the MMC impedance is investigated by performing measurements and simulations of the MMC TB and the LS model with and without the converter transformer. In the case of the simulations, the transformer is simply removed from the model, and the MMC TB requires additional switches to bypass the transformer. Therefore, a switch is added to the laboratory setup that disconnects the transformer and provides a direct connection between the transformer's primary and secondary sides. Additionally, the supplied grid voltages are changed from 400 V to 208 V to provide nominal AC RMS phase-to-phase voltage level of the MMC station.

Figure 5.2 shows the impedances of the MMC TB and the LS model in grid-following control mode without transformers (green lines), in addition to the measurements and simulations including the converter transformers that are also presented before in Figure 5.1 (blue lines). The comparison demonstrates that the converter transformer causes the high currents that are observed at perturbation frequencies below 20 Hz, leading to a decrease in the MMC TB magnitudes at lower frequencies. Although the differences in magnitude and phase remain, the MMC TB and LS model impedances show a similar trend without the transformer, deviating at a constant factor that can be contributed to differences in the parameters of the MMC TB and LS model. However, differences at frequencies above 3 kHz remain the same and are not caused by the transformer because, at those frequencies, no differences can be seen except that the impedances with transformers have a higher impedance magnitude due to the additional transformer impedance. These differences are likely caused by additional resistive parts of the arm inductances that dominate the frequency behavior at frequencies above 3 kHz.

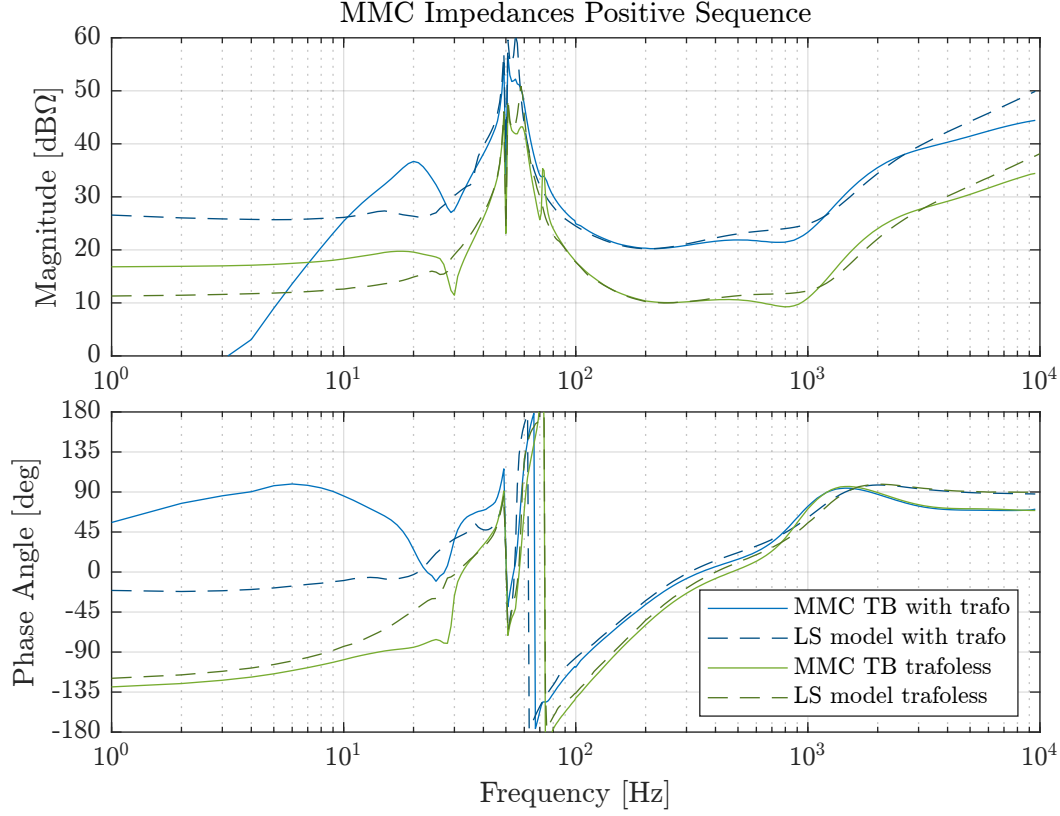


Figure 5.2: MMC TB and LS model impedance with and without transformer.

5.1.3 Conclusion

The investigations demonstrate that the validity of an MMC model depends on the frequency range in which converter control interactions might cause unstable grid situations. While at low frequencies, the overall converter impedance is highly impacted by the transformer, and at high frequencies by the arm inductances, at frequencies above the fundamental frequency and below 3 kHz, the control system dominates the frequency behavior, and the MMC model is able to replicate the laboratory converter impedance. Thus, the following conclusions can be made:

- Impedances measured utilizing laboratory MMCs can be used to detect phenomena influencing the impedance trajectory, which are not captured in models due to underlying assumptions and simplifications. However, impedances measured based on laboratory converters need to be scaled if they are used for interaction studies of the transmission system. A suitable scaling approach is presented in chapter A.2.
- In frequency areas in which the control system dominates the frequency behavior, simplified MMC models can also be used to model the converter's electrical structure. As a result, the impedances derived in this thesis on the basis of the FS MMC model can be used for identifying unstable grid situations caused by fast interactions at frequencies between 30 Hz and 3 kHz.

- An exact representation of the control system is crucial for accurate stability analysis. Hence, converter control replicas provide a means to investigate converter control interactions of real MMCs where the exact control system is unknown to the system operator.
- At very low or high frequencies, the electrical structure and components of the MMC play an important role in the frequency behavior. Consequently, non-detailed MMC models should be considered carefully when interactions are observed at those frequencies.

5.2 Impedance-based Stability Assessment Validation

Relying on linearized models, the investigation of converter control interactions in the frequency domain using the IbSC requires validation of the stability assessment. Hence, a test case is presented that demonstrates a stable, as well as an unstable system. Both systems are modeled in Matlab/Simulink to verify the implemented method and the stability predictions. The MMC is represented by the FS model with the parameters given in chapter 3 Table 3.3 and the converter operating in grid-following control mode. Varying the short circuit power of the AC grid, and accordingly, the grid impedance $Z_{grid}(f)$ enables the creation of stable, as well as unstable test cases. The short circuit power $S_{SC,grid}$ of the AC grid is given relative to the nominal power rating $P_{n,MMC}$ of the MMC by means of the SCR so that

$$S_{SC,grid} = SCR \cdot P_{n,MMC}. \quad (5.1)$$

The parameters for the resulting impedance $Z_{grid}(f) = R_{grid} + j2\pi f_{grid}L_{grid}$ are given in Table 5.1.

Table 5.1: Grid impedance parameters.

SCR	$S_{SC,grid}$ [GVA]	R_{grid} [Ω]	L_{grid} [mH]
4	4.8	3.3168	105.58
5	6.0	2.6534	84.46
6	7.2	2.2112	70.38
7	8.4	1.8953	60.33

5.2.1 Frequency-Domain Stability Assessment

Deriving the output current, $I(f)$, of the equivalent circuit diagram presented in Figure 5.3 to

$$I(f) = \left(I_{MMC}(f) - \frac{U_{grid}(f)}{Z_{MMC}(f)} \right) \cdot \frac{1}{1 + \frac{Z_{grid}(f)}{Z_{MMC}(f)}} \quad (5.2)$$

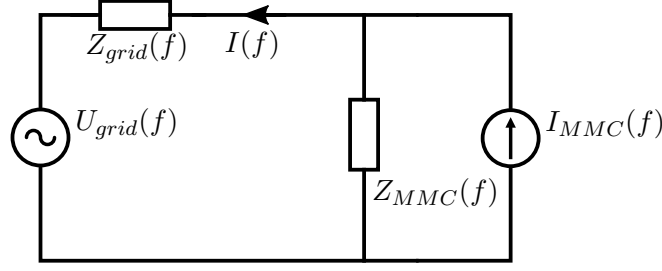


Figure 5.3: Equivalent circuit diagram of AC grid connected to MMC.

determines the feedback loop gain $\frac{Z_{grid}(f)}{Z_{MMC}(f)}$, that allows for the calculation of the system's phase margins, facilitating a graphical stability assessment of the validation test case. The frequencies of the phase margin, f_{PM} , are determined by the intersection points of the grid and MMC impedance where

$$|Z_{grid}(f_{PM})| = |Z_{MMC}(f_{PM})|. \quad (5.3)$$

The difference in the phase angles at f_{PM} , that is subtracted from 180° gives the PMs such that

$$PM = 180^\circ - \arg\left(\frac{Z_{grid}(f_{PM})}{Z_{MMC}(f_{PM})}\right) = 180^\circ - (\arg(Z_{grid}(f_{PM})) - \arg(Z_{MMC}(f_{PM}))). \quad (5.4)$$

Figure 5.4 and Figure 5.5 show the positive- and negative-sequence MMC impedance and the AC grid impedances for different SCRs.

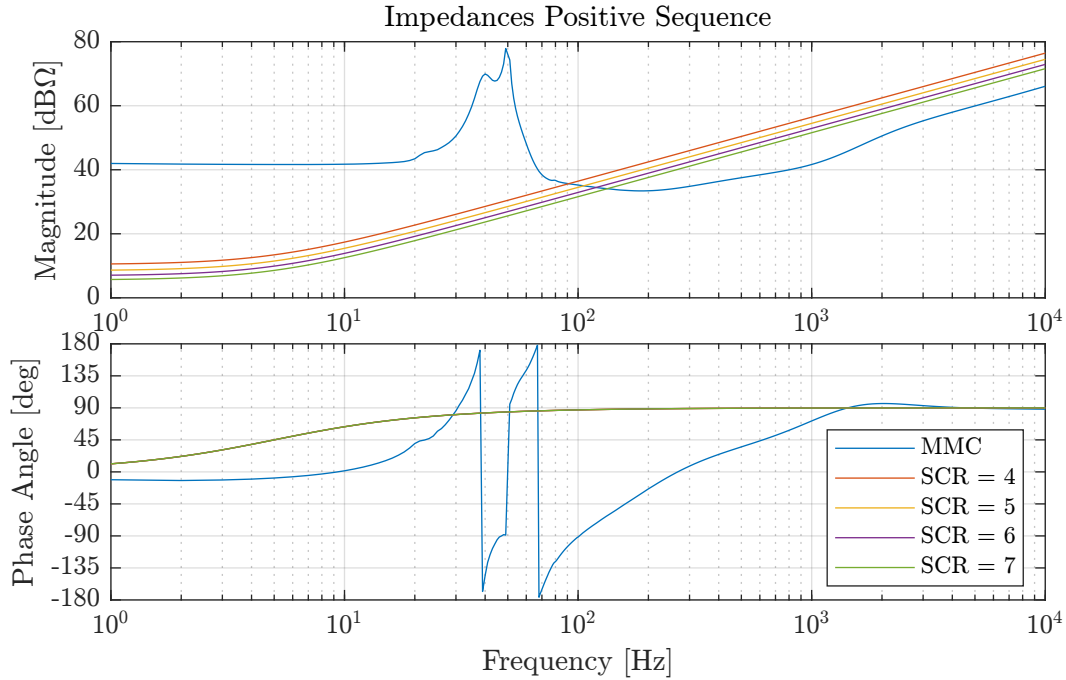


Figure 5.4: MMC impedance and AC grid impedance for different SCRs.

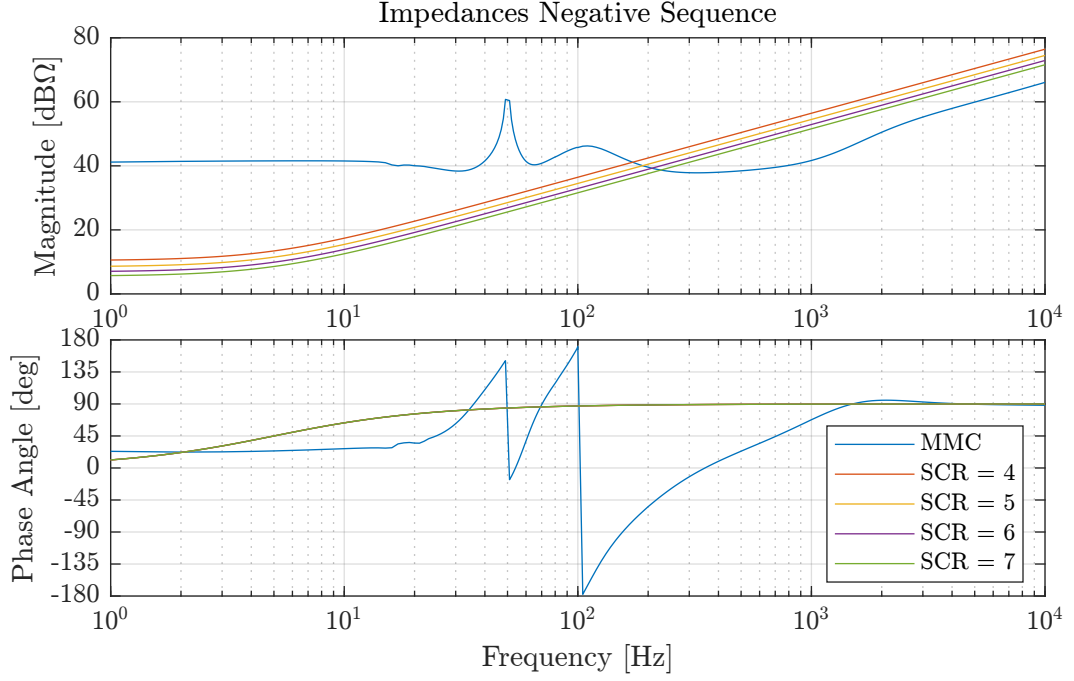


Figure 5.5: MMC impedance and AC grid impedance for different SCRs.

The depicted Bode plots show that the magnitudes of the grid impedances intersect those of the MMC around 100 Hz for a positive-sequence impedance and around 200 Hz for a negative-sequence impedance. Having phase differences close to 180° at the intersection frequencies, the graphical analysis indicates that the stability of a test case might be threatened. Figure 5.6 allows for a closer look at the impedances around the intersection frequencies and the resulting PMs by highlighting the intersections of the magnitudes with a circle and marking the corresponding phase values with a cross. The magnitude plots show that with increasing SCR, the intersection points of the MMC and the grid impedance move to the left, and therefore lower frequencies. Meanwhile, the phase plots show that at lower frequencies, the phase difference of the positive-sequence MMC impedance and the grid impedances increases and exceeds 180° for $SCR = 4$. Hence, calculating the PMs as in (5.4), results in a negative PM for positive-sequence impedances. This can be seen in Table 5.2 that shows the positive-sequence impedance's PM PM^p at frequency f_{PM}^p and the negative-sequence impedance's PM PM^n at f_{PM}^n . Thus, the IbSC predicts an unstable system for an AC grid with SCR equal to four, and stable systems for AC grids with SCRs higher than four.

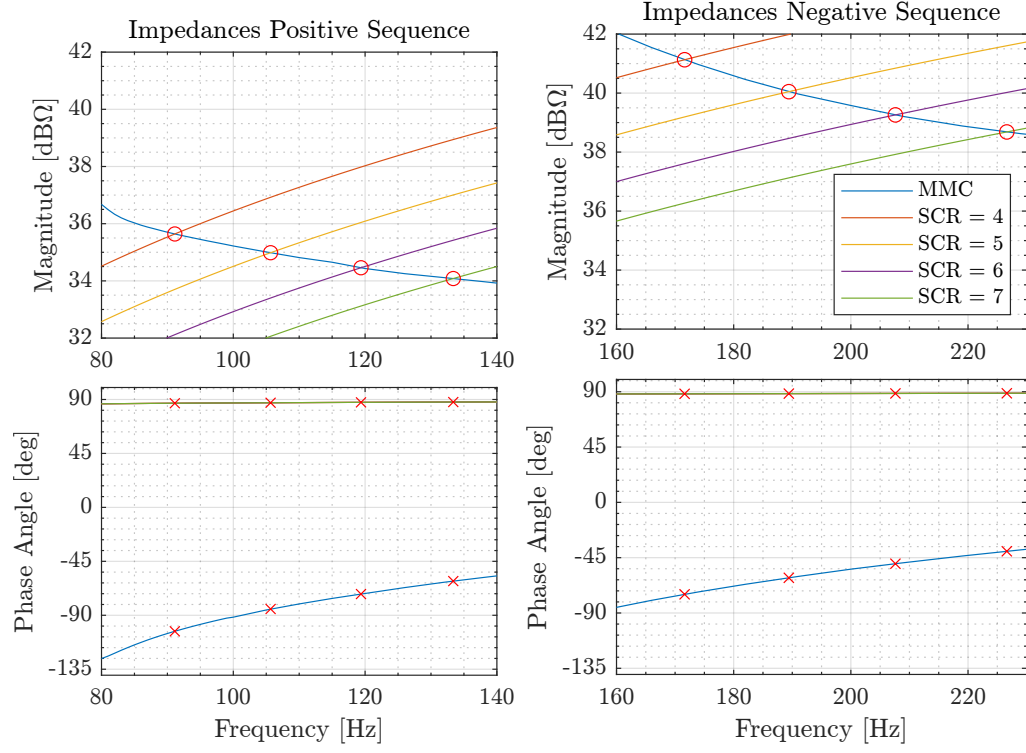


Figure 5.6: Close up of MMC and AC grid impedance intersections.

Table 5.2: Phase Margins for positive- and negative-sequence impedances.

SCR	f_{PM}^p	PM^p	f_{PM}^n	PM^n
4	91 Hz	-10°	172 Hz	17°
5	106 Hz	8°	189 Hz	30°
6	119 Hz	20°	208 Hz	41°
7	133 Hz	31°	227 Hz	51°

Encircling the $(-1, j0)$ point for the positive-sequence impedance and for a $SCR = 4$, but not for a $SCR = 5, 6$ and 7 , the Nyquist plots shown in Figure 5.7 confirm the stability assessment based on evaluating the phase differences.

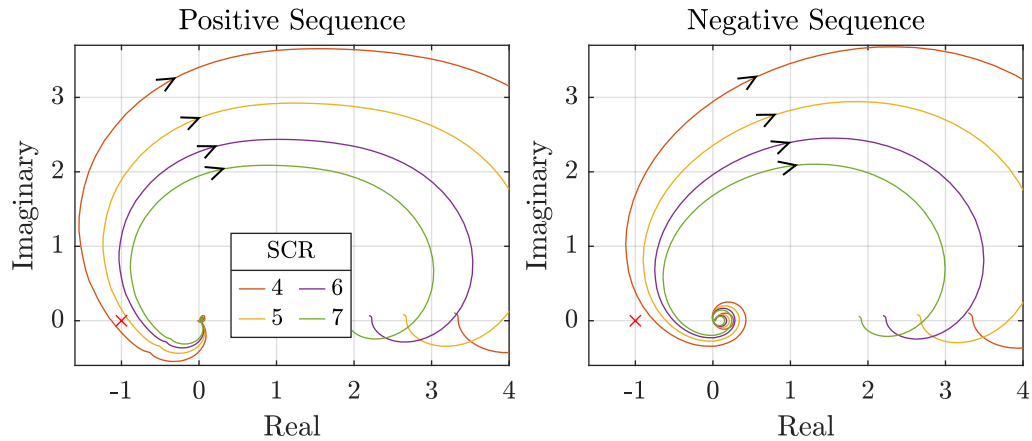


Figure 5.7: Nyquist plots of AC grid and MMC impedance ratio.

5.2.2 Time-Domain Validation

The aforementioned stability predictions are verified by simulating equivalent systems in the time domain, using Matlab/Simulink. After energizing and starting the MMC, the MMC is switched to an AC grid with decreasing SCR every two seconds. Figure 5.8 shows the current at the AC terminal of the MMC.

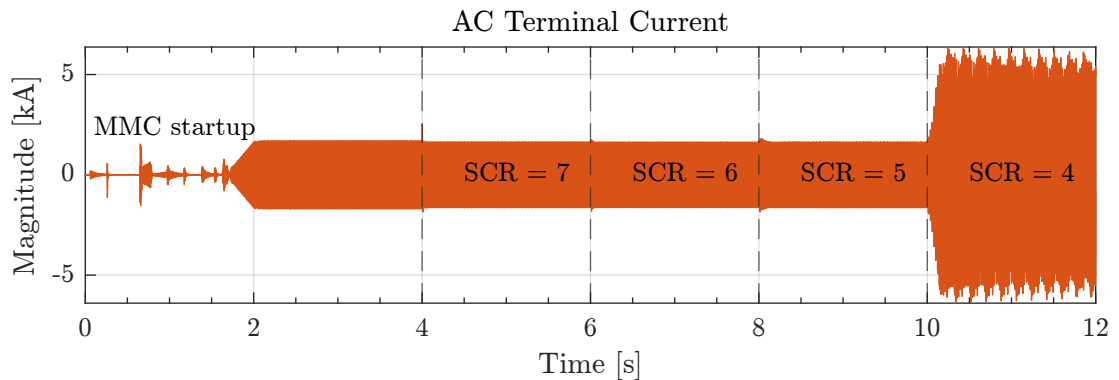


Figure 5.8: AC terminal current when MMC is connected to AC grids with decreasing SCR.

At $t = 4$ s transients resulting from the startup process are assumed to have decayed and the MMC is connected to an AC grid with $\text{SCR} = 7$. Switching the MMC consecutively to AC grids with $\text{SCR} = 6$ and 5 shows that the MMC remains in a stable operation as predicted by the IbSC. However, at $t = 10$ s, the current sharply increases after the MMC is connected to an AC grid with $\text{SCR} = 4$. A close up of the three-phase current and voltage signals in Figure 5.9 reveals that immediately after switching, both current and voltage oscillate at frequencies other than the grid frequency of 50 Hz. Thus, a lower SCR is not solely responsible for causing the instabilities.

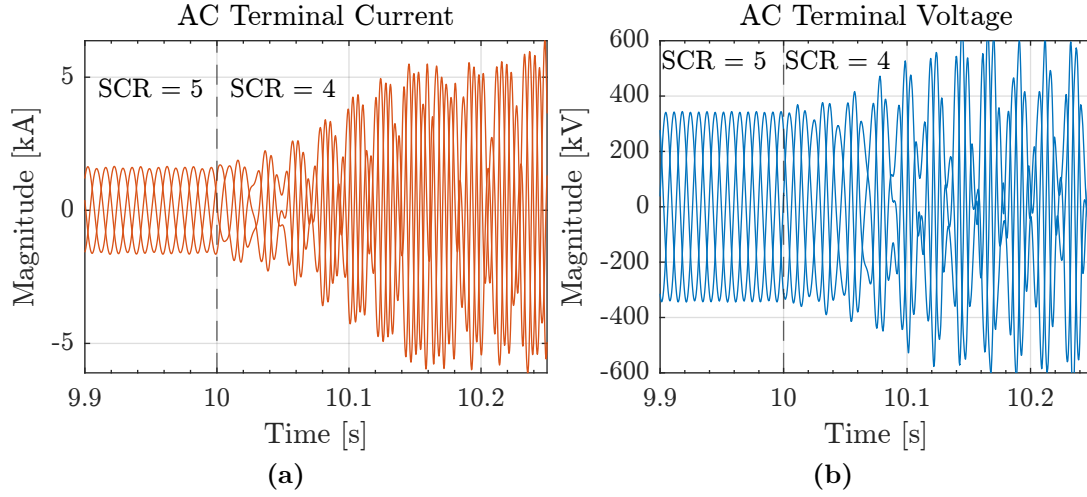


Figure 5.9: AC terminal current (a) and voltage (b) when switched to an AC grid with SCR = 4.

Calculating the Fourier spectrum of one phase of the current and voltage signals determines the frequencies of the observed oscillations. Figure 5.10 shows that the spectrum of voltage and current has the highest magnitude at $f = 94.5$ Hz matching the instability prediction at $f = 91$ Hz by the frequency domain stability assessment. Moreover, the spectrum analysis shows that at this frequency, the current magnitude is 270 % of the 1.5 kA grid current, and the grid voltage has an additional 80 % component relative to the 320 kV magnitude at 50 Hz. As a result, current and voltage are heavily distorted with a Total Harmonic Distortion (THD) of 297.42 % and 87.42 %, respectively. Thus, the time-domain validation demonstrates that the developed impedance measuring method facilitates an accurate stability prediction using the IbSC.

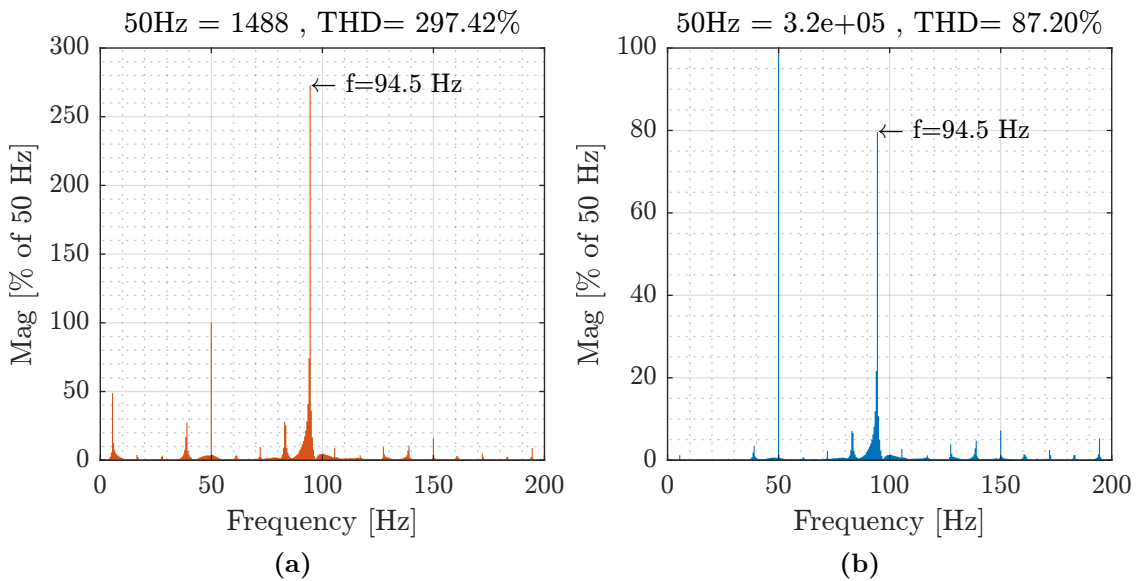


Figure 5.10: Fourier spectrum of phase a of the AC terminal current (a) and voltage (b) when connected to an AC grid with SCR = 4.

5.3 Stability Assessment

Several test cases are investigated to identify unstable grid situations, applying the impedance derivation method that is developed in chapter 4 and validated in chapter 5.2.

5.3.1 Test Case Analysis

While the lower-level controls remain the same, the upper-level controls can be varied during operation of the MMC to account for different grid situations and power flow requirements. Moreover, the outer loops of the upper-level controls are often neglected in MMC impedance models [92] and thus, merit investigation.

As the analysis in chapter 5.1 concluded that the impedance models derived on the basis of the EMT time-domain model are valid between 30 Hz and 3 kHz and the validation test case showed oscillations around 91 Hz, the FS model impedances are used for the stability assessment.

Varying the control variables of the outer loop of the upper-level control system and consequently changing the reference values i_d^* and i_q^* for the current controller results in the test cases presented in Table 5.3. Additionally, the reference values for P_{AC} are varied throughout the test cases to investigate different active power transmissions. The reference values for the DC and AC voltage are not varied and set to 1 pu. If the reactive power is controlled, its reference value is set to zero. In test case 1 to 3 the MMC is operated in active and reactive power control mode with varying power flows. In test case 4 to 6, 0.5 pu of active power is transmitted by the MMC while changing the controlled reference variables to P_{AC} & U_{AC} in test case 4, U_{DC} & U_{AC} in test case 5 and U_{DC} & Q_{AC} in test case 6. Lastly, test case 7 investigates converter control interactions with the AC grid if the MMC operates in grid-forming control mode, defining the AC voltage magnitude and frequency.

Table 5.3: Control and powerflow settings.

Test Case	Reference Variables	Powerflow at MMC 1 [pu]
1	P_{AC} & Q_{AC}	$P_{AC} = 0.5$
2	P_{AC} & Q_{AC}	$P_{AC} = 1$
3	P_{AC} & Q_{AC}	$P_{AC} = -0.5$
4	P_{AC} & U_{AC}	$P_{AC} = 0.5$
5	U_{DC} & Q_{AC}	$P_{AC} = 0.5$
6	U_{DC} & U_{AC}	$P_{AC} = 0.5$
7	f_{grid} & $ U_{AC} $	$P_{AC} = 0.5$

In addition to the influence of different operation modes of the converter, the impact of varying grid strengths on stability is investigated by changing the SCR and hence, the AC grid impedance for each test case. Figure 5.11 depicts an overview of the executed test cases and shows the results of the frequency-domain stability assessments. Unstable grid situations, indicated by a red square, are identified by the Nyquist plots of the MMC and AC grid impedance ratio around $(-1, j0)$. The full Nyquist plots are given in appendix A.5 in Figure A.8 to A.14. In addition, the PM of the stable grid situations is calculated to identify grid situations close to instability and provide a stability margin. PMs below 10° are indicated by a light red square, PMs between 10° and 20° are depicted by a light green square while the remaining grid situations with PMs above 20° are shown as dark green squares in the stability matrix.

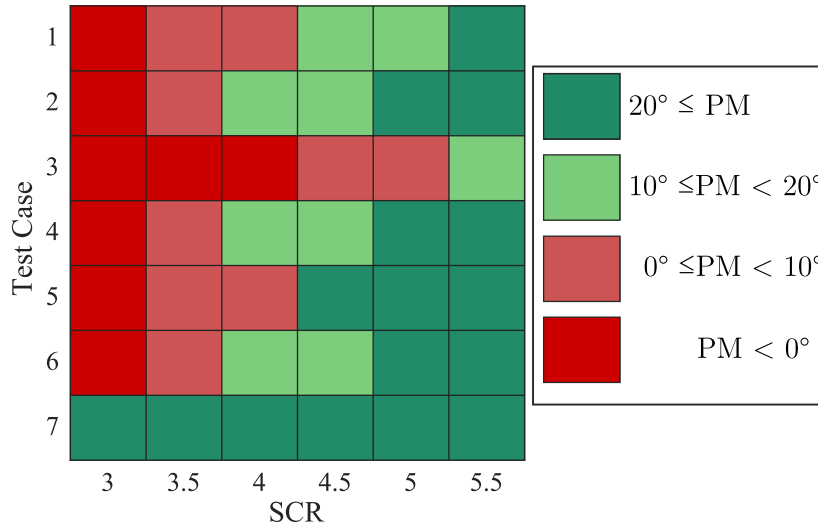


Figure 5.11: Stability matrix of conducted test cases.

Whereas the stability matrix in Figure 5.11 illustrates the overall stability assessment, Figure 5.12 to Figure 5.18 show how stable and unstable grid situations are identified by depicting the Nyquist plots and the Bode plots of the test cases. Unstable grid situations are ascertained if either the positive-sequence (solid line) or the negative-sequence (dashed line) Nyquist plots of the loop gains, derived for each SCR, encircle the point $(-1, j0)$ in the complex plane. The PMs are obtained from the Bode plots for each SCR. For simplicity, only the rightmost light-red SCR of the stability matrix is depicted in the Bode plots, showing the first system close to instability. While the positive- and negative-sequence AC grid impedance is the same, the MMC impedances differ. Therefore, the positive-sequence impedance, (MMC^p) , and the negative-sequence impedance, (MMC^n) , are shown in one Bode plot. Determining the phase difference of AC grid and MMC impedance (depicted by crosses) at the magnitude intersection frequency (depicted by a circle) for both positive- and negative-sequence results in the PMs where the minimum PM determines the stability assessment depicted in Figure 5.11.

Test case 1 is used as the reference test case, with the MMC operating in grid-following

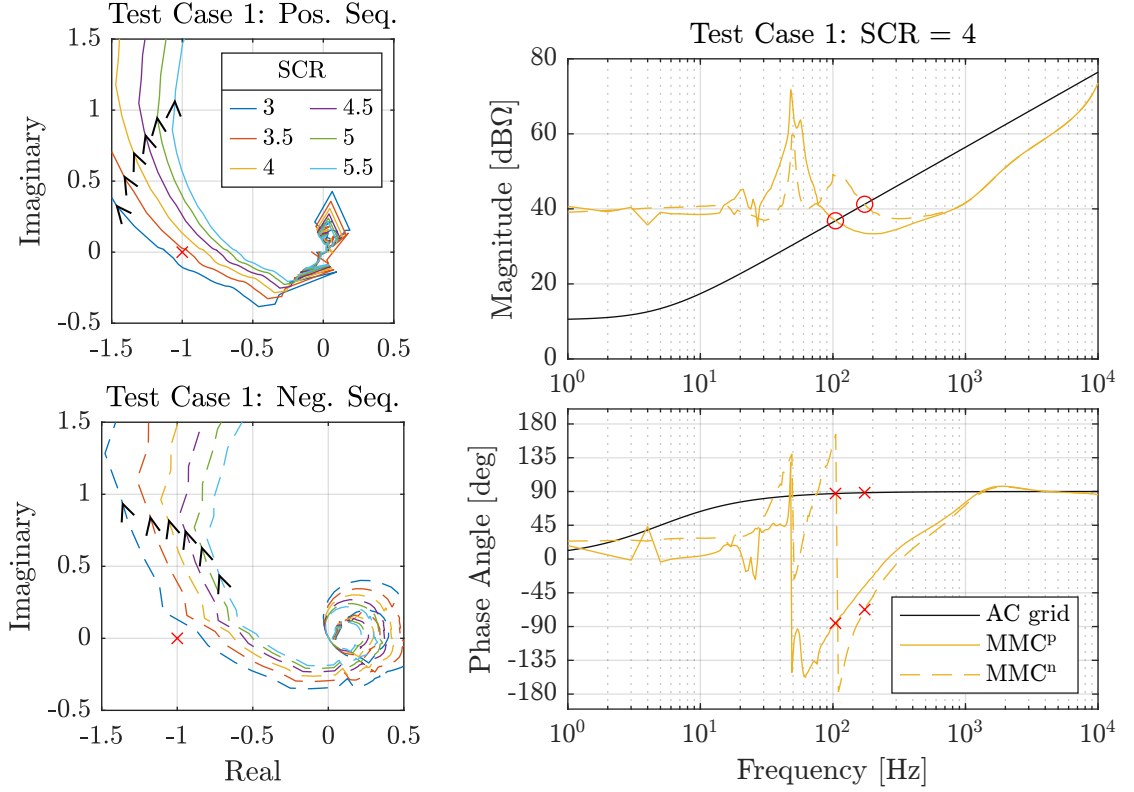


Figure 5.12: Nyquist plots and Bode diagram of test case 1.

control mode while controlling the active power to 0.5 pu and the reactive power to zero. The results of the frequency domain stability analysis can be seen in Figure 5.12. As illustrated in the stability matrix depicted in Figure 5.11, the test case demonstrates an unstable grid situation for a SCR = 3. The Nyquist plots show that the positive- but not negative-sequence impedance ratio encircles the $(-1, j0)$ point for a SCR = 3 and therefore causes the instability. The adjacent bode diagram depicts the AC grid impedance as well as the positive- and negative-sequence MMC impedance for a SCR = 4. This SCR results in a very low PM that can lead to a marginally stable system with sustained oscillations. Intersecting the grid impedance magnitude around 100 Hz for the positive-sequence MMC impedance, the corresponding phase difference of AC grid and MMC impedance is already close to 180° due to the MMC impedance phase angle that reaches values below -90° for decreasing frequencies. In contrast, the negative-sequence MMC impedance intersects the grid impedance at a frequency close to 200 Hz due to a bump in the impedance magnitude around 100 Hz. Consequently, the phase difference is significantly smaller compared to the positive-sequence impedance.

Varying the reference variable for the active power control, test case 2 investigates the stability for a MMC with double the active power being transmitted. The Nyquist plots in Figure 5.13 show that, in contrast to test case 1, the negative-sequence impedance ratio indicates the unstable system for a SCR = 3. However, the Nyquist plot of the positive-sequence impedance ratio also almost encircles $(-1, j0)$, indicating how close also the positive-sequence impedance ratio is to instability. The Bode plot analysis

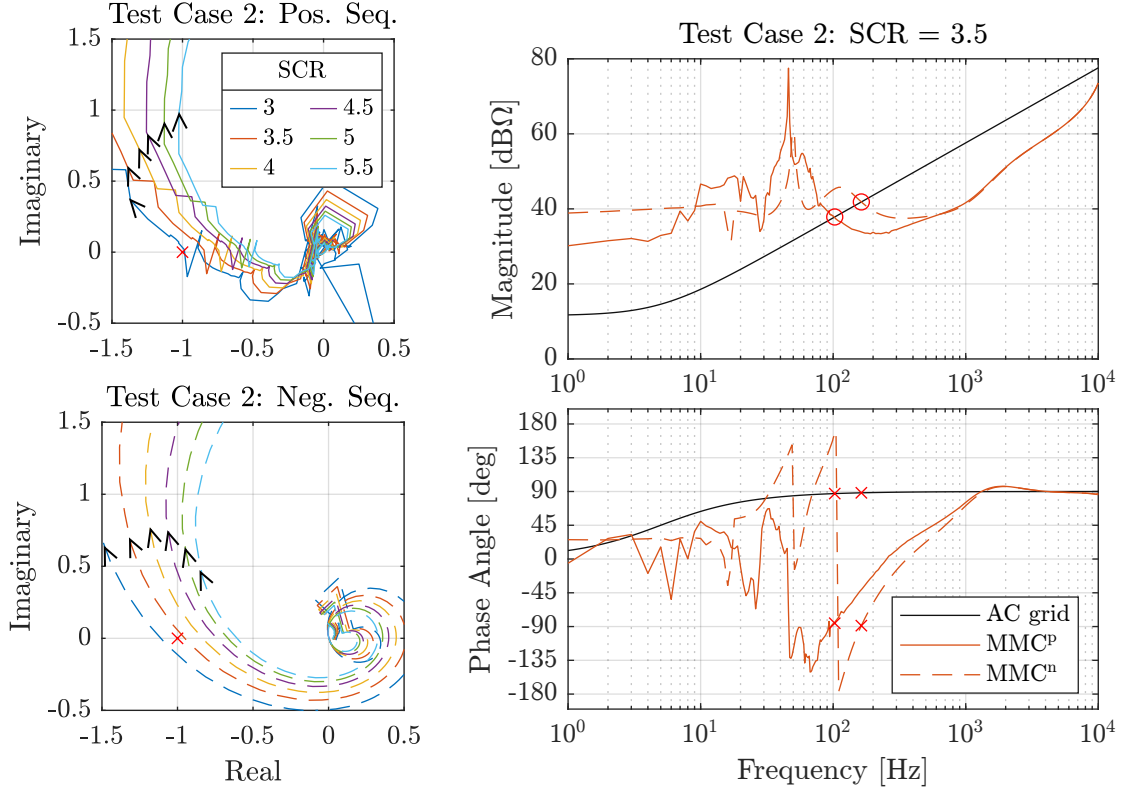


Figure 5.13: Nyquist plots and Bode diagram of test case 2.

confirms that both positive-sequence and negative-sequence MMC impedances are close to instability for an AC grid impedance with $\text{SCR} = 3.5$. Whereas the positive-sequence impedance magnitude intersects that of the grid impedance at lower frequencies, the negative-sequence impedance phase angle is comparably lower, resulting in similar phase differences. Whereas only slightly impacting the impedance trajectory, the test case shows that an increased power flow setting can result in different PMs with the stability matrix showing a PM below 10° only for an AC grid impedance with $\text{SCR} = 3.5$ instead of 4.

Test case 3 demonstrates the impact of a power flow reversal with the MMC controlling the active power to -0.5 pu compared to 0.5 pu as in the reference test case. The Nyquist plots in Figure 5.15 show for the positive-sequence impedance ratio an encirclement of the $(-1, j0)$ point for a grid impedance with $\text{SCR} = 4$. The corresponding Bode plots further demonstrate that even for a grid impedance with $\text{SCR} = 4$, the system's stability is threatened due to converter control interactions as the phase difference reaches 180° and the resulting PM is close to zero.

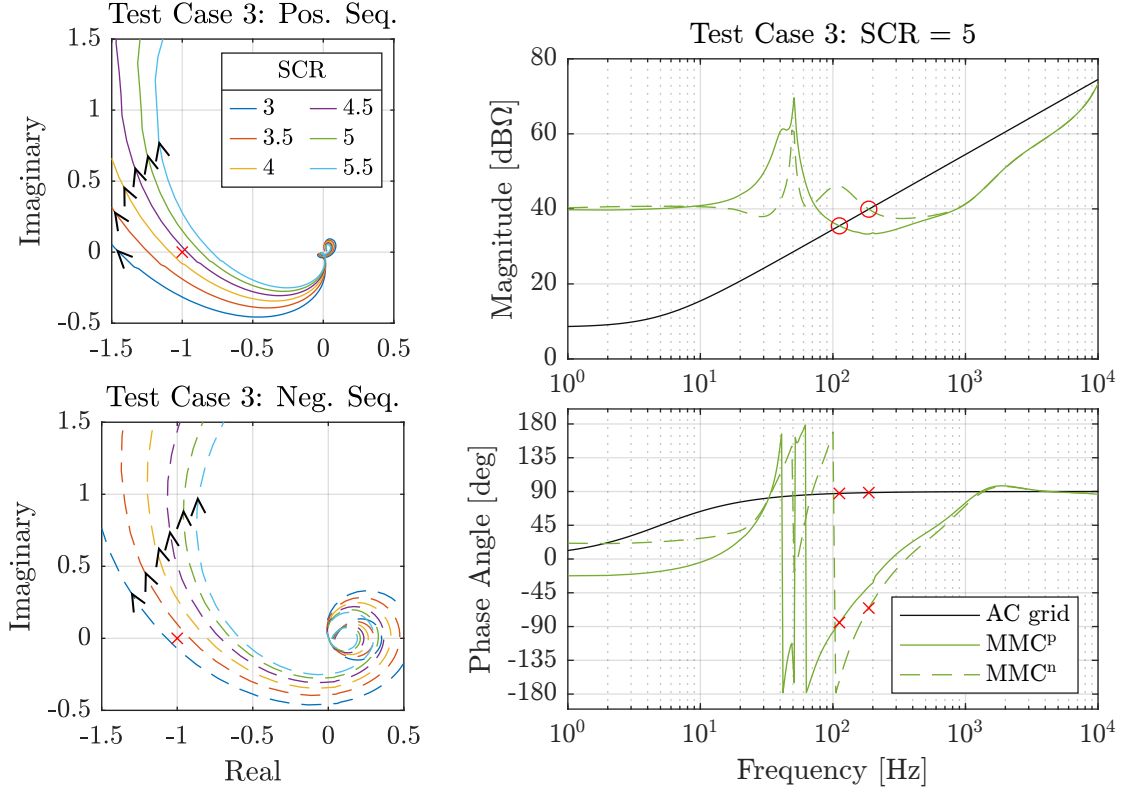


Figure 5.14: Nyquist plots and Bode diagram of test case 3.

Test case 4 has the same power flow settings as the reference test case. However, the MMC controls the AC voltage at the converter terminal instead of the reactive power flow. The results of the stability assessment can be seen in Figure 5.15. It can be seen that changing the outer control loops causes a different frequency behavior of the MMC. In particular, the magnitude of the positive-sequence impedance has a differing trajectory, having a local minimum around the fundamental frequency instead of a maximum, as can be seen in the other test cases. Also, this system becomes unstable for an AC grid impedance with $SCR = 3$; the Nyquist plots show that the encirclement of the $(-1, j0)$ point results from the negative-sequence impedance ratio. Moreover, a low PM and therefore low stability margin can be seen for a slightly lower SCR. Thus, there is also a change in the reference value for the i_q current of the upper-level controls that impacts the stability of the system in certain grid situations.

Test case 5 investigates the stability when the MMC controls the DC voltage instead of the active power. To ensure a comparable power flow, the second MMC connected to the DC line controls the active power to 0.5 pu. Both positive- and negative-sequence impedance ratios indicate an unstable system for a $SCR = 3$ as it can be seen in the Nyquist plots of Figure 5.16. The Bode plots further demonstrate a low stability margin for a system with an AC grid impedance with an $SCR = 4$. Although having different impedance trajectories, the stability assessment shows the same results as the reference test case due to similar phase angles at the intersection frequency.

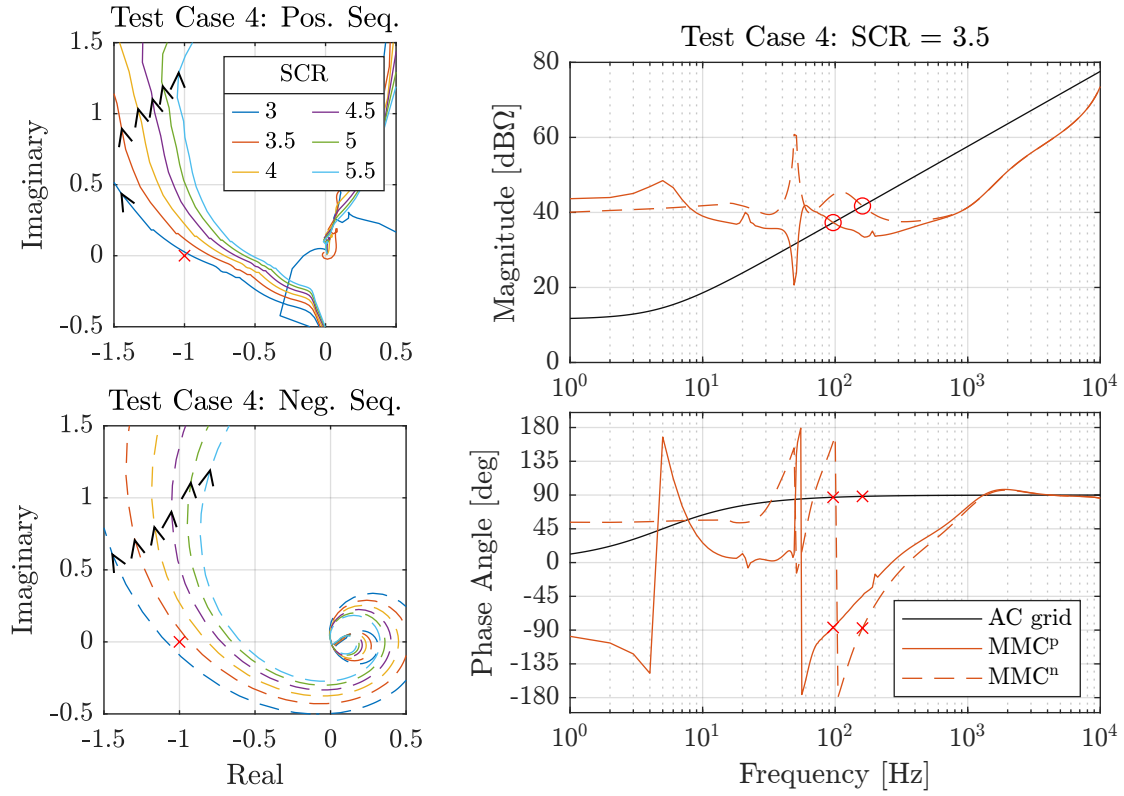


Figure 5.15: Nyquist plots and Bode diagram of test case 4.

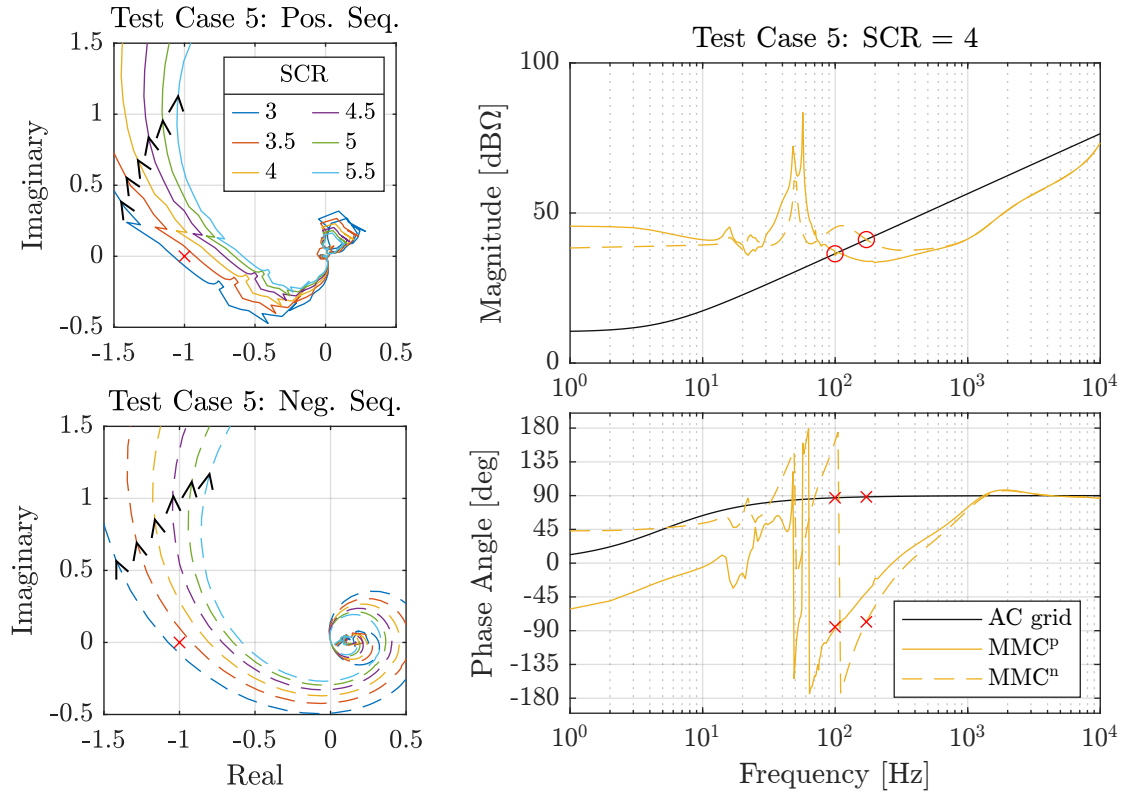


Figure 5.16: Nyquist plots and Bode diagram of test case 5.

Controlling the DC voltage and AC voltage is the basis of test case 6 that is shown in Figure 5.17. Analogous to test case 4, it can be seen that the AC voltage control causes the positive-sequence impedance to decrease initially and then increase after 50 Hz. The test case has a significantly deviating frequency behavior compared to the other control modes, resulting in additional intersection frequencies around the fundamental frequency. However, this does not affect stability assessment. The phase difference at these intersection frequencies is not close to 180° . Therefore, they do not indicate an unstable system as also shown in the Nyquist plots that do not clockwise encircle $(-1, j0)$ for a SCR = 4.5.

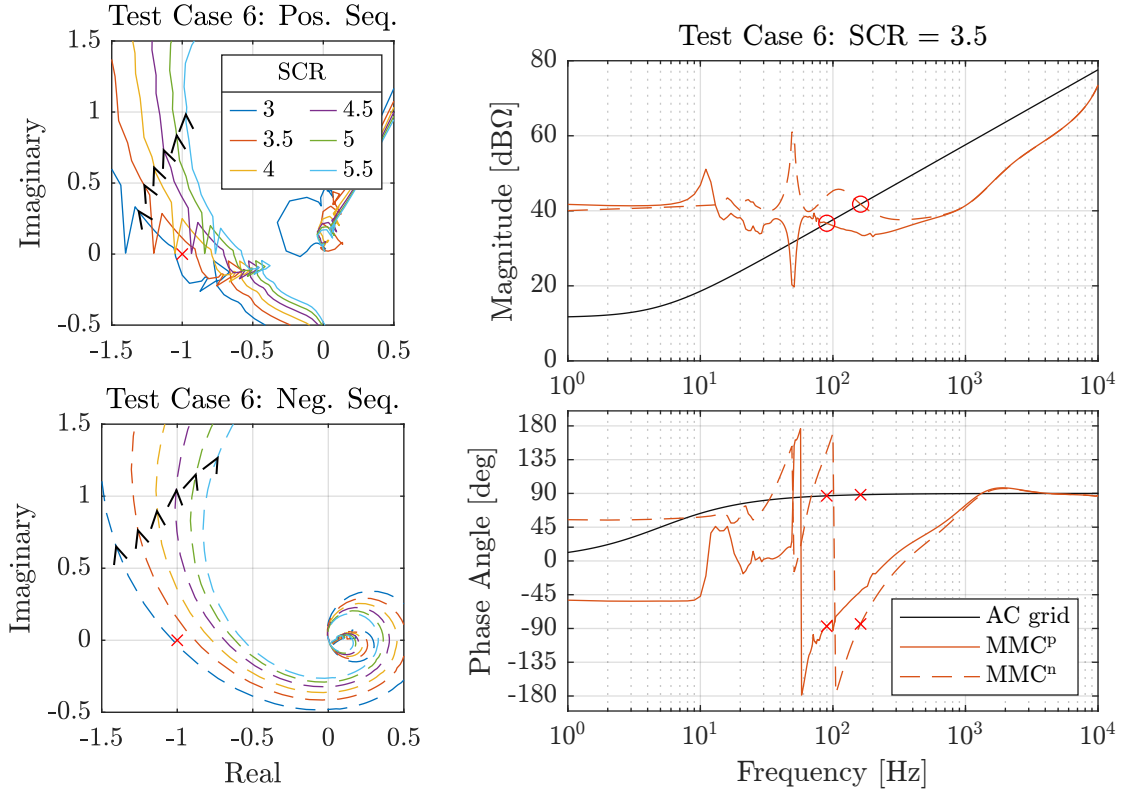


Figure 5.17: Nyquist plots and Bode diagram of test case 6.

Lastly, test case 7 investigates the stability of an MMC operating in grid-forming control mode, providing the AC voltage magnitude and frequency. The Nyquist plots in Figure 5.17 demonstrate that even for an AC grid impedance with SCR = 3, $(-1, j0)$ is not encircled and the overall system is assumed to be stable. Examining the Bode plots further reveals that the MMC phase angles of the MMC impedances never fall below -90° . Because the phase angle of the AC grid impedance does not exceed 90° , no negative PM and therefore unstable grid situation can be reached. Moreover, phase angles are only close to -90° near the fundamental frequency so that no low PM are observed. Thus, the test case demonstrates that for an AC grid that is dominated by inductive behavior, the grid-forming control mode of the MMC does not cause unstable grid situations.

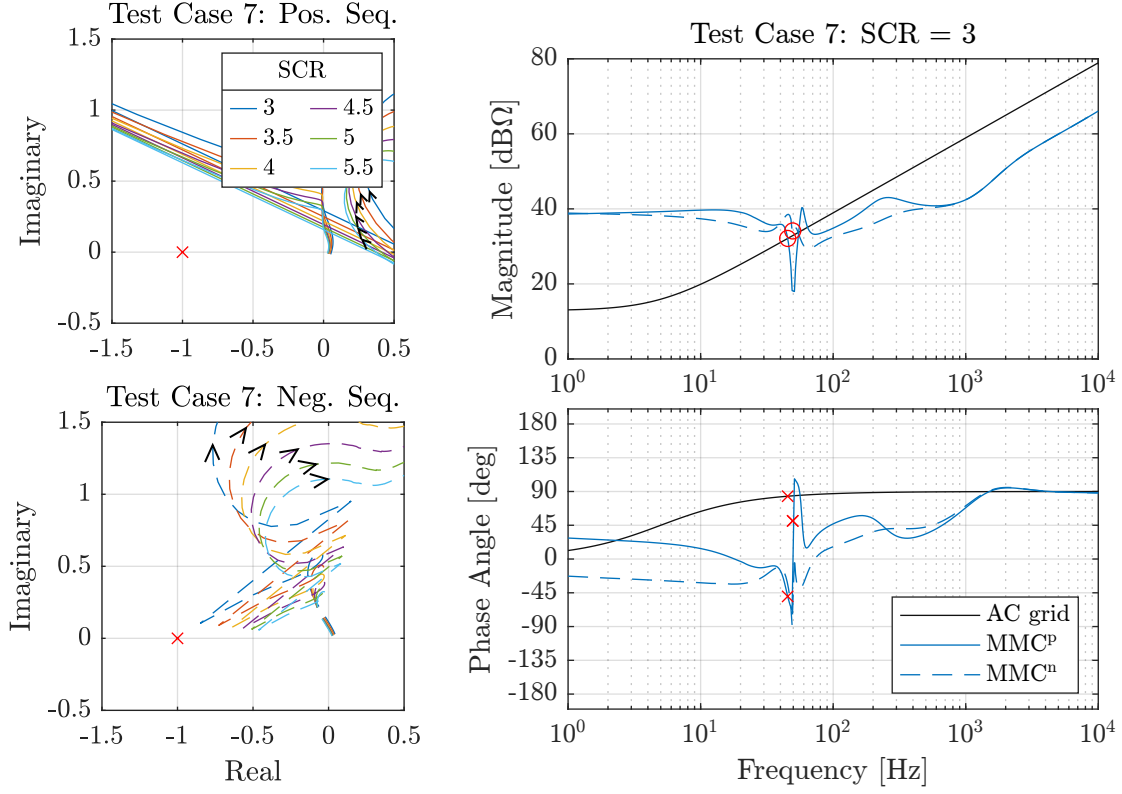


Figure 5.18: Nyquist plots and Bode diagram of test case 7.

5.3.2 Conclusion

Investigating converter control interactions in the frequency domain using the IbSC makes it possible to assess the stability of a large number of test cases due to insignificant computation times. As a result, the executed test cases are used to assess the stability of different grid situations and identify those with low or negative PMs. Based on the conducted test cases, the following conclusions can be made:

- The magnitude of the MMCs impedance in grid-following control mode intersects that of the AC grid at frequencies between 90 Hz and 200 Hz when the SCR is below 5. Due to the phase angle of the MMCs that is potentially below -90° in these frequency regions, the phase difference of MMC and AC grid can exceed 180° and lead to an unstable system. Hence, it has been shown that MMCs in grid-following control mode are prone to interact with an inductive AC grid and low short-circuit power.
- The outer control loops of the upper-level controls influence the impedance in the frequency regions where the MMC magnitude intersects the AC grid magnitude so that the stability of the AC grid depends not only on the SCR but also on the control mode of the MMC and therefore, should not be neglected in the impedance models.
- The impedance trajectory also depends on the operation point of the MMC. In

particular, a power reversal can cause interactions with AC grids at a lower or a higher short-circuit power.

- The stability assessment can be influenced by the frequency-coupling effect that plays a role in the frequency regions, in which interactions occur. Consequently, modeling of the coupling effect should be considered for stability analysis.
- MMC control systems should be tuned that they do not reach phase angles below -90° when the MMC is known to operate in weak and inductive AC grids.
- Having no PLL for grid synchronization and current controllers, the impedance trajectory of the MMC in grid-forming control mode deviates significantly from that of MMCs in grid-following control mode. In particular, the phase angle does not fall below -90° so that also for AC grids with low short-circuit power, no interactions with the control system occur.
- The test cases demonstrate interactions at frequencies below 100 Hz that explain the oscillations at super-synchronous frequencies observed in the China Southern Grid [25]. Moreover, the results support the authors' argument that weak grids can cause interactions at these frequencies, and that they can be avoided with increasing SCR. Contrarily, the high-frequency oscillations cannot be explained by the conducted test cases. However, in the literature, delays due to signal processing of real controllers are considered as cause for these oscillations [172]. Because the developed method can be applied to converter control replicas, it can be used to identify these cases.

6 Summary and Outlook

6.1 Summary

The increasing share of HVDC systems in the transmission system has shown that VSCs, in particular MMCs, can interact with the AC grid and cause oscillations at a wide range of frequencies, leading to a new electromagnetic stability phenomenon. Several incidents at offshore wind farms, onshore HVDC links, and systems with STATCOMs showed the potential of the converter control system to interact with the AC grid. Although investigations show that the control system is mainly responsible for these interactions, converter manufacturers do not disclose them due to IPR concerns. Using a converter control replica, the TSO can perform tests and studies based on the original converter control system. Nonetheless, the control system remains a black-box to the system operator. Consequently, methods and approaches are required to investigate these converter control interactions that do not require full knowledge of the converter and its control system.

Electromagnetic phenomena are mainly investigated using EMT time-domain simulation tools. However, the duration of simulations can be excessive, and any conclusions require studying the simulated waveforms, making it practically impossible to use EMT simulations for investigating a larger number of grid situations. Alternatively, converter control interactions can be investigated in the frequency domain using the IbSC. It models two potentially interacting systems as frequency-dependent impedances and subsequently assesses the stability by applying the Nyquist-criterion or analyzing the Bode plots of the subsystems' impedance ratio. This approach allows for direct and straightforward stability assessments and also provides stability margins for stable systems. Moreover, numerous investigations can be performed due to insignificant computational times and the suitability for automation. The impedance models can be derived analytically based on the converter structure and control diagrams. Contrary, numerical derivation approaches can perturb a converter at different frequencies and derive the frequency-dependent impedance based on current and voltage signals. The latter also facilitates the use of encrypted, black-boxed systems, such as converter control replica. The impedance models can be derived in the dq-domain or positive and negative sequence to account for the three-phase nature of the power system. Whereas in the dq-domain, coupling effects are considered through the off-diagonal elements in the impedance matrix, the impedance models need to be aligned with a common reference angle that makes it difficult to be used in practical applications when the models are black-boxed and not developed by the same entity. Because the stability can be assessed interdependently for positive and negative sequences, the sequence domain offers a more straightforward approach that can also be used when different parties derive the impedance models. However, it needs to

be monitored carefully if coupling effects are to be neglected or need to be additionally dealt with.

In this thesis, a numerical method is developed to derive positive and negative-sequence impedance models that are used to assess stability and interactions in the frequency domain. The method can be applied both to physical laboratory MMCs as well as to time-domain converter models. It is based on an online approach that perturbs the MMC sequentially for a time period and determines the frequency-dependent impedances at the same time. Determining the impedance online and not in post-processing makes it possible to significantly reduce the recorded data's sampling time, and therefore the file sizes while ensuring high accuracy. That way, numerous impedances can be derived for different operating points and controller settings of the MMC. The method and its accuracy are verified by comparing the magnitude of a derived impedance to that of a standard method using an FFT algorithm that calculates the impedance using high-resolution voltage and current data. Frequency coupling is considered by additionally recording the coupling current, $i(f_{grid} - 2f_{pert})$, that is used to model the coupling effect as an additional parallel admittance. If impedance models that capture the frequency behavior of physical laboratory-scaled MMC are intended to be used for interaction studies, a method is developed that can be used to scale up the impedance according to the electrical parameters of the converters.

Comparing the impedance of a small-scale laboratory MMC, the MMC TB, to the impedance obtained from a Type 5 MMC model shows that the model can be used to represent the electrical structure of the converter when interactions are assessed that cause oscillations between 30 Hz and 3 kHz. The comparison further reveals that the converter transformer causes a deviating frequency behavior at sub-synchronous frequencies that cannot be seen in the EMT model.

The validity of stability assessments in the frequency domain using the IbSC is demonstrated by comparing the frequency-domain stability and instability predictions to results of equivalent time-domain simulations. The comparison shows that the IbSC is able to predict unstable cases accurately and predict the frequencies at which the oscillations occur. Seven test cases are investigated to identify unstable grid situations for different controller settings and operation modes of the MMC. It is shown that, dependent on the AC grid strength, the grid-following control system of the converter interacts with the grid, causing oscillations below 100 Hz. The results further demonstrate that the stability also depends on the power flow as well as the outer control loops of the upper-level controls. Moreover, when in grid-forming control mode, the MMC remains stable for lower grid strengths that lead to interactions when the MMC is operating in grid-following control mode.

6.2 Outlook

The developed method can be used to investigate converter interactions, considering the frequency behavior of real control systems. For this purpose, the measuring method can be executed on a RTS and coupled with control replica systems that contain the actual controllers of a MMC station.

Moreover, the developed method can be used to investigate interactions between MMCs and large-scale AC grids, with consideration to the impact of the topology. This requires an interface and aggregation algorithm that defines various interfaces and divides a system into several potential subsystems. Aggregating the two systems' impedance models into two impedances that represent each subsystem makes it possible to assess the overall system stability by consecutively assessing the stability for each subsystem division and the resulting aggregated impedances. An interface algorithm has been developed to assess an offshore wind farms' stability with radial topology using a recursive algorithm [173]. However, onshore AC grids require different approaches to account for the transmission system's meshed structure.

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List of Publications

Journal Articles

1. Andreas Roehder, Bernhard Fuchs, Janek Massmann, **Matthias Quester**, Armin Schnettler. "Transmission System Stability Assessment Within an Integrated Grid Development Process", in *Cigré Science & Engineering*, Jun. 2017.
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6. **Matthias Quester**, Viswaja Yellisetti, Fisnik Loku, Ralf Puffer. "Assessing the Impact of Offshore Wind Farm Grid Configuration on Harmonic Stability", in *2019 IEEE Milan PowerTech*, Milan, Italy, Jun. 2019.
7. Mashhood Raza, Markus Kaiser, **Matthias Quester**. "Comparison of Electromagnetic and Eletromechanical Transient Simulations for AC Power System", in *2019 54th International Universities Power Engineering Conference (UPEC)*, Bucharest, Romania, Sept. 2019.
8. **Matthias Quester**, Fisnik Loku, Viswaja Yellisetti, Ralf Puffer. "Online Impedance Measurement of a Modular Multilevel Converter", in *2019 IEEE PES Innovative Smart Grid Technologies Europe (ISGT-Europe)*, Bucharest, Romania, Sept. 2019.

9. **Matthias Quester**, Fisnik Loku, Viswaja Yellisetti, Albert Moser. “Frequency Behavior of an MMC Test Bench System”, in *2020 6th IEEE International Energy Conference (ENERGYCon)*, Tunis, Tunisia, Sept. 2020.
10. Sandor Simon, Philipp Linnartz, Dominik Willenberg, **Matthias Quester**, Antonello Monti. “Experimental investigation of mirror frequency currents emitted by low voltage inverters under distorted grid voltage conditions”, accepted at *IEEE PES PowerTech Madrid 2021*, 2021.
11. Sandor Simon, **Matthias Quester**, Sriram Karthik Gurumurthy, Antonello Monti. “Experimental Measurements of Frequency-Dependent Active Component Impedances for Harmonic Stability Assessment”, accepted at *CIREN 2021 Conference*, 2021.
12. Viswaja Yellisetti, David Seibert, **Matthias Quester**, Albert Moser. “Vector fitting algorithms for impedance modelling of converter based systems”, accepted at *2021 56th International Universities Power Engineering Conference (UPEC)*, 2021.

List of Supervised Theses

The following theses have been written during my time as research associate: I want to thank all students for their efforts and research. In particular, I want to thank Viswaja Yellisetti, Otmane El Azzati, and Xinrui Dong for their research and work in helping me to investigate converter control interactions.

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- Hui Zhou, Investigation on HVDC-Diode Rectifiers Connected to Offshore DC Grids, 2017.
- Chenbo Xu, Modeling of a Voltage Sourced Converter for Harmonic Stability Studies, 2018.
- Viswaja Yellisetti, Harmonic Stability Analysis of Offshore Grid Connected Wind Turbines, 2018.
- Daniel Manrique-Ortiz, Investigating the Harmonic Stability of HVDC Systems Connecting Offshore Wind Farms, 2019.
- Xinrui Dong, Investigating the Frequency Behaviour of Modular Multi-Level Converters, 2020.

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- Jonas Bonus, Untersuchung der Auswirkungen von Netzanschlussregeln von HVDC-Systemen auf die Systemstabilität des Übertragungsnetzes, 2016.
- Paul Raffelsiefen, Untersuchung von Resonanzfrequenzen in Offshore Windkollektornetzen, 2017.
- Carsten Docktor, Technische Analyse zur Messung und Simulation von Oberschwingungseinträgen durch wechselrichterbasierte Erzeugungsanlagen in Mittel- und Hochspannungsnetzen, 2017.
- Lothar Brixius, Numerische Umrichtermodellierung zur Bewertung der harmonischen Stabilität im Frequenzbereich, 2019.
- Otmane El Azzati, Investigating Harmonic Stability Problems of Modular Multi-Level Converters, 2020.

List of Figures

1.1	Voltage measurements at TenneT wind farm [18].	3
1.2	On-site voltage measurements INELFE project [20].	4
1.3	1270 Hz resonance at the Luxi back-to-back VSC-HVDC system [23]. . . .	4
1.4	2.5 Hz and 97.5 Hz oscillations in a transmission system of China Southern Grid [25].	5
1.5	MMC electrical structure and control system.	6
1.6	Research questions and thesis approach.	10
2.1	Power system dynamic phenomena classification [48].	12
2.2	Power system stability classification [50].	13
2.3	Additional stability categories due to the increase of power electronic interfaced technologies [56].	15
2.4	EMT simulation showing unstable system.	19
2.5	State-space of a small-signal unstable system.	23
2.6	Converter connected to AC grid (a) and equivalent circuit diagram (b). .	24
2.7	Equivalent circuit seen as feedback loop.	25
2.8	Nyquist plot of unstable system.	26
2.9	Bode plot of an unstable system.	27
2.10	Potential interfaces defining subsystems for impedance-based approach. .	28
2.11	dq-domain circuit.	34
2.12	Positive-sequence circuit (a) and negative-sequence circuit (b).	36
2.13	Average value model of a grid-connected 2-level VSC.	37
2.14	Electrical structure of an MMC.	42
2.15	Investigation approach.	50
3.1	MMC TB setup.	53
3.2	Power Amplifier (a) [136] and MMC rack (b) [137].	54
3.3	Type 5 model representation of the electrical structure of an MMC. . . .	57
3.4	Equivalent circuit diagram of AC grid model.	58
3.5	High-level overview of the MMC control system.	59
4.1	Method Overview.	64
4.2	Impedance measurement perturbation frequency series.	65
4.3	Voltage and current perturbation circuit.	66
4.4	Components of the implemented DDS sine wave. Adapted from [159]. . .	67
4.5	Developed phase accumulator logic.	68
4.6	Phase accumulator increments for f_{pert} and $2f_{pert}$	69
4.7	1 kHz sine wave created by CPU and FPGA.	69
4.8	DFT window.	70

4.9	Fourier implementation.	71
4.10	FPGA design of signal processor.	72
4.11	Impedance calculation of an exemplary load.	74
4.12	MMC Test Bench at RWTH Aachen University.	75
4.13	MMC Test Bench impedance measurement setup.	75
4.14	Simulated MMC impedance measurement setup.	76
4.15	MMC impedance magnitude determined by the developed measuring method and by the voltage and current spectrum obtained from a FFT algorithm in post-processing.	77
4.16	MMC current response spectrum with positive-sequence (a) and negative- sequence perturbation (b).	80
4.17	MMC current response to a voltage perturbation with frequency f_{pert} . . .	81
4.18	MMC voltage response spectrum with positive-sequence (a) and negative- sequence perturbation (b).	82
4.19	MMC voltage response to a current perturbation with frequency f_{pert} . . .	82
4.20	Influence of coupling current on perturbation current [169].	83
4.21	Shifting of admittances as in (4.31).	85
4.22	MMC impedance model with and without integration of coupling effect. .	86
4.23	Exemplary AC grid impedance.	87
5.1	MMC TB and LS model impedance for grid-following and grid-forming control.	90
5.2	MMC TB and LS model impedance with and without transformer.	92
5.3	Equivalent circuit diagram of AC grid connected to MMC.	94
5.4	MMC impedance and AC grid impedance for different SCRs.	94
5.5	MMC impedance and AC grid impedance for different SCRs.	95
5.6	Close up of MMC and AC grid impedance intersections.	96
5.7	Nyquist plots of AC grid and MMC impedance ratio.	97
5.8	AC terminal current when MMC is connected to AC grids with decreasing SCR.	97
5.9	AC terminal current (a) and voltage (b) when switched to an AC grid with SCR = 4.	98
5.10	Fourier spectrum of phase a of the AC terminal current (a) and voltage (b) when connected to an AC grid with SCR = 4.	98
5.11	Stability matrix of conducted test cases.	100
5.12	Nyquist plots and Bode diagram of test case 1.	101
5.13	Nyquist plots and Bode diagram of test case 2.	102
5.14	Nyquist plots and Bode diagram of test case 3.	103
5.15	Nyquist plots and Bode diagram of test case 4.	104
5.16	Nyquist plots and Bode diagram of test case 5.	104
5.17	Nyquist plots and Bode diagram of test case 6.	105
5.18	Nyquist plots and Bode diagram of test case 7.	106

A.1	Measurement data and fitted transfer function of the power amplifier frequency response.	139
A.2	Scaling factors $ k $ and $ k_\varphi $	141
A.3	Comparison of simulated LS and scaled up LS model positive-sequence impedance to that of the FS.	142
A.4	MMC Test Bench current response spectrum with positive sequence (a) and negative sequence perturbation (b).	143
A.5	MMC Test Bench current response to a voltage perturbation with frequency f_{pert}	143
A.6	MMC Test Bench and LS model impedance for grid-following and grid-forming control.	144
A.7	MMC TB and LS model impedance with and without transformer in grid-following control mode.	145
A.8	Test case 1 positive-sequence (a) and negative-sequence (b) Nyquist plots.	145
A.9	Test case 2 positive-sequence (a) and negative-sequence (b) Nyquist plots.	146
A.10	Test case 3 positive-sequence (a) and negative-sequence (b) Nyquist plots.	146
A.11	Test case 4 positive-sequence (a) and negative-sequence (b) Nyquist plots.	146
A.12	Test case 5 positive-sequence (a) and negative-sequence (b) Nyquist plots.	147
A.13	Test case 6 positive-sequence (a) and negative-sequence (b) Nyquist plots.	147
A.14	Test case 7 positive-sequence (a) and negative-sequence (b) Nyquist plots.	147

List of Tables

2.1	Categories of power quality phenomena according to [63].	17
2.2	Comparison of analysis methods.	29
3.1	Power amplifier parameters.	55
3.2	MMC Test Bench parameters.	55
3.3	Full-scale and lab-scale model parameters.	57
3.4	SCR classification.	58
4.1	Relative difference of impedance magnitude of measuring method and FFT algorithm.	77
4.2	Output data file size for a 30 s measuring series.	78
5.1	Grid impedance parameters.	93
5.2	Phase Margins for positive- and negative-sequence impedances.	96
5.3	Control and powerflow settings.	99

A Appendix

A.1 Power Amplifier Bandwidth Verification

Potentially distorting the impedance measurements, the PAs' specified bandwidth is verified by additional measurements. The resulting transfer function is compared to the specified bandwidth at -3 dB. The frequency response of the PA is obtained by applying a signal generated by a frequency generator with a bandwidth of 100 MHz to the analog input of the power amplifier. Using a 200 MHz oscilloscope, the output signals are measured. The frequency response of the PA is then determined by the ratio of output to input signals as shown in Figure A.1 [174]. Moreover, the measurement data allows for

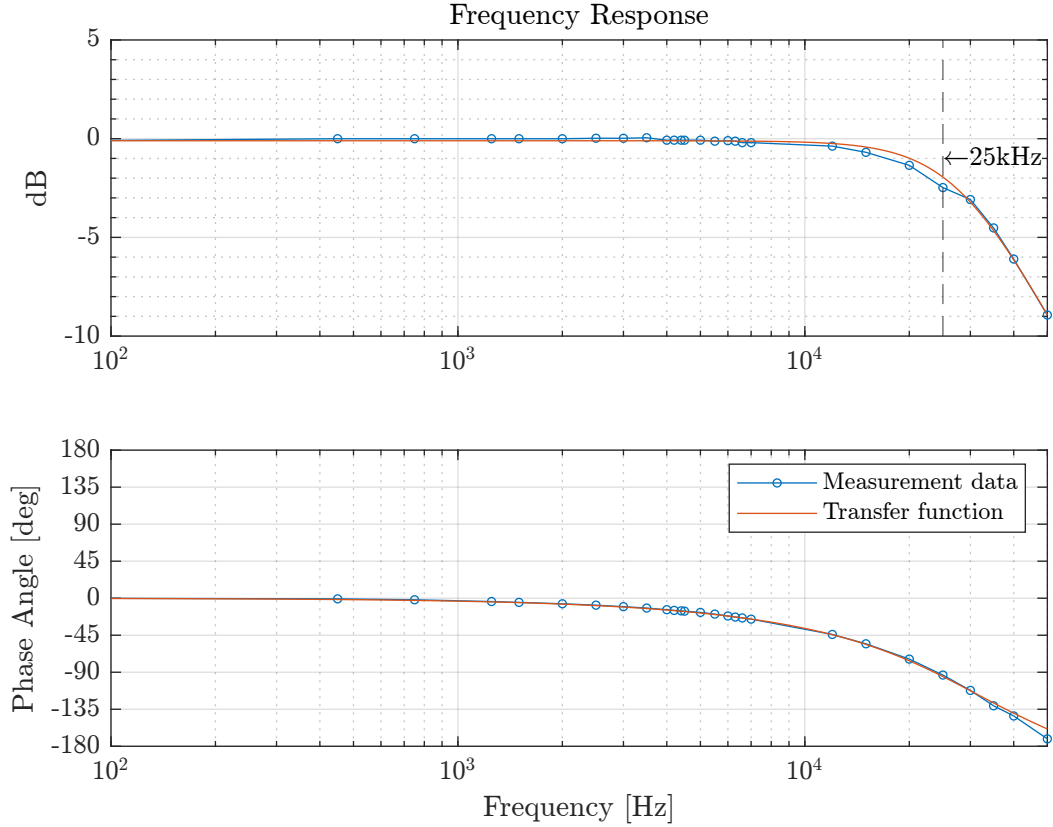


Figure A.1: Measurement data and fitted transfer function of the power amplifier frequency response.

the calculation of the PA's transfer function that is depicted in Figure A.1 as well. Using the *vector fitting algorithm*¹ [175–177], a second-order transfer function $H(s)$ is fitted to

¹Vector fitting is a numerical and publicly available method that uses measurement data such as frequency responses to approximate rational transfer functions by placing poles and residues.

the measurement data with

$$H(s) = \frac{-6.191 \cdot 10^4 \cdot s + 3.264 \cdot 10^{10}}{s^2 + 2.66 \cdot 10^5 \cdot s + 3.303 \cdot 10^{10}}. \quad (\text{A.1})$$

The results verify the specified PA parameters, demonstrating a frequency response between -2 dB and -3 dB at 25 kHz. However, it can be seen that starting with 1 kHz, the PA might affect the phase angle of a measured signal.

A.2 Scaling Up Laboratory-Scaled Impedances

The impedance models derived based on the FS MMC model can be directly applied for identifying unstable grid situations of transmission systems. In contrast, MMC TB impedance models need to be scaled up so they can be used to assess interactions with a transmission system. This is necessary in case simplified EMT models do not capture the real component behavior in frequency areas where interactions occur. Thus, an approach is developed and tested for scaling-up MMC impedances for different power ratings and voltage levels.

Using the measured MMC Test Bench impedance for assessing the stability of high voltage systems requires the scaling of the MMC TB impedance so that the impedance's magnitude and phase are similar to those of full-scale converter systems. Because the control system is in p.u., differences in the impedances depend on the electrical parameters only. Consequently, the electrical parameters of the MMC TB and its equivalent LS simulation model are related to those of the full-scale MMC model to calculate a complex scaling factor $\underline{k}$ that, when multiplied with the MMC TB impedance $\underline{Z}_{TB}$ or the LS model impedance $\underline{Z}_{LS}$, shifts the impedance into regions of an equivalent full-scale system [154]. This approach enables the use of measured MMC TB impedances for investigating interactions of MMCs with high-voltage systems. The scaling factor is divided into $|k|$ and k_φ so that for an impedance $\underline{Z} = |Z|\angle\varphi_Z$, $|Z_{scaled}| = |Z| \cdot |k|$ and $\varphi_{Z_{scaled}} = \varphi_Z - k_\varphi$. The factors can be then calculated to

$$|k| = \sqrt{\frac{r_{FS}^2 + (\omega \cdot l_{FS})^2}{r_{LS,TB}^2 + (\omega \cdot l_{LS,TB})^2}} \quad (\text{A.2})$$

and

$$k_\varphi = \arctan \frac{r_{FS}}{\omega \cdot l_{FS}} - \arctan \frac{r_{LS,TB}}{\omega \cdot l_{LS,TB}}. \quad (\text{A.3})$$

The equivalent resistances, $r_{LS,TB}$ and r_{FS} are calculated according to the resistive parameters of the LS model, the MMC TB, and the FS model, where

$$r = R_{Primary} + R'_{Secondary} + R'_{Arm} + n_{SM} \cdot R'_{on} \quad (\text{A.4})$$

The equivalent inductances, $l_{LS,TB}$ and l_{FS} , are likewise calculated according to the

inductive parameters of the LS model, the MMC TB, and the FS model:

$$l = L_{Primary} + L'_{Secondary} + L'_{Arm}. \quad (\text{A.5})$$

Variables denoted with $'$ are referred from the transformer's secondary side to the primary side. The equivalent resistance, r_{FS} and inductance l_{FS} , are calculated according to the electrical parameters of the full-scale model. In contrast, the resistance, $r_{LS,TB}$, and inductance, $l_{LS,TB}$, are based on the specifications of the MMC TB and match those of the simulated LS model. The angular frequency, ω , is a vector and equals the product of 2π and the perturbation frequency, f_{pert} , meaning that the calculated scaling factor, $\underline{k}$, is a vector as well and needs to be multiplied by $\underline{Z}$ for every measured frequency point. Figure A.2 shows $|k|$ and $|k_\varphi|$ calculated for the MMC TB, and simulated LS model according to the parameters given in Table 3.3.

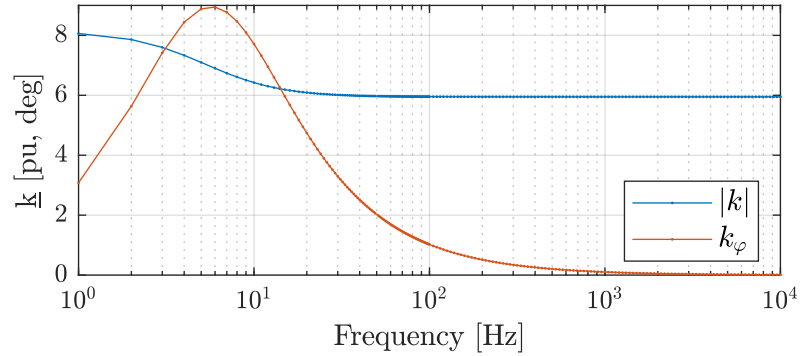


Figure A.2: Scaling factors $|k|$ and $|k_\varphi|$.

With increasing frequency, $|k|$ converges to six as the impedance is dominated by the inductive parameters, indicating a six times larger magnitude of the FS system compared to the LS. In contrast, only at low frequencies does scaling the MMC impedances affect the phase angle as it can be seen that for high frequencies k_φ converges to zero degrees. The phase angle of the impedance only deviates slightly because the ratio of the inductive and resistive parameters of the FS and LS system remains similar. For high frequencies, ω dominates $\underline{k}$ resulting in $|k|$ being dependent on l_{FS}/l_{LS} only and k_φ not being affected by the electrical parameters.

To verify the scaling approach, the impedance of the simulated LS model is scaled up and compared to that of the FS model as shown in Figure A.3 (a) for grid-following control and in Figure A.3 (b) for grid-forming control. The comparison shows that the scaling approach can shift the magnitude of the lab-scale impedance in regions of the full-scale model, demonstrating that the scaling approach can be applied to scale up the MMC TB impedances for full-scale system stability studies.

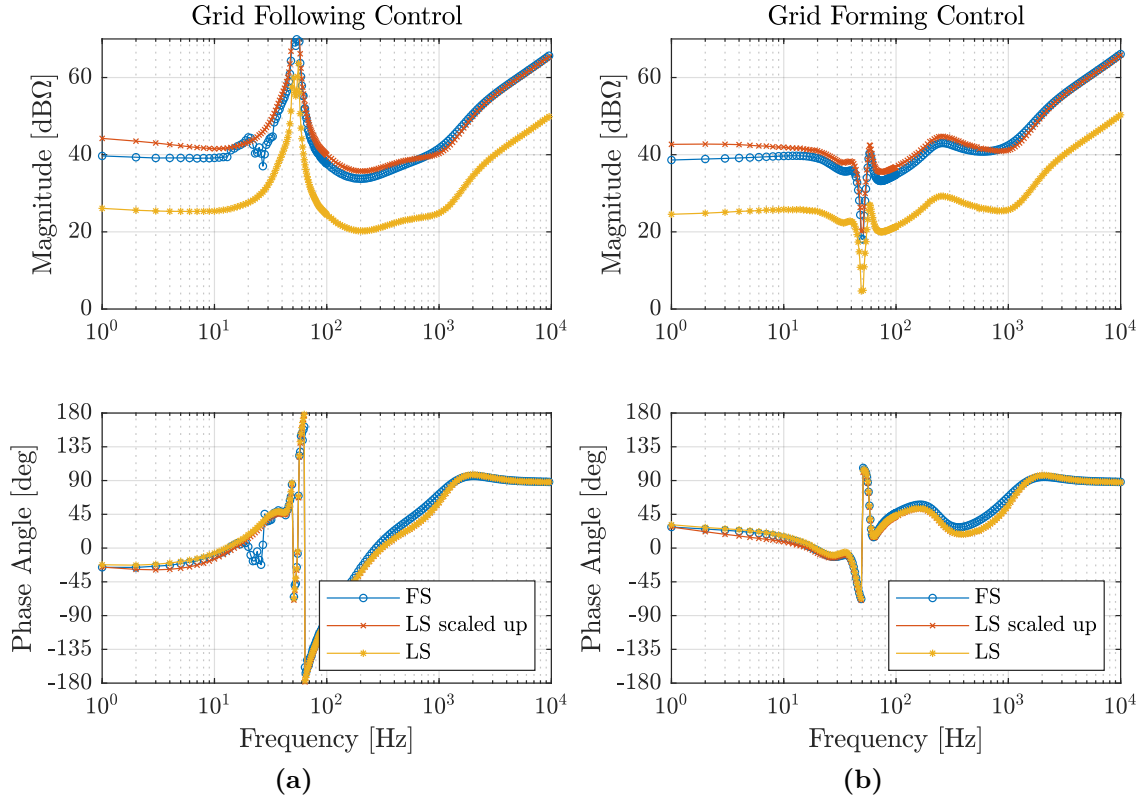


Figure A.3: Comparison of simulated LS and scaled up LS model positive-sequence impedance to that of the FS.

A.3 MMC Test Bench Frequency Response

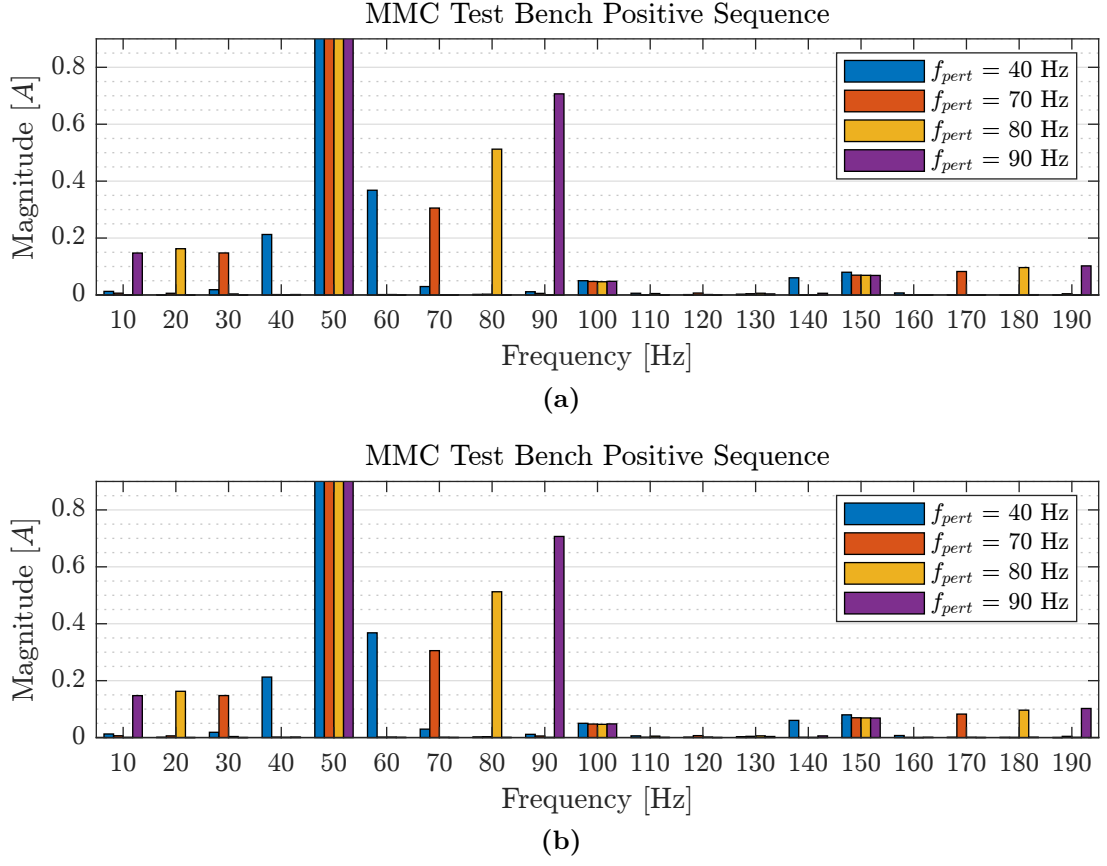


Figure A.4: MMC Test Bench current response spectrum with positive sequence (a) and negative sequence perturbation (b).

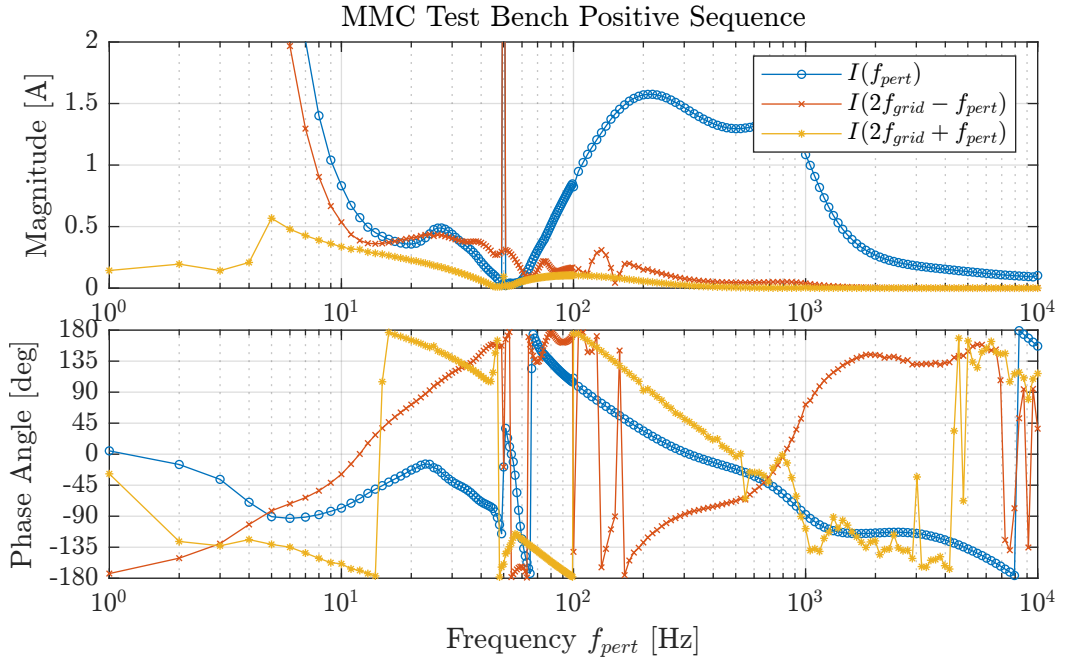


Figure A.5: MMC Test Bench current response to a voltage perturbation with frequency f_{pert} .

A.4 MMC TB and LS Model Negative-Sequence Impedances

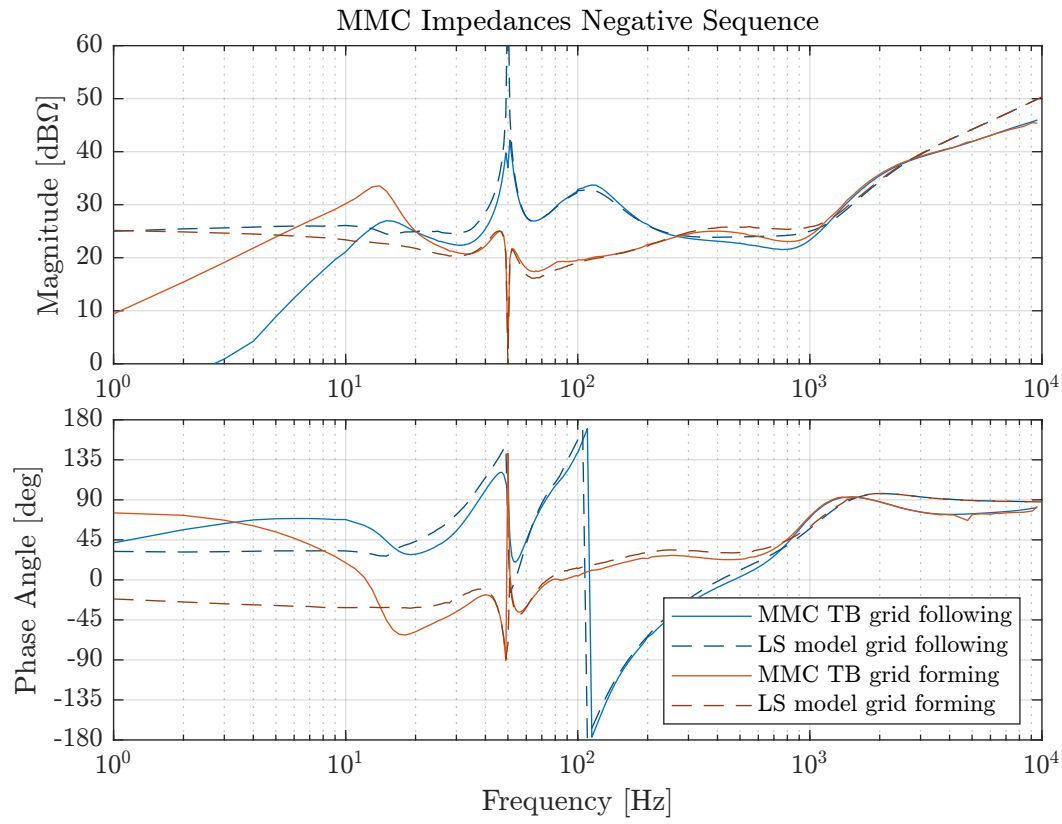


Figure A.6: MMC Test Bench and LS model impedance for grid-following and grid-forming control.

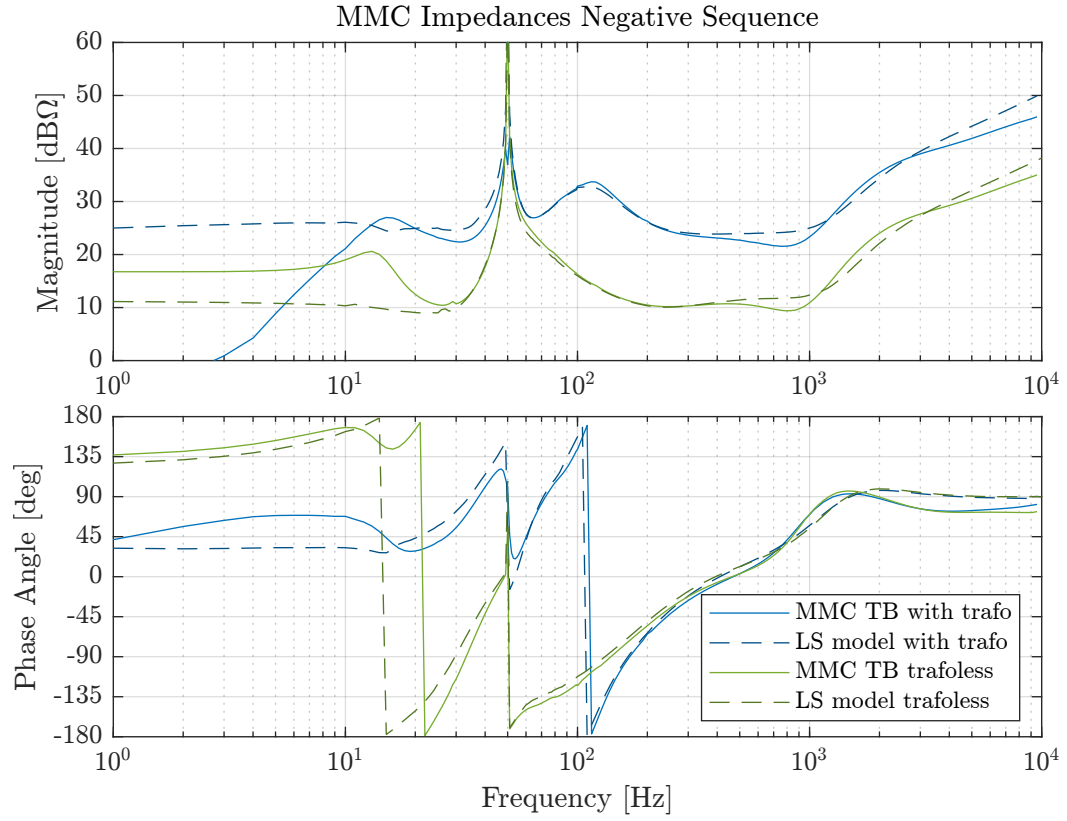


Figure A.7: MMC TB and LS model impedance with and without transformer in grid-following control mode.

A.5 Stability Assessment Nyquist Plots

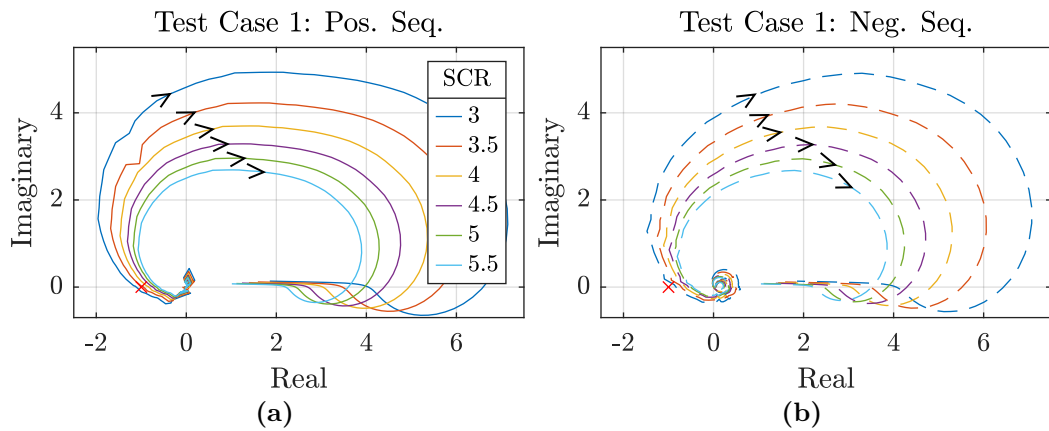


Figure A.8: Test case 1 positive-sequence (a) and negative-sequence (b) Nyquist plots.

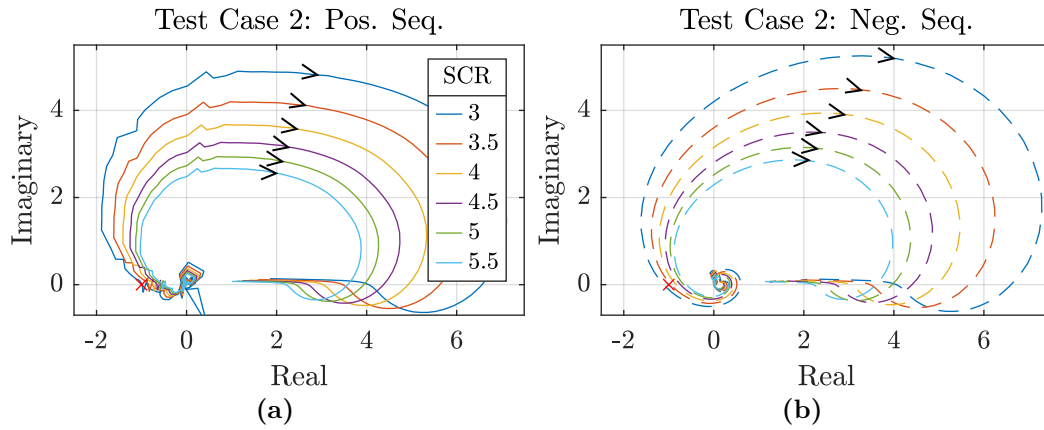


Figure A.9: Test case 2 positive-sequence (a) and negative-sequence (b) Nyquist plots.

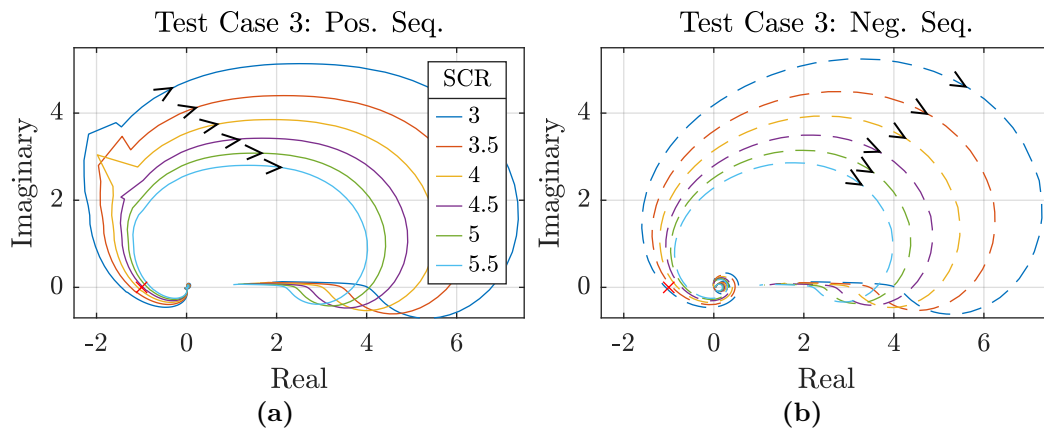


Figure A.10: Test case 3 positive-sequence (a) and negative-sequence (b) Nyquist plots.

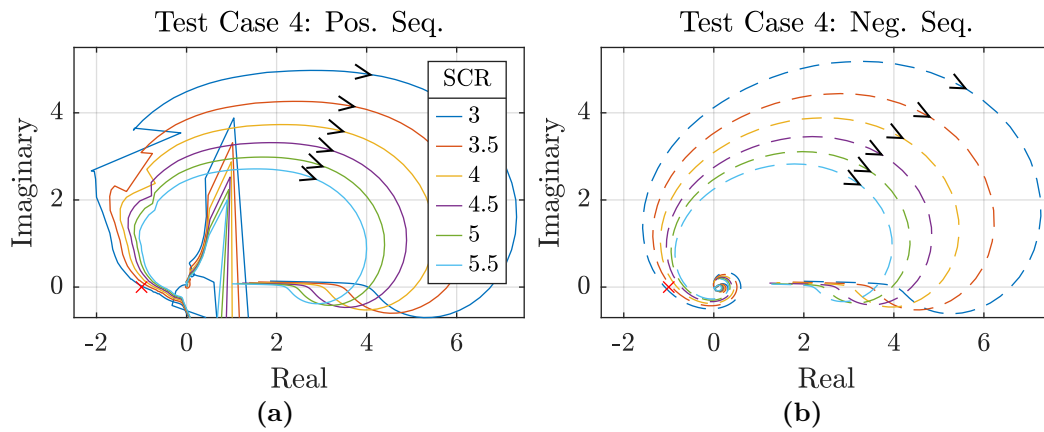


Figure A.11: Test case 4 positive-sequence (a) and negative-sequence (b) Nyquist plots.

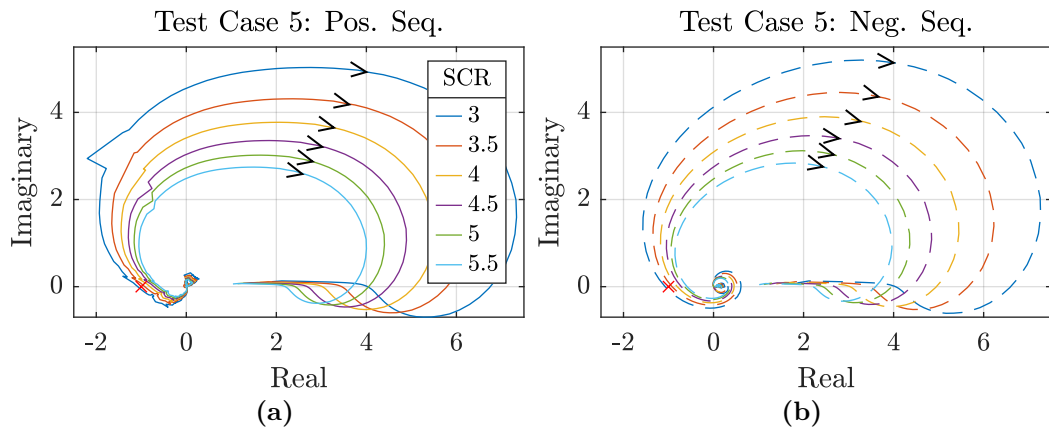


Figure A.12: Test case 5 positive-sequence (a) and negative-sequence (b) Nyquist plots.

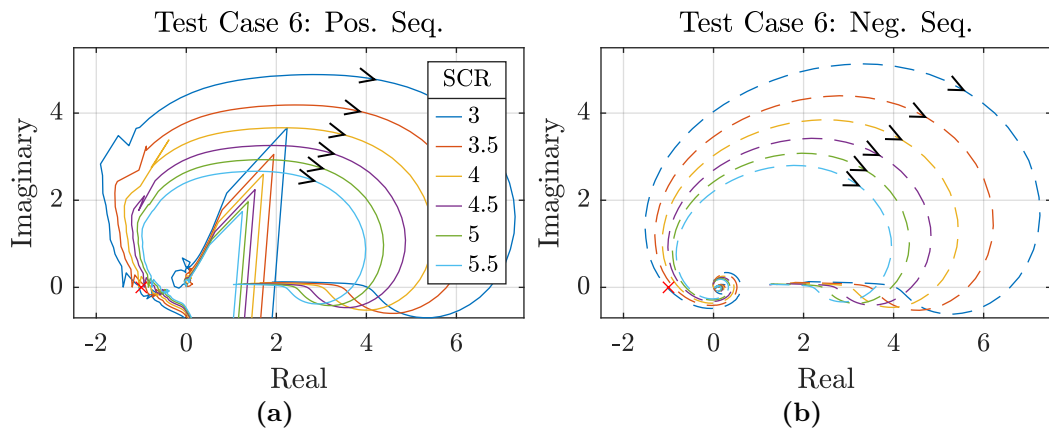


Figure A.13: Test case 6 positive-sequence (a) and negative-sequence (b) Nyquist plots.

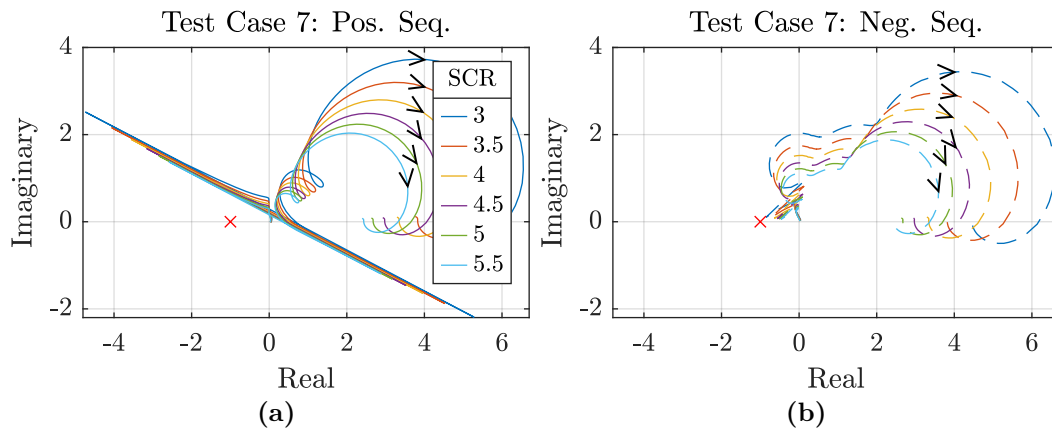


Figure A.14: Test case 7 positive-sequence (a) and negative-sequence (b) Nyquist plots.

