

High-Power Medium-Voltage DC-DC Converters: Design, Control and Demonstration

Nils Soltau



High-Power Medium-Voltage DC-DC Converters: Design, Control and Demonstration

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

vorgelegt von

Diplom-Ingenieur
Nils-Jörgen Soltau
aus Lippstadt

Berichter:

Univ.-Prof. Dr. ir. Dr. h. c. Rik W. De Doncker
Univ.-Prof. Antonello Monti, Ph.D.

Tag der mündlichen Prüfung: 3. März 2017

**Diese Dissertation ist auf den Internetseiten der Hochschulbibliothek
online verfügbar.**

Bibliographische Information der Deutschen Nationalbibliothek

Die Deutsche Nationalbibliothek verzeichnet diese Publikation in der Deutschen Nationalbibliografie; detaillierte bibliografische Daten sind im Internet über <http://dnb-nb.de> abrufbar.

D 82 (Diss. RWTH Aachen University, 2017)

Herausgeber:

Univ.-Prof. Dr.ir. Dr. h. c. Rik W. De Doncker

Direktor E.ON Energy Research Center

Institute for Power Generation and Storage Systems (PGS)

E.ON Energy Research Center

Mathieustraße 10

52074 Aachen

E.ON Energy Research Center I 43. Ausgabe der Serie

PGS I Power Generation and Storage Systems

Copyright Nils Soltau

Alle Rechte, auch das des auszugsweisen Nachdrucks, der auszugsweisen oder vollständigen Wiedergabe, der Speicherung in Datenverarbeitungsanlagen und der Übersetzung, vorbehalten.

Printed in Germany

ISBN: 978-3-942789-42-4

1. Auflage 2017

Verlag:

E.ON Energy Research Center, RWTH Aachen University

Mathieustraße 10

52074 Aachen

Internet: www.eonerc.rwth-aachen.de

E-Mail: post_erc@eonerc.rwth-aachen.de

Herstellung:

Druckservice Zillekens

Rainweg 19

52224 Stolberg

Vorwort

Diese Dissertation entstand während meiner Tätigkeit als wissenschaftlicher Mitarbeiter am Institute for Power Generation and Storage Systems (PGS) des E.ON Energy Research Centers der Rheinisch-Westfälischen Technischen Hochschule Aachen. Die Arbeit befasst sich mit der Steuerung, Auslegung und dem Aufbau eines Hochleistungs-Gleichspannungswandlers für Mittelspannung. Bei meiner Arbeit haben mich viele Menschen unterstützt, denen ich an dieser Stelle danken möchte.

Meinem Doktorvater Professor Rik W. De Doncker möchte ich herzlich für seine Betreuung und Unterstützung danken. In den Jahren am Institut hatte ich die Möglichkeit an vielen interessanten Themen zu arbeiten und eine Menge zu lernen. Insbesondere möchte ich mich für das Vertrauen bedanken, welches mir während der spannenden Zeit als Oberingenieur durch Herrn Professor De Doncker geschenkt wurde. Herrn Professor Monti möchte ich für die Übernahme des Korreferats, die wertvollen Anmerkungen und das Interesse an meiner Arbeit danken.

Meinen Freunden und Kollegen am PGS und am Institut für Stromrichtertechnik und Elektrische Antriebe (ISEA) möchte ich ebenfalls ganz herzlich danken für eine unvergessliche Zeit. Insbesondere bedanke ich mich bei Marco Stieneker, auf dessen Unterstützung ich immer bauen konnte; insbesondere als wir gemeinsam als Oberingenieure unsere Dissertationen finalisiert haben. Weiterhin danke ich Stefan Engel für die spannende und schöne Zusammenarbeit an unseren gemeinsamen Veröffentlichungen. Bei Robert Lenke, Hanno Stagge und Johannes Voss möchte ich mich für das Korrekturlesen meiner Doktorarbeit und die hilfreichen Anmerkungen bedanken. Darüber hinaus haben mich tatkräftige Studenten während meiner Promotion unterstützt, insbesondere danke ich Hafiz Siddique, Marius Mechlinski, Andreas Nies und Julian Lange.

Schließlich möchte ich mich bei meinen Eltern, meiner Familie und Freunden bedanken. Mein wesentlicher Dank gilt zudem Kathi für Rückhalt, Verständnis und Ermutigung. Vielen Dank!

Düsseldorf, im April 2017

Nils Solla

Abstract

In the last years, power generation from renewable energy sources has increased steadily. On the one hand, this enables the independence from fossil fuels and the potential for decreasing electric-energy cost. On the other hand, the share of volatile power generation in the electricity grid increases. Moreover, while in the past power has been generated centrally in large power plants and injected into the high-voltage transmission grid, now vast amounts of electric energy are fed into the low- and medium-voltage grid. Consequently, the electricity grid needs to handle fluctuating power generation and reversal power flows from the lower to the higher voltage grids more often. Since the classical electricity distribution grid has been designed for unidirectional power flow and only has limited control capabilities, measures have to be taken.

In modern applications, direct current (dc) has proven advantageous compared to the classical alternating current (ac) technology. It allows the transfer of more power since the utilization of cables and lines is higher. Moreover, renewable energy sources, which are mostly dc or variable ac sources, can be connected more efficiently to a dc system. Since these dc systems use power-electronic converters, power flows can be controlled actively to allow optimal utilization of lines and the limitation of fault currents. Moreover, since these converters operate at relatively high frequencies, valuable materials like copper and steel, used for 50 Hz transformers, can be saved. The utilization of dc in the medium-voltage distribution grid would increase the capacity of the existing cables and adds control capability to cope with the volatile power generation from renewables [1].

This work is dedicated to the implementation of a high-power medium-voltage dc-dc converter and the construction of a full-scale mega-watt prototype. This converter is used to step up or step down dc voltages and is therefore considered as the equivalent of the transformer in the classical ac system. Also based on the previous work of Lenke [2], the dual-active bridge (DAB) has been identified as optimal dc-dc-converter topology for a grid application. It allows bidirectional power flow, provides galvanical isolation and utilizes zero voltage (ZV) or zero current (ZC) switches, which reduce the losses in the converter significantly.

Chapter 1 gives the motivation for dc systems. Based on a short historical digres-

sion, the advantages of dc are discussed in detail. Moreover, several applications on low, medium and high voltage levels are introduced, in which dc is contemplated or has been implemented successfully already. Different dc-dc converters suitable for high-power medium-voltage applications are introduced and an overview on different research works is given.

In chapter 2, the DAB converter is modeled to allow the proper design of voltage and current control. Based on the findings, a fast current controller is introduced and demonstrated. Moreover, a balancing control is presented compensating the effects of asymmetric transformers.

Chapter 3 discusses the soft-switching nature of the converter. The use of an auxiliary resonant-commutated pole (ARCP) to ensure soft-switching in the entire operating range is discussed in detail. Different devices in the ARCP are compared in terms of losses and costs. Finally, the pay-back period of such a system is calculated for an exemplary application based on the costs for the developed full-scale prototype.

The high-power medium-frequency transformer is discussed in chapter 4. Through its high switching frequency, the transformer enables the high power density of the dc-dc converter. Based on a short introduction, different core and winding materials are discussed for the use in the high-power high-frequency transformer prototype. Since the voltage waveforms in the DAB are square shape, the impact of the higher harmonics on the core losses is evaluated in detail. Finally, different methods to increase the stray inductance of the transformer are compared. The stray inductance of the transformer shall serve as an integrated filter element for the converter. The findings result in the design of a 2.2 MVA transformer to be operated in a 5 kV dc-dc converter.

Finally, chapter 5 presents the construction and the setup of the high-power medium-voltage demonstrator of the DAB converter. This chapter combines the findings from the previous chapters and gives some impression on the construction and the commissioning of the converter. The medium-voltage dc-dc converter is rated for 5 kV dc-link voltage and a power of 5 MW.

The work is completed with a summary and outlook.

Contents

Abstract	iii
1 Introduction	1
1.1 “The War of Currents”	1
1.2 DC in a Nutshell	2
1.3 Low- and High-Voltage DC Applications	4
1.4 Medium-Voltage DC Applications	5
1.4.1 Target Applications	5
1.4.2 State-of-the-Art	13
2 Control of the Three-Phase Dual-Active Bridge Converter	19
2.1 Operation Principle	19
2.2 Different Operating Areas	22
2.3 Modeling and Dynamic Behavior	22
2.3.1 State-Space and State-Variable Averaging	23
2.3.2 First-Harmonic Approximation	27
2.3.3 Verification and Comparison	29
2.4 Instantaneous Current Control	30
2.4.1 Applying the Clarke Transformation	30
2.4.2 The Instantaneous Current Control (ICC)	32
2.4.3 The Improved Instantaneous Current Control (I^2CC)	37
2.4.4 Verification	40
2.5 Closed-Loop Control	45
2.6 Balancing Control	46
2.6.1 Motivation	46
2.6.2 Impact of Asymmetric Impedances on the Converter	47
2.6.3 Compensation of Asymmetric Impedances	50
2.6.4 Optimization Objectives	51
2.6.5 Practical Implementation	53
3 Power Semiconductors and Soft-Switching Operation	59
3.1 Preferable Power-Electronic Devices	59
3.2 Ensuring Soft-Switching Operation Using the Auxiliary Resonant-Commutated Pole (ARCP)	63
3.2.1 ARCP Operation	64

3.2.2	Design Considerations	65
3.2.3	Loss Modeling	67
3.2.4	Simulation Results	68
3.2.5	Measuring Results	71
3.2.6	Advantages of Applying an ARCP	75
4	Medium-Frequency High-Power Transformer	81
4.1	Introduction	81
4.1.1	Review on different applications	81
4.1.2	Windings	83
4.1.3	Core Materials	85
4.2	Core Losses in a Dual-Active Bridge Application	90
4.2.1	Core Loss Measurement	90
4.2.2	Analytical Validation	92
4.2.3	Core Losses Under Load Conditions	94
4.3	Stray Inductance Design Considerations	97
4.3.1	Stray Inductance Calculation	97
4.3.2	Integrated Stray Inductance	104
4.3.3	Adjustment Precision and Tolerances	105
4.3.4	Distinct Series Inductances	114
5	Medium-Voltage High-Power Demonstrator	119
5.1	Control Implementation	119
5.1.1	FPGA Implementation	120
5.1.2	PowerPC Implementation	122
5.2	Converter Design	125
5.2.1	Clamping Circuit	126
5.2.2	DC Link	128
5.2.3	Auxiliary Power Supply	128
5.3	Medium-Frequency Transformers	128
5.4	Commissioning	131
5.4.1	Single-Phase	132
5.4.2	Influence HF Litz	136
5.4.3	Three-Phase Configuration	139
6	Conclusion and Outlook	145
	Bibliography	149
A	Glossary	167
B	List of Symbols	169

C	List of Figures	171
D	List of Tables	175

1 Introduction

This chapter gives the motivation for dc technology and high-power medium voltage (MV) dc-dc converters. An overview of dc technology and dc-dc converters is given. This overview includes advantages, challenges and the current state of research regarding dc technology.

1.1 “The War of Currents”

Today’s transmission and distribution of electric energy mainly relies on ac grids operated at a frequency of 50 Hz or 60 Hz. At the end of the 19th century, ac was established since, with the invention of the transformer [3, 4], it allowed, compared to dc, an easier and more efficient voltage transformation at that time.

Soon people realized that the transformation to higher voltage levels is essential to efficiently transmit electric energy over long distances. In Germany, the first three-phase high-voltage alternating current (HVAC) transmission line, connecting the cities of Lauffen and Frankfurt, was commissioned in 1891 on the occasion of the International Electrotechnical Exhibition. The rugged ac technology enabled a power transfer across 176 km at an operating voltage of 15-25 kV. From the high voltage (HV) generator located in Lauffen to the low voltage side of the transformer placed in Frankfurt a peak efficiency of 74.5 % was achieved [5].

In those years, HVAC generators and efficient transformers made the success of HV energy transmission possible. This decided the so-called “war of currents” and established ac technology for transmitting and distributing electric energy [6].

In the present days, the advances in materials research and semiconductor devices allow an efficient voltage conversion for dc as well. This motivates the revision of ac versus dc and the advantages, disadvantages and potential of dc technology.

1.2 DC in a Nutshell

Although today's energy distribution relies on ac, dc is already advantageous in many applications. The main drawbacks of the classical ac are reactive power demand, inflexibility and skin and proximity effects. At the same time, dc technology allows a higher utilization of materials. Finally, the main problem with dc, the absence of dc circuit breakers, has been solved and demonstrated through full-scale prototypes [7]. In the following, these issues are introduced to illustrate the motivation for dc technology concisely.

As said before, one of the main drawbacks of ac systems is the required reactive power demand. Reactive currents generate losses, however, they are not transferring active power. They result from the reactive impedances of transmission lines, cables and the loads. The reactive line impedance and its reactive power demand increase with the length of the line or the cable. Long distance cable connections have a large reactive power demand, which impedes an efficient energy transfer. Hence, the connection of offshore wind farms far out in the sea with ac undersea cables is not economic.

The skin effect is another disadvantage in an ac systems. It is caused by the magnetic field of ac currents flowing through conductors, causing a displacement of the currents. This current-displacement effect reduces the effective area of the conductor and increases its resistance consequently. The effect, which is also explained in detail in chapter 4.1, is especially critical for higher frequencies. For conductor diameters larger than 20 mm, however, the skin effect is noticeable even at 50 Hz, thereby decreasing the efficiency [8].

Another issue with the classical ac power system is that since market liberalization the mix of power generation drastically changed: the share of decentralized and renewable energy source connected to the grid increases steadily. Most of these sources are volatile and vary their power generation constantly, e.g. with the weather or the season. To compensate these fluctuation, power flows have to be controlled and rerouted in the power system. The classical power system, however, only has little control capability. It has been built at a time, when generation was well determined solely through large power plants. Moreover, the large and heavy generators in the power plants have offered sufficient inertia to stabilize the grid during short transients. Now, different measures have to be taken in ac grids to control power flows and to ensure its stability and efficient operation. The measures are numerous: "phase angle regulators", "subsynchronous resonance damper", "static volt-ampere reactive (VAR) compensators", "static synchronous compensators (STATCOMs)", "static synchronous series compensators (SSSCs)".

Condensed in the expression flexible ac transmission system (FACTS), these devices are required to maintain the power quality of modern ac networks [9, 10].

In dc systems, many of the above-mentioned issues are not present like reactive power demand and skin effect. Moreover, power electronics in dc converters allow active power flow control in (meshed) grids.

Furthermore, the power-transfer capability of dc is higher as well. This is due to the fact that the insulation of ac cables is rated for the maximal peak voltage. For the power transfer, however, the root-mean-square (RMS) value of the voltage is decisive. This is illustrated by a MV cable suitable for ac as well as dc: HELUTRAIN 4GKW-AXplus [11]. In an ac system, this exemplary cable is rated for 3 kV RMS line-to-line voltage. For operation in a dc system, the cable is approved for ± 2.7 kV. Comparing the maximal power that can be transferred per number of wires n , whereas the ac system is operated with three wires and the dc system is operated with two wires, leads to:

$$\frac{P/n|_{\text{dc}}}{P/n|_{\text{ac}}} = \frac{\frac{2 \cdot 2.7 \text{ kV} \cdot I}{2}}{\frac{\sqrt{3} \cdot 3 \text{ kV} \cdot I}{3}} = 1.56. \quad (1.1)$$

This best-case scenario neglects the influence of the skin effect and the reactive power demand. Although, the power-transfer capability using dc can be increased by 56 % compared to ac [12].

In a similar manner, dc can increase the power-transfer capability of overhead power lines. According to [13] the same tower structure can be used for operation with ± 380 kV dc instead of 220 kV ac. This corresponds to a three-fold power increase.

Interrupting fault currents in a dc system has been an issue for a long time. Breaking current in an inductive line generates an arc between the opening contacts, where the contacting medium might be air, vacuum or sulfur hexafluoride (SF_6) [14–16]. The arc extinguishes when the current crosses zero [17]. In an ac system, the current crosses zero naturally about every 10 ms, leading to the extinction of the arc. In a dc system, a natural zero crossing is not present. Consequently, the approach of conventional ac circuit breakers is not working for dc. Instead, the zero crossing has to be generated artificially.

Although breaking fault currents in a dc system, has been considered as the biggest

challenge, there are different circuit breaker approaches today. DC circuit breaker can be roughly divided into three categories: mechanical, solid-state and hybrid circuit breakers [18, 19].

Mechanical breakers are relatively slow. Thereby, the fault current to be disconnected is comparatively high. This makes mechanical dc circuit breaker rather interesting for low-voltage applications. However, the mechanical switches offer low on-state losses and they are cheaper compared to solid-state switches.

Solid-state breakers consist mainly of power-electronic devices [20–22]. Due to the absence of moving parts, solid-state breakers are faster. However, their conduction losses are considerably higher because of the higher on-state voltage of the semiconductor compared to a mechanical switch.

The hybrid circuit breaker has a mechanical and a solid-state switch. The mechanical switch offers low losses during normal operation. The actual breaking of the fault current is mainly achieved by power-electronic devices. Hybrid circuit breakers have been demonstrated for MV applications [23]. Moreover, ABB has first presented a 9 kA breaker operating at 320 kV for high-voltage direct current (HVDC) applications [7].

Consequently, many promising concepts have been identified and demonstrated in practical setups. Now, through further development and experience in the field, dc circuit breaker have to become as reliable and cost competitive as their ac equivalents.

1.3 Low- and High-Voltage DC Applications

There are numerous applications in which the advantages of dc could be demonstrated.

On a low voltage level around 400 V, dc is especially promoted for data and telecommunication centers. Since servers and IT equipment in these buildings are dc loads, hard switched PWM inverters can be saved if a dc distribution is used. Using a scaled prototype, 7 % energy savings were demonstrated using 400 V dc instead of 480 V ac [24]. Moreover, [25] shows that dc data centers have an increased availability and a lower probability of failures over time.

In aircrafts, the electrical power consumption increases emphasized by the improved

performance, reliability and life cycle cost of electrical systems compared to mechanical, hydraulic and pneumatic systems. Whilst traditionally these airplanes are operated at high-frequency ac (400-800 Hz), the trend is towards dc distribution systems operating at 270 V or 540 V due to the higher power density [26–28].

As mentioned before, ac transmission is not economic for long distances due to the reactive power demand. If ac is not economic, HVDC is used instead. The connection of an offshore wind park via cable is a popular HVDC application. Especially for cables, due to the relative proximity of the conductors or shielding, the reactive power demand is higher than for overhead lines [29, 30]. A point-to-point HVDC connection, with its large fixed cost for the power-electronic converters, is considered more economic for wind parks farther than around 100 km from the shore (depending on many different factors like soil conditions) [31].

1.4 Medium-Voltage DC Applications

One can assume that applications at MV levels benefit similarly from dc. Consequently, MV applications and MV dc-dc converters are investigated further in the following.

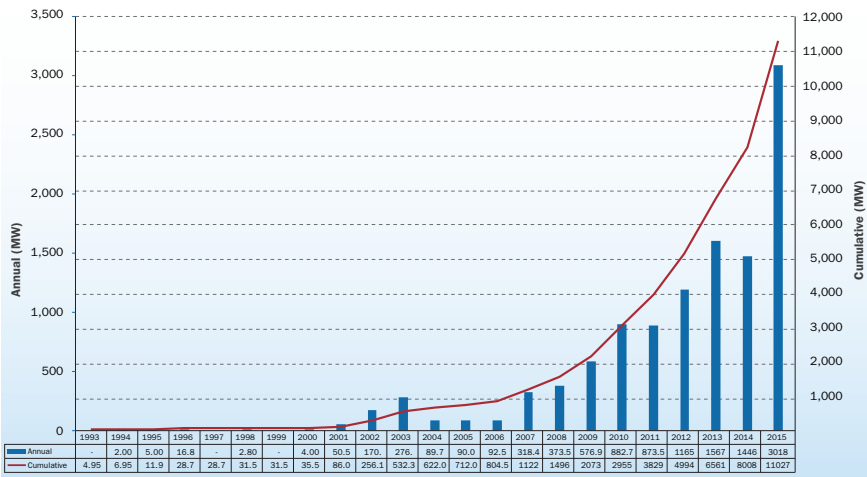
1.4.1 Target Applications

1.4.1.1 Offshore wind farms

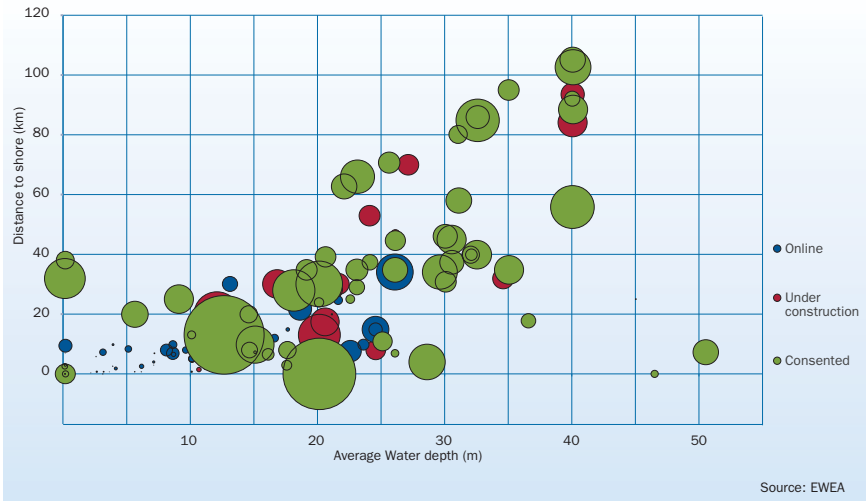
The capacity of offshore-wind electricity generation has heavily increased over the last years, as illustrated in Fig. 1.1 (a).

As technology develops and experience is being gained, the trend is to move large-scale wind farms further away from coast lines, i.e. into deeper waters [32]. Figure 1.1 (b) shows the distance from the shore and the water depth of wind farms, planned for development after 2015. Consequently, to be economical, the number of parks that need an HVDC connection increases.

A central platform transforms the wind energy to HVDC level to transmit it to the shore. The collector grid connects every wind turbine of the park with this converter platform as illustrated in Fig. 1.2. It should be noted that the schematic assumes the use of full-power converter topology. Since it allows the use



(a) Installed offshore wind capacity in Europe (1993-2015)



(b) Distance and depth of planned offshore wind farms; bubble size represents wind-farm capacity (status in 2013)

Figure 1.1: Offshore wind facts (Source: [32, 33])

of standard components, the collector grid is usually run at 50 Hz. AC collector grids require filter components, which are quite heavy and bulky at the given frequency.

Furthermore, a converter stage is needed to connect the wind turbines to the collector grid since the wind generators run at variable frequency.

A more efficient and reliable approach is to remove the inverters and use a dc collector grid instead [18]. Several dc-collector-grid topologies are depicted in Fig. 1.3. The optimal topology depends on the wind-park power and its dimensions. These issues are discussed in detail in [34].

Besides low losses, also heavy and bulky 50 Hz components become redundant in a dc collector grid, which is especially desirable in an offshore application. Furthermore, a dc collector field is able to integrate dc sources like battery energy storage systems and electrolyzers more efficiently.

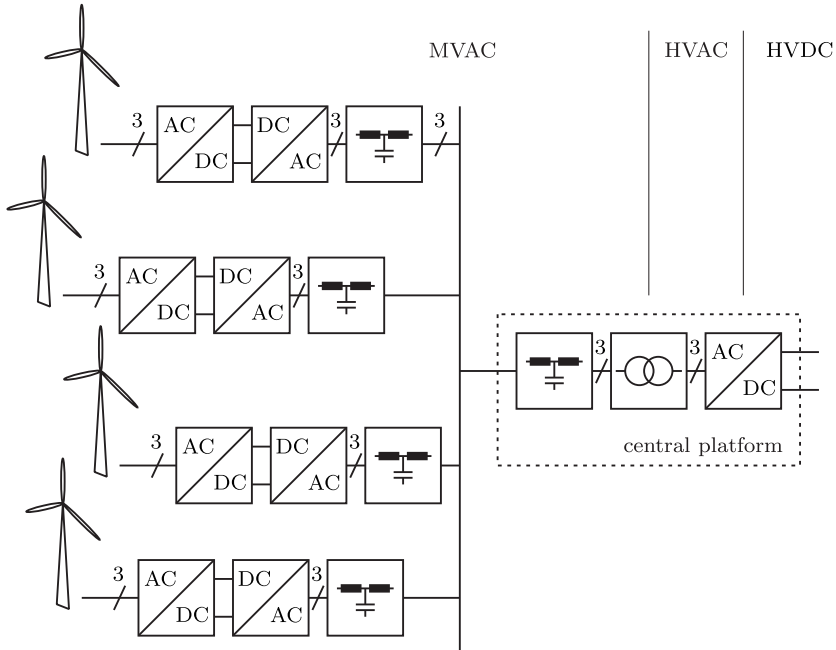


Figure 1.2: Typical ac offshore collector grid

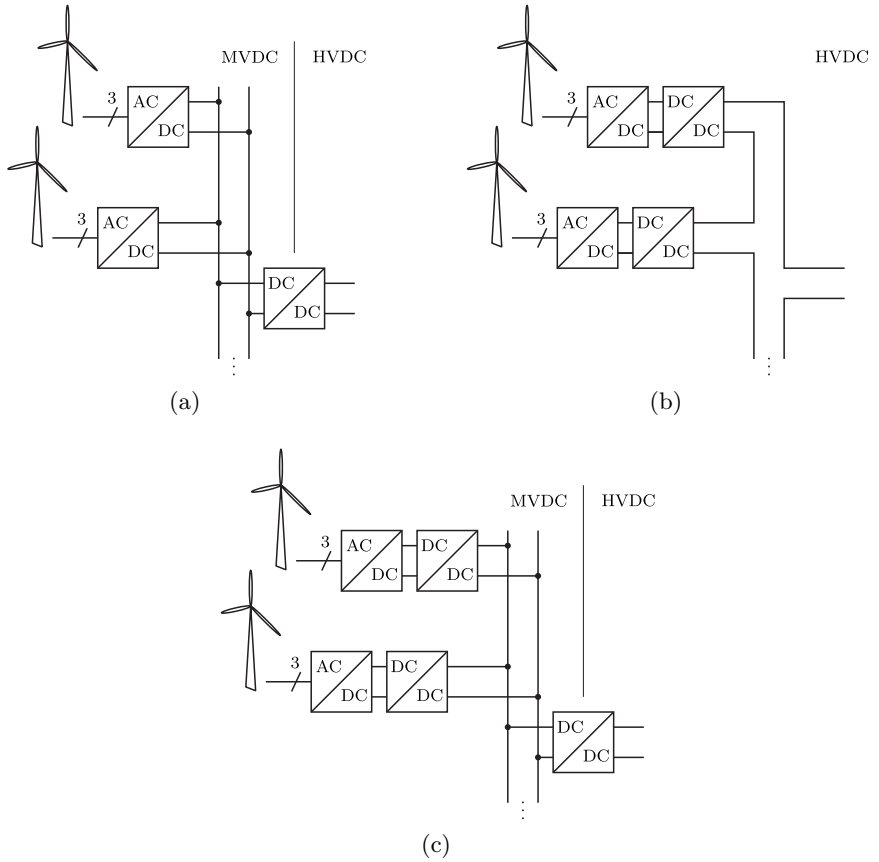


Figure 1.3: Different topologies of a dc collector grid

1.4.1.2 Photovoltaic

To a large extent, photovoltaic (PV) generation is fed in the low-voltage grid. However in Germany, more than 30 % of PV is also injected into the 20 kV MV grid as illustrated in Fig. 1.4.

Figure 1.5 (a) shows a possible layout of these large PV parks. Multiple PV modules are clustered to a subfield. Per subfield one low-voltage inverter generates a 50 Hz ac voltage. Subsequently, the energy from the subfields is collected in an ac grid, stepped up and injected into the MV grid.

In an alternative approach each PV subfield is connected to a common medium-voltage direct current (MVDC) collector grid through a dc-dc converter as shown in Fig. 1.5 (b). From the dc collector grid one central MV inverter feeds the energy into the grid.

Since one large inverter usually operates more efficiently than several smaller ones and since cable losses are decreased, the PV power plant becomes more efficient. In [36] both options are compared for a particular PV layout. Consequently, the European efficiency index rises from 96.3 % to 97.9 %. Furthermore, the dc solution brings even more advantages when the plant is connected to an MVDC or HVDC grid.

Another issue in large PV plants can be the interference between numerous ac filters. The work of Dotter et. al. demonstrates that weak grid conditions and a large number of filters can lead to harmful oscillating currents [37]. This effect still has to be investigated for large dc facilities. However, it can be assumed that the smaller dc filters are less critical concerning mutual interference.

1.4.1.3 Transmission and Distribution Grid

To cope with the increasing share of distributed and volatile generation in the ac grid, solid-state transformers (SST) are an effective solution to control the power flows in future ac grids.

The patent to achieve voltage transformation through high-frequency using power electronics has been filed by McMurray in 1970. Back then, he called these devices “electric transformers” [38]. The principle of the SST today remains the same. It achieves the ac voltage transformation through a high-frequency ac-link using

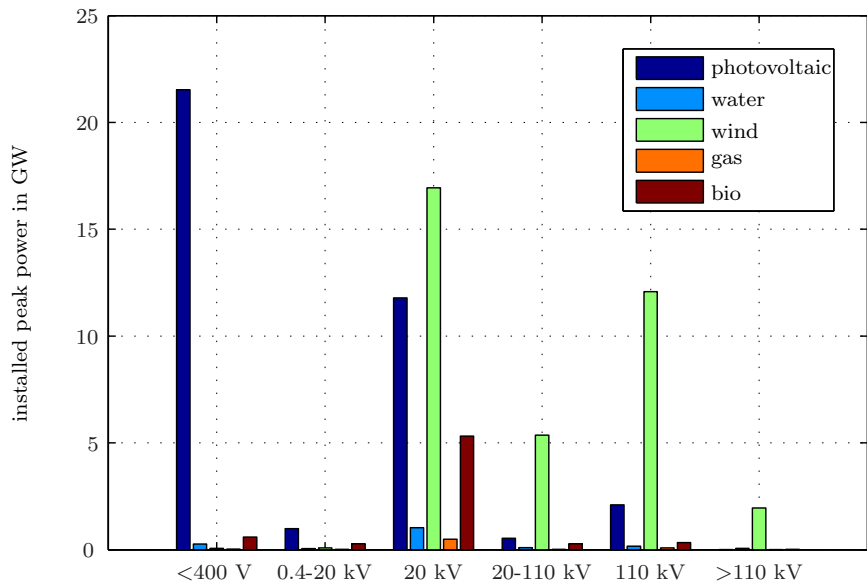
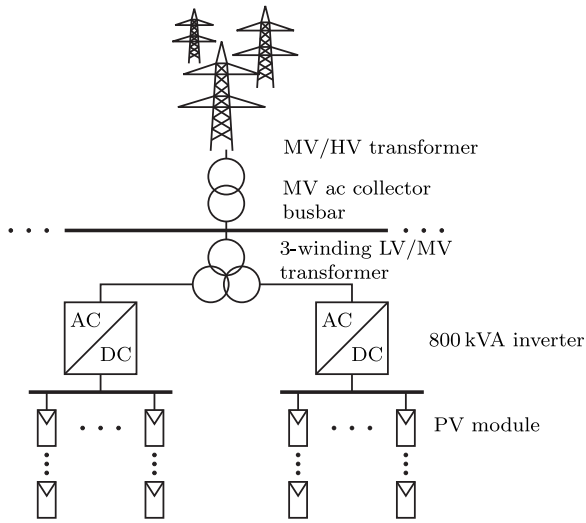
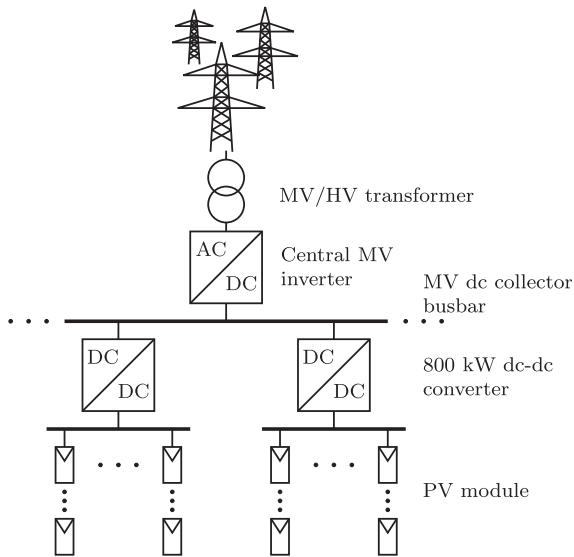


Figure 1.4: Share of the voltage levels on the integration of renewables in Germany in November 2014 (data source: [35])



(a) AC collector grid



(b) DC collector grid

Figure 1.5: Collector grid topologies for a PV application

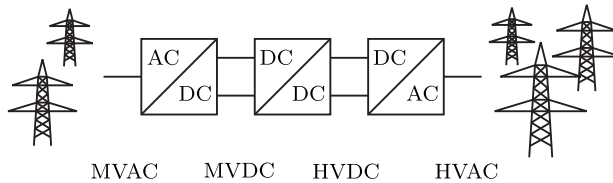


Figure 1.6: Structure of a solid-state ac transformer

power electronics. Therefore, the size of the transformer can be reduced. Moreover, the SST provides power-flow control, voltage-sag compensation and fault-current limitation [39].

Fig. 1.6 exemplarily shows the SST for a connection between medium-voltage alternating current (MVAC) and HVAC. A direct conversion from ac to ac without distinct dc-dc converter stage is another approach. However, this requires special reverse-blocking semiconductor devices [40].

Converting the distribution grid to dc inherently offers the features of the SST [41, 42]. An MVDC infrastructure enhances the stability of the grid under the influence of volatile energy generation [43]. Moreover, renewable energy sources are more efficiently integrated into a dc grid since renewables are mostly dc or variable ac sources [44]. Finally, the increasing urbanization makes the higher power capability of dc systems a decisive feature for densely populated areas [45].

Similarly, the transmission grid benefits from the advantages of a dc infrastructure. In a scenario with a large number of HVDC-operated offshore wind farms, it seems attractive to connect these to a common dc grid [46]. Similarly, projects like “Desertec” integrate solar power from desert regions with the help of dc “superhighways” [47, 48]. Moreover, in new initiatives like Europe’s “Supergrid” the system’s dc voltage is increased further [49, 50].

1.4.1.4 Electric Ship

The electric ship has become a popular application for MVDC. Recently, the Institute of Electrical and Electronics Engineers (IEEE) published a standard for the recommended practice for 1-35 kV dc systems on ships [51]. Besides the already mentioned advantages of dc, some aspects are especially interesting for naval applications. Using dc, the system can be easily separated into several sub-systems that have different priorities and safety measures. Through active power

routing, the supply of critical and vital loads is ensured in case of emergency. The dc system is able to recover faster since generators and motors might not require synchronization. The use and the efficient integration of high-speed generators and motors and resulting volume savings are especially interesting in naval application. Furthermore, redundancy and emergency power supply, like batteries, can be established easily throughout the dc system [52].

Moreover, there are some analogies between the power grid of an electric ship and the electricity grid: different voltage buses, relatively high power ratings and wide proportions (making fault detection and isolation challenging). This promotes the transfer of technology and findings from the electric ship to grid applications [53–56].

1.4.1.5 Railway

Particularly in German-speaking countries, where railways are operated at a frequency of 16.7 Hz, the electrification of trains and locomotives with dc enable great material and weight savings. Due to the low frequency, 16.7 Hz transformers and machines, that need to be carried by the locomotives, are especially heavy and bulky.

Through a power-conversion stage at medium-frequency, the size of passive components can be reduced strongly [57–59]. Exemplarily, in [59] the power conversion at 1.75 kHz is compared to a conventional drive train. Through the medium-frequency conversion the power density increases from 0.2–0.35 kVA/kg to 0.5–0.75 kVA/kg and the efficiency is increased by 2–4 % on average.

1.4.2 State-of-the-Art

As stated in the previous chapter, dc makes power distribution more efficient and cheaper. Furthermore, the advantages are utilizable in many different MV applications. One of the key enablers for dc distribution systems are highly efficient dc-dc converters. They are required to adapt voltage levels, control power flows and / or to achieve galvanic isolation. The dc-dc converter can be considered therefore as the “smart” equivalent of a transformer in an ac system. Some topologies are even capable to directly limit fault currents. The following chapter presents the current research foci regarding these converters.

Table 1.1 presents current research on MV dc-dc converters published since 2012⁽ⁱ⁾. While some dc-dc converters are investigated based on simulation solely, others are actually realized as prototypes. Often larger power ratings are achieved by a connection of multiple smaller converters in series or in parallel. Then, the ratings of the smallest converter unit are given to allow comparability. For better visualization the converters from Table 1.1 are mapped in Fig. 1.7.

As also discussed in the previous chapter, there is clearly a demand for mega-watt dc-dc converters. However, single dc-dc-converter prototypes are at the moment limited to 250 kW (cf. Fig. 1.7). Higher power ratings are realized subsequently by interconnection of multiple converter modules. Further advancements could be achieved through new materials, semiconductor devices as well as new topologies.

Regarding material development, especially wide band-gap materials, like silicon carbide (SiC), shall be highlighted. Semi-conductor devices using these materials allow dc-link voltage up to 15 kV while avoiding a series-connection of devices [60]. However, the resulting common-mode currents, electro-magnetic interference (EMI) and the impact on the insulation materials (cf. chapter 4.1) needs to be investigated further.

Regarding converter topologies, many different ones are under research at the moment. A few are depicted in Fig. 1.8 and Fig. 1.9. These topologies mainly differ in their power-flow direction (uni- or bidirectional), soft-switching capability and galvanic isolation. Especially, (inspired by the modular multilevel converter (MMC) [61–63]) the modular dc-dc converters depicted in Fig. 1.9 allow very high voltage levels by a massive series connection of sub-modules [64]. Each sub-module is equipped with an individual dc-link capacitor. By controlling the capacitor voltage, proper voltage sharing among the switches is ensured. These topologies are especially interesting for HVDC applications. Moreover, it should be noted that galvanically isolated dc-dc converters can be connected in series to achieve very high voltage levels [34].

This work focuses on the three-phase dual-active bridge (DAB3) [65] since it offers bidirectional power flow, galvanic isolation and soft-switching operation over a large operating range. Within the scope of this work the DAB3 is investigated for the use in high-power MV applications. This is demonstrated through the construction of a full scale MV dc-dc converter.

⁽ⁱ⁾see [2] for data before 2012

Table 1.1: Overview of published research works on MV dc-dc converters

Power in MW	HV side in kV	LV side in kV	Frequency in kHz	Topology	Isolated	Realized	Reference
0.001	0.225	0.075	1	Kenzelmann-DAB1	☒	☒	[64]
0.002	3	0.2	10	DAB3	☒	☒	[66]
0.005	1.5	0.1	10	DAB1	☒	☒	[67]
0.0076	0.95	0.19	1.4	Jovic	☐	☒	[68]
0.01	6.1	0.4	10	DAB1	☒	☒	[69]
0.024	10.2	0.6	7.5	Resonant switched capacitor	☐	☒	[70]
0.1	22	0.8	10	DAB3	☒	☐	[66]
0.15	6.5	3.3	1.75	LLC	☒	☒	[59]
0.2	2.5	2.5	5	DAB1	☒	☒	[71]
0.25	6.5	3.3	3.6	LLC	☒	☒	[72]
0.25	3.16	0.38	1	M2DC	☐	☐	[73]
0.36	25.2	1.2	1	Kenzelmann-DAB1	☒	☐	[64]
10	13.2	1.65	1.35	Kenzelmann-DAB1	☒	☐	[74]
10	30	5	1.2	Kenzelmann-DAB1	☒	☐	[75]
5	5	5	1	DAB3	☒	☐	[2]

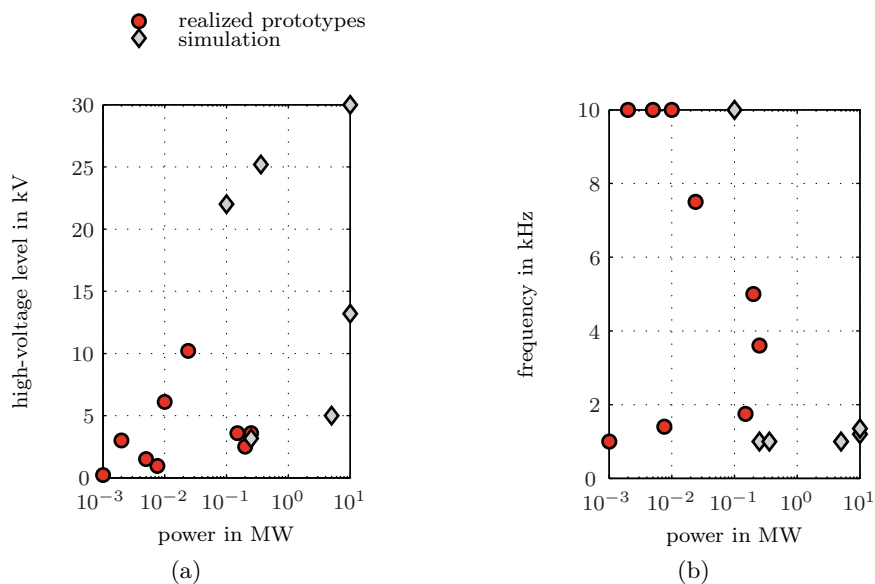
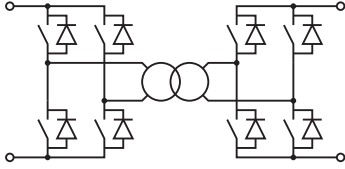
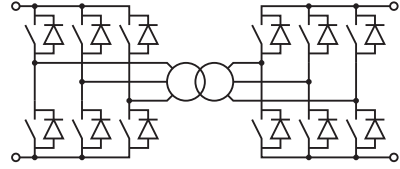


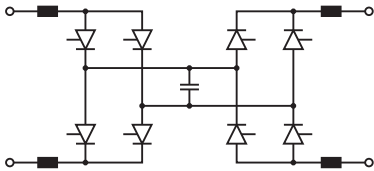
Figure 1.7: Recent research activities on dc-dc converters



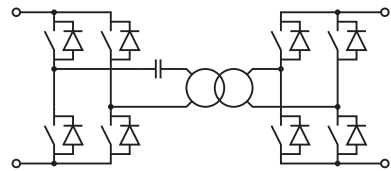
(a) Single-phase dual-active bridge (DAB1) converter



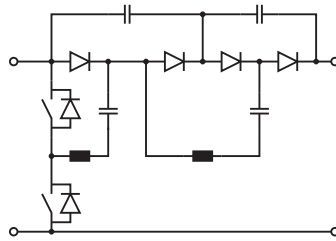
(b) DAB3 converter [65]



(c) DC-DC converter according to Jovcic [76]



(d) LLC converter



(e) Resonant switched capacitor converter

Figure 1.8: Different dc-dc-converter topologies

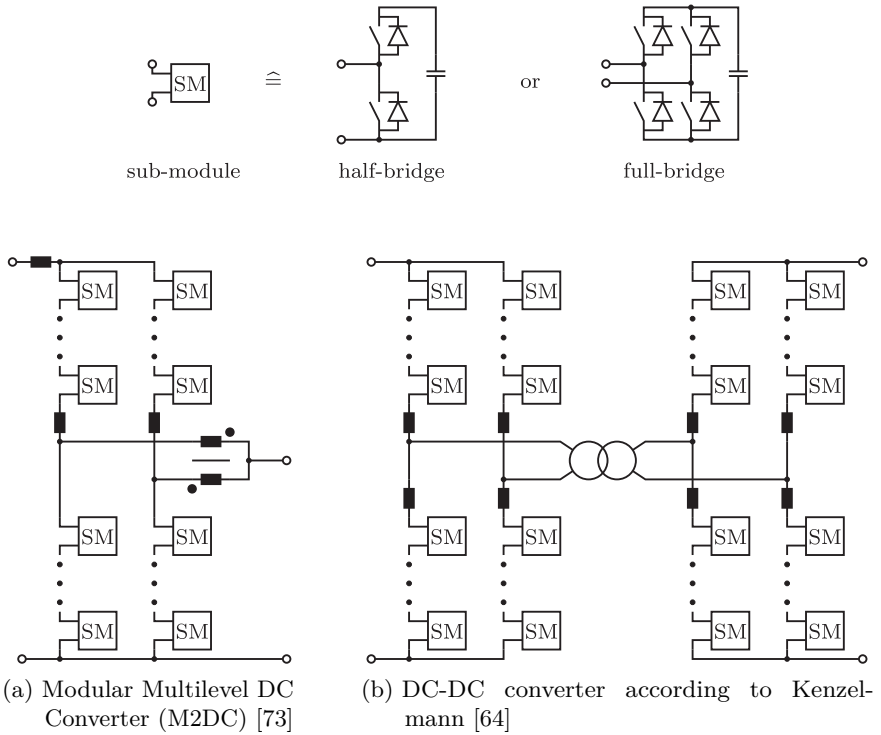


Figure 1.9: Modular dc-dc converters

2 Control of the Three-Phase Dual-Active Bridge Converter

This chapter describes the operating principle of a DAB3. Subsequently, modeling principles as well as control structures are discussed. A modulation strategy is introduced, which allows the setting up of an arbitrary output current within one third of a switching period. This chapter closes with a control scheme to compensate the effect of asymmetric transformers in the ac link of the dc-dc converter.

2.1 Operation Principle

As shown in Fig. 2.1 the DAB3 consists of two three-phase full bridge converters. The one is referred as primary while the other as secondary-side bridge. Both bridges are connected via a transformer. This transformer provides voltage transformation and galvanic isolation, one of the main features of the DAB3. A static voltage conversion ratio can be implemented easily through the transformer's winding ratio

$$n = \frac{N_p}{N_s} \quad (2.1)$$

with N_p and N_s being the number of turns on the primary and secondary respectively. Moreover, the galvanic isolation allows the connection of multiple converters in series to step up the voltage. Therefore, unity turns ration ($n = 1$) is regarded in the following.

Both bridges are operated in fundamental-frequency modulation. Consequently, a six-step voltage waveform is applied to the transformer as depicted in Fig. 2.2. To transfer power, both voltages are phase shifted by the power-electronic bridges. Analogies to a synchronous machine are obvious, where the angle between stator and rotor is determined by the load. Correspondingly, the phase shift in a DAB3 is referred to as load angle φ in the following. According to [65], the power

transferred by the DAB3 is

$$p_s = \frac{u_p^2}{\omega_s L_s} d \varphi \left(\frac{2}{3} - \frac{\varphi}{2\pi} \right) \quad \text{for} \quad 0 \leq \varphi \leq \frac{\pi}{3} \quad (2.2)$$

$$p_s = \frac{u_p^2}{\omega_s L_s} d \left(\varphi - \frac{\varphi^2}{\pi} - \frac{\pi}{18} \right) \quad \text{for} \quad \frac{\pi}{3} < \varphi \leq \frac{2\pi}{3}. \quad (2.3)$$

With $f_s = \omega_s/2\pi$ being the switching frequency, L_s being the series inductance in the ac link and $d = u_s/u_p$ being the dynamic voltage conversion ratio. The series inductance L_s is a major design parameter of the DAB3. It is either a dedicated inductance connected in series to the transformer, or the stray inductance of the transformer used as an internal filter element. This is discussed further in chapter 4.

Figure 2.3 shows the output characteristics of the DAB3. Depending on the dynamic voltage conversion ratio, the diagram gives the output power as a function of the load angle according to (2.2) and (2.3). In the load-angle range from 60-90° the power increases only slightly. However, the apparent power in the transformer in relation to the transferred active power increases disproportionately high [2]. Consequently, the operating range above 60° is considered rather to provide temporary overload.

From the symmetrical structure of the converter it is evident that it is capable of bidirectional power flow. For negative load angles the power flow reverses.

Apart from the static voltage conversion ratio by the transformer or the mentioned series connection, the converter allows a dynamic voltage conversion ratio d . To a large extent, this ratio decides whether the converter operates in hard or soft-switching operation. Therefore, the operation is particularly efficient when the dynamic voltage variation is small like in grid applications. This issue is also discussed later and a circuit is introduced to maintain an efficient operation.

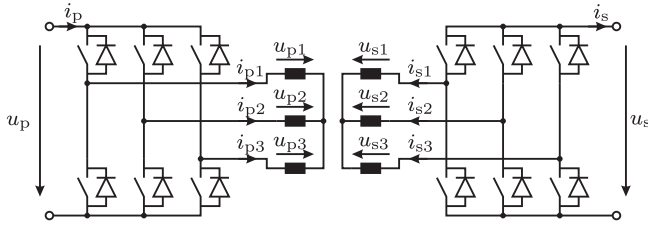


Figure 2.1: Schematic of the DAB3

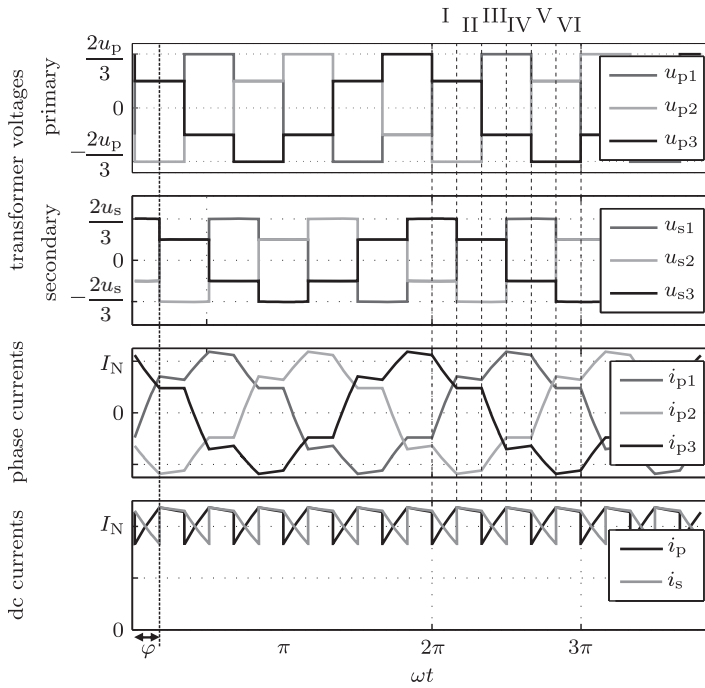


Figure 2.2: Characteristic waveforms in a DAB3

2.2 Different Operating Areas

As stated before, the DAB3 is soft-switched in a large operating area. Consequently, the turn-on of the active switches and the turn-off of the diodes are under zero-voltage conditions. Hence, the reverse recovery losses of the diode and the consequential extra losses in the active switch are negligible. In these soft-switching operating points, the active switches can be equipped with lossless du/dt snubbers to minimize the turn-off losses as well. This topic is discussed separately in chapter 3.2 in more detail.

Figure 2.3 demonstrates that hard switching occurs either in the primary or in the secondary bridge. Especially, hard switching occurs at low-load conditions for large dynamic voltage variations. This makes the DAB3 especially interesting for grid applications, where voltages vary only slightly in the range of $\pm 5\text{-}10\%$.

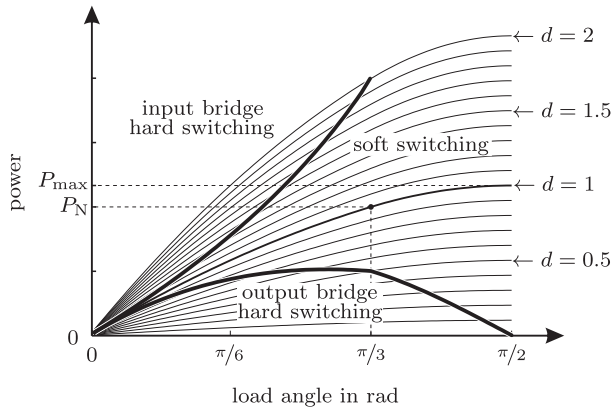


Figure 2.3: Boundaries of the hard and soft-switched operating areas depending on the load angle, the transferred power and the dynamic voltage conversion ratio $d = u_s/u_p$

2.3 Modeling and Dynamic Behavior

The modeling of the converter supports the understanding of its dynamic behavior as basis for the control design. Two different approaches are presented based on state-space averaging (SSA) and first-harmonic approximation (FHA). Finally,

the resulting models are compared with a detailed circuit simulation. The two modeling approaches are developed in the scope of this work and have been presented first in [77, 78].

2.3.1 State-Space and State-Variable Averaging

In the first modeling approach two techniques, SSA and state-variable averaging (SVA), are utilized. Already in 1976, SSA has been used to analyze the dynamics of switched converters [79, 80]. It has been applied to the DAB1 in [81–83]. Later in the modeling approach, SVA is used to reduce the complexity of the model. It has been applied to other three-phase dc-dc converters in [84–86]. In the scope of this work, it has been applied to the DAB3 for the first time.

As indicated in Fig. 2.2, six states (I–VI) over a half switching period can be identified. Within each state the voltages applied to the transformer windings are constant. These six states are used for the SSA approach. The order of the system and with it its complexity is determined by the number of state variables. Since inductor currents and capacitor voltages are used as state variables, the system order is determined by the total number of these devices. Neglecting the transformer's main magnetizing inductance and taking into account the three series inductances and the secondary dc capacitor, the system depicted in Fig. 2.4 is a fourth-order system. The matrix inversion, as part of the SSA, is computationally intensive for such a system.

Consequently, SVA is used to reduce the order of the system. In a DAB, the currents in the series inductances are fast changing compared to the capacitor voltage. Consequently, the fast changing currents are substituted with their short-term average in each state. Hence, the system order reduces to one.

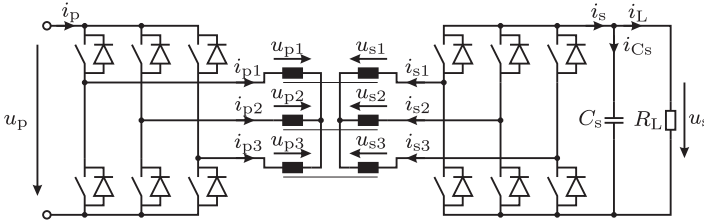


Figure 2.4: Equivalent circuit diagram for the modeling approaches

The averaged state-space representation of the system is

$$\dot{\underline{x}} = A \cdot \underline{x} + B \cdot \underline{u} \quad (2.4)$$

$$\underline{y} = C \cdot \underline{x} + D \cdot \underline{u} \quad (2.5)$$

with

$$\underline{x} = u_s, \quad \underline{y} = u_s, \quad \underline{u} = U_p, \quad C = 1, \quad D = 0. \quad (2.6)$$

In the following, the system matrix A and the input matrix B are derived for $\varphi \geq 0$. A similar calculation for $\varphi < 0$ yields the same matrices.

The node equation of the dc capacitor is given by

$$\frac{du_s}{dt} = \frac{1}{C_s}(i_s - i_L) = \frac{i_s}{C_s} - \frac{u_s}{C_s R_L}. \quad (2.7)$$

According to the SVA, for each state $M \in \{I, \dots, VI\}$, the output current i_s is substituted by the average output current

$$\overline{i_{sM}} = \sum_{x=1}^3 (S_{sxM} \overline{i_{pxM}}). \quad (2.8)$$

S_{sxM} gives the value of the switching function in state M of the secondary bridge of phase $x \in \{1, 2, 3\}$. For $S_{sxM} = 1$ and $S_{sxM} = 0$ the upper and lower switch is conducting, respectively. Since the magnetizing inductance is neglected, the absolute value of the primary and secondary-side transformer currents are equal ($i_{pxM} = -i_{sxM}$).

The averaged phase currents are calculated assuming constant dc voltages in each state and steady-state operation. Consequently, the phase currents at the beginning of mode I equal the negative phase currents at the end of mode VI:

$$\begin{aligned} i_{px}(0) &= -i_{px}(\pi) = - \left(i_{px}(0) + \sum_{M=I}^{VI} \Delta I_{pxM} \right) \\ &= -\frac{1}{2} \sum_{M=I}^{VI} \Delta I_{pxM}, \end{aligned} \quad (2.9)$$

Table 2.1: Transformer voltages and secondary switching function for states M according to Fig. 2.2

M	I	II	III	IV	V	VI
u_{p1M}	$1/3$	$1/3$	$2/3$	$2/3$	$1/3$	$1/3$
u_{s1M}	$-1/3$	$1/3$	$1/3$	$2/3$	$2/3$	$1/3$
S_{s1M}	0	1	1	1	1	1
u_{p2M}	$-2/3$	$-2/3$	$-1/3$	$-1/3$	$1/3$	$1/3$
u_{s2M}	$-1/3$	$-2/3$	$-2/3$	$-1/3$	$-1/3$	$1/3$
S_{s1M}	0	0	0	0	0	1
u_{p3M}	$1/3$	$1/3$	$-1/3$	$-1/3$	$-2/3$	$-2/3$
u_{s3M}	$2/3$	$1/3$	$1/3$	$-1/3$	$-1/3$	$-2/3$
S_{s3M}	1	1	1	0	0	0
$\omega \Delta t_M$	φ	$\pi/3 - \varphi$	φ	$\pi/3 - \varphi$	φ	$\pi/3 - \varphi$

where

$$\Delta I_{p x M} = \frac{1}{L_s} (u_{p x M} - u_{s x M}) \cdot \Delta t_M \quad (2.10)$$

is the current change in phase x during state M . $u_{p x M}$ and $u_{s x M}$ are the primary and secondary transformer voltages of phase x in state M , respectively. Δt_M is the duration of the state M .

Due to the linear current slopes, the average phase current $\overline{i_{p x M}}$ in a state is given by the mean value of the initial current and final current in state M .

$$\overline{i_{p x M}} = i_{p x}(0) + \sum_{m=1}^{M-1} (\Delta I_{p x m}) + \frac{1}{2} \Delta I_{p x M} \quad (2.11)$$

The averaged secondary dc currents for state I and II can be calculated using (2.8)

and the values in Table 2.1:

$$\begin{aligned}\overline{i_{sI}} &= \overline{i_{p3I}} \\ &= \frac{1}{9\omega L_s} \left((3\varphi - \pi) U_s + \left(\frac{3\varphi}{2} + \pi \right) U_p \right),\end{aligned}\quad (2.12)$$

$$\begin{aligned}\overline{i_{sII}} &= \overline{i_{p1II}} + \overline{i_{p3II}} \\ &= \frac{1}{9\omega L_s} (3\varphi U_s + 3\varphi U_p).\end{aligned}\quad (2.13)$$

Due to symmetry, the averaged secondary currents of the even and odd states are equal. Consequently, also the system matrices of the even states (A_e , B_e) and of the odd states (A_o , B_o) are equal, respectively. Therefore, the equations (2.12), (2.13) and (2.7) yield

$$A_o = \frac{-\frac{9\omega L_s}{R_L} + 3\varphi - \pi}{9\omega L_s C_s}, \quad B_o = \frac{\frac{3\varphi}{2} + \pi}{9\omega L_s C_s}, \quad (2.14)$$

$$A_e = \frac{-\frac{9\omega L_s}{R_L} + 3\varphi}{9\omega L_s C_s}, \quad B_e = \frac{3\varphi}{9\omega L_s C_s}. \quad (2.15)$$

As part of the SSA, the matrices are weighted with the duration of the according mode. As also the duration Δt_M of the even and the odd modes is equal, the averaged system and input matrix are

$$A = \frac{3}{\pi} (A_o \cdot \varphi + A_e \cdot (\pi/3 - \varphi)) = \frac{1}{R_L C_s}, \quad (2.16)$$

$$B = \frac{3}{\pi} (B_o \cdot \varphi + B_e \cdot (\pi/3 - \varphi)) = \frac{\varphi \left(\frac{2}{3} - \frac{\varphi}{2\pi} \right)}{\omega L_s C_s}. \quad (2.17)$$

By perturbation of the input and output around a certain operating point, the small-signal transfer functions are determined. The effect on the output is then linearized to derive the transfer functions for the DAB3. This practice in general is described in [79]. The steady-state model, which can also be derived from the characteristic equation of the DAB3 [65] is given by

$$\frac{U_s}{U_p} = \frac{R_L \cdot \varphi \left(\frac{2}{3} - \frac{\varphi}{2\pi} \right)}{\omega L_s}. \quad (2.18)$$

The small-signal transfer function from the input to the output voltage represents

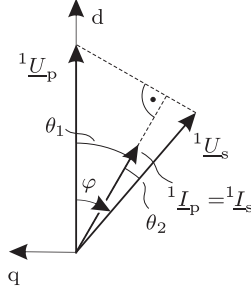


Figure 2.5: Fundamental FHA model of a DAB3

the sensitivity of the output voltage with respect to disturbances of the input voltage:

$$\left. \frac{\tilde{y}}{\tilde{u}} \right|_{\tilde{\varphi}=0} = \frac{\tilde{u}_s}{\tilde{u}_p} = \frac{U_s/U_p}{R_L C_s s + 1}. \quad (2.19)$$

Finally, the small-signal transfer function from the load angle to the output voltage is determined:

$$\left. \frac{\tilde{u}_s}{\tilde{\varphi}} \right|_{\tilde{u}_p=0} = \frac{U_p R_L}{\omega L s} \left(\frac{2}{3} - \frac{\varphi}{\pi} \right) \frac{1}{R_L C_s s + 1}. \quad (2.20)$$

2.3.2 First-Harmonic Approximation

The FHA is a common approach to analyze ac quantities in power-electronic converters. Although these quantities might consist of many harmonics (like the square wave forms in a DAB), the FHA only considers the fundamental oscillation - the first harmonic. Consequently, the system can be described by simple ac-circuit analysis using space vectors and phasors.

The FHA has been applied first to the DAB3 in [65]. It has been used to describe the steady-state behavior of the converter. Contrary, time dependent space vectors instead of stationary phasors are used in this work and originally in [77]. Therefore, the FHA is able to represent the dynamic behavior of the DAB3.

Exemplarily, the fundamental space-vector diagram of a DAB3 is shown in Fig. 2.5. The angle between the primary and secondary transformer voltage corresponds to the load angle φ . The voltage difference ${}^1\underline{U}_p - {}^1\underline{U}_s$ applies across the series inductance L_s of the ac link. Consequently, the currents

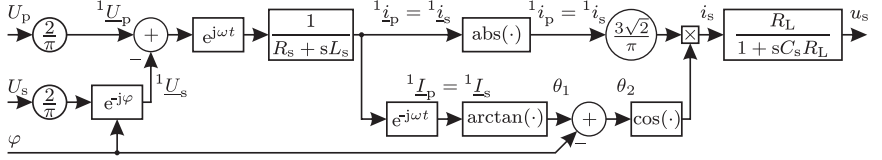


Figure 2.6: Dynamic DAB3 model based on FHA

$$^1I_p = ^1I_s = \frac{^1U_p - ^1U_s}{j\omega L_s} \quad (2.21)$$

are perpendicular to the voltage difference as also indicated in Fig. 2.5.

As also discussed further in chapter 3.2, the DAB3 operates in soft-switching when the current phasor is located between the voltage phasors. Hence, the FHA is also a good tool to quickly estimate a certain operation point.

In the following, the FHA approach is used to model the dynamic behavior of the DAB3. The structure of this model is depicted in Fig. 2.6. Primary and secondary voltage are scaled by $2/\pi$ resulting in the RMS value of the first harmonic. Rotating the secondary voltage phasor by φ and subtracting both voltages give the voltage difference across the transformer. This phasor is transformed to the stationary reference frame to achieve a voltage vector rotating according to the switching frequency. The transformer vector is fed into a transformer model giving the first harmonic of the ac-link current. Finally, the current component which is in phase with the secondary side voltage is calculated and fed into the model of the output filter. The model of the output filter, which is in this example an RC load, results in the dynamic secondary dc voltage.

Since the principle of this dynamic FHA model can be depicted in two figures (Fig. 2.5 and Fig. 2.6) basically, it is very illustrative and especially useful to understand the operation principle of a DAB3. Moreover, due the modularity of this dynamic model, the transformer model or the output filter can be changed easily. This allows a quick comparison of different output filters for example. However, the accuracy of the model is limited as discussed in the following section.

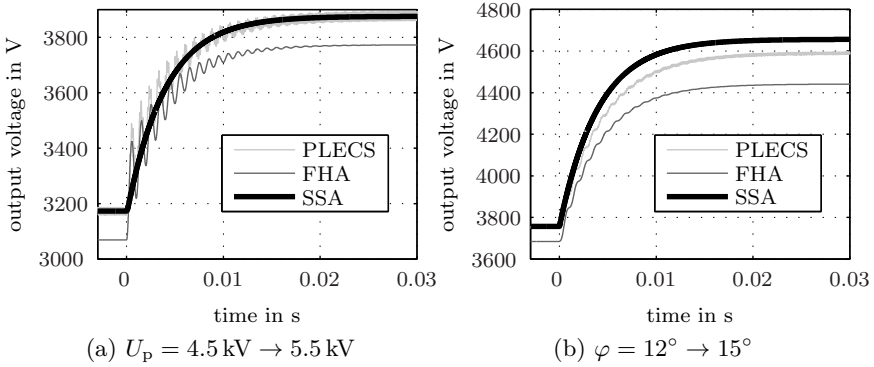


Figure 2.7: Comparison of the models by introducing a step on the primary dc voltage and the load angle respectively.

2.3.3 Verification and Comparison

In the following, the two models are verified and compared in terms of accuracy. As a reference, a detailed model is built using the circuit simulator software Piecewise Linear Electrical Circuit Simulation (PLECS) [87], a toolbox for Matrix Laboratory (MATLAB) [88].

The SSA model is implemented through equations (2.18), (2.19) and (2.20) in MATLAB.

The implementation of the dynamic FHA model is according to Fig. 2.6 using MATLAB and Simulink[®]. The transformer is modeled with its transfer function in the s-domain. While stray inductance and winding resistance are considered in the transfer function, the magnetizing inductance neglected. This ensures also a proper comparability with the SSA model, which neglects the magnetizing inductance as well. According to Fig. 2.4, a dc-link capacitor and a resistor representing a load are connected.

The models are compared by stepping up a certain input variable and observing the system output, which is the secondary dc-link voltage u_s . Figure 2.7 shows the results of the comparison. In (a), a step of the input voltage from 4.5 kV to 5.5 kV is considered, while (b) evaluates a step in the load angle from 12° to 15°.

Considering the step of the input voltage, the SSA model achieves very good

accuracy. Both the dynamic behavior and the steady-state values agree with the detailed PLECS circuit simulation. This is reasonable since the system matrices are independent of the input voltage. Regarding the step of the load angle, the steady-state value of the output voltage is overestimated. The reason lies in the quadratic relationship between load angle and secondary voltage, which is linearized during the process of the SSA.

From Fig. 2.7 it is obvious that the FHA model underestimates the transferred power. This results in a lower output voltage. The FHA neglects the higher harmonics which contribute to the power transfer in a DAB3. Consequently, the dynamic FHA model rather suits an illustrative representation of the converter than a precise modeling of the dynamic power transfer. However, if additional higher harmonics were to be considered, this model could be very well suitable for real-time simulation of dc grids. Since most commercial real-time simulators have their origin in the calculation of ac grids, they can compute a model based on harmonic approximations very efficiently. This is also motivated by the fact that the dynamic behavior of the model matches the circuit simulation very well.

From both models it is evident that the dynamic behavior of the converter is mainly determined by the dc-link capacitors. Since the current in the ac-link is limited by a comparably small series inductance, its dynamic is much higher. This and the observation of the oscillations in the dc currents after an abrupt stepwise change of voltage or load angle have led to the investigations on the instantaneous current control (ICC).

2.4 Instantaneous Current Control

The ICC, which is presented in the following section, allows the control of the output current of a DAB3 within one third of a switching period. It has been presented first in [78, 89, 90]. Based on the ICC, the improved instantaneous current control (I²CC) has been developed, which was presented first in [91, 92].

2.4.1 Applying the Clarke Transformation

Equation (2.22) and (2.23) represent the well-known Clarke transformation [93]. Presented in 1943, the Clarke transformation allows the representation of a three-phase system in the two-dimensional $\alpha\beta$ -plane. In the following, it is applied to the quantities of a DAB3.

$$\underline{i}_p = i_{p\alpha} + j i_{p\beta} = \frac{2}{3} (i_{p1} + \underline{a} i_{p2} + \underline{a}^2 i_{p3}) \quad \text{with} \quad \underline{a} = e^{j120^\circ} \quad (2.22)$$

or

$$\begin{bmatrix} i_{p\alpha} \\ i_{p\beta} \end{bmatrix} = \frac{2}{3} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} i_{p1} \\ i_{p2} \\ i_{p3} \end{bmatrix} \quad (2.23)$$

Applying the Clarke transformation to the ac currents in a DAB3 and plotting the resulting currents $i_{p\alpha}$ and $i_{p\beta}$ in the $\alpha\beta$ -plane result in a hexagon as depicted in Fig. 2.8.

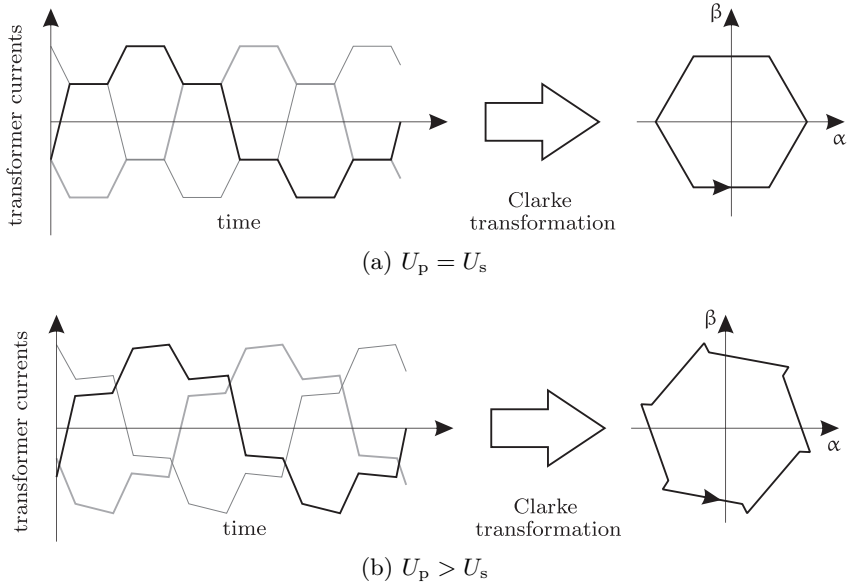


Figure 2.8: Applying the Clarke transformation [93] to the ac currents in a DAB3 results in a hexagon in the $\alpha\beta$ -plane

Figure 2.9 shows the formation of the trajectory within a half switching cycle. In each state (cf. chapter 2.3), the vectors of the transformer voltage are constant in amplitude and phase. The current vector develops along a trajectory which is indicated by a dashed line in the figure. Note that the current vector itself is not shown. It ranges from the point of origin to a point on the trajectory.

Neglecting the winding resistance and the transformer's main inductance, the current vector moves in parallel to the difference of the voltage vectors. In each state, the length of the trajectory is proportional to the duration of the state. Therefore, the main edges of the current trajectory are proportional to the load angle φ . In the following, the trajectory is observed during a load step.

2.4.2 The Instantaneous Current Control (ICC)

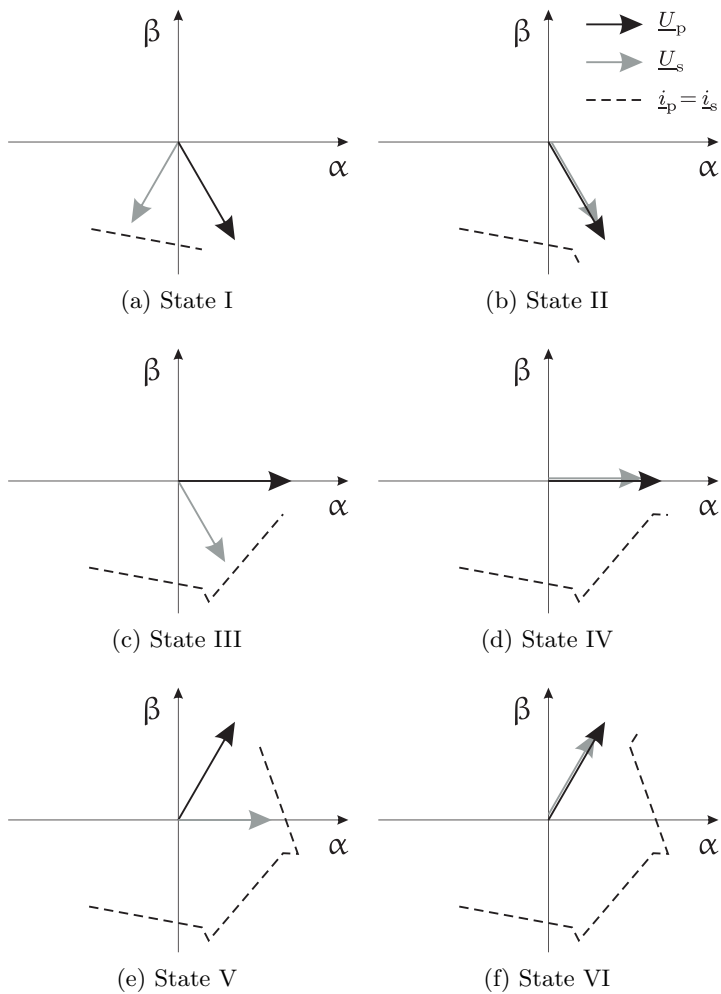
If the load angle is changed abruptly from φ_1 to φ_2 the current trajectory will leave the origin as depicted in Fig. 2.10 (a). This shift results in diverging phase currents and, consequently, in an oscillation of the dc-link currents. After the displacement, the trajectory returns to the center in an exponential manner. The time constant of the exponential decay $\tau = \frac{L_s}{R_s}$ is determined by the series resistance and the inductance of the ac link.

As stated before, the main edges of the trajectory are proportional to the load angle. Consequently, setting the load angles of each converter phase independently, the trajectory can be centered immediately after the transition. This is the principle of the ICC. From infinite combinations to control the phase angles during the transitions, Fig. 2.10 (b) and (c) depict two particular methods. The load angles in the time domain using these methods are shown in Fig. 2.11.

The two-step method is the fastest way to move from one trajectory to another. The transition is completed after one third of a switching period. It may however lead to an overshoot, in particular when the power shall be reversed as shown later. Implementing the two-step method, two intermediate load angles φ_{t1} and φ_{t2} are set during the transition. Regarding a load-angle transition from φ_1 to φ_2 , the intermediate load angles are $\varphi_{t1} = \varphi_2$ and $\varphi_{t2} = \varphi_1$ as depicted in Fig. 2.11. This corresponds to the application of the new load angle at either a rising or falling edge in all three phases.

Using the three-step method, a third intermediate load angle is introduced. For convenience all three transition load angles can be chosen equal $\varphi_{t1} = \varphi_{t2} = \varphi_{t3} = \frac{\varphi_1 + \varphi_2}{2}$. The angle that is being set during the transition is the mean value of the initial and the target load angle.

Figure 2.12 shows a simulation comparing an abrupt change of the load angle and ICC three-step method. The proof of both ICC methods can be found in [78]. It should be noted that the ICC works equally well when the sign of the load angle changes to reverse the power flow.

Figure 2.9: Voltage vectors in the $\alpha\beta$ -plane and generation of the current trajectory

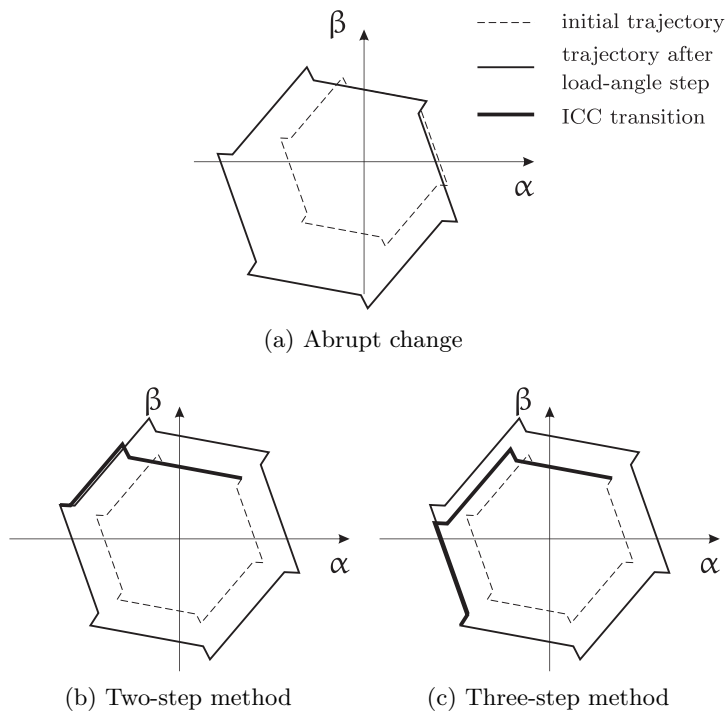
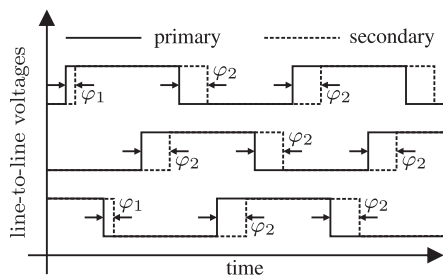
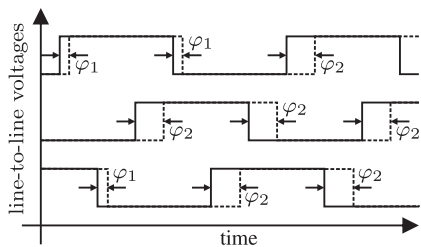


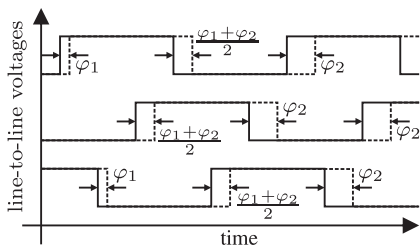
Figure 2.10: Load-angle step without ICC and two different ICC methods



(a) Abrupt change



(b) Two-step method



(c) Three-step method

Figure 2.11: Load angles in the time domain with and without ICC

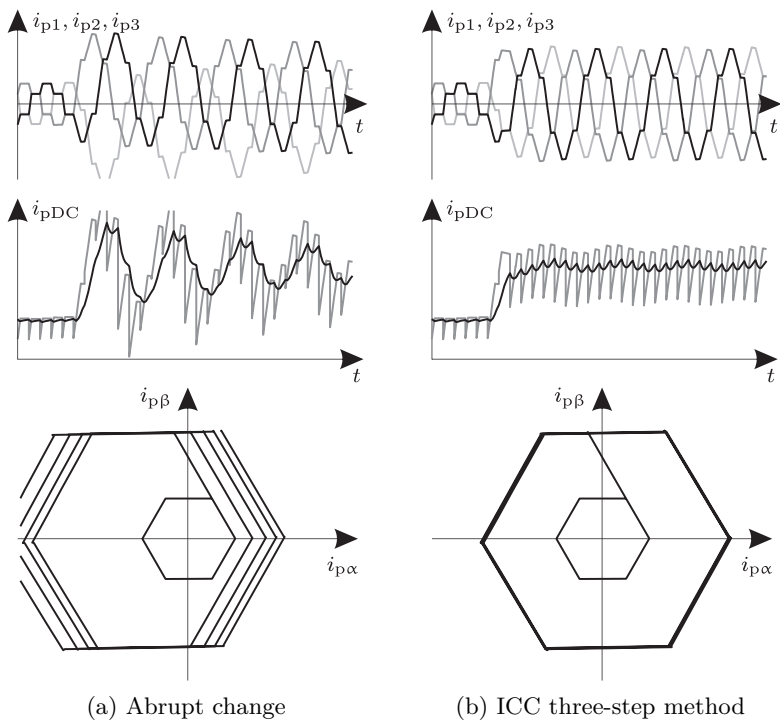


Figure 2.12: Comparison of a load step with and without ICC

2.4.3 The Improved Instantaneous Current Control (I²CC)

The ICC neglects the series resistance in the ac-link. It achieves very good results for highly efficient transformers so that $\tau = L_s/R_s \gg 1/f_s$. In case τf_s becomes smaller, there might still occur unwanted oscillations on the dc currents using the ICC. For these scenarios the I²CC has been developed and presented first in [91, 92].

Similar to the ICC, there are two methods, a two-step and a three-step method, using two and three intermediate transition load angles respectively. However, compared to the ICC, the transition load angles have to be corrected by an additional factor

$$\kappa = e^{-\frac{1}{6f_s\tau}}. \quad (2.24)$$

The equations of the I²CC involve logarithmic and exponential terms which are computational complex regarding the implementation on a digital signal processor (DSP). Therefore, the equations are approximated using the first element of the Taylor series around point $(\varphi_1, \varphi_2) = (0, 0)$.

Considering the two-step method, the transition load angles are:

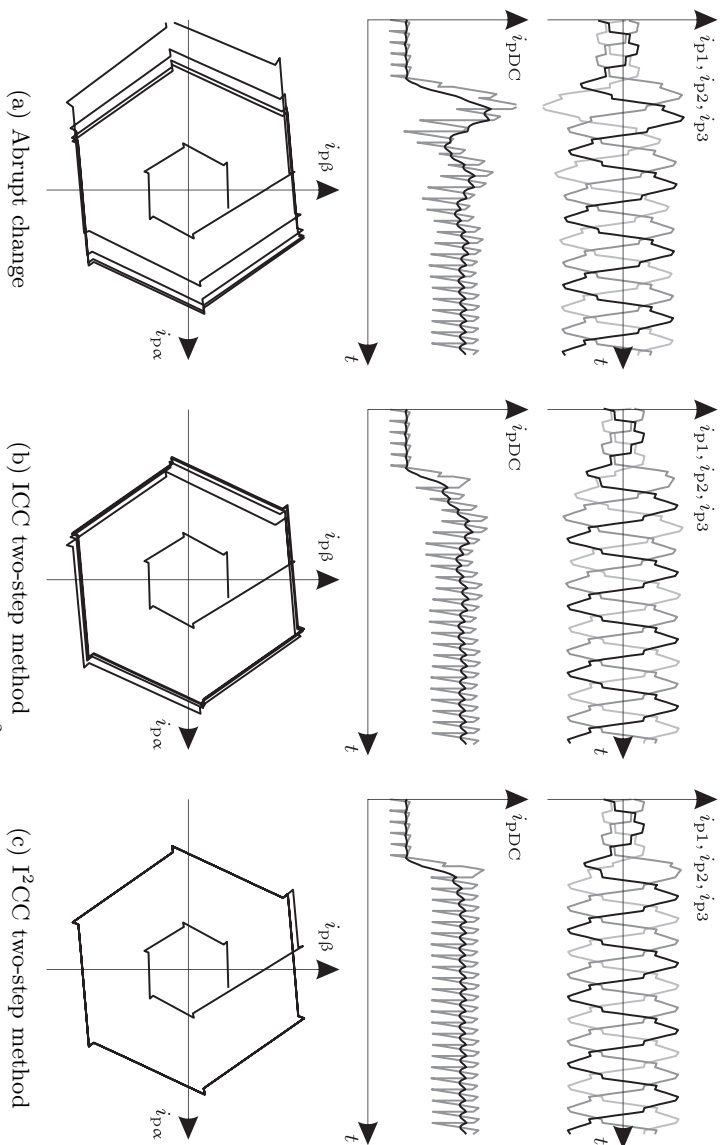
$$\varphi_{t1}(\varphi_1, \varphi_2) = 2\pi f_s \tau \ln \left(\frac{(\kappa^2 - \kappa)e^{\frac{\varphi_1}{2\pi f_s \tau}} + e^{\frac{\varphi_2}{2\pi f_s \tau}}}{\kappa^2 - \kappa + 1} \right) \approx \frac{(\kappa^2 - \kappa)\varphi_1 + \varphi_2}{\kappa^2 - \kappa + 1} \quad (2.25)$$

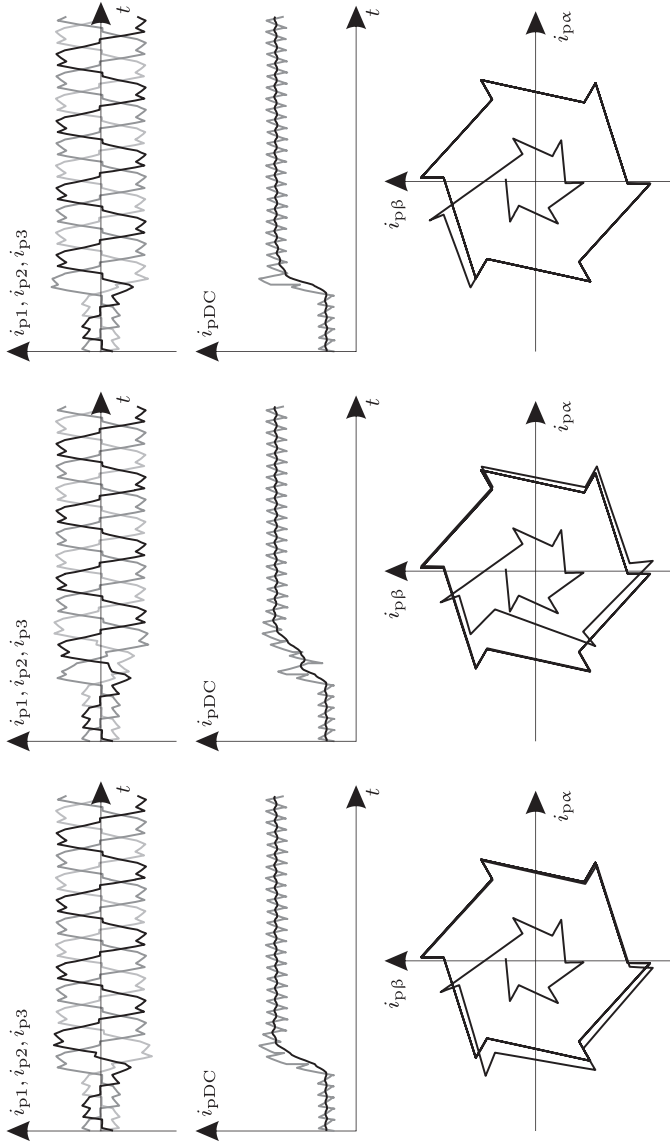
$$\varphi_{t2}(\varphi_1, \varphi_2) = 2\pi f_s \tau \ln \left(\frac{\kappa^2 e^{\frac{\varphi_1}{2\pi f_s \tau}} + (1 - \kappa)e^{\frac{\varphi_2}{2\pi f_s \tau}}}{\kappa^2 - \kappa + 1} \right) \approx \frac{\kappa^2 \varphi_1 + (1 - \kappa)\varphi_2}{\kappa^2 - \kappa + 1}. \quad (2.26)$$

Like for the ICC, the three-step method uses three constant transition load angles:

$$\begin{aligned} \varphi_{t1}(\varphi_1, \varphi_2) &= \varphi_{t2}(\varphi_1, \varphi_2) = \varphi_{t3}(\varphi_1, \varphi_2) \\ &= 2\pi f_s \tau \ln \left(\frac{\kappa^3 e^{\frac{\varphi_1}{2\pi f_s \tau}} + e^{\frac{\varphi_2}{2\pi f_s \tau}}}{\kappa^3 + 1} \right) \end{aligned} \quad (2.27)$$

$$\approx \frac{\kappa^3 \varphi_1 + \varphi_2}{\kappa^3 + 1}. \quad (2.28)$$

Figure 2.13: Comparison between ICG and I²CG with $\tau_{fs} = 1$



(a) Abrupt (b) ICC two-step method (c) I²CC two-step method
Figure 2.14: Comparison between ICC and I²CC with $\tau f_s = 0.25$

2.4.4 Verification

In the following, the ICC and I²CC are demonstrated using a 50 V small-scaled converter. The converter uses insulated-gate bipolar transistors (IGBTs) (Infineon 1KW30N60) that are switched at 10 kHz.

2.4.4.1 Measurements of the ICC

Firstly, the load angle is changed from 30° to 45° .

As evident from Fig. 2.15 (a), the abrupt change of the load angle leads to a shift of the hexagon from the origin. Using the ICC (three-step method), the new load angle can be set fast while avoiding the shift of the hexagon.

Next, the ICC is demonstrated for a reversal of the power flow. The load angle is changed from -30° to 45° .

Changing the load angle in an abrupt manner, shifts the entire hexagon out of the origin leading to intensively diverging phase currents and oscillations on the dc-link currents. As demonstrated in Fig. 2.16 (b), the ICC is very effective even when the power flow has to be reversed.

The remaining transient is caused mainly by a tilting of the hexagon. This results from the forward voltage drop of the power-electronic devices, which is non-linear and different for IGBT and diode. During the power-flow reversal this effect is especially noticeable since for different directions of power flow different devices are mainly conducting the current. Moreover, the ICC neglects the influence of the main inductance and the resistive losses in the transformer.

However, this measurement corresponds to a load jump from -58% to 81% of the nominal power. Regarding this very extreme scenario, the ICC still performs fast and accurately.

Startup and Emergency Stop Analogously, the ICC can be used during the startup and emergency stop. In Fig. 2.17, the system sets a load angle of 45° immediately after the startup. Again the ICC settles instantaneously on the target trajectory. Only a small transient can be observed, which (additionally to the

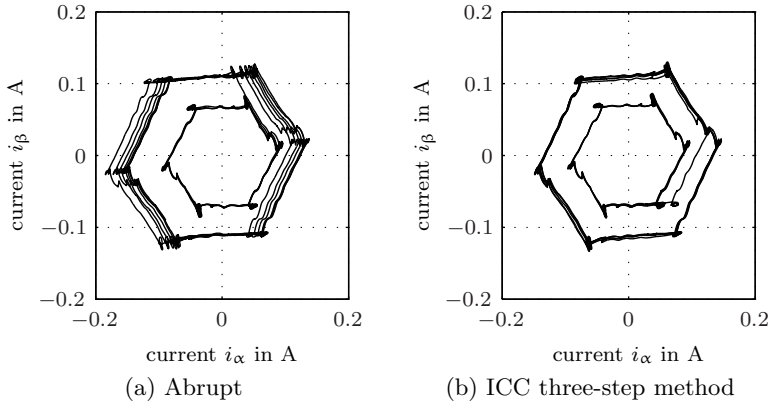


Figure 2.15: Measurement without change of power-flow direction

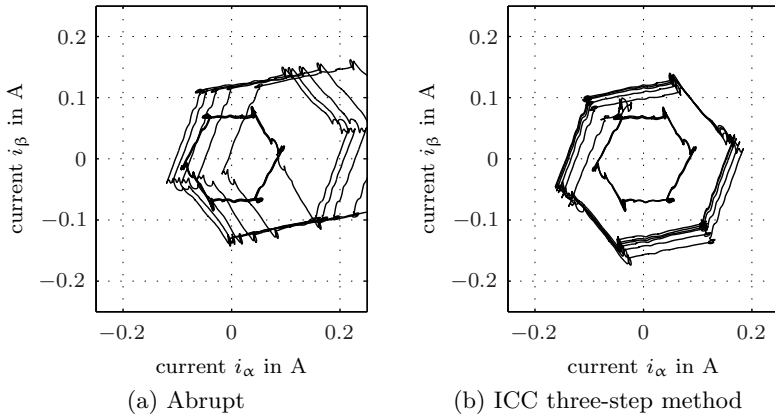


Figure 2.16: Measurement with change of power-flow direction

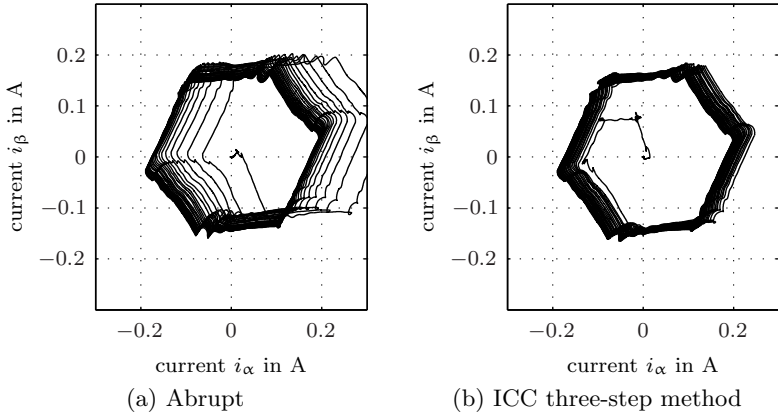


Figure 2.17: Measurement of a startup

reasons named above) results from the main inductance of the transformer which has to be magnetized during startup.

Similarly, the ICC can be used during an emergency stop to demagnetize the series inductances very fast and remove the energy in the system as quickly as possible.

2.4.4.2 Measurements of the I²CC

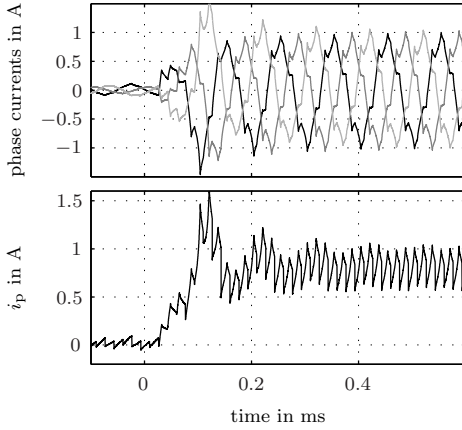
The I²CC is verified as well using the small-scale laboratory prototype. To show the advantages of the I²CC over the ICC, the unfiltered dc current before the dc-link capacitor is calculated from the phase currents $i_{p1} \dots i_{p3}$ and the switching function of the according bridge.

The transformer has a series inductance of $L_s = 250 \mu\text{H}$ and a resistance of $R_s = 0.8 \Omega$ leading to $f_s \tau = 3.12$ at 10 kHz switching frequency. The before mentioned voltage drop of the semiconductor devices, however, leads to an effective $f_s \tau$ of 0.44.

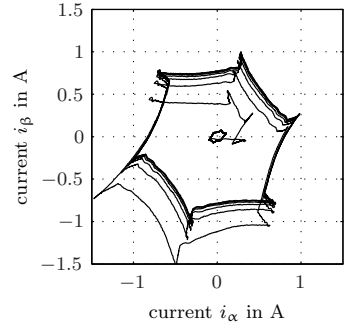
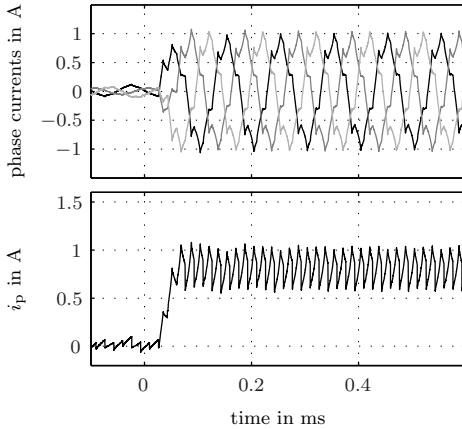
The load-angle step is from 0° to 40° . This leads to the transition load angle for the ICC of $\varphi_{t1} = \varphi_{t2} = \varphi_{t3} = 20^\circ$. Within the experiment the κ is set to achieve a fast transition to the target load angle. Consequently, the transition load angle

of $\varphi_{t1} = \varphi_{t2} = \varphi_{t3} = 31.2^\circ$ is set for the I²CC three-step method.

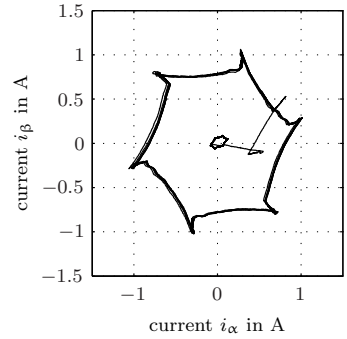
The results of the measurement are depicted in Fig. 2.18.



(a) ICC three-step method - time domain

(b) ICC three-step method - $\alpha\beta$ -plane

(c) I2CC three-step method - time domain

(d) I2CC three-step method - $\alpha\beta$ -planeFigure 2.18: Comparison of ICC and I²CC

2.5 Closed-Loop Control

The ICC and I²CC are very fast open-loop current-control methods. Closed-loop current and voltage controller that utilize the fast dynamics of the two ICCs are discussed in the following. Parts of this work have been presented also in [77].

Since the dc current can be set within one third of a switching period, it can be assumed that the dynamic of the converter is mainly limited by the switching frequency and the dc filter, namely the dc link capacitors. This can be concluded from the modeling approaches in section 2.3 as well.

A straightforward approach to design a closed-loop voltage controller is a cascaded control structure [94]. They consist of an inner and an outer control loop. The current controller, with its higher dynamic, regulates the inner loop, while the voltage is controlled in the outer loop. Cascaded control structures have two main advantages. Disturbances in the inner loop are compensated directly independent from the slower outer loop. Moreover, the cascaded controller consists of two controllers. Since these controllers can be designed subsequently, the design process is more intuitive. On the contrary, the use of two cascaded controllers results in a lower bandwidth of the outer control loop.

Figure 2.19 shows one possible controller implementation. As stated, the current is controlled in the inner loop while the voltage is controlled in the outer. Moreover, a feed forward in the current control is used which consists of the power equation of the DAB3. The share of the load angle contributed by the feed forward (cf. Fig. 2.19) is

$$\Phi = \frac{2\pi}{3} - \sqrt{\left(\frac{2\pi}{3}\right)^2 - \frac{2\pi\omega L_s}{U_p} i_s^*}. \quad (2.29)$$

Since the power equation neglects the transformer's main inductance, the winding resistance and the forward voltage drop of the power semiconductors, a feed-back control should be implemented as well. This can be for example a proportional-integral (PI) controller.

The voltage is controlled in the outer control loop. The gain of the voltage controller, which is a PI controller in the example, is tuned according to the dc-link capacitor and the load.

The proposed feed-back control is a straightforward approach, which is easy to implement and to design. A current limiter can be implemented easily by limiting

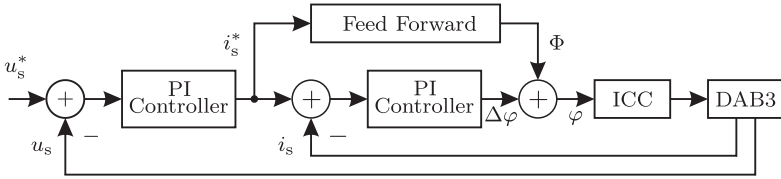


Figure 2.19: Measurement of a startup

the reference current. Note that an anti windup mechanism should be implemented when a control variable, either the current or the load angle, is limited.

For most applications, the proposed control is well suitable. If higher control bandwidth is required or the switching frequency is very low, techniques like disturbance feed forward, dead-time compensation or adaptive control can be implemented [95–97].

2.6 Balancing Control

In the following section, the effect of asymmetric series impedances in the ac-link and the consequences on the converter are discussed. Moreover, a control method is introduced to compensate the effect of asymmetric impedances. Parts of this work have also been published in [78, 98].

2.6.1 Motivation

Often the stray inductance of the transformer L_σ is used as internal filter element. Hence, the stray inductance of the transformer serves as series inductance L_s . In this case, a careless design of the transformer can lead to asymmetric impedances in the ac link.

This is demonstrated in the following. Figure 2.20 (a) gives the winding and core arrangement of a standard three-phase three-leg transformer. Considering this arrangement, the transformer is designed for the use in a 5 kV DAB3 operated at a switching frequency of 1 kHz. Increasing the distance between the primary winding P_i and secondary winding S_i of each phase $i \in \{1, 2, 3\}$, the stray inductance is increased to serve as integrated series inductance. This distance is set for each phase equally.

A finite-element method (FEM) model of this design is created using the simulator Finite Element Method Magnetics (FEMM) [99]. To determine the stray inductance of each phase, the secondary winding is shorted while the primary winding is supplied. Consequently, a magnetic field is generated between primary and secondary winding. The energy stored in the magnetic field corresponds to the energy stored in the stray inductance of the transformer's equivalent circuit. This way the value of the stray inductance can be determined from the FEM simulation. Alternatively, one can calculate the stray inductance from the current and voltage phasors, which leads to the same result.

Figure 2.20 (b) shows the result from the simulation to determine the stray inductance of the first phase. Carrying out this simulation for the other phases as well leads to

$$L_{\sigma 1} = 104.5 \mu\text{H} \quad (2.30)$$

$$L_{\sigma 2} = 112.1 \mu\text{H} \quad (2.31)$$

$$L_{\sigma 3} = 104.5 \mu\text{H}. \quad (2.32)$$

Although the transformer is perfectly symmetrical, the second phase located on the central core has a higher stray inductance. This results solely from the use of a three-legged core. Evident from the simulation in Fig. 2.20 (b), the magnetic field is denser in the winding windows resulting from the yoke above and below the winding. In the outer part of the transformer, the stray field spreads wider. Consequently, the stray inductance of the second phase is larger since the magnetic field increases.

This example illustrates that a very marginal variation in the geometry already has a significant effect on the stray inductance of the transformer. Including manufacturing tolerances, $\pm 5 - 10\%$ variation of the stray inductance between the phases is reasonable.

2.6.2 Impact of Asymmetric Impedances on the Converter

The impact of an asymmetric series impedance in the ac link is investigated through a laboratory prototype. The converter is operated with a primary- and secondary-side dc-link voltage of 60 V at a switching frequency of 10 kHz. The main inductance of the transformer's three phases are $L_{h1} = 2.7$ mH, $L_{h2} = 2.1$ mH and $L_{h3} = 2.4$ mH. Moreover, the stray inductance of each phase is $L_{\sigma 1} = 394 \mu\text{H}$, $L_{\sigma 2} = 233 \mu\text{H}$ and $L_{\sigma 3} = 248 \mu\text{H}$. The load angle in that measurement is $\varphi = 38^\circ$.

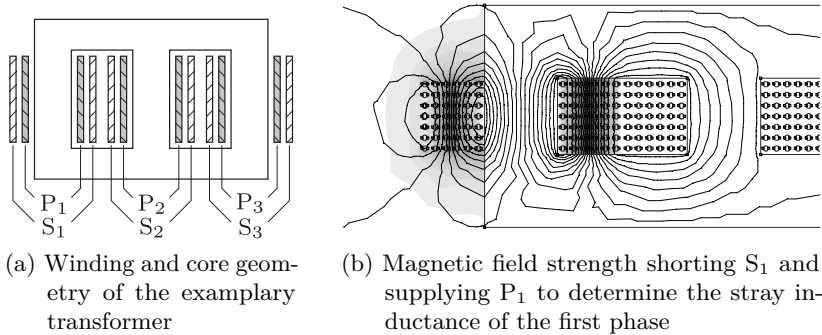
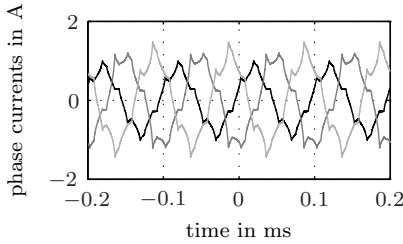


Figure 2.20: Stray-inductance analysis through an FEM simulation

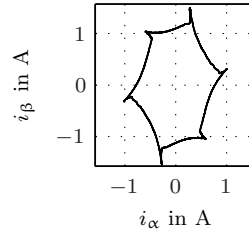
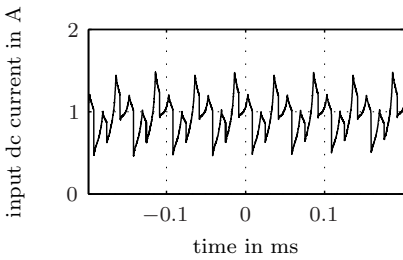
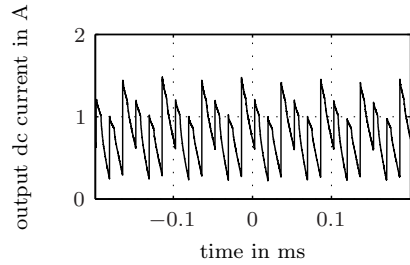
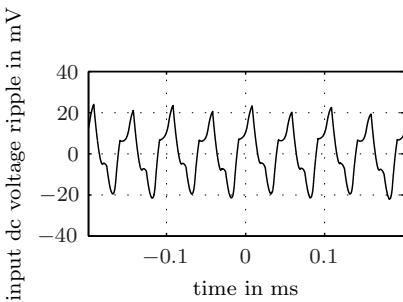
Figure 2.21 shows the result of the measurement. Due to the asymmetric transformer, the phase currents are unequally distributed among the phases. The phase with the smallest series inductance carries most current. If the phase currents are transformed to the $\alpha\beta$ -plane, the resulting hexagon will be deformed as depicted in Fig. 2.21 (b). From the dc input and output current, shown in Fig. 2.21 (c) and (d) respectively, one can see that the current ripple increases. The current is superimposed by a second order harmonic with respect to the switching frequency. The dc input and output voltage ripple, depicted in Fig. 2.21 (e) and (f) respectively, is calculated from the dc current and the assumption the dc-link capacitor being 100 μF . The second-order harmonic can be observed in the dc voltage as well.

Due to the asymmetry in the transformer the phases do not share the current equally. Consequently, the windings of the transformer and the power-electronic devices might show different temperatures for the three phases. Ultimately, the converter would need to be derated according to the devices sharing the highest current. The harmonics on the dc current increase leading to higher losses in the capacitors. Moreover, the voltage ripple on the dc-links is increased unnecessarily.

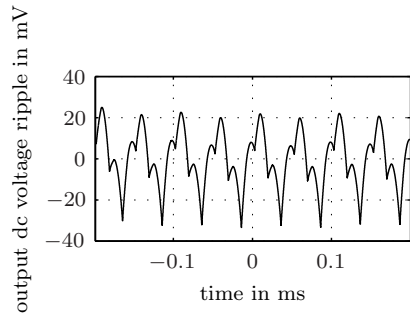
The length of the hexagon's edges is proportional to the load angle of the particular phase. As the hexagon in the $\alpha\beta$ -plane deforms under the influence of the asymmetry (cf. Fig. 2.21 (b)), it is possible to correct it using individual load angles for each phase. This is the main principle of the balancing control.



(a) Phase currents

(b) Phase currents represented in the $\alpha\beta$ -plane(c) Input current i_p (d) Output current i_s 

(e) Input voltage ripple



(f) Output voltage ripple

Figure 2.21: Influence of an asymmetric transformer on a DAB3

2.6.3 Compensation of Asymmetric Impedances

Having asymmetric impedances in the transformer and introducing individual load angles, two issues have to be kept in mind.

Usually the potential of the transformer's star point is either plus or minus $\frac{1}{6}U_p$ with respect to the mid point of the balanced dc-link. For an asymmetric system, however, the voltage is

$$U_{NP,I} = \frac{U_p}{6} + \Delta U_{NP,I} = \frac{U_p}{6} + U_s \frac{\frac{L_{\sigma 1} L_{\sigma 2}}{6} + \frac{L_{\sigma 1} L_{\sigma 3}}{6} + \frac{L_{\sigma 2} L_{\sigma 3}}{3}}{L_{\sigma 1} L_{\sigma 2} + L_{\sigma 1} L_{\sigma 3} + L_{\sigma 2} L_{\sigma 3}}. \quad (2.33)$$

In the following, an impedance variation of $\pm 5\%$ is considered. Consequently, $\Delta U_{NP,I}$ is less than $\pm 1\%$ of the dc-link voltage. For the sake of simplicity, this effect is neglected and phase voltages of $\pm \frac{1}{3}U_{DC}$ and $\pm \frac{2}{3}U_{DC}$ are assumed.

Secondly, one has to note that when different load angles for the three phases are set, 120° phase shift between the phases can not be kept up for the primary and secondary side at the same time. In other words, besides the load angles, two additional degrees of freedom come along: the phase shift of two phases with respect to the reference phase. For the first investigation the phase shift is set symmetrically on the primary and secondary bridge. Consequently, the gating signals on the primary and secondary bridge are shifted by $-\frac{\varphi}{2}$ and $\frac{\varphi}{2}$ respectively.

Equalizing the rms value of the ac currents, for example, ensures a balanced distribution of the winding losses in the transformer. This is verified through a simulation using MATLAB and PLECS with the following parameters:

$$\begin{array}{lll} U_p = 5 \text{ kV} & L_{\sigma 1} = 230 \text{ } \mu\text{H} & R_s = 15 \text{ m}\Omega \\ U_s = 5.5 \text{ kV} & L_{\sigma 2} = 160 \text{ } \mu\text{H} & L_h = 100 \text{ mH} \\ f_s = 1 \text{ kHz} & L_{\sigma 3} = 200 \text{ } \mu\text{H} & \end{array}$$

The simulation results are depicted in Fig. 2.22. Initially, the reference output current is set to 500 A. All phases are operated with the same load angle. As discussed before, the transformer currents are unequally distributed among the phases. At $t_0 = 0$ ms the balancing control is switched on. Consequently, the phase currents converge while the power-controller keeps the total output power stable. At $t_1 = 40$ ms the power controller receives a new reference value of 1000 A.

Note that during the transition, the phase currents remain stable. Finally, at $t_2 = 70$ ms the balancing control is switched off and the transformer currents diverge.

The balancing-control structure is depicted in Fig. 2.23. One dedicated controller achieves the power regulation. In the given example, this controller sets the load angle of the third phase φ_3 . The balancing controllers set the load angles for the first and second phase φ_1 and φ_2 . They are controlled in a way that the RMS value of the three ac currents is equal.

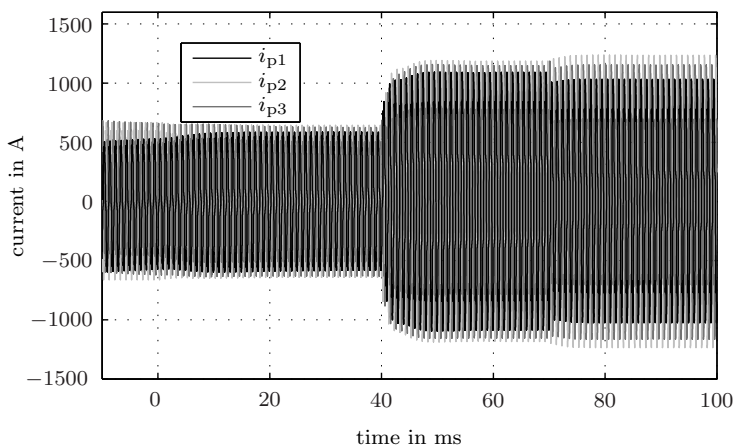
It should be noted that the three phase currents and load angles are strongly coupled by the nature of the DAB3 and the controllers interact with each other. Since the dynamic of the balancing controllers, however, can be very low, it is unlikely that the balancing and power controllers build up oscillations. In fact, very good results are achieved with the balancing controller being pure integral controllers.

2.6.4 Optimization Objectives

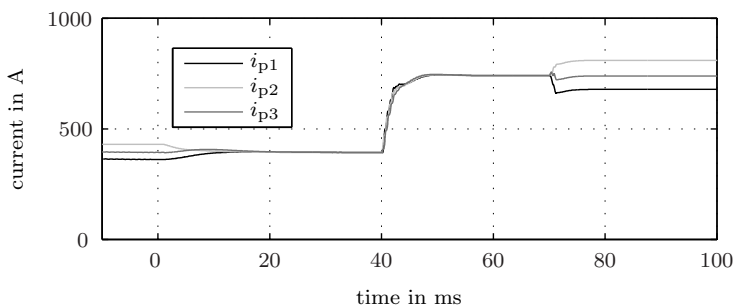
Figure 2.24 gives a close-up view of the simulation in Fig. 2.22. Although the RMS currents are balanced, there is still a second harmonic on the dc currents. Actually, there are multiple optimization goals:

- equal RMS phase currents,
- minimal dc voltage ripple on the primary,
- minimal dc voltage ripple on the secondary,
- minimal capacitor RMS current on the primary,
- minimal capacitor RMS current on the secondary,
- equal turn-off currents (main contributor to the switching losses),
- compensation of the second harmonic on the primary dc current and
- compensation of the second harmonic on the secondary dc current.

It has been shown that, for example, it is not possible to minimize the dc-voltage ripple on the primary and on the secondary at the same time [98]. Rather, a single optimization goal has to be found on the one hand. For example, due to the operating range of a converter the primary dc-capacitor current is always larger than the secondary. Consequently, the primary dc-capacitor current shall be minimized.



(a) Phase currents



(b) RMS currents

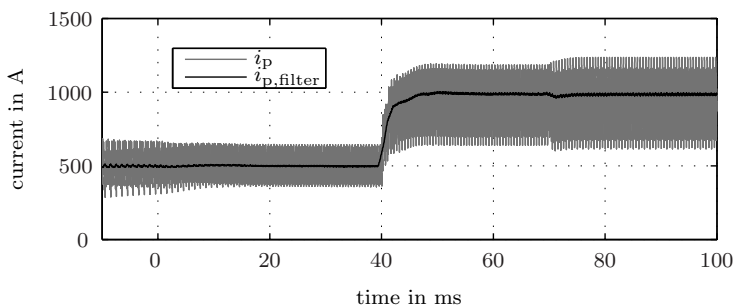
(c) Input current i_p

Figure 2.22: Balancing feed-back control

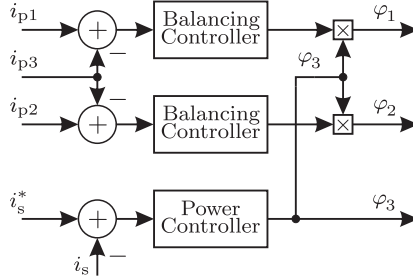


Figure 2.23: Structure of the balancing control

On the other hand, one can optimize the system taking multiple objectives into account. This includes the definition of a cost function which is minimized during the optimization process. The definition of this cost function depends on the specific application of the converter.

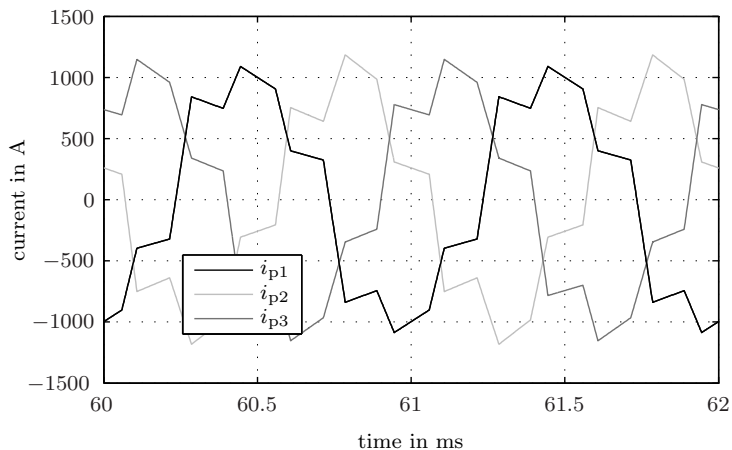
2.6.5 Practical Implementation

Due to the elevated frequency, it is challenging to determine the RMS value accurately. Consequently, an alternative control scheme is investigated beside the feed back of the RMS currents.

To balance the transformer currents, three characteristic current values (one for each phase) are fed back to a current controller. Ideally, these current values can be acquired via a single sampling point for each phase without further post processing. To ensure a robust implementation, these points should not be sampled at a switching action concurrently. Consequently, the midpoints of the linear current trajectories are analyzed for suitability.

Figure 2.25 shows different sampling points that are intended as reference input for the feed-back control. For the different investigated control methods either the currents marked with a cross (i_{+}), a dot ($i_{\bullet}$) or an 'x' ($i_{\times}$) are equalized by the controller.

In the following comparison, the primary and secondary dc-voltage ripples (Δu_p and Δu_s) are used to benchmark and compare the different approaches. A comparison is conducted using $U_p = 5 \text{ kV}$ and ranging U_s from 4.5 kV to 5.5 kV . The stray inductance of the transformer is $L_{\sigma 1} = 230 \mu\text{H}$, $L_{\sigma 2} = 160 \mu\text{H}$, $L_{\sigma 3} =$



(a) Phase currents

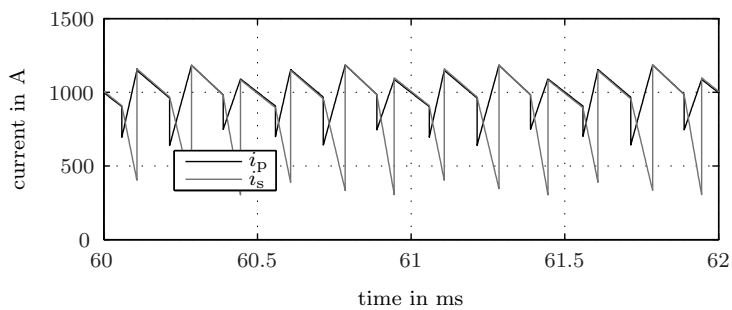
(b) DC currents i_p , i_s

Figure 2.24: Close-up view of the simulation given in Fig. 2.22

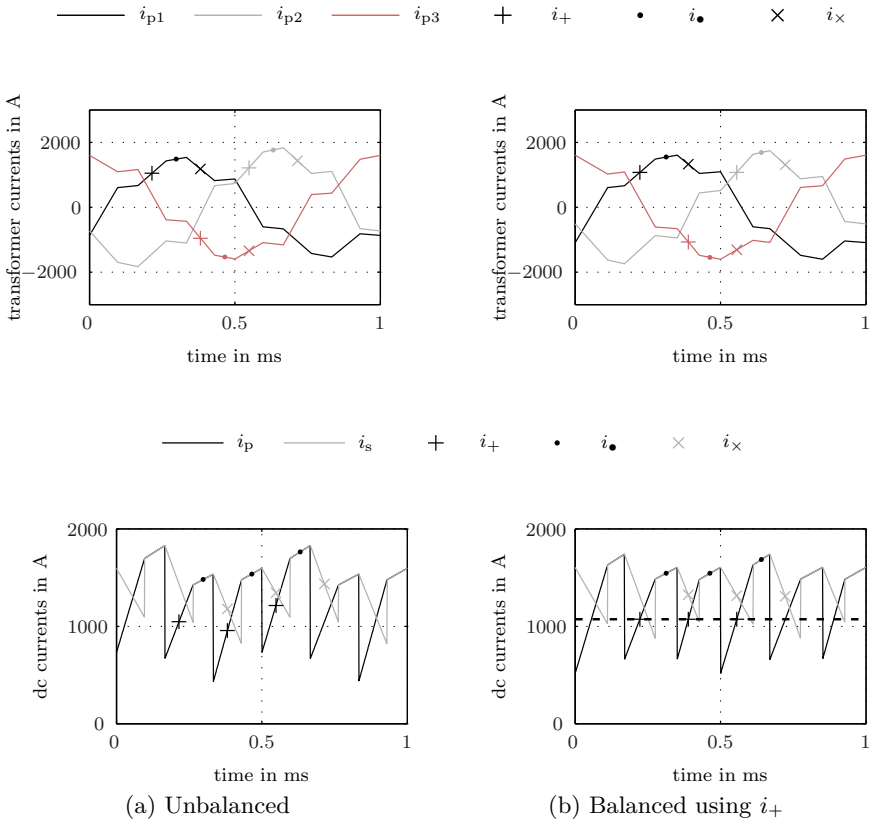


Figure 2.25: Possible sampling points indicated in phase currents and dc currents

200 μH for the asymmetric case and $L_s = 200 \mu\text{H}$ for the symmetric case. The switching frequency is assumed $f_s = 1 \text{ kHz}$ and the dc-link capacitances on the primary and secondary side are 1 mF each. The comparison of the different input variables is shown in Fig. 2.26. The balancing scheme based on controlling the current $i_•$ achieves good results only for low load angles ($\phi < 15^\circ$). Moreover for higher load angles, the voltage ripple on the primary and secondary dc is even higher than the case without balancing control (indicated by "w/o balancing" in the graphic). Equalizing the currents i_+ or i_x leads to much better results. The choice between i_+ and i_x does not influence the balancing significantly. The performance is comparable to a perfectly symmetrical transformer as indicated in the figures as well. Consequently, the negative influence of an asymmetric

transformer can be compensated nearly completely.

It should be noted that especially for low load angles, balancing i_+ might not be practical since switching actions near the sampling point influence the current measurement. However, balancing the currents $i_\bullet$ shows good results for low load angles. This motivates to use $i_\bullet$ for low load angles and switch to i_+ for higher load angles.

Using the same laboratory demonstrator with the same parameters as for the measurement in Fig. 2.21, the balancing control has been implemented. The results are shown in Fig. 2.27. The steady-state load angles are $\varphi_1 = 44.5^\circ$, $\varphi_2 = 33^\circ$ and $\varphi_3 = 31^\circ$. Figure 2.27 demonstrates that the dc-current ripple can be reduced effectively compared to the measurement shown in Fig. 2.21.

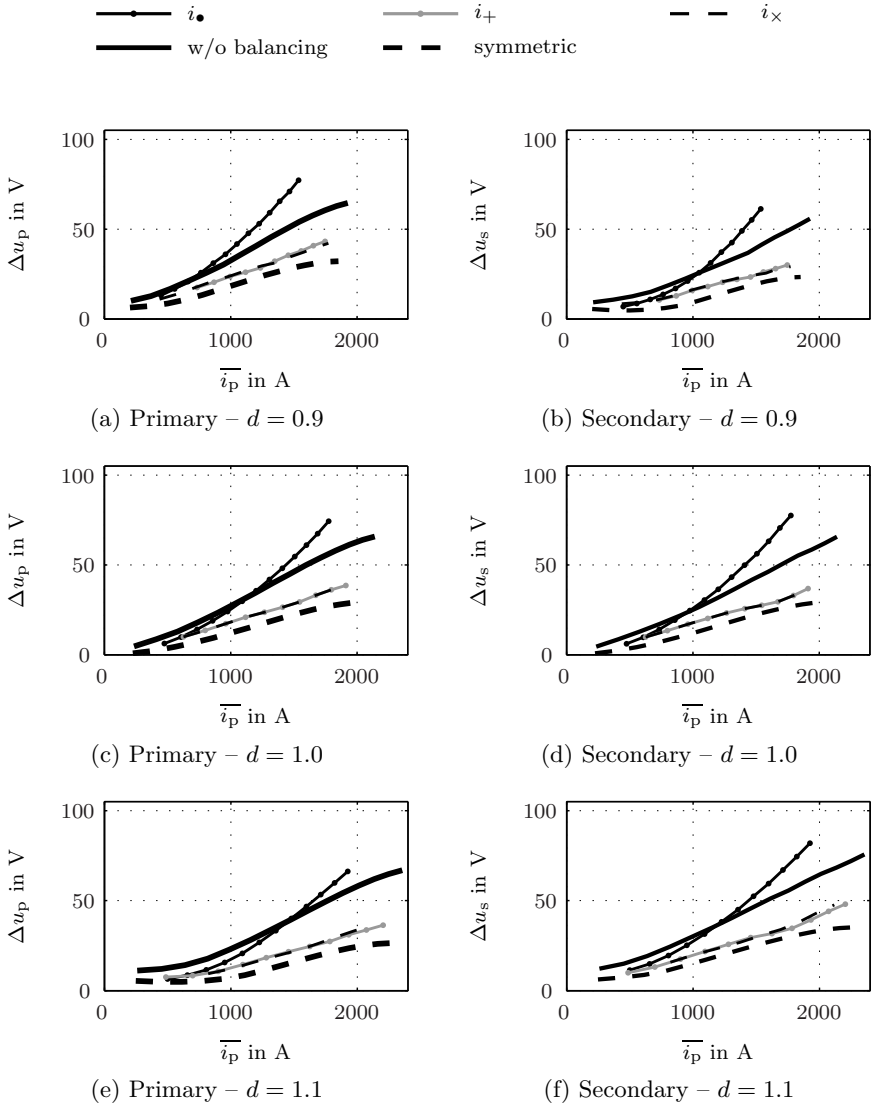
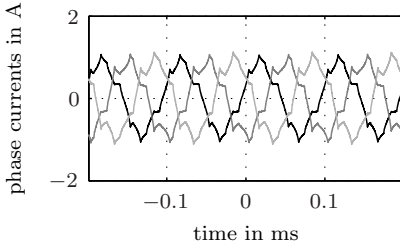
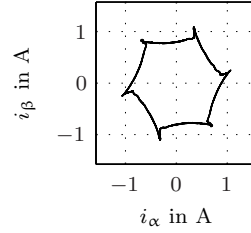
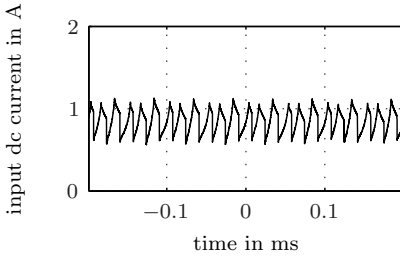
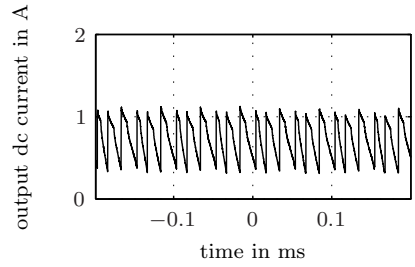
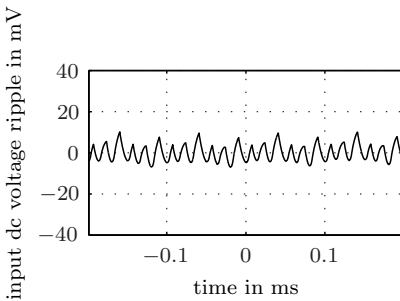


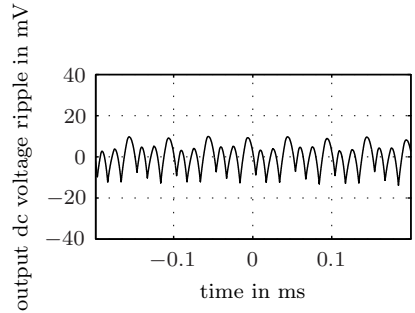
Figure 2.26: Primary and secondary-side dc-voltage ripple for different voltage ratios $d = \frac{U_s}{U_p}$



(a) Phase currents

(b) Phase currents represented in the $\alpha\beta$ -plane(c) Input current i_p (d) Output current i_s 

(e) Input voltage ripple



(f) Output voltage ripple

Figure 2.27: Influence of an asymmetric transformer on a DAB3

3 Power Semiconductors and Soft-Switching Operation

As discussed before, the DAB3 is a soft-switched converter in a wide operating range. Regarding the realization of this dc-dc converter as an MV prototype, the different applicable power semiconductors are discussed in the following. Moreover, an ARCP is designed which allows soft-switching in the entire operating range of the converter.

Parts of this chapter have been published as well in [12, 100].

3.1 Preferable Power-Electronic Devices

The power-electronic switches in a DAB3 require active turn-off capability. Consequently, IGBTs, metal-oxide-semiconductor field-effect transistors (MOSFETs), integrated gate-commutated thyristors (IGCTs) and many other devices are applicable in general. Since the DAB3 is a soft-switched converter, switching losses are heavily reduced by the converter topology. Therefore, one can suggest that devices, which are optimized for a very good conduction performance, are the ideal choice for this converter [2].

This shall be verified in a brief investigation. Through a PLECS simulation the losses of different semiconductors are compared. The simulation parameters are listed in Table 3.1. For the comparison, three IGBT and IGCT devices are chosen rated for 4500 V. A module and a press pack IGBT are compared. Both of them have the highest rating in their class. Regarding the IGCT, the device is chosen which already prevailed for a DAB3 application in [2]. The devices, together with some characteristic values, are given in Table 3.2. The loss data of each device is found in the data sheets of the corresponding manufacturers. Moreover, the data of the IGCT using turn-off snubbers have been measured by Lenke [2].

Figure 3.1 gives the sum of the semiconductor losses in a stacked bar plot. The losses of the active switches and the diodes are included. Since 2.8 kV devices are used, two of them per inverter arm are connected in series. Consequently, the

Table 3.1: Simulation parameters

primary dc voltage	$U_p = 5 \text{ kV}$
voltage conversion ratio	$d = 1$
switching frequency	$f_s = 1 \text{ kHz}$
series inductance	$L_s = 200 \text{ } \mu\text{H}$

Table 3.2: Semiconductor devices

Type	Model	$U_{\max}$	ΔU^a	E_{off}^b
Presspack IGBT	ABB 5SNA 2000K451300	4500 V	3.7 V	10 J
Module IGBT	Infineon FZ1200R45KL3	4500 V	4 V	8.5 J
IGCT	ABB 5SHY 35L4511	4500 V	2.55 V	8.5 J
Diode	Infineon D1031SH ^c	4500 V	3.8 V	
Diode	Internal diode FZ1200R45KL3	4500 V	3.2 V	

^a at 2000 A^b at 2000 A^c Except of the module IGBT, which is using its internal diode

losses are given for 24 IGBTs or IGCTs and 24 diodes. The results are broken down in conduction losses P_{cond} and switching losses P_{sw} .

The performance of the IGCT is superior to both IGBT devices. The lower forward voltage drop and the reduced switching losses decrease the overall power loss. Compared to the module IGBT and the press-pack IGBT, the use of the IGCT reduces the losses by 26 kW and 46 kW, respectively, at a power level of 7 MW.

Moreover, the influence of additional lossless snubbers, which are introduced in chapter 3.2, can be compared. Figure 3.1 (d) and (e) show the semiconductor losses if $1 \text{ } \mu\text{F}$ and $2 \text{ } \mu\text{F}$ are connected in parallel to each IGCT. Consequently, $1 \text{ } \mu\text{F}$ reduces the switching losses significantly. A further increase of the snubber capacitance to $2 \text{ } \mu\text{F}$, however, reduces the switching losses only slightly. This “saturation effect” is also described in [101]. Moreover, the turn-off losses under quasi zero-voltage conditions are strongly affected by the carrier lifetime of the bipolar devices [102] and their internal structure [103]. This motivates the special design of bipolar devices, both IGBT and IGCT, for the use in soft-switched converters.

Regarding the relative losses in relation to the transferred power, the IGCT allows an efficiency increase of roughly 0.7 percentage points compared to the press-pack IGBT. If the IGCT is equipped with lossless snubbers, additional 0.5 percentage points can be gained.

Usually, the performance increase has to be judged against the additional features of the IGBT as e.g.

- the short-circuit detection and turn-off capability,
- the simplified series connection of devices,
- the simpler gate drive unit,
- the absence of a clamping circuit in hard-switched applications [104],
- several different manufacturers and
- possibility for in-house gate driver design.

However, the fact that many of these features are not required in soft-switched converters motivates the use of IGCTs in the DAB3 even more. This and the assurance of the soft-switching operation in the entire operating range is discussed in the following chapter.

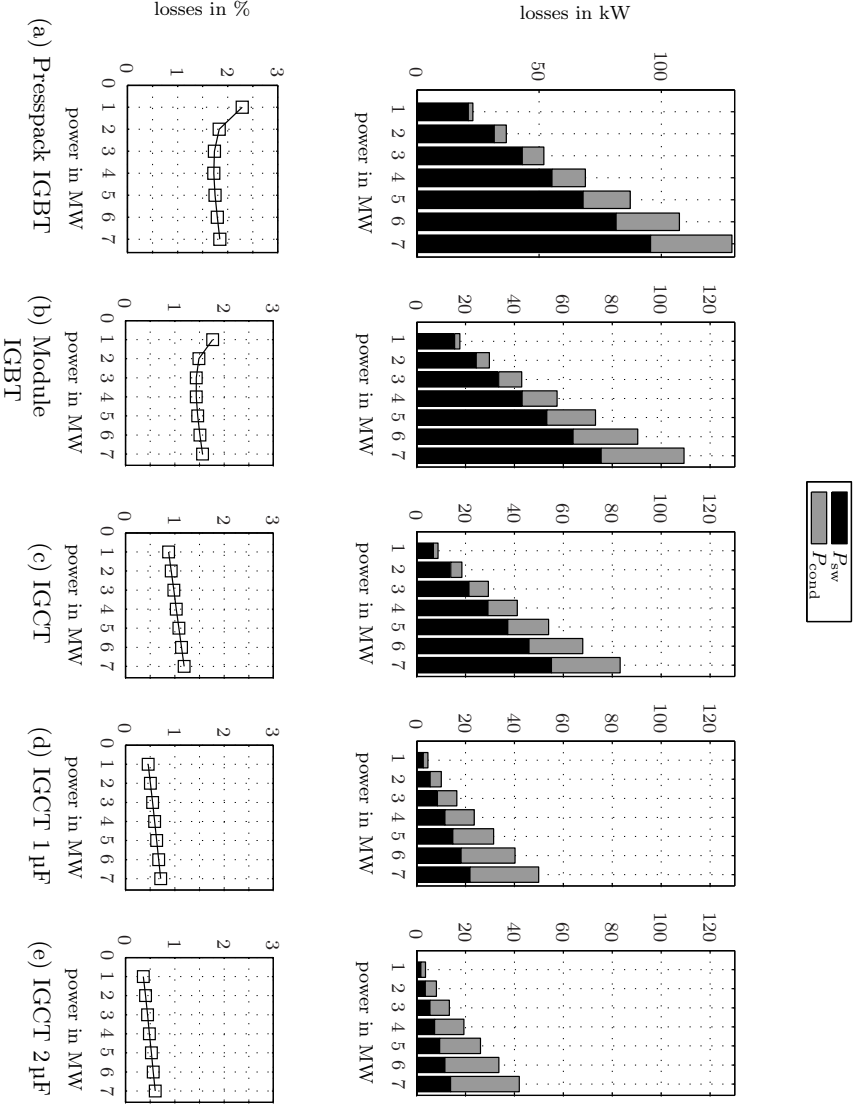


Figure 3.1: Losses of different power-electronic devices in a DAB3

3.2 Ensuring Soft-Switching Operation Using the Auxiliary Resonant-Commutated Pole (ARCP)

In general, snubber circuits are used to reduce the switching losses in power-electronic devices by limiting the current or voltage transients during device turn-on or turn-off respectively. An RC snubber for example is depicted in Fig. 3.2 (a). The capacitor C_{sn} limits the voltage slope. Hence, virtually zero-voltage turn-off is achieved. The resistor R_{sn} limits the inrush current, occurring when the device turns and shorts the charged capacitor C_{sn} . Although the losses in the semi-conductor switch are reduced, the total losses increase due to the snubber loss [105]. Consequently, it is rather a shift of the generated losses from the semi-conductor device to the snubber, which allows a simpler cooling.

Since the DAB3 provides soft-switching in a wide operating range, each power-electronic device turns on when virtually zero voltage is applied across it. Consequently, the resistor R_{sn} (used to limit the inrush current during turn-on) is not required. Hence, the snubber reduces the turn-off losses while no additional snubber losses occur. Therefore, this principle is referred to as lossless snubber in the following.

Applying the lossless snubber in the DAB3, soft-switching operation has to be ensured throughout all operating points. In the following, an ARCP is analyzed and designed for the use in a particular dc-dc converter prototype of 5 kV and 5 MW. Parts of this work have been presented in [100].

Since 1989, the ARCP is used to maintain soft-switching in power-electronic converters [106–108]. Mainly, it has been applied to inverters. However, it has also been used in three-level converters [109], multi-level flying capacitor converters [110] as well as matrix converters [111]. Moreover, in [2] first investigations have been made to apply the ARCP to a DAB3.

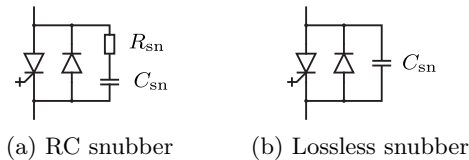


Figure 3.2: Turn-off snubbers

3.2.1 ARCP Operation

A single phase leg of an ARCP is depicted in Fig. 3.3 (a). Figure 3.3 (b) shows an exemplary commutation from the lower main switch $S_{\text{main}-}$ to the upper main switch $S_{\text{main}+}$.

The operation principle of the ARCP according to [107] is introduced briefly:

- Phase I** To initiate the commutation S_{aux} is turned on. Consequently, the current i_{aux} rises linearly according to the dc-link voltage and the inductance L_{aux} .
- Phase II** When the current i_{aux} equals the load current i_{p1} , the second phase begins. Now the load current is carried by the ARCP branch completely. Within phase II additional boost current is injected into the ARCP branch. The boost current provides further energy for the commutation to compensate losses.
- Phase III** The lower main switch $S_{\text{main}-}$ is turned off. According to the resonance of auxiliary inductance L_{aux} and the snubber capacitances $C_{\text{sn}+}$ and $C_{\text{sn}-}$, the snubber capacitors reload. When the antiparallel diode of the upper main switch $S_{\text{main}+}$ becomes forward biased, $S_{\text{main}+}$ is able to turn on at zero-voltage conditions.
- Phase IV** The commutation is completed when the inductance L_{aux} is demagnetized.

It should be noted that there are different operating points for the ARCP. In the given example, the load current has the improper sign for the commutation from the lower to the upper voltage rail. Therefore, the ARCP takes over the full load current in phase I. This ARCP operation mode is referred to as “mode 2” in the following and corresponds to the hard-switched operating point of the DAB3. In “mode 1” the sign of the load current is suitable to reload the snubber. However, the energy provided by the load current is insufficient to commutate the voltage within a given time frame. “Mode 1” can be split into two further modes. In “mode 1a” the current is not high enough to commutate timely within the dead time. Yet, it provides enough energy to compensate the losses during the commutation. Therefore, no additional boost time is needed and S_{aux} is turned on at the same time the according main switch turns off. In “mode 1b”, the load current is even too low to compensate the losses. Consequently, the ARCP has to provide extra boost energy. The different modes in relation to the operation boundaries are depicted in Fig. 3.4. Comparing to Fig. 2.3, the operating range of “mode 2” is identical to the hard switched operating points of the DAB3 without ARCP.

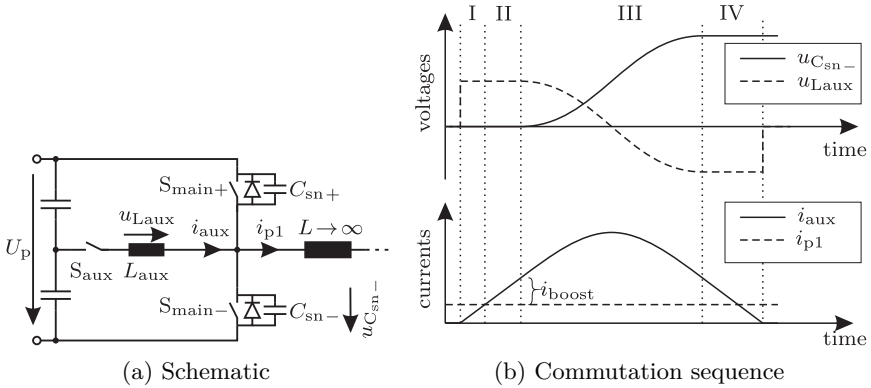


Figure 3.3: Single phase leg of an ARCP

The integration of the ARCP-concept into the DAB3 for the primary side is shown in Fig. 3.5. The additional switch S_{aux} and the inductance L_{aux} are connected per phase. Especially in a DAB3, they are considered as auxiliary devices since they are rated for a small fraction of the total converter power. This results from the fact, that the DAB3 is inherently soft-switched. If the DAB3 operates in a formerly hard-switched operating point, the ARCP will need to provide small reactive current only. This is also shown later, when an ARCP for a medium-voltage dc-dc converter is designed and constructed. Compared to the use in an inverter application, this is one main advantage of the ARCP use in a DAB3.

3.2.2 Design Considerations

In this chapter, the ARCP is designed for a 5 MW DAB3 operated at a nominal dc-link voltage of 5 kV considering a dynamic voltage variation of $\pm 10\%$. The dc-dc converter is operated at 1 kHz.

The auxiliary switch S_{aux} has to be bidirectional in terms of voltage-blocking capability and current conductivity. Moreover, it has to be rated for the half dc-link voltage. The RMS current in the ARCP branch is fairly low compared to the power rating of the converter. The exact current rating, however, depends on the chosen inductance L_{aux} . The switching frequency of S_{aux} equals the switching frequency of the dc-dc converter. Consequently, being S_{aux} a thyristor, a sufficient recovery time has to be fulfilled.

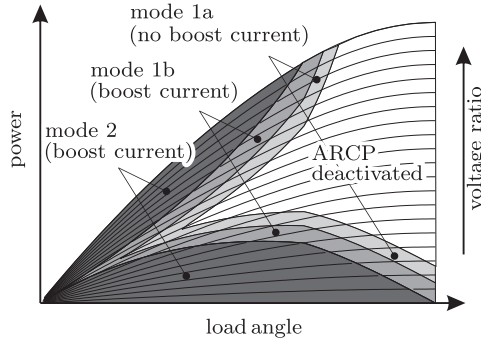


Figure 3.4: Operating modes of the ARCP in a DAB3 application

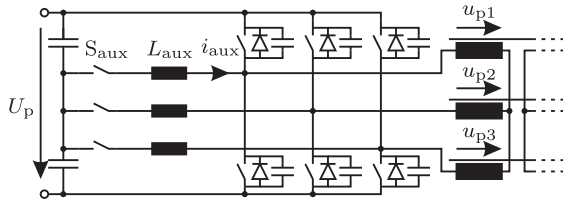


Figure 3.5: Integration of the ARCP in one bridge of the DAB3

Also the passive components, the snubber capacitance C_{sn} and the auxiliary inductor L_{aux} , need to be designed. These components influence the turn-off losses of the main switches, the operating range with deactivated ARCP, the peak auxiliary current, the RMS auxiliary current, the commutation time, the reverse-recovery losses of S_{aux} , the voltage transients across the auxiliary switch, the voltage transients imposed to the medium-frequency transformer, the auxiliary-inductor size and the inductor losses. Designing the ARCP, a good trade-off needs to be found. From an efficiency point of view, the auxiliary inductance should be as large as possible while still allowing commutation within a specified dead time. Hence, the peak and RMS auxiliary current are minimal. Yet, this leads to a heavier and more costly inductor.

During the design of the ARCP, the suitability of different technologies shall be evaluated and compared. This includes different semiconductor devices (thyristor vs. turn-off IGCT), different materials (silicon (Si) vs. SiC) and different topologies. The considered semiconductor devices that are compared are listed in Table 3.3. SiC diodes with the specified ratings are commercially not available. Consequently, a series and parallel connection of the commercial diodes D_2 and D_3 is utilized

Table 3.3: Compared semiconductor devices

Symbol	Model	Material	Rev. Block.	Snubber
T_1	GeneSiC GA060TH65	SiC	no	no
T_2	Westcode R1127	Si	yes	yes
IGCT	Mitsubishi GCT GCU15CA-130	Si	yes	no
D_1	ABB HiPak 5SLD 0600J650100	Si	—	—
D_2	CREE C2D20120D	SiC	—	—
D_3	CREE C4D40120D	SiC	—	—
D_4	Infineon D1031 SH	Si	—	—
D_5	Infineon DD600S65K1	Si	—	—
D_6	6.5 kV, 1 kA prototype [112]	SiC	—	—

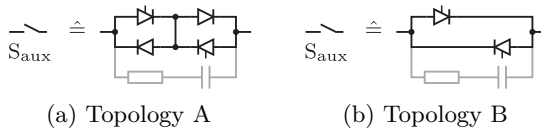


Figure 3.6: Topologies of the bidirectional switch, depicted with thyristors

respectively. Alternatively, a prototype diode presented in [112] is considered. In the end, the results of both approaches are similar. Moreover, Table 3.3 indicates whether a thyristor is able to block reverse voltages. If this is not the case, antiparallel diodes are used as shown in Fig. 3.6 (a) to block the reverse voltage. In the following, this is considered as “topology A”. Furthermore, as the reverse-recovery behavior of S_{aux} is determined by the diodes, the use of additional diodes might be reasonable especially when the thyristor’s recovery charge is very high. In contrast to this, “topology B” omits these diodes. Figure 3.6 also shows an optional RC snubber. It might be required by the active device to limit the rise of the forward voltage. Whether an RC snubber is required, is also indicated in Table 3.3.

3.2.3 Loss Modeling

The losses of the various ARCP implementations shall be evaluated for different power ratings and ratios of the primary and secondary dc voltage. The loss

Table 3.4: Input parameters

DC-link voltage	$U_p = 5 \text{ kV}$
Switching frequency	$f_s = 1 \text{ kHz}$
Main snubber capacity	$C_{sn+} = C_{sn-} = 0.68 \text{ }\mu\text{F}$
Auxiliary inductance	$L_{aux} = 8.3 \text{ }\mu\text{H}$
Dead time of main switches	$50 \text{ }\mu\text{s}$

calculation has to be as precise as possible. Especially, reverse recovery losses of S_{aux} shall be considered. Since these are not satisfactorily calculated by available simulation tools, an appropriate ARCP model is built up, which uses the loss data from the manufacturers' data sheets. This is encouraged furthermore by the relatively simple voltage and current waveforms in an ARCP.

Figure 3.7 shows the principle flow chart of the loss calculation. It includes the input parameters, manufacturers' data-sheet values and the analytical loss model. The input parameters of the following investigation are given in Table 3.4. During the input-side switching instant the main current $i_{p1}(0)$ is assumed to be constant. Dependent on the operating point of the DAB3, the current is calculated to

$$i_{p1}(0) = \begin{cases} \frac{U_p}{3\omega L_s} \cdot \left(\frac{2\pi d}{3} - d\varphi - \frac{2\pi}{3} \right) & \text{for } 0 \leq \varphi \leq \frac{\pi}{3} \\ \frac{U_p}{3\omega L_s} \cdot \left(\pi d - 2d\varphi - \frac{2\pi}{3} \right) & \text{for } \frac{\pi}{3} < \varphi \leq \frac{\pi}{2} \end{cases} \quad (3.1)$$

$$i_{s1}(\varphi) = \begin{cases} -\frac{U_p}{3\omega L_s} \cdot \left(\frac{2\pi d}{3} + \varphi - \frac{2\pi}{3} \right) & \text{for } 0 \leq \varphi \leq \frac{\pi}{3} \\ -\frac{U_p}{3\omega L_s} \cdot \left(\frac{2\pi d}{3} + 2\varphi - \pi \right) & \text{for } \frac{\pi}{3} < \varphi \leq \frac{\pi}{2} \end{cases} \quad (3.2)$$

according to [2]. Note that $i_{s1}(\varphi)$ is the main current for a commutation in the output bridge. For comparison, also the losses for a DAB3 without an ARCP are calculated with the same model.

3.2.4 Simulation Results

Considering the different power-electronic devices in Table 3.3, the loss savings by the ARCP are calculated. For this purpose, a load profile is assumed as depicted in Fig. 3.8. It considers part-load conditions as well as dynamic voltage variations. Although this fictitious load profile does not correspond to an actual application, it is a suitable benchmark for the different ARCP configurations.

Table 3.5 shows the energy saved by the ARCP compared to a setup without ARCP. If no ARCP is used, the main switches are not applied with the lossless

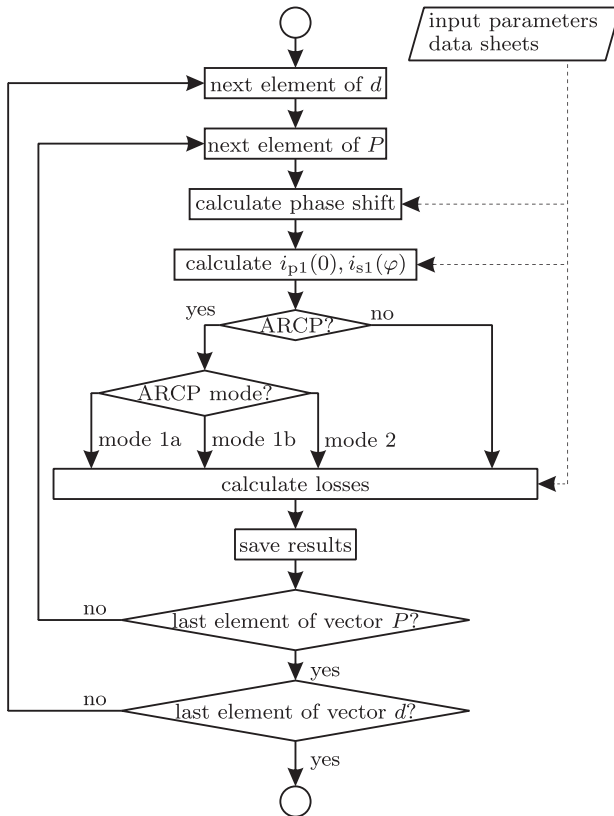


Figure 3.7: ARCP loss calculation

snubbers. Moreover, the table gives the average energy saving per phase and per switching period. The following conclusions can be drawn from the results.

The sole use of two silicon thyristors T_2 has no positive effect on the average losses. This is basically due to the high turn-off losses of the thyristor and the losses in the RC snubber that occur also when the ARCP is inactive. If the diode D_5 is connected in series, to limit the recovery charge (topology A), the ARCP performs very well and saves in total (for all three phases) an energy of around 11.3 J per switching cycle. For this configuration, which is named “ T_2D_5 ”, Fig. 3.9 (a) shows the losses in the whole operation range. The figure illustrates the switching losses generated in one phase leg in the input bridge per switching period. The

black curve represents the switching losses without an ARCP. As soft-switching is not ensured, snubber capacitors are not considered at the main switches in this case. Moreover, the black curve includes losses from a di/dt snubber that has to be applied for IGCT converters. It prevents the diodes from destructive reverse-recovery currents [101]. The gray curve depicts the commutation losses with the use of an ARCP. The di/dt snubber can be omitted as the diodes do not turn off hard. It is evident that the ARCP configuration has lower losses. However, it suffers from the no-load losses of the thyristor's snubber.

The configuration “T₂D₅” does not inject any boost current, if the load current is sufficient to compensate the commutation losses. Therefore, the comparison between configuration “T₂D₅” and “T₂D₅(Boost)” gives the benefit of the introduced “mode 1a”. A significant part of the loss is generated in the RC snubber of the thyristor, even though the ARCP is deactivated. “T₂D₅(Decouple)” separates the RC snubber from the ARCP if it is deactivated using an additional solid-state switch or relay. As losses can be reduced significantly, the extra effort is worth to consider.

Furthermore, it is an interesting option to use SiC thyristors due to their small reverse-recovery charge. However, as they are not reverse blocking capable, an antiparallel diode is needed as shown in Fig. 3.6 (a). This diode has to be an SiC device as well since the diode determines the reverse-recovery behavior of the series connection of thyristor and diode. Using an Si diode would negate the advantage of the SiC thyristor. Figure 3.9 (b) shows the results in case SiC thyristors and SiC diodes are used as a bidirectional switch. Due to the absence of an RC snubber, the efficiency is improved especially in the naturally soft-switched operating area. When the ARCP is active, the higher conduction losses of the SiC devices impair the small turn-off losses.

The option “IGCT” uses an IGCT instead of a thyristor. The IGCT is turned off actively when the current crosses zero. This comes with two advantages: Firstly, the IGCT does not need an RC snubber. Secondly, there is no presence of a reverse-recovery charge. As the conduction losses are lower, this configuration achieves a higher efficiency than the SiC solution (cf. Fig. 3.9 (c)). However, additional sensor and control effort are required to detect the zero crossing of the current.

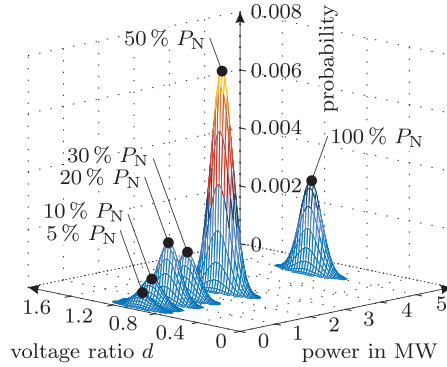


Figure 3.8: Load profile based on the Euro Efficiency [113] for PV applications

Table 3.5: Saved energy by the ARCP (per phase, averaged over entire load profile)

Configuration	Input Bridge	Output Bridge
T_2	-0.65 J	0.00 J
T_2D_5	1.75 J	2.03 J
$T_2D_5(\text{Boost})$	1.54 J	1.85 J
$T_2D_5(\text{Decouple})$	2.57 J	2.90 J
T_1D_5	1.66 J	1.95 J
T_1D_3	3.50 J	3.69 J
T_1D_6	3.54 J	3.72 J
IGCT	3.73 J	3.88 J

3.2.5 Measuring Results

To demonstrate the feasibility of the concept, full-scale prototypes are designed and built for a single phase. Moreover, the additional effort in control, device volume and cost can be judged.

The configurations “ T_2D_5 ” and “IGCT” are implemented since they are the most cost effective and most efficient solution respectively. The ARCP is designed for the DAB3 parameters given in Table 3.1 except for the dc-link voltages, which are now considered dynamically in the range of 4.5 to 5.5 kV.

For this application, the auxiliary inductance is determined to be 8.3 μH . The peak current in the inductor is 1.8 kA, while the RMS current does not exceed 380 A.

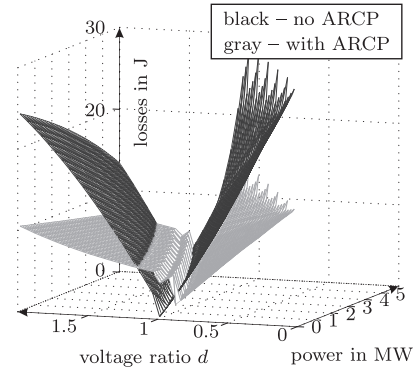
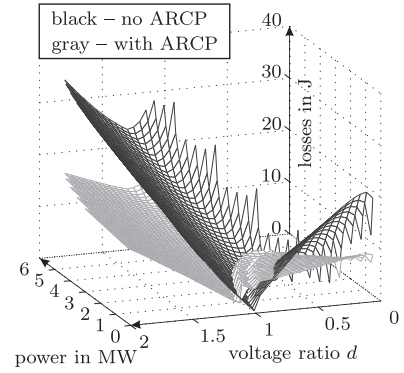
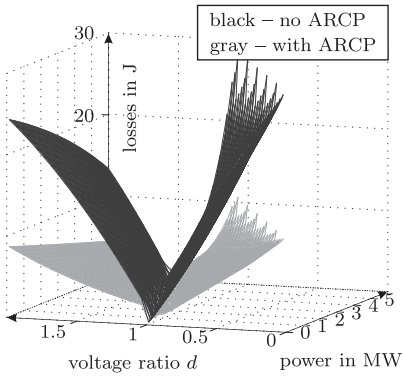
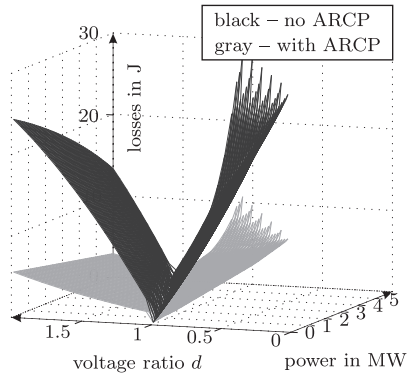
1.36 μF are connected in parallel to each main switch as snubber capacitance. Due to availability reasons for the configuration “IGCT”, the same devices for the auxiliary switch have been used as for the main switches, “5SHY 3545” from ABB Switzerland. A reverse blocking IGCT, like the Mitsubishi GCT GCU15CA-130, however, would achieve higher efficiencies since it does not require the antiparallel diodes. The logic to determine the zero-current crossing has been implemented as well based on the work of [114].

For the thyristor-based configuration (“ T_2D_5 ”), the fast silicon thyristors “R1127 NC 36 R” from IXYS Westcode are used [115]. These press-pack devices offer turn-off times below 200 μs at a maximal voltage of 3.6 kV. To reduce the effective reverse-recovery charge, the Infineon diode “DD600S65K1” is connected as antiparallel diode to each thyristor (cf. Fig. 3.6 (a)). While the reverse-recovery charge of the diode is given with 1050 μC when turning off 600 A at 2000 A/ μs and 125 $^\circ\text{C}$, the thyristor’s theoretical value lies beyond 5000 μC . Therefore, the switching losses of the auxiliary switch are supposed to be decreased heavily.

Firstly, the setup is commissioned in the laboratory as depicted in Fig. 3.10 (a). Subsequently, the ARCP is also connected to the medium-voltage dc-dc converter rated for 5 MW as shown in Fig. 3.10 (b). The successful tests ensure that the ARCP also works with the increased parasitics of a medium-voltage setup.

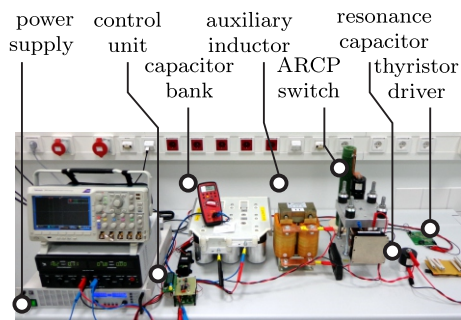
Figure 3.10 (c) shows the corresponding measurement with the configuration “ T_2D_5 ”. The load current, which is not depicted in the measurement, is $i_{p1} = -1.5$ A. Without ARCP, this would result in a hard-switched operating point and a short-circuit of the snubber capacitor when the main switch turns on. However, the ARCP injects a boost current of about 150 A using a fixed boost time. Therefore, the snubber capacitors are reloaded and the upper main switch $S_{\text{main}+}$ can be turned on under ZV conditions.

Moreover, it can be seen from the measurements that the reverse-recovery effect is solely determined by the diode D_5 . From the measurements it is evident that during the reverse recovery the entire voltage is blocked by the diode connected in series to the thyristor. Moreover, the other diode (connected in antiparallel to the thyristor) prevents the reverse voltage from building up across the thyristor. Consequently, the charge carrier are not flushing out of the device as reverse-recovery current but recombining inside the device. Consequently, the reverse-recovery charge is solely determined by the diode and the large reverse-recovery charge of the thyristor can be suppressed effectively.

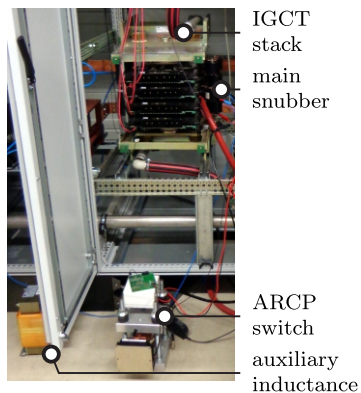
(a) Configuration T_2D_5 (Si) – Input Bridge(b) Configuration T_2D_5 (Si) – Output Bridge(c) Configuration T_1D_6 (SiC) – Input Bridge

(d) Configuration IGCT – Input Bridge

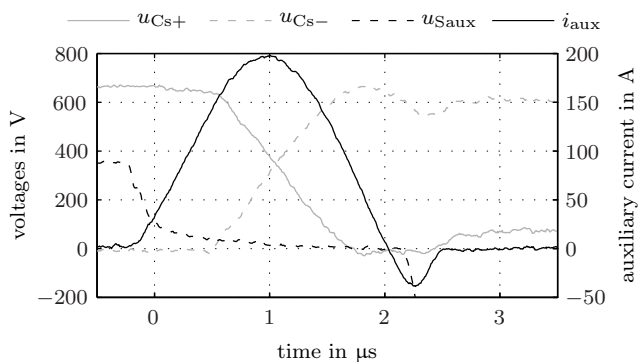
Figure 3.9: Loss comparison with and without ARCP



(a) Commissioning in the lab



(b) Commissioning at the converter



(c) Measuring results in the converter

Figure 3.10: Prototype verification

3.2.6 Advantages of Applying an ARCP

Ensuring soft-switching operation through the ARCP comes with several advantages, which shall be investigated further.

Efficiency Improvement and Payback Period

As already discussed, the ARCP saves switching losses in the converter by ensuring soft-switching. The improvement of efficiency is one of the main advantages of the ARCP. The increase, however, depends on the load profile of the converter. Especially when operated at partial-load conditions or at unequal dc voltages, the dc-dc converter benefits from the ARCP.

One of the preferable applications for high-power medium-voltage dc-dc converters is the connection of wind turbines to an MVDC collector grid [116]. Therefore, as an example, the energy savings for a wind application shall be investigated in the following. Subsequently, the payback period can be calculated easily.

To determine the load profile, data of the wind farm “Alpha Ventus” located in the North Sea are considered. Moreover, the power profile of a 5 MW “Areva Multibrid M5000” wind turbine as depicted in Fig. 3.11 is considered [117].

The weather conditions of “Alpha Ventus” are provided by the research platform “FINO1” [118]. “FINO1” provides weather data as the statistical distribution of different wind speeds v from the last 10 years. The occurrence probability p can be represented by a Weibull distribution [119]

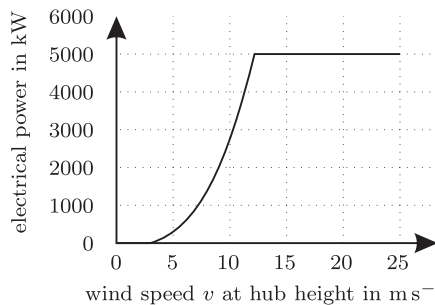


Figure 3.11: Power curve of an “Areva Multibrid M5000” wind turbine

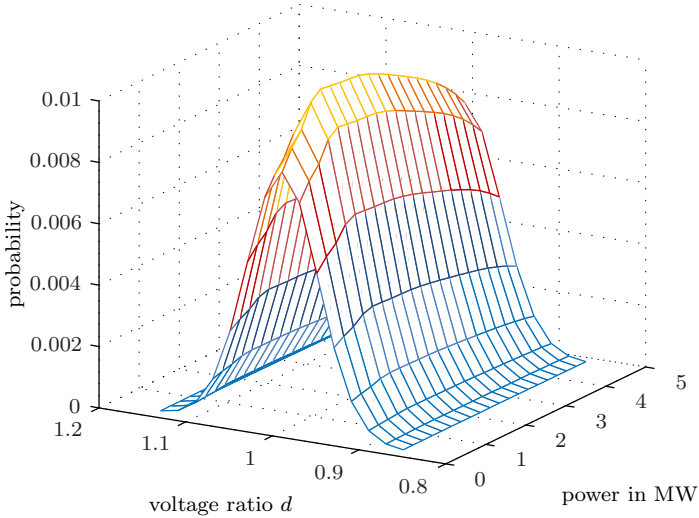


Figure 3.12: Load profile resulting from the Weibull distribution in the North Sea and the power profile of an “Areva Multibrid M5000”

$$p(v) = \frac{k}{\lambda} \left(\frac{v}{\lambda} \right)^{k-1} e^{-(v/\lambda)^k} \text{ with } k = 2.2 \text{ and } \lambda = 10.5. \quad (3.3)$$

The voltage ratio d is approximated using a Gaussian distribution with a standard deviation of $\sigma_d = 0.04$. The resulting load profile is illustrated in Fig. 3.12.

Multiplying the load profile with the saved commutation energy as depicted in Fig. 3.9 gives the averaged energy saving per switching period. Multiplying it with the switching frequency of 1 kHz gives the average loss savings. Consequently, an ARCP with the configuration “T₂D₅” saves 8.4 kW on average.

To evaluate the amortization period, a feed-in tariff of 0.19 €/kWh is considered according to the German Renewable Energy Act [120].

The additional costs for the ARCP components (configuration “T₂D₅”) are approximately 1600 €. The costs are determined by the built prototype and therefore do not include volume discount. The following components are included in the costs: thyristors, diodes and RC snubber (for the switch S_{aux}), the inductor L_{aux} and the capacitors C_s . About 70 % of the costs are semiconductor cost. To equip

six inverter legs of a 5 kV DAB3 with the ARCP technology, the investment of 9600 € is required.

Accordingly, the period of amortization is

$$T_{T_2D_5} = \frac{9600 \text{ €}}{0.19 \text{ €/kWh} \cdot 8.4 \text{ kW}} \cdot \frac{1}{365 \cdot 24 \text{ h/a}} = 0.69 \text{ a.} \quad (3.4)$$

An analogue investigation for the configuration “IGCT” leads to

$$T_{IGCT} = \frac{28\,000 \text{ €}}{0.19 \text{ €/kWh} \cdot 21.6 \text{ kW}} \cdot \frac{1}{365 \cdot 24 \text{ h/a}} = 0.78 \text{ a.} \quad (3.5)$$

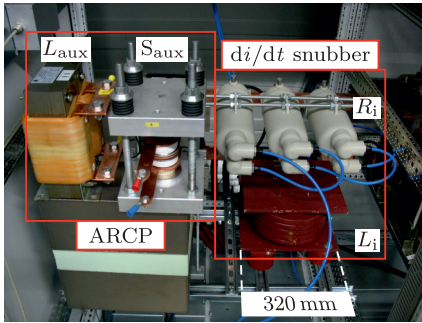
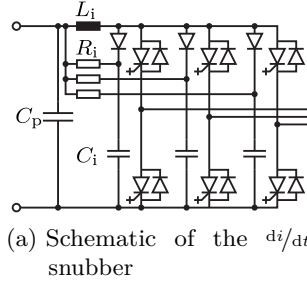
Consequently, the costs for the additional components amortize after 8 to 9 months in case of the thyristor-based solution. If the ARCP is realized based on IGCT devices, it amortizes after 9 to 10 months and allows higher savings afterwards.

Saving IGCT related di/dt Snubbers

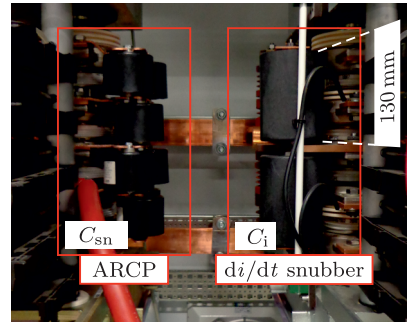
One of the major drawbacks of IGCTs is the need for a di/dt snubber as depicted in Fig. 3.13 (a). It is required to limit the di/dt during the IGCT turn-on and the diode turn-off respectively. Without the snubber, the steep diode turn-off would lead to a destructive reverse-recovery current peak. Moreover, a homogeneous turn-on of the IGCT is ensured without causing hot spots in the silicon.

For soft-switched operation points, however, the di/dt snubber is not required since the diode turns off naturally when the current crosses zero. Hence, the ARCP, ensuring soft-switching in the entire operating range, allows the omission of the di/dt snubber.

Fig. 3.13 (b) and (c) depict the devices for the single-phase ARCP prototype and the di/dt snubber for a 5 MW medium-voltage converter. Since the di/dt snubber can be saved, the installation of an ARCP does not require much additional space. Moreover, losses and cooling effort are reduced due to saving the water-cooled resistors in the di/dt snubber.



(b) S_{aux} and L_{aux} vs. di/dt snubber's inductor and resistors



(c) C_s vs. di/dt snubber's capacitance

Figure 3.13: Size comparison of the ARCP with the obsolete di/dt snubber

Improved Converter Safety

One critical fault in power-electronic converters is a failure of a semiconductor device. In contrast to module devices, the press-pack devices always fail short. Hence, the behavior during a failure is determined and redundancies can be implemented easier.

A failure of a single device is not very crucial initially. However, when the conjunctive device of the same phase turns on, the dc link is short circuited. This in turn leads to excessive over currents and to the potential destruction of the converter.

In zero voltage (ZV) switched converters, the voltage in the turn-on instant of a semiconductor device is zero in general. If a single device fails, however, the

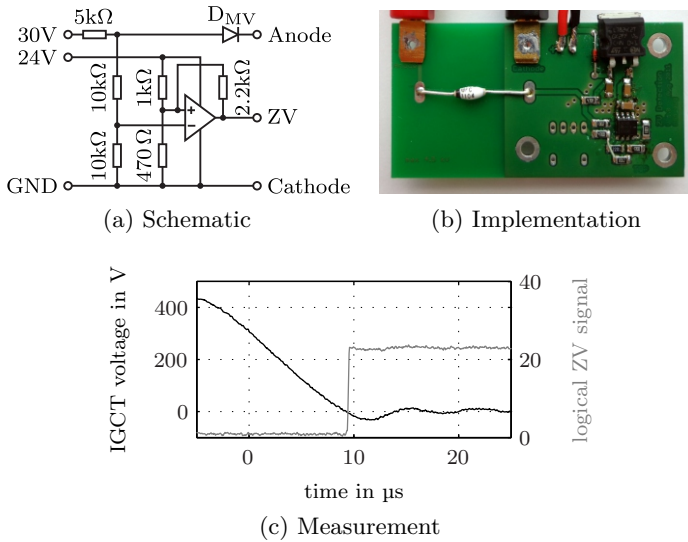


Figure 3.14: Zero-voltage detection circuit

voltage across the complementary device of the same phase leg will not decay to zero. By the short-circuit of the single device, the voltage will be clamped to the dc-link voltage. If the ZV condition is detected before the switch turns on, the dangerous dc-link short circuit can be avoided.

Figure 3.14 (a) shows such a ZV detection circuit. It is based on the circuit presented by Köllensperger [121] with adapted parameters for the given application. As soon as the IGCT voltage falls below 30 V, diode D_{MV} becomes forward biased and the device voltage is applied to the subsequent circuitry. If the voltage falls below a certain limit, the schmitt-trigger indicates the ZV condition by a 24 V logic signal. This signal can be fed directly to a controller or to an optical transmitter.

This circuit has been implemented on a 35 mm by 70 mm printed-circuit board (PCB) as shown in Fig. 3.14 (b).

In Fig. 3.14 (c), a measurement shows the IGCT voltage in an ARCP during commutation. It demonstrates a clear detection of the ZV condition. The schmitt-trigger circuit (cf. Fig. 3.14 (a)) ensures a certain tolerance against noise.

4 Medium-Frequency High-Power Transformer

The following chapter is dedicated to the medium-frequency transformer. Connecting both semiconductor bridges, it is one of the key components in the DAB3. It provides galvanic isolation and ensures high efficiency even at large voltage-conversion ratios.

The mega-watt prototypes in Chapter 5 are built using proven design methodologies assuming sinusoidal voltage and current waveforms. Therefore, this chapter investigates the core losses for sinusoidal voltages and rectangular voltages generated by power electronics.

Chapter 2 highlighted the tight tolerances on the stray inductance for dc-dc converters. Consequently, several transformer designs, suitable for MV applications, are compared in terms of achievable stray inductance and sensitivity on physical dimensions.

The high power density and the requirements on the stray inductance make the transformer design challenging. Alternatively, the series inductance could be realized through separate inductors connected in series to the medium-frequency transformer. This alternative option is discussed for the MV dc-dc converter.

Some of the following content has been published as well in [12, 122, 123].

4.1 Introduction

4.1.1 Review on different applications

Transformers in general are used for a huge variety of applications whenever ac voltages need to be stepped up or down efficiently or if galvanic isolation shall be established. Regarding the design and the used materials, three decisive parameters can be identified: current, voltage and frequency.

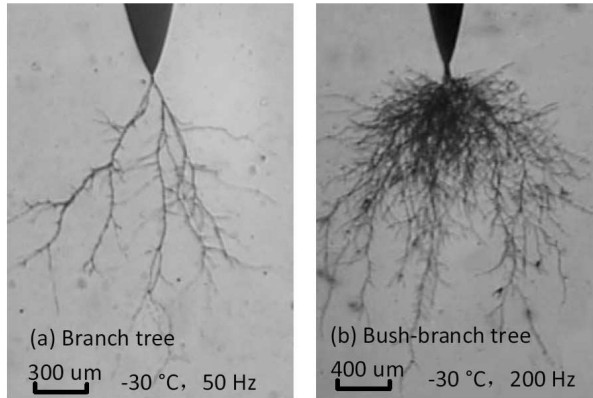


Figure 4.1: Treeing in silicone rubber at 50 Hz and 200 Hz (Source: [134])

The current rating of the transformer or (more precisely) the RMS current through the phases determines the cross-sectional area of the transformer wires and with it space required in the winding window of the transformer. Moreover, at higher frequency additional effects, namely the skin and proximity effects, increase the ohmic losses further. To suppress these influences, usually more space is required in the winding window since copper-fill factors are reduced.

The operating voltage at the windings and the voltage withstand between primary, secondary and ground, determine the requirements of the transformer insulation. At high voltage levels, in the order of 3 kV and above, partial discharge (PD) has to be considered [124]. PD is either the cause or an indicator for the aging of the insulation material being it gaseous [125], liquid [126, 127] or solid material [128]. Especially in solids like polyimides, PD causes a degeneration of the insulation material. Reasons for this are chemical processes and physical damaging through particle bombardment [129]. This results in treeing while a quicker growth of the tree, i.e. a faster aging, with increasing frequency has been observed (cf. Fig. 4.1). However, this effect has not been understood fully [130–134].

As discussed before, elevated frequencies result in complexer winding configurations and probably a faster degeneration of insulation material. Moreover, higher eddy currents occur in the core materials increasing the core losses. Also the higher power density comes with additional challenges in the thermal design, since less surface is available to dissipate heat.

Consequently, voltage, current and frequency mainly determine the feasibility

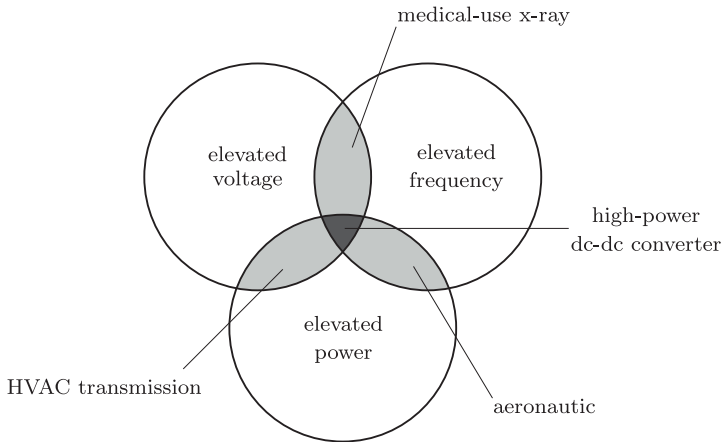


Figure 4.2: State-of-the-art transformers and their applications

of a transformer design. Additionally, high voltage, current and / or frequency ratings make the transformer design challenging so that often only two of these parameters are of a high rating. In aeronautics, for example, transformers are operated at elevated power and frequency while the voltage level, however, is fairly low. Figure 4.2 gives additional examples.

In contrast, the transformers in high-power dc-dc converters need to achieve a high power throughput at high frequency and fairly high voltage. The following chapter is dedicated to that kind of transformer.

4.1.2 Windings

Considering the core and the windings, there is a big variety of materials and styles that can be considered for a given application.

As winding material usually copper is used since its specific resistance is very low allowing a very compact design. Increasingly, aluminum is used since it is less expensive and features a lower density than copper. However, aluminum is more difficult to process - especially connecting copper and aluminum parts with each other. Moreover, the specific electrical resistance of aluminum is higher compared to copper. This fact, however, might be especially interesting at higher frequencies. At high frequencies, current displacement named skin effect results in

a poor utilization of the material inside e.g. a round conductor. The utilization is given by the skin depth [135]

$$\delta \approx 2\rho\sqrt{\frac{\epsilon}{\mu}} \quad \text{for } \omega \gg 1/\rho\epsilon, \quad (4.1)$$

which in turn depends the electrical resistance ρ , the electric permittivity ϵ and the magnetic permeability μ of the material.

In the end, the skin depth of an aluminum conductor is deeper compared to copper at the same frequency leading to a higher utilization. This is demonstrated using the software FEMM [99]. Figure 4.3 shows two simulations in which a sinusoidal current having an RMS value of 100 A and a frequency of 1000 Hz is fed to a round wire with a diameter of 10 mm. While in Fig. 4.3 (a) and (c) the wire is made from copper, the simulation in (b) and (d) results from an aluminum wire. It is demonstrated that current distribution is more homogeneous and, hence, the utilization of the aluminum wire is higher. In medium-frequency high-power application, aluminum might be the favorable winding material finally, since thicker litz wires or foil can be used (as explained below).

Besides the already mentioned skin effect, the proximity effect is another source of current displacement. Different from the skin effect, it originates from the high-frequency currents in the neighboring windings. Figure 4.4 depicts the proximity effect of two neighboring windings carrying the same current but in different directions. Due to the magnetic field between the windings and the Lorentz force, the current is displaced. As before, the effect is more intense considering copper material.

Both, skin and proximity effect, lead to a current displacement in the conductor. Consequently, the losses, being proportional to I^2 , increase as well. To suppress these effects, the geometry of the conductor should be in the order of the before mentioned skin depth. In practice, two kinds of windings are usually used for these applications, foil winding or high frequency (HF) litz wire.

The foil winding is a very cheap and very effective solution. The winding consists of a thin foil that is wrapped around the core. The thickness of the foil is chosen to be not more than twice the skin depth to suppress the eddy currents. Due to the geometry, however, the foil winding suffers from large capacitance values from windings to core and within the winding. The effect on the converter should be regarded beforehand.

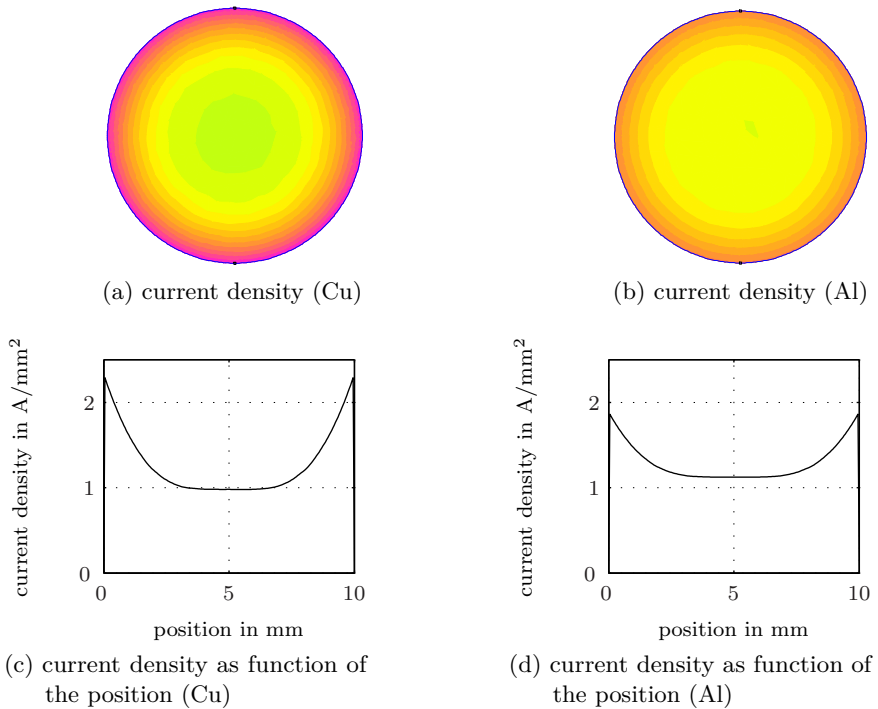


Figure 4.3: Current distribution due to skin effect comparing copper and aluminum

As an alternative, HF litz wire can be used. One wire consists of several hundreds to thousands isolated parallelized strands, each being small compared to the skin depth. Therefore, displacement currents are suppressed. Due to the use of this many strands, the copper fill factor is not very high, leading to a high stray inductance. As already discussed, however, a certain stray inductance is desired in a dc-dc converter application. Consequently, it might be a promising solution. Unfortunately, litz wire is expensive compared to foil windings.

4.1.3 Core Materials

The choice of the core material mainly depends on the frequency components of the voltage and the power rating. Limiting factors are the core size, the costs

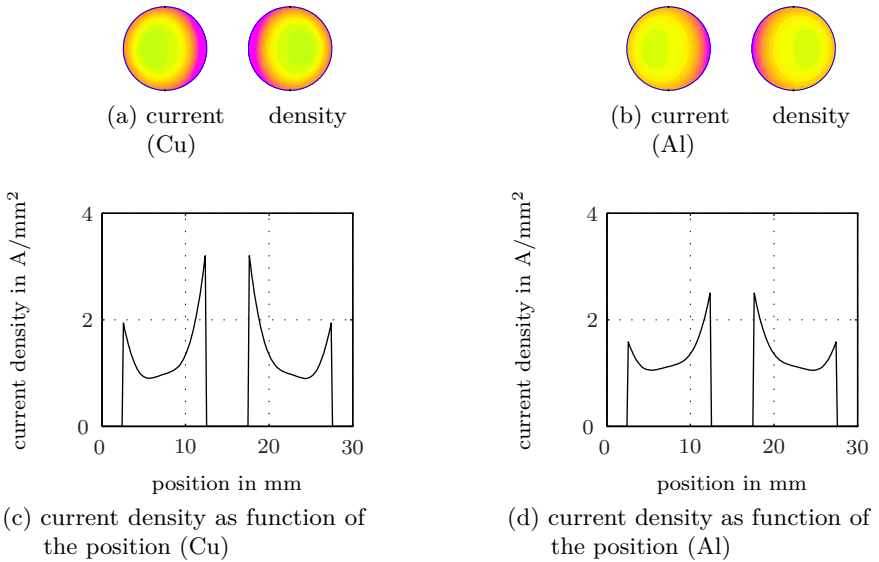


Figure 4.4: Current distribution due to the proximity effect comparing copper and aluminum

and the losses. The core size is determined by the saturation field strength B_{sat} . The costs increase nearly linearly with the core weight. The core losses consist of the hysteresis losses and eddy-current losses. While the coercive magnetic-field strength H_c is a measure for the hysteresis losses, eddy-current losses depend on the resistivity of the material, the lamination thickness or the particle size.

For the given application, the interesting soft-magnetic materials can be roughly classified as silicon steel, amorphous iron, nanocrystalline material, soft-magnetic composite (SMC) and ferrite. Each material has different advantages and shall be introduced briefly in the following. At the same time, the key facts of some prominent core materials are enlisted in Table 4.1.

Grain oriented or non-oriented electrical steel is a classical and well known material for transformers and inductor cores. Usually, a fraction of silicon (typically 3 % to 6.5 %) is added to the iron to increase the resistivity of the material. Consequently, eddy current can be reduced. This material is available in sheets or ribbons with different thicknesses. While lamination thicknesses of 230-350 μm are typically used in 50 Hz applications, thicknesses of 50-180 μm are especially interesting for

medium-frequency applications. One advantage of silicon steel is the maximal flux density of 2 T, which is the highest compared to the other core materials. This gives the chance for compact transformer designs since less core material is required. Another advantage is the low price, which is due to the well known manufacturing processes and the fairly large required volume. However, the price increases with decreasing thickness. In comparison to the 180 μm material, the price per kilogram increases by a factor of 4 and 12 for the 100 μm and 50 μm materials respectively.

Amorphous iron is produced through rapid solidification. The melted metal alloy is cooled down rapidly by flowing on metallic drum which is water cooled. This results in a solid ribbon with a very small thickness of around 20 μm . Moreover, through a fairly small coercivity the hysteresis losses are small as well. Compared to silicon steel, the core losses can be reduced significantly as also indicated by the figures in Table 4.1. Also the price has come down with time. Today the price per kilogram of amorphous material is in the range of silicon steel. However, one big disadvantage of amorphous iron is the high magnetostriction, meaning the linear expansion of the material under the influence of magnetic fields. In the application, magnetostriction results in increased noise emission and additional losses. Moreover, if a lap joint is used to close the magnetic core, its temperature increases significantly in comparison to the rest of the core [2]. Therefore, the entire transformer needs to be derated to keep the lap-joint temperature within the limits.

In the first steps, producing nanocrystalline materials is similar to amorphous metals. Through an additional annealing process at for example 500–600 $^{\circ}\text{C}$, however, the amorphous structure is converted to a crystalline structure. Since the size of the crystals is in the range of 10 nm, it is referred to as nanocrystalline. Compared to amorphous material, nanocrystalline compositions have a smaller saturation flux density as well as smaller coercivity. Considering the same flux density, the losses of the nanocrystalline core are $1/5$ compared to an exemplary amorphous core [136]. Moreover, nanocrystalline materials do not have the issues of magnetostriction as indicated in Table 4.1. The price per kilogram, in turn, is four times higher compared to amorphous material.

For comparison, a “high polarization” ferrite core material is enlisted in Table 4.1 as well. The saturation flux density is very low compared to the other materials. This is also the reason, why the target frequencies for ferrite material is 10 kHz and above.

SMC might serve as a compromise. Commercial materials offer a saturation flux density equal to amorphous metals, but the manufacturing process is similar to

ferrites using a sintering process. Moreover, it could be that the sintering process allows alternative core geometries and very low costs in a series production. On the other hand, the sintering process limits the maximal core size. Pressures of 5-8 t/cm² are required for the sintering while industrial presses maximally achieve forces equivalent to 800 t [137]. In the end, the geometry is limited by the maximal capability of industrial presses. Another benefit is that the material losses can be tailored for different applications: through different grain sizes, an optimal compromise between permeability and hysteresis losses can be set. Moreover, different metallic powders can be used. Using powders of iron (Fe) and cobalt (Co) for example, high saturation-flux densities of above 2 T have been reported while the grain size is less than 3 μm [138]. Since the eddy current losses of the material are proposed to be very low [139], it might be especially interesting when the operating frequency is increased further.

Table 4.1: Soft-magnetic materials

Category	Name	B_{sat} ^a in T	H_c in A/m	resistivity in $\mu\Omega$ cm	ribbon thickness in μm	magneto- striction in ppm	losses ^b in W/kg	Ref.
Si steel	TKES Power Core H	2.03	5	48	180	-	34	[123, 140]
Amorph	JFE 10JNEX900	1.8	15	82	100	0.1	18.7	[141, 142]
	VAC vitrovac 6030 F	0.82	-	130	17-25	<0.2	-	[143, 144]
	Metglas 2605SA1	1.56	<4	130	23	27	5	[145, 146]
	"Fe-based amorphous"	1.5	-	130	25	27	5.5	[142]
Nano- cryst.	VAC vitroperm 500	1.2	<3	115	18	0.01-1	-	[147]
	Hitachi Finemet FT-3H	1.23	0.6	120	18	0.1	-	[136]
Ferrite	Hitachi MB22D (Mn-Zn)	0.51	7.5-15	500	-	-	-	[148]
SMC	X-Somaloy 700 HR	1.57	120	70 000	-	-	104	[149]
	Höganäs Somaloy 700HR-5P	1.53	-	-	-	-	63	[139]
	PM4EM 10	1.56	249	-	-	-	168	[150]

^a or $B(H = 10 \text{ kA/m})$ ^b at 1 T and 1 kHz

4.2 Core Losses in a Dual-Active Bridge Application

In the past, many researchers have investigated the core losses under non-sinusoidal excitation. Most of them regard drive applications using pulse-width modulation (PWM) [151–154] or fundamental frequency modulation at 50–60 Hz [155]. Since the frequency in the DAB3 is one to two decades higher involving additional loss mechanism, this effect is investigated further in the following. This work has been published first in [123].

4.2.1 Core Loss Measurement

For this investigation laser-processed grain-orientated steel with a lamination thickness of 180 μm is measured at frequencies up to 10 kHz. The measurements have been performed at the Institute for Electrical Machines (IEM), RWTH Aachen University. The test bench is able to generate arbitrary flux-density forms in the magnetic material. Consequently, a sinusoidal flux can be compared with the piece-wise linear flux in a DAB3 application. In the following, the two test cases are referred to as “sinusoidal” and “DAB3” respectively. Figure 4.5 shows the phase voltage, the magnetic-flux density and the hysteresis loop for both cases. Already the hysteresis loop reveals that the core loss density, represented by the spanned area of the loop, is slightly lower for the DAB3 case. Figure 4.5 (d) gives the specific iron losses P_s measured by the test bench as a function of the peak flux density $\hat{B}$. Some points are given separately in Table 4.2. It is evident that the core losses for a DAB3 application are lower compared to a sinusoidal excitation with equal peak flux density. This results from the lower rate-of-change $\frac{dB}{dt}$ of the magnetic flux density. This effect shall be validated using state-of-the-art analytical models.

Table 4.2: Measured specific core losses

$\hat{B}$ in T	Specific core losses in W/kg		Relative deviation $\frac{ P_{s,\text{sinus}} - P_{s,\text{DAB3}} }{P_{s,\text{DAB3}}}$
	$P_{s,\text{sinus}}$	$P_{s,\text{DAB3}}$	
0.1	0.54	0.41	32 %
0.5	9.61	8.59	12 %
1.0	33.96	31.09	9 %
1.5	73.90	69.73	6 %

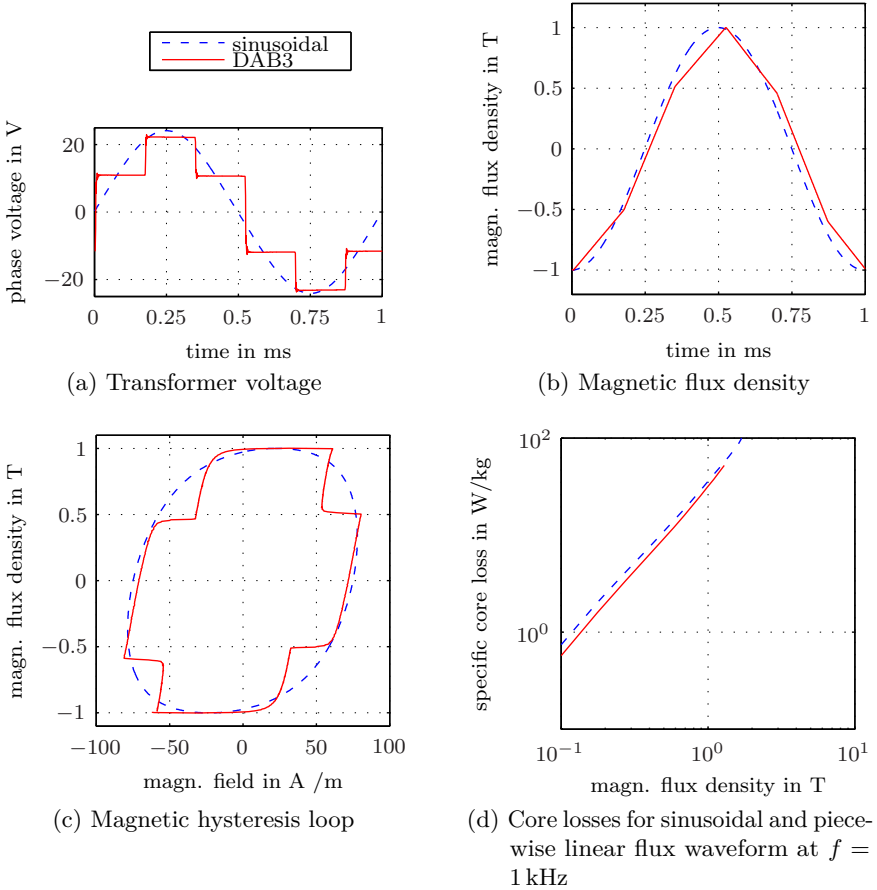


Figure 4.5: Comparison of sinusoidal flux excitation and flux density in a DAB3

4.2.2 Analytical Validation

For the modeling of core losses under non-sinusoidal, unbiased excitation the improved generalized Steinmetz equation (iGSE) [156] has proven. It uses Steinmetz parameters α , β and k according to the original Steinmetz equation (OSE) [157]

$$P_s = k f^\alpha \hat{B}^\beta \text{ with } [P_s] = \frac{\text{W}}{\text{kg}}. \quad (4.2)$$

The Steinmetz parameters are determined through several measurements at different frequencies and flux densities during sinusoidal excitation. While the frequency is set to 1 kHz, 5 kHz, 7 kHz and 10 kHz, the peak flux density is varied from 0.1 to 1.8 T in 0.1 T steps. From the measurements, the following Steinmetz parameters can be determined for the chosen silicon steel material with a thickness of 180 μm :

$$\alpha = 1.6155 \quad \beta = 1.7021 \quad k = 5.2 \cdot 10^{-4}. \quad (4.3)$$

Figure 4.6 shows the measuring points together with the lines of best fit resulting from the evaluated Steinmetz parameters.

Equations (4.4) and (4.5) represent the well-known improved generalized Steinmetz equations as published in [156].

$$P_s = \frac{k_i (\Delta B)^{\beta-\alpha}}{T} \int_0^T \left| \frac{dB}{dt} \right|^\alpha dt \quad (4.4)$$

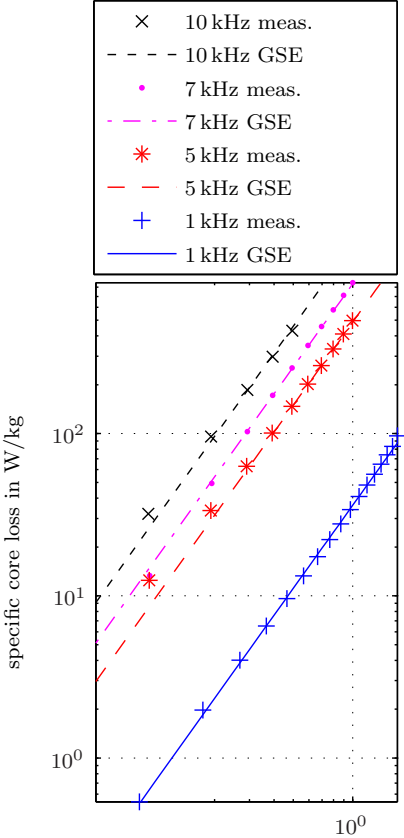
with

$$k_i = \frac{k}{(2\pi)^{\alpha-1} \int_0^{2\pi} |\cos \theta|^\alpha \cdot 2^{\beta-\alpha} d\theta}, \quad (4.5)$$

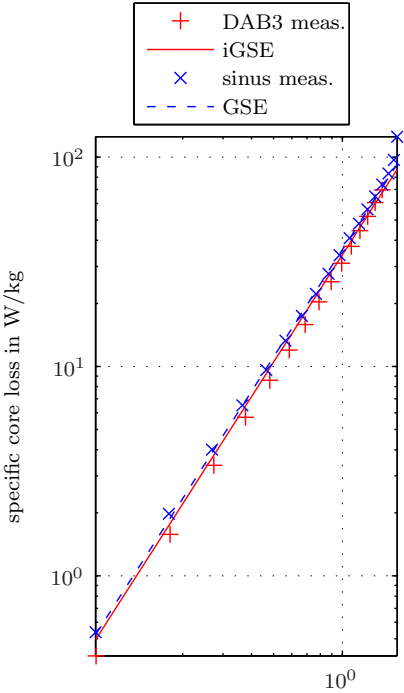
where ΔB is the peak-to-peak value of the flux density, T is the period time of the flux density and α , β and k are the Steinmetz parameters.

For piece-wise flux densities, (4.4) simplifies to [156]:

$$P_s = \frac{k_i (\Delta B)^{\beta-\alpha}}{T} \sum_m \left| \frac{B_{m+1} - B_m}{t_{m+1} - t_m} \right|^\alpha (t_{m+1} - t_m), \quad (4.6)$$



magn. peak flux density in T
Figure 4.6: Specific core losses in silicon steel for sinusoidal excitation



magn. peak flux density in T
Figure 4.7: Specific core losses for sinusoidal and piece-wise linear course - measurement and iGSE

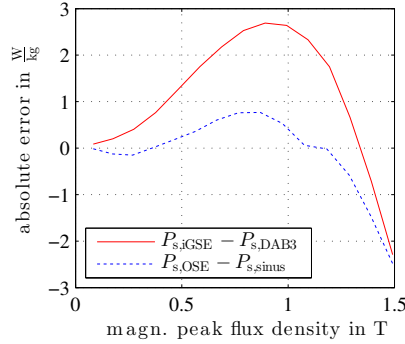


Figure 4.8: Error between the measured losses and the one calculated by OSE and iGSE respectively

where (t_m, B_m) are the supporting points of the piece-wise linear flux.

Now, using the Steinmetz parameters in (4.3) and the iGSE according to (4.6), the losses are calculated for the piece-wise flux in a DAB3. Figure 4.7 shows the results from OSE and iGSE together with the original measurement at a frequency of 1 kHz. While the results from the OSE are apparently in good accordance with the sinusoidal measurement, the iGSE slightly overestimates the losses of the DAB3 flux waveform. Figure 4.8 shows the errors between OSE and iGSE with the respective measurement. It can be seen that there is already a slight error in the OSE. It propagates to the iGSE which uses the same Steinmetz parameters. This error might be related to variations of the permeability for different flux densities and especially the complex permeability of silicon steel at the regarded frequencies [158]. Moreover, eddy currents which occur in the material at the considered frequencies, are not covered by the Steinmetz formulas [159]. The relative error, however, is very small. Consequently, the iGSE delivers accurate core losses for a DAB3 application.

4.2.3 Core Losses Under Load Conditions

The previous discussion assumed no-load condition in the core. In the following, the influence of the load condition on the core losses is further investigated.

Under load condition, both power-electronic bridges are phase shifted according to the load angle. Consequently, also the magnetic flux generated by each bridge

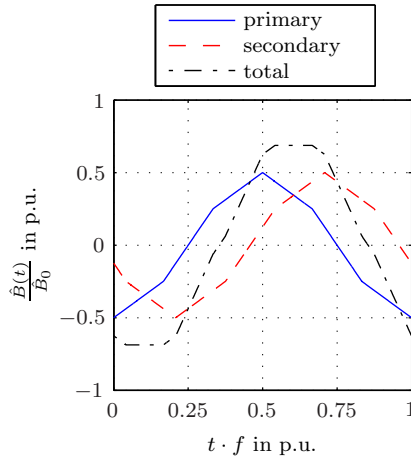


Figure 4.9: Magnetic flux density for a DAB3 at $\phi = 75^\circ$ and $d = 1$

is phase shifted according to Fig. 4.9. Subsequently, the magnetic fluxes generated by primary and secondary winding superpose in the transformer core. Hence, the peak flux density under load is smaller compared to no-load condition. The maximal flux density at no load is referred to as $\hat{B}_0$ in the following.

The iGSE is now used to calculate the core losses at load conditions. Consequently, magnetic relaxation during the phase of constant flux is not considered [160]. Figure 4.10 gives the core losses calculated by the iGSE. Please note that even under load conditions, the losses are plot versus the peak-flux density under no-load conditions $\hat{B}_0$. It is evident that the core losses decrease with increasing power transfer of the converter. Moreover, the core losses decrease faster for load angles larger than 60° , since the flux waveforms overlap in a way that the maximum $\frac{dB}{dt}$ of the total flux is reduced. This effect has been evaluated through measurements. Figure 4.11 shows the measuring results for $\hat{B}_0 = 1$ T. In the figure, two best fit lines are drawn as well, calculated for the intervals $[0^\circ \ 60^\circ]$ and $[60^\circ \ 90^\circ]$ respectively. Also the core losses calculated by the iGSE are given. An almost constant offset between the iGSE and the measurement can be observed, which has been present at no-load conditions as well (cf. Fig. 4.8). Apart from this offset, iGSE and measurement are in good agreement. Therefore, the effect could be verified in practice and should be considered when designing a DAB3 converter.

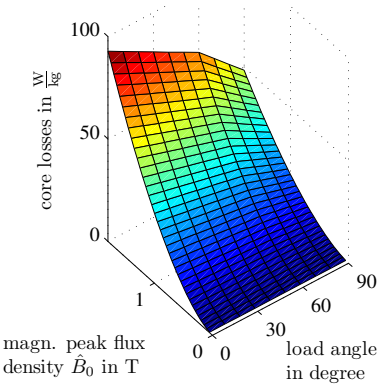


Figure 4.10: Specific core losses in a DAB3 under load condition calculated by the iGSE

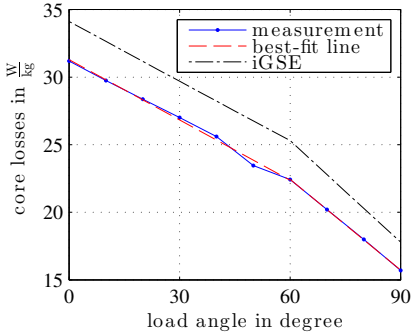


Figure 4.11: Measured specific core losses in a DAB3 under load condition

4.3 Stray Inductance Design Considerations

It shall be noted that the very basics of transformer design are not repeated within this thesis. Only the definition of the transformer's stray inductance is recapitulated due to the importance for this work. For deeper insides in the transformer design, the following references are recommended: Mohan, Undeland, Robbins [161], McLyman [162] and Hurley [135].

4.3.1 Stray Inductance Calculation

Several methods to determine the stray inductance have been evaluated and compared in this work. For simple geometries and a first rough estimation, the stray inductance can be calculated by an analytical approach. On the contrary, FEM simulations can be used to calculate the stray field. Subsequently, through the stray field energy, the equivalent stray inductance can be determined. The FEM simulation can be either performed considering two (2D) or three dimensions (3D). These different methods shall be compared in the following.

Analytical Calculation

As also presented in [161], the procedure to calculate the stray inductance analytically is explained briefly. The analytically derived stray inductance is then compared to the FEM simulation.

To derive the stray inductance, the magnetic energy stored in the stray field shall be calculated according to the volume integral of magnetic field strength H and flux density B . Subsequently, the stray inductance L_σ is the inductance value leading to the same energy amount.

Figure 4.12 shows the cross section of an exemplary winding configuration. It is assumed that the secondary winding is short circuited. In the primary-side winding the current flows out of the drawing plane indicated by a dot. In the secondary-side winding the current has a contrary direction represented by a cross. The winding thickness is a , the wire insulation is b and the coil insulation is c . The course of the magnetic field strength is depicted below. The field increase due to a wire is simplified. It is assumed linear. Moreover, the magnetic field in the

core is assumed to be zero. According to the ampère's circuit law, the maximal magnetic field strength in the winding window is

$$\hat{H} = \frac{N_p \cdot \hat{i}_p}{h_w}, \quad (4.7)$$

where h_w is the height of the winding window.

In the following it is assumed that the relative permeability of the materials in the winding window is 1. Consequently, for this two-layer configuration, the magnetic energy is

$$\begin{aligned} W_m &= \frac{1}{2} \int_V (H(x) \cdot B(x)) dV = \frac{1}{2} \mu_0 \int_V H^2(x) dV = \frac{1}{2} \mu_0 l_m h_w \int_0^{x_8} H^2(x) dx \\ &= \frac{\mu_0 h_w l_m}{2} \hat{H}^2 \left[\int_{x_1}^{x_2} \left(\frac{1}{2} \frac{x - x_1}{x_2 - x_1} \right)^2 dx + \int_{x_2}^{x_3} \left(\frac{1}{2} \right)^2 dx \right. \\ &\quad + \int_{x_3}^{x_4} \left(\left(\frac{1}{2} \right)^2 + \left(\frac{1}{2} \frac{x - x_3}{x_4 - x_3} \right)^2 \right) dx + \int_{x_4}^{x_5} 1 \cdot dx \\ &\quad + \int_{x_5}^{x_6} \left(\left(\frac{1}{2} \right)^2 + \left(\frac{1}{2} \frac{x_6 - x}{x_6 - x_5} \right)^2 \right) dx \\ &\quad \left. + \int_{x_6}^{x_7} \left(\frac{1}{2} \right)^2 dx + \int_{x_7}^{x_8} \left(\frac{1}{2} \frac{x_8 - x}{x_8 - x_7} \right)^2 dx \right] \\ &= \frac{\mu_0 h_w l_m}{2} \hat{H}^2 \left[\frac{2}{2^2} \cdot \left(\frac{a}{3} + b + \left(a + \frac{a}{3} \right) \right) + (b + c) \right] \\ &= \frac{\mu_0 h_w l_m}{4} \hat{H}^2 \left[\frac{5}{3} a + 3b + 2c \right] \\ &= \frac{\mu_0 l_m N_p^2}{4 h_w} \hat{i}_p^2 \left[\frac{5}{3} a + 3b + 2c \right] \quad (4.8) \\ &\stackrel{!}{=} \frac{1}{2} L_\sigma \hat{i}_p^2, \quad (4.9) \end{aligned}$$

where l_m is the mean winding length and N_p the number of primary-side turns.

This gives for the stray inductance considering the 2-layer arrangement

$$L_\sigma = \frac{\mu_0 l_m N_p^2}{2 h_w} \left[\frac{5}{3} a + 3b + 2c \right]. \quad (4.10)$$

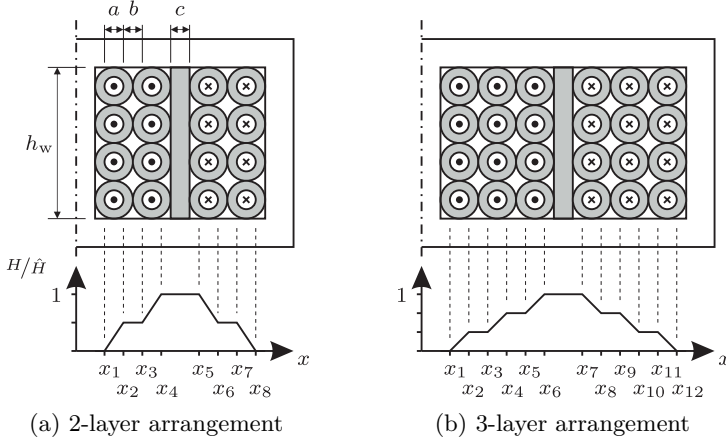


Figure 4.12: Magnetic field in the winding window

Analogously, the stray inductance for the 3-layer arrangement is

$$L_{\sigma} = \frac{\mu_0 l_m N_p^2}{h_w} \left[\frac{14}{9}a + \frac{19}{9}b + c \right]. \quad (4.11)$$

Deriving the formulas for the stray inductances, many assumptions have been made:

- The current in each winding is assumed perfectly homogeneous,
- the course of the magnetic field correlates to a rectangular wire, not a round one,
- each winding has the same length l_m and
- the geometry of the transformer is assumed cylindrical.

Although, the approach depicts the influence of transformer geometry and stray inductances very clearly, it is questionable whether these formulas give trustful results for a practical implementation. Consequently, two- and three-dimensional FEM simulations are investigated to determine the transformer's stray inductance.

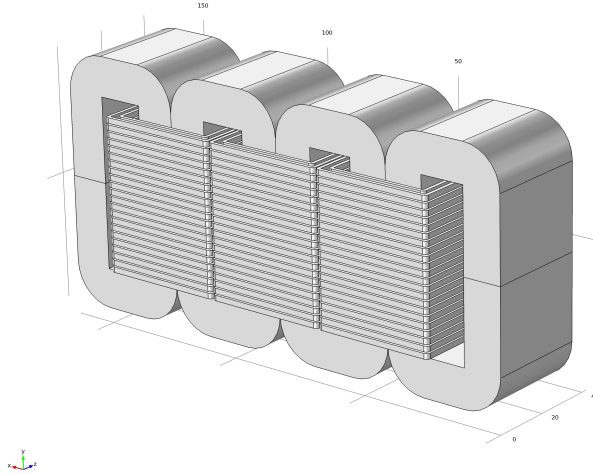


Figure 4.13: Three-phase transformer serving as example for the FEM comparison

3D FEM Simulation

To rate the performance of the FEM tools for determining the stray inductance, a low-power transformer is modelled. The transformer is part of a small-scaled dual-active bridge serving as laboratory prototype. It is a three-phase five-legged arrangement as depicted in Fig. 4.13.

Due to the computational burden, one phase of this transformer is simulated using the software Comsol [163]. The dimensions of the investigated transformer core are (regarding Fig. 4.14)

$$\begin{aligned} h_{\text{core}} &= 13 \text{ mm} & h_{\text{wndw}} &= 56 \text{ mm} & d_{\text{core}} &= 35 \text{ mm} \\ w_{\text{core}} &= 13 \text{ mm} & w_{\text{wndw}} &= 15 \text{ mm}. \end{aligned} \quad (4.12)$$

Besides the parameters of the transformer cores, the winding geometry is defined by three parameters: $d_{\text{core,pri}}$ is the distance between the inner winding and the core, $d_{\text{pri,sec}}$ is the distance between the primary and secondary winding and d_y is the distance between two turns of the same winding. For the following simulation

the winding parameters are

$$d_{\text{core,pri}} = w_{\text{core,pri}} = 3.35 \text{ mm} \quad d_{\text{pri,sec}} = w_{\text{pri,sec}} = 1.5 \text{ mm} \quad d_y = 1 \text{ mm}. \quad (4.13)$$

The litz wire is made of copper and has a width of 1.2 mm and a height of 3.6 mm. Regarding the 11 turns of each winding and the insulation d_y between two turns, the height and width of the coils are

$$h_{\text{pri}} = h_{\text{sec}} = 49.6 \text{ mm} \quad w_{\text{pri}} = w_{\text{sec}} = 1.2 \text{ mm}. \quad (4.14)$$

The stray inductance is determined through a short-circuit test. Both primary and secondary carry a peak current density of 8 A/mm^2 , however, in different directions according to Lenz's law. The short-circuit test results in an RMS current of 24.4 A at 1 kHz.

The energy stored in the stray field during the short-circuit test corresponds to the stray inductance. The results of the simulation are depicted in Fig. 4.15. The color map gives the magnetic-field strength in A/m while the stream lines indicate the direction of the magnetic field. The total magnetic energy in the model is calculated to $81.5 \text{ } \mu\text{J}$. Since the model represents a quarter of the whole transformer, the total energy for a single phase of the transformer is $326 \text{ } \mu\text{J}$. This results in the stray inductance

$$L_{\sigma,3D} = \frac{2W_m}{I^2} = \frac{2 \cdot 326 \text{ } \mu\text{J}}{(24.4 \text{ A})^2} = 1.09 \text{ } \mu\text{H}. \quad (4.15)$$

This value corresponds very well to the measured stray inductance of the lab prototype. Therefore it can be concluded that the FEM software models the reality sufficiently to calculate the stray inductance already in the design stage of the transformer.

However, especially considering more complex systems, like a three-phase transformer, the computational effort increases extensively. Therefore, the same design is simulated in 2D in the following chapter. The accuracy and simulation time is compared with the 3D software subsequently.

To compare qualitatively with the 2D simulation, Fig. 4.16 shows the reference transformer in the 3D simulation from its front and side view.

2D FEM Simulation

In the following, it shall be investigated to what extend 2D simulations are suitable to determine the transformer's stray inductance. For the 2D simulation the software FEMM is used. Extruding the simulation result of the plane by the depth, the magnetic energy in this virtual arrangement can be calculated. Although this method is occasionally referred to as 2.5-D, in this work the term 2D is kept.

Figure 4.17 (a) shows the simulation result of a single phase of the transformer in the front view. Using these results and assuming the depth of the transformer, the energy amount of the stray field inside the winding window is covered. However, it is clear that also the end winding, laying outside the winding window, has a strong contribution to the stray field. Consequently, in a second step, the transformer is simulated from the side view. The corresponding result is depicted in Fig. 4.17 (b).

The simulation of the front view and the side view give a magnetic energy of

$$W_{m,f} = 127.719 \mu\text{J} \quad \text{and} \quad W_{m,s} = 93.247 \mu\text{J} \quad (4.16)$$

respectively.

Consequently, the inductance values

$$L'_\sigma = 0.4277 \mu\text{H} \quad \text{and} \quad L''_\sigma = 0.3122 \mu\text{H} \quad (4.17)$$

can be calculated as depicted in Fig. 4.14.

Figure 4.14 also illustrates that after performing the two simulations there is still a share of the stray inductance missing. This share comes from the four corners and is represented by L'''_σ . To take it into consideration, the magnetic energy per unit length is investigated first:

$$W'_{m,f} = \frac{W_{m,f}}{d_{\text{core}}} = \frac{127.719 \mu\text{J}}{35 \text{ mm}} = 3.649 \mu\text{J}/\text{mm} \quad (4.18)$$

$$W'_{m,s} = \frac{W_{m,s}}{2w_{\text{core}}} = \frac{93.247 \mu\text{J}}{26 \text{ mm}} = 3.586 \mu\text{J}/\text{mm} \quad (4.19)$$

Due to the spreading of the magnetic field, the stray inductance per unit length

in the winding windows is about 1.76 % larger than in the end winding in front of the core. This effect has also been discussed in chapter 2.6 where it results in asymmetric stray inductances.

Since the corners are also located outside the winding window it seems reasonable to use $W'_{m,s}$ to calculate the inductance contribution of the corners. Since the mean length of a single corner is

$$2 \cdot (d_{\text{core,pri}} + d_{\text{pri}} + 1/2 \cdot d_{\text{pri,sec}}), \quad (4.20)$$

it follows for the stray inductance

$$L''_{\sigma} = \frac{2 \cdot 2 \cdot (d_{\text{core,pri}} + d_{\text{pri}} + 1/2 \cdot d_{\text{pri,sec}})}{2w_{\text{core}}} L''_{\sigma} = 0.2546 \mu\text{H}. \quad (4.21)$$

Hence, the total stray inductance of this configuration is

$$L_{\sigma,2D} = L'_{\sigma} + L''_{\sigma} + L'''_{\sigma} = 0.9945 \mu\text{H}. \quad (4.22)$$

In this approach, L'''_{σ} is calculated indirectly calculating first the inductance per unit length $L'''_{\sigma}/2w_{\text{core}}$. As an alternative, the side-view simulation can be performed using the depth of

$$w_{\text{core}} + 2 \cdot 2 \cdot (d_{\text{core,pri}} + d_{\text{pri}} + 1/2 \cdot d_{\text{pri,sec}}). \quad (4.23)$$

This considers the influence of the corners directly.

Compared to the 3D simulation, there is a difference of

$$\frac{L_{\sigma,3D} - L_{\sigma,2D}}{L_{\sigma,3D}} = \frac{1.09 \mu\text{H} - 0.9945 \mu\text{H}}{1.09 \mu\text{H}} = 8.76 \%. \quad (4.24)$$

Although, the difference might already have an influence on the operation of the converter, the advantage of the 2D approach is its computation time.

The simulation time required on an Intel® Core™ i7-4600U CPU and 8 GB of system memory is shown in Table 4.3. The simulation time for the 2D arrangement is only 1 % of the 3D simulation. Therefore, the 2D approach is extremely interesting at an early design stage when many iterations are required to make the design more concrete.

Table 4.3: Simulation time

		time
3D Comsol		437 s
2D FEMM	front	3.4 s
	side	1.4 s
	total	4.8 s

4.3.2 Integrated Stray Inductance

In literature many approaches can be found to increase the stray inductance to use it as an integrated filter element. At first, these approaches are introduced with their individual advantages and disadvantages. Moreover, they are evaluated regarding their suitability for high-power medium-voltage applications.

Enlarged Winding Distance

The most intuitive approach is to simply increase the distance between the coaxial windings of the primary and secondary as also depicted in Fig. 4.12. As discussed in chapter 4.3.1, the magnetic energy $W_m = \frac{1}{2} \mu_0 \int_V H^2(x) dV$ (cf. (4.8)) is adjusted by increasing the volume of the winding and by reducing the height of the transformer.

This approach is very simple and straight forward. For first design steps, the stray inductance can be approximated by a single formula due to the simple geometry. Increasing the volume of the winding, leads of course to a higher mean length of turns and hence results in higher winding losses. Moreover, the space within the winding window is limited and with it the maximal achievable distance between the windings.

Stacked Winding

In contrary to the coaxial winding arrangement, the stacked windings are located on top of each other. It has been applied to dc-dc converters e.g. in [164]. The

stray inductance value of this winding arrangement is probably higher than for the coaxial winding arrangement. This is being evaluated below.

Stray Channel

In both mentioned approaches the space between the windings is air or rather transformer oil. Since the stray-channel does not contain any soft magnetic material, it is not able to saturate, which might be required in some applications. Moreover, there is no dependence of the stray inductance on the current.

As an alternative, soft-magnetic material might be added to the stray channel. For example, this has been proposed in [165, 166] also. As an advantage a higher stray inductance can be achieved. However, the channel might saturate which would lead to an excessive current rise and might destroy the converter ultimately. Moreover, the manufacturing of the transformer becomes more complicated.

Core and Shell Type

The different winding arrangements and stray-channel concepts can be applied to a core-type or shell-type transformer. While the core-type transformer has exterior windings, the windings of the shell-type transformer are enclosed by core material (cf. Fig. 4.19).

The shell-type transformer has a more complicated core geometry. However, the core-type transformer requires split primary and secondary windings considering a single-phase transformer.

The shell-type transformer has a lower stress under short-circuit conditions. However especially in three-phase transformers mostly core types are used due to the easier mechanical construction [167, 168].

4.3.3 Adjustment Precision and Tolerances

In the following, the presented approaches are compared for their suitability in a high-power converter. Hence, single-phase transformers are designed and compared in terms of their achievable stray inductance value and in terms of manufacturing tolerances.

In the following, four different geometries are compared. The transformer core is either a core type or a shell type. The windings are positioned on top of each other (stacked) or coaxially.

The transformers are designed for an apparent power of 2.2 MVA and a phase current of 670 A. At a frequency of 1000 Hz, the core requires a minimum area product of $A_p = 1.737 \cdot 10^{-3} \text{ m}^4$ (cf. [135]). For the following comparison, 28 turns per winding and a current density of 1 A/mm^2 are used.

The simulations are performed in 2D using FEMM and the previously explained method. Figures 4.18 to 4.21 show the simulations of the four considered transformer geometries. Each figure shows the simulation of the front and side view respectively.

As indicated by the color map, the magnetic field strength of the stacked winding is higher than for the coaxial winding. This leads to a higher inductance value as also shown later. Moreover, it can be seen that the stray field in the near environment of the transformer is larger for the stacked winding. This has to be considered also in the later construction of the transformer. Placing conductive material near the transformer might induce eddy current. However, this effect needs to be investigated further.

Comparing the transformer cores, the core-type seems to be more self-contained from the front view. From the side, core and shell type show similar characteristics.

In the following, the achievable stray inductances for each geometry are investigated. The core size and the insulation between each layer of a coil are kept constant. The insulation per layer is set to 10 mm. However, the distance between the primary and secondary coil is varied. It is set to 10 mm initially. Subsequently, it is increased step-wise until the winding-window area is utilized fully. The results of this evaluation are depicted in Fig. 4.22 showing the stray-inductance value as a function of the distance between primary and secondary winding.

First of all, it is evident that each setup covers a different range of stray-inductance values. Consequently, depending on the desired stray inductance one would choose the core-winding arrangement. As supposed before, the stacked windings achieve the highest stray-inductance values using the same transformer cores compared to the coaxial windings.

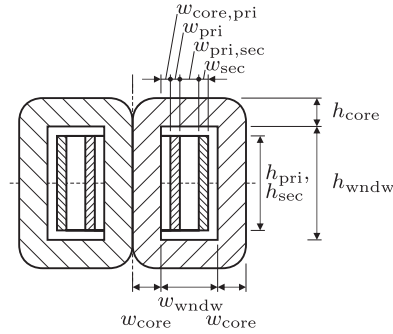
In a next step the influence of tolerances in the winding placement on the stray inductance shall be investigated. Since the distance between the primary and

Table 4.4: Inductance variation on the distance between primary and secondary

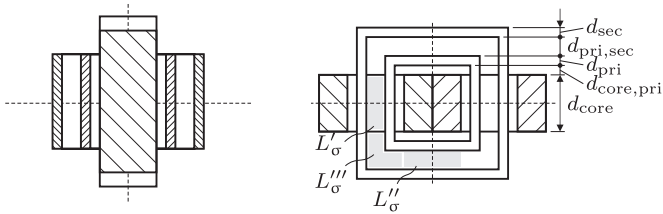
Core	Winding	Inductance variation
Shell type	stacked	1.38 $\mu\text{H}/\text{mm}$
Core type	stacked	0.87 $\mu\text{H}/\text{mm}$
Shell type	coaxial	1.47 $\mu\text{H}/\text{mm}$
Core type	coaxial	0.63 $\mu\text{H}/\text{mm}$

secondary has the highest influence on the stray inductance, the tolerances of this distance are investigated. Hence the slope of the best-fit lines of the curves in Fig. 4.22 are regarded. The values are given in Table 4.4. Consequently, the core-type transformers (0.63-0.87 $\mu\text{H}/\text{mm}$) are less sensitive to geometry variations than the shell-type transformers (1.38-1.47 $\mu\text{H}/\text{mm}$).

It is evident from the simulations however that some stray field is generated in the outer part of the transformer. This can be seen as fringing flux due to the very large air gap. Bringing magnetic material between the primary and secondary the flux can be guided and the fringing flux should be reduced. Moreover, analog to an usual inductor, a concentrated or distributed air gap can be implemented. Therefore, the stray inductance value can be set more precisely and less distance is required between the windings. The magnetic material might be brought into space between the winding locally or along the full length of the winding. The later ends up with some kind of toroidal inductor which is fully integrated into the transformer, leading probably to the design with the highest power density regarding a single phase transformer.



(a) front view



(b) side view

(c) top view

Figure 4.14: Transformer layout and the different stray-inductance contributions

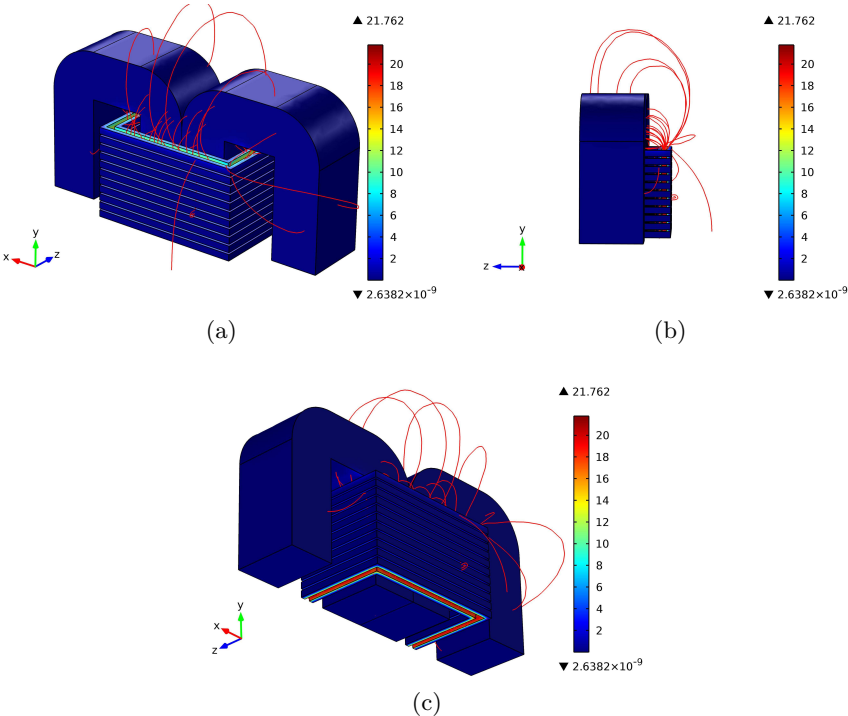


Figure 4.15: 3D FEM simulation of the reference transformer, color map showing the magnetic-field strength in A/m

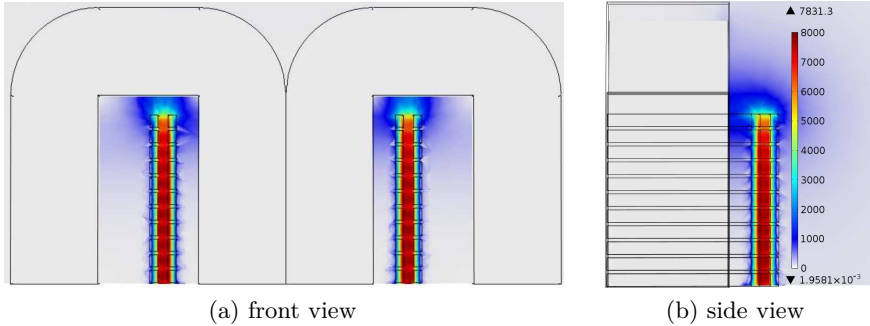


Figure 4.16: Front and side view of the 3D FEM simulation, color map showing the magnetic-field strength in A/m

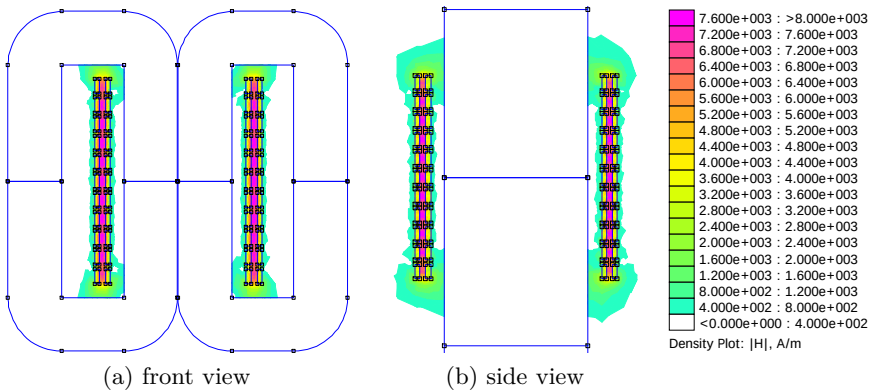


Figure 4.17: 2D simulation of the front and side view using FEMM, color map showing the magnetic-field strength in A/m

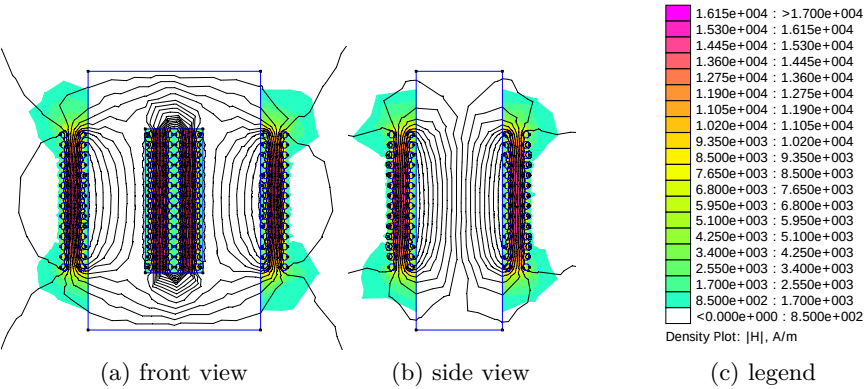


Figure 4.18: Core-type transformer with coaxial winding

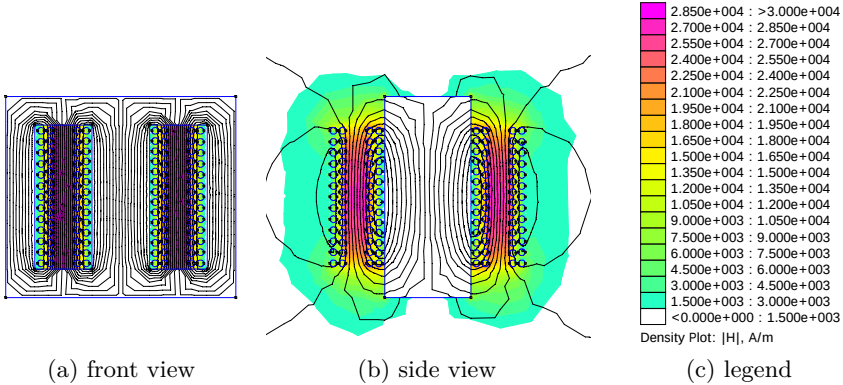


Figure 4.19: Shell-type transformer with coaxial winding

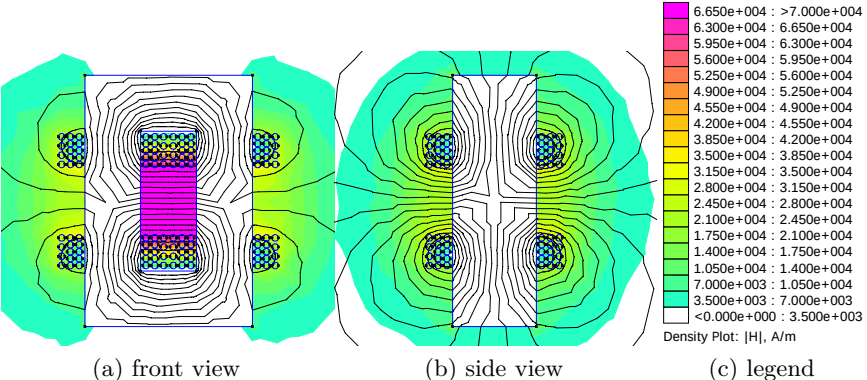


Figure 4.20: Core-type transformer with stacked winding

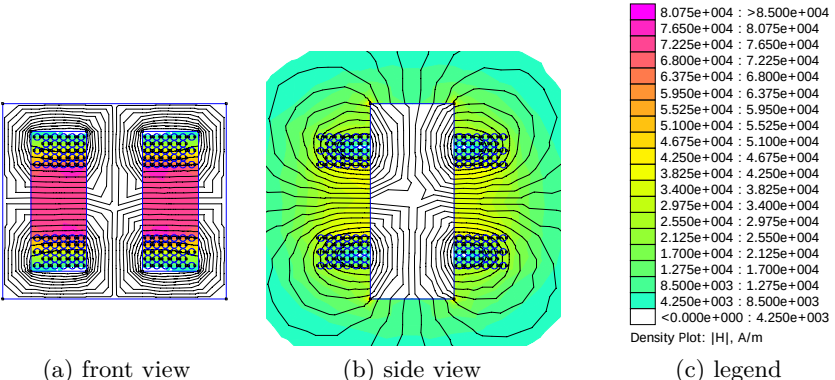


Figure 4.21: Shell-type transformer with stacked winding

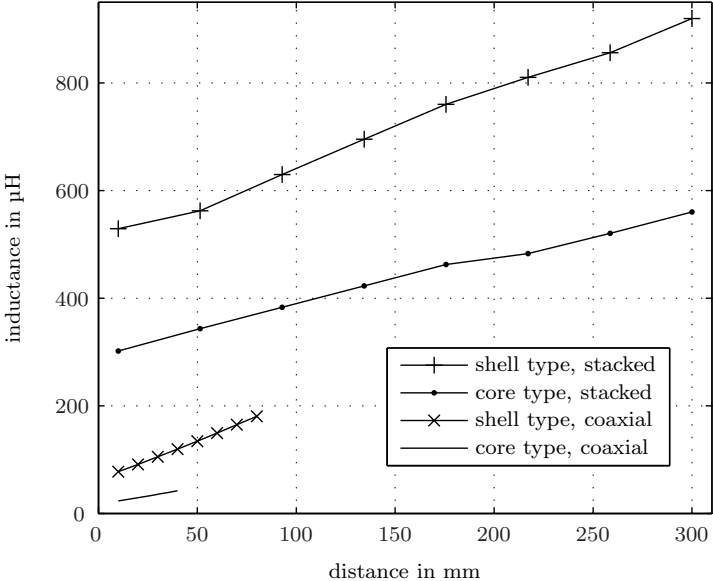


Figure 4.22: Transformer stray inductance as a function of the winding distance

4.3.4 Distinct Series Inductances

It might not be desired to use the stray inductance of the transformer as series inductance of the dc-dc converter. Reasons for it might be that the design becomes too complicated or the stray inductance can not be set precisely. Moreover, it might be an approach to design the transformer stray inductance as low as possible and, subsequently, use distinct inductances to set the series inductance precisely but also flexibly for different applications. This section investigates the use of these distinct series inductances. The additional costs and losses due to the inductances are analyzed for different core materials. Parts of this section are also published in [169].

The inductances are designed according to Hurley [135] using the parameters given in Table 4.5. For a quick estimation of the inductor size, double E cores are assumed with the optimized dimensions given in [161].

The regarded core materials are:

- Thyssen Krupp PowerCore H90 (TK): silicon steel, 180 μm lamination thickness
- JFE 10JNEX900 (10JNEX): silicon steel, 100 μm lamination thickness
- JFE 10JNHF600 (10JNHF): silicon steel, 100 μm lamination thickness
- JFE 10JNEX900 (20JNHF): silicon steel, 200 μm lamination thickness
- Hitachi 2605SA1 (H2605): amorphous iron
- Hitachi NanoCrystalline (NC): nanocrystalline

The transformers have been designed in a way that the sum of the winding and core losses become minimal [169]. Figure 4.23 shows the inductor for different core material. The silicon steels TK, 10JNHF and 20JNHF result in an impractical design since the required air gap is larger than the actual core height. However, the

Table 4.5: Basic input data for the inductor design

Objective Parameters	Value
Operation frequency	1000 Hz
RMS current	900 A
Inductance value	100 μH
Ambient temperature	40 $^{\circ}\text{C}$
Temperature rise	55 K
Heat transfer	10 W/($\text{m}^2 \text{ K}$)

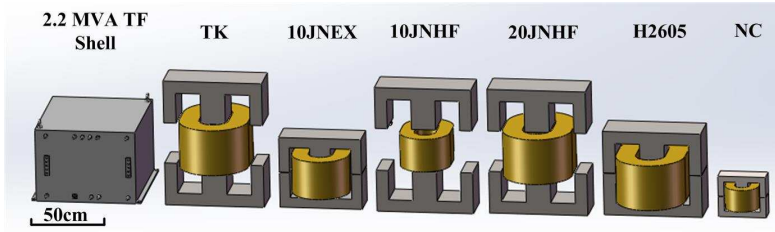


Figure 4.23: Volume comparison of inductors using different core materials in relation to the medium-frequency transformer prototype

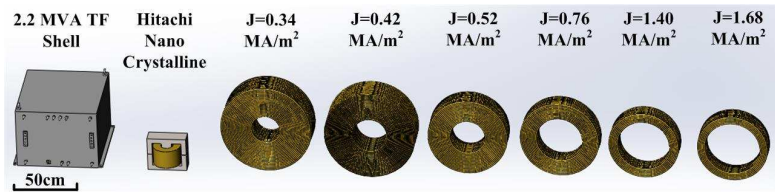


Figure 4.24: Volume comparison of air coils at different current densities in relation to the medium-frequency transformer prototype and the cored inductor using nanocrystalline material

material 10JNEX demonstrates that there are more suitable silicon steel materials for the considered application. The amorphous inductor using H2605 leads to a similar core size. The inductor from nanocrystalline material can be built very compact due to the low core loss density.

For the regarded medium-voltage high-power dc-dc converter two of those inductors would be required. In relation to the medium-frequency transformer, going for two additional inductors might be an option when the transformer design can be simplified through this approach.

In the following, it is evaluated if air coils might be an alternative solution. For different current densities, air-coil designs are depicted in Fig. 4.24. Increasing the current density reduces the volume of course. However, also the ohmic losses increase of the inductor.

The losses and the costs of the presented inductors are investigated in the following. The losses in the core and in the winding result from the inductor design directly. The total costs C take the material price C_{mat} and the costs due to the energy loss C_{el} into consideration. The costs due to energy loss consider an operation

time of 10 years at nominal power and an electricity price of 0.19€/kWh. It shall be noted that these costs are highly dependent on the later application and the according load profile. The material costs have been calculated by the masses of the winding and core and the specific material prices listed in Table 4.6.

A distinct inductor using silicon steel weights 450 kg and increases the total system weight significantly. Even though the specific costs for silicon steel material are comparably low, these inductors have the highest investment costs due to their larger volume.

Due to the lower specific core losses of amorphous and nanocrystalline materials, the corresponding inductors are more compact and require less material. Although the nanocrystalline material is 3-4 times more expensive than standard silicon steel or amorphous material, the compact design results in the lowest inductor costs finally. Moreover, the inductor made from nanocrystalline material would offer the highest efficiency and the lowest electricity costs consequently. The producibility of such large cores from nanocrystalline material, however, needs to be investigated.

Figure 4.26 compares air inductors with different current densities. Higher current density allows a more compact design while the losses in the winding increase. Consequently, the best compromise for a certain application has to be found. For high current densities the design becomes similarly compact like the nanocrystalline core. However, the efficiency is worth resulting in higher electricity costs. Therefore, air inductors are especially interesting in applications that operate often in part load conditions. Another advantage is that air inductors will not saturate and offer a constant inductance throughout the operating range. This is especially interesting regarding overload capability of the converter.

Table 4.6: Material prices used for the costs comparison

Material	Price in €/kg ^a
Copper	5.70
TK	4
10JNEX	15
10JNHF	15
20JNHF	4
H2605	3
NC	12

^a Average market prices in Germany in March 2015; prices might vary with time and purchase quantity.

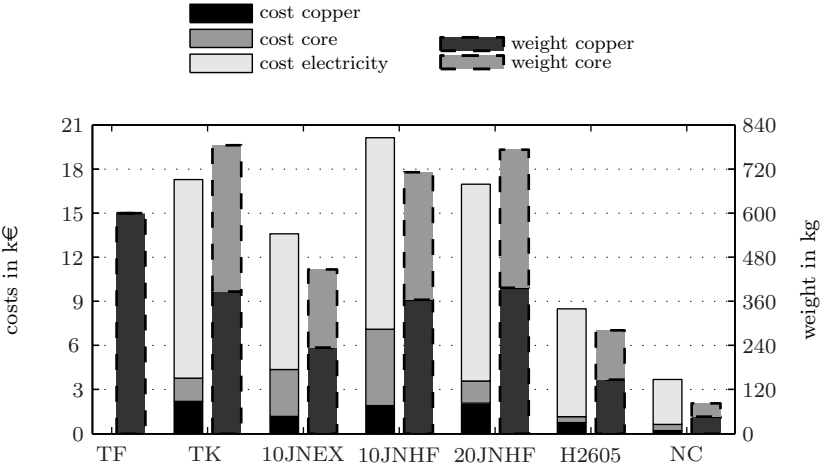


Figure 4.25: Weight and costs of the inductors using different core materials

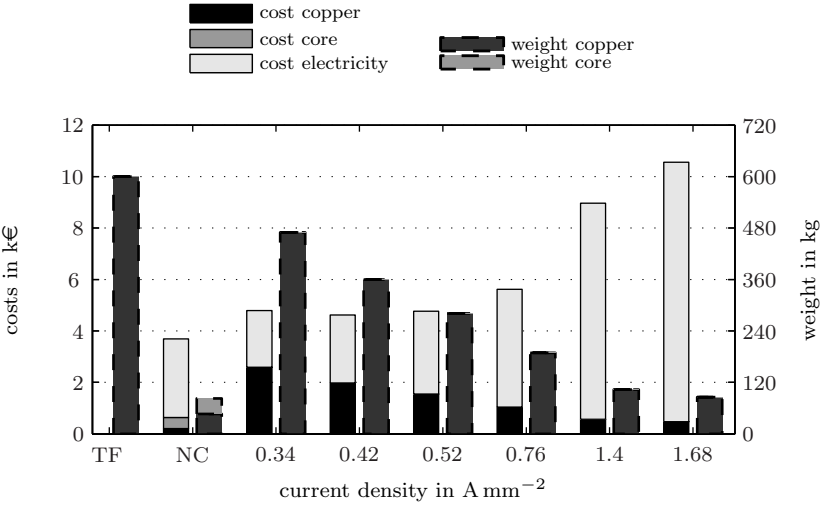


Figure 4.26: Weight and costs of the air inductors using different current densities

5 Medium-Voltage High-Power Demonstrator

The following chapter is dedicated to the construction of a full-scale megawatt medium-voltage dc-dc converter. Consequently it will give some impressions on the design and the commissioning of the converter. Parts of this chapter have been published in [12, 122].

5.1 Control Implementation

The modulation of the DAB3 is quite different from conventional PWM converters. Moreover, special control methods like the ICC and the balancing control result from this work. In the following it is demonstrated that the control of a DAB3 can be implemented on a conventional control hardware designed to operate PWM inverters usually.

The considered hardware with the model number PC D247 is from the manufacturer ABB (cf. Fig. 5.1). As depicted in Fig. 5.2, it consists of a performance optimization with enhanced RISC performance chip (PowerPC) and a field-programmable gate array (FPGA). The PowerPC is intended for tasks with a cycle time of 100-1000 μ s. The Xilinx Spartan3 FPGA is operated with a 40 MHz clock and has a time resolution of 25 ns consequently. Both devices are interconnected via a dual-ported random-access memory (DPRAM) for data exchange.

Regarding a DAB3 application, the PowerPC is responsible for the voltage and current regulation. Receiving the load angle from the PowerPC, the FPGA sets the actual switching signals for the IGBTs. The FPGA introduces, moreover, the dead time in the switching signals and implements the ICC.

In the following, the implementation is explained briefly.

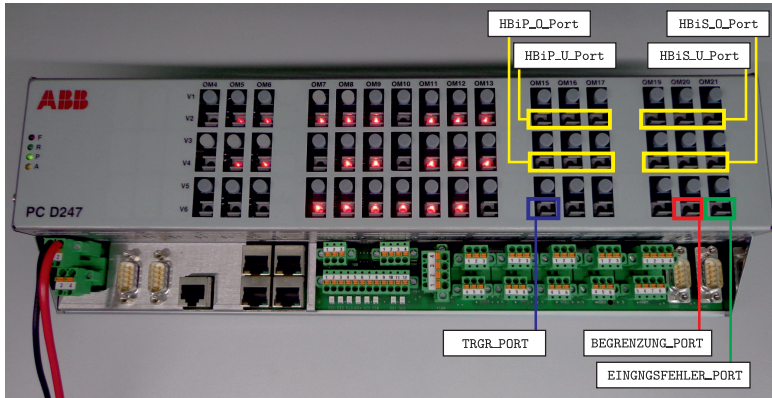


Figure 5.1: Conventional control hardware from ABB to control the DAB3

5.1.1 FPGA Implementation

To control the DAB3, six half-bridge switching signals with arbitrary phase shifts are required. Conventional control units, that are designed to drive three-phase PWM inverters, might require a modification of its FPGA firmware.

On the FPGA of the PC D247 a customizable applications block together with a framework is provided by ABB. The framework ensures the communication between FPGA, PowerPC, sensors and other interfaces. The application block has to be created by the user according to the particular application.

In this work, a modular control for a DAB3 has been created using Very High Speed Integrated Circuit Hardware Description Language (VHDL). It can be implemented on any other control hardware utilizing programmable logic devices. Moreover, due to its modularity it can be adapted and extended easily.

With the internal clock and the resolution of the counters, certain switching frequencies can be achieved. In this work, the implemented DAB3 control set allows switching frequencies in the range 868-27 770 Hz. The range was chosen regarding high-power applications and can be adapted easily.

The ICC method 2 is implemented in the control set. It is available for load angles up to 60° . The ICC is not only performed when load angles are changed, but also when the control turns on or off. Especially during an emergency off, the ICC ensures that the series inductances in the ac link of the converter are

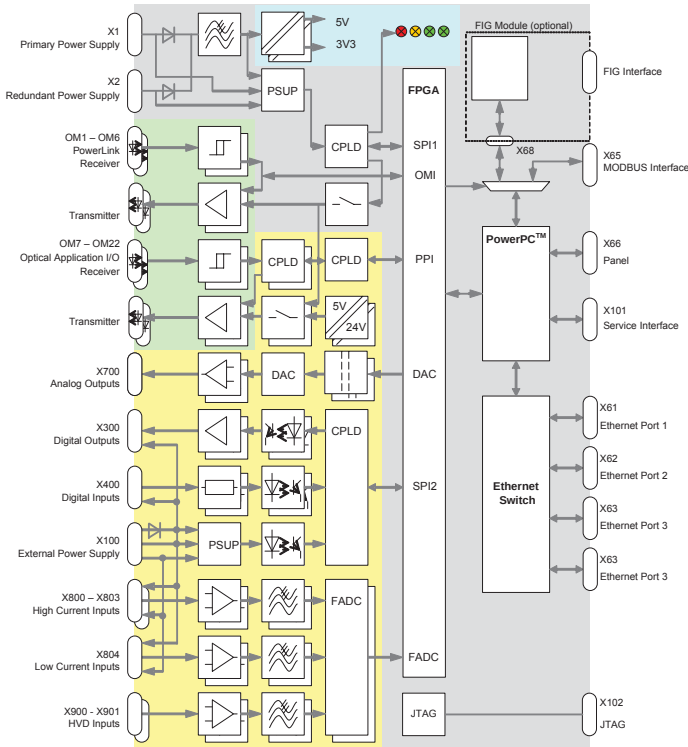


Figure 5.2: System structure of the control hardware (Source: [170])

demagnetized as quickly as possible.

Moreover, the balancing control is implemented on the controller. For each phase, offset angles from -90° to 90° can be set which add to the original load angle.

For testing purpose the features ICC and balancing control can be turned off using distinct signals.

The interface of the created control block is depicted in Fig. 5.3 (a). Written by typescript, the figure gives the names of entities, signals or components as in the original source code. On the left-hand side, signals from the PowerPC, like the load angles, emergency stop or dead time, are transferred by the DPRAM. On

the right-hand side, switching signals for the semiconductor devices are sent via an internal periphery bus and sent to the fiber optics. Figure 5.3 (b) shows the internal structure of the developed application block. It shall be noted that the original names as used in the source code are given together with the English translation. This schematic shall also demonstrate the modular nature of the code. Each block represents an instance of a VHDL component which can be adapted or modified according to future needs.

5.1.2 PowerPC Implementation

Next to the FPGA, the PowerPC is responsible for various tasks of higher cycling times: human-machine interface, current and voltage regulation or protection. The functionality of the PowerPC is implemented using MATLAB, Simulink resulting in a block set as shown in Fig. 5.4. Subsequently, the block set is translated to C++ using the MATLAB toolbox “Real-Time Workshop”. Finally, the code is transferred to the PC D247 hardware and executed on the PowerPC.

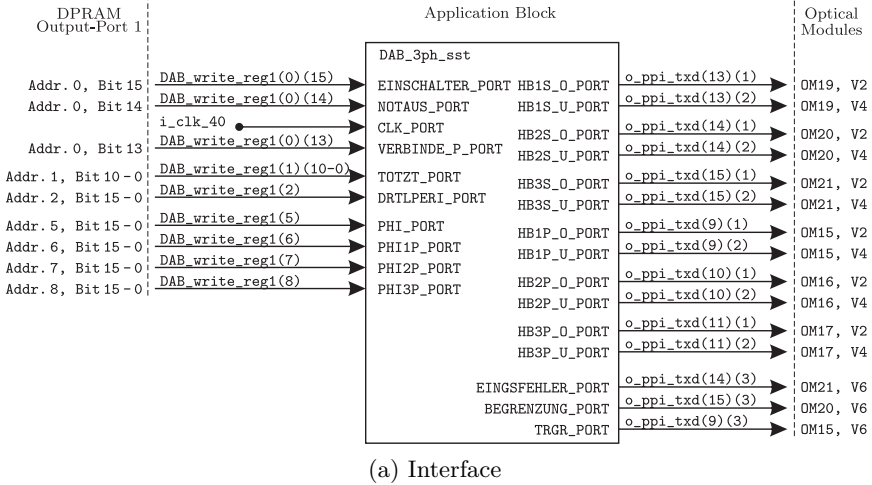
Figure 5.4 shows the implementation of the DAB3 control. Main part of the program are three interrupt service routines (ISRs) that run with different cycle times between 250 μ s and 5 ms. An interrupt control coordinates and synchronizes the calling of the ISRs.

The fastest ISR is executed with a cycle time of 250 μ s. It contains over-current and over-voltage control. Moreover it reads the external emergency stop.

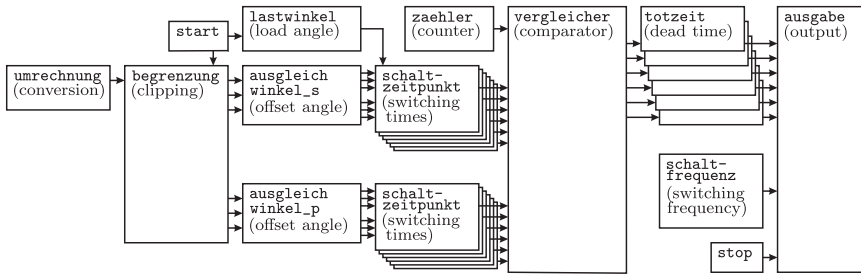
The second ISR is executed every 1 ms which corresponds to the reciprocal switching frequency of the converter. This routine for example sets the load angles and transfers them to the FPGA.

The slowest ISR reads the sensors related to the hydraulics of the water cooling like water temperature or pressure.

A fourth main task is unsynchronized and is permanently executed when no interrupt is worked off. This routine contains tasks that are not time critical like the visualization of sensor data or the input of control parameters.



(a) Interface



(b) Internal structure

Figure 5.3: Developed DAB3 control set

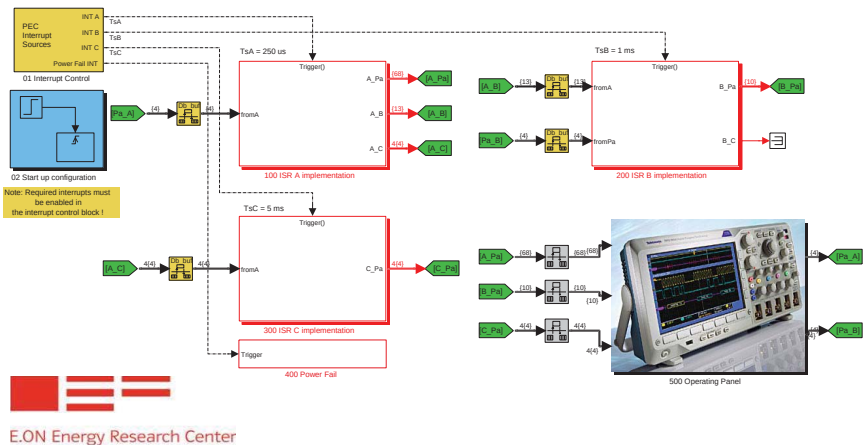


Figure 5.4: PowerPC implementation

5.2 Converter Design

In the scope of this work a dc-dc converter for medium-voltage applications and power ratings in the mega-watt range is constructed. The target specifications are primary and secondary voltage of 5 kV, while considering a voltage variation of 10 % and power of 5 MW. The fact that input and output voltage are equal helps the commissioning in the laboratory since the output is connected to the input. Therefore, the output power is recuperated and only the power loss (a small fraction of the transferred power) has to be supplied by the medium-voltage source to operate the setup. A voltage transformation, however, could be achieved easily by applying a transformer with an according winding ratio or by connection of multiple dc-dc converter in series and in parallel at the input and output respectively.

As discussed before, IGBTs are the chosen device technology due to their low forward-voltage drop off and reduced conduction losses therefore. In this project the devices “SHY 3545L0001” have been kindly provided by ABB Switzerland. The free-wheeling diodes are Infineon “D1031SH45TS02”. Both devices are rated for 2.8 kV according to 100 failure in time (FIT). Hence, two devices per converter arm have to be connected in series. In the final setup the dynamic voltage balancing is ensured by the lossless snubber as introduced in chapter 3.2. As an alternative, conventional three-level neutral-point clamped (NPC) or other multi-level converters can ensure dynamic voltage balancing while the extra level might give additional freedom for the operation of the DAB3 [171, 172].

Due to the soft-switching nature of the DAB3, the switching frequency of the IGBTs is chosen to be 1000 Hz. Regarding the loss dissipation in the semiconductor devices this is not critical. However, also other auxiliary components like the gate driver or the below-mentioned clamping circuit (usually operated at lower frequencies of 400-600 Hz) have to cope with the increasing switching frequency. Four IGBTs are integrated in a water-cooled stack called power-electronic building block (PEBB). Each PEBB represents one phase leg. Consequently, six PEBBs are used for the entire converter. Figure 5.5 shows the demonstrator during an earlier construction phase. The six PEBBs are visible as well as the water cooling auxiliary on the right hand side. The water is deionized. Since the water is very aggressive, special materials need to be used. Consequently, the pipes are made of stainless steel for example.

Each dc-link consists of two 2.27 mF capacitors that are connected in series. Compared to a hard-switched inverter with similar voltage and power rating, the dc-link capacitors in the DAB3 are very small as also demonstrated by Fig. 5.5.

This results from the three-phase nature of the converter and the comparatively high switching frequency.

5.2.1 Clamping Circuit

The converter is commissioned without applying the ARCP. Consequently, soft-switching operation can not be ensured and a clamping circuit is required as also mentioned in section 3.2.6.

The clamping circuit is needed to limit the di/dt for three reasons mainly:

- to limit the peak reverse-recovery current of the diode during turn-off,
- to ensure homogeneous firing of the IGCT during turn-on and
- to limit the diode's forward recovery voltage as well as an induced voltage across the parasitic inductance of the diode, which would appear as reverse voltage across the IGCT and might lead to a reverse avalanche.

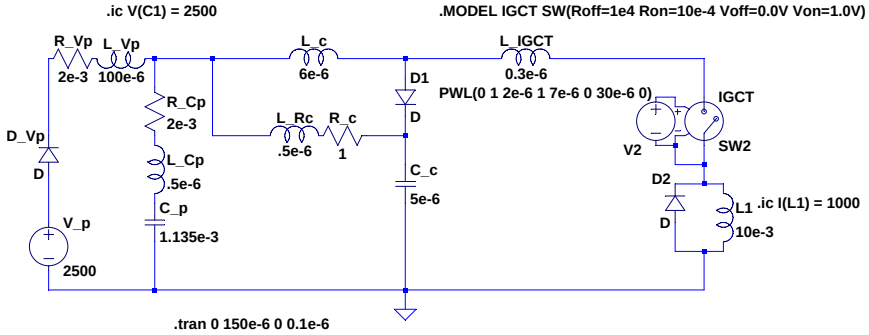
Each clamping circuit consists of a water-cooled air inductor with $L_c = 5 \mu\text{H}$. It shall be noted that these inductors sit in the main path of the dc-current. Moreover several clamping capacitors of $C_c = 5 \mu\text{F}$ are used. Each being located directly at one PEBB to ensure low-inductive connection to the IGCTs. Finally, a damping resistor R_c is required to dump the energy stored in the air inductor when the according IGCT turn-off. Especially at elevated frequency a lot of energy has to be dissipated by these resistors. Therefore they are water-cooled as well.

To design the clamping circuit, a spice simulation as shown in Fig. 5.6 is used.

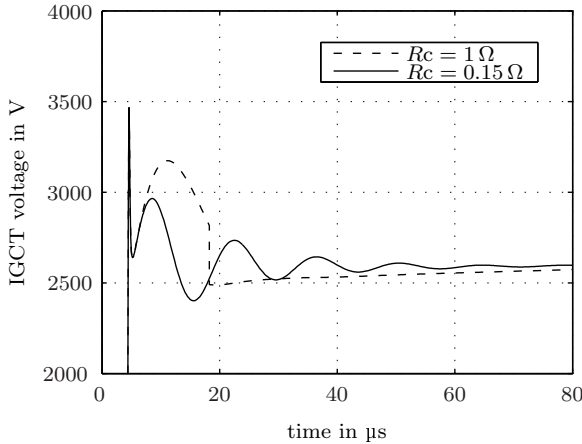


Figure 5.5: DC-DC converter during an early construction phase

The values for the parasitic inductances are partly from data sheets and partly estimations from practical experience. Regarding the mentioned values for the clamping inductors and capacitors, the optimal damping-resistor value R_c is $1\ \Omega$. Due to availability reasons, however, a lower resistance value is used. As evident from Fig. 5.6 (b) this leads to an oscillation which can also be observed in the later commissioning. It is, however, not harmful to the devices since the voltage stays below the maximum repetitive peak voltage of $4.5\ \text{kV}$.



(a) Spice model



(b) Simulation result

Figure 5.6: Spice circuit simulation of the IGCT turn-off process at $1\ \text{kA}$ to derive the optimal clamping circuit values

5.2.2 DC Link

As stated above, the primary and secondary-side dc link consists of two 2.27 mF capacitors each. Connecting them in series and applying two high-ohmic resistors for static voltage balancing, the mid-point between the capacitors can be used to connect the ARCP or the mid-point of a three-level converter. Due to the clamping circuit and the including of the clamping inductance, the dc-link has not been constructed with the aim for low inductance design as usually in voltage-source converters. The used film capacitors are DKTFM 3K302277 from MUECAP rated for 3.3 kV nominal voltage. During the commissioning using the single-phase setup, as described below, the maximal RMS current of 255 A has to be considered.

5.2.3 Auxiliary Power Supply

The IGBTs require a galvanically isolated dc voltage or ac voltage in the range of $U_{\text{GDU}} = 28 - 40 \text{ V}$ as supply voltage. The ac voltage shall be a square-wave voltage with 15-100 kHz. The power required by the gate driver to turn off the current I at the temperature T is

$$P = U_{\text{GDU}} \cdot Q_{\text{G}}(I, T) \cdot f_{\text{s}}, \quad (5.1)$$

where Q_{G} is the gate charge required to turn off the IGBT. Therefore, the power consumption rises linearly with the frequency. Turning off 1 kA at a repetition rate of 1 kHz, the gate drive unit of each IGBT requires approximately 130 W [122, 173].

5.3 Medium-Frequency Transformers

As mentioned before, the series inductance in the ac link is very important since it affects many other parameters of the converter. Therefore it can be optimized for several different optimization goals. One optimization goal might be the minimization of the ac-link RMS current for a given operating range as proposed by Lenke [2]. For this purpose, Lenke uses an iterative process to simulate several operation points at different input and output voltages. This results in very illustrative maps as given in Fig. 5.7 (a) exemplarily. These maps nicely show the influence of the series inductance on the RMS current. However, the generation process takes a lot of time since many operating points need to be simulated.

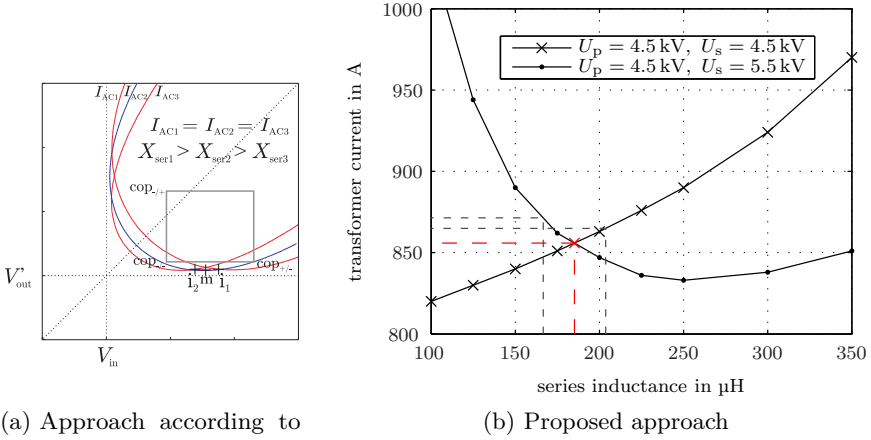


Figure 5.7: Series-inductance optimization regarding minimal RMS currents

Lenke already shows that the significant operating points are the two or three corner points of the operating range. Therefore, it is proposed in this work to limit the investigation on these three points. Instead the series inductance is varied within the simulation as shown in Fig. 5.7 (b). Subsequently, the crossing point indicates the optimal stray inductance regarding minimal RMS current, which is 185 μH for the given target specifications leading to an RMS current of 855 A.

Moreover, this method is advantageous if a certain tolerance in the series inductance has to be taken into account. Figure 5.7 (b) illustrates the current overrating which has to be implemented when the series inductance exemplarily has a manufacturing tolerance of $\pm 10\%$.

Finally, instead of simulating all series inductance values, a simple Newton-Raphson method can be used to solve this optimization problem using a small number of iterations.

Although the minimization of the RMS currents is an obvious optimization criterion, the series inductance in the ac link has many other influencing factors on the dc-dc converter. Figure 5.8 tries to name and cluster most of these factors. According to Fig. 5.8, the previous optimization method considers the “sensitivity on d (the dynamic voltage ratio) regarding ... RMS current in the ac link” which are traded for the “reactive power” in the transformer. However,

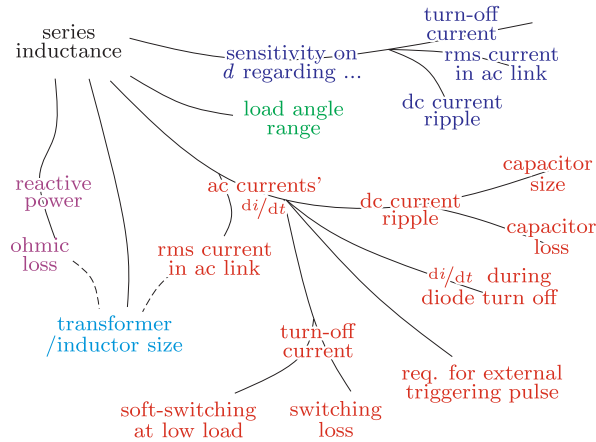


Figure 5.8: Influences of the series inductance on the converter design

it shall be demonstrated that the series inductance has many other important influencing factors, like the turn-off current of each semiconductor or the ripple in the dc currents affecting the capacitor size and losses. Moreover, the di/dt of the currents during the zero crossing is determined by the series inductance. Consequently, the semiconductor losses are influenced and at high di/dt ratings the IGCT might require a dedicated triggering pulse to ensure homogeneous firing of the semiconductor [173].

Regarding the medium-voltage demonstrator, a 1 kHz medium-frequency transformer prototype has been developed together with the German company Sangl [174]. The 5 MW DAB3 requires an apparent power of 6.6 MVA assuming a voltage variation of $\pm 10\%$. However, the transformer design comes with many challenges as also discussed in chapter 4. Since the design and conception of the transformer come with some uncertainties, three single-phase transformers, each rated for 2.2 MVA, have been setup in the scope of this work. The consecutively design, manufacturing and testing of these single-phase transformers have allowed revision and adjustment when necessary.

Especially challenging have become the medium- and high-frequency currents at the given power rating requiring HF litz wires with a significant copper cross-section area. Hence, two of the largest available litz wires have been connected in parallel. Moreover, the winding has been casted in resin to achieve an insulation,

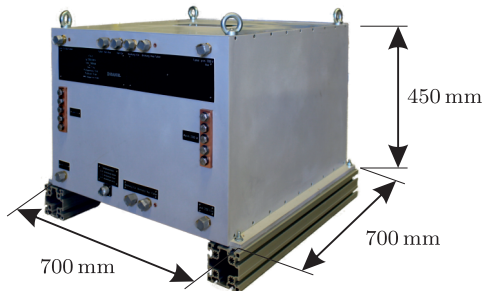


Figure 5.9: Single-Phase 2.2 MVA 1 kHz transformer

which is free from PD until 12 kV (PD measurement performed at 50 Hz). This was the aim since the effect of high-frequent PD on the aging of the transformer insulation is not understood yet as discussed in chapter 4.1. Casting the windings, the litz wire needs a special enveloping layer which allows a perfect cast without air bubbles.

The transformer core is made of silicon steel. Manufacture and handling of this material is very well known. Therefore it is cheaper than the nanocrystalline materials. Moreover magnetostriction is much lower than for the amorphous irons. Compared to a conventional silicon steel, the 180 μm thickness of each sheet is chosen to suppress eddy currents in the core sufficiently.

Figure 5.9 shows one of the three 2.2 MVA transformers. The given dimensions and the weight of 607 kg make the transformer very compact. It illustrates nicely the advantages of dc systems and the voltage conversion at higher frequencies. A 50 Hz transformer of equivalent power and voltage ratings would be approximately 10 times larger and heavier than the 1 kHz prototype (cf. Fig. 5.9). Since these transformers consist of steel and copper, valuable materials can be saved [34, 175].

5.4 Commissioning

The demonstrator is commissioned in several stages. First the converter is operated in a single-phase configuration to check the basic functionality and to verify the transformer design. Subsequently, the transformer is redesigned and evaluated again. Finally, the converter is operated in a three-phase configuration using three

single-phase transformers. The different steps are introduced in the following.

5.4.1 Single-Phase

Figure 5.10 shows the converter arrangement for the single-phase commissioning. Due to the transformer turns ratio of 1:1, the dc output can be connected to the input. Consequently, the main power stream is circulating in a loop. The dc power supply only needs to compensate the losses in order to keep the dc-link voltage constant. Hence, the power rating of the supply is fairly low and a separate dc load is not required.

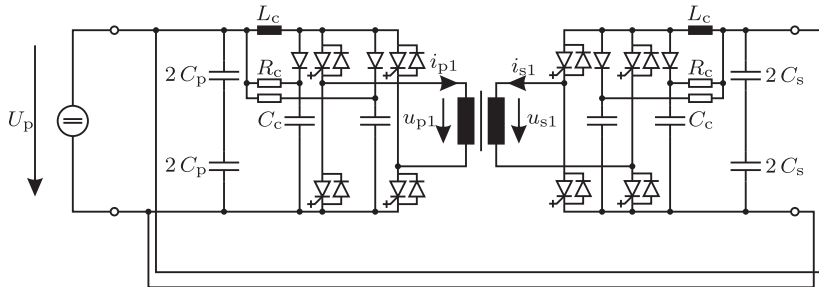


Figure 5.10: Arrangement for the single-phase commissioning

Two different measurements at different dc-link voltages are presented in Fig. 5.11 and Fig. 5.12. In general, the IGBTs do not show noticeable problems in this soft-switched converter. The measurement shows a ringing in the voltage resulting from the clamping circuit. Changing the water-cooled damping resistors, as discussed in 5.2.1, could damp the oscillation stronger. However, it is not harmful to any devices. The amplitude of the overshoot, which is independent from the dc-link voltage, is within the normal range and appears larger at lower dc-link voltages. The zero-current turn-on of the IGBTs is allowed by the transformer's stray inductance which limits the current slope during turn-on.

The transformers stray inductance, however, is lower compared to the target. According to the measurement the effective series inductance is around 45 μH . The main issues of the stray inductance has been the limited space in the winding window while achieving a proper cast of the winding. Moreover it has turned out that the cast affects the effective inductance since the stray inductance decreases significantly after the casting process. Due to a lack of measurements during manufacturing, reasons can only be assumed. The obvious reason might be that

the winding is not fixed perfectly and due to the casting the windings are displaced slightly. Another explanation might be that due to the cast the inter-winding capacitance is increased which, in turn, lowers the effective stray inductance. It demonstrates, however, that due to the tight tolerances of dc-dc converters and the higher power density more research is necessary to integrate a stable stray inductance into high-power transformers. The alternative, using separate inductors instead, is discussed in chapter 4.3.4.

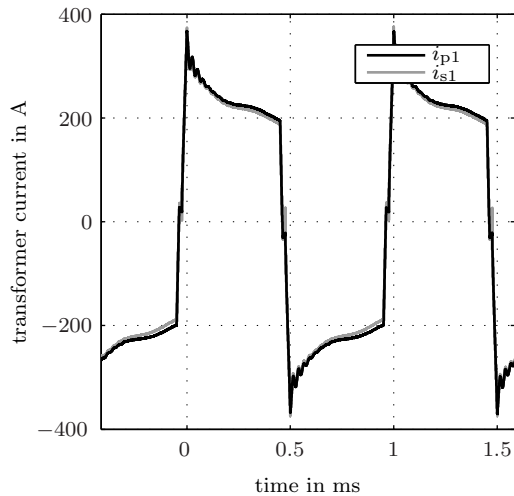
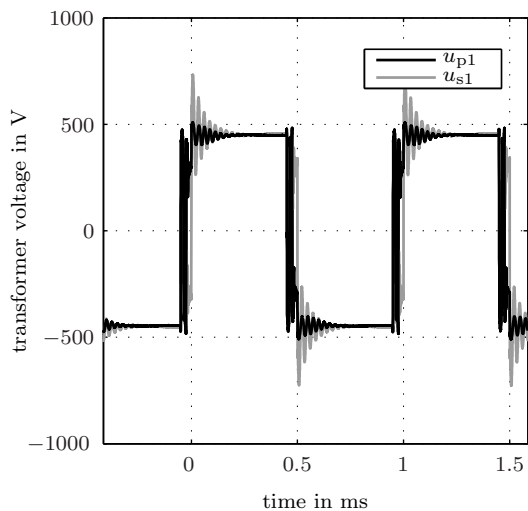
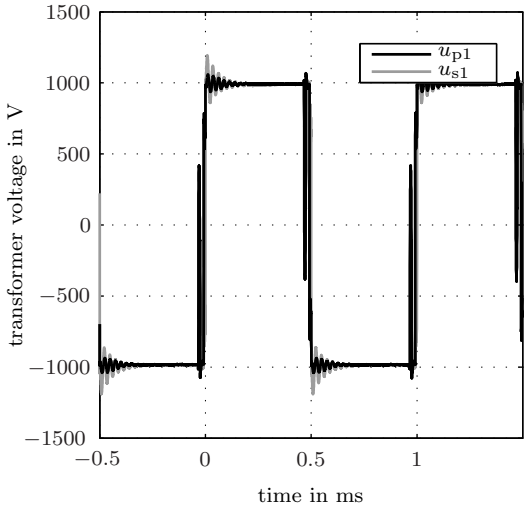
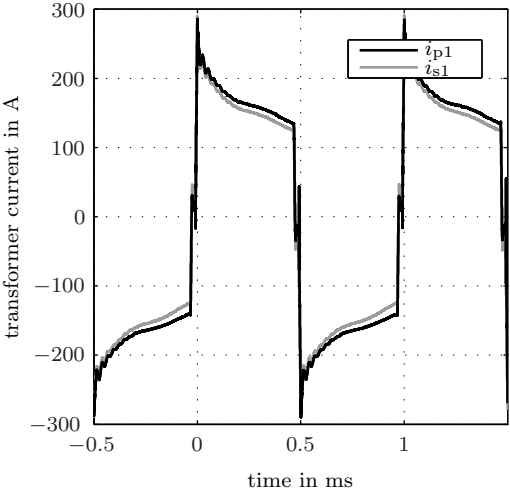


Figure 5.11: Single-phase measurement ($U_p = 450$ V)



(a) transformer voltage



(b) transformer current

Figure 5.12: Single-phase measurement ($U_p = 1000$ V)

5.4.2 Influence HF Litz

After the first experience in a converter application, the transformer is redesigned. While the first prototype was built with non-isolated litz wire, isolated litz wire, intended for high-frequency applications, is used in the second transformer design. Is it therefore referred to as HF litz in the following. To cope with the power rating of the transformer, two of the largest available HF litz wires are arranged in parallel and then casted jointly.

The stray inductances of both transformers are compared using a Hameg “HM8118” LCR meter. As shown in Fig. 5.13, the stray inductance of the transformer using HF litz tends to be higher. Although, regarding the frequency range below 10 kHz, the difference of less than 2.5 % is not significant. For higher frequencies above 20 kHz, however, the effect becomes more pronounced.

Since the stray inductance of the transformer has not varied significantly, the operation of the dc-dc converter is not affected by the HF litz as shown in Fig. 5.14 for an exemplary operating point.

The transformer losses for standard and HF litz wire are shown in Fig. 5.15 (a) and (b) respectively. The HF litz increases the efficiency of the transformer. Hence, a larger operating range can be covered for the same power consumption.

Figure 5.15 (c) shows the ratio between the losses with HF and standard litz wire. Consequently, the transformer losses are reduced by 27 % for the measured operating range by the use of HF litz. The losses reduce steadily with increasing current, while it is not affected significantly by the voltage. Since the voltage mainly affect the core losses and not the winding losses, this is reasonable.

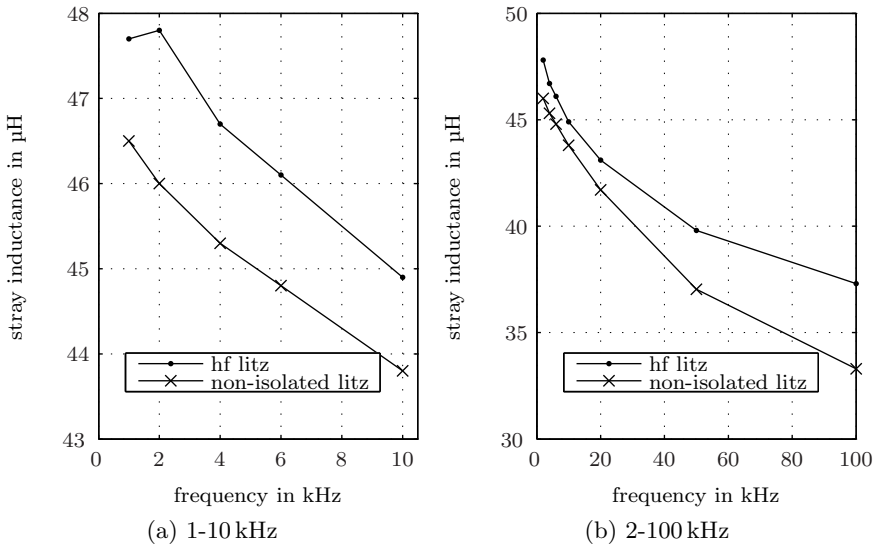


Figure 5.13: Comparison of the stray inductance with and without the use of litz wire

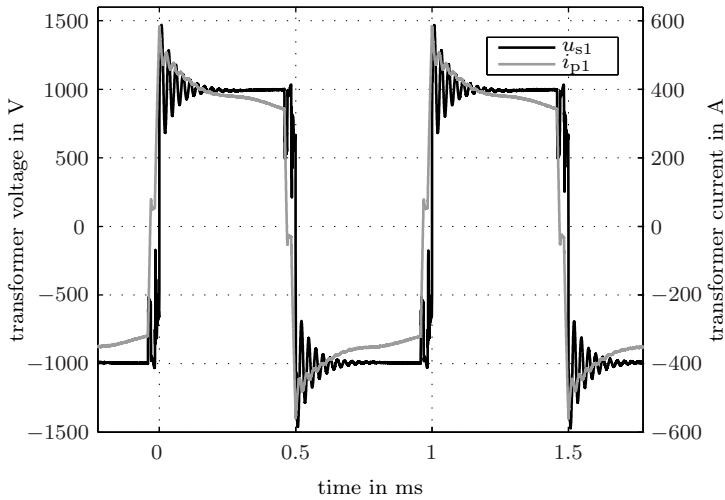
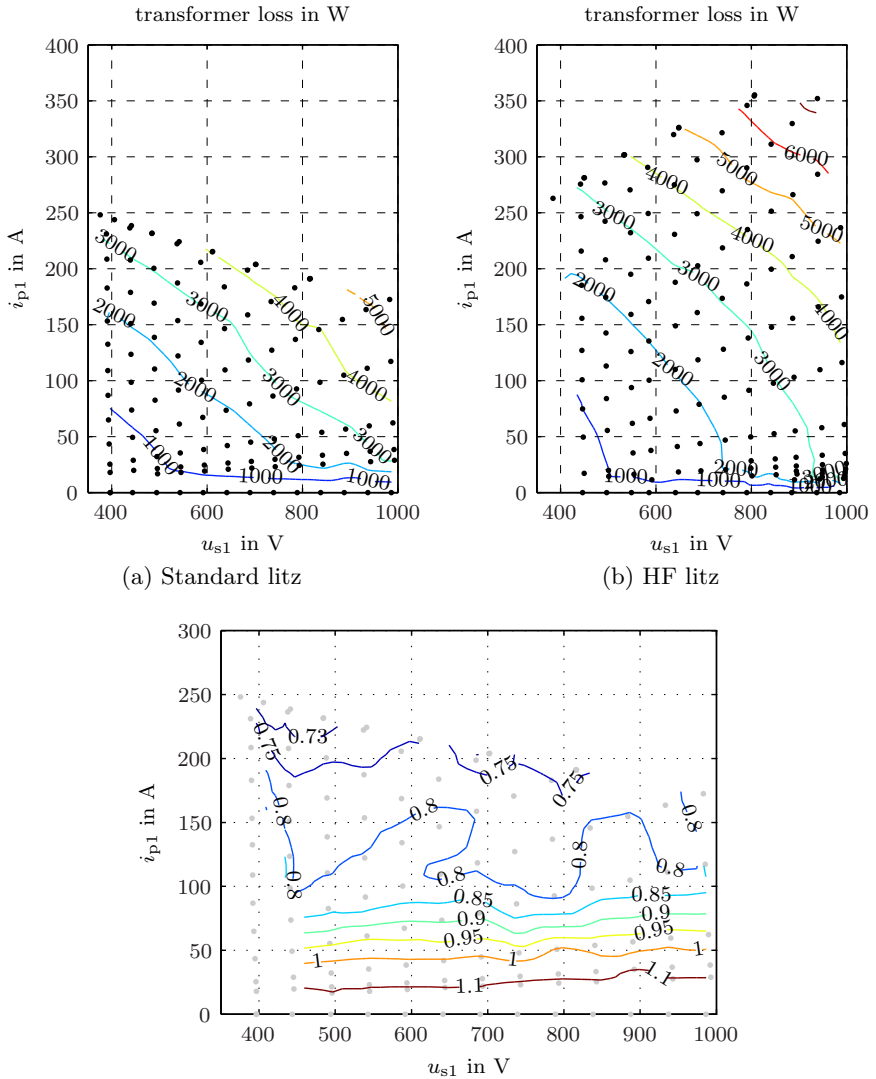


Figure 5.14: Single-phase measurement after transformer redesign using HF litz wires ($U_p = 1000 \text{ V}$)



(c) Ratio between losses with HF and standard litz wire
 Figure 5.15: Transformer losses with and without the use of litz wire

5.4.3 Three-Phase Configuration

Finally, the converter is commissioned in a three-phase configuration as depicted in Fig. 5.16. For this purpose, three single-phase transformers are used each of them having an apparent power rating of 2.2 MVA. In the setup one transformer is used with a standard wire and two transformers with HF litz wire are used. Consequently, the influence of the litz wire can be compared directly in the same setup. Moreover, the balancing control might be tested later.

The measurement is made at a dc-link voltage of $U_p = 1000$ V and with a load angle of 9° first. Figure 5.17 shows the voltage and current waveforms. There is no evidence for abnormal behavior. Also the voltages and currents in the $\alpha\beta$ plane, shown in Fig. 5.19, are as expected.

The next measurement is performed with a load angle of 15° . According to Fig. 5.18 (a) there is an asymmetry in the voltages. In the $\alpha\beta$ plane, the voltages build a symmetric hexagon as expected (cf. Fig. 5.20 (a)). Therefore, the asymmetry results from an irregular displacement of the transformers' star points as also measured in Fig. 5.18 (b). One reason for the star-point displacement lies in the fact that the transformers are different. While two are made of HF litz for example, one transformer uses non-isolated litz wire. Moreover, it is assumed that due to a slight dc offset in the transformer voltages, introduced by the power electronics, the operating points of transformer cores are shifted. This would lead to a different main inductances and would explain the irregular voltage distribution.

In contrast, the currents in the ac link shown in Fig. 5.18 (c) do not show any abnormality. A small asymmetry is observable resulting from the asymmetric transformers as well as from the asymmetric voltages. Figure 5.20 (b) shows the currents in the $\alpha\beta$ plane that are as expected as well.

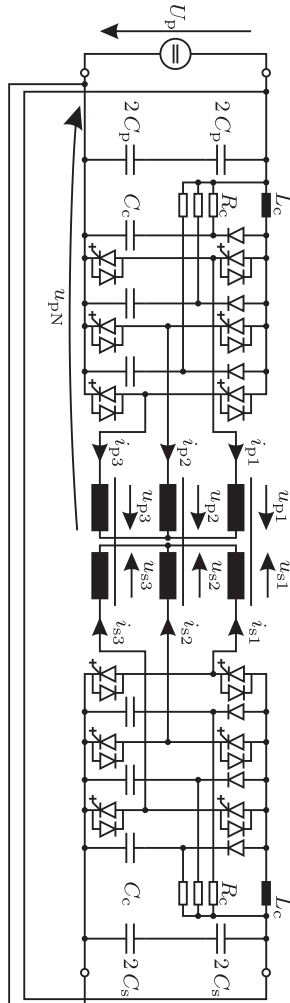
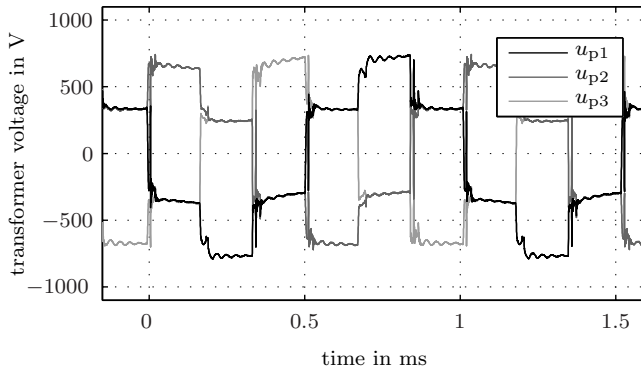
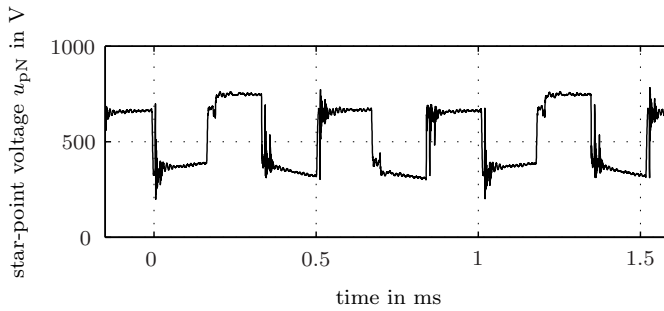


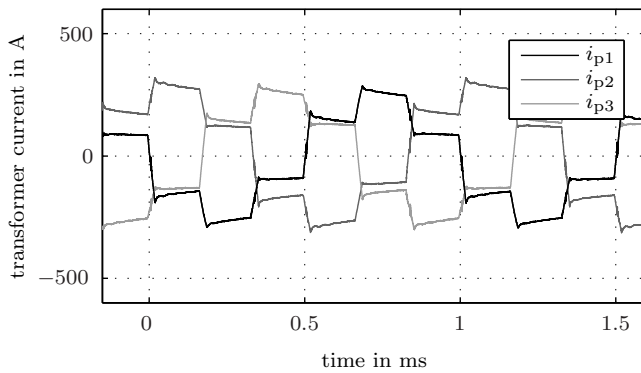
Figure 5.16: Arrangement for the three-phase commissioning



(a) primary transformer voltages

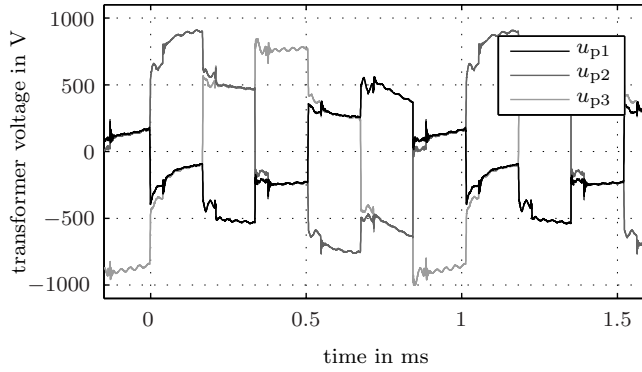


(b) transformer's primary star-point voltages

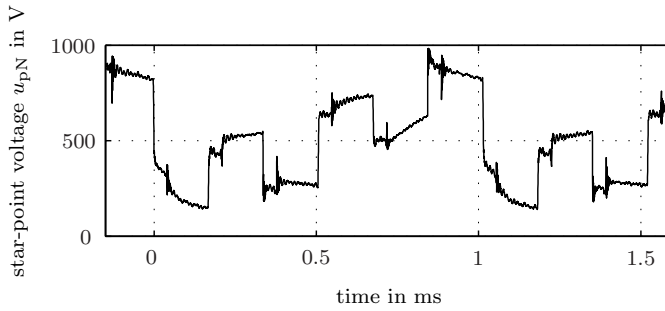


(c) transformer currents

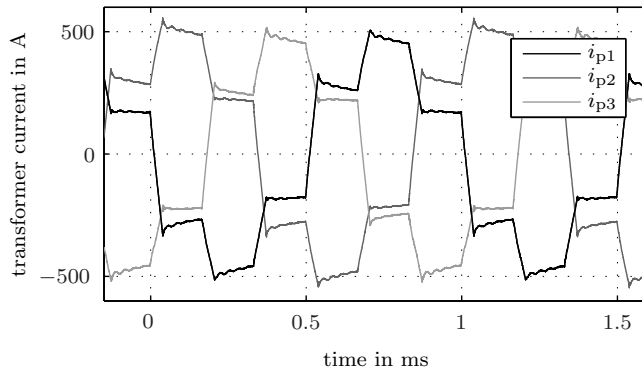
Figure 5.17: Three-phase measurement ($U_p = 1000$ V, $\varphi = 9^\circ$)



(a) primary transformer voltages

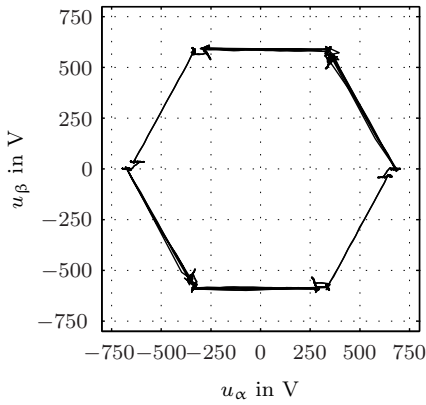


(b) transformer's primary star-point voltages

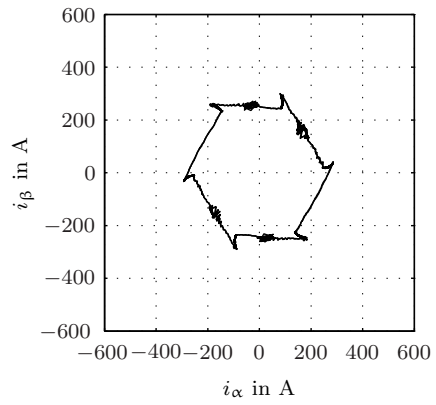


(c) transformer currents

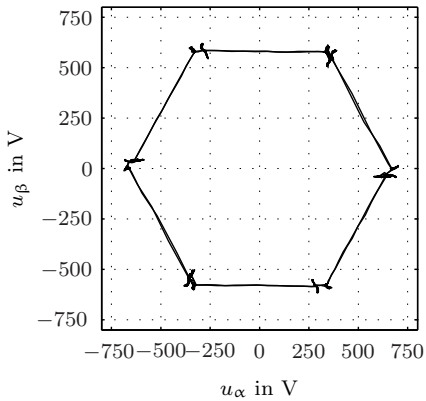
Figure 5.18: Three-phase measurement ($U_p = 1000 \text{ V}$, $\varphi = 15^\circ$)



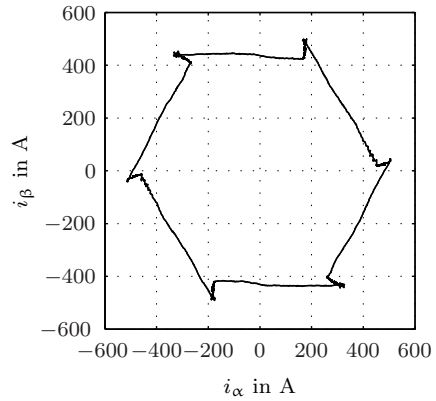
(a) primary transformer voltages



(b) transformer currents

Figure 5.19: Transformer quantities in the $\alpha\beta$ plane ($U_p = 1000$ V, $\varphi = 9^\circ$)

(a) primary transformer voltages



(b) transformer currents

Figure 5.20: Transformer quantities in the $\alpha\beta$ plane ($U_p = 1000$ V, $\varphi = 15^\circ$)

6 Conclusion and Outlook

In this thesis, high-power dc-dc conversion in the mega-watt range has been demonstrated. Due to its advantages in high-power applications, the three-phase dual-active bridge (DAB3) is chosen as most suitable converter topology. It offers galvanical isolation and achieves high efficiency through soft-switching operation. Particularly important for high-power applications is the reduced dc-link capacitor, the smaller transformer and the reduced turn-off currents compared to the single phase dual-active bridge.

Two different modeling approaches are investigated: a state-space averaging (SSA) method and a method based on first-harmonic approximation (FHA). It is found that the SSA method is able to describe the dynamic behavior of the converter accurately. This gives valuable information for the design of current and voltage controllers. The FHA underestimates the power transferred by the converter. Both voltage and currents in the DAB3 have higher harmonics which contribute to the power transfer. However, these harmonics are neglected by the FHA. Taking these harmonics into consideration might lead to a modeling approach which might be especially interesting for real-time simulation. This model might be more effective for these simulators (usually used to calculate alternating current (ac) systems) than models based on switching functions. From the modeling approach, the instantaneous current control (ICC) and the improved instantaneous current control (I^2CC) have been developed and tested on a small-scaled laboratory demonstrator. The ICC is an open-loop control which allows transition to a new operating point within one third switching period. The I^2CC has been developed for transformers with a large winding resistance and low leakage inductance. Both ICCs are enabled by setting the load angles for each phase individually. Applying this in steady-state operation, asymmetries in the transformer or between the phases in general can be compensated. This results in a balancing control which has been developed and demonstrated in this work as well.

The DAB3 is a converter which offers soft-switching in a large operating range. Particularly in light-load conditions and when the dc-link voltages vary from their nominal value, the converter might enter hard-switching operation. Lenke already proposed to use an auxiliary resonant-commutated pole (ARCP) to enable soft-switching operation in the entire operating range. This approach has been

investigated further in this work. The focus has been on the technical feasibility of the ARCP for a medium-voltage high-power dc-dc converter especially. Since the ARCP is operated at the same switching frequency like the converter, suitable thyristors had to be identified that are fast enough. Subsequently, an economical study has been performed. The additional components of the ARCP have been related to the efficiency increase of the converter for a wind application. As a result, the ARCP has a payback period of 8 to 9 month in the given example.

The transformer in the DAB3 links the two power-electronic bridges. According to the power rating of the converter, the transformer is rated for elevated voltage and current values. Furthermore, to be very compact, the transformer is operated by the power electronics with a fairly high frequency. The combination of voltage, current and frequency ratings is the major challenge in the transformer design. Moreover, the stray inductance of the transformer shall be used as an integrated filter element for the dc-dc converter. Firstly, it is investigated how the voltage waveforms of the DAB3 influence the core losses in the transformer. Consequently, the square-shaped voltage waveforms lead actually to a higher utilization of the core material, compared to a sinusoidal excitation of the material. Therefore, the core losses are slightly lower at the same peak flux density. Subsequently, different winding and core arrangements have been investigated so that the stray inductance of the transformer can serve the converter as a filter element. This study is based on 2D and 3D finite-element methods (FEMs) using conventional simulation tools. Depending on the required value for the stray inductance, different core and winding configurations are identified as basis for the more detailed transformer design. Since the stray-inductance considerations lead to additional effort in the design and also increase the losses in the transformer, separate inductances are investigated as an alternative. Both cored and air-coil inductors are investigated for this purpose. For the regarded dc-dc converter, each inductor has about the same size like the transformer itself. Saving additional costs and volume motivates to further improve the design tools of medium-frequency transformers for the use in dc-dc converters.

The final chapter is dedicated to the construction of the high-power demonstrator. Firstly, the ICC and the balancing control have been implemented on a conventional industrial control hardware. Integrated gate-commutated thyristors (IGCTs) are used as power-electronic switches. They offer lowest conduction losses, while the switching losses are reduced by the soft-switched nature of the DAB3. The IGCTs are operated with a switching frequency of 1 kHz. Due to the complexity of a three-phase medium-frequency transformer, three single-phase 2.2 MVA transformers have been designed instead. They are rated for a 5 kV 5 MW dc-dc converter. As core material, silicon steel with a very thin lamination thickness of 180 μm is used. The material is very cost effective and large core

geometries can be realized, required for medium-voltage high-power applications. Using high frequency (HF) litz wire, the winding losses can be reduced significantly at higher current ratings. The commissioning revealed as well the asymmetric displacement of the transformer's star points. Resulting from the asymmetric transformers. Moreover, it is assumed that the power electronics influence the operating point of each transformer core introducing different main inductances. This effect needs closer investigation in the future.

Bibliography

- [1] De Doncker, R. W. ‘Towards a Sustainable Energy Supply - The New Landscape of Energy Technologies’. In: *Panasonic Technical Journal* 57.4 (Jan. 2012), pp. 236–242.
- [2] Lenke, R. U. ‘A Contribution to the Design of Isolated DC-DC Converters for Utility Applications’. PhD thesis. Institute for Power Generation and Storage Systems, RWTH Aachen University, 2012.
- [3] Gaulard, L. and Gibbs, J. D. UK 4362. 1882.
- [4] Gaulard, L. and Gibbs, J. D. ‘System of Electric Distribution’. US 351589 A. 1886.
- [5] Hughes, T. P. and Hughes, T. *Networks of Power: Electrification in Western Society, 1880-1930*. Baltimore, USA: Johns Hopkins University Press, Mar. 1, 1993. 488 pp. ISBN: 978-0801846144.
- [6] Jonnes, J. *Empires of Light: Edison, Tesla, Westinghouse, and the Race to Electrify the World*. Random House Publishing Group, Dec. 10, 2004. ISBN: 9780375758843.
- [7] Callavik, M., Blomberg, A., Haefner, J., and Jacobson, B. ‘The Hybrid HVDC Breaker’. In: *ABB Grid Systems*. Ed. by ABB. 2012.
- [8] Monteiro, J., Coelho M.Costa, E., Pinto, A., Kurokawa, S., Gatous, O., and Pissolato, J. ‘Simplified skin-effect formulation for power transmission lines’. In: *Science, Measurement Technology, IET* 8.2 (Mar. 2014), pp. 47–53. ISSN: 1751-8822. DOI: 10.1049/iet-smt.2013.0072.
- [9] Hingorani, N. ‘Flexible AC transmission’. In: *Spectrum, IEEE* 30.4 (Apr. 1993), pp. 40–45. ISSN: 0018-9235. DOI: 10.1109/6.206621.
- [10] Edris, A. ‘FACTS technology development: an update’. In: *Power Engineering Review, IEEE* 20.3 (Mar. 2000), pp. 4–9. ISSN: 0272-1724. DOI: 10.1109/39.825623.
- [11] HELUKABEL GmbH. *HELUnews*. Newsletter. Nov. 2011. URL: http://www.helukabel.de/media/publication/de/cor_docs/ge_23/GE_COR-DOCS_HELUnews-5_EN.pdf.

- [12] Soltau, N., Stagge, H., De Doncker, R., and Apeldoorn, O. 'Development and demonstration of a medium-voltage high-power DC-DC converter for DC distribution systems'. In: *Power Electronics for Distributed Generation Systems (PEDG), 2014 IEEE 5th International Symposium on*. June 2014, pp. 1–8. DOI: 10.1109/PEDG.2014.6878696.
- [13] Macleod, N., Barker, C. D., and Kirby, N. 'Connection of renewable energy sources through grid constraint points using HVDC power transmission systems'. In: *Transmission and Distribution Conf. and Exposition*. Apr. 2010, pp. 1–7. DOI: 10.1109/TDC.2010.5484208.
- [14] Strokes, A. D. and Oppenlander, W. T. 'Electric arcs in open air'. In: *Journal of Physics D: Applied Physics* 24.1 (1991). DOI: 10.1088/0022-3727/24/1/006.
- [15] Das, J. 'Arc-Flash Hazard Calculations in LV and MV DC Systems - Part I: Short-Circuit Calculations'. In: *Industry Applications, IEEE Transactions on* 50.3 (May 2014), pp. 1687–1697. ISSN: 0093-9994. DOI: 10.1109/TIA.2013.2288416.
- [16] Das, J. 'Arc-Flash Hazard Calculations in LV and MV DC Systems - Part II: Analysis'. In: *Industry Applications, IEEE Transactions on* 50.3 (May 2014), pp. 1698–1705. ISSN: 0093-9994. DOI: 10.1109/TIA.2013.2288423.
- [17] Dikshit, D. K. *Handbook of Switchgears*. Tata McGraw-Hill Publishing Company Limited, 2005. Chap. 1. ISBN: 978-0-07-053238-0.
- [18] Meyer, C. 'Key Components for Future Offshore DC Grids'. PhD thesis. Aachen, Germany: Institute for Power Electronics and Electrical Drives, RWTH Aachen University, 2007.
- [19] Ganhao, Z. 'Study on DC Circuit Breaker'. In: *Intelligent Systems Design and Engineering Applications (ISDEA), 2014 Fifth International Conference on*. June 2014, pp. 942–945. DOI: 10.1109/ISDEA.2014.208.
- [20] Schmerda, R., Cuzner, R., Clark, R., Nowak, D., and Bunzel, S. 'Shipboard Solid-State Protection: Overview and Applications'. In: *Electrification Magazine, IEEE* 1.1 (Sept. 2013), pp. 32–39. ISSN: 2325-5897. DOI: 10.1109/MELE.2013.2273395.
- [21] Kempkes, M., Roth, I., and Gaudreau, M. 'Solid-state circuit breakers for Medium Voltage DC power'. In: *Electric Ship Technologies Symposium (ESTS), 2011 IEEE*. Apr. 2011, pp. 254–257. DOI: 10.1109/ESTS.2011.5770877.
- [22] Corzine, K. and Ashton, R. 'A New Z-Source DC Circuit Breaker'. In: *Power Electronics, IEEE Transactions on* 27.6 (June 2012), pp. 2796–2804. ISSN: 0885-8993. DOI: 10.1109/TPEL.2011.2178125.

- [23] Liu, L., Zhuang, J., Wang, C., Jiang, Z., Wu, J., and Chen, B. 'A Hybrid DC Vacuum Circuit Breaker for Medium Voltage: Principle and First Measurements'. In: *Power Delivery, IEEE Transactions on* PP.99 (2014), pp. 1–1. ISSN: 0885-8977. DOI: 10.1109/TPWRD.2014.2384023.
- [24] Pratt, A., Kumar, P., and Aldridge, T. 'Evaluation of 400V DC distribution in telco and data centers to improve energy efficiency'. In: *Telecommunications Energy Conference, 2007. INTELEC 2007. 29th International*. Sept. 2007, pp. 32–39. DOI: 10.1109/INTLEC.2007.4448733.
- [25] Sithimolada, V. and Sauer, P. 'Facility-level DC vs. typical ac distribution for data centers: A comparative reliability study'. In: *TENCON 2010 - 2010 IEEE Region 10 Conference*. Nov. 2010, pp. 2102–2107. DOI: 10.1109/TENCON.2010.5686625.
- [26] Emadi, A. and Ehsani, M. 'Aircraft power systems: technology, state of the art, and future trends'. In: *IEEE Aerospace and Electronic Systems Magazine* 15.1 (Jan. 2000), pp. 28–32. ISSN: 0885-8985. DOI: 10.1109/62.821660.
- [27] Izquierdo, D., Barrado, A., Raga, C., Sanz, M., and Roldan, A. 'Protection Devices for Aircraft Electrical Power Distribution Systems: State of the Art'. In: *Aerospace and Electronic Systems, IEEE Transactions on* 47.3 (July 2011), pp. 1538–1550. ISSN: 0018-9251. DOI: 10.1109/TAES.2011.5937248.
- [28] Baumann, C., Piquet, H., Roboam, X., and Bru, E. 'A Mixed Function for Actuation and Power Flow Control in Embedded Networks'. In: *IEEE Transactions on Industrial Electronics* 59.9 (Sept. 2012), pp. 3596–3603. ISSN: 0278-0046. DOI: 10.1109/TIE.2012.2188870.
- [29] Bernstein, B., Thue, W. A., Landinger, C., Hiivala, L., Medek, J. D., and Parmar, D. *Electrical Power Cable Engineering, Third Edition*. Ed. by Thue, W. A. United States of America, 2011. ISBN: 9781439856437.
- [30] Guidi, G. and Fosso, O. B. 'Investment cost of HVAC cable reactive power compensation off-shore'. In: *Energy Conference and Exhibition (ENERGYCON), 2012 IEEE International*. Sept. 2012, pp. 299–304. DOI: 10.1109/EnergyCon.2012.6347771.
- [31] Arapogianni, A., Moccia, J., Williams, D., and Phillips, J. *Wind in our Sails. The coming of Europe's offshore wind energy industry*. Report. 2011.
- [32] Arapogianni, A., Genachte, A.-B., Ochagavia, R. M., Vergara, J. P., Castell, D., Tsouroukdissian, A. R., Korbijn, J., and Bolleman, N. *Deep Water. The next step for offshore wind energy*. Report. 2013.
- [33] Ho, A., Mbistrova, A., and Corbetta, G. *The European offshore wind industry. Key trends and statistics 2015*. Ed. by Pineda, I. Report. 2015.

- [34] Stieneker, M. ‘Analysis of Medium-Voltage Direct-Current Collector Grids in Offshore Wind Parks’. PhD thesis. Institute for Power Generation and Storage Systems, RWTH Aachen University, 2017.
- [35] Deutsche Gesellschaft fuer Sonnenenergie e.V. *Energy Map*. Online. 2014. URL: <http://www.energymap.info> (visited on 01/18/2014).
- [36] Siddique, H., Ali, S., and De Doncker, R. ‘DC collector grid configurations for large photovoltaic parks’. In: *Power Electronics and Applications (EPE), 2013 15th European Conference on*. 2013, pp. 1–10. DOI: 10.1109/EPE.2013.6631799.
- [37] Dotter, G., Ackermann, F., Bihler, N., Grab, R., Rogalla, S., and Singer, R. ‘Stable operation of PV plants to achieve fault ride through capability - Evaluation in field and laboratory tests’. In: *Power Electronics for Distributed Generation Systems (PEDG), 2014 IEEE 5th International Symposium on*. June 2014, pp. 1–8. DOI: 10.1109/PEDG.2014.6878642.
- [38] McMurray, W. ‘Power converter circuits having a high frequency link’. US3517300 A. 1970.
- [39] She, X., Huang, A., and Burgos, R. ‘Review of Solid-State Transformer Technologies and Their Application in Power Distribution Systems’. In: *Emerging and Selected Topics in Power Electronics, IEEE Journal of* 1.3 (2013), pp. 186–198. ISSN: 2168-6777. DOI: 10.1109/JESTPE.2013.2277917.
- [40] Qin, H. and Kimball, J. ‘Solid-State Transformer Architecture Using AC-AC Dual-Active-Bridge Converter’. In: *Industrial Electronics, IEEE Transactions on* 60.9 (2013), pp. 3720–3730. ISSN: 0278-0046. DOI: 10.1109/TIE.2012.2204710.
- [41] Heinemann, L. ‘Analysis and design of a modular, high power converter with high efficiency for electrical power distribution systems’. In: *Power Electronics Specialists Conference, 2002. pesc 02. 2002 IEEE 33rd Annual*. Vol. 2. 2002, 713–718 vol.2. DOI: 10.1109/PSEC.2002.1022538.
- [42] Karlsson, P. ‘Dc distributed power systems’. PhD thesis. Lund University, 2002.
- [43] Bloh, J. v. ‘Multilevel-Umrichter zum Einsatz in Mittelspannungs Gleichspannungsuebertragungen’. PhD thesis. Germany: Institute for Power Electronics and Electrical Drives (ISEA), RWTH Aachen University, 2001.
- [44] Bragard, M., Soltan, N., Thomas, S., and De Doncker, R. ‘The Balance of Renewable Sources and User Demands in Grids: Power Electronics for Modular Battery Energy Storage Systems’. In: *Power Electronics, IEEE Transactions on* 25.12 (Dec. 2010), pp. 3049–3056. ISSN: 0885-8993. DOI: 10.1109/TPEL.2010.2085455.

- [45] Jiang, H. and Ekstrom, A. 'Multiterminal HVDC systems in urban areas of large cities'. In: *Power Delivery, IEEE Transactions on* 13.4 (Oct. 1998), pp. 1278–1284. ISSN: 0885-8977. DOI: 10.1109/61.714496.
- [46] Veal, C. *European Offshore Supergrid Proposol*. Ed. by Airtricity. 2006. URL: http://www.trec-uk.org.uk/resources/airtricity_supergrid_V1.4.pdf (visited on 02/17/2014).
- [47] Krogmann, C., ed. *Dream of Desertec is over [Desertec-Traum geplatzt]*. German. News. Norddeutscher Rundfunk, Oct. 14, 2014. URL: <http://www.tagesschau.de/wirtschaft/desertec-aus-101.html> (visited on 01/25/2015).
- [48] Foundation, D. *Clean power from deserts - the DESERTEC concept for energy, water and climatesecurity*. Ed. by TREC, T.-M. R. E. C. 4th ed. Protex Verlag, 2009.
- [49] Hammons, T., Lescale, V., Uecker, K., Haeusler, M., Retzmann, D., Staschus, K., and Lepy, S. 'State of the Art in Ultrahigh-Voltage Transmission'. In: *Proceedings of the IEEE* 100.2 (2012), pp. 360–390. ISSN: 0018-9219. DOI: 10.1109/JPROC.2011.2152310.
- [50] Fairley, P. 'Germany jump-starts the supergrid'. In: *Spectrum, IEEE* 50.5 (2013), pp. 36–41. ISSN: 0018-9235. DOI: 10.1109/MSPEC.2013.6511107.
- [51] 'IEEE Recommended Practice for 1 kV to 35 kV Medium-Voltage DC Power Systems on Ships'. In: *IEEE Std 1709-2010* (Nov. 2010), pp. 1–54. DOI: 10.1109/IEEESTD.2010.5623440.
- [52] Arcidiacono, V., Monti, A., and Sulligoi, G. 'Generation control system for improving design and stability of medium-voltage DC power systems on ships'. In: *Electrical Systems in Transportation, IET* 2.3 (Sept. 2012), pp. 158–167. ISSN: 2042-9738. DOI: 10.1049/iet-est.2011.0016.
- [53] Reed, G., Grainger, B., Sparacino, A., and Mao, Z.-H. 'Ship to Grid: Medium-Voltage DC Concepts in Theory and Practice'. In: *Power and Energy Magazine, IEEE* 10.6 (Nov. 2012), pp. 70–79. ISSN: 1540-7977. DOI: 10.1109/MPE.2012.2212613.
- [54] Li, H., Li, W., Luo, M., Monti, A., and Ponci, F. 'Design of Smart MVDC Power Grid Protection'. In: *Instrumentation and Measurement, IEEE Transactions on* 60.9 (Sept. 2011), pp. 3035–3046. ISSN: 0018-9456. DOI: 10.1109/TIM.2011.2158152.
- [55] Li, W., Monti, A., and Ponci, F. 'Fault Detection and Classification in Medium Voltage DC Shipboard Power Systems With Wavelets and Artificial Neural Networks'. In: *Instrumentation and Measurement, IEEE Transactions on* 63.11 (Nov. 2014), pp. 2651–2665. ISSN: 0018-9456. DOI: 10.1109/TIM.2014.2313035.

- [56] Li, W. 'Fault Detection and Protection in Medium Voltage DC Shipboard Power Systems'. PhD thesis. Institute for Automation of Complex Power Systems, RWTH Aachen University, 2013.
- [57] Steiner, M. and Reinold, H. 'Medium frequency topology in railway applications'. In: *Power Electronics and Applications, 2007 European Conference on*. Sept. 2007, pp. 1–10. DOI: 10.1109/EPE.2007.4417570.
- [58] Martin, J., Ladoux, P., Chauchat, B., Casarin, J., and Nicolau, S. 'Medium frequency transformer for railway traction: Soft switching converter with high voltage semi-conductors'. In: *Power Electronics, Electrical Drives, Automation and Motion, 2008. SPEEDAM 2008. International Symposium on*. June 2008, pp. 1180–1185. DOI: 10.1109/SPEEDHAM.2008.4581221.
- [59] Zhao, C., Dujic, D., Mester, A., Steinke, J., Weiss, M., Lewdeni-Schmid, S., Chaudhuri, T., and Stefanutti, P. 'Power Electronic Traction Transformer - Medium Voltage Prototype'. In: *Industrial Electronics, IEEE Transactions on* 61.7 (July 2014), pp. 3257–3268. ISSN: 0278-0046. DOI: 10.1109/TIE.2013.2278960.
- [60] Madhusoodhanan, S., Mainali, K., Tripathi, A., Kadavelugu, A., Patel, D., and Bhattacharya, S. 'Thermal design considerations for medium voltage power converters with 15 kV SiC IGBTs'. In: *Power Electronics for Distributed Generation Systems (PEDG), 2015 IEEE 6th International Symposium on*. June 2015, pp. 1–8. DOI: 10.1109/PEDG.2015.7223105.
- [61] Lesnicar, A., Hildinger, J., and Marquardt, R. 'Modulares Stromrichterkonzept für Netzkupplungsanwendungen bei hohen Spannungen [Modular Converter Concept for Grid Interconnection at High Voltage Levels]'. In: *ETG-Fachtagung*. 2002 (in German).
- [62] Lesnicar, A. and Marquardt, R. 'A new modular voltage source inverter topology'. In: *International Conference and Exposition on Electrical and Power Engineering (EPE), 2003*. 2003, pp. 1–10.
- [63] Marquardt, R. and Lesnicar, A. 'New Concept for High Voltage - Modular Multilevel Converter'. In: *IEEE Power Electronics Specialists Conference (PESC), 2004*. 2004.
- [64] Kenzelmann, S., Rufer, A., Dujic, D., Canales, F., and Novaes, Y. d. 'Isolated DC/DC Structure Based on Modular Multilevel Converter'. In: *Power Electronics, IEEE Transactions on* 30.1 (Jan. 2015), pp. 89–98. ISSN: 0885-8993. DOI: 10.1109/TPEL.2014.2305976.
- [65] De Doncker, R., Divan, D., and Kheraluwala, M. 'A three-phase soft-switched high-power-density DC/DC converter for high-power applications'. In: *Industry Applications, IEEE Transactions on* 27.1 (Jan. 1991), pp. 63–73. ISSN: 0093-9994. DOI: 10.1109/28.67533.

- [66] Tripathi, A., Mainali, K., Patel, D., Bhattacharya, S., and Hatua, K. 'Control and performance of a single-phase dual active half bridge converter based on 15kV SiC IGBT and 1200V SiC MOSFET'. In: *2014 Twenty-Ninth Annual IEEE Applied Power Electronics Conference and Exposition (APEC)*. Mar. 2014, pp. 2120–2125. DOI: 10.1109/APEC.2014.6803599.
- [67] Tripathi, A., Mainali, K., Patel, D., Kadavelugu, A., Hazra, S., Bhattacharya, S., and Hatua, K. 'Design considerations of a 15kV SiC IGBT enabled high-frequency isolated DC-DC converter'. In: *2014 International Power Electronics Conference (IPEC-Hiroshima 2014 - ECCE-ASIA)*. May 2014, pp. 758–765. DOI: 10.1109/IPEC.2014.6869673.
- [68] Filsoof, K., Hagar, A., and Lehn, P. 'A transformerless modular step-up dc-dc converter for high power applications'. In: *IET Power Electronics* 7.8 (Aug. 2014), pp. 2190–2199. ISSN: 1755-4535. DOI: 10.1049/iet-pel.2013.0409.
- [69] Wang, F., Yao, G., Huang, A., Song, W., and Ni, X. 'A 3.6kV high performance solid state transformer based on 13kV SiC MOSFET'. In: *Power Electronics for Distributed Generation Systems (PEDG), 2014 IEEE 5th International Symposium on*. June 2014, pp. 1–8. DOI: 10.1109/PEDG.2014.6878693.
- [70] Chen, W., Huang, A. Q., Li, C., Wang, G., and Gu, W. 'Analysis and Comparison of Medium Voltage High Power DC/DC Converters for Offshore Wind Energy Systems'. In: *IEEE Transactions on Power Electronics* 28.4 (Apr. 2013), pp. 2014–2023. ISSN: 0885-8993. DOI: 10.1109/TPEL.2012.2215054.
- [71] Matsuoka, Y., Takao, K., Wada, K., Nakahara, M., Sung, K., Ohashi, H., and Nishizawa, S. '2.5kV, 200kW bi-directional isolated DC/DC converter for medium-voltage applications'. In: *2014 International Power Electronics Conference (IPEC-Hiroshima 2014 - ECCE-ASIA)*. May 2014, pp. 744–749. DOI: 10.1109/IPEC.2014.6869671.
- [72] Dujic, D., Steinke, G., Bellini, M., Rahimo, M., Storasta, L., and Steinke, J. 'Characterization of 6.5 kV IGBTs for High-Power Medium-Frequency Soft-Switched Applications'. In: *Power Electronics, IEEE Transactions on* 29.2 (Feb. 2014), pp. 906–919. ISSN: 0885-8993. DOI: 10.1109/TPEL.2013.2259264.
- [73] Ferreira, J. 'The Multilevel Modular DC Converter'. In: *IEEE Transactions on Power Electronics* 28.10 (Oct. 2013), pp. 4460–4465. ISSN: 0885-8993. DOI: 10.1109/TPEL.2012.2237413.

- [74] Sasongko, F., Hagiwara, M., and Akagi, H. 'A front-to-front (FTF) system consisting of multiple modular multilevel cascade converters for offshore wind farms'. In: *Power Electronics Conference (IPEC-Hiroshima 2014 - ECCE-ASIA), 2014 International*. May 2014, pp. 1761–1768. DOI: 10.1109/IPEC.2014.6869822.
- [75] Kenzelmann, S., Dujic, D., Canales, F., Novaes, Y. d., and Rufer, A. 'Modular DC/DC converter: Comparison of modulation methods'. In: *Power Electronics and Motion Control Conference (EPE/PEMC), 2012 15th International*. Sept. 2012. DOI: 10.1109/EPEPMC.2012.6397405.
- [76] Jovicic, D. 'Bidirectional, High-Power DC Transformer'. In: *IEEE Transactions on Power Delivery* 24.4 (Oct. 2009), pp. 2276–2283. ISSN: 0885-8977. DOI: 10.1109/TPWRD.2009.2028600.
- [77] Soltau, N., Siddique, H., and De Doncker, R. 'Comprehensive modeling and control strategies for a three-phase dual-active bridge'. In: *Renewable Energy Research and Applications (ICRERA), 2012 International Conference on*. Nov. 2012, pp. 1–6. DOI: 10.1109/ICRERA.2012.6477408.
- [78] Engel, S. P., Soltau, N., Stagge, H., and De Doncker, R. W. 'Dynamic and Balanced Control of Three-Phase High-Power Dual-Active Bridge DC-DC Converters in DC-Grid Applications'. In: *Power Electronics, IEEE Transactions on* 28.4 (Apr. 2013), pp. 1880–1889. ISSN: 0885-8993. DOI: 10.1109/TPEL.2012.2209461.
- [79] Middlebrook, R. D. and Cuk, S. 'A General Unified Approach to Modelling Switching-Converter Power Stages'. In: *Proceedings of the IEEE Power Electronics Specialists Conference*. June 1976, pp. 73–86.
- [80] Sanders, S., Noworolski, J., Liu, X., and Verghese, G. 'Generalized averaging method for power conversion circuits'. In: *Power Electronics Specialists Conference, 1990. PESC '90 Record., 21st Annual IEEE*. June 1990, pp. 333–340. DOI: 10.1109/PESC.1990.131207.
- [81] Bai, H., Mi, C., Wang, C., and Gargies, S. 'The dynamic model and hybrid phase-shift control of a dual-active-bridge converter'. In: *Industrial Electronics, 2008. IECON 2008. 34th Annual Conference of IEEE*. Nov. 2008, pp. 2840–2845. DOI: 10.1109/IECON.2008.4758409.
- [82] Krismer, F. and Kolar, J. 'Accurate Small-Signal Model for the Digital Control of an Automotive Bidirectional Dual Active Bridge'. In: *Power Electronics, IEEE Transactions on* 24.12 (Dec. 2009), pp. 2756–2768. ISSN: 0885-8993. DOI: 10.1109/TPEL.2009.2027904.

- [83] Zhao, C., Round, S., and Kolar, J. 'Full-order averaging modelling of zero-voltage-switching phase-shift bidirectional DC-DC converters'. In: *Power Electronics, IET 3.3* (May 2010), pp. 400–410. ISSN: 1755-4535. DOI: 10.1049/iet-pel.2008.0208.
- [84] Jacobs, J., Averberg, A., and De Doncker, R. 'Multi-Phase Series Resonant DC-to-DC Converters: Stationary Investigations'. In: *Power Electronics Specialists Conference, 2005. PESC '05. IEEE 36th.* June 2005, pp. 660–666. DOI: 10.1109/PESC.2005.1581697.
- [85] Jacobs, J., Averberg, A., Schroder, S., and De Doncker, R. 'Multi-Phase Series Resonant DC-to-DC Converters: Transient Investigations'. In: *Power Electronics Specialists Conference, 2005. PESC '05. IEEE 36th.* June 2005, pp. 1972–1978. DOI: 10.1109/PESC.2005.1581902.
- [86] Jacobs, J. 'Multi-Phase Series Resonant DC-to-DC Converters'. PhD thesis. Institute for Power Electronics and Electrical Drives (ISEA), RWTH Aachen University, 2006.
- [87] *Plexim Inc.* accessed April 03, 2015. URL: <http://www.plexim.com>.
- [88] *The MathWorks Inc.* accessed April 03, 2015. URL: <http://www.mathworks.com>.
- [89] Engel, S., Soltau, N., and De Doncker, R. 'Instantaneous current control for the three-phase dual-active bridge DC-DC converter'. In: *Energy Conversion Congress and Exposition (ECCE), 2012 IEEE.* Sept. 2012, pp. 3964–3969. DOI: 10.1109/ECCE.2012.6342162.
- [90] Engel, S. and Doncker, R. W. D. 'Stromregelung fuer Gleichspannungswandler'. DE102012204035, WO2013135811. 2012.
- [91] Engel, S., Soltau, N., Stagge, H., and De Doncker, R. 'Improved instantaneous current control for the three-phase dual-active bridge DC-DC converter'. In: *ECCE Asia Downunder (ECCE Asia), 2013 IEEE.* June 2013, pp. 855–860. DOI: 10.1109/ECCE-Asia.2013.6579204.
- [92] Engel, S., Soltau, N., Stagge, H., and De Doncker, R. 'Improved Instantaneous Current Control for High-Power Three-Phase Dual-Active Bridge DC-DC Converters'. In: *IEEE Transactions on Power Electronics* 29.8 (Aug. 2014), pp. 4067–4077. ISSN: 0885-8993. DOI: 10.1109/TPEL.2013.2283868.
- [93] Clarke, E. *Analysis of A-C Power Systems. Symmetrical and Related Components.* Vol. 1. John Wiley and Sons, 1943.
- [94] Franks, R. G. and Worley, C. W. 'Quantitative Analysis of Cascade Control'. In: *Industrial and Engineering Chemistry* 48.6 (1956), pp. 1074–1079. DOI: 10.1021/ie50558a034.

- [95] Segaran, D., McGrath, B., and Holmes, D. ‘Adaptive dynamic control of a bi-directional DC-DC converter’. In: *Energy Conversion Congress and Exposition (ECCE), 2010 IEEE*. Sept. 2010, pp. 1442–1449. DOI: 10.1109/ECCE.2010.5618258.
- [96] Segaran, D., Holmes, D., and McGrath, B. ‘Enhanced Load Step Response for a Bidirectional DC-DC Converter’. In: *Power Electronics, IEEE Transactions on* 28.1 (Jan. 2013), pp. 371–379. ISSN: 0885-8993. DOI: 10.1109/TPEL.2012.2200505.
- [97] Richter, S. ‘Digitale Regelung von PWM-Wechselrichtern mit niedrigen Trägerfrequenzen [Digital Control of PWM Inverters with Low Carrier Frequency]’. (in German). PhD thesis. Institute for Power Electronics and Electrical Drives, RWTH Aachen University, 2013.
- [98] Soltau, N., Engel, S., Stagge, H., and De Doncker, R. ‘Compensation of asymmetric transformers in high-power DC-DC converters’. In: *ECCE Asia Downunder (ECCE Asia), 2013 IEEE*. June 2013, pp. 1084–1090. DOI: 10.1109/ECCE-Asia.2013.6579243.
- [99] Meeker, D. C. *Finite Element Method Magnetics*. accessed May 25, 2015. URL: <http://www.femm.info>.
- [100] Soltau, N., Lange, J., Stieneker, M., Stagge, H., and De Doncker, R. ‘Ensuring soft-switching operation of a three-phase dual-active bridge DC-DC converter applying an auxiliary resonant-commutated pole’. In: *Power Electronics and Applications (EPE’14-ECCE Europe), 2014 16th European Conference on*. Aug. 2014, pp. 1–10. DOI: 10.1109/EPE.2014.6910857.
- [101] Bernet, S., Loescher, M., and Steimer, P. K. ‘IGCTs in Soft Switching Power Converters’. In: *European Conf. on Power Electronics and Applications (EPE)*. 1999.
- [102] Fujii, K., Koellensperger, P., and De Doncker, R. ‘Characterization and Comparison of High Blocking Voltage IGBTs and IEGTs Under Hard- and Soft-Switching Conditions’. In: *Power Electronics, IEEE Transactions on* 23.1 (Jan. 2008), pp. 172–179. ISSN: 0885-8993. DOI: 10.1109/TPEL.2007.911771.
- [103] Pendharkar, S. and Shenai, K. ‘Zero voltage switching behavior of punchthrough and nonpunchthrough insulated gate bipolar transistors (IGBT’s)’. In: *IEEE Transactions on Electron Devices* 45.8 (Aug. 1998), pp. 1826–1835. ISSN: 0018-9383. DOI: 10.1109/16.704385.
- [104] Alvarez, R., Filsecker, F., Buschendorf, M., and Bernet, S. ‘Characterization of 4.5 kV/2.4 kA press pack IGBT including comparison with IGCT’. In: *Energy Conversion Congress and Exposition (ECCE), 2013 IEEE*. Sept. 2013, pp. 260–267. DOI: 10.1109/ECCE.2013.6646709.

- [105] McMurray, W. 'Selection of Snubbers and Clamps to Optimize the Design of Transistor Switching Converters'. In: *Industry Applications, IEEE Transactions on* IA-16.4 (July 1980), pp. 513–523. ISSN: 0093-9994. DOI: 10.1109/TIA.1980.4503823.
- [106] McMurray, W. 'Resonant snubbers with auxiliary switches'. In: *Industry Applications Society Annual Meeting*. 1989, 289–834 vol.1. DOI: 10.1109/IAS.1989.96608.
- [107] De Doncker, R. and Lyons, J. 'The auxiliary resonant commutated pole converter'. In: *Conf. Record of the 1990 IEEE Industry Applications Society Annual Meeting*. 1990, 1228–1235 vol.2. DOI: 10.1109/IAS.1990.152341.
- [108] McMurray, W. 'Resonant snubbers with auxiliary switches'. In: *IEEE Transactions on Industry Applications* 29.2 (1993), pp. 355–362. ISSN: 0093-9994. DOI: 10.1109/28.216544.
- [109] Rizet, C., Ferrieux, J., Le Moigne, P., Delarue, P., and Lacarney, A. 'A Simplified Resonant Pole for Three-Level Soft-Switching PFC Rectifier Used in UPS'. In: *IEEE Transactions on Industrial Electronics* 57.8 (2010), pp. 2739–2746. ISSN: 0278-0046. DOI: 10.1109/TIE.2009.2038392.
- [110] Turpin, C., Forest, F., Richardeau, F., Meynard, T., and Lacarney, A. 'Switching faults and safe control of an ARCP multicell flying capacitor inverter'. In: *IEEE Transactions on Power Electronics* 18.5 (2003), pp. 1158–1167. ISSN: 0885-8993. DOI: 10.1109/TPEL.2003.816197.
- [111] Teichmann, R. and Oyama, J. 'ARCP soft-switching technique in matrix converters'. In: *IEEE Transactions on Industrial Electronics* 49.2 (2002), pp. 353–361. ISSN: 0278-0046. DOI: 10.1109/41.993268.
- [112] Filsecker, F., Alvarez, R., and Bernet, S. 'Comparison of 6.5 kV silicon and SiC diodes'. In: *IEEE Energy Conversion Congress and Exposition (ECCE)*. 2012, pp. 2261–2267. DOI: 10.1109/ECCE.2012.6342433.
- [113] Heesen, H. t., Baron, M., Kurz, C., and Pfatischer, R. 'Definition of the euro efficiency of solar modules'. In: *Proc. of the 25th European Photovoltaic Solar Energy Conf. and Exhibition*. Valencia, Spain, 2010.
- [114] Kollensperger, P., Lenke, R., Schroder, S., and De Doncker, R. 'Design of a Flexible Control Platform for Soft-Switching Multilevel Inverters'. In: *Power Electronics, IEEE Transactions on* 22.5 (Sept. 2007), pp. 1778–1785. ISSN: 0885-8993. DOI: 10.1109/TPEL.2007.904216.
- [115] *IXYS UK Westcode Ltd.* accessed July 5, 2015. URL: <http://www.westcode.com>.

- [116] Stieneker, M., Averous, N. R., Soltau, N., Stagge, H., and Doncker, R. W. D. 'Analysis of wind turbines connected to medium-voltage dc grids'. In: *Proc. of the 16th European Conf. on Power Electronics and Applications (EPE)*. 2014.
- [117] AREVA Wind GmbH. *M5000 Technical data*. [Accessed 2015-07-12]. July 2010. URL: http://windturbine.co.kr/tt/board/images/AREVAwind_TechnicalData.pdf.
- [118] Beeken, A., Neumann, T., and Esterhellweg, A. 'Five years of operation of the first offshore wind research platform in the German Bight - FINO1'. In: *Proc. of the 9th German Wind Energy Conf. (DEWEK)*. 2008.
- [119] Tuzuner, A. and Yu, Z. 'A theoretical analysis on parameter estimation for the weibull wind speed distribution'. In: *Power and Energy Society General Meeting - Conversion and Delivery of Electrical Energy in the 21st Century, 2008 IEEE*. July 2008, pp. 1–6. DOI: 10.1109/PES.2008.4596167.
- [120] *Gesetz fuer den Vorrang Erneuerbarer Energien (Erneuerbare-Energien-Gesetz EEG)*. German Law. [Accessed 2014-05-13]. 2012. URL: http://www.bundesanzeiger-verlag.de/fileadmin/BIV-Portal/Dokumente/eeg_2012_bf.pdf.
- [121] Kollensperger, P., Bloh, J. v., Schroder, S., and De Doncker, R. 'A GCT-driver optimized for soft-switching high-power inverters with short circuit protection'. In: *Power Electronics Specialists Conference, 2004. PESC 04. 2004 IEEE 35th Annual*. Vol. 1. June 2004, 105–111 Vol.1. DOI: 10.1109/PESC.2004.1355723.
- [122] Soltau, N., Lenke, R., and De Doncker, R. W. 'High-Power DC-DC Converter'. In: *E.ON Energy Research Center Series 5.5* (2013), pp. 1–85.
- [123] Soltau, N., Eggers, D., Hameyer, K., and De Doncker, R. 'Iron Losses in a Medium-Frequency Transformer Operated in a High-Power DC-DC Converter'. In: *IEEE Transactions on Magnetics* 50.2 (Feb. 2014), pp. 953–956. ISSN: 0018-9464. DOI: 10.1109/TMAG.2013.2283733.
- [124] HVPD High Voltage Partial Discharge Limited, ed. *Introduction to Partial Discharge*. URL: <http://www.hvpd.co.uk/technical/> (visited on 07/19/2015).
- [125] Piccin, R., Mor, A., Morshuis, P., Girodet, A., and Smit, J. 'Partial discharge analysis of gas insulated systems at high voltage AC and DC'. In: *Dielectrics and Electrical Insulation, IEEE Transactions on* 22.1 (Feb. 2015), pp. 218–228. ISSN: 1070-9878. DOI: 10.1109/TDEI.2014.004711.
- [126] Devins, J. C., Rzaad, S. J., and Schwabe, R. J. 'Breakdown and prebreakdown phenomena in liquids'. In: *Journal of Applied Physics* 52 (1981). DOI: 10.1063/1.329327.

- [127] Sarathi, R. and Archana, M. 'Investigation of partial discharge activity by a conducting particle in transformer oil under harmonic AC voltages adopting UHF technique'. In: *Dielectrics and Electrical Insulation, IEEE Transactions on* 19.5 (Oct. 2012), pp. 1514–1520. ISSN: 1070-9878. DOI: 10.1109/TDEI.2012.6311495.
- [128] Sili, E., Cambronne, J., Naude, N., and Khazaka, R. 'Polyimide lifetime under partial discharge aging: effects of temperature, pressure and humidity'. In: *Dielectrics and Electrical Insulation, IEEE Transactions on* 20.2 (Apr. 2013), pp. 435–442. ISSN: 1070-9878. DOI: 10.1109/TDEI.2013.6508745.
- [129] Dissado, L. A. and Fothergill, J. C. *Degradation and Breakdown in Polymers*. Ed. by Stevens, G. C. Peter Peregrinus Ltd. on behalf of the Institution of Electrical Engineers, 1992. ISBN: 0 86341 196 7.
- [130] Crine, J.-P. and Jow, J. 'Influence of frequency on water treeing in polyethylene'. In: *Electrical Insulation and Dielectric Phenomena, 2000 Annual Report Conference on*. Vol. 1. 2000, 351–354 vol.1. DOI: 10.1109/CEIDP.2000.885298.
- [131] Bao, M., Yin, X., and He, J. 'Structure characteristics of electrical treeing in XLPE insulation under high frequencies'. In: *Physica B: Condensed Matter* 406.14 (2011), pp. 2885–2890. ISSN: 0921-4526. DOI: <http://dx.doi.org/10.1016/j.physb.2011.04.055>. URL: <http://www.sciencedirect.com/science/article/pii/S0921452611004248>.
- [132] Sarathi, R., Nandini, A., and Tanaka, T. 'Understanding electrical treeing phenomena in XLPE cable insulation under harmonic AC voltages adopting UHF technique'. In: *Dielectrics and Electrical Insulation, IEEE Transactions on* 19.3 (June 2012), pp. 903–909. ISSN: 1070-9878. DOI: 10.1109/TDEI.2012.6215093.
- [133] Saeternes, H., Aakervik, J., and Hvidsten, S. 'Water treeing in XLPE insulation at a combined DC and high frequency AC stress'. In: *Electrical Insulation Conference (EIC), 2013 IEEE*. June 2013, pp. 494–498. DOI: 10.1109/EIC.2013.6554297.
- [134] Du, B., Han, T., and Su, J. 'Effect of low temperature on tree characteristics in silicone rubber with different power frequency'. In: *Dielectrics and Electrical Insulation, IEEE Transactions on* 21.4 (Aug. 2014), pp. 1880–1886. ISSN: 1070-9878. DOI: 10.1109/TDEI.2014.004312.
- [135] Hurley, W. G. and Woelfle, W. H. *Transformers and Inductors for Power Electronics: Theory, Design and Applications*. John Wiley and Sons Ltd., 2013. ISBN: 978-1-119-95057-8.

- [136] Hitachi Metals. *Nanocrystalline soft magnetic material FINEMET*. accessed August 07, 2015. URL: <http://hilltech.com/pdf/hl-fm10-cFinemetIntro.pdf>.
- [137] Szabo, C. RE: *Daten RE: Somaloy request*. Peronal E-mail. Höganäs GmbH, 12. August 2015.
- [138] Sourmail, T. 'Near equiatomic FeCo alloys: Constitution, mechanical and magnetic properties'. In: *Progress in Materials Science* 50.7 (2005), pp. 816–880. ISSN: 0079-6425. DOI: <http://dx.doi.org/10.1016/j.pmatsci.2005.04.001>. URL: <http://www.sciencedirect.com/science/article/pii/S0079642505000204>.
- [139] Höganäs AB. *Somaloy quick guide*. Document Reference 1122HOG. Jan. 2015.
- [140] ThyssenKrupp Electrical Steel (TKES). *PowerCore H - Typical physical properties*. accessed August 07, 2015. URL: http://www.tkes.com/web2010/tkeswebcms.nsf/www/en_typische_Werkstoffeigenschaften_H.html.
- [141] JFE Steel Corporation. *Super Core - Magnetic Property Curves*. accessed August 07, 2015. URL: <http://www.jfe-steel.co.jp/en/products/electrical/catalog/f2e-001.pdf>.
- [142] JFE Steel Corporation. *Super Core - Electrical steel sheets for high-frequency application*. accessed August 07, 2015. URL: <http://www.jfe-steel.co.jp/en/products/electrical/catalog/f1e-002.pdf>.
- [143] Vacuumschmelze GmbH & Co. KG. *VITROVAC - Magnetic Properties*. accessed August 07, 2015. URL: <http://www.vacuumschmelze.com/en/products/materials-parts/soft-magnetic/amorphous-nanocrystalline/vitrovac/vitrovac-magnetic-properties.html>.
- [144] Vacuumschmelze GmbH & Co. KG. *VITROVAC 6030 F - Tape-Wound Cores in Power Transformers for Switched Mode Power Supplies*. accessed August 07, 2015. URL: <http://www.vacuumschmelze.de/fileadmin/documents/broschueren/kbbrosch/Pk003.pdf>.
- [145] Metglas, Inc. *Metglas 2605SA1 & 2605HB1M Magnetic Alloy*. accessed August 07, 2015. URL: http://metglas.com/products/magnetic_materials/2605SA1.asp.
- [146] Metals, H. and Metglas, Inc. *Amorphous Alloys for Transformer Cores*. accessed August 07, 2015. URL: <http://metglas.com/assets/pdf/2605sa1.pdf>.

- [147] Vacuumschmelze GmbH & Co. KG. *Nanocrystalline cut cores - made of vitroperm 500 for transformers*. accessed August 07, 2015. URL: http://www.vacuumschmelze.de/fileadmin/Medienbibliothek_2010/Produkte/Kerne_und_Bauelemente/Anwendungen/Kerne/Schnittbandkerne/Cut_Cores_flyer_2011.pdf.
- [148] Hitachi Metals. *Material: MB22D*. accessed August 07, 2015. URL: http://www.hitachimetals.com/product/softferrites/mn_zn_materials/PDFs/MB22D.pdf.
- [149] Asef, P. 'Design, characteristic analysis of PM wind generator based on SMC material for small direct-drive wind energy conversion system'. In: *Renewable Energy Research and Application (ICRERA), 2014 International Conference on*. Oct. 2014, pp. 41–47. DOI: 10.1109/ICRERA.2014.7016437.
- [150] GKN Sinter Metals. *Soft Magnetic Composites (SMC)*. accessed August 07, 2015. URL: <http://www.gkn.com/sintermetals/capabilities/soft-magnetic-pm/Documents/GKN%20Soft%20Magnetic%20and%20SMC%20Materials.pdf>.
- [151] Namikawa, M., Ninomiya, H., Tanaka, Y., and Takada, Y. 'Magnetic properties of 6.5% silicon steel sheets under PWM voltage excitation'. In: *Magnetics, IEEE Transactions on* 34.4 (1998), pp. 1183–1185. ISSN: 0018-9464. DOI: 10.1109/20.706469.
- [152] Sagarduy, J., Moses, A., and Anayi, F. 'Eddy Current Losses in Electrical Steels Subjected to Matrix and Classical PWM Excitation Waveforms'. In: *Magnetics, IEEE Transactions on* 42.10 (2006), pp. 2818–2820. ISSN: 0018-9464. DOI: 10.1109/TMAG.2006.879138.
- [153] Kaihara, H., Takahashi, N., Nakano, M., Kawabe, M., Nomiyama, T., Shiozaki, A., and Miyagi, D. 'Effect of Carrier Frequency and Circuit Resistance on Iron Loss of Electrical Steel Sheet Under Single-Phase Full-Bridge PWM Inverter Excitation'. In: *Magnetics, IEEE Transactions on* 48.11 (2012), pp. 3454–3457. ISSN: 0018-9464. DOI: 10.1109/TMAG.2012.2197599.
- [154] Kawabe, M., Nomiyama, T., Shiozaki, A., Kaihara, H., Takahashi, N., and Nakano, M. 'Behavior of Minor Loop and Iron Loss Under Constant Voltage Type PWM Inverter Excitation'. In: *Magnetics, IEEE Transactions on* 48.11 (2012), pp. 3458–3461. ISSN: 0018-9464. DOI: 10.1109/TMAG.2012.2198199.
- [155] Boglietti, A., Ferraris, P., Lazzari, M., and Profumo, F. 'Iron losses in magnetic materials with six-step and PWM inverter supply [induction motors]'. In: *Magnetics, IEEE Transactions on* 27 (1991). ISSN: 0018-9464. DOI: 10.1109/20.278830.

- [156] Venkatachalam, K., Sullivan, C., Abdallah, T., and Tacca, H. 'Accurate prediction of ferrite core loss with nonsinusoidal waveforms using only Steinmetz parameters'. In: *Computers in Power Electronics, 2002. Proceedings. 2002 IEEE Workshop on*. 2002, pp. 36–41. DOI: 10.1109/CIPE.2002.1196712.
- [157] Steinmetz, C. P. 'On the Law of Hysteresis (Part II.) and Other Phenomena of the Magnetic Circuit'. In: *American Institute of Electrical Engineers, Transactions of the IX.1* (1892), pp. 619–758. ISSN: 0096-3860. DOI: 10.1109/T-AIEE.1892.5570469.
- [158] Abeywickrama, K. G. N. B., Daszczynski, T., Serdyuk, Y., and Gubanski, S. 'Determination of Complex Permeability of Silicon Steel for Use in High-Frequency Modeling of Power Transformers'. In: *Magnetics, IEEE Transactions on* 44.4 (2008), pp. 438–444. ISSN: 0018-9464. DOI: 10.1109/TMAG.2007.914857.
- [159] Popescu, M., Ionel, D., Boglietti, A., Cavagnino, A., Cossar, C., and McGilp, M. 'A General Model for Estimating the Laminated Steel Losses Under PWM Voltage Supply'. In: *Industry Applications, IEEE Transactions on* 46.4 (2010), pp. 1389–1396. ISSN: 0093-9994. DOI: 10.1109/TIA.2010.2049810.
- [160] Muhlethaler, J., Biela, J., Kolar, J., and Eklebe, A. 'Improved core loss calculation for magnetic components employed in power electronic system'. In: *Applied Power Electronics Conference and Exposition (APEC), 2011 Twenty-Sixth Annual IEEE*. 2011, pp. 1729–1736. DOI: 10.1109/APEC.2011.5744829.
- [161] Mohan, N., Undeland, T. M., and Robbins, W. P. *Power Electronics: Converters, Applications, and Design*. 3rd ed. John Wiley & Sons, 2002. ISBN: 978-0471226932.
- [162] McLyman, T. *Transformer and Inductor Design Handbook*. 4th ed. CRC Press, 2011. ISBN: 978-1439836873.
- [163] COMSOL. *Official Site of COMSOL Multiphysics*. accessed August 11, 2015. URL: <https://www.comsol.de/>.
- [164] Choi, H. 'Analysis and Design of LLC Resonant Converter with Integrated Transformer'. In: *Applied Power Electronics Conference, APEC 2007 - Twenty Second Annual IEEE*. Feb. 2007, pp. 1630–1635. DOI: 10.1109/APEX.2007.357736.
- [165] Pavlovsky, M., Haan, S. d., and Ferraira, J. 'Winding Losses in High-Current, High-Frequency Transformer Foil Windings with Leakage Layer'. In: *Power Electronics Specialists Conference, 2006. PESC '06. 37th IEEE*. June 2006, pp. 1–7. DOI: 10.1109/PESC.2006.1711861.

- [166] Cougo, B. and Kolar, J. 'Integration of Leakage Inductance in Tape Wound Core Transformers for Dual Active Bridge Converters'. In: *Integrated Power Electronics Systems (CIPS), 2012 7th International Conference on*. Mar. 2012, pp. 1–6.
- [167] Charley, R. 'Recent progress in large transformers'. In: *Electrical Engineers, Journal of the Institution of* 69.418 (Oct. 1931), pp. 1189–1207. DOI: 10.1049/jiee-1.1931.0131.
- [168] Swihart, D. and McCormick, L. 'Short Circuit Vibration Analysis of a Shell Form Power Transformer'. In: *Power Apparatus and Systems, IEEE Transactions on* PAS-99.2 (Mar. 1980), pp. 800–810. ISSN: 0018-9510. DOI: 10.1109/TPAS.1980.319688.
- [169] Soltau, N., Zhan, S., and Doncker, R. W. D. 'Design of Series Inductances for High-Power DC-DC Converters'. In: *Proc. of the 4th International Conference on Renewable Energy Research and Applications (ICRERA)*. 2015.
- [170] Datasheet. *PC D247 A PEC80-INT PM*. Ed. by Ltd., A. S. 2009.
- [171] Moonem, M. A. and Krishnaswami, H. 'Analysis and control of multi-level dual active bridge DC-DC converter'. In: *Energy Conversion Congress and Exposition (ECCE), 2012 IEEE*. Sept. 2012, pp. 1556–1561. DOI: 10.1109/ECCE.2012.6342628.
- [172] Thomas, S. 'A Medium-Voltage Multi-Level DC/DC Converter with High Voltage Transformation Ratio'. PhD thesis. Lehrstuhl und Institut fuer Stromrichtertechnik und Elektrische Antriebe (ISEA), RWTH Aachen University, 2013. 226 pp.
- [173] Setz, T. and Luescher, M. *Applying IGCTs*. Application Note 5SYA 2032-03. ABB Switzerland Ltd., 2013.
- [174] *Sangl Spezialtransformatoren*. accessed November 15, 2015. URL: <http://www.sangl.de>.
- [175] Doncker, R. W. D. 'Power electronic technologies for flexible DC distribution grids'. In: *Power Electronics Conference (IPEC-Hiroshima 2014 - ECCE-ASIA), 2014 International*. May 2014, pp. 736–743. DOI: 10.1109/IPEC.2014.6869670.

A Glossary

ac	alternating current.
ARCP	auxiliary resonant-commutated pole.
Co	cobalt.
DAB	dual-active bridge.
DAB1	single-phase dual-active bridge.
DAB3	three-phase dual-active bridge.
dc	direct current.
DPRAM	dual-ported random-access memory.
DSP	digital signal processor.
EMI	electro-magnetic interference.
FACTS	flexible ac transmission system.
Fe	iron.
FEM	finite-element method.
FEMM	Finite Element Method Magnetics.
FHA	first-harmonic approximation.
FIT	failure in time.
FPGA	field-programmable gate array.
HF	high frequency.
HV	high voltage.
HVAC	high-voltage alternating current.
HVDC	high-voltage direct current.
I ² CC	improved instantaneous current control.
ICC	instantaneous current control.
IEEE	Institute of Electrical and Electronics Engineers.
IGBT	insulated-gate bipolar transistor.
IGCT	integrated gate-commutated thyristor.
iGSE	improved generalized Steinmetz equation.
ISR	interrupt service routine.

MATLAB	<u>Matrix Laboratory.</u>
MMC	modular multilevel converter.
MOSFET	metal-oxide-semiconductor field-effect transistor.
MV	medium voltage.
MVAC	medium-voltage alternating current.
MVDC	medium-voltage direct current.
NPC	neutral-point clamped.
OSE	original Steinmetz equation.
PCB	printed-circuit board.
PD	partial discharge.
PEBB	power-electronic building block.
PI	proportional-integral.
PLECS	Piecewise Linear Electrical Circuit Simulation.
PowerPC	performance optimization with enhanced RISC performance chip.
PV	photovoltaic.
PWM	pulse-width modulation.
RMS	root-mean-square.
Si	silicon.
SiC	silicon carbide.
SMC	soft-magnetic composite.
SSA	state-space averaging.
SSSC	static synchronous series compensator.
SST	solid-state transformers.
STATCOM	static synchronous compensator.
SVA	state-variable averaging.
VAR	volt-ampere reactive.
VHDL	Very High Speed Integrated Circuit Hardware Description Language.
ZC	zero current.
ZV	zero voltage.

B List of Symbols

A_e, B_e	system matrices of the even-numbered states.
A_o, B_o	system matrices of the odd-numbered states.
A_p	area product of the core-cross section area and the winding window of a transformer.
$\hat{B}$	peak flux density.
$\hat{B}_0$	peak flux density at no-load condition.
B_{sat}	saturation flux density.
C_c	Clamping circuit capacitor.
C_s	secondary dc-link capacitor.
C_{sn}	snubber capacitor.
$d_{\text{core,pri}}$	distance between core and inner winding of a transformer.
Δt_M	duration of state M .
$d_{\text{pri,sec}}$	distance between the primary and secondary winding.
d_y	distance between two turns of a winding.
f_s	switching frequency.
H_c	coercive magnetic-field strength.
i_{Cs}	current flowing into the secondary dc capacitor.
i_L	current flowing into the load.
j	imaginary unit according to $j^2 = -1$.
I_N	nominal current.
i_p	primary dc current.
i_{p1}, i_{p2}, i_{p3}	primary-side transformer currents.
$i_+, i_-, i_\times$	Command currents for the balancing control.
i_{pXM}, I_{pXM}	primary transformer current of phase x in state M .
i_s	secondary dc current.
i_{s1}, i_{s2}, i_{s3}	secondary-side transformer currents.
i_{sM}	secondary dc current in state M .
i_{sXM}, I_{sXM}	secondary transformer current of phase x in state M .
φ	load angle.
L_c	Clamping circuit inductor.

$L_h, L_{h1}, L_{h2}, L_{h3}$	Transformer's main inductance.
$L_\sigma, L_{\sigma1}, L_{\sigma2}, L_{\sigma3}$	Transformer's stray inductance.
L_s	series inductance in ac link.
M	switching state for the state-space averaging.
N_p, N_s	turns of primary and secondary transformer winding.
p_p, p_s	primary and secondary instantaneous power.
P_s	specific core losses.
Q_G	Required gate charge to turn of an IGCT.
ρ_c	resistivity of a soft-magnetic core material.
R_c	Clamping circuit resistor.
R_L	load resistor.
R_s	series resistance in ac link.
R_{sn}	snubber resistor.
S_{sxM}	switching function of the secondary bridge of phase x in state M .
α	Steinmetz parameter.
β	Steinmetz parameter.
k	Steinmetz parameter.
n	transformer's turns ratio.
U_{GDU}	Supply voltage of the gate drive unit.
u_p	primary dc voltage.
u_{p1}, u_{p2}, u_{p3}	primary-side transformer voltages.
u_{pxM}, u_{sxM}	primary and secondary transformer voltage of phase x in state M .
u_s	secondary dc voltage.
u_{s1}, u_{s2}, u_{s3}	secondary-side transformer voltages.

C List of Figures

1.1	Offshore wind facts (Source: [32, 33])	6
1.2	Typical ac offshore collector grid	7
1.3	Different topologies of a direct current (dc) collector grid	8
1.4	Share of the voltage levels on the integration of renewables in Germany in November 2014 (data source: [35])	10
1.5	Collector grid topologies for a photovoltaic (PV) application	11
1.6	Structure of a a solid-state ac transformer	12
1.7	Recent research activities on dc-dc converters	16
1.8	Different dc-dc-converter topologies	17
1.9	Modular dc-dc converters	18
2.1	Schematic of the DAB3	21
2.2	Characteristic waveforms in a DAB3	21
2.3	Boundaries of the hard and soft-switched operating areas depending on the load angle, the transferred power and the dynamic voltage conversion ratio $d = u_s/u_p$	22
2.4	Equivalent circuit diagram for the modeling approaches	23
2.5	Fundamental FHA model of a DAB3	27
2.6	Dynamic DAB3 model based on FHA	28
2.7	Comparison of the models by introducing a step on the primary dc voltage and the load angle respectively.	29
2.8	Applying the Clarke transformation [93] to the ac currents in a DAB3 results in a hexagon in the $\alpha\beta$ -plane	31
2.9	Voltage vectors in the $\alpha\beta$ -plane and generation of the current trajectory	33
2.10	Load-angle step without ICC and two different ICC methods	34
2.11	Load angles in the time domain with and without ICC	35
2.12	Comparison of a load step with and without ICC	36
2.13	Comparison between ICC and I ² CC with $\tau f_s = 1$	38
2.14	Comparison between ICC and I ² CC with $\tau f_s = 0.25$	39
2.15	Measurement without change of power-flow direction	41
2.16	Measurement with change of power-flow direction	41
2.17	Measurement of a startup	42
2.18	Comparison of ICC and I ² CC	44

2.19	Measurement of a startup	46
2.20	Stray-inductance analysis through an FEM simulation	48
2.21	Influence of an asymmetric transformer on a DAB3	49
2.22	Balancing feed-back control	52
2.23	Structure of the balancing control	53
2.24	Close-up view of the simulation given in Fig. 2.22	54
2.25	Possible sampling points indicated in phase currents and dc currents	55
2.26	Primary and secondary-side dc-voltage ripple for different voltage ratios $d = \frac{U_s}{U_p}$	57
2.27	Influence of an asymmetric transformer on a DAB3	58
3.1	Losses of different power-electronic devices in a DAB3	62
3.2	Turn-off snubbers	63
3.3	Single phase leg of an ARCP	65
3.4	Operating modes of the ARCP in a DAB3 application	66
3.5	Integration of the ARCP in one bridge of the DAB3	66
3.6	Topologies of the bidirectional switch, depicted with thyristors	67
3.7	ARCP loss calculation	69
3.8	Load profile based on the Euro Efficiency [113] for PV applications	71
3.9	Loss comparison with and without ARCP	73
3.10	Prototype verification	74
3.11	Power curve of an “Areva Multibrid M5000” wind turbine	75
3.12	Load profile resulting from the Weibull distribution in the North Sea and the power profile of an “Areva Multibrid M5000”	76
3.13	Size comparison of the ARCP with the obsolete di/dt snubber	78
3.14	Zero-voltage detection circuit	79
4.1	Treeing in silicone rubber at 50 Hz and 200 Hz (Source: [134])	82
4.2	State-of-the-art transformers and their applications	83
4.3	Current distribution due to skin effect comparing copper and aluminum	85
4.4	Current distribution due to the proximity effect comparing copper and aluminum	86
4.5	Comparison of sinusoidal flux excitation and flux density in a DAB3	91
4.6	Specific core losses in silicon steel for sinusoidal excitation	93
4.7	Specific core losses for sinusoidal and piece-wise linear course - measurement and improved generalized Steinmetz equation (iGSE)	93
4.8	Error between the measured losses and the one calculated by original Steinmetz equation (OSE) and iGSE respectively	94
4.9	Magnetic flux density for a DAB3 at $\phi = 75^\circ$ and $d = 1$	95
4.10	Specific core losses in a DAB3 under load condition calculated by the iGSE	96

4.11	Measured specific core losses in a DAB3 under load condition . . .	96
4.12	Magnetic field in the winding window	99
4.13	Three-phase transformer serving as example for the FEM comparison	100
4.14	Transformer layout and the different stray-inductance contributions	108
4.15	3D FEM simulation of the reference transformer, color map showing the magnetic-field strength in A/m	109
4.16	Front and side view of the 3D FEM simulation, color map showing the magnetic-field strength in A/m	110
4.17	2D simulation of the front and side view using Finite Element Method Magnetics (FEMM), color map showing the magnetic-field strength in A/m	110
4.18	Core-type transformer with coaxial winding	111
4.19	Shell-type transformer with coaxial winding	111
4.20	Core-type transformer with stacked winding	112
4.21	Shell-type transformer with stacked winding	112
4.22	Transformer stray inductance as a function of the winding distance	113
4.23	Volume comparison of inductors using different core materials in relation to the medium-frequency transformer prototype	115
4.24	Volume comparison of air coils at different current densities in relation to the medium-frequency transformer prototype and the cored inductor using nanocrystalline material	115
4.25	Weight and costs of the inductors using different core materials . .	117
4.26	Weight and costs of the air inductors using different current densities	117
5.1	Conventional control hardware from ABB to control the DAB3 . .	120
5.2	System structure of the control hardware (Source: [170])	121
5.3	Developed DAB3 control set	123
5.4	PowerPC implementation	124
5.5	DC-DC converter during an early construction phase	126
5.6	Spice circuit simulation of the IGCT turn-off process at 1 kA to derive the optimal clamping circuit values	127
5.7	Series-inductance optimization regarding minimal root-mean-square (RMS) currents	129
5.8	Influences of the series inductance on the converter design	130
5.9	Single-Phase 2.2 MVA 1 kHz transformer	131
5.10	Arrangement for the single-phase commissioning	132
5.11	Single-phase measurement ($U_p = 450$ V)	134
5.12	Single-phase measurement ($U_p = 1000$ V)	135
5.13	Comparison of the stray inductance with and without the use of litz wire	137
5.14	Single-phase measurement after transformer redesign using HF litz wires ($U_p = 1000$ V)	137

5.15	Transformer losses with and without the use of litz wire	138
5.16	Arrangement for the three-phase commissioning	140
5.17	Three-phase measurement ($U_p = 1000\text{ V}$, $\varphi = 9^\circ$)	141
5.18	Three-phase measurement ($U_p = 1000\text{ V}$, $\varphi = 15^\circ$)	142
5.19	Transformer quantities in the $\alpha\beta$ plane ($U_p = 1000\text{ V}$, $\varphi = 9^\circ$) . . .	143
5.20	Transformer quantities in the $\alpha\beta$ plane ($U_p = 1000\text{ V}$, $\varphi = 15^\circ$) . .	143

D List of Tables

1.1	Overview of published research works on medium voltage (MV) dc-dc converters	15
2.1	Transformer voltages and secondary switching function for states M according to Fig. 2.2	25
3.1	Simulation parameters	60
3.2	Semiconductor devices	60
3.3	Compared semiconductor devices	67
3.4	Input parameters	68
3.5	Saved energy by the ARCP (per phase, averaged over entire load profile)	71
4.1	Soft-magnetic materials	89
4.2	Measured specific core losses	90
4.3	Simulation time	104
4.4	Inductance variation on the distance between primary and secondary	107
4.5	Basic input data for the inductor design	114
4.6	Material prices used for the costs comparison	116

CURRICULUM VITAE

NILS-JÖRGEN SOLTAU

PERSONAL INFORMATION

Date of birth	March 26 th , 1985
Place of birth	Lippstadt, Germany
Nationality	German

EDUCATION

since 02/2010	PhD in Electrical Engineering and Information Technology, RWTH Aachen University, Supervisor: Prof. Rik W. De Doncker, Co-Supervisor: Prof. Antonello Monti
10/2004–01/2010	Studies in Electrical Engineering and Information Technology, RWTH Aachen University
10/1995–09/2004	University-Entrance Diploma, Marienschule Lippstadt, Germany

WORK EXPERIENCE

since 04/2017	Application Engineering Semiconductors, Mitsubishi Electric Europe, Ratingen, Germany
10/2014–03/2017	Chief Engineer at the Institute for Power Generation and Storage Systems of RWTH Aachen University
02/2010–09/2014	Research Associate at the Institute for Power Generation and Storage Systems of RWTH Aachen University

E.ON ERC Band 1**Streblow, R.**

Thermal Sensation and
Comfort Model for
Inhomogeneous Indoor
Environments

1. Auflage 2011

ISBN 978-3-942789-00-4

E.ON ERC Band 2**Naderi, A.**

Multi-phase, multi-species
reactive transport modeling as
a tool for system analysis in
geological carbon dioxide
storage

1. Auflage 2011

ISBN 978-3-942789-01-1

E.ON ERC Band 3**Westner, G.**

Four Essays related to Energy
Economic Aspects of
Combined Heat and Power
Generation

1. Auflage 2012

ISBN 978-3-942789-02-8

E.ON ERC Band 4**Lohwasser, R.**

Impact of Carbon Capture and
Storage (CCS) on the European
Electricity Market

1. Auflage 2012

ISBN 978-3-942789-03-5

E.ON ERC Band 5**Dick, C.**

Multi-Resonant Converters as
Photovoltaic Module-
Integrated Maximum Power
Point Tracker

1. Auflage 2012

ISBN 978-3-942789-04-2

E.ON ERC Band 6**Lenke, R.**

A Contribution to the Design of
Isolated DC-DC Converters for
Utility Applications

1. Auflage 2012

ISBN 978-3-942789-05-9

E.ON ERC Band 7**Brännström, F.**

Einsatz hybrider RANS-LES-
Turbulenzmodelle in der
Fahrzeugklimatisierung

1. Auflage 2012

ISBN 978-3-942789-06-6

E.ON ERC Band 8**Bragard, M.**

The Integrated Emitter Turn-
Off Thyristor - An Innovative
MOS-Gated High-Power
Device

1. Auflage 2012

ISBN 978-3-942789-07-3

E.ON ERC Band 9**Hoh, A.**

Energiebasierte Bewertung
gebäudetechnischer Anlagen

1. Auflage 2013

ISBN 978-3-942789-08-0

E.ON ERC Band 10**Köllensperger, P.**

The Internally Commutated
Thyristor - Concept, Design
and Application

1. Auflage 2013

ISBN 978-3-942789-09-7

E.ON ERC Band 11**Achtnicht, M.**

Essays on Consumer Choices
Relevant to Climate Change:
Stated Preference Evidence
from Germany

1. Auflage 2013

ISBN 978-3-942789-10-3

E.ON ERC Band 12**Panašková, J.**

Olfaktorische Bewertung von
Emissionen aus Bauprodukten

1. Auflage 2013

ISBN 978-3-942789-11-0

E.ON ERC Band 13**Vogt, C.**

Optimization of Geothermal
Energy Reservoir Modeling
using Advanced Numerical
Tools for Stochastic Parameter
Estimation and Quantifying
Uncertainties

1. Auflage 2013

ISBN 978-3-942789-12-7

E.ON ERC Band 14**Benigni, A.**

Latency exploitation for
parallelization of
power systems simulation

1. Auflage 2013

ISBN 978-3-942789-13-4

E.ON ERC Band 15**Butschen, T.**

Dual-ICT – A Clever Way to
Unite Conduction and
Switching Optimized
Properties in a Single Wafer

1. Auflage 2013

ISBN 978-3-942789-14-1

E.ON ERC Band 16**Li, W.**

Fault Detection and
Protection in Medium
Voltage DC Shipboard
Power Systems

1. Auflage 2013

ISBN 978-3-942789-15-8

E.ON ERC Band 17**Shen, J.**

Modeling Methodologies for
Analysis and Synthesis of
Controls and Modulation
Schemes for High-Power
Converters with Low Pulse
Ratios

1. Auflage 2014

ISBN 978-3-942789-16-5

E.ON ERC Band 18**Flieger, B.**

Innenraummodellierung einer Fahrzeugkabine in der Programmiersprache Modelica

1. Auflage 2014

ISBN 978-3-942789-17-2

E.ON ERC Band 19**Liu, J.**

Measurement System and Technique for Future Active Distribution Grids

1. Auflage 2014

ISBN 978-3-942789-18-9

E.ON ERC Band 20**Kandzia, C.**

Experimentelle Untersuchung der Strömungsstrukturen in einer Mischlüftung

1. Auflage 2014

ISBN 978-3-942789-19-6

E.ON ERC Band 21**Thomas, S.**

A Medium-Voltage Multi-Level DC/DC Converter with High Voltage Transformation Ratio

1. Auflage 2014

ISBN 978-3-942789-20-2

E.ON ERC Band 22**Tang, J.**

Probabilistic Analysis and Stability Assessment for Power Systems with Integration of Wind Generation and Synchrophasor Measurement

1. Auflage 2014

ISBN 978-3-942789-21-9

E.ON ERC Band 23**Sorda, G.**

The Diffusion of Selected Renewable Energy Technologies: Modeling, Economic Impacts, and Policy Implications

1. Auflage 2014

ISBN 978-3-942789-22-6

E.ON ERC Band 24**Rosen, C.**

Design considerations and functional analysis of local reserve energy markets for distributed generation

1. Auflage 2014

ISBN 978-3-942789-23-3

E.ON ERC Band 25**Ni, F.**

Applications of Arbitrary Polynomial Chaos in Electrical Systems

1. Auflage 2015

ISBN 978-3-942789-24-0

E.ON ERC Band 26**Michelsen, C. C.**

The *Energiewende* in the German Residential Sector: Empirical Essays on Homeowners' Choices of Space Heating Technologies

1. Auflage 2015

ISBN 978-3-942789-25-7

E.ON ERC Band 27**Rolfs, W.**

Decision-Making under Multi-Dimensional Price Uncertainty for Long-Lived Energy Investments

1. Auflage 2015

ISBN 978-3-942789-26-4

E.ON ERC Band 28**Wang, J.**

Design of Novel Control algorithms of Power Converters for Distributed Generation

1. Auflage 2015

ISBN 978-3-942789-27-1

E.ON ERC Band 29**Helmedag, A.**

System-Level Multi-Physics Power Hardware in the Loop Testing for Wind Energy Converters

1. Auflage 2015

ISBN 978-3-942789-28-8

E.ON ERC Band 30**Togawa, K.**

Stochastics-based Methods Enabling Testing of Grid-related Algorithms through Simulation

1. Auflage 2015

ISBN 978-3-942789-29-5

E.ON ERC Band 31**Huchtemann, K.**

Supply Temperature Control Concepts in Heat Pump Heating Systems

1. Auflage 2015

ISBN 978-3-942789-30-1

E.ON ERC Band 32**Molitor, C.**

Residential City Districts as Flexibility Resource: Analysis, Simulation, and Decentralized Coordination Algorithms

1. Auflage 2015

ISBN 978-3-942789-31-8

E.ON ERC Band 33**Sunak, Y.**

Spatial Perspectives on the Economics of Renewable Energy Technologies

1. Auflage 2015

ISBN 978-3-942789-32-5

E.ON ERC Band 34**Cupelli, M.**

Advanced Control Methods for Robust Stability of MVDC Systems

1. Auflage 2015

ISBN 978-3-942789-33-2

E.ON ERC Band 35**Chen, K.**

Active Thermal Management for Residential Air Source Heat Pump Systems

1. Auflage 2015

ISBN 978-3-942789-34-9

E.ON ERC Band 36**Pâques, G.**

Development of SiC GTO

Thyristors with Etched

Junction Termination

1. Auflage 2016

ISBN 978-3-942789-35-6

E.ON ERC Band 42**Huber, M.**

Agentenbasierte

Gebäudeautomation für

raumluftechnische Anlagen

1. Auflage 2016

ISBN 978-3-942789-41-7

E.ON ERC Band 37**Garnier, E.**

Distributed Energy Resources

and Virtual Power Plants:

Economics of Investment and

Operation

1. Auflage 2016

ISBN 978-3-942789-37-0

E.ON ERC Band 38**Cali, D.**

Occupants' Behavior and its

Impact upon the Energy

Performance of Buildings

1. Auflage 2016

ISBN 978-3-942789-36-3

E.ON ERC Band 39**Isermann, T.**

A Multi-Agent-based

Component Control and

Energy Management System

for Electric Vehicles

1. Auflage 2016

ISBN 978-3-942789-38-7

E.ON ERC Band 40**Wu, X.**

New Approaches to Dynamic

Equivalent of Active

Distribution Network for

Transient Analysis

1. Auflage 2016

ISBN 978-3-942789-39-4

E.ON ERC Band 41**Garbuzova-Schiftler, M.**

The Growing ESCO Market for

Energy Efficiency in Russia: A

Business and Risk Analysis

1. Auflage 2016

ISBN 978-3-942789-40-0

In direct-current (dc) networks, dc-dc converters are used to step up or step down dc voltages and are therefore considered as the equivalent of the transformer in the classical alternating-current (ac) system. This dissertation is dedicated to the implementation of a medium-voltage high-power dc-dc converter.

After modelling the converter, a fast current controller is demonstrated that sets a reference current within one third of a switching period. Moreover, a balancing control scheme is developed which increases the utilization of the converter.

To increase the converter efficiency, an auxiliary circuit has been designed for a medium-voltage high-power application. The prototype demonstrates a compact design and an amortization in less than one year for a wind application.

The medium-frequency transformer is a key component in the dc-dc converter. The high power density and the requirements on the stray inductance make its design challenging. Therefore, especially the core losses under the influence of power electronics and the design of the stray inductance are analyzed. Moreover, a 2.2 MVA transformer operated at 1 kHz is constructed and demonstrates a ten-fold power density in comparison to a 50 Hz transformer.

Finally, the dissertation presents the construction and the setup of the high-power medium-voltage demonstrator of the converter. The demonstrator is rated for 5 kV dc-link voltage and a power of 5 MW. From the commissioning, valuable experience on an industrial construction of the converter are concluded, as well as future research topics.

ISBN 978-3-942789-42-4