Investigation of DC Collection Networks for Offshore Wind Farms

Vogel, Stephan; Rasmussen, Tonny Wederberg; El-Khatib, Walid Ziad

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Investigation of DC Collection Networks for Offshore Wind Farms

Stephan Vogel

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Supervisor: Tonny W. Rasmussen
Co-supervisor: Walid Z. El-Khatib
Abstract

The possibility to connect offshore wind turbines with a collection network based on Direct Current (DC) instead of Alternating Current (AC) gained attention in the scientific and industrial environment. There are many promising properties of DC components that could be beneficial such as: smaller dimensions, less weight, fewer conductors, no reactive power considerations, and less overall losses due to absence of proximity and skin effects. On the other hand, challenges arise due to low grid impedance, high input and output current ripple of DC/DC converters, and high fault currents in case of an incident.

To provide a solid foundation for a DC collection network simulation, intensive research of electrical representations of components was performed, and suitable models were identified and implemented in the transient simulation program PSCAD. By utilizing this tool, transient effects and fault scenarios in wind turbine converters, cables, and in the High Voltage Direct Current (HVDC) link, were studied. The steady-state simulations of the system showed high ripple current due to the low grid impedance and the high input and output ripple of the converter. Three different solutions have been proposed to reduce the ripple. In general, the proposed DC grid shows a good transient response to disturbances, and steady-state conditions are regained after the faults. As expected, peak current and voltages during fault conditions reached high magnitudes up to 15 times the nominal value.

Additionally, each model of the DC grid has been evaluated regarding losses, and an overall efficiency of the DC collection network was found to be 94.9%.
Muligheden for at tilslutte offshore vindmøller med et sammensat netværk baseret på jævnstrøm (DC) i stedet for vekselstrøm (AC) har fået opmærksomhed i både det videnskabelige og industrielle miljø. Der er mange lovlige egenskaber ved DC komponenter, som kunne være fordelagtige, såsom: Mindre dimensioner, lavere vægt, færre ledere, ingen forbehold for reaktive effekter og færre samlede tab på grund af manglende proximity- og skineffekter. På den anden side opstår der udfordringer grundet den lav eventræksimpedans, højt in- og output ripplestrøm fra DC/DC konvertere og store fejlstrømme i tilfælde af driftssvigt.

For at sikre en solid baggrund for en DC samlenetværkssimulering, er der blevet foretaget grundig research af elektriske modeller af komponenter. Passende modeller var identificeret og implementeret i det transiente simuleringsprogram PSCAD. Ved at anvende dette værktøj, blev de transiente effekter, fejlscenarier i vindmøllekonvertere, kabler og højspændt jævnstrøm (HVDC) links studeret. Ved simulering af ligevægtstilstanden viste systemet høj ripplestrøm grundet den lave netværksimpedans og den høje input og output ripple fra konverteren. Tre forskellige ripple-reduceringsløsninger er blevet foreslået. Generelt viser det foreslåede DC netværk god transient respons ved forstyrrelser og ligevægtsforholdene opnås igen efter udfald. Som forventet opnår strøm- og spændingsmaksima størrelser på over 15 gange den nominelle værdi ved disse udfald. Ydermere er DC netværkene blevet evalueret for tab og den samlede effektivitet af DC samlingsnetværket er fundet til at være 94,9 %.
I would like to thank my supervisor, Tonny Wederberg Rasmussen, for his support and guidance during this master project, and the special course I attended previously. The door to his office was always open, and whenever questions appeared, his competent and analytic approach always helped to find a solution.

Furthermore, I would like to express my gratitude to Walid Ziad El-Khatib for introducing me to the DC grid concept in the course "Transients in Power Systems". Additionally, he was always accessible if question during the thesis appeared. His positive attitude and smart feedback did not just help to solve problems, it enriched the daily working environment at DTU.

Finally, I would like to embrace Hilary Hansen and Tobias M. Nørbo for corrections, proofreading, and Danish translations.
Acronyms

AC  Alternating Current.

BIDITC  Bidirectional Thyristor Converter.

CL  Conduction Loss.

CSC  Current Sourced Converter.

DAB  Dual Active Bridge.

DC  Direct Current.

DPS  Dual Phase Shift.

FT  Fast Thyristor.

HVAC  High Voltage Alternating Current.

HVDC  High Voltage Direct Current.

HVS  High Voltage Side.

IEEE  Institute of Electrical and Electronics Engineers.

LVS  Low Voltage Side.

MMC  Modular Multi-Level Converter.
NPC Neutral Point Clamped.

PCC Point of Common Coupling.

PCT Phase Controlled Thyristor.

PWM Pulse Width Modulation.

RMS Root Mean Square.

SL Switching Loss.

SLG Single Line to Ground.

SPS Single Phase Shift.

TPDAB Three Phase Dual Active Bridge.

VSC Voltage Source Converter.

WTG Wind Turbine Generator.

ZCS Zero Current Switching.
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Chapter 1

Introduction

Offshore wind farms have a lot of potential to provide electricity for thousands of domestic households without having a negative impact on society. The wind resources at open sea appear to have higher wind speeds and are less turbulent compared to the circumstances on land. Massive projects are planned in the North- and Baltic seas in the next years, and especially Germany and the United Kingdom have consented to expand their capacity [1]. For 2050, the global offshore wind power installation is estimated to be around 700 GW [2]. The demand that is created in Northern Europe helps to decrease the global costs of turbines, foundation, and the installation process. Reduced costs and greater experience, increase the attractiveness of this technology to investors outside of Europe, leading to a further increase in installed capacity. Cost scenarios predict a further decrease of approximately 35% to prices in the next 35 years [2].

One main issue of offshore wind energy is the dispatch of the electrical energy from the sea to the land. Long distances require high transmission voltages in order to economically dispatch the energy without high losses. Two different methods, High Voltage Alternating Current (HVAC) and HVDC, are considered. Recent studies have proven that HVAC connections are the best economical choice for transmission distances below 80 km [3]. As the distance increases, the benefits of HVDC are dominate the economical properties [4]. Fur-
thermore, the active power transmission capability is not limited \cite{5}, and the implementability into a recently proposed multi-terminal \textit{HVDC} grid is given \cite{6} \cite{7}.

If a collection grid for offshore wind farms is reviewed, the state of the art transmission technology to the \textit{HVDC} platform is AC. This has been adapted from onshore wind farm installations; however, since the power is not fed in a regular 50 Hz grid, but further transmitted with HVDC, the continuing use of 50 Hz AC in the collection grid is questionable.

The possibility to use a DC-collection grid instead of AC-collection grid inside an offshore wind farm has been introduced in several scientific reports in recent years \cite{8} \cite{9} \cite{10} \cite{11}. Benefits, like material and weight savings in cables and transformers and a general higher efficiency, enable this technology to become superior in the future. Since an efficient transmission of current requires a voltage above the state of the art generator voltage of 0.69\textit{kVAC} (1\textit{kVDC}), the internal DC grid voltage needs to be boosted to a medium voltage. DC/DC converters enable this process, and different topologies are suggested such as the \textit{Dual Active Bridge (DAB)} \cite{12} and the \textit{Bidirectional Thyristor Converter (BIDITC)} \cite{13}. However, market information about the development progress of these innovative converters are hard to find (06/2014). Furthermore, some main challenges have been identified when DC collection grids are considered. First of all, since DC currents naturally do not cross the zero current mark in time, they are more difficult to break if a fault current appears. Special breakers need to be implemented that are able to break the high currents and market availability of those switches needs to be established. Secondly, the lack of field experience leads avoidance by many wind farm developers. Without experience, the first applicant of a new technology has to deal with start-up problems that might be expensive. Thirdly, the lack of standardisation for DC grids hampers the development of the components. Therefore, further research about the various system configuration needs to be performed and documented.
Chapter 2

Collection Grids for Offshore Applications

2.1 Introduction

In this chapter, the reader is going to be introduced to the concepts of AC and DC collection grids for offshore wind farms. At first, the state of the art AC technology is revised and the basic concepts are presented. Afterwards, the focus is put to the DC grid. An overview of the grid shows the wind farm with HVDC-connection. The specific control responsibilities of the converter entities are illustrated, and the voltage levels are defined in Figure 2.2. Subsequently, two different connection strategies for wind turbines with DC output are explained. To conclude, the chapter ends with an overview of merits of DC grids, and the challenges that need to be tackled.
2.2 AC Collection Grid

The state of the art technology to interconnect offshore wind turbines is via a medium voltage, AC grid. The technology has been adapted from onshore wind farms and is approved and well known to grid designers. The transformation of the power can be described in several steps. The typical generator output of a offshore wind turbine is $V_{G_{out}} = 0.69kV$. With an assumed wind turbine size of $S = 5MW$, the Root Mean Square (RMS) current output of the generator at full load conditions can be estimated with: $I_{G_{out}} = 4.16kA$; however, the current magnitude of the generator depends on the wind conditions and hence on the load to the rotor. The voltage level $V_{G_{out}}$ is too low to be used as transmission voltage to the collector offshore platform, since conduction losses would be too high due to the high currents. Therefore, the voltage is stepped-up to 10-33 kV medium voltage by utilizing a 50 Hz transformer. Between generator and transformer, an AC/DC/AC converter is operating to ensure a stable 50 Hz voltage for the transformer and provide controllability for the generator. Several converter typologies are used in practice, but for offshore applications, the trend to full scale converters seem to dominate.

![Figure 2.1: Offshore AC-Grid - 1 Feeder](image)

It can be seen from Figure 2.1 that each turbine is separated from the medium voltage collection grid by a 50 Hz transformer. Therefore, all components on the Wind Turbine Generator (WTG) side are galvanically isolated from the grid and need to withstand just a fraction of the medium grid voltage.

Offshore wind turbines are connected in parallel arrangements and form a feeder. Each feeder transmits the power output of several turbines to the main platform. This enables a more economical cable layout in the park. At the main platform, the voltage level is further increased to a high voltage transmission level. Depending on the transmission distance to the shore and the power rating of the wind farm, either HVAC or HVDC technologies is preferred.

In AC collection grids the voltage in the turbine is stepped-up utilizing a
transformer. In DC Collection Grids, on the other hand, the possibility of a transformer-less solution can be considered. This will imply that the converter needs to withstand the full DC voltage directly with the semiconductors.

2.3 DC Collection Grid for HVDC Applications

The development to built wind farms far away from the coast enables HVDC connections to become superior to the usual HVAC systems. Capacitive effects from long undersea cables do not affect the direct current transmission and therefore, the active power can be dispatched more effectively. Additional reactive VAR compensation is not essential at the PCC; however, the grid operator might demand ancillary services to support the operability of the local grid.

In the last recent years, the scientific community elaborated about the possibility to connect offshore wind farms with a DC instead of an AC collection grid. Challenges arise especially when new DC/DC converter systems are designed, built and tested. However, benefits like material, weight and size reduction of components, and lower losses enable DC grids to possibly dominate AC grids in the future.

2.3.1 Overview

The main structural difference in a DC collection grid compared to an AC collection grid is the DC/DC boost converter that provides a DC output at the wind turbine terminals. Several topologies were proposed in the scientific community to fulfill this purpose. In this thesis, two promising converter concepts are elaborated and compared regarding operability, control, and losses (See Chapter 3.3).

![Figure 2.2: DC grid Template - 1 Feeder](image)

**Figure 2.2: DC grid Template - 1 Feeder**
In Figure 2.2 an overview of an offshore wind farm has been illustrated that utilizes a DC grid. The 0.69 kV AC synchronous generator output is rectified with a passive AC/DC diode rectifier. Capacitive filters operate to stabilize the current and voltage. The output of the diode rectifier is boosted with a DC/DC converter system up to 30 kV. This medium voltage has been chosen to compare the losses with the existing 33 kV AC grids. The power is collected in the DC collection network and transmitted to the HVDC-platform, where the voltage is further stepped up. On land, the HVDC/AC inverter transforms the DC current to AC and injects it into the local AC grid. The grid codes for the wind farm apply at the Point of Common Coupling (PCC) and need to be controlled by the HVDC/AC inverter. The orange arrows in Figure 2.2 indicate the control responsibilities of the single converter blocks. The low voltage is controlled by the internal wind turbine converter, whereas the DC grid voltage is maintained from the HVDC converter.

2.3.2 Connection Strategies

Extensive work regarding offshore wind farm layouts has been published in [14]. Even though, the authors main focus was based on AC collection grids, many methods can be applied to DC collection grids as well. The most promising DC structures for HVDC applications are presented below.

Parallel Bus Design

In the radial design the wind turbines are parallel connected to the medium voltage bus and forming a string that is connected to the main platform. The maximum transmitted power through a cable is given by the location in the bus and the output power of the turbine. Therefore, cables have to be designed to carry the maximum load current at a specific position. As can be seen from Figure 2.3, the turbines are separable by switchgear components. Each turbine is required to be equipped with a set of breakers that can isolate the turbine during a fault. DC switchgear components are under development and deal with specific problems such as no zero crossing of the current in regular operation. Different methods to break DC currents are elaborated in [15], and intensive research is performed in this field. Therefore, it can be assumed that a suitable breaker technology will be available on the market when the first DC grid field test are performed.
To improve availability of the turbines, a redundant cable can be installed inside the park layout. Such a cable could provide a second current path if a failure in one of the feeders occurs. The possibility exists that several feeders share the same redundant cable, since the probability of two faulted feeders is very unlikely. However, the optimal design layout depends on disturbance studies as well on economic aspects and is not investigated further in this work. More information can be found in [13].

Series Bus Design

A second interesting approach to connect wind turbines is the serial bus design. In this case, the turbines are connected in series and hence the voltage output of each turbine adds up. This implies that a DC/DC step-up converter is not essential for this grid layout. Therefore, the possibility to save costs is given. In Figure 2.4 an illustration of the possible design is shown.
The drawback of this design is clearly the fault vulnerability. In case of a technical failure in a turbine, the voltage in the feeder will drop, because the turbine needs to be excluded from the grid. A lower grid voltage implies higher currents in the wind turbine rectifier, in the cables and in the HVDC boost converter terminal. This creates a natural limit of the amount of malfunctioning turbines, before the whole feeder needs to be disconnected to the HVDC-platform. Additionally, the amount of turbines in series determines the collection grid level. This means, with a state of the art generator voltage output of 0.69kV (1kV DC), several turbines need to be connected together in order to reach an adequate transmission voltage of 10-30kV DC. To reduce the amount of turbines per feeder, a DC/DC boost converter could be used between the rectifier and the grid to increase the collection grid voltage. However, this would introduce another boost converter and hence negate the economical advantage. Additionally, the AC/DC rectifier would need to have an active Voltage Source Converter (VSC) structure to control the generator.

In this work, the parallel wind turbine layout has been used for further elaboration, since the uncertainties of the serial grid design are too high.

2.3.3 Characteristic of DC Collection Grids

The purpose of this chapter is to highlight the factors that enable the DC collection grid to be come superior to existing AC collection grids. In addition, the challenges that need to be addressed are listed as well.
2.3 DC Collection Grid for HVDC Applications

Advantages

- No heavy 50 Hz transformer in the turbine necessary
- No contribution of reactive power in the grid
- 2 phases instead of 3 phases
- No skin and proximity effects - only resistive losses in the cables
- DC/DC converters can have modular structure
- DC/DC conversion with medium frequency decreases component size

Challenges

- Technology is untested on large scale MW application
- No DC-standards - regulation essential
- Challenges arise with medium frequency transformers
- Fast DC-breaker to limit short-circuit currents are essential

The list above is used as a guideline in this report.
Chapter 3

Electrical Models for Collection Grids

3.1 Introduction

The goal of this chapter is to find appropriate electrical representations, that can describe the behaviour of components that are employed in a medium voltage DC-grid. Nearly exceptionless, all the information are originated from the Institute of Electrical and Electronics Engineers (IEEE) database and references are given to the specific papers. The models have been implemented in the electrical simulations software PSCAD/EMTDC, that seems to be the leading power simulation software for transient studies. To operate converters with high switching frequencies and see the influence of lightnings into the grid, the solution time step has been set to $t_{ets} = 1 \mu s$. This will ensure good transient response during the entire simulation.
3.2 Wind Turbine - Electrical Representation and Rectifier

In this study, a four-pole synchronous generator has been used to provide the electrical conversion of torque to electric power. The drive train, rotor and wind turbulence is not a part of the simulation, since it would extend the simulation time for the whole system drastically. Furthermore, the dynamic behaviour of the mechanical systems consists of time constants with longer intervals compared to the fast responses of electrical transients. Therefore, it is valid to neglect this representation for short time intervals.

The generator model is presented in Figure 3.1. It consists of synchronous machine with rated power $S = 5 \text{ MVA}$ and an exciter model. Both models have been utilized from the PSCAD model library. The torque of the machine can be adjusted by a slider and is set to 1 pu during the simulation. Since the machine does not synchronize with a grid, the rotational speed of the machine can vary. This is a necessary feature for a variable speed wind turbine. The output of the machine is connected to a standard diode bridge rectifier. No active power components are necessary. The current of the machine is controlled by the DC/DC converter.

![Synchronous Machine Representation with Diode Rectifier](image)

Figure 3.1: Synchronous Machine Representation with Diode Rectifier

The key information of the diodes used in the rectifier have been summarized in Table 3.1. With this information, a loss estimation can be performed. Two different methods are proposed to estimate the loss. The first includes only
the (dominant) conduction loss, whereas the second considers also the reverse recovery effect.

<table>
<thead>
<tr>
<th>Rectifier Diode</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Manufacturer</td>
<td>ABB</td>
</tr>
<tr>
<td>Number</td>
<td>5SDD 5SDD 24F2800</td>
</tr>
<tr>
<td>Collector-emitter voltage [V]</td>
<td>2800</td>
</tr>
<tr>
<td>Max. Average on-state current [A]</td>
<td>2596</td>
</tr>
<tr>
<td>Max. RMS on-state current [A]</td>
<td>4078</td>
</tr>
<tr>
<td>Max. Peak on-state current [A]</td>
<td>8156</td>
</tr>
<tr>
<td>DIODE ON-Resistance [mΩ]</td>
<td>0.135</td>
</tr>
<tr>
<td>DIODE Forward Voltage Drop [V]</td>
<td>0.906</td>
</tr>
</tbody>
</table>

Table 3.1: Diode Properties for 6 Pulse-Rectifier Bridge

The conduction loss through a diode can be determined by equation 3.3. To determine $I_{rms}$ and $I_{avg}$, the apparent peak current $I_{peak} = 6kA$ has been determined by investigating the results of the simulation from the model in Figure 3.1.

The $I_{rms}$ and $I_{avg}$ currents can be estimated with:

$$I_{rms} = \frac{I_{peak}}{2} = \frac{6kA}{2} = 3kA$$ (3.1)

and

$$I_{avg} = \frac{I_{peak}}{\pi} = 1.91kA$$ (3.2)

if sinusoidal currents are with an conduction angle of 180° are assumed. Therefore,

$$P_{Lcond} = U_f * I_{avg} + R_d * I_{rms}^2$$ (3.3)

$$P_L = 0.906V * 1910A + 0.135\Omega * 3000A^2 = 2.946kW$$ (3.4)

The loss has to be multiplied with the amount of diodes (6) to receive the total loss of the diode bridge,

$$P_L = 6 * 2.946kW \approx 18kW$$ (3.5)
Another option to determine the loss is the forward power loss vs. average forward current characteristic in the data-sheet of the diode. This method can be used to determine both, conduction loss and reverse recovery energy. Notice, that Figure 3.2 is valid for a certain junction temperature (85°). It can be seen, that the loss estimation from the characteristic is slightly higher compared to the calculated value.

![Graph](image)

**Figure 3.2:** Forward Power-Loss vs. Average Forward Current Characteristic - Alternative Approach to Determine the Loss [Figure ABB]

### 3.3 Line Model - Cables

#### 3.3.1 MVDC Cable Model

The wind farm internal cabling consists of commercially available medium voltage cables. The distance between the turbines can be estimated as 450m and the distance between the wind turbine feeders as 850m. These numbers are based on the Nysted offshore wind farm. The wind farm cabling has a bipolar structure.

Since the transmission offshore is most likely realized with multi-layered cables rather than overhead-transmission lines, the capacitive and inductive parasitic effects need to be taken into consideration. In [16], an overview of three different
DC cable modelling methods for VSC-HVDC transmission systems are provided. It is proven that the simplest model, the $\pi$-model, can provide results that are as accurate as more complex frequency-dependant phase model for short distances up to 15 km. Therefore, the cable model inside the wind farm is realized with a lumped $\pi$-model. The parameters for a MVDC cable are shown in Table 3.2. It is assumed that the diameter of the internal cables is fixed through the entire park and does not depend on the position of the turbine. The values for resistance, inductance, and capacitance depend on the cable type and the diameter of the conductor. The diameter of the conductor, in turn, depends on the maximum conducted current. The values provided in Table 3.2 are obtained from the cable manufacturer Nexans [17].

<table>
<thead>
<tr>
<th>MVDC Cable</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Rated Phase to Ground Voltage [kV]</td>
<td>± 15</td>
</tr>
<tr>
<td>Cable Length [km]</td>
<td>0.45</td>
</tr>
<tr>
<td>Steady State Current [kA]</td>
<td>$4 \times 0.166 = 0.664$</td>
</tr>
<tr>
<td>Cross Sectional Area [mm²]</td>
<td>500</td>
</tr>
<tr>
<td>Line Resistance [Ω/km]</td>
<td>0.0366</td>
</tr>
<tr>
<td>Line Inductance [mH/km]</td>
<td>0.357</td>
</tr>
<tr>
<td>Parasitic Capacitance [μF/km]</td>
<td>0.3</td>
</tr>
</tbody>
</table>

Table 3.2: Cable-parameter for MVDC

Note, that shunt conductance can be neglected because of the rather small dielectric loss factor for modern insulation materials like XLPE [18].

![Bipolar Π-Circuit](image)

Figure 3.3: Model Bipolar Π-Circuit

Generally, it can be said that DC transmission inherits lower conduction losses, since skin and proximity effects do not appear, and no phase shift causes reactive power. Moreover, AC cables are insulated for the peak voltage, even tough just
the RMS voltage is effective, which does not apply to DC cables. Therefore, it can be stated that a single conductor 22 kV DC cable has the same intrinsic loss as a three conductor 33 kV cable, and is hence more efficient [10].

The loss of the cable is defined by the resistance of the cable and the DC current that is flowing through it. The current magnitude in the cable differs from the location of the cable in the wind farm, since parallel bus design is assumed.

\[ P_L = R_{\text{cable}} * I_{\text{DC}}^2 \]  

(3.6)

3.3.2 HVDC Cable Model

The economic benefits of HVDC are superior compared to HVAC from transmission distances of 80 km, when a 300 MW wind farm is connected via cables (±150 kV) to the grid [3]. Therefore, newly commissioned wind farms and power transmission over long distances utilize HVDC technology. One benefit of HVDC is that the capacitive effects of cables do not limit the active power transmission in steady state. However, as the transmission distance increases, the capacitive effects need to be taken into account to accurately simulate transient processes in the system.

Voltages and currents can be seen as travelling waves in cables. They have a finite travelling speed, and their progression along the line may cause reflections and refractions at the end of the line. Those occurrences can further create local peak points of current and voltage; however, the natural resistance of cables dampens out the peak values of the wave. The lumped π-model is not able to model this effect of a travelling wave along the line, since the capacity has been lumped to the front and end of the line. It might be sufficient to simulate short cable segments inside the wind farm, but if the transmission distance increases, another model needs to be introduced.

Fortunately, PSCAD provides several more elaborated models to accurately take long distance transient responses into account. For this purpose the "Frequency Depended Phase Model" has been implemented into the simulation. "Frequency Depended Models are distributed travelling wave models. However, the system resistance R is distributed across the system length (along with L and C) instead of lumped at the end points. More importantly, the FD models are solved at a number of frequency points, thereby including the frequency dependence of the system [19]."

The cable parameters illustrated in Table 3.3 have been chosen by using the cable ratio of 158 and recalculated to a 800mm² cross-section cable. The cable structure consists of six layers: main conductor, insulator 1, sheath, insulator 2, armour and insulator 3. The dimensions between the layers can be seen in Fig-
The distance between the positive and the negative pole in the seabed is assumed to be 20 m.

Table 3.3: Cable-parameter for HVDC

![Cable Model with Dimensions](image)

<table>
<thead>
<tr>
<th>Technology</th>
<th>Bipolar HVDC</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rated Phase to Ground Voltage [kV]</td>
<td>± 150</td>
</tr>
<tr>
<td>Cable Length [km]</td>
<td>200</td>
</tr>
<tr>
<td>Steady State Current 1 Feeder [kA]</td>
<td>0.066</td>
</tr>
<tr>
<td>Steady State Current 15 Feeder [kA]</td>
<td>0.999</td>
</tr>
<tr>
<td>Cross Sectional Area [mm²]</td>
<td>800</td>
</tr>
</tbody>
</table>

3.4 Protection Equipment

In Figure 2.3, an overview of the breakers pattern is shown. It can be seen that several circuit breakers need to be installed in order to protect the system in case of faults. The absence of a natural zero crossing of the current is hereby problematic. Therefore, the breakers need to be proactive to disrupt the current. A second problem occurs when DC grids are considered. AC grids have naturally high inductances, since massive transformers are a part of the system.
This limits occurring short circuit current automatically. DC grids, on the other hand, are based on converter technology to step-up the voltage. The inductance of the system depends on the topology of the converters; however, a lower inductance will be apparent. Short-circuit currents are not limited by a high grid impedance. The rise-time of the currents might be considerably shorter, implying higher $\frac{di}{dt}$. Peak currents up to 20 times the steady state current can appear. Wind farm installations are exposed to this mechanical and thermal stress if no appropriate breaker is installed. Therefore, it can be concluded that a DC-breaker needs to be able to interrupt the current, and interrupt the current in a very short time [15].

DC breakers have been a research topic especially for naval applications in recent years, since the benefits for a MVDC system on military ships with electric propulsion are huge [20]. Research is still ongoing, and information about the availability and development status are hard to find. The following technologies are promising to solve the problem in the future:

- Hybrid breaker
- Hybrid breaker with forced commutation
- Mechanical circuit breaker with snubber circuit
- Solid-state-circuit breaker

A thorough study of these breaker types and their merits and drawbacks have been performed in [15] and shall not be a part of this study. A brief summary between the different methods are illustrated in Figure 3.5. It is highlighted in the paper that the solid-state breaker combines both the best reaction time and price relation, for grid voltages above 20 kV. Below 20 kV, the mechanical circuit breaker with snubber circuit provides the best price relation. Hybrid breakers seem to underperform in economical, as well as technical performance aspects. On the contrary, ABB has published several papers about an novel developed hybrid breaker for MVDC in 2012 and seem to pursue this technology. For illustration purpose, Figure 3.6 show the concept of an solid state and an hybrid breaker.
3.5 Bidirectional DC/DC Converters

In this section, two different bidirectional converter concepts are investigated. The purpose is to determine the most efficient topology and to verify the operational behaviour. Please notice, that it is necessary to clarify certain positions in the converter when analysing the operation. For this reason, the phrase [Low Voltage Side (LVS)] and [High Voltage Side (HVS)] define the input and output full-bridge of the converter, respectively. Moreover, an illustration to distinguish between the phrases "switch", "valve" and "leg" in the converter has been put in Appendix A.1 and should be reviewed there, before the reader continues with this section.
3.5.1 Converter Selection and Requirements

The purpose of a DC/DC converter is to step-up the voltage from a lower to a higher level. This is necessary because currents that are too high would create unreasonable losses in the transmission lines and the converter itself, leading to an uneconomical operation of the wind farm. Therefore, intensive literature research has been performed to find converter topologies that can be operated in following conditions:

- The voltage is stepped up from 1 kV DC to 30 kV DC.
  Those values are chosen, since the state of the art generator for wind turbines supply $V_{LL} = 0.69kV$. By utilizing a three-phase diode rectifier, this voltage can be transformed to $V_{DC} = 1.35V_{LL} = 1kV$ [21]. The DC grid voltage of 30 kV DC is a trade-off between the losses in the transmission lines and the voltage withstanding capability of the components. A higher DC grid voltage would decrease the resistive losses in the transmission lines, but increase the amount of switches that need to be stacked in the converter. This increases the inherent conduction loss in the valves. 30 kV is chosen, since existing AC collection grids employ voltages between 24 kV and 36 kV.

- Bi-directional power flow in the converter.
  The turbine needs to be supplied with electrical energy when the wind conditions are below the cut-in wind speed and the turbine is in idle state. The DC/DC converter needs to be able to reverse the current direction and supply the axillary services. It should be noted that the current rating of the components in reversed direction needs to be just a fraction of the primary direction.

- Input current control capability.
  The DC/DC converter is able to control the output current of the AC/DC diode rectifier from the wind turbine generator.

- The losses of the converter and the estimated costs for components are reasonable.
  The converter needs to be able to boost the voltage with low losses and perform this task as price-efficient as possible

Two promising candidates have been identified, modelled, and described in the following two chapters. Furthermore, the converters have been investigated regarding operability, losses, and number of components. The first topology employs a medium-frequency transformer and phase shift control between the low
and the high voltage side. The second topology benefits from a transformer-less architecture, and hence, one expensive key component is saved. This converter uses a frequency control to regulate the power. The most suitable topology is used for the simulation in Chapter 4.

### 3.5.2 Dual Active Bridge Converter with a Medium Frequency Transformer

#### 3.5.2.1 Working Principle

The Dual-Active Bridge (DAB) converter, first proposed in [12], is a high power density, low loss, and easy to control DC/DC converter that has gained some attention in the last two decades. The converter is able to be operated bidirectionally, in step-down or step-up voltage mode. In Figure 3.7, the main topology is illustrated. The converter consists of two active full bridges that are interconnected via a small and light medium-frequency transformer. Furthermore, an inductor is essential in the AC circuit of the topology that may be inherent in the transformer reactances. The active power flow is controlled by varying the phase shift displacement between the two bridges over this inductance. Figure 3.8 shows a simplified model for the converter. The two bridges operate with two-level square wave switching, where the two valves in one converter leg are shifted by 180°. This provides the converter with a significantly speed increase compared to pulse-width modulated control strategies, since no sine wave has to be modelled.

The current through the inductance can be determined the following equation

$$ I \angle \Theta = \frac{V_{f1} \angle \varphi_1 - V_{f0} \angle \varphi_{ref}}{j\omega L} $$

(3.7)

if the voltages $V_1$ and $V_2$ appear to be purely sinusoidal. Unfortunately, this relation is not correct for this converter type, since square voltage waves are used; however, as proposed in [12], the square waves can be replaced by their fundamental frequencies, since the phase shift $\Theta$ determines the active power flow (See Figure 3.8). Therefore, equation 3.7 can be used. The RMS fundamental component $V_{f1}$ is given by:

$$ V_{f1} = \frac{2\sqrt{2}V_1}{\pi} $$

(3.8)
where $V_1$ represents the square-wave input voltage of peak amplitude. Similarly, for the output fundamental component $V_{f2}$ with the square-wave output voltage $V_2$.

$$V_{f0} = \frac{2\sqrt{2}V_2}{\pi}$$

(3.9)

A second, more accurate, approach to calculate the active output power is achieved by determining the Fourier representation of the voltages $V_1$ and $V_2$. By individually adding those contributions, the transmitted active power of the converter can be determined [22].

$$P = \sum_{n=1}^{n=\infty} \left( \frac{4V_1}{n\pi} \frac{4V_2}{n\pi} \sin(n\varphi) \right) \left( n\omega L \right)$$

(3.10)

In addition to the phase shift between the low voltage and high voltage bridge $\Theta = D_1$ ($D_1$ in [RAD]), an inner phase shift $D_2$ between the converter valves has been implemented. This means that IGBT $I_1$ and IGBT $I_4$ do not share the same gate signal, but are phase shifted by $D_2$ (See Figure 3.7). Advantages include the possibility to eliminate reactive power in the AC-circuit, causing lower peak currents in the converter, improved system stability, and limited inrush currents. The improved control structure has been introduced in [23] and further elaborated in [24].
In Figure 3.7, the modular converter structure is shown. There are three reasons to divide the converter structure into several stages. Firstly, the power level of a 5 MVA wind turbine forces very high currents through the low voltage inverter stage, since the input DC voltage of the DAB converter is set to 1 kV. After the first design of the DAB was tested with just one module, the concept has been withdrawn due to the high conduction losses in the switches. The second simulated approach used a modular design with a parallel input connection on the LVS and a series connection of the modules on the HVS. With this design, the currents on the LVS and the voltage on the HVS are divided by $N$ number of modules. The second benefit from a modular converter structure is the scalability and replaceability. Other power ratings can easily be achieved for other wind turbine sizes, and in case of a failure, a module of a converter is easier to replace than the whole converter. The third advantage relates to the medium frequency transformer design. Since the HVS is connected in series, the individual modules act like series connected voltage sources and therefore, one module does not need to span the full voltage ratio of $1kV/30kV$, but are divided by $N$

$$\frac{N_2}{N_1} = \frac{30kV}{1kVN}$$

This enables an improved transformer design as shown in [25].

One crucial part of the DAB converter is the the medium-frequency, high-power transformer. It enables the converter to step-up/down the voltage and provides galvanic isolation between the low and high voltage side. Moreover, isolation between the high and low voltage sides are ensured. Compared to normal 50/60 Hz transformers, the size of a medium frequency transformer in a DAB converter is significantly decreased due to the dependency of the magnetic core cross
sectional area $A_c$ to the frequency

$$A_c = \frac{V_{\text{rms}1}}{\sqrt{2\pi} f B N_1}$$  \hspace{1cm} (3.12)\

where $V_{\text{rms}1}$ is the root mean square value of the primary voltage, $B_p$ is the peak flux density, $N_1$ is the number of primary turns, and $f$ is the frequency. The magnetic core dimensions can be decreased with increased operational frequency, and still the rated power can be applied.

Increasing the operation frequency causes skin, proximity, hysteresis, and dielectric losses to be significantly increased. Therefore, a careful investigation of the temperature distribution in the transformer has to be performed, and new transformer materials need to be considered. More information can be found in Section 3.5.2.4.

### 3.5.2.2 Main Component size selection

The inductance in the intermediate AC circuit can be determined with [25]:

$$L_1 = \frac{V_1 V_2 \varphi (\pi - \varphi)}{2 P_{\text{out}} \pi^2 f_s N}$$  \hspace{1cm} (3.13)\

$V_1$ and $V_2$ are the in- and output voltage, respectively, $\varphi = D_1$ is the phase shift between the bridges, $P_{\text{out}}$ is the output power, $f_s$ is the switching frequency, and $N$ the transformer ratio. As mentioned before, the inductance $L_1$ might be designed internally as a part of the transformer reactances.

The size of the in- and output capacitors depends on the required voltage ripple and have been adjusted as needed during the simulation.

The IGBT selection depends on the voltage and current that is apparent in the individual leg of the converter. The dimensioning is documented in Section 3.5.2.4.
3.5 Bidirectional DC/DC Converters

3.5.2.3 Control

The control of the DAB converter can be executed easily with Single Phase Shift (SPS) between the two bridges. However, this control strategy has some disadvantages regarding reactive power and harmonics. Therefore, the Dual Phase Shift (DPS) control system has been implemented [23].

In Figure 3.10, the control schematic is illustrated. As can be seen, the base of the control is a reference saw tooth signal (X) that is operating at a constant frequency ($f_s = 5kHz$) with a peak value of $saw_{peak} = 2\pi$. This saw tooth signal (X) is fed into the SinPhi block to generate the reference sine wave for all gate signals. The phase displacement, $\varphi$, of the sine wave depends on the output of two separated PI-controllers. Generally:

$$\text{SinPhi} = \sin(X + \varphi) \quad (3.14)$$

![Control and Start-up Behaviour- DAB Converter](image)

**Figure 3.9:** Start-up Behaviour DAB Converter

In this control pattern, two separate PI-controllers are used to generate the phase shift between the bridges, $D_1$, and the phase shift between the gates, $D_2$. The input for both PI’s is the input current $I_1$ compared to a reference current $I_{1ref}$. The output of $PI_1 = D_1$ shifts the gate signals from the LVS in relation to the HVS. The output of $PI_2 = D_2$ generates the displacement between the valves $I_5/I_7$ in relation to $I_5/I_6$, as well as $I_3/I_4$ in relation to $I_1/I_2$. The converter is set to be operated in $0 \leq D_2 \leq D_1 \leq 1$ conditions. This means that the phase shift between the gates is always bigger than the phase shift between the valves (See Figure 3.9). As detailed reported in [24], the transmitted power
can be calculated with:

\[ P_t = \frac{nV_1 V_2}{2f_s L} \left[ D_1 (1 - D_1) - \frac{1}{2} D_2^2 \right] \]  

(3.15)

Additionally, the possibility to define deadtimes between the gate signals is given to avoid short-circuits in the converter legs.

**Figure 3.10:** DAB Control Schematic

### 3.5.2.4 Loss Estimation and Frequency Selection

Theoretically, the DAB converter topology can be used in many frequency ranges. The selection of the switching frequency depends on many aspects and especially the resistive losses are closely related to it. On the one hand, the designer of the converter system is pushed to increase the frequency of the system as much as possible to decrease the volume of the transformer. This will decrease the weight and material cost of the device. On the other hand,
3.5 Bidirectional DC/DC Converters

a smaller transformer implies a higher energy-density factor. This makes the cooling process of the transformer more complex, and additionally, the switching loss of the power electronic switches is increased. Moreover, it will be proven, in the upcoming sections that the total loss will increase with higher frequencies.

In the following chapters, an approach to determine the highest possible switching frequency is made. At first, the medium frequency transformer is investigated regarding core and winding losses. Afterwards, the IGBT switches are examined, and switching and conduction loss is determined. Based on the results, a suitable switching frequency is suggested in Section 3.3.2.5

Transformer losses, switching losses, and conduction losses are included in the investigation.

The losses of the DAB converter can be subdivided into the following paragraphs.

Transformer Losses Currently, there is a lot of research regarding high-power, medium-frequency transformers. Many publications are published in recent years. The key challenges for a proper design are to combine suitable voltage isolation requirements with low thermal stress. The latter depends mainly on the losses inside the transformer, which can be subdivided into core and winding losses. The impact of the frequency on the transformer design is huge, and parameters have to be chosen carefully. A detailed study regarding frequency behaviour of transformers has been performed by [26], [27], [25], and [28]. A detailed analysis of this topic is out of the scope for this thesis, but information from the sources above have been gathered to justify a reasonable switching frequency.

Core Losses In 50/60 Hz transmission systems, the transformer core materials are often laminated silicon-steel or nickel-steel. These materials are relatively cheap and readily available. The disadvantage is high core loss.

New materials have been discovered that combine a high saturation flux with less specific losses. The properties of nanocrystalline and amorphous cores are superior in comparison to the old silicon-steel materials, as seen in Table 3.4. However, the price of the new materials is considerably higher due to their elaborate manufacturing process. [26]

As stated in [27], the nanocrystalline core material are superior in comparison
Table 3.4: Core Materials in High-Power Medium Frequency Transformers. 

<table>
<thead>
<tr>
<th>Series</th>
<th>Sat. Flux</th>
<th>Sp. losses</th>
<th>Manufacturer</th>
</tr>
</thead>
<tbody>
<tr>
<td>Microlite (2605SA1)</td>
<td>1.56 T</td>
<td>1.5 kW/kg</td>
<td>Metglas</td>
</tr>
<tr>
<td>Powerlite (2605SA1)</td>
<td>1.56 T</td>
<td>0.6 kW/kg</td>
<td>Metglas</td>
</tr>
<tr>
<td>Namglass</td>
<td>1.59 T</td>
<td>0.34 kW/kg</td>
<td>Magnet</td>
</tr>
<tr>
<td>Vitrovac (6030F)</td>
<td>0.82 T</td>
<td>0.19 kW/kg</td>
<td>VAC</td>
</tr>
<tr>
<td>Finemet (FT-3M)</td>
<td>1.23 T</td>
<td>0.14 kW/kg</td>
<td>Hitachi</td>
</tr>
<tr>
<td>Vitroperm (500F)</td>
<td>1.2 T</td>
<td>0.07 kW/kg</td>
<td>VAC</td>
</tr>
<tr>
<td>Nanoperm</td>
<td>1.2 T</td>
<td>0.04 kW/kg</td>
<td>Magnetec</td>
</tr>
<tr>
<td>Namglass 4</td>
<td>1.23 T</td>
<td>0.04 kW/kg</td>
<td>Magnet</td>
</tr>
<tr>
<td>Arnon 7 (3-6%Si,Fe)</td>
<td>1.53 T</td>
<td>1.6 kW/kg</td>
<td>Arnold</td>
</tr>
<tr>
<td>Arnon 5 (3-6%Si,Fe)</td>
<td>1.48 T</td>
<td>1.06 kW/kg</td>
<td>Arnold</td>
</tr>
</tbody>
</table>

Generally, core losses can be described as hysteresis loss. The hysteresis of magnetic field intensity, \(H\), and flux density, \(B\), are used to describe the effects of the time varying flux, \(\Phi\), inside the curve. With increased frequency, the magnetic dipoles in the core need to rearrange faster, and more energy is needed. This implies a higher loss-density factor. In the hysteresis loop, the enclosed surface represents the dissipated energy during each cycle. In [27], the effects of frequency have been studied extensively, and Figure 3.11 has been provided, to illustrate the loss behaviour.

It can be seen, that the core losses from 1 kHz to 10 kHz do not change significantly. However, above 10 kHz, the body of the hysteresis becomes bigger, and hence, more loss is apparent. In general, hysteresis loss is denoted with

\[
P_h = K_h \ast f \ast B_m^n
\]

where \(K_h\) is a proportionality constant for the material and shape of the core, \(f\) is the frequency, \(B_m\) the maximum flux density, and the exponent \(n\) ranges from 1.5 to 2.5 which is material specific [29].

The second loss apparent in the core is eddy current loss that is generated if a
conductive material is exposed to a time-varying magnetic field. Those current loops flow perpendicular to the flux and heat-up the core [25][30]. Moreover, eddy currents decrease the skin-depth of the available core surface. At 50/60 Hz, the effects of eddy currents are negligible and are prohibited by laminations in the core. However, with increased frequency, the losses from eddy currents will have a bigger contribution, since they vary with the square of the frequency.

\[ P_e = K_e \times (B_m \times f \times \delta)^2 \]  

(3.17)

\( K_e \) is again a proportionality constant for the material, \( B_m \) the maximum flux density, \( f \) the frequency and \( \delta \) is the lamination thickness [29].

The equations presented above should not be the first choice if a detailed investigation is necessary, as they are based on empirical studies and have restricted validity for non-sinusoidal current waveforms. Moreover, the frequency and flux density spectrum might be limited. A method to estimate exact core losses is proposed in [27], but the complexity exceeds the applicability to this master thesis.

In summary, it can be said that the core losses in the frequency range up to 10 kHz are not considerable since new, low-loss, materials are used. Attention should be paid so that the core is designed to operate at 10-20% below the
saturation flux density. It is important to minimize hysteresis losses to keep the core losses at a reasonable level.

**Winding Losses**  The winding loss from the primary and secondary side is defined by the product of the square current RMS value and the inherent DC resistance of the winding. For DC and low frequency RMS currents, the resistance can be calculated with:

\[
R_{DC} = \frac{N \times \rho \times l_{mean}}{A_w}
\]  (3.18)

where \(N\) is the number of turns, \(\rho\) is the conductor resistivity at a certain temperature, \(l_{mean}\) is the average length per turn, and \(A_w\) is the cross section of the wire.

With increased frequency, the current distribution in a wire changes. Skin and proximity effects influence the current behaviour and change the resistance of the wire. Due to the rapidly changing magnetic fields in the wires, the actual conducting surface is decreased and leads to an increased resistance. To avoid these parasitic effects, often stranded conductor wires, also known as Litz-wires, are used. Therefore, the winding losses can be controlled by a careful design of the conductor shape and the frequency does not have a big impact in the range up to 10 kHz.

[25] presented a comprehensive analysis of high-power, medium-voltage, transformer in his licentiate report.

**Conclusion Transformer Losses**  In the previous chapters, a summary has been given to provide information about the challenges that arise when implementing a medium frequency transformer. The main losses of such a transformer have been explained. In order to provide reasonable losses for the transformer, the methodology of [25] has been used. In his project, a detailed determination has been proposed to evaluate the losses. The author applied his work to a DAB converter and evaluated the losses. Therefore, non-sinusoidal current and voltage waveforms are applied, resulting in a situation applicable to this project.

The summary of the work in [25] can be stated as:
3.5 Bidirectional DC/DC Converters

- System requirements: \( P = 5 \text{MW}, V1/V2 = 3/30kV, f = 5kHz \) turbine with isolate [DAB] converter
- Proposed modular converter design with five 1 MW converter modules
- For one converter module, the losses of the transformer have been calculated by setting fixed parameters, calculating basic requirements, and optimizing free parameters, leading to an optimized design

By utilizing this information, a good loss estimation can be provided and included in the model for PSCAD. The highest efficiency for a 1 MW transformer module was found to be \( \eta_{\text{opt}} = 99.74 \). This implies a total loss of \( P_{\text{Ltot}} = 2637.5 \text{W} \), subdivided into winding loss \( P_{Lw} = 1639 \text{W} \), core loss \( P_{Lc} = 979 \text{W} \) and dielectric loss \( P_{Ld} = 19.5 \text{W} \). Since the latter has a minor share in the energy loss, it has been neglected in further studies. With the knowledge about the share in between the losses, a suitable PSCAD model in \([\text{pu}]\) could be derived.

<table>
<thead>
<tr>
<th>Rating One Module</th>
<th>Rating Five Module</th>
<th>( \sum P_{-L} ) in ([\text{pu}])</th>
</tr>
</thead>
<tbody>
<tr>
<td>( P_{Ltot} )</td>
<td>2637.5</td>
<td>13188</td>
</tr>
<tr>
<td>( P_{Lw} )</td>
<td>1639</td>
<td>8195</td>
</tr>
<tr>
<td>( P_{Lc} )</td>
<td>979</td>
<td>4895</td>
</tr>
</tbody>
</table>

**Table 3.5:** Determined Transformer Losses

The information gathered in Table 3.5 has been used to simulate the transformer. \( P_{Lw} \) determines the copper losses, whereas \( P_{Lc} \) accounts for the no-load losses in the program. Saturation effects have been neglected since the transformer should be designed to avoid those effects.

It should be emphasized that the total volume of a 1 MW, single phase, 5 kHz, 3 kV / 6 kV power transformer can be as large as 45 litres [25]. This represents a huge decrease of material and weight compared to a 1 MW, three phase, 50 Hz, 0.4 kV / 10 kV transformer which has a volume of \( \approx 3250 \text{litres} \) [31].

**Switch related losses** The [DAB] topology utilizes active switches to transform different voltage levels through an intermediate AC link. Each of those switches have inherit individual conduction and switching loss.

**Conduction Loss (CL)** Electrical switches, such as thyristors or IGBT’s, can not be modelled as perfect switches. They consist of several doped semicon-
ductor layers that allow control of current flow through the device by applying a voltage to the gate terminal. During the conduction state, a small resistance, $R_d$, can be noticed that leads to conduction loss in form of heat. Furthermore, the forward voltage drop $U_f$ in a semiconductor influences the conduction loss from a device. The relation between conduction loss can be expressed with

$$P_{L\text{cond}} = U_f \cdot I_{\text{avg}} + R_d \cdot I_{\text{rms}}^2$$

(3.19)

where $I_{\text{avg}}$ and $I_{\text{rms}}$ denote the average and rms current, respectively. Both, the forward voltage drop, $U_f$, and the on resistance, $R_d$, can be determined by the data-sheet of the semiconductor manufacturer. It should be mentioned that conduction losses depend on the operating temperature of the device. For this analysis, the parameters have been determined for a temperature of 125°C. Below, the on-state characteristics for the two different IGBT’s are illustrated.

As can be seen from Figure 3.12, the on resistance can be determined with following relation.

$$\frac{1}{R_d} = \frac{\Delta y}{\Delta x} = \frac{\Delta I_C}{\Delta V_{CE}}$$

(3.20)

The on-resistance is the inverse of the gradient in the linear regression of the on-state characteristic. The gradient is used from the point of operation of $I_C$. Furthermore, the forward voltage drop $U_F$ can be derived from the data-sheet.
by determination of the crossing of the linear regression graph with the axis of abscissa. Both values have been determined from Figure 3.12 and presented in Table 3.6. Note that the on-state characteristics are different for the IGBT and the anti-parallel diode, but both are provided in the data-sheet. The values for the diodes have been determined in the same manner.

It should be noticed, that current and voltage requirements are different for each half bridge since a transformer separates both sides. Therefore, the properties of IGBT 1 are optimized for high current and a lower voltage, whereas IGBT 2 has a higher DC voltage blocking capability and needs to conduct less current. Table 3.6 summarizes the properties for the chosen ABB IGBT switches with determined conduction loss properties.

<table>
<thead>
<tr>
<th></th>
<th>IGBT 1</th>
<th>IGBT 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>Type</td>
<td>HiPak IGBT Module</td>
<td>HiPak IGBT Module</td>
</tr>
<tr>
<td>Manufacturer</td>
<td>ABB</td>
<td>ABB</td>
</tr>
<tr>
<td>Number</td>
<td>5SNA 3600E170300</td>
<td>5SNA 0400J650100</td>
</tr>
<tr>
<td>Purpose</td>
<td>Low Voltage Side</td>
<td>High Voltage Side</td>
</tr>
<tr>
<td>Collector-emitter voltage [V]</td>
<td>1700</td>
<td>6500</td>
</tr>
<tr>
<td>Rated DC collector current [A]</td>
<td>3600</td>
<td>400</td>
</tr>
<tr>
<td>Rated peak collector current [A]</td>
<td>7200</td>
<td>800</td>
</tr>
<tr>
<td>IGBT ON-Resistance [mΩ]</td>
<td>0.57</td>
<td>7.71</td>
</tr>
<tr>
<td>IGBT Forward Voltage Drop [V]</td>
<td>1.05</td>
<td>2.55</td>
</tr>
<tr>
<td>DIODE ON-Resistance [mΩ]</td>
<td>0.313</td>
<td>5.0</td>
</tr>
<tr>
<td>DIODE Forward Voltage Drop [V]</td>
<td>0.85</td>
<td>1.65</td>
</tr>
</tbody>
</table>

**Table 3.6:** Summary of IGBT used in DAB. Top: Basic Information, Bottom: Information regarding conduction loss derived from Figure 3.12

The current and voltage ratings of the semiconductor switches have to be noticed carefully. In order to deliver accurate predictions about the loss behaviour, the components need to be stacked/placed in parallel regarding the converter requirements. Figure 3.13 illustrates the voltage and current distributions of the DAB converter in full load conditions for one IGBT switch.
(a) Pulses, Voltage and Current in the Low Voltage Side of the DAB Converter

(b) Pulses, Voltage and Current in the High Voltage Side of the DAB Converter

Figure 3.13: Steady State, Full Load Conditions
It can be seen, that currents up to 1.2 kA appear on the low voltage side (Figure 3.13a). Furthermore, the main current flow is conducted through the IGBT. The diode is used for commutation, and enables Zero Current Switching (ZCS) for turn-on. Figure 3.13b indicates the higher voltage and the lower current for the high voltage side. The main current is conducting through the anti-parallel diode, that is used as a rectifier.

To improve the loss estimation quality, the converter system has been dimensioned with a safety margin. It is common practice to double voltage and current in order to withstand voltage transients without damaging components.

Table 3.7 states voltage and current magnitudes that are apparent under full load conditions. From this values, the ratings for the rectifier and inverter stage are determined. The amount of stacked/parallel switches are stated afterwards. Finally, the total on-resistance and the forward voltage drop of the converter is calculated. These values have been implemented in PSCAD simulation.

<table>
<thead>
<tr>
<th></th>
<th>Full-Bridge LVS</th>
<th>Full-Bridge HVS</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak Voltage [kV]</td>
<td>1</td>
<td>6</td>
</tr>
<tr>
<td>Peak Current [A]</td>
<td>1200</td>
<td>200</td>
</tr>
<tr>
<td>Parallel Connection</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>Series Connection</td>
<td>1</td>
<td>2</td>
</tr>
<tr>
<td>Max. Voltage Rating DC [kV]</td>
<td>1.7</td>
<td>13 (2x 6.5kV)</td>
</tr>
<tr>
<td>Max. DC Current Rating [A]</td>
<td>3600</td>
<td>400</td>
</tr>
<tr>
<td>$\sum$ IGBT ON-Resistance per Valve [mΩ]</td>
<td>0.57</td>
<td>15.42</td>
</tr>
<tr>
<td>$\sum$ IGBT Forward Voltage Drop [V]</td>
<td>1.05</td>
<td>5.1</td>
</tr>
<tr>
<td>$\sum$ DIODE ON-Resistance per Valve [mΩ]</td>
<td>0.313</td>
<td>10</td>
</tr>
<tr>
<td>$\sum$ DIODE Forward Voltage Drop [V]</td>
<td>0.85</td>
<td>3.3</td>
</tr>
</tbody>
</table>

Table 3.7: Conduction ON-resistance for Simulation of DAB. 1st: Apparent Voltage and Current, 2nd: Parallel/Series IGBT’s, 3rd: Total Withstand Capability for one Module 4th: Accumulated On-resistance

Notice that the rated values from Table 3.7 have been implemented with a safety margin of 1.8. As can be seen, two values in series are applied in the high voltage circuit. This leads to a higher total resistance per valve. However, since the current is lower than in the low voltage side, the energy dissipated during the conduction cycle will be lower as well. Note that the forward voltage drop of the parallel connected IGBT’s is the same on the low voltage side. In practical
applications, it is important to ensure equal current sharing between the IGBT modules.

In general, it can be observed in the PSCAD simulations that the highest conduction loss per IGBT appears in the low voltage side due to the high currents.

Since the full-load steady-state current distribution is known from Figure 3.13, the currents $I_{avg}$ and $I_{rms}$ are determined with.

$$I_{avg} = \frac{1}{T} \int_{t-T}^{t} i(t)dt$$  \hspace{2cm} (3.21)

and

$$I_{rms} = \sqrt{\frac{1}{T} \int_{t-T}^{t} i(t)^2dt}$$  \hspace{2cm} (3.22)

for both high and low voltage side.

<table>
<thead>
<tr>
<th></th>
<th>Low Voltage Side</th>
<th>High Voltage Side</th>
</tr>
</thead>
<tbody>
<tr>
<td>$I_{avg}[kA]$</td>
<td>0.54</td>
<td>0.09</td>
</tr>
<tr>
<td>$I_{rms}[kA]$</td>
<td>0.77</td>
<td>0.14</td>
</tr>
</tbody>
</table>

**Table 3.8:** Average and RMS Current for Steady State, Full Load Conditions in DAB

By utilizing equation 3.19 and the information from Table 3.7 and 3.8 the conduction loss in one valve, in one module, and for the whole converter can be approximated as seen in Table 3.9.

<table>
<thead>
<tr>
<th></th>
<th>Low Voltage Side</th>
<th>High Voltage Side</th>
<th>Total</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P_{CL_{switch}}[kW]$</td>
<td>0.904</td>
<td>0.246</td>
<td></td>
</tr>
<tr>
<td>$P_{CL_{valve}}[kW]$</td>
<td>0.904</td>
<td>2 * 0.246 = 0.492</td>
<td></td>
</tr>
<tr>
<td>$P_{CL_{side}}[kW]$</td>
<td>4 * 0.904 = 3.616</td>
<td>4 * 0.493 = 1.972</td>
<td></td>
</tr>
</tbody>
</table>

$\sum P_{CL_{module}}$ 5.588

$\sum P_{CL}[kW]$ 5 * 3.616 = 18.08 5 * 1.972 = 9.860 27.94

**Table 3.9:** Expected Conduction Loss in DAB Converter

**Switching Loss (SL)** The switching loss per cycle determines the amount of energy to break a current with a certain voltage. The amount of energy needed
depends on the level of voltage applied between the collector and emitter as well as the switching frequency. In the data-sheets of the switches, information about turn-on and turn-off energy are given which are valid under certain conditions. Table 3.10 gathers information about the switching loss energy of the two mentioned switches.

<table>
<thead>
<tr>
<th></th>
<th>IGBT 1</th>
<th>IGBT 2</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Type</strong></td>
<td>HiPak IGBT</td>
<td>HiPak IGBT</td>
</tr>
<tr>
<td><strong>Manufacturer</strong></td>
<td>ABB</td>
<td>ABB</td>
</tr>
<tr>
<td><strong>Number</strong></td>
<td>5SNA 3600E170300</td>
<td>5SNA 0400J650100</td>
</tr>
<tr>
<td><strong>Purpose</strong></td>
<td>Low Voltage Side</td>
<td>High Voltage Side</td>
</tr>
<tr>
<td><strong>IGBT Turn-on energy</strong></td>
<td>$E_{on}$ [J]</td>
<td>$E_{on}$ [J]</td>
</tr>
<tr>
<td></td>
<td>1.10</td>
<td>2.8</td>
</tr>
<tr>
<td><strong>IGBT Turn-off energy</strong></td>
<td>$E_{off}$ [J]</td>
<td>$E_{off}$ [J]</td>
</tr>
<tr>
<td></td>
<td>1.60</td>
<td>2.120</td>
</tr>
<tr>
<td><strong>Diode Reverse recovery energy</strong></td>
<td>[J]</td>
<td>[J]</td>
</tr>
<tr>
<td></td>
<td>1.26</td>
<td>1.38</td>
</tr>
<tr>
<td><strong>Voltage $V_{base}$ [V]</strong></td>
<td>900</td>
<td>3600</td>
</tr>
<tr>
<td><strong>Collector Current $I_{base}$ [C]</strong></td>
<td>3600</td>
<td>400</td>
</tr>
<tr>
<td><strong>Temperature [$^\circ$ C]</strong></td>
<td>125</td>
<td>125</td>
</tr>
</tbody>
</table>

**Table 3.10:** Summary of IGBT used in DAB. Top: Basic Information, Middle: Information regarding switching loss, Bottom: Switching Conditions

Generally, both turn-on and turn-off losses have to be accounted for carefully. The switch energies provided in Table 3.10 are only valid if the switch is operating with the mentioned conditions. Therefore, an adjustment factor needs to be included.

$$E_{on} = E_{on}^* \frac{V_m}{V_{base}} \frac{I_m^2}{I_{base}^2} \quad (3.23)$$

and respectively

$$E_{off} = E_{off}^* \frac{V_m}{V_{base}} \frac{I_m^2}{I_{base}^2} \quad (3.24)$$

where $V_m$ and $I_m$ are the voltages and currents that are apparent during switching conditions and $V_{base}$ and $I_{base}$ are the rated conditions for the switch (See Table 3.10).

In order to find $V_m$ and $I_m$, PSCAD has been used to investigate the switching in steady-state full-load conditions. At first, the low-voltage high-current side is investigated.
Switching Loss in the Low-Voltage High Current Side  Plots from the switching sequence (IGBT 1 - Low Voltage Side) are presented in Figure 3.13a. It can be observed that the current during turn-on in the IGBT is zero, because the anti-parallel diode from IGBT 2 conducts the current in the transition process. This enables ZCS conditions, implying no losses during turn-on state in the low-voltage side.

However, during turn-off, a current of \( I_m \approx 1.2kA \) has to be disrupted. This can be seen in Figure 3.13a. The red line shows the sampled current in the IGBT during turn-off. By using equation 3.23 with the voltage applied and the sampled current, the energy loss per switch can be calculated. For the situation described in Figure 3.13a:

\[
E_{off} = 1.6J \times \frac{1kV}{0.9kV} \times \frac{1200A^2}{3600A^2} = 0.2J
\] (3.25)

The current that is flowing in the diode is low compared to the IGBT current. Therefore, the assumption is valid that no loss will be produced in the diode.

The switching power loss can now be calculated by multiplication with the frequency.

Switching Loss in the High-Voltage Low Current Side  The steady state current and voltage distribution for the high voltage side is illustrated in Figure 3.13b. The voltage is boosted to 6 kV, and the current magnitude decreased by \( \frac{1}{6} \).

The most charge is conducted in the diode and not in the IGBT which is used for commutation purposes during the switching process. However, the IGBT has to break a current of \( I_m = 0.2kA \) during turn-off, which will imply a switching loss. Therefore, the return recovery energy in the diode and the IGBT turn-off is calculated using equation 3.23 with high voltage, low current, conditions. Notice, that two IGBT’s are stacked in a valve and the voltage per switch is 6kV/2 = 3kV. According to [32], the dissipated energy in the diode during turn-on is negligible, however, during turn-off the reverse recovery energy needs to be accounted for.

For the IGBT turn-off switching loss:

\[
E_{off} = 2.8J \times \frac{3kV}{3.6kV} \times \frac{200A^2}{400A^2} = 0.5833J
\] (3.26)
and for the diode return recovery current:

\[ E_{off} = 1.38 J \times \frac{3kV}{3.6kV} \times \frac{200A^2}{400A^2} = 0.288J \]  \hspace{1cm} (3.27)

**Conclusion Switching Loss**  In the previous chapters, the total energy dissipation for one turn-off switch cycle has been determined. As mentioned before, no turn-on switching loss is generated since ZCS conditions apply. Therefore, the total switching loss can be determined by multiplication with the frequency \( f \).

\[ P_{sw} = E_{off} \times f \]  \hspace{1cm} (3.28)

The relation between switching loss and frequency is linear. Figure 3.14 illustrates the dissipated energy of one IGBT switch on the LVS and HVS respectively. Doubled switching frequency implies doubled losses. Notice that this investigation has been performed for full-power steady-state. During partial load conditions, the current distribution can be different. As can be seen, the highest losses arise on the HVS of the converter.

![Switching Loss vs. Switching Frequency](image)

**Figure 3.14:** Switching Loss as a Function of Frequency - Separated per IGBT

A detailed listing of the energy dissipated per switching cycle and the total energy loss is presented in Table 3.11. Since two IGBTs are stacked in the HVS.
per valve, the switching loss is doubled.

<table>
<thead>
<tr>
<th></th>
<th>Low Voltage Side</th>
<th>High Voltage Side</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>IGBT</td>
<td>Diode</td>
</tr>
<tr>
<td>$P_{SL_{\text{switch}}} \frac{f}{J_{\text{sw}}}$</td>
<td>0.2</td>
<td>-</td>
</tr>
<tr>
<td>$P_{SL_{\text{valve}}} \frac{f}{J_{\text{sw}}}$</td>
<td>0.2</td>
<td>2 * 0.5833 = 1.1666</td>
</tr>
<tr>
<td>$P_{SL_{\text{side}}} \frac{f}{J_{\text{sw}}}$</td>
<td>4 * 0.2 = 0.8</td>
<td>4 * 1.1666 = 4.4664</td>
</tr>
</tbody>
</table>

$\sum P_{SL_{\text{module}}} \frac{f}{J_{\text{sw}}}$ = 7.5704
$\sum P_{SL} \frac{f}{J_{\text{sw}}}$ = 5 * 7.5704 = 37.852

Table 3.11: Overview - Expected Switching Loss in DAB Converter

3.5.2.5 Frequency Selection Conclusion

The last chapters provided information about factors that influence the losses of a [DAB]. It can be summarized that due to novel core materials, the transformer losses are not the main driver for the frequency selection. With an intelligent design, the transformer losses can be kept at a very low level. However, the losses in the IGBT's need to be considered carefully. Since the voltage and current levels are given parameters for the simulation, the conduction loss will be constant with varying frequency. However, switching loss increases linearly with the frequency and is the main loss in the converter. According to [33] the maximum dissipated power by an IGBT device can be estimated at around 6kW. This value is valid for steady-state operating temperature and is limited by the surface of the IGBT and the efficiency of the heat sink. This provides a limitation to the switching frequency of the [DAB] converter.
Figure 3.15: Accumulated Loss for one IGBT as a Function of Frequency - CL=Conduction Loss, SL=Switching Loss

Figure 3.15 provides an overview of the maximum possible switching frequency. In the configuration studied in this project, the maximum switching frequency can be estimated with 6.5kHz due to the high switching losses in the high voltage side. Frequencies above this limit force the IGBT to dissipate more than 6 kW of heat and are therefore not realizable.

The total dissipated energy has been calculated by adding switching and conduction loss together.

\[
P_{L_{\text{total}}} = P_{CL} + P_{SL} \tag{3.29}
\]

Therefore, after a careful investigation, the switching frequency of the DAB converter has been set to 5kHz in the simulation. This provides some safety margin for the component.

3.5.2.6 Total Loss DAB Converter

In this section, the calculated results from the previous section are listed together. The assumption has been made that the passive components such as the input and output filters account for 5 kW of loss.

The total losses calculated account for \( \approx 5\% \) of the total power compared to the output of 5 MVA. From Table 3.12, it is obvious that the switching loss in the IGBT accounts for the highest share of loss in the converter. More specifically, the stacked IGBT's in the HVS are responsible for this high loss.
### Table 3.12: Total Estimated Loss for a 5 MVA DAB Converter

<table>
<thead>
<tr>
<th>Type of Loss</th>
<th>Loss in [kW]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Transformer Winding</td>
<td>13.19</td>
</tr>
<tr>
<td>Transformer Core</td>
<td>8.20</td>
</tr>
<tr>
<td>Transformer Insulation</td>
<td>4.900</td>
</tr>
<tr>
<td>IGBT Conduction</td>
<td>27.94</td>
</tr>
<tr>
<td>IGBT Switching</td>
<td>189.26</td>
</tr>
<tr>
<td>Passive Components</td>
<td>5.00</td>
</tr>
<tr>
<td><strong>Total</strong></td>
<td><strong>248.49</strong></td>
</tr>
</tbody>
</table>

This value could be decreased significantly if the modular converter consisted of 6 instead of 5 modules. The voltage share between the modules would decrease from $30kV/5 = 6kV$ to $30kV/6 = 5kV$. Therefore, it could be assumed, that one $6.5kV/400A$ IGBT could handle the $5kV$ voltage level with a $1.5kV$ safety margin. Therefore, instead of two IGBT’s, one IGBT would operate and the switching loss is halved.

A second possibility to lower the loss is simply to reduce the switching frequency.

### 3.5.3 Bi-Directional Thyristor Converter

In this chapter, the BIDITC, firstly proposed in [34] and further elaborated in [13], has been investigated. The voltage levels have been adapted to the limitations of the system. The choice to pick this converter topology has been made, since a transformer-less structure is possible. Therefore, one of the main cost drivers can be saved. Moreover, the topology is completely soft-switched and no switching loss appears.

#### 3.5.3.1 Working Principle

The converter, illustrated in Figure 3.16, consists of a low and a high voltage circuit that is connected through a capacitor $C_r$. The circuit has a bipolar structure with a common ground connection in the middle. This enables bipolar energy transmission and is suitable for HVDC with Current Sourced Converter (CSC) or VSC technology.

In order to step-up the voltage, a switched-capacitor circuit is used. By alternately firing thyristors T1 and T2 on the LVS, the capacitors are charged through the resonant inductance $L_1$, leading to a boosted voltage across the ca-
pacitor. The charge in the capacitors, and hence the voltage, can be controlled by firing thyristors T3 and T4 on the high voltage side. The size of the resonant inductor $L_2$ affects the derivatives of the output current and defines the peak capacitor voltage. In step-up mode, thyristors T5-T8 are turned-off and have no influence in the operation; however, in step-down mode, thyristors T1-T4 are turned-off and thyristors 5-8 start to operate. It needs to be noticed that thyristors T5-T8 only need to be rated for the idle current of the turbine.

$$f_{\text{max}} = \frac{1}{2t_q} \quad (3.30)$$

**Figure 3.16: Topology Bi-directional Converter**

Soft switching of the current is caused by the resonance operation between the inductor $L_1$, the resonance capacitor $C_r$ and the inductor $L_2$. It decreases the switching losses and reduces the harmonics in the system.

Table 2 indicates the objective of different thyristor modules.

<table>
<thead>
<tr>
<th>T1/T2</th>
<th>T3/T4</th>
<th>T5/T6</th>
<th>T7/T8</th>
</tr>
</thead>
<tbody>
<tr>
<td>Purpose</td>
<td>Main Power-Flow</td>
<td>Aux. Supply</td>
<td></td>
</tr>
</tbody>
</table>

**Table 3.13: Purpose of Thyristors in Converter**

### 3.5.3.2 Main component size selection

The size of the components can be determined by applying three basic equations. The first step is to determine the nominal switching frequency:

$$f_{\text{max}} = \frac{1}{2t_q} \quad (3.30)$$
where $t_q$ is the circuit-commutated turn-off time of the thyristor. This information is found in the data-sheet of the component. Afterwards, the resonance capacitor and inductor can be calculated with the equations provided in [13].

\[
\frac{I_2(V_2 - V_1)}{V_1 V_2} = 2C_r f_s
\]  

(3.31)

and

\[
L_{1cr} = \frac{1}{(\pi^2 f_s^2 C_r)} \quad L_1 \leq L_{cr1}
\]  

(3.32)

The critical size of the inductor $L_{1cr}$ defines the maximum charge that is transmitted to the capacitor. In order to ensure discontinuous conduction mode, the value of $L_1$ needs to be chosen below the resonance inductor $L_{1cr}$. Equations [3.30] [3.31] and [3.32] imply the following characteristics for the high power converter:

- The resonant operation between $L_1$ and $C_r$ generates high voltages by precisely switching the thyristors
- With increased switching frequency $f_s$, the component size of $C_r$ and $L_1$ can be decreased.
- Losses consist mainly of conduction loss in the thyristors, since switching takes place at zero current. Moreover, snubber circuits are unnecessary.

In offshore applications, a bi-directional power flow is necessary to ensure the supply of energy in the turbine in idle mode. In reverse current mode, the rated current values for the semiconductors could be decreased to a fraction of the main current direction values, because the auxiliary systems do not demand much electrical energy. However, each converter leg needs to withstand the AC blocking voltage $V_{ac}$. This applies to the low voltage and the high voltage side. Therefore, a practical converter would consist of several stacked thyristor modules.

### 3.5.3.3 Control System

To ensure a stable operation of the converter, the controller proposed in [13] has been implemented. An illustration is presented in Figure [3.17]. The con-
trol pattern uses a frequency regulation principle that is synchronized with the alternating voltage, $E_a$, over the switched capacitor. By utilizing a phase-locked-loop, the input/output voltage or current wave forms can be aligned even during transient states. This enables the converter to react fast during faults or imbalances in the grid.

**Figure 3.17:** Step-Up Control System for Thyristor Converter

In the proposed converter topology, the input current is controlled with a PI-controller. However, the output current or voltage could be controlled if necessary. The proportional and integration constants for the controller have been determined by simulation and adjusting the parameters iteratively. The output of the PI-controller is added up with the error signal from the phase-locked loop and is used as the input for the integrator. This component ramps up the signal according to the input and resets the input after $360^\circ$ has been reached. Depending on the magnitude of the input, the frequency of the ramp, and hence the switching frequency of the thyristors is controlled. The shape of the ramp will be continuous and have the same gradient in steady-state.

In voltage step-up mode, the output of the integrator is used to fire the thyristors at predefined reference angles $\alpha_{lv}$ and $\alpha_{hv}$. As can be observed, the gate signals T1 and T2 are fired at a fixed angle of $\alpha_{LVT1} = 0^\circ$ and $\alpha_{LVT2} = 180^\circ$, respectively. Gate Signals T3 and T4, however, are fired at the angle $\alpha_{HVT3} = 139^\circ$ and $\alpha_{HVT4} = 139 + 180^\circ$, respectively. The angle $\alpha_{HV}$ is defined by the relation:
\( V_{ac} > V_2 \) \hspace{1cm} (3.33)

to fire the high voltage thyristors with a positive bias and ensure the current flow from the low voltage to the high voltage side. In other words, the intermediate AC voltage of the converter needs to be larger than the output voltage of the converter to activate the conduction state of the thyristors. Notice, that thyristors 5-7 are permanently turned-off in this operation state.

Figure 3.18: Intermediate AC Circuit Waveforms with Output Voltage \( V_2 \)

The control system for the voltage step-down mode is not actually implemented in the control pattern in PSCAD, but is implied in the control pattern in Figure [3.17]. When the converter operates in step-down mode, Thyristors T1-T4 become turned-off and the opposed thyristors T5-T8 are active. The predefined fire angle, \( \alpha_{lv} \), is kept by the thyristors on the low voltage side, but the fire angle on the high voltage side, \( \alpha_{hv,down} \), is adapted in order to reverse the power flow. The full derivation of the control system is documented in [13].

3.5.3.4 Thyristor Development and Maximum Switching Frequency Limitation

The following chapter contains background information about thyristor development, especially in relation to the circuit commutated turn-off time \( t_q \). This value determines the maximum switching frequency that can be achieved with thyristor valves and is therefore crucial to consider. The evolution of thyristor valves began in the 1970’s to find a replacement for the former gas-filled high-power switches. Thorough the years, the voltage and current rating increased [35].
This has been gained especially due to the improvements of silicon industry. The first thyristors had a diameter of 33 mm and were able to withstand a voltage of 1.6 kV with a direct current capability of 1 kA. Nowadays, 152 mm inch Phase Controlled Thyristors (PCTs) are available in the market that allow to withstand voltages up to 8 kV and a direct current of 4.5 kA. However, these switches are usually used for HVDC based on CSC. The switching frequency of these switches is limited by the circuit commutated turn-off time $t_q$. This value defines the time-frame that is needed to remove the load carriers in the switched gate-layer of the thyristor and therefore, defines the time when the thyristor regains the blocking capability. Since the current ratings for thyristors are increasing, the diameter increases as well. Therefore, tendentially, it takes longer for larger switches to regain blocking capability and this leads to a higher turn-off time $t_q$. The commutated turn-off time of PCT with high current capability are in in the range of 450 – 1050 $\mu$s, allowing maximum switching frequencies of 0.5 – 1.1 kHz.

When thyristors are applied to medium voltage converters, special care needs to be taken of the current and voltage rating. Since the voltage level is low compared to HVDC converters, high current peaks can arise.

Recently, Fast Thyristors (FTs) became commercially available. These switches are specially designed for a faster circuit commutated turn-off times up to 25 – 100 $\mu$s, allowing switching frequencies of 5 – 20 kHz. Higher switching frequencies directly effect the size of the inductive and capacitive components main components of BIDITC leading to a cheaper design of the converter. The disadvantage of FTs is lower voltage and current ratings compared to PCTs.
The limitations can be estimated with a blocking voltage of $2 - 3kV$ and a conduction RMS current of $1 - 2kA$. Furthermore, parasitic effects can occur that need to be taken into consideration. Therefore, a careful analysis of the peak voltages has to be performed if FTS are considered.

### 3.5.3.5 Converter Practical Design

In order to estimate the losses for an applied converter, the amount of thyristors stacked and/or in parallel has to be determined. Therefore, data-sheets from leading power component manufacturers have been utilized. Moreover, as mentioned in the previous chapter, the circuit commutated turn-off time $t_q$ needs to be considered to ensure that the thyristor is able to operate with switching frequency $f_s$.

In the BIDITC, the DC voltage level on the low level side is 1 kV. Therefore, the highest current will occur in this position and the expected DC value of the current for a 5 MVA wind turbine can be estimated with:

$$I_{rms_{LV}} = \frac{S}{V_1} = \frac{5\text{MVA}}{1\text{kV}} = 5\text{kA}$$  \hspace{1cm} (3.34)

The current capability for the high voltage side is defined as:

$$I_{rms_{LV}} = \frac{S}{V_2} = \frac{5\text{MVA}}{30\text{kV}} = 0.166\text{kA}$$  \hspace{1cm} (3.35)

The voltage rating for each one of the converter legs is determined by the alternating voltage $V_{ac}$ in the intermediate AC circuit. Both, the LVS and the HVS need to withstand the full AC voltage $V_{ac}$. According to ABB [36], the practical voltage rating of a thyristor can be estimated with:

$$V_{DSM} = \sqrt{2} \times V_{il} \times k$$  \hspace{1cm} (3.36)

with an AC line to line RMS voltage of 30 kV and a safety factor $k = 2$. The safety factor is important to protect the valves from overvoltages, like the non-repetitive peak off-state voltage $V_{DSM}$ illustrated in Figure 3.20:

$$V_{DSM} = \sqrt{2} \times 30\text{kV} \times 2 = 90\text{kV}$$  \hspace{1cm} (3.37)
The voltages and currents determined above provide the foundation for the amount of valves per converter leg. As can be seen, the required blocking voltage and current capability are both high. Therefore, PCTs need to be used. The application of FT would lead to an uneconomical amount of thyristors that need to be stacked and/or in parallel connection to fulfill the ratings in each leg. However, the possibility to use higher frequencies and decrease the component size would be beneficial for economic considerations. Therefore, the development of the next generation of FTs with higher voltage and current ratings should be observed carefully in the future.

Three different thyristors have been selected that represent commercially available components. It is necessary to differentiate between the required current conduction capability in the individual legs. The most important key specification of the valves, as well as the mounting position in the converter, is indicated in Table 3.14. Notice, that the position of the thyristors are related to Figure 3.16.

The current ratings $I_{RMS}$ and $I_{avg}$ can be used to calculate the maximum allowed repetitive peak current in the thyristor. This has been performed for thyristor 1 (See Table 3.14).

Figure 3.21 illustrates a simplified schematic of each single thyristor in the converter.

Therefore, the relation between peak and average current can be derived by integration with:

$$I_{avg} = \frac{1}{2\pi} \int_0^{\pi} I_{peak} \sin(\omega t) dt = \frac{I_{peak}}{2\pi} (-\cos \pi + \cos \alpha) = \frac{I_{peak}}{2\pi} (1 + \cos \alpha) \quad (3.38)$$
### Table 3.14: Thyristor Properties Used for the BIDITC

<table>
<thead>
<tr>
<th></th>
<th>Thyristor 1</th>
<th>Thyristor 2</th>
<th>Thyristor 3</th>
</tr>
</thead>
<tbody>
<tr>
<td>Manufacturer</td>
<td>Infinion</td>
<td>Powerex</td>
<td>Powerex</td>
</tr>
<tr>
<td>Type</td>
<td>PCT</td>
<td>PCT</td>
<td>PCT</td>
</tr>
<tr>
<td>Peak forward/reverse blocking voltage [V]</td>
<td>7000...8000</td>
<td>4500 V</td>
<td>6500 V</td>
</tr>
<tr>
<td>Average on-state current [A]</td>
<td>2800</td>
<td>1200</td>
<td>325</td>
</tr>
<tr>
<td>Max RMS on-state current [A]</td>
<td>4400</td>
<td>1700</td>
<td>511</td>
</tr>
<tr>
<td>Circuit commutated turn-off-time [μs]</td>
<td>550</td>
<td>500</td>
<td>500</td>
</tr>
<tr>
<td>Peak non-repetitive surge current [A]</td>
<td>93000</td>
<td>36500</td>
<td>4243</td>
</tr>
<tr>
<td>Forward Voltage Drop [V]</td>
<td>1.35</td>
<td>1.262</td>
<td>1.169</td>
</tr>
<tr>
<td>Slope resistance [mΩ]</td>
<td>0.274 - 0.336</td>
<td>0.368</td>
<td>0.326</td>
</tr>
<tr>
<td>Mounting Position</td>
<td>T1/T2</td>
<td>T3/T4</td>
<td>T5/T6/T7/T8</td>
</tr>
</tbody>
</table>

The relation between RMS and peak current can be derived with [38]:

$$I_{RMS} = \sqrt{\frac{1}{2\pi}} \int_{0}^{\alpha} (I_{peak} \cdot \sin(\omega t))^{2} dt = \frac{I_{peak}}{2} \sqrt{\frac{1}{\pi} \left( \frac{\pi - \alpha + \sin(2\alpha)}{2} \right)} \quad (3.39)$$

Notice, that sinusoidal currents and voltages in the AC-circuit are assumed. The maximum average and RMS current are given at $\alpha = 0$. Therefore, the equations [3.38] and [3.39] simplify to

$$I_{RMS} = \frac{I_{peak}}{2} \quad (3.40)$$

and

$$I_{avg} = \frac{I_{peak}}{\pi} \quad (3.41)$$

The maximum allowable peak currents for the Thyristor 1 (See [3.14])

$$I_{peak} = 2I_{RMS} = \pi I_{avg} \quad (3.42)$$

Therefore:

$$I_{T1,peak} = 8.8kA \quad I_{T2,peak} = 3.2kA \quad I_{T3,peak} = 1.022kA \quad (3.43)$$

The converter is designed to only operate in discontinuous mode. This implies that no hard switching occurs and ZCS will apply. Therefore, no higher switching transients are expected since high $\frac{di}{dt}$ are avoided.
3.5 Bidirectional DC/DC Converters

![Diagram of Half Wave Thyristor Application](image)

**Figure 3.21:** Half Wave Thyristor Application

![Figure 3.22: Simulated Current through Thyristors](image)

**Figure 3.22:** Simulated Current through Thyristors

With the information gathered from Table 3.14, 3.15 and the simulated current and voltage waveforms seen in Figure 3.18, 3.22, the amount of series and parallel connection for the converter can be calculated.

Notice, that the current flow in the LVS is higher than in the HVS. See Table 3.15.

### 3.5.3.6 Losses

Similar to the DAB converter, the losses for BIDITC are calculated with the threshold voltage $U_f$ and the slope resistance $R_d$ as well as the average and
### Table 3.15: Peak Withstand and Dimensioned Values for Voltage for BIDITC

<table>
<thead>
<tr>
<th></th>
<th>Low Voltage DC Side</th>
<th>High Voltage DC Side</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak Withstand Voltage [kV]</td>
<td>45</td>
<td>45</td>
</tr>
<tr>
<td>Peak Current [A]</td>
<td>8000</td>
<td>600</td>
</tr>
<tr>
<td>Rated Peak Voltage [kV]</td>
<td>90</td>
<td>90</td>
</tr>
<tr>
<td>Rated Peak Current [A]</td>
<td>8800</td>
<td>3400</td>
</tr>
</tbody>
</table>

### Table 3.16: Amount of Stacked and Parallel Thyristors in BIDITC

<table>
<thead>
<tr>
<th>Valve</th>
<th>T1/T2</th>
<th>T5/T6</th>
<th>T3/T4</th>
<th>T7/T8</th>
</tr>
</thead>
<tbody>
<tr>
<td>Thyristor Model</td>
<td>1</td>
<td>3</td>
<td>2</td>
<td>3</td>
</tr>
<tr>
<td>Series</td>
<td>13</td>
<td>20</td>
<td>14</td>
<td>20</td>
</tr>
<tr>
<td>Parallel</td>
<td>1</td>
<td>1</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>Voltage Capability [kV]</td>
<td>91</td>
<td>90</td>
<td>91</td>
<td>91</td>
</tr>
<tr>
<td>Current Capability RMS [A]</td>
<td>8800</td>
<td>325</td>
<td>2820</td>
<td>325</td>
</tr>
</tbody>
</table>

\[ \sum \text{Resistance} = 4.368 \text{m} \Omega \quad 6.52 \text{m} \Omega \quad 5.152 \text{m} \Omega \quad 6.52 \text{m} \Omega \]

\[ \sum \text{Voltage Drop [V]} = 17.55 \quad 23.38 \quad 17.668 \quad 23.38 \]

\[ P_{CL} = (17.67V \times 2.8kA) + (4kA^2 \times 4.368m\Omega) = 119.4kW \] (3.45)

This estimates the conduction loss for one LVS valve. To calculate the full conduction loss for the LVS, the number has to be multiplied by four.

\[ \sum_{iv} P_{CL} = 4 \times 119.4kW = 477.6kW \] (3.46)
However, the losses would be drastically reduced if the voltage output of the generator is stepped up, and indeed, the trend to higher generator voltages is apparent. For example, the Multibird M5000 utilizes a 3 kV output voltage instead of the regular 0.69 kV. This would quarter the current, leading to significantly lower conduction losses:

\[ P_{CL} = (17.67V \times 0.7\,kA) + 1\,kA^2 \times 4.368\,m\Omega = 16.737\,kW \]  

(3.47)

This would provide a very attractive converter concept; however, since one goal of this work is the comparability to existing offshore wind farms, the grid output generator voltage is set to 0.69 kV. Therefore, the BIDITC is not suitable in this grid setup and is withdrawn from the overall simulation. It needs to be highlighted that the operability of this converter type should be studied further with different conditions.

### 3.5.3.7 Simplified Converter Structure with Buck-Converter for Auxiliary Services

A huge simplification of the converter topology would be apparent, if the auxiliary services for the turbine could be provided from another source. This could be realized with an efficient second DC/DC step-down converter in the turbine that is connected to the 30 kV grid.

![Figure 3.23: Structure of DC Grid with Auxiliary Buck Converter in Turbine](image)

The structure of the main converter would simplify a lot as presented in Figure 3.24.

Instead of eight legs of stacked thyristors, a topology of two thyristor and two diode legs are applicable. This would reduce the cost of the converter significantly.


Figure 3.24: One-Directional Thyristor Converter with Diode Bridge Rectification

3.6 HVDC-Converter Models

In order to study the transient response of the DC grid, the offshore wind farm model needs to be connected to a 50 Hz grid. As specified in Chapter 2, the most economical dispatch technology is HVDC for long distances and high power outputs. Therefore, a suitable HVDC boost converter and inverter model has been implemented in the PSCAD model. For these models, a detailed loss estimation has not been performed, since they are not a part of the DC grid. However, a good transient response to the medium voltage DC grid is established, as can be revised in Chapter 4.

3.6.1 HVDC Boost Converter

The purpose of the HVDC-boost converter is to raise the internal 30 kV grid voltage to a higher transmission potential. The schematic of the proposed boost converter can be seen in Figure 3.25. The topology is rather simple and well known from low power boost converters; however, the boost converter is scalable and needed power components are readily available on the market. Notice that IGBT and diode needs to withstand the full ± 150 kV, resulting in a stacked components. The component dimensioning of the inductor has been performed according to [39].

The derivation of conduction resistance and the forward voltage drop is similar to Section 3.5. Both values have been implemented into the PSCAD model.
The model proposed in Figure 3.25 is suitable to step-up the voltage only. Hence, only unidirectional power flow is possible. A bi-directional topology could be designed as well, however, the efficiency and costs of such a converter might not be as competitive since, all the components need to be dimensioned for the rated voltage or current, depending on the component location. On the other hand, a separate buck converter can be dimensioned for just the power level the wind farm needs in idle mode. Therefore, the assumption is made that a separate buck converter takes over operation if the wind conditions demand that the wind farm is an electrical consumer and hence, the power direction is reversed. The step down converter has not been modelled in this work because the idle mode of the farm is not a case of this project.

Additionally to step-up the voltage, the boost converter controls the internal DC voltage as illustrated in Figure 2.2. A simple voltage regulator structure with a PI-controller has been implemented. The control circuit of the structure can be seen in Figure 3.26.
### 3.6.2 HVDC-Inverter

The HVDC-inverter is the key component that connects the wind farm to the grid on land. The model consists of a two-level IGBT full-bridge inverter (VSC-Converter) that is connected via an inductor to a three-phase grid (See Figure 3.27). By controlling the currents in the inductors, the active and reactive power flow between the inverter and the grid can be controlled. Bidirectional power flow is possible. The voltage on the converter side is inverted with Pulse Width Modulation (PWM) techniques that allow a close emulation of the converter voltage in relation to the grid voltage.
As can be seen in Figure 3.27, a simple two-level converter structure is chosen. Nowadays, HVDC-VSC normally use Modular Multi-Level Converter (MMC) structures that have the ability to share the voltage between several IGBT’s per valve. The focus on this master thesis, however, is directed to the DC grid and therefore, just the simple two-level model is applied.

To control the HVDC-inverter, a vector-control structure published in [40] is used. In this paper, the line voltages and currents are transformed into dq0-reference frame, where steady state values appear to be constant. By exploiting this feature, regular PI-controller can be used to independently control active and reactive power as well as the DC-link voltage with the equivalent currents $i_d$ and $i_q$. The controllers have been tuned manually after the control was implemented. It has to be mentioned that the proposed control system in [40] includes some reference errors, that relate to the dq0 terms. However, those mistakes have been eliminated and the control system is fully working.

The voltage balance across the inductors can be stated with:

\[
\begin{bmatrix}
v_a \\ v_b \\ v_c
\end{bmatrix} = R \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} + L \frac{d}{dt} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} + \begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix}
\]

(3.48)

The dq0-reference frame representation can be performed regarding to Appendix A.3. Therefore, the voltage at angular speed $\omega_e$ are:

\[
\begin{align*}
v_d &= R i_d + L \frac{di_d}{dt} - \omega_e L i_q + v_d^1 \\
v_q &= R i_q + L \frac{di_q}{dt} + \omega_e L i_d + v_q^1
\end{align*}
\]

(3.49)

The active and reactive power can then be calculated.

\[
\begin{align*}
P &= 3(v_d i_d + v_q i_q) \\
Q &= 3(v_d i_q + v_q i_d)
\end{align*}
\]

(3.50)

Due to the complexity, the PSCAD model illustrations for the control system have been shifted to Appendix A.2. The losses of the HVDC inverter have not been part of this study. A two level inverter design has been chosen, because it provides a realistic transient response to the DC grid that is not too complex in structure. However, a detailed loss estimation would exceed the boundaries of this master thesis. Furthermore, it is likely that newly commissioned projects utilize (MMC) or Neutral Point Clamped (NPC) technology with improved harmonic behaviour.
4.1 Introduction

In this section, the overview of the simulated DC-collection grid is introduced. All components from Chapter 3 are assembled together to form the network. An illustration of the system is given in Figure 4.1. The grid consists of four wind turbines that are connected with one DC feeder. Every turbine is protected by DC breakers that are located at the terminal connection point (See Figure 4.1). Additionally, each cable segment between the turbines can be isolated. With this arrangement, in case of a cable fault, it is possible to exclude the cable segment while maintaining the operation of the turbines that are not affected. Unfortunately, it was only possible to simulate one feeder due to the complexity of the models and the resulting simulation time.

The initial goal of the thesis was to compare the behaviour of two different DC/DC converter topologies. Since the losses of the BIDITC were uneconomically high for the input voltage $V_{dc} = 1kV$ (see section 3.5.3.6), the focus of this study has been put to the DAB DC/DC converter.
Four different fault incidents are described in this thesis. Fault scenario 1 and 2 are directed to short circuits inside a wind turbine in different locations in the grid. Fault scenario 3 simulates a cable fault between two wind turbines and the effect of a Single Line to Ground (SLG) fault is observed. The last scenario deals with the influence of a lightning strike from the AC grid on land. Two different lightning strike intensities are illustrated. The simulations explained above are assumed to represent the most common fault types in a wind farm \[41\].

All faults occur at simulation time \(t_1 = 0.6\). Afterwards, the time to clear the fault is comprised of three contributions:

(a) the time \(t_{detc}\), the current needs to reach the critical short circuit value of this position \(i_{scr} = 5i_{steady\text{state}}\)

(b) a delay for the breaker to react \(t_{react} = 60 \mu s\)
(c) the time to break the current $t_{break} = 800 \mu s$

It can be estimated that the breaker action will occur around $1...1.5ms$ after the fault. Breaker reaction times have been found in [15].

Before the fault simulations, the system steady state conditions of a wind farm with a DC grid are presented.
4.2 Steady State

To approach a meaningful fault analysis, the first step is to investigate the steady state conditions. The purpose is to show currents and voltages in the system and to look for irregularities in the patterns. Figure 4.2 shows the system in steady state conditions. As can be seen, all voltages are on a stable and intended level.

Figure 4.2: Steady State Currents and Voltages
\[ V_{\text{gen}} = 1\, \text{kV} \]
\[ V_{\text{bus}} = 30\, \text{kV} \]
\[ V_{\text{HVDC}} = 300\, \text{kV} \] (4.1)

However, the currents from the generator and in the DC bus seem unexpected oscillating and are full of harmonics. This is unproblematic for the simulation itself, since the controllability of the system is still given. Furthermore, reactive power consideration can be neglected in a DC bus. In real life, on the contrary, this may cause troubles with the durability of the DC-cables and capacitors, and the overshoot in current will cause additional conduction loss in cables and components. Additionally, unwanted current resonances between the turbines might arise that overload the HVDC-boost converter terminal.

To identify the reason for such oscillating currents, the individual components have been investigated separately with resistive loads. Currents appeared to be non-distorted and without harmonics. However, as soon as the models were connected together, the currents fluctuated. The first suspicion was related to resonance between the in- and output filters of the models, but it proved to be incorrect.

The reason why the currents are so distorted can be explained with two factors.

- The DAB converter is a topology with high input and output current ripple. Naturally, the DAB converter is a topology with a high input and output ripple (See Figure 4.3). This has been stated from the first paper published by [12]. The control of this type of converter is utilized with fast switching square waves instead of the often used pulse width modulation. With this control approach faster switching cycles can be achieved, because no sine wave needs to be modulated. This enables the topology to reduce passive component sizes (e.g., the inductor \( L_1 \), integrated in the transformer). This is one of the main advantages of the topology.

- The impedance of the grid is very low

Unlike in offshore AC-grids, the DC-grids do not inherit additional inducances from full scale VSC or transformers. Therefore, the grid impedance is generally very low. This implies, that changes in current will be limited only by cable and parasitic impedances in the system. Those values are naturally very low.
Both factors together create a system that delivers a very direct response to outer influences such as switching. This is one of the main differences of DC-grids compared to AC-grids and future grid designers need to pay special attention, when dimensioning the first layouts.

At this point of this work, a decision has to be made, if and in which way the distorted currents influences the operation of the wind farm. Basically, the fault analysis of the DC-grid could be performed with distorted currents, since the DC-voltage is stable and controlled. On the contrary, fluctuating currents imply higher peak currents that increase losses in the system. Moreover, the lifetime of components could be significantly reduced. Therefore, it is assumed that such conditions are not acceptable in the system and need to be eliminated as good as possible. Probably the easiest way to reduce the ripple current in the system is to introduce additional reactors that increase the impedance of the system. It needs to be mentioned that such a step influences the fault behaviour of the system tremendously. The outcome of the fault study will be different compared to a scenario without additional reactors. A second approach to reduce the ripple in the current could be an intelligent converter control. The next section introduces three approaches to limit the current distortion.

### 4.2.1 Approach to reduce the ripple current

This section provides suggestions how to tackle the current ripple problem. Three different approaches are explained.
1. Method: Additional Inductances in the System

This section documents the attempt to smoothen out the currents in the system. Since the impedance of the grid is very low, every switching process distorts the currents. The easiest way to prevent this behaviour is to introduce some additional inductances. Figure 4.4 show the generator current and voltage, after an inductance of \( L = 3\, \text{mH} \) has been added per pole between the generator and the DAB converter. As can be seen, the current ripple is distinctly reduced compared to the unfiltered steady state conditions in Figure 4.2. Notice that additional inductances in the system need to be compliant with the input and output capacities of the modules in the full operation spectrum of the generator.

![Figure 4.4: Influence of Inductances in Generator Output](image)

For the 30 kV grid, three different inductor positions are possible.

1. The inductors are placed on the output terminals of the DAB converters
2. The inductors are placed on the input terminal of the HVDC boost converter
3. The inductors are placed on both, the output terminal of the DAB converter and the HVDC boost converter

The results of the simulations are shown in Figure 4.5. As can be seen, the current shapes are improved in every scenario compared with the steady state results in Figure 4.2. If the inductor is placed close to the wind turbine terminal, the harmonic contribution of the current output for each single turbine (red graphs) is significantly reduced. If the inductor is placed at the HVDC boost converter terminal, mainly the overall line current from all four turbines (blue graphs) is reduced. Therefore, it can be stated that additional inductances in the 30 kV grid can improve the current quality. The best conditions are noticeable
in scenario 3, when both turbine output and HVDC boost converter input are equipped with $L = 10\, \text{mH}$ inductors.

2. Method: Phase Shift between the Converter Modules to limit harmonics

A second possible approach how to limit the switching current has been observed while studying the DAB converter behaviour. This method utilizes the modular structure of the topology that has been utilized and explained in section 3.5.2. In Figure 4.6 the input current flow of the DAB converter is illustrated. On the left hand side, the input currents of all five modules are shown in conventional operation mode. As can be seen, the current inflow happens simultaneously between all five modules. The total input current is comprised of the sum of the currents $I_1...I_5$. 

Figure 4.5: Influence of Additional Inductances - 3 Different Scenarios
Comparison of Concurrent and Distributed Current-Flow in DAB Converter

On the right hand side of Figure 4.6, the modules 2, 3, 4 and 5 operate with a fixed time delay $t_m$ referred to the reference module 1. With this scheme, the current ripple in the total input current is decreased, a higher DC-offset is achieved and the peak current is reduced. The time delay between the modules can be calculated from the switching frequency of the DAB $f_s$, the total amount of sub modules $n$ and the indices of the sub module $m$ that needs to be calculated.

$$t_m = \left( \frac{1}{2f_n} \right) (1 - m) \quad (4.2)$$

Figure 4.7 illustrates the process schematically.

**Figure 4.6:** Comparison of Concurrent and Distributed Current Inflow

**Figure 4.7:** Current Flow through DAB Converter Modules
This process above is illustrated for the LVS of the DAB converter, however, this method can improve the current quality of the HVS and hence for the DC-grid as well. One drawback of this approach is that the modules on the HVS can not be connected in series connection, because of the distributed output voltage. Therefore, the modules need to be connected in parallel and the medium frequency transformers need to perform the whole voltage transformer ratio of \( \frac{30\text{kV}}{1\text{kV}} \) instead of \( \frac{6\text{kV}}{1\text{kV}} \times 5 \).

Unfortunately, due to time limitations, a proper study about the properties of this method is out of scope the for this thesis, but should be performed in the future.

3. Method: Different Converter Topology

If the additional inductor option is not feasible or the reduction of ripple current is not enough, another DC/DC converter topology might be used to limit the ripple in the DC-grid. In [12], a step-up converter topology is introduce that employs a Three Phase Dual Active Bridge (TPDAB) system in the intermediate AC-circuit. Apart from the three phases, the converter topology is similar to the DAB converter and utilizes a phase shift between the bridges. Figure 4.8 illustrates the converter topology with in- and output waveforms. It can be observed, that input and output current ripple is obviously reduced compared to the DAB converter seen in Figure 4.3.

**Figure 4.8:** Voltage and Current for TPDAB - \( i_i = \) Input Current, \( i_o = \) Output Current, with Reduced Ripple [12]
Beside the costs of additional switches in this converter, the leakage inductances in the transformer need to be exactly the same size in order to create a constant DC output current. This might be easy to create in a simulation programm, however, in real life it is rather challenging.

4.2.2 Conclusion Current Quality

For the following fault simulation of the system, the decision has been made to implement current filtering utilizing inductors. Additional inductors, however, will increase the impedance of the DC grid significantly and hence impact the whole fault studies in terms of resonant oscillations, voltage and current decay times (time constants) and peak magnitudes. It needs to be mentioned clearly, that the following fault scenario would look different if no inductors are added to the system.

In a practical application, filter inductors will create resistive losses in the system, that could be avoided. Furthermore, the grid will become more expensive. On the other hand, reduced harmonic currents will create lower losses in cables and switches and components won’t be exposed to high changes of current in time $\frac{di}{dt}$. Furthermore, filter capacitor are not exposed to the ripple. This increases the durability.

The fault simulations are performed with additional inductors at the turbine output terminals. This refers to the suggestion 1 of the previous chapter. The exact position of the filter inductors are marked in Figure 4.1.
4.3 Fault Scenario 1: Short Circuit in Wind Turbine Terminal 1

The first two scenario investigated a fault that appears in the wind turbine. This represents probably the most common fault type in an offshore wind farm. A trigger for this phenomena could be dust in the converter, an insulation failure that leads to a short circuit between the phases or simply a malfunction in IGBT, snubber or gate driver. According to [41], 40% of all offshore system faults are related to converter faults. As indicated in the overview of Figure 4.1, the fault appears at the high voltage terminal of the DAB converter of turbine terminal 1. After the fault is detected by over-current sensors, the breaker at bus bar 1 disconnects the turbine 1 from the rest of the grid.

Figure 4.9: Short Circuit in Wind Turbine Terminal 1
The graphs illustrated in Figure 4.9 show the behaviour of the positive phase in different positions in the grid after the fault is applied at $t = 0.6$ and resolved at $t \approx 0.6015$. In the upper graph the currents are indicated. During the fault, the normally positive currents from wind turbine 1 and the boost converter reverse the current direction and become negative. This can be explained with the sudden appearing current path through the short circuit. Figure 4.10 illustrates the incident. Therefore, the breaker needs to be able to operate in both the conventional and the reversed current direction. Notice that the current measurement in the simulation is performed close in the breaker. The magnitude of the negative current in the positive phase is determined by the breaker reaction time. The filter-inductor limits the current flow from the DC-grid side. As can be seen, current oscillations from the input of the boost converter appear after the breaker separated the turbine from the grid.

![Diagram](image)

**Figure 4.10:** Current Flow in Phase to Phase Terminal Fault

The voltages $V_{BB}$ at the different bus bars decrease slowly due to the filter inductances between the short circuit and the bus bar. After the breaker reacted and isolated the turbines, transient oscillations occur due to the sudden change of current flow. The magnitude of the DC voltage reaches $V_{BB} = 120$ kV.

The behaviour of the HVDC connection can be seen in the bottom part of 4.9. It is observable, that the voltage drops first, but stabilizes back to the reference voltage of 300 kV. As can be expected, the current drops after the turbine fault to around $\frac{3}{4}$ of the steady state value.

Generally, it can be said, that the fault in the turbine generates high voltages and current stress for the grid. It's necessary that all components need to withstand those magnitudes for a short duration. However, it can be said that the control system of the components are able to recover the voltages and currents after the fault to the steady state values and the wind farm can operate further.
4.4 Fault Scenario 2: Short Circuit in Wind Turbine Terminal 4

The second scenario is very similar to scenario one, with the difference that the fault appears in wind turbine terminal 4. Therefore, the grid impedance at the fault position is slightly different from scenario one since the wind turbine is located at the end of the feeder. However, as can be seen from Figure 4.11, the behaviour of the current and voltages are very similar. The only difference compared to scenario 1 are the slightly higher peak values for current and voltages due to the slightly lower grid impedance at the end of the feeder.

Figure 4.11: Short Circuit in Wind Turbine Terminal 4
4.5 Fault Scenario 3: MVDC Cable Fault between Turbine 2 and Turbine 3

The third fault scenario simulates a SLG fault in the positive phase between turbine 2 and turbine 3 (See Figure 4.1). This could refer to an insulation failure due to wear or ageing of the cable or a ship, that crosses the wind farm with the anchor set.

Figure 4.12: Single Line to Ground Fault between Turbine 2 and 3
The reaction of the short-circuit in the cable is illustrated in Figure 4.12. The fault is applied at \( t_{\text{fault}} = 0.6 \text{s} \) and the breaker isolates the faulted turbines 3 and 4 at \( t_{\text{brk}} \approx 0.601 \text{s} \). Afterwards, distinct current oscillations with a fundamental frequency of \( f_{\text{osc}} = 225 \text{Hz} \) are noticeable at the output terminals of the remaining wind turbines after the fault. Furthermore, high frequency peak voltages of \( V_{BB_2} \) are detected that decay in between \( t_{\text{decay}} = 0.25 \text{ms} \). Afterwards, the 30 kV grid voltage is restored.

Figure 4.13 shows the currents of the line terminals during the fault. It can be seen that fault currents at line terminal 3 increases in the positive direction with a peak value of 2.4 kA. After 1.3 ms the fault is cleared. The currents from line terminal 2 reverse the current direction in case of a fault, because the ground connection provides suddenly a lower potential. A schematic of the process is shown in Figure 4.14. It is observable that the current reaches peak values up to -14.4 kA during the fault. The higher peak current can be explained with the sudden de-charge of the filter capacitor in the high voltage boost converter. In the simulation, it was necessary to implement a big filter capacitor to reduce the ripple of the DC grid voltage. A 600\( \mu \text{F} \) capacitor is able to store

\[
Q = CU \\
Q = 600 \mu \text{F} \times 30 \text{kV} \\
Q = 18 \text{As} = 18 \text{kAms}
\]  

(4.3)

Since the grid impedance is solely determined by the properties of cables, the capacitor time constant can be expected to be low. Note that the simulation does not include parasitic influences of capacitors that might further decrease the magnitude of the peak currents.
4.5 Fault Scenario 3: MVDC Cable Fault between Turbine 2 and Turbine 3

The reason, why the breaker at line terminal 2 reacts a faster, can be explained with the faster current rise time of line terminal 2. Therefore, the short-circuit detection threshold value is reached earlier.

The influence of the HVDC voltage and current is observable in the bottom part of Figure 4.12. The voltage decreases first, but stabilizes after $t = 0.2s$ at the 300 kV reference. As expected, the transmitted HVDC current is halved after the two turbines are isolated.
4.6 Fault Scenario 4: Influence of Lightning Faults from AC-Grid

The fourth scenario refers to lightning occurrences that might appear in overhead lines in the AC-grid connection on land. The fault current might be transmitted through the HVDC connection into the offshore wind farm and cause disturbances in the converter. Figure 4.15 shows two scenarios of different lightning strengths, that are exposed into the AC grid close to the HVDC-inverter terminal. The first simulated stroke has a current amplitude of $I_{\text{peak}1} = 40\, \text{kA}$ with a rise time $t_1 = 2\, \mu\text{s}$ and a decay time $t_2 = 50\, \mu\text{s}$, whereas the parameters of the second stroke are $I_{\text{peak}2} = 120\, \text{kA}$ with $t_1 = 1\, \mu\text{s}$ with $t_2 = 100\, \mu\text{s}$ (See Figure 4.16).

![Lightning Fault from AC Grid - HVDC Voltage and Currents](image)

**Figure 4.15:** Lightning Strikes in AC Grid
4.6 Fault Scenario 4: Influence of Lightning Faults from AC-Grid

As can be seen, the difference of the transient state between weak and heavy lightning strike are distinct. Voltage and current oscillations in the HVDC connection are observable, but the control structures are able to regulate the system back to steady state operation after \( \approx 0.15 \) s. In the simulation, no effects of the lightning strike were able to be noticed in the DC-grid. The 30 kV DC-bus voltage and line current remains without oscillations. The reason for this occurrence can be explained with the topology of the boost converter (See Figure 3.25). The big line filter capacities absorb the voltage oscillation and the diode from the boost-converter prevents reversed current flow in the phase of the DC-grid.

![Lightning Strike PSCAD Model](image)

**Figure 4.16: Lightning Model Approach**

Figure 4.16 illustrates the method and location, the lightning strikes are introduced to the system. Basically, a timed current source is used to apply a defined current pulse into one of the phases. This representation can be seen as an approximation of a lightning strike, since the behaviour is more complex in nature.
Overview Loss and Efficiency for DC Grid

This short chapter provides an overview of the losses of the models presented in Chapter 3. In Table 5.1, the individual contributions of loss from the models are listed. Note that the HVDC boost-converter and inverter are not a part of this investigation, since this thesis is directed to medium voltage DC grids. The purpose of these models was to provide a transient response with the AC-grid on land. Furthermore, the BIDTC has been withdrawn from the evaluation due to uneconomical conduction losses in the predefined operating conditions (See Section: 3.5.1). However, this converter type might be suitable for different operating conditions.

With the simulated conditions, the overall efficiency can be stated as 94.4%. This number is determined with the assumption that four 5 MW turbines are connected in parallel and form a feeder to the HVDC boost-converter platform.

It can be stated that the main loss is created in the DAB converter. There are two possibilities to decrease the value:

(a) change the amount of parallel operating sub-modules in the converter
(b) increase the generator output voltage
This provides the possibility to increase the overall efficiency of the DC grid. More details regarding DAB losses can be found in Section 3.5.2. With the detailed analysis and the elucidation in Chapter 3.5.2.6, it is valid to assume that the DAB losses can be halved in future designs. This would boost the overall efficiency of the system to 96.9%.

The losses in the cables depend on the length of the cable and the conducting current. Since the cable layout is a bipolar design, the losses need to be doubled. To decrease cable losses, the medium voltage of 30 kV could be increased to a higher level. A further decrease of the distance between the turbines is not an option, since the recommended distance between them can be estimated as 5 to 9 rotor diameters [12]. Assuming a 5 MW wind turbine with a rotor diameter of 90 m, the limit is already reached with the assumed 450 m distance.

\[
\begin{array}{|c|c|c|}
\hline
\text{Loss [kW]} & \text{Wind Turbine} & \text{Cables} \\
\hline
\text{1 Turbine} & \text{4 Turbines} & \\
\hline
\text{Diode Rectifier} & 18 & 72 \\
\text{DAB Converter} & 250 & 1000 \\
\hline
\text{Loss Turbines [kW]} & & 1072 \\
\hline
\text{Loss Cables [kW]} & & 40.1 \\
\hline
\text{Total Loss [kW]} & & 1112.1 \\
\text{Efficiency} & & 0.944 \\
\hline
\end{array}
\]

\textbf{Table 5.1}: Overview of Losses in a DC grid - Full-Load Conditions
Chapter 6

Future Work

In principle, the evaluation of a DC grid is a very complex task and many parameters influence the operation of such a system. With this master thesis, an overview of possible layouts, converter structures, and transient responses is given. Moreover, an overall loss estimation of grid and converter is provided. However, in order to prepare the way for a practical offshore DC grid, several problems need to be tackled.

- Further research regarding the influence of the low grid impedance
  As has been mentioned in this work, the low grid impedance in DC collection grids leads to heavy current fluctuations due to the switching of the converters. A way to reduce this ripple has been proposed in this work, but additional inductors create losses and are expensive. Therefore, if possible, a different solution should be found.

- DC grid fault study without inductors
  The fault studies should be repeated without inductors, if the current oscillations can be decreased. Current rise and fall times, as well as peak values, are expected to change.

- DC breaker operation and detection methods
  Breaker operation and detection methods for DC grids have just been mentioned in this work and comprise a whole research topic on their own. The
possibility to isolate faulty lines with a redundant cable can be interesting when several wind turbines are connected to one feeder.

- **Evaluation of BIDITC with higher input voltage**
  The BIDITC is an interesting converter concept and should be studied when applied to a DC grid with higher generator output voltages. Conduction losses would decrease significantly, and the converter concept could be economical.

- **Improvement of DAB converter**
  The possibility to operate the DAB converter modularly provides room for improvements. The best amount of modules for a given apparent power is determined by the individual application. An optimization study about converter loss, versus economical aspects and modular stages, is possible. The converter operation might be more economical with 20 sub-modules instead of 5 sub-modules. Furthermore, the possibility of timed current inflow between the sub-modules should be studied closer.

- **Influence of faults to other feeders**
  Since this study investigated just one feeder of an offshore wind farm, the effect of faults between the feeders is lacking.

- **Effect of varying wind conditions**
  A study that investigates the grid operation with changing loads to the generator should be performed to see effects on current and voltage.
Chapter 7

Conclusion

The main objective of the thesis was to analyse and validate DC collection grids for offshore applications regarding operability, fault behaviour, and efficiency. To achieve this goal, the first step was to define boundary conditions for the network and to illustrate the main differences between AC and DC collection grids. An overview of merits and challenges that need to be faced when considering DC grids summarizes Chapter 2.

The next step comprised the detailed study of electrical representations of the individual components in a DC grid. The goal was to find models that are able to reflect realistic behaviour of the components. Therefore, state of the art technologies have been reviewed, and newly proposed concepts have been evaluated. One main focus of this work was to find appropriate bidirectional DC/DC converter topologies. The two most promising concepts have been identified as the DAB and BIDITC converter. The DAB converter has a simple layout, an easy control pattern, a good transient response, and the possibility to share the load between several modules. The drawback of this concept is the high input and output current ripple. The BIDITC provides the possibility of a transformerless converter design that is operated with thyristors as switches. Furthermore, the frequency controlled converter was able to operate in discontinuous current conduction mode, resulting in no switching losses. Unfortunately, both the low and high voltage side need to withstand the full medium voltage $V_{ac} = \sqrt{2}V_{dc}$, leading to a high amount of stacked thyristors in the converter. Moreover, the high conducted currents lead to unreasonably high losses with an input volt-
Due to these reasons, the BIDITC has been withdrawn from the fault simulation; however, the converter system might be a suitable topology for higher generator voltages. The wind turbine representation is realized with a synchronous machine that is loaded with nominal torque $T = 1pu$, and the AC output is rectified with a passive six-pulse diode bridge. The medium voltage cables are represented with a lumped $\pi$-model that is proven to be as accurate as more complex models for short distances. The HVDC cables are modelled with the PSCAD frequency deepened model. Finally, the HVDC inverter and rectifier models provide a good transient response to the main grid on land.

In Chapter 4, after all models had been chosen and tested individually, the DC grid simulation in PSCAD was created. Since computation limitations restricted the simulation of a whole wind farm, just one feeder that connected four wind turbines was evaluated. First, the steady-state voltages and currents were monitored. It was observed that the voltages were stable and did not possess a high ripple; however, the currents in the DC grid were full of harmonics. Two reasons have been identified for this occurrence. Firstly, the grid impedance of the collection network is determined solely by the intrinsic medium voltage cables and can be estimated with a very low value. This means that, unlike in AC grids where huge transformers characterise the grid impedance, the response to switching will be more distinct. Secondly, the high input and output ripple of the investigated DAB converter propagates the disturbance for the currents further. Three different approaches to solve the ripple current issue were proposed, and the introduction of additional inductors was an efficient method to reduce the ripple current, which has been implemented in the PSCAD simulation in the end. With the improved operating condition, the fault analysis was conducted. Four different fault scenarios were investigated and explained, which reflect the most common fault types in offshore wind farms. The first two scenarios are directed to wind turbine converter faults that are located at the beginning and the end of the feeder. The third scenario investigates a SLG cable fault between two turbines that might happen if a ship damages the cable with the anchor down. The last scenario studies the effect to the DC grid if a lightning strike hits the AC connection on land. Every turbine and cable segment was protected with a set of DC breakers that were able to detect and isolate faulty turbines or cable segments. In case of a turbine converter fault, the breaker detected the over-current and isolated the turbine. High current and voltage oscillations appeared, but the control of the grid was able to be restored to the rated voltage in $\approx 3ms$. The study of the cable fault in the third scenario resulted in high current oscillations, but after the fault location was isolated, the system was able to return to steady state conditions. Lightning strikes did not influence the DC grid with the proposed converter setup; however, distinct HVDC current and voltage oscillations were observable.

Chapter 6 summarised both Chapter 4 and 5 and presented the calculated losses for the DC grid. With the given layout, the efficiency of the network was determined to be $\eta = 94.4\%$. In addition, a proposal regarding an improved DAB converter setup was presented.
converter operation is stated to further increase the efficiency up to $\eta = 96.9\%$. 
A.1 Terminology for Converter Setup

Converter related Terminology

Figure A.1: General Terminology used in the report
A.2 Control for HVDC-Inverter

Figure A.2: Conversion of reference voltages and current from AC side

Figure A.3: Controller for HVDC
A.3 ABC to DQ0 Transformation

\[
\begin{bmatrix}
v_d \\
v_q \\
v_0
\end{bmatrix} = \frac{2}{3} \begin{bmatrix}
\cos(\omega t) & \cos(\omega t - \frac{2\pi}{3}) & \cos(\omega t + \frac{2\pi}{3}) \\
-\sin(\omega t) & -\sin(\omega t - \frac{2\pi}{3}) & -\sin(\omega t - \frac{2\pi}{3}) \end{bmatrix}
\]  

(A.1)

Source Matlab 2013b
A.4 Overview of the DC Grid in PSCAD
Dear Stephan

...

For high power IGBTs the data sheet usually gives 10-12kW as the maximum loss for a component with the condition that the temperature of the semiconductor is below the maximum temperature. Due to the area of the baseplate on the component and the efficiency of the heat sink calculation shows that the maximum continuous power dissipation will be about 6kW.

Tonny W. Rasmussen
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