Increasing the Functional Safety of Safety-Critical Systems in the 48 V Network
By Redundant Supply of the 48 V Network through the 12 V Network

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The undersigned hereby certify that they have read and recommend to the Faculty of Electrical Engineering, Mathematics and Computer Science (EEMCS) for acceptance a thesis entitled

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Abstract

In 2011, AUDI AG and four other German car manufacturers decided to implement an additional 48 V on-board power network inside cars. This offers the opportunity to increase the number of electrical applications in the car, to increase the fuel efficiency and to decrease the CO₂ emissions. In order to achieve this, the current electromechanical systems have to be moved to the 48 V network.

Safety critical systems may also be operated from the 48 V network. In case of failure, these systems are required to deliver a minimum level of performance. The objective of this research is to make an overview of possible failures, to obtain a method to operate these safety critical systems in case of these failures, and thus to increase the functional safety.

Possible failures of safety-critical systems have been analysed using a Fault Tree Analysis (FTA). A failure in the battery appeared to be the most frequently occurring fault that could lead to dysfunctional safety-critical systems. In this case, the Belt-driven Starter Generator (BSG) has to be shut off, as it cannot regulate its generated power as dynamically as the dynamic load demand of 48 V loads. This leads to a power failure of the 48 V network.

The proposed solution for this fault is a 12 V redundant supply, by connecting the 12 V network to the 48 V through two back-to-back MOSFET-type switching modules. One concept is to power the Belt-driven Starter Generator (BSG) with the 12 V network for a short time, so that the BSG can start the Internal Combustion Engine (ICE), and generate afterwards. The other concept is to power the safety-critical systems with the 12 V battery directly. Therefore, the concept of switching a constant power load between 48 V and 12 V was tested. This proved to be possible, although there is room for improvement.

The theoretical and practical performance of the 48 V BSG at nominal and reduced voltage have also been investigated. After the identification of the parameters of the machine, its performance has been modelled. Then, the real performance was tested. The performance of the BSG at reduced voltage was worse than expected, due to the control of the machine.

The current BSG may already be able to start an ICE in most of the cases, and it is expected that the machine will be able to start the ICE much more smoothly with improved control. Whether this really is possible should be determined in a test in practice.

The added functional safety of the proposed solution is dependent on the performance of the safety-critical systems on 12 V for the investigated topology. As it was not possible to test these systems, the added functional safety for the proposed solution is difficult to determine.

However, the concept has other interesting applications. In the current iHEV topology, the concept could serve as a redundant power generation application. Furthermore, in the case of a safety-requirement on the starting of an ICE, the concept could be implemented to meet this safety-requirement.
# Table of Contents

Acknowledgements xiii  

1 Introduction 1  
1-1 State of the art technology 2  
1-1-1 On-board power networks 2  
1-1-2 The 48 V power system 3  
1-2 Research objective 4  
1-3 Thesis outline 6  

2 Safety and safety analysis of the 48 V power network 7  
2-1 Functional safety 7  
2-1-1 The ISO 26262-1:2011 standard 7  
2-1-2 Safety critical systems in the 48 V network 8  
2-2 Safety analysis 9  
2-2-1 Safety analysis techniques 9  
2-2-2 Determination of an ASIL 10  
2-3 General means to increase safety 12  

3 Fault tree analysis of safety-critical components in the 48 V power network 15  
3-1 Assumptions made before analysis 15  
3-2 Fault tree analysis of safety-critical systems 15  
3-2-1 Failure of the power supply 16  
3-2-2 Failure of safety-critical systems 18  
3-2-3 Most frequently occurring fault 21
# Table of Contents

## 4 Overview of solutions for identified faults

4-1 Increasing the fault tolerance of machine drives ............................. 25  
  4-1-1 Continued operation after fault .............................................. 25  
  4-1-2 Fault tolerance through redundancy .................................... 26  
  4-1-3 Fault-tolerant permanent magnet synchronous machine design 27  
4-2 Fault tolerance of the power network ....................................... 27  
  4-2-1 Possible solutions to increase fault tolerance ......................... 27  
4-3 Proposed solution ...................................................................... 28  
  4-3-1 Type of switch ..................................................................... 29  
  4-3-2 Location of switch ............................................................... 30  
  4-3-3 Switch topology ................................................................... 31

## 5 Validation of reduced voltage operation of electrical machines

5-1 The belt-driven starter generator ................................................. 35  
  5-1-1 Choice of the machine ............................................................. 35  
  5-1-2 The design of the machine ....................................................... 36  
  5-1-3 Modelling the dual-three phase hybrid excited machine ......... 38  
5-2 Operating regions of the machine .............................................. 41  
  5-2-1 Non-salient machine ............................................................... 42  
  5-2-2 Salient machine ................................................................... 43  
5-3 Determination of machine parameters ......................................... 46

## 6 Measurements of real machine performance

6-1 Test set-up .............................................................................. 53  
6-2 Measurement results ................................................................. 55  
6-3 Interpreting the measurement results ....................................... 63  
  6-3-1 Comparison of model and results ........................................... 63  
  6-3-2 Starting the Internal Combustion Engine .............................. 63  
  6-3-3 Steering and braking ............................................................. 65

## 7 Implementing and testing the switching topology

7-1 The set-up used for testing ........................................................ 69  
7-2 Results .................................................................................. 71

## 8 Conclusions and recommendations

8-1 Conclusions .......................................................................... 79  
8-2 Contributions of work .............................................................. 80  
8-3 Recommendations for further work ......................................... 81

## A A d,q model for the dual three-phase machine

## B Analysis of machine losses
# Table of Contents

C Additional measurement results 93
D Q-Diode driver circuit 99

Glossary 109
  List of Acronyms ........................................... 109
  List of Symbols ............................................ 109
List of Figures

1-1 Example of a typical 12 V network supplemented with a 48 V network. 3
1-2 Voltage levels of the 48 V system according to LV 148. 5
1-3 The layout of this thesis. 6

2-1 Schematic representation of the 48 V iHEV topology 9

3-1 Top-level overview of faults of a system in the 48 V network. 16
3-2 Overview of faults of the power supply in the 48 V network. 17
3-3 Possible faults of the electrical machine. 20
3-4 Possible faults of the inverter of the electrical machine. 23
3-5 Possible faults of the control of the electrical machine. 23

4-1 Different possible connection points in the 48 V iHEV topology. 30
4-2 Overview of different possible switching topologies. 33

5-1 An exploded view of a claw-pole synchronous generator. 36
5-2 The tested Belt-driven Starter Generator. 37
5-3 The flux path inside a claw-pole synchronous machine. 37
5-4 A simplified overview of the magnetic flux path of the magnets inside the rotor. 38
5-5 An example of the combined delta-wye winding. 39
5-6 The inverters of the dual three-phase machine. 39
5-7 The operating regions of a non-salient machine. 43
5-8 The operating regions of a non-salient machine, for different values of field winding current. 44
5-9 The operating regions of a salient machine. 45
5-10 The operating regions of a salient machine, for different values of field winding current. 46
5-11 Measurement set-up for the the open circuit test of the machine. 48
5-12 The excitation flux of the investigated machine. ......................... 50
5-13 Ideal torque-speed characteristic of the machine, operated at 40 V and at 12 V . 50
5-14 Magnetization curves for three different types of steel. .................. 51

6-1 The test setup used for measurements. ................................. 54
6-2 The torque-speed characteristic of the machine, operated at 40 V. .... 57
6-3 The torque-speed characteristic of the machine, powered by the battery. .... 58
6-4 The control signals of the machine, that indicate that the performance of the machine is constrained by software. ......................... 59
6-5 Overview of mechanical power and dc link current of the machine at 40 V and when supplied by a battery. ............................ 60
6-6 Overview of power loss and efficiency of the machine at 40 V and when supplied by a battery. ................................. 62
6-7 Comparison of the ideal and real torque speed characteristics. It is assumed that the inverters of the machine are operating in square wave pulse width modulation. The dashed lines correspond to the modelled torque speed characteristic of the machine when it is powered by the battery. Only the saturated 12 V model had a maximum speed < 6000 RPM, after which field weakening could be applied to increase the operating range. This has not been included in this graph. .... 64
6-8 An example of the results from the starting test involving the starting of an ICE. 67

7-1 A schematic overview of the setup used for testing. .......................... 71
7-2 Switching waveform of the Q-Diode of a 100 A constant current load with a 4.7 mF capacitance parallel to the load. ........................ 72
7-3 Switching characteristics for a constant current load of 100 A at different capacitances. .................................................. 74
7-4 Switching characteristics of different constant power loads at a constant capacitance of 9.4 mF. .................................................. 75
7-5 Switching characteristics for a constant power load of 1.5 kW at different capacitances. .................................................. 77
7-6 The switching characteristics of a CAN-bus controlled Q-Diode. ........ 78

C-1 The voltage of the field winding of the machine, when powered by a battery. .... 93
C-2 The voltage of the field winding of the machine, when powered by a 40 V voltage supply. .................................................. 94
C-3 A capture of the waveform of the gate voltage of a MOSFET pair for sinusoidal pulse width modulation. .................................. 94
C-4 A capture of the waveform of the gate voltage of a MOSFET pair for square wave pulse width modulation. ................................ 95
C-5 Overview of the torque-speed characteristic for different voltages and temperatures. .................................................. 96
C-6 The electrical energy that the 48 V BSG can generate at 12 V. ............... 97

D-1 The circuit used to drive the Q-Diode. ........................................ 101
List of Tables

2-1 The different classes of severity for determining the ASIL. ................................. 11
2-2 The different classes of probability of exposure during operational situations for determining the ASIL .......................................................... 11
2-3 The different classes of controllability for determining ASIL. .............................. 11
2-4 The table used to attribute an ASIL to different events. ....................................... 11
4-1 Overview of possible solutions for a power network related fault. ....................... 29
4-2 Overview of characteristics of the different connection points between the 12 V and 48 V networks. ................................................................. 31
4-3 Topologies, their effectiveness and the respective number of necessary switches. .... 32
6-1 Components used for the machine measurements. ............................................. 54
6-2 The specifications of the prime mover. ............................................................... 54
6-3 Summary of the starting capability of the BSG for different situations. .............. 65
7-1 Overview of different components used for the test set-up. ............................... 71
B-1 Calculated losses for the machine for a stator current of $I = 70$ A. ................. 91
B-2 Calculated losses for the machine for a stator current of $I = 90$ A. ................. 91
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Chapter 1

Introduction

In the last decade, research and development efforts on the Plug-in Hybrid Electric Vehicle (PHEV), the Electric Vehicle (EV) and Fuel Cell Electric Vehicle (FCV) have increased significantly. Even though conventional vehicles with an Internal Combustion Engine (ICE) have the highest market share, with approximately 96.5% in the USA in 2013, the sales of hybrid and electric vehicles have grown significantly, with around 68% in the period from 2007 to 2013 in the USA [1]. This growth is predominantly driven by two factors [2].

First, the increasing global population has increased demand for personal transportation, and therefore demand for oil. With several large countries such as China and India rapidly developing, this demand is expected to increase even more in years to come.

In addition, governments and institutions are restricting fuel consumption and CO₂ emissions of cars. In order to meet these requirements, car manufacturers were forced to new approaches for improvement, and could no longer solely further mature the ICE technology.

Over the years, car manufacturers have also implemented more electrical applications in cars, and this has put pressure on the power requirements of the on-board power network. It has become increasingly difficult to implement new functionalities in the 12 V on-board power network due to both technological and economical reasons [3]. Therefore, car manufacturers recently decided to raise the voltage level of the on-board power network.

A raise of the voltage level of the on-board power network is not a new phenomenon. In 1955 car manufacturers decided to raise the voltage level from 6 V to 12 V. During the 1990’s, a new 42 V electrical system standard was proposed to increase the power, and to enable new electrical functions [4].

The 42 V standard was implemented in only a few car models, and the concept was never broadly adapted. The then existing battery technology was not yet ready [5]. A couple of years later in 1999, it was expected that vehicles with 42 V would remain a niche product.

Later, reduction of CO₂ emissions is an additional driver to implement the 48 V network [6]. Governments and institutions imposed new restrictions, forcing car manufacturers to further pursue CO₂ emission reduction possibilities. At that point, new developments of lithium-ion technology have partially overcome the previous hurdles caused by battery technology [6].
Therefore, in 2011 AUDI AG and four other German car manufacturers agreed on implementing an additional 48 V on-board power network, including other common architectural elements. Due to this additional network, cars are able to facilitate additional high-power applications now and in the future, e.g. electrical roll stabilization, steer-by-wire and brake-by-wire. Furthermore, it is possible to recuperate more energy otherwise lost through braking, which increases fuel efficiency and reduces CO₂ emissions [6].

The remainder of this introduction will elaborate on the state-of-the-art technology of the on-board power networks of automotive vehicles and the 48 V power network specifically. Then, the research objective and approach are discussed, concluded with an outline of this thesis.

1-1 State of the art technology

1-1-1 On-board power networks

There are currently several common types of on-board power networks. The 12 V network is currently the most frequently occurring network. In this topology a starter, a generator, and the loads are connected with a battery power source. The loads can be connected and disconnected with the power source through a switch. As the load requirements on the automotive power supply are increasing, this network is increasingly less suited due to both technological and economical reasons [3] and this drives the automotive industry to a power network with a higher voltage.

In addition, the topology containing both a 12 V and a 42 V network was proposed, where both voltage networks are coupled through a DC/DC converter. In this topology, the high-power loads are connected on the 42 V network, whereas low-power loads are connected to the 12 V side of the network. Networks both with a single battery on the 42 V side and batteries on both sides had been proposed [7].

The combination of both voltage levels reduces implementation costs and ensures a smooth transition with regards to redesigning the entire system including its loads to 42 V. The latter would not only increase the cost of all components, but it would have also required immediate availability of components with new voltage ratings. At the time however, the existing battery technology did not suffice on aspects including range, charging infrastructure and costs [5] and therefore this system was not broadly adapted.

In 2011, a successive topology containing both a 12 V and a 48 V network was proposed by AUDI AG and four other German car manufacturers. The value of 48 V was chosen, as it is the highest nominal voltage below the limit for shock protection of 60 V while having a buffer for unintended overvoltages. Exceeding 60 V would significantly increase costs due to extra safety measures that would have to be taken. Cars with this on-board power network are often called mild hybrids.

The implementation of the 48 V system is expected to be inevitable, due to the combination of the increasing electrical load requirements, the requirements on the reduction of CO₂ emissions, and battery technology developments, as existing cell-based accumulators fulfill the requirements [6]. A typical network is illustrated in figure 1-1. The high and low voltage networks are again coupled with a DC/DC converter, and high power functions are connected to the high voltage side of the network.
Last, some HEVs and many PHEVs include a power network with an even higher voltage. In these cars, the 12 V, or 12 V and 42 V power system is combined with another high voltage power level. This high voltage level is primarily used to drive the electric machine used for traction. The level of this voltage differs per car, e.g. the '04 and later model of the Toyota Prius have a high voltage network with a rated voltage of around 200 V, and in the RX 400h Lexus this network has a rated voltage level of 288 V.

1-1-2 The 48 V power system

The 48 V power system has only recently been proposed and currently only few loads have been moved to this side of the power network. Eventually many, if not all, high-power loads will be connected to the 48 V side. This increases both the comfort for the consumer and the fuel efficiency of the vehicle. The 48 V power system increases the available power for new systems for extra comfort, e.g. roll stabilization and extra infotainment systems, and for systems that make the car more fuel efficient, such as EPS and the ability to go periodically into sailing mode. During sailing mode the engine is turned off, and this increases overall efficiency. In addition, the system can reduce losses due to the fact that the current in the wires is a factor of four times lower for the same delivered power. Therefore, the diameter of the wires can also reduced and weight can be saved.

The standard published by AUDI AG and other German car manufacturers includes information about the different voltage levels of the 48 V system, which are shown in figure 1-2. The figure shows each voltage level and the corresponding mode of operation of the system at that voltage level. The system should operate below 60 V and above 20 V. The choice of the 48 V is a compromise between improvement of the system and cost, as a voltage higher than 60 V requires extra safety measures. The standard provides the following information about the voltage levels:

![Diagram of a typical 12 V network supplemented with a 48 V network.](image)
• > 60 V is the shock protection voltage not to be exceeded according to ECE-R 100\(^1\);
• 58 V - 60 V is a 2 V safety margin;
• 54 V - 58 V is the overvoltage range, that includes all tolerances. In this range, overvoltage protection must be active;
• 52 V - 54 V is the upper operating range with restricted performance. This range is dedicated for the calibration of the accumulator and the absorption of recuperation energy;
• 36 V - 52 V is the operating range without full performance;
• 24 V - 36 V is the lower operating range with restricted performance. Operating in this range is only temporarily allowed and countermeasures have to be taken to return to the operating range without functional degradation;
• 20 V - 24 V is defined as the undervoltage range;
• < 20 V is the storage protection voltage, where the system is not operational.

The connection of high-power loads to the 48 V power network is an important step in the electrification of the car, and also an important factor to satisfy the requirements on CO\(_2\) emissions and fuel efficiency. In this process, safety-critical systems will be connected to the 48 V network. It is therefore important that the network is robust and that these safety-critical systems are fault-tolerant, so that they can still be operated after a fault has occurred.

The 48 V standard is relatively new, and not much research has been done on this subject. However, research has been published on the 42 V standard, which very much resembles the 48 V standard. The research is predominantly devoted to the network in general, the design of its components, and the optimisation of the power management of and energy storage within the network [8–12]. In addition, general research has been published on faults and fault-tolerance of electrical machines and its inverters [13–16], even for automotive applications [17,18], where different designs and implementations have been investigated.

The combination of two voltage networks may offer new opportunities to increase functional safety. From literature it is not apparent which designs or solutions to increase functional safety are applicable in a combined power network in such an automotive application with two voltage levels. The robustness of this network, and the fault-tolerant operation of these safety-critical systems specifically in this network in case of a failure, are important for the application of the network with its subsystems, and are interesting topics for research.

1-2 Research objective

In a Plug-in Hybrid Electric Vehicles (PHEVs) and Hybrid Electric Vehicles (HEVs), more and more mechanical systems have been replaced by electromechanical systems. The replacement of mechanical systems includes safety critical-systems and these may be operated from the 48 V on-board power network. It is necessary that these systems can still operate in case of

\(^1\)ECE-R 100 is Regulation 100, a document on battery electric vehicle safety proposed by the European Commission for Europe of the United Nations.
The research objective of this thesis is to make an overview of possible failures of safety-critical systems in the 48 V network, and, if possible, to derive methods to operate these safety-critical systems in case of a failure such that the machines can carry out the minimum required level of performance and a higher level of functional safety is achieved. The research objective of this thesis can be divided in the following separate research questions:

1. Which systems are safety-critical, and what are the requirements on safety and fault tolerance of the 48 V network for these systems?

2. What are current possible failures of safety-critical systems in the 48 V network?

3. Which possible solutions are there to prevent or mitigate these failures, which of these are worth investigating further and how can they best be implemented?

4. How is the performance of these proposed solutions, and do these proposed solutions offer the minimum required level of performance and extra functional safety in practice?

To answer the first research questions, the safety critical systems in a car are identified and the requirements on safety inside a car are discussed. The second research question will be answered with the help of a Fault Tree Analysis (FTA) of the safety-critical systems in the 48 V network. Then, possible feasible solutions are assessed and the best solutions are considered, including how they can best be implemented. This forms the answer on question three. Finally, the performance of these proposed solutions in practice will be assessed and their added functional safety is evaluated.
1-3 Thesis outline

The thesis outline is shown in figure 1-3. This introduction forms the first chapter of this thesis. The next chapter discusses the concepts of functional safety and safety analysis, and identifies the safety-critical systems in the car. Chapter three identifies possible failures of safety-critical systems within the 48 V power network through a fault tree analysis. Chapter four will discuss possible solutions for these failures, and conclude with solutions for further exploration. Then, the theoretical difference in performance of an electrical machine at normal and reduced voltage is investigated in chapter five. The real difference in performance of an electrical machine is investigated by means of measurements in chapter six. The concept of the switching of a load on the 48 V to the 12 V network is investigated in chapter seven. The thesis concludes with conclusions and recommendations in chapter eight.

![Figure 1-3: The layout of this thesis.](image)
Chapter 2

Safety and safety analysis of the 48 V power network

The objective of this chapter is to explain how safety is incorporated in the design process in the automotive industry, and to determine which systems in the 48 V power network are safety-critical. This is explained by discussing the used standard for functional safety, the safety-critical systems in the 48 V network, the analysis of safety, the determination of the required level of safety, and how extra safety is obtained in general in practice.

2-1 Functional safety

2-1-1 The ISO 26262-1:2011 standard

In the automotive industry, the ISO 26262-1:2011 standard titled 'Road vehicles - Functional Safety' is commonly used as a framework to acquire the required functional safety levels for different events. The process of obtaining such a level is discussed in detail further in this chapter. The standard is relatively abstract, and leaves room for interpretation for the user.

ISO 26262-1:2011 defines standards for functional safety, which it defines as the absence of unreasonable risk due to hazards caused by malfunctioning behaviour of electrical systems. The standard does not strictly define fault tolerance. It utilizes four different Automotive Safety Integrity Levels (ASIL) to classify the necessary requirements and safety measures that have to be applied for avoiding an unreasonable residual risk of an item or element [20]. Unreasonable risk is defined as risk that is judged to be unacceptable in a certain context according to valid societal moral concepts [20]. An item is defined as a system or an array of systems to implement a function at the vehicle level, while an element is defined as a system or part of a system including components, hardware, software, hardware parts and software units.

An ASIL is assessed for each hazardous event that can be identified for an item. It is determined through a combination of the result of hazard analysis and risk assessment. The
severity, probability and controllability of hazardous events is estimated, based on a defined rationale, for each event. The ASIL is assigned based on the results of these estimations, where ASIL D is the most stringent level, and ASIL A is the least stringent level. In addition, for events without safety relevance there is the level Quality Management (QM). In this case, only standard quality management processes are required. An example of the assignment of an ASIL level is covered later in this chapter.

2-1-2 Safety critical systems in the 48 V network

The considered system is the intelligent Hybrid Electric Vehicle (iHEV) topology, designed by Audi, including an Electrical Power Steering (EPS) and a braking system. The 48 V topology consists of a 48 V belt-driven starter generator, an electrical roll stabilisation, an electrical turbocharger and a DC/DC converter. The EPS and braking system has been added to the topology, and the whole system is shown schematically in figure 2-1. The 12 V network has already been implemented in cars for a very long time and much effort has already been put into increasing the functional safety of this network to a high level. Therefore, the focus of the research is solely on the 48 V side of the power network.

The sole safety-critical systems of a vehicle are its steering and braking systems. Furthermore, it is debated whether the eTurbo, the electrical turbocharger, and eAWS, the active roll stabilization, are safety critical systems. The eTurbo system provides extra air to the engine in the case of acceleration. In the case that a driver is overtaking another vehicle, a sudden lower than expected acceleration can lead to a dangerous situation. The roll stabilization stabilizes the vehicle in case of forces to sideways movement, and a unexpected loss of this stabilization can lead to dangerous situations in a curve.

For these safety critical systems, Audi requires that the driver is able to switch at least two lanes before safely putting the car to a stop at the side of the road. Therefore, the following safety-critical systems are identified in this network:

- Electric power steering.
- Braking system.

The following systems, that could be relevant for safety, are present in the 48 V power network:

- eTurbo.
- Active roll stabilization.

Other systems provide comfort to the driver, and are not necessary for the driver to bring him or herself to safety. Therefore, these are not safety critical systems.

There are multiple causes of failures of these safety-critical systems. Each possible cause will emerge from the Fault Tree Analysis that will be carried out. To put fault tree analysis in the perspective of the total design, fault tree analysis and other safety analysis techniques will be first discussed.
2-2 Safety analysis

2-2-1 Safety analysis techniques

During the design and operation of engineering systems such as aircraft, automotive, railway, space, and military, specific procedures were developed to meet certain safety requirements. For the investigation of the effect of faults on the reliability and safety during the design phase and operation of the system, multiple safety analysis techniques have been developed [21,22], including:

- Event Tree Analysis (ETA) and Fault Tree Analysis (FTA)
- Failure Mode and Effects Analysis (FMEA)
- Hazard Analysis (HA)
- Risk classification

These analysis techniques can be combined to assure maximum reliability and safety. Failure Mode and Effects Analysis (FMEA) identifies all components, failures, causes and effects. The Fault Tree Analysis (FTA) is used for single failures to determine the causes and the logical interconnections on a component level. These failure causes can then be used to design the overall reliability of the system.

Hazard Analysis (HA) extracts safety-critical failures from the FMEA. These failures are then analysed with a fault tree analysis, so that the risk of hazardous faults can be lowered through designing the system at lower levels in such a way that it is safe.

At this stage in the design process, remaining faults are then usually classified in their...
risk. If necessary, techniques to reduce the risk to an acceptable level are then determined. There are many risk classifying standards and methods for different industries, but the automotive industry uses the Automotive Safety Integrity Level (ASIL) defined by the standard ISO 26262-1:2011. The determination of these levels will now be discussed and is illustrated with an example.

2-2-2 Determination of an ASIL

The severity of potential harm of a hazardous event can be categorized in the classes S0, S1, S2 or S3. Hazardous events that cause no injuries are classified as S0, while life-threatening injuries and fatal injuries are classified as S3. Hazardous events classified as S0 are not assigned an ASIL. Table 2-1 shows all classes of severity.

Furthermore, the probability of exposure of each hazardous event during operation is estimated, based on a defined rationale. The probability is categorized according to the probability classes E0, E1, E2, E3 and E4. For classes E1 to E4, the difference in probability to the next class is an order of magnitude. The class E0 may be used for situations that are considered extremely unlikely or incredible, and thus hazards assigned to E0 require no ASIL assignment. Table 2-2 summarizes the classes and their probability.

The controllability of a hazardous event by the driver or other persons potentially at risk is the third and last factor in determining the ASIL. This controllability is again estimated based on a defined rationale and is categorized in the classes C0, C1, C2 and C3. The class C0 may be used for hazards addressing the unavailability of the item if the item does not affect the safe operation of the vehicle. Class C0 may also be assigned if dedicated regulations exist that specify functional performance with respect to a defined hazard. A hazard assigned to C0 does not require an assigned ASIL. Table 2-3 summarizes the classes and their controllability.

Now, the combination of the severity, probability and controllability of an event determines the associated ASIL according to table 2-4. For example, an event with a severity of S2, a probability of E4 and a controllability of C3 is assigned ASIL C. Hazardous events classified as S0 are not assigned an ASIL. Table 2-1 shows all classes of severity.

An ASIL is only a guideline for a failure rate, and a car can always be designed safer than required. In general, adding extra safety measures increases costs unnecessarily. The loss of an assisted steering system during parking will likely not be harmful, and therefore reducing this event’s occurrence increases costs unnecessarily. Reducing the occurrence of harmful events can done in multiple ways, and general means to do so will be discussed now.
### Table 2-1: The different classes of severity for determining ASIL [20].

<table>
<thead>
<tr>
<th>Severity class</th>
<th>S0</th>
<th>S1</th>
<th>S2</th>
<th>S3</th>
</tr>
</thead>
<tbody>
<tr>
<td>No injuries</td>
<td>Light and moderate injuries</td>
<td>Severe and life-threatening injuries</td>
<td>Life-threatening injuries (survival uncertain), fatal injuries</td>
<td></td>
</tr>
</tbody>
</table>

### Table 2-2: The different classes of probability of exposure during operational situations for determining the ASIL [20].

<table>
<thead>
<tr>
<th>Probability class</th>
<th>E0</th>
<th>E1</th>
<th>E2</th>
<th>E3</th>
<th>E4</th>
</tr>
</thead>
<tbody>
<tr>
<td>Incredible</td>
<td></td>
<td>Very low probability</td>
<td>Low probability</td>
<td>Medium probability</td>
<td>High probability</td>
</tr>
</tbody>
</table>

### Table 2-3: The different classes of controllability for determining ASIL [20].

<table>
<thead>
<tr>
<th>Controllability class</th>
<th>C0</th>
<th>C1</th>
<th>C2</th>
<th>C3</th>
</tr>
</thead>
<tbody>
<tr>
<td>Controllable in general</td>
<td></td>
<td>Simply controllable</td>
<td>Normally controllable</td>
<td>Difficult to control or uncontrollable</td>
</tr>
</tbody>
</table>

### Table 2-4: The table used to attribute an ASIL to different events [20].

<table>
<thead>
<tr>
<th>Severity class</th>
<th>Probability class</th>
<th>Controllability class</th>
</tr>
</thead>
<tbody>
<tr>
<td>S1</td>
<td>E1</td>
<td>C1 QM QM QM</td>
</tr>
<tr>
<td></td>
<td>E2</td>
<td>C1 QM QM QM</td>
</tr>
<tr>
<td></td>
<td>E3</td>
<td>C1 QM QM A</td>
</tr>
<tr>
<td></td>
<td>E4</td>
<td>C1 QM A B</td>
</tr>
<tr>
<td>S2</td>
<td>E1</td>
<td>C1 QM QM QM</td>
</tr>
<tr>
<td></td>
<td>E2</td>
<td>C1 QM QM A</td>
</tr>
<tr>
<td></td>
<td>E3</td>
<td>C1 QM A B</td>
</tr>
<tr>
<td></td>
<td>E4</td>
<td>C1 A B C</td>
</tr>
<tr>
<td>S3</td>
<td>E1</td>
<td>C1 QM QM A</td>
</tr>
<tr>
<td></td>
<td>E2</td>
<td>C1 QM A B</td>
</tr>
<tr>
<td></td>
<td>E3</td>
<td>C1 A B C</td>
</tr>
<tr>
<td></td>
<td>E4</td>
<td>C1 B C D</td>
</tr>
</tbody>
</table>
2-3 General means to increase safety

There are in general four different techniques to increase the reliability and safety of a system [23]. These techniques are fault prevention, fault forecasting, fault removal, and fault tolerance.

Fault prevention

Traditional fault-prevention involves specification of the reliability of the system and studying how to increase the Mean Time to Failure (MTTF) of the system, so that it meets the required reliability. This is obtained by quality control techniques during the design and manufacturing of hardware and software.

Fault forecasting

Fault forecasting is aimed at fault prediction, followed by prevention. It is conducted by performing an evaluation of the system behaviour with respect to fault occurrence. The evaluation is obtained through computations using measured values of the system.

Fault removal

Fault removal involves finding and removing faults, which can occur before the system is developed and produced or when the system is already in use. In the latter case, this is possible through corrective maintenance and preventive maintenance. Preventive maintenance uncovers and removes faults before they might cause errors during normal operation. This can be physical faults that have occurred since the last preventive maintenance or design faults that have led to errors in other similar systems. Corrective maintenance is often performed in stages, where the fault is first isolated by a workaround and then removed. These forms of maintenance can be applied to both non-fault-tolerant systems as well as fault-tolerant systems. For fault-tolerant-systems, the maintenance can be performed both during service outage or without interrupting service delivery.

Fault tolerance

Fault tolerance involves mechanisms to tolerate a fault, such that the system will still deliver the correct service in case of a fault. This technique is often implemented by inserting one or more redundant systems to counteract a fault, and is a widely adopted technique for safety-critical systems. In case of failure, the system can then still deliver the required service.

There are five important aspects to consider when making a system fault tolerant [24]. First of all, the system has to be built such that a system can still continue its service while parts of the system have failed. Second, the fault has to be detected when it occurs. This is required for the system to react on the event of a fault. Third, it is important that the system adapts its control so that proper service delivery is ensured. In addition, the system has to isolate the fault, so that it cannot propagate to other parts of the system or influence other parts of the system. Last, it is important that the fault is reported after detection, so
that the faulty system can be replaced.

Fault tolerant systems can be categorized depending on their fault-tolerance requirements. The following categories exist [22]:

- **Fail-operational (FO)** systems tolerate one failure and stay operational. The breaking and steering systems in a car belong to this category.

- **Fail-safe (FS)** systems tolerate one or multiple failures and then reach a safe state. This can be reached passively, without external power, or actively by a special action, with external power.

- **Fail-silent (FSIL)** systems become passive after one or multiple failures. They switch off and stop influencing other components.

Now different safety techniques have been discussed, the 48 V power network will be analysed. Then, solutions will be proposed based on the identified faults.
Chapter 3

Fault tree analysis of safety-critical components in the 48 V power network

This chapter identifies possible faults in safety-critical systems in the 48 V power network, both on power system level and system level, including components. These possible faults are categorized based on the components within a system. Section 3-1 discusses the assumptions made prior to the analysis. Then, section 3-2 identifies the possible failures that could occur in a safety critical system with a FTA.

3-1 Assumptions made before analysis

For this analysis, the iHEV topology with EPS and a braking system is considered. This topology is shown in figure 2-1. It is assumed that the car is under normal operation conditions. It is assumed that no faults have already occurred and that no safety mechanisms related to a car crash have been activated. Furthermore, it assumed that the battery management system works normally and is capable of detecting the ageing of the battery, by which it will be replaced during maintenance. Furthermore, the typical expected lifetime of the Lithium-ion battery is expected to be approximately 8 years. In addition, it was assumed that the lifetime of a car is 15 years, with an annual operating time of 500 hours per year. As a result, the total operating time is assumed to be 7,500 hours.

3-2 Fault tree analysis of safety-critical systems

The failure causes of these safety-critical system are categorized in two different categories, namely power related and system related. This is shown in figure 3-1. First, the failure of the power supply will be discussed in section 3-2-1. Then, the failure of the system itself will be discussed in section 3-2-2.
3-2-1 Failure of the power supply

The 48 V side of the power network has multiple components that can supply power to the network. The Belt-driven Starter Generator (BSG) is the main source of instantaneous power. Excess generated energy is absorbed by the Li-Ion battery, or directly transferred to the 12 V network through the DC/DC converter. Furthermore, the DC/DC converter and the battery can feed power back in the 48 V network. In addition, the roll stabilization can also feed energy back to the network.

Malfunction of these components, the loads and the wiring harness can lead to either undervoltage or overvoltage inside the network. The causes of under- or overvoltage are presented in the FTA as shown in figure 3-2 and will now be highlighted shortly.

Overvoltage in the 48 V network

For an overvoltage in the power network to occur, an overvoltage has to be generated and there has to be no load that is able to absorb the overvoltage. An overvoltage can be both temporary or continuous. If a 48 V load feeds energy back into the network, or when a load disconnects, an overvoltage can be generated. In addition, the DC/DC converter or the BSG can generate an overvoltage due to a fault on a continuous basis.

For an overvoltage to occur, it is required that the battery as well as loads in the network cannot absorb the overvoltage. In the case of a disconnection or failure of the battery, however, the 48 V BSG and DC/DC converter are shut off. This is due to the fact that the BSG cannot control its output fast enough to respond to the high dynamic loads present in the 48 V network. Therefore, the BSG and DC/DC converter are shut off when the battery malfunctions.

Figure 3-1: Top-level overview of faults of a system in the 48 V network.
Figure 3-2: Overview of faults of the power supply in the 48 V network.
Undervoltage in the 48 V network

Undervoltage in the power network can occur due a short circuit, due to a loose connection or due to a battery malfunction. A short circuit can occur due to a connection between the 48 V and 12 V network and due to a short to ground. A connection between the 48 V and 12 V networks is possible in case of abrasion of insulation between the different networks at the same location, or due to a connection of connectors at a device that is connected to both voltage levels, such as the DC/DC converter. In theory, a short to ground is possible due to an abrasion of the insulation layer, a failed (dis)connect of a relay under load, or an incorrect blowing of a fuse. In the latter two cases, the relay contacts could melt or the fuse does not blow correctly and still conducts.

A disconnection can be caused due to a broken wire or connector, or because a fuse is blown unintentionally. At voltage levels higher than approximately 15 V, an arc will occur that does not extinguish by itself. If this arc is not extinguished quickly, it is likely that a fire will start in the car. Therefore, when a serial arc is detected, a safety mechanism will trip and this mechanism disconnects or shuts down the power supplying components, e.g. the DC/DC converter, the battery and BSG.

Lack of sufficient power fed to the network is only dependent on the failure of the battery. The loads in the 48 V system are highly dynamic and thus require power very dynamically. The BSG is unable to control its generated output power this rapidly, which creates a situation where the voltage of the 48 V network is unstable, and overvoltages and undervoltages may occur. Therefore, the BSG and DC/DC converter are shut down when the battery is disconnected.

The battery on its own is able to supply enough power for the loads on the 48 V network. Therefore if both the generator and the converter fail, the safety critical loads can still function long enough to drive to a safe place and stop the car, and the driver is not endangered.

A battery failure that is not detected, will also lead to an undervoltage in the network. This is due to the fact that the car goes into sailing mode whenever the driver is not pressing the gas pedal. The sailing mode is an automatic mechanism to increase fuel efficiency that will be implemented with new cars that have the iHEV energy network. The mechanism shuts off the engine whenever someone is not giving gas, and thus then the BSG is no longer generating energy. In this case, the DC/DC converter is operating in its standard mode operation, which is feeding energy into the 12 V network as there is no generator in the 12 V network. In this situation, loads are demanding power, while there is no power generated nor any energy can be supplied. Therefore, the voltage will drop quickly and all systems in the 48 V network, including braking and steering, will stop operating.

3-2-2 Failure of safety-critical systems

Possible failure modes on a system level of safety-critical systems can be categorized in three different locations. The fault can occur in the electrical machine, in the inverter circuit or in the control of the machine. The faults for each of these locations will be shortly highlighted in their respective section.

The analysis investigated the failure rates of machines of systems that were still in development, and therefore there was no data available on the failure rate of these machines. However, currently the safety requirement on the braking and steering systems at 12 V is
ASIL D. Obviously the systems will have the same safety requirement when they are supplied by the 48 V network.

At Audi, there was only little information available on the cause of the failures. For the 12 V BSG, a failure rate analysis based on field data and theoretical modelling was carried out. The study already gave an understanding of the possible causes and their frequency of the faults, but in addition literature was studied.

Faults in electrical machines

The faults in electrical machines can be classified in four types [25], namely bearing faults, stator faults, rotor faults and eccentricity related faults. The fault tree of faults in electrical machines is shown in figure 3-3.

Bearing faults, for example due to fatigue, corrosion, improper lubrication or improper installation of the bearing. Bearing related faults account for 51% of all faults, and 69% of motor related faults according to [26]. This was for totally enclosed fan cooled squirrel cage induction motors in the petroleum and chemical industry.

Stator faults, such as open and short circuits in the windings due to insulation failure. This includes phase to ground, phase to phase and turn to turn short circuits. Stator faults account for 16% of all faults, and 21% of motor related faults of mentioned induction motors [26].

Rotor faults, such as broken rotor bars and end-ring breakage. This can have several causes, of which examples are thermal stresses, magnetic stresses, environmental stresses or mechanical stresses. Rotor faults account for 5% of all faults, and 7% of motor related faults of mentioned induction motors [26].

Eccentricity related faults due to an unequal air gap between the stator and the rotor. This can be either static or dynamic air gap eccentricity. A fixed unequal air gap, e.g. due to incorrect positioning of rotor, is called static eccentricity. Dynamic eccentricity is a result of a situation in which the center of the rotor is not at the center of the rotation, which leads to an rotating position of minimum air-gap. If eccentricity becomes large, it can cause stator to rotor rub, leading to damage of the stator and rotor.

Faults in inverter circuit

A fault in the electronics of the inverter circuit can also lead to a possible failure of the system. Possible faults are a fault in the MOSFET, in the MOSFET driver and in the DC link capacitor [15]. The possible failures are summarized in figure 3-4.

According to the internal analysis of Audi, faults in the inverter circuit are the most frequently occurring faults in the power electronics of the machine for a BSG at 12 V, with around two thirds of the total faults for the power electronics. However, in [27] it is reported that the power converter circuits account for 38% of the problems of adjustable speed drives, and the control circuits for 53%. In addition, in [28] a failure analysis of a permanent magnet synchronous motor in the motor industry is carried out, both with field data and with data from RDF 2000, the reliability handbook of Union Technique de L’Electricite, which is the French organisation for normalisation in the domain of electronics. Here, it is reported that faults in the MOSFETs account for approximately 60% of all faults in the electronics.
Figure 3-3: Possible faults of the electrical machine.
Faults in control circuit

Finally, a fault in the control circuit can also lead to a malfunction of the system. In [27], it is reported that a fault in the control accounts for 53% of the faults in the power electronics of an adjustable speed drive. A fault in the control circuit is possible due to a malfunction of the Electronic Control Unit (ECU), due to a loss of ground, or due to lack of communication. The latter comprises of either a broken connection between the ECU and the communication bus, or a broken connection between the ECU and the semiconductor switches. A loss of ground of an ECU at the 48 V domain could forward the voltage on to the communication bus line, which could possible lead to a destruction of devices at the 12 V side. A malfunction of the ECU can be caused by a software related fault, by sensor related fault, by a microcontroller related fault or due to loss of power of the ECU. The possible faults are summarized in the fault tree shown in figure 3-5.

3-2-3 Most frequently occurring fault

With the Fault Tree Analysis (FTA), an overview of the possible faults for safety-critical systems in the 48 V network is given. Manufacturers have taken several measures to decrease the failure rates of components. In practice, the 12 V and 48 V cables are separated physically by a certain distance. This prevents a short circuit between both voltage levels. In addition, loss of common ground is also prevented by separation of power wires and separation of connectors on dual voltage electronics. It is however possible to analyse the failure rates more quantitatively.

The chance of a failure in an electrical machine in a safety-critical system, that is so severe that the electrical machine is no longer able to deliver the minimum required service, is very low. The steering and braking systems on the 12 V side are currently designed with an ASIL D safety requirement. Future steering and braking systems on the 48 V network will therefore also be classified as an ASIL D, and these systems must also have a failure rate of $10^{-9}$ or lower per hour.

An overvoltage in the 48 V network can only occur when two faults occur at the same time, which is the generation of an overvoltage and the inability to absorb the overvoltage. An overvoltage cannot be absorbed when both the battery is disconnected, and no load can absorb the overvoltage. The generation of an overvoltage is either due to the BSG, the DC/DC converter, the backfeed of a 48 V load, or the disconnection of a load. Thus the risk that an overvoltage is generated is dependent on the failure of multiple systems at the same time, and therefore lower than the failure of each of these systems on their own.

The risk of an undervoltage in the network is relatively high. This is due to the fact that an undervoltage in the 48 V network is solely dependent on the failure of a single component, which is the battery. If the battery fails, the BSG is unable to control its output power dynamically enough with reference to the high dynamic loads present in the 48 V network. Therefore, the BSG is shut off as a safety measure and no energy storing or generating system is present in the 48 V network. This will lead to an undervoltage in the 48 V network, and all systems in the 48 V network will stop operating.

From this analysis, it follows that an undervoltage in the network is by far the fault with the highest risk. This is due to the fact that this consequence is solely dependent on a failure of
the battery. Furthermore, there is no ASIL level assigned to the functioning of the battery management system. In other words, there is no requirement on the maximum failure rate of the battery management system. Therefore, a failure rate of \(1000 \cdot 10^{-9}\), which is the requirement for ASIL A, or higher may well be possible.

To conclude, a loss of supplied voltage due to insufficient supplied power is a severe fault that has the highest likelihood to occur. This is mainly due to a fault in the battery, either because the battery breaks down, or because the battery is actually not functioning properly, while the energy management system thinks the opposite.
Figure 3-4: Possible faults of the inverter of the electrical machine.

Figure 3-5: Possible faults of the control of the electrical machine.
Overview of solutions for identified faults

The objective of this chapter is to give an overview of possible solutions for the previously identified faults. Different techniques to increase the fault-tolerance of electrical machines will be discussed first. Then, solutions to faults related to the power network are discussed.

4-1 Increasing the fault tolerance of machine drives

Currently, a wide variety of machines are used in cars, both ranging from different types of electrical machines to different rated powers. Because switched reluctance machines, induction machines, and permanent magnet machines are currently built in cars, the increase of fault-tolerance of each of these types of machines is investigated. As the automotive industry tries to keep costs as low as possible, doubly-fed induction machines and a redundant motor as solution have not been investigated. Solutions to the faults identified in the fault tree analysis will be discussed afterwards.

As the topologies of induction machines, permanent magnet synchronous machines, and switched reluctance machines can be similar, the identified possible faults and their solutions for these topologies will be discussed together. First of all, it is possible to operate the machine after a fault has occurred. In addition, fault tolerance can be increased by making use of redundancy. These different types of solutions will be shortly highlighted.

4-1-1 Continued operation after fault

There are different methods to operate an induction machine and specifically designed permanent magnet synchronous machine after a fault. One method is to operate a star coupled induction motor in single-phase operation, where the neutral point of the star coupled induction machine is not connected [29]. When phase $a$ is disconnected due to a fault, the
remaining phases $b$ and $c$ can be phase-shifted so that $i_{bref} = -i_{cref}$. The machine can operate with a torque with ripple, and therefore this method does not suffice for high dynamic and precision performances. Further research developed an algorithm that attenuated the pulsating torque [30]. This is obtained by injecting harmonics.

Continued operation of the induction machine after a fault has occurred is a very cost-efficient method to increase the fault tolerance of a system. If the converters are not overdesigned, the machine will deliver reduced torque. Otherwise, if the rating of the converters is overdesigned, it can tolerate higher currents, and the machine will be able to deliver the same torque as before the fault has occurred. In both cases, the control algorithm has to be adapted so that the system functions well after a fault has occurred.

### 4-1-2 Fault tolerance through redundancy

In addition, it is possible to create a fault-tolerant induction machine with redundancy in the converter topology. The following topologies are options that provide fault-tolerance:

- A switch-redundant topology, with up to four switches and three fuses. A topology with all switches and fuses can clear a single shorted switch, a single opened switch, and an opened phase fault. The neutral of the motor and the lines are connected with the midpoint of the dc bus with a switch [31,32]. When a fault occurs, the mid point of the dc-bus capacitor eliminates the pulsating torque that normally occurs with an open phase operation.

- A double-switch redundant topology, with four inverter legs, two fuses and two switches per phase leg, and two capacitors [33]. In case of a fault, the faulty leg is disconnected and the redundant leg is connected to the star point of the machine, and with control a balanced post-fault operation can be maintained [14].

- A phase redundant topology, where a fourth phase leg can replace a faulty phase leg. Here, there are only fault isolating switches and fuses in three legs of the inverter. When a fault occurs, the faulty phase leg is disconnected, and the fourth phase leg is connected. The machine can resume normal operation [33,34].

- A multi-phase topology, with for example nine phases [35]. In this system, one or more phases can fail before the machine stops delivering its intended service. This topology has several benefits, for which the reader is referred to [35]. The system can continue operation at reduced torque, or at the same torque. For the latter case, the rating of components has to be overdesigned. Furthermore, the higher losses in the machine due to the higher currents can limit the post-fault output torque. This topology can also be designed so that every phase can have its own single-phase converter [36].

These topologies, including different versions, are summarized in [37]. These solutions require extra components or higher rated components, so that a higher current can be tolerated in case a fault occurs. These extra or higher rated components come at a higher cost, which is downside in the automotive industry where costs are important. If maximum performance is not required after a fault has occurred, it could be decided to operate the faulty system with downgraded performance, that is, with reduced torque, and a possibly pulsating torque.
Nonetheless, redundancy is an important factor in functional safety. As can be concluded from the options mentioned previously, the redundancy can be achieved either with hardware or with software. In either case, the system can still fulfill its role - either with full or with downgraded performance. In the automotive industry, the most interesting option is likely to not implement redundant hardware, but to operate the system at reduced performance after a fault has occurred. Then, the system can be replaced during maintenance.

In practice, the requirements on the fault tolerance of the system heavily depend on the necessity of the system. A system is always required to have a minimum operational life, and it can be required to be fail-operational, fail-safe or fail-silent. The necessity of the system determines how severe the requirement on fault tolerance is.

### 4-1-3 Fault-tolerant permanent magnet synchronous machine design

For a permanent magnet synchronous machine, a phase has to be electrically, magnetically, and thermally isolated from other phases in order for faulted phases to have a marginal effect on other phases. This is achieved by designing a permanent magnet machine with single-layer concentrated windings, with a spacer tooth in between phases to avoid direct contact between phases and a 1 per unit phase inductance. Furthermore, each phase is switched through a full bridge converter. In this design, phases are electrically, magnetically and thermally isolated [38–40]. If a short circuit in the switches occurs, the fault current will be limited to 1 per unit as the phase inductance is 1 per unit.

### 4-2 Fault tolerance of the power network

Aside from failure of the safety critical systems themselves, the power supply to these system can also fail. This section discusses several measures that have already been taken, and possible solutions that could address other faults.

#### 4-2-1 Possible solutions to increase fault tolerance

Several measures have already been taken to decrease the failure rates of several components. From the analysis in chapter 3, it was clear that the event of undervoltage due to a broken connection, a short circuit or insufficient supplied power are the most probable faults in the power network. Therefore, solutions are discussed for each of these three faults. An overview of the possible solutions is given in table 4-1.

A redundant set of cables with switches, to switch between the cable that is being used, could supply power in the case of a broken connection or a short circuit. The downside is this solution does not prevent a lack of generated power by 48 V components, and an extra set of cables in a car is a relatively heavy and costly solution to implement.

A redundant energy supply can provide energy to the grid in case of lack of energy in the grid. This could be both a supercapacitor or a lithium ion battery, that is connected when necessary. The former can only supply power for a short time, but this time could be enough to bring the car in safety. Obviously, this depends on the rating of the capacitor.

Currently, the braking and steering systems in a car require a power of around 4 kW
when both systems are fully active. In the best case, a reasonable time to switch two lanes and put the car to a standstill is around 10 seconds. This is the minimum requirement by Audi for functional safety. During this period, both systems are not continuously in use, but it is assumed that they are in use for half the time. This would require an energy saved of at least 20 kJ. The energy stored in a capacitor between two different voltages is given by:

\[ E = \frac{1}{2} \cdot C \cdot (V_{\text{high}}^2 - V_{\text{low}}^2) \]  

(4-1)

If it is assumed that \( V_{\text{high}} \) is 48 V, and \( V_{\text{low}} \) is 24 V, which is the start of the undervoltage range, and the capacitance module has an efficiency of 100 %, the capacitance module is required to have around 25 F. Due to the high voltage however, several capacitors have to be placed in series. The cell voltage normally lies around 3 V, which means the solution would require around 16 individual cells in series with a rating of 400 F each. Therefore, the module is expected to be relatively bulky. Furthermore, it does not provide a solution for a broken connection or a short circuit.

A combination of both a redundant energy supply, and a redundant set of cables, covers all possible faults under investigation. Of course, in order to be able to switch between the feeders, switches have to be implemented as well. This system comes at a much higher cost due to the redundancy for both systems.

Furthermore, it could be possible to change the design of the DC/DC converter such that critical loads on the 48 V side can be powered by the 12 V battery. Whether the 48 V critical loads perform well enough, has to be investigated.

In the current design, this connection is not possible. Furthermore, the energy management system of the 48 V network, which controls the DC/DC converter, takes too long to react on changes in the system. Within this time, the voltage has already dropped to such a low level that a safety mechanism trips and shuts the DC/DC converter down to protect itself. This solution does not protect against broken connections, short circuits or a breakdown of the battery.

In addition, a solution could be to connect the 12 V network with the critical loads on the 48 V network through a switch. For this solution to be viable, the performance of these safety-critical 48 V systems at 12 V has to be investigated. Furthermore, the battery on the 12 V side has to be able to deliver this power. The solution requires only a switch between 48 V and 12 V, and an extra connection, and therefore appears to be relatively cheap. With more switches, it could disconnect the 48 V network from the load, and protect against short circuits in the network.

From the overview, it follows that a connection between 12 V and 48 V through a switch appears to be the most promising. Not only are its costs low compared with other possibilities, it can also be very small and light. This solution will therefore be investigated further.

### 4-3 Proposed solution

The Fault Tree Analysis (FTA) from chapter 3 determined that the most frequently occurring fault, is a disconnect or failure of the battery. The BSG will then be disconnected, as it cannot regulate the voltage to the highly dynamic load. This will lead to an undervoltage in the network. Therefore the proposed solutions will consider this fault.
Table 4-1: Overview of possible solutions for a power network related fault.

<table>
<thead>
<tr>
<th>Solution</th>
<th>Broken connection</th>
<th>Short circuit 48 V</th>
<th>No power</th>
</tr>
</thead>
<tbody>
<tr>
<td>Redundant feeder</td>
<td>✓</td>
<td>✓</td>
<td>x</td>
</tr>
<tr>
<td>Redundant supply</td>
<td>x</td>
<td>x</td>
<td>✓</td>
</tr>
<tr>
<td>Redundant feeder and supply</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
</tr>
<tr>
<td>DC converter let through</td>
<td>x</td>
<td>x</td>
<td>✓</td>
</tr>
<tr>
<td>12 V connection</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
</tr>
</tbody>
</table>

Furthermore, based on the analysis of the fault tolerance of the power network, the connection of the 12 V network with the 48 V network through a switch seems to be the most promising concept that could solve this problem. This concept could work in two ways, of which one is to connect the 12 V network directly to the safety-critical loads in the 48 V network. The other concept is to switch the 12 V network to the BSG, so that it can start the motor and then generate energy. The latter case does not solve a broken connection or short circuit fault. For both methods the different possibilities for the type of the switch, its location and the topology are now analysed.

4-3-1 Type of switch

Important factors for deciding the switch topology are cost, volume and weight, as was also important for deciding the solution in general. Self-evident, the proposed solution should contribute to functional safety and should not make the system less reliable. The types of switches under consideration are mechanical relays, solid state relays, and MOSFETs. Diodes would be the easiest option to implement, but the voltage drop across the diode of 0 to 0.7 V causes problems. With a current ranging from 100 A to 300 A or higher, this could lead to a power loss of 30 to 210 W. This requires a large heat sink, which would make the solution bulky.

Mechanical relays appear not be an interesting solution either. On one side, the switching time is relatively high with for example a switching time of < 10 ms according to its datasheet [41]. Furthermore, as the mechanical relay has moving contacts, there is a chance of an arc occurring inside the relay during switch off if the switching voltage is higher than 15 V.

The remaining solution is using a semiconductor type switch, e.g. parallel MOSFETs or a solid state relay. In general, semiconductor switches can react very quickly and have a very low conducting resistance, so that there is only a small voltage drop across the switch. Therefore, losses in the switch are low and a large heat sink is not required. In addition, there are no moving contacts, and thus arcing cannot occur.

Using semiconductor switches also has several downsides. The currents can be very high, and therefore a couple of switches have to be placed in parallel to meet the current rating. Furthermore, MOSFETs have an internal body diode that has to be taken into account for the choice for a topology. In addition, the semiconductor switches have to be able to block the maximum possible voltage between 48 V and 12 V, even in the case of transients.

To conclude, multiple parallel semiconductor switches are the best application for this solution. At the time of writing, Audi was using a switching module called 'Q-Diode', after quasi-diode, that comprised of multiple parallel MOSFETs that could switch high currents. The Q-Diode module consisted not only of parallel MOSFETs and a driver circuit, but it is...
30 Overview of solutions for identified faults

Figure 4-1: Different possible connection points in the 48 V iHEV topology.

actually an embedded system of many components, including microcontrollers, transistors, capacitors and resistors. In the car, the Q-Diode is used in the 12 V network to protect the 12 V network during motor start. Because these switching modules were available and met the previously mentioned requirements, these switching modules have been used in the further research. There were two types available, one was controlled analogue and the other by the CAN-bus.

4-3-2 Location of switch

The location of the switch between the 48 V and 12 V voltage levels was to be chosen such that it comes at the lowest cost, size and weight. There were three possible placed identified in the 48 V network, which are shown by the numbers 1, 2, and 3 in figure 4-1. The fuses, that are normally present between the network and their loads, have been shown for the steering and braking system.

Connection point number one is between the fuse and the load. The benefit is that it can still operate the load if the fuse is blown. This can be beneficial if the fuse is blown unintentionally. Most likely, the fuse will blow to a short circuit in the load so that the added benefit is only minor. In addition, it is possible to disconnect the 48 V network in case of a short circuit in the network, either by blowing the fuse or by actively disconnecting the network with switches. However, this solution requires the double amount of switches, which makes it relatively expensive.

Connection point two is at a load distributor, the closest common point of both critical loads. The benefit of these connection point is that it offers most of the advantages of connection point one, with reduced costs. It can power both critical loads with one connection point, it can power the loads in most cases of a broken connection (as long, of course, this happens between connection point two and the grid), and it could also protect against short circuits, depending on the switching topology.
Table 4-2: Overview of characteristics of the different connection points between the 12 V and 48 V networks.

<table>
<thead>
<tr>
<th>Location</th>
<th>Problem</th>
<th>Occurrence</th>
<th>Solvable</th>
<th>Remarks</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Short circuit 48 V</td>
<td>≈ 3 – 10 FIT</td>
<td>≈ 95%</td>
<td>Switch off of 48 V required</td>
</tr>
<tr>
<td></td>
<td>Loose connection</td>
<td>≈ 4.5 FIT</td>
<td>≈ 95%</td>
<td>N.A.</td>
</tr>
<tr>
<td></td>
<td>No power</td>
<td>High</td>
<td>✓</td>
<td>N.A.</td>
</tr>
<tr>
<td>2</td>
<td>Short circuit 48 V</td>
<td>≈ 3 – 10 FIT</td>
<td>≈ 75%</td>
<td>Switch off of 48 V required</td>
</tr>
<tr>
<td></td>
<td>Loose connection</td>
<td>≈ 4.5 FIT</td>
<td>≈ 60%</td>
<td>N.A.</td>
</tr>
<tr>
<td></td>
<td>No power</td>
<td>High</td>
<td>✓</td>
<td>N.A.</td>
</tr>
<tr>
<td>3</td>
<td>Short circuit 48 V</td>
<td>≈ 3 – 10 FIT</td>
<td>×</td>
<td>Possible with extra switches</td>
</tr>
<tr>
<td></td>
<td>Loose connection</td>
<td>≈ 4.5 FIT</td>
<td>≈ 25%</td>
<td>N.A.</td>
</tr>
<tr>
<td></td>
<td>No power</td>
<td>High</td>
<td>✓</td>
<td>N.A.</td>
</tr>
</tbody>
</table>

Connection point three is at a general point in the 48 V power network. It is again a cheaper solution than connection one. However, the distance to the load is higher, so that there will be a larger voltage drop over the wire to the load. Assuming the chance of a broken wire is equally distributed along the wire, this connection point will less likely protect against loose connections. This is also valid for short circuit protection, as disconnecting the short circuit in the 48 V network while still supplying power to the safety-critical systems is difficult.

To conclude, connection point two offers the best protection at lowest cost. The loss of sufficient supplied power could have a failure rate of more than 100 FIT, making it the most occurring failure. The three different connection points have been summarized in table 4-2, which clearly shows why connection point two is the most interesting one. The switching topology is now investigated.

4-3.3 Switch topology

Now that the connection point has been chosen, configuration of the switches is further investigated. There are multiple topologies possible, each having their own advantages and disadvantages. Each of these topologies and their advantages and disadvantages will be shortly discussed. The topologies are shown in figure 4-2, and table 4-3 gives a clear overview of which topology solves which problem.

The most simple topology with one switch, 4-2a, has the least amount of switches and is therefore the cheapest option. Furthermore, the body diode will start to conduct the moment the voltage at the load drops below the voltage at the battery minus the voltage drop over the diode. However, this solution makes the overall system less reliable in case of a short circuit in the 48 V network or in the load. The body diode will then start to conduct high currents.

A fuse in series with the Q-Diode will disconnect the 12 V network in case of a short circuit in the 48 V network, but a very high current will flow for a short time nonetheless. It is expected that the 48 V network requires high currents in the case of a connection of the 12 V network. Therefore, a fuse with a high current tolerance is required. Therefore, a short circuit in the 48 V network will still have great impact on the 12 V and must be prevented. A Q-Diode with a fuse in series is therefore no option.

The back-to-back MOSFET topology, 4-2b, solves the short circuit problem. At the
cost of an extra switch, a possible short-circuit current to the 48 V network or to the load is prevented. In the case of a short circuit, there is still no possibility to power to the load.

Two MOSFETs can also be placed as show in figure 4-2c, which offers a solution for each of the investigated faults. However, a short circuit at the load causes currents from both networks to start to flow. The fuse present at the 48 V side of the load will blow, but in the current design there is no fuse present at the 12 V side. Therefore, this option is not a good solution.

The topology of figure 4-2d has the same short circuit problem as the topology shown in figure 4-2a, because a current to the 48 V network can always flow when there is a short circuit in the 48 V network. Therefore, this topology is not interesting.

The topology shown in figure 4-2e is able to mitigate all the investigated problems. The two back-to-back connected MOSFETs can power the load in the case that there is no power, or that the connecting wire is broken. The extra MOSFET between the load and the 48 V network enables the topology to mitigate short circuits in the 48 V network. Furthermore, a short-circuit in the load will lead to a blown fuse on the 48 V network side, and in these situations the sole MOSFET will suffice.

The last topology shown in figure 4-2, figure 4-2f, may add extra reliability to the solution. It offers the same protection as the topology shown in figure 4-2e. However, in the case of a short circuit in the load, it could be more desirable to switch off the high current instead of letting the fuse blow. This comes at a relatively high cost of an extra switch.

To wrap up, topology 4-2b is the topology with most benefits regarding increased reliability at the lowest amount of costs possible. Furthermore increasing the amount of switches could lead to a possible additional reduction of 3 to 10 FIT, but this also comes at 50% costs. The problem with the highest chance of occurring, is the lack of enough power in the grid. Therefore, topology 4-2b is the most interesting one, and will therefore be investigated further.

<table>
<thead>
<tr>
<th>Topology</th>
<th>Short circuit</th>
<th>Lost connection</th>
<th>No power</th>
<th>Number of switches</th>
</tr>
</thead>
<tbody>
<tr>
<td>a</td>
<td>✔️</td>
<td>✔️</td>
<td>✔️</td>
<td>N</td>
</tr>
<tr>
<td>b</td>
<td>✔️</td>
<td>✔️</td>
<td>✔️</td>
<td>2N</td>
</tr>
<tr>
<td>c</td>
<td>✔️</td>
<td>✔️</td>
<td>✔️</td>
<td>2N</td>
</tr>
<tr>
<td>d</td>
<td>✔️</td>
<td>✔️</td>
<td>✔️</td>
<td>2N</td>
</tr>
<tr>
<td>e</td>
<td>✔️</td>
<td>✔️</td>
<td>✔️</td>
<td>3N</td>
</tr>
<tr>
<td>f</td>
<td>✔️</td>
<td>✔️</td>
<td>✔️</td>
<td>4N</td>
</tr>
</tbody>
</table>
Figure 4-2: Overview of different possible switching topologies.
Chapter 5

Validation of reduced voltage operation of electrical machines

This chapter discusses the concept of the operation of the investigated machine at reduced voltage, which is a synchronous hybrid excitation machine. The objective of this chapter is provide an model of the performance of the machine. First, the decision of the machine and its design are elaborated upon in section 5-1. The theory of the operation of the machine and its operating limits is discussed in section 5-2. Then, the method that has been used to determine the relevant parameters of the machine is described in section 5-3.

5-1 The belt-driven starter generator

5-1-1 Choice of the machine

Due to availability constraints, a Belt-driven Starter Generator (BSG) was tested. Unfortunately, it was not possible to test a 48 V braking or steering system at 12 V. Ideally, these systems could be tested at 12 V, so that test results would align well with what one concept of the proposed solution. These systems were still in a pre-development phase, and no working prototypes of these systems existed at the time of writing. Therefore, testing these systems was not possible.

The electronic roll stabilisation and the electronic turbocharger systems present in the iHEV topology could not be used for testing purposes either. These systems are programmed to automatically shut off at a lower voltage, so that the more vital systems in the 48 V network receive all the power that is available in the system. The software was written by a third party, which would make a change of the software very time-consuming and costly. Therefore, it was decided not to pursue this option.

It was possible to test two similar Belt-driven Starter Generators (BSG), where one operated at 48 V and one operated at 12 V. Both the design of the machine and its power electronics were equal, except for the control of the machines. The software of the power
electronics of one machine was changed so that one machine operated at 12 V. During the research, all the tests were carried out on these two machines and their attached power electronics.

The BSG inside a car can function as a generator to generate electricity, and as a motor to start the engine. With a BSG it is possible to quickly start the engine to come back from sailing mode or as part of a start-stop system. A car is in sailing mode when the engine is disconnected during driving, which is also called coasting. By turning off the engine when it is not necessary, even while driving, fuel efficiency can be improved.

5-1-2 The design of the machine

Figure 5-1 shows a similar machine as the one investigated. The pulley of the machine is connected with the motor through a rubber belt. The field windings are enclosed by the two claw poles and they are electronically connected through slip rings. These slip rings are enclosed by the power electronics. Between the claw poles on the rotor, permanent magnets have been added.

The machine that has been investigated is shown in figure 5-2. The machine is built in a similar way as shown in figure 5-1, but it is closed as it is cooled with coolant instead of air. The integrated power electronics at the back side of the BSG can be easily identified. The BSG is controlled through the connectors, of which CAN signals can be sent. The machine is supplied with power through the connection point on the right side below the cover of the coolant tubes in figure 5-2b.

The machine is a dual-three phase hybrid excited claw pole synchronous machine with combined delta-wye windings. Due to the presence of both magnets and a field winding, the machine is a hybrid excited synchronous machine. Furthermore, the machine is of a claw pole type. In addition, the stator consists of two three-phase windings that are electrically shifted.
by 30°, and each phase winding set comprises of combined delta-wye three-phase windings.

The flux path of the claw-pole synchronous machine is shown in figure 5-3. In this schematic, the magnets are omitted. The field winding induces flux, which travels to the stator via one claw pole. There, the flux travels to the stator through the air gap, after which the flux travels back to the other claw pole via the air gap.

On the rotor of the investigated machine, magnets are placed between the claw poles. The flux path from these magnets is parallel to the flux path of the field winding when the field is excited, and the magnets increase the flux and the torque output of the machine. This is illustrated in figure 5-4. When the rotor field winding is not excited, the flux of the magnet passes through the rotor. However, when the rotor field is excited, the iron of the rotor saturates and the flux of the magnet is forced through the stator. Now the flux of the magnet is superimposed on the flux of the field winding, leading to an increase in flux and torque. When the polarity of the field of the rotor field winding is reversed, the magnet and field winding flux will oppose each other. For more information, the reader is referred to [43] and [28].
Validation of reduced voltage operation of electrical machines

Figure 5-4: A simplified overview of the magnetic flux path of the magnets inside the rotor. When the field is not excited, the magnet flux passes through the rotor (dotted red line). When the field is excited, the flux of the magnet no longer passes through the rotor, but through the stator (solid red line).

The combination of a delta-wye winding provides several benefits, leading to an overall higher efficiency of the machine [45–49]. The combined delta-wye winding is illustrated in figure 5-5. Normally, a distributed winding is used to reduce the harmonic content of the magnetomotive force. With the combined winding harmonic content is reduced, so that a double-layer winding is no longer necessary and a single-layer winding can be used [45]. In addition, the winding reduces the ohmic losses of the machine. This leads to either a higher efficiency, or a reduction of material usage. Furthermore, an improvement of the heat distribution is reported [46] and according to [47] there are only low no-load losses due to the circulating current in this case. The articles [45–49] provide more information on the combined delta-wye winding and its benefits.

The dual three-phase design of the machine also has several benefits, of which the most interesting are a lower required converter rating and a higher fault tolerance [50]. Due to the dual three-phase design of the machine, the current of the machine is split across a higher number of phases, which decreases the required converter rating. In addition, the machine is fault tolerant due to an extra phase winding set. In the case of a concentrated winding distribution, higher stator current harmonics can be injected to further increase the torque as proposed by [51]. Furthermore, the dual three-phase machine has a lower torque ripple than a similar three-phase machine. For more information including review on multiphase electric machines, the reader is referred to [50].

5-1-3 Modelling the dual-three phase hybrid excited machine

In order to create a model for the performance of the machine, the machine is analysed in different steps. First, a model in the \( d,q \) reference frame is constructed for the dual-three phase hybrid excited machine. Then, this model is used to analyse the operating limits of the
Figure 5-5: An example of the combined delta-wye winding.

Figure 5-6: The inverters of the dual three-phase machine, based on [52]. The dc link is shared by each inverter, and each phase is fed by its own inverters.
Validation of reduced voltage operation of electrical machines

From the model and measured parameters of the machine, the performance of the machine can be modelled.

First of all, the combined delta-star winding is transformed to a single star for the analysis of the machine. The delta windings can be transformed to a star by multiplying the resistances of the adjacent windings, and dividing the value by the sum of the total winding resistances. By adding the windings in series, an equivalent star winding can be obtained.

From previous work on these types of machines [53], it is expected that the machine is salient and that $L_d > L_q$ when the machine is not deeply saturated. Therefore, a model of the dual three-phase machine in the $d, q$ reference frame will be constructed.

The modelling of a dual three-phase machine is very similar to the modelling of a classical three-phase machine. For the modelling of a dual three-phase machine, two different approaches can be identified [28, 54–56]. One method considers the machine as two three-phase machines that are magnetically coupled. Applying the classical transformations to the dual three-phases results in a model with magnetic coupling. This magnetic coupling between the both three-phase winding sets makes the model difficult to analyse.

The other approach considers the machine as a polyphase machine with equally distributed phases. Applying an orthonormal transformation in this case introduces the classical coupling between the $d, q$ reference frame, and no coupling between the $d, q$ reference frame and the remaining $z_1, z_2, z_3, z_4$ reference frames. As there is no coupling between the both phase winding sets, it is easier to use this model for analysis. Therefore, a model has been constructed using the latter approach. The full construction of the model is done in appendix A. Here, for readability only the results of the analysis is shown. For the model, it is assumed that:

- The three-phase winding sets in the stator are identical and shifted electrically by an angle of $30^\circ$
- The magnetomotive forces (MMF) are sinusoidally distributed
- Only the first space harmonics are considered, and the mutual inductances are characterized by their fundamental component alone
- Magnetic saturation is not regarded

The approach leads to a model in the referential $dqz_1z_2z_3z_4$ frame. The model is constructed in appendix A. From appendix A, the voltage equations are decomposed in different subspaces and the voltage equations for these subspaces are:

$$
\begin{bmatrix}
v_d \\
v_q
\end{bmatrix} = R_a \begin{bmatrix}
i_d \\
i_q
\end{bmatrix} + \frac{d}{dt} \begin{bmatrix}
\lambda_d \\
\lambda_q
\end{bmatrix} + \begin{bmatrix}
0 & -\omega_e \\
\omega_e & 0
\end{bmatrix} \begin{bmatrix}
\lambda_d \\
\lambda_q
\end{bmatrix}
$$  \hspace{1cm} (5-1)

$$
\begin{bmatrix}
v_{z_1} \\
v_{z_2} \\
v_{z_3} \\
v_{z_4}
\end{bmatrix} = R_a \begin{bmatrix}
i_{z_1} \\
i_{z_2} \\
i_{z_3} \\
i_{z_4}
\end{bmatrix} + L_\sigma \frac{d}{dt} \begin{bmatrix}
\lambda_{z_1} \\
\lambda_{z_2} \\
\lambda_{z_3} \\
\lambda_{z_4}
\end{bmatrix}
$$  \hspace{1cm} (5-2)

$$
v_f = R_f i_f + \frac{d}{dt} \lambda_f
$$  \hspace{1cm} (5-3)
The equations for the flux linkages are

\[
\begin{bmatrix}
\lambda_d \\
\lambda_q \\
\lambda_f
\end{bmatrix} =
\begin{bmatrix}
L_d & 0 & \sqrt{3}M_{SF} \\
0 & L_q & 0 \\
\sqrt{3}M_{SF} & 0 & L_f
\end{bmatrix}
\begin{bmatrix}
i_d \\
i_q \\
i_f
\end{bmatrix} +
\begin{bmatrix}
\sqrt{3}\lambda_{pm}
\end{bmatrix}
\tag{5-4}
\]

where \(L_\sigma\) is the leakage inductance, and \(L_d\) and \(L_q\) are given by

\[
L_d = L_\sigma + 3M_{ss} + 3M_{sfm}
\]
\[
L_q = L_\sigma + 3M_{ss} - 3M_{sfm}
\tag{5-5}
\]

where \(M_{ss}\) and \(M_{sfm}\) are both terms that describe the magnetic coupling between the two three-phase windings. As can be seen from the equations, only the \(d,q\) currents contribute to the electromechanical energy conversion. The equations in the \(z_1z_2z_3z_4\) subspace contribute only to the losses in the machine [56]. Therefore, the dual three-phase machine can be considered as a three-phase machine in a \(dqz_1\) subspace. From appendix A, the torque of the machine is given by:

\[
T = p\left(\sqrt{3}i_q(M_{SF}i_f + \lambda_{pm}) + (L_d - L_q)i_di_q\right)
\tag{5-6}
\]

If the machine is non-salient, the right part of the equation disappears due to \(L_d = L_q = L\), and the torque is given by

\[
T = p\sqrt{3}i_q(M_{SF}i_f + \lambda_{pm})
\tag{5-7}
\]

Now that the theoretical background of the machine is known, the operating limits of the machine will be analysed.

### 5-2 Operating regions of the machine

In order to model the performance of the machine, the operation limits of the machine are investigated. The maximum available value of voltages and currents restrict the performance of the machine to specific operating regions. The current rating of the inverter and the thermal rating of the stator windings can further limit the operation of the machine. In this analysis, it is assumed that the machine operates in steady state with a supply of sinusoidal currents and voltages with constant amplitudes and frequency. Furthermore, at first it is assumed that the current of the field winding is at its maximum value and remains constant. Then, the case of decreasing the field winding current is investigated.

The limit of the operating region of the machine in the \(d,q\) reference frame due to the maximum voltages and currents can be expressed as

\[
I_d^2 + I_q^2 \leq I_N^2
\tag{5-8}
\]
\[
V_d^2 + V_q^2 \leq V_N^2
\tag{5-9}
\]

where \(I_N\) respectively \(V_N\) are the maximum current and voltage. First, the operating limits of the non-salient machine are discussed, followed by the operating limits of the salient machine.
5-2-1 Non-salient machine

The analysis will be done for the machine in case of salient and non-salient behaviour. For the non-salient machine \( L_d = L_q = L \), and therefore the reluctance component in the torque equation of 5-6 disappears and the expression for the torque becomes equation 5-7. The following voltage equations can be written down for the voltage for the non-salient pole machine in steady state:

\[
\begin{align*}
V_d &= RI_d - \omega_e LI_q \\
V_q &= RI_q + \omega_e LI_d + \omega_e \lambda_{\text{rotor}}
\end{align*}
\] (5-10)

where \( \lambda_{\text{rotor}} = M_s F I_f + \lambda_{\text{pm}} \). For high speed operation, the ohmic voltage drop can be neglected as \( \omega_e L \gg R \). Then, the equations 5-10 and 5-11 can be inserted in 5-9 and the following equation for the voltage limit circle is obtained after rearranging

\[
I_q^2 + \left( I_d + \frac{\lambda_{\text{rotor}}}{L} \right)^2 \leq \frac{V_N^2}{(\omega_e L)^2}
\] (5-12)

From this it follows that the voltage limit of the machine is given by a circle with a radius less than or equal to \( V_N^2/\omega_e L \). The centre of this circle is given by

\[
\begin{align*}
I_d &= -\frac{\lambda_{\text{rotor}}}{L} = -\frac{M_s F I_f - \lambda_{\text{pm}}}{L} \\
I_q &= 0
\end{align*}
\] (5-13)

This has been graphically shown in figure 5-7. The voltage limit ellipse is shown for various speeds, and it can be seen that the ellipse decreases as the speed increases. The current limit region is shown as a circle with radius \( I_N \) with its centre at \( I_q = I_d = 0 \). The limit is shown both for continuous operation, and a temporary operating limit for a higher current.

The torque of the machine is a straight line that is only dependent on the field winding current, the flux from the permanent magnets and current \( I_q \), as shown by equation 5-7. The base speed of the machine is reached when the speed can no longer increase without decreasing the torque of the machine. At the speed at which base speed is reached, the torque is still at its maximum value. Therefore, the current \( I_q \) is maximum, thus \( I_q = I_N \) and \( I_d = 0 \). By rearranging equation 5-12 the base speed of the machine can be obtained by

\[
\omega_{e,\text{base}} = \frac{V_N^2}{\sqrt{(LI_N)^2 + (\lambda_{\text{rotor}})^2}}
\] (5-15)

Furthermore, the machine may have a maximum speed as it is assumed that the field winding remains constant at its maximum. At this speed, the right side of the voltage limit circle only just crosses the left side of the current limit circle. At this point, the current \( I_q = 0 \). After rearranging equation 5-12, the following formula can be obtained for the maximum speed of the machine

\[
\omega_{e,\text{max}} = \frac{V_N}{\lambda_{\text{rotor}} - LI_d}
\] (5-16)

Until now, the machine’s capability to reduce the field current was not included. In practice, the machine will reduce the field winding current to further increase the maximum speed. The effect of a decrease in field winding current for a non-salient machine can be seen in 5-10,
where the speed is assumed to be constant for different values of field winding current.

A reduction in the field winding shifts the centre of the voltage limit circle to the right. As the flux of the rotor consists of the flux due to the field winding and the permanent magnet, \( \lambda_{\text{rotor}} = M_s F I_f + \lambda_{\text{pm}} \), a decrease in field winding current leads to a decrease in rotor flux. It is known that the centre of the voltage limit circle is given by equations 5-13 and 5-14. From these equations it follows that the more the field current is decreased, the more the centre of the voltage circle shifts to the right. At some point, the centre of the voltage limit circle is inside the current limit circle. Then, assuming an ideal machine without losses, the maximum speed of the machine is infinite. Assuming the speed stays constant, a decrease in rotor flux does not lead to a change in the voltage limit circle due to equation 5-12. However, it does lead to a decrease in torque as can be seen from equation 5-7.

5-2-2 Salient machine

A similar analysis will now be done for the machine in the case of a salient machine. For a salient machine, \( L_d \neq L_q \), and therefore the following equations are valid for the voltages:

\[
V_d = RI_d - \omega_L q I_q \\
V_q = RI_q + \omega_L d I_d + \omega_L \lambda_{\text{rotor}}
\]

where \( \lambda_{\text{rotor}} = M_s F I_f + \lambda_{\text{pm}} \). For high speed operation, the ohmic voltage drop can again be neglected. The following equation is obtained for the voltage limit, by inserting 5-17 and 5-18 in 5-9 and rearranging

\[
\left( \frac{L_d I_q}{L_d} \right)^2 + \left( I_d + \frac{\lambda_{\text{rotor}}}{L_d} \right)^2 \leq \frac{V_N^2}{(\omega_L L_d)^2}
\]
Figure 5-8: The operating regions of a non-salient machine, for different values of field winding current. The speed of the machine is constant, and therefore the radius of the circle is constant. In addition, a decrease of field winding current leads to a decrease in torque due to equation 5-7.

Contrary to the non-salient machine, the voltage limit of a salient machine is given by an ellipse instead of a circle. As expected, the voltage limit of the machine is speed dependent. Furthermore, the centre of this voltage limit ellipse is given by

\[ I_d = \frac{-\lambda_{\text{rotor}}}{L_d} = \frac{-M_{sF} - \lambda_{\text{pm}}}{L_d} \]  (5-20)

\[ I_q = 0 \]  (5-21)

This can be shown graphically as in figure 5-9. The voltage limit ellipse is shown for various speeds, and it can be seen that the ellipse decreases as the speed increases. The current limit region is again shown as a circle with radius \( I_N \) with its centre at \( I_q = I_d = 0 \). Due to equation 5-6 for the torque, equal torque lines are hyperboles with asymptotes \( I_q = 0 \) and \( I_d = \lambda_{\text{rotor}}/(L_q - L_d) \).

The torque of the machine, as given by equation 5-6, is dependent both on \( I_d \) and \( I_q \). When the machine has reached its base speed, the voltage limit ellipse of the machine will decrease further when speed increases. As visible from the graph, it can be seen that the torque then starts to decrease. The base speed is given by the speed at which the torque is still maximum, and the voltage limit is reached. By rearranging 5-19, the following equation is obtained for the base speed

\[ \omega_{c,\text{base}} = \frac{V_N^2}{\sqrt{(L_q I_q)^2 + (L_d I_d + \lambda_{\text{rotor}})^2}} \]  (5-22)

Furthermore, it was assumed that the field current is again constant at its maximum value. Then, the maximum speed of the machine may reached when the voltage limit limit ellipse crosses
the current limit circle. At this point, $I_q = 0$. Therefore, the left side of equation 5-19 is zero, and the maximum speed of the machine can be calculated by rearranging the rest of the equation

$$\omega_{e,\text{max}} = \frac{V_N}{\lambda_{\text{rotor}} - L_d I_d}$$  \hfill (5-23)

The flux weakening capability of the machine was neglected, as it was assumed that the field winding current was at maximum and constant value. In practice, however, the machine will reduce the field winding current at higher speeds. The decrease of field current is shown graphically in figure 5-10, where the speed is assumed to be constant for these different values of field winding current.

A reduction in the field current again shifts the centre of the voltage ellipse to the right in the $I_d, I_q$ plane. As the flux of the rotor consists of the flux due to the field winding and the permanent magnet, $\lambda_{\text{rotor}} = M_s F_I + \lambda_{\text{pm}}$, a decrease in field winding current leads to a decrease in excitation flux. As mentioned before, the centre of the ellipse is given by equations 5-20 and 5-21. From these equations it is clear that when field current is decreased more, the centre of the voltage ellipse shifts more to the right, as can be seen from equation 5-20. If the centre of the ellipse is inside the current limit circle, then in this case the maximum speed of the machine is also infinite for an ideal machine. In addition, from 5-6 it can be seen that a decrease in field winding current will lead to a lower torque.

![Operating regions of a salient machine](image)

**Figure 5-9:** The operating regions of a salient machine. It is assumed that the field winding current remains constant.
46 Validation of reduced voltage operation of electrical machines

\[ I_d, I_q, I_{N, continuous}, I_{N, max}, \omega = \omega_b, T_N \]

\[ T < T_N - T_N \]

Generator operation
Motor operation
Decreasing \(-\lambda_{rotor}/L_d\)

\[ \text{Figure 5-10: The operating regions of a salient machine, for different values of field winding current. The speed of the machine is constant, and therefore the radius of the circle is constant. In addition, a decrease of field winding current leads to a decrease in torque due to equation 5-6.} \]

5-3 Determination of machine parameters

In order to approximate the performance of the machine, it is necessary to experimentally determine the aforementioned parameters. Therefore, a third machine was disassembled in order to have access to the phases directly. The machine was identical to the other two.

The direct and quadrature axis inductances have been determined using the voltage step response of the machine. The machine was connected with a voltage source through a switch, and the voltage and current were measured with a LeCroy WaveJet 324A oscilloscope. Furthermore, the resistance of the stator windings was also determined. As the time constant of the current rise time was known, it was possible to calculate the inductance using

\[ \tau = \frac{L}{R} \]  

(5-24)

where \( \tau \) is the time the current reaches \( 1 - e^{-1} \) of its maximum value. The measurements were done when the field winding was not excited. First, the rotor was aligned to either the direct or quadrature axis, and then the measurement was carried out.

The other methods to determine the direct and quadrature axis inductance were not possible or very time consuming. Due to the lack of presence of equipment, it was not possible to determine the parameters through a slip test. Determining the direct and quadrature inductances is also possible by performing a load angle test. During the test, the machine is connected to an ohmic load, and will be driven at multiple speeds for different field winding currents. During the test, phase currents, phase voltages, the phase shift between the currents and the voltages, and the load angle are measured. Furthermore, an open circuit measurement
has to be performed to determine the induced EMF per phase at each operation point. With the measured variables, the inductances can then be calculated. Even though a load angle test could have given a better overview of the inductances at different operation points, it would have taken a very long to perform these measurements. At this point in time of doing the research, time was limited. Therefore, it was not decided to carry out a load angle test.

Furthermore, a Standstill Frequency Reponse (SSFR) test was performed. However, the results proved to be inaccurate and impractical for usage in a model. At low frequencies the inductance became negative due to measurement inaccuracies, at which the results can no longer have a physical meaning. Therefore, the step response measurement was carried out and these results were used in the model.

From the average of five step response measurements, the values $L_d \approx 100 \, \mu H$ and $L_q \approx 160 \, \mu H$ have been obtained. The measurements were carried out with an error margin, as the rotor had to be kept still manually. Furthermore, the beforehand determined stator resistance varied slightly during measurements. Therefore, the obtained values are to be considered with an error margin.

The obtained results are contrary to expectations, as it is expected that $L_d > L_q$. The air gap along the $q$-axis is larger than along the $d$-axis, and therefore the reluctance is higher along the $q$-axis. Reluctance is given by

$$ R = \frac{l}{\mu_0 \mu_r A} \quad (5-25) $$

where $l$ is the length, $\mu_0$ is the permeability of vacuum, $\mu_r$ is the permeability of the material, and $A$ is the cross-sectional area. Indeed, if the length of the air gap is higher, the reluctance $R$ should be higher.

If the reluctance is higher, the inductance is lower along the $q$-axis, as the inductance is given by

$$ L = \frac{N^2}{R} \quad (5-26) $$

where $N$ is the number of turns. In addition, also from previous work [53] it follows that $L_d > L_q$. The reason that $L_d < L_q$ was measured is due to the fact that the rotor field winding was not excited during the measurement. In this case, most of the magnet flux flows through the rotor, and not through the stator, and superimposes a flux in the opposite direction of the flux generated by the step response. Therefore, the obtained value for $L_d$ is not correct, as the inductance is given by:

$$ L = \frac{N \Phi}{i} \quad (5-27) $$

From this equation it is clear that when less net flux flows through the rotor, the obtained value for $L_d$ is lower than when the flux of the permanent magnet does not oppose the flux through the stator.

The inductance $L_d$ however, has been determined more accurately. As the reluctance of the rotor is much lower than the reluctance of the air gap, it is assumed that all the magnetic flux of the magnets flows through the rotor and there is not much leakage flux. Therefore, the measurement of the $L_q$ is not influenced as much as the inductance $L_d$. As it was not possible to repeat the measurement, the same ratio between $L_d$ and $L_q$ as in [53] was been assumed and $L_d$ was estimated using the ratio of [53].
The flux generated by the permanent magnets and the stator - field mutual inductance have been obtained by an open circuit test of the synchronous machine. The connection diagram of this measurement is shown in figure 5-11. The phases of the machines were not connected, and the voltage was measured with a LeCroy WaveJet 324A oscilloscope. The machine was driven at its rated speed of 2,000 RPM, and the excitation voltage was varied from −60 V to 60 V in small incremental steps. Although the voltage range of −60 V to 60 V contains the operating range of the machine, in practice the voltage is actually not allowed to exceed 52 V. As a matter of fact, the voltage for motor operation in general is around 40 V. With this data, it is possible to calculate the excitation flux using the following equation [57]:

\[ E_0 = \frac{\omega_n}{\sqrt{2}} \lambda_{rotor} \]  

(5-28)

where \( E_0 \) is the induced RMS phase voltage at no load and \( \lambda_{rotor} \) is the flux that links with the stator windings of one phase. The magnitude of the EMF first harmonic is used for the calculation. The permanent magnet flux \( \lambda_{pm} \) is the flux when the field current is zero; the stator - field mutual inductance can be found by linearising the slope of the flux. A non-saturated and a saturated value for \( M_{sF} \) can be found. The excitation flux of the machine is shown in figure 5-12. For both cases, the operating point of the machine is indicated. As the battery voltage drops when a high current is demanded, an operating voltage of 11 V is expected in practice.

In the figure, the effect of the magnet parallel to the flux path as described in section 5-1-2 is clearly visible. When the field winding is excited, the flux of the magnet is superimposed on the flux of the field winding. This leads to a higher excitation flux. On the contrary, when the flux of the field winding opposes the flux of the magnet, the magnet weakens the excitation flux. Indeed, a much lower excitation flux is measured when the field opposes the magnet.

Now, the maximum value for \( I_N \) and \( V_N \) will be determined based on the rating of the machine. The maximum dc link current rating is \( I_{max} = 250 \) A, which is the total current for the inverters of three-phase winding sets. The maximum rating for the inverters of one
three-phase windings is therefore \( I_{\text{max}} = 125 \text{ A} \). Furthermore, the machine operates at a dc link voltage of 40 V. The phase voltages and currents are generated with Pulse Width Modulation (PWM). Therefore, for normal modulation and sinusoidal PWM the RMS line to line voltage is given by [57]:

\[
V_{LL,RMS} = \sqrt{3} \frac{(\hat{V}_{AN})_1}{2\sqrt{2}} = \sqrt{3} \frac{m_a V_{dc}}{2\sqrt{2}} \approx 0.612 m_a V_{dc} 
\]

(5-29)

where \((\hat{V}_{AN})_1\) is the fundamental frequency component that varies linearly with the amplitude modulation ratio for \(m_a < 1.0\), and \(V_{dc}\) is the dc link voltage. For square wave modulation, the voltage can be increased to:

\[
V_{LL,RMS} = \frac{\sqrt{3} 4 V_{dc}}{\sqrt{2} \pi} \approx 0.78 V_{dc} 
\]

(5-30)

If losses are neglected, the following can be written for the dc to ac inverter [57]:

\[
V_{dc} I_{dc} = 3 V_{ph} I_{ph} \cos \phi
\]

(5-31)

where \(\cos \phi\) is the phase angle by which the phase current lags the inverter phase voltage. If equation 5-31 is rearranged, the following equation can be written for the phase current

\[
I_{ph} = \frac{V_{dc} I_{dc}}{3 V_{ph} \cos \phi}
\]

(5-32)

Now, we calculated the root mean square value of \(I_{ph}\). From appendix A it follows that \(I_N = \sqrt{3} i_{ph}\). Figure 5-12 and the maximum value \(I_N\) for the currents \(i_d\) and \(i_q\) can be used to determine the maximum value of the torque both for normal operation and operation at reduced voltage. The real excitation flux of the machine was used for the model. Both electrical and mechanical losses in the machine are neglected.

The torque-speed curve can now be approximated using the values of the parameters, equations 5-6 and 5-22. An example of the result for square wave pulse width modulation is shown in figure 5-13, where it is assumed that no saturation takes place. In this figure, the performance of the machine is idealized and it is assumed that, as there are no losses in the machine, all electrical energy is converted to mechanical energy. For both 12 V as well as 40 V, the flux of the rotor is smaller than \(L_d I_d\). From equation 5-23 it follows that the maximum speed of the machine is infinite as the losses of the machine are neglected, as described in 5-2-2.

In practice, the maximum torque and speed of the machine is limited. The torque of the machine will be lower than the modelled torque, due to the saturation of the iron inside the machine as shown in figure 5-14. The iron saturation is not included in the model. When iron saturations, increasing the current does not increase the flux by as much. As the torque of the machine is dependent on the flux, increasing the current then also does not increase the torque as much.
Figure 5-12: The excitation flux of the investigated machine. The operating points of the machine, both at 40 V and when powered with a battery, have been indicated. Due to the fact that the battery voltage decreases with high torque outputs, an operating voltage of 11 V has been assumed. The asymmetry is caused by the magnets parallel to the flux path of the field winding.

Figure 5-13: Ideal torque-speed characteristic of the machine, operated at 40 V and at 12 V. In the ideal case, the machine has no maximum speed as explained in section 5-2-2 by equation 5-23, as $\lambda_{\text{rotor}} < L_d I_d$. 
Figure 5-14: Magnetization curves for three different types of steel, from [58].
Chapter 6

Measurements of real machine performance

The objective of this chapter is to describe the test set-up that has been used, and to discuss and explain the obtained measurement results. The test set-up will be discussed first in section 6-1. Then, the obtained results are discussed in 6-2. These results will then be evaluated and compared with model in section 6-3.

6-1 Test set-up

A test set-up available at Audi has been used to perform the measurements on the machine and its power electronics. A schematic overview of the set-up is given in 6-1. The shaft of the machine was connected to another machine, the prime mover. Furthermore, a supercapacitor was connected in parallel with the power supply, so that the peak power of the BSG in machine operation could be handled by the voltage supply. In addition, a load was connected in parallel to the machine in generator operation. The machines were cooled with coolant, and the temperature of the coolant was set at 90° Celsius, which is the normal temperature of the coolant in a car. The ambient air temperature was regulated at 25° Celsius. The devices used in the test set-up are listed in table 6-1, and the specifications of the used prime mover are given in table 6-2.

The current to the machine, the voltage across the machine, the revolutions per minute (RPM), and the torque of the machine were measured. Therefore, also the mechanical power delivered by the machine was known. In addition, the terminal voltage of the voltage supply was measured, and the current from the voltage supply. Furthermore, the temperature of the coolant to and from the machine, including the flow rate of the coolant, was measured.

In order to control the desired torque of the machine, a laptop was connected to the machine through a CANcaseXL interface using the programme CANoe, both produced by the company Vector. In this way, it was possible to demand a torque from the machine using a laptop. In this way different values for torque were demanded at different speeds. The torque was varied from −60 Nm to 60 Nm, while the speed was varied from 0 to 10,000 RPM.
Table 6-1: Components used for the machine measurements.

<table>
<thead>
<tr>
<th>Device</th>
<th>Type number</th>
</tr>
</thead>
<tbody>
<tr>
<td>Torque sensor</td>
<td>Kistler 4503A100</td>
</tr>
<tr>
<td>Current transducer</td>
<td>LEM Transducer, 1000A DC ±1A</td>
</tr>
<tr>
<td>Climate room</td>
<td>Vötsch VMV 06</td>
</tr>
<tr>
<td>Coolant pump</td>
<td>Huber Unistat 150W</td>
</tr>
<tr>
<td>Supercap</td>
<td>Maxwell BMOD0165 P048 BXX</td>
</tr>
<tr>
<td>Electronic load controller</td>
<td>Zentro Elektrik, type number unknown</td>
</tr>
<tr>
<td>Oscilloscope</td>
<td>YokoGawa WT3000</td>
</tr>
</tbody>
</table>

Table 6-2: The specifications of the prime mover.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Motor type</td>
<td>Siemens 1PH8107 Asynchronous Motor</td>
</tr>
<tr>
<td>Voltage rating</td>
<td>326 V</td>
</tr>
<tr>
<td>Current rating</td>
<td>38 A</td>
</tr>
<tr>
<td>Power rating</td>
<td>13,0 kW</td>
</tr>
<tr>
<td>Maximum shaft speed</td>
<td>18000 rpm</td>
</tr>
<tr>
<td>Maximum torque</td>
<td>170 Nm</td>
</tr>
</tbody>
</table>

Figure 6-1: The test setup used for measurements. The current to the machine, the voltage across the machine, the RPM of the shaft, the torque on the shaft, and the mechanical power were measured by the test setup. In addition, the terminal voltage of the voltage supply, the current from the voltage supply, the temperature of the coolant from and to the machine, including the flow rate of the coolant, were measured.


6-2 Measurement results

The objective of the measurement is to assess the performance of a machine at reduced voltage operation. One backup function is to power the safety-critical systems directly in case of a failure. For this application, it is interesting to investigate whether the safety-critical systems could still function sufficiently when they are powered by a 12 V supply. The other backup function is to power the BSG, so that the machine can start the Internal Combustion Engine (ICE). Then, the BSG can generate energy. In this case, it should be assessed whether it is possible to start a car engine with a 48 V BSG powered by a 12 V battery. To answer this question, both the required torque-speed characteristic by the engine and the delivered torque-speed characteristic of the BSG are required. For the first concept, ideally it would have been able to test these safety-critical systems. However, as only the BSG was available, the concept of starting the ICE with a BSG is investigated. The results, however, may provide insight in how safety-critical systems with a 12 V supply may perform.

The torque speed characteristic of the BSG for the different operating voltages is plotted in figure 6-2 respectively figure 6-3. These characteristics zoom in on the motor operation of the machine. After 5,000 RPM the output torque of the machine is restricted by its software. The purpose of the machine is to bring an ICE up to a high enough speed to start the engine. When the ICE has started, the BSG functions as a generator. In practice, in general a BSG does not exceed a speed of 3,000 to 4,000 RPM in motor operation. Therefore, the torque of the machine in motor operation is limited after 5,000 RPM in this case. In generator operation, the BSG is able to exceed a speed of 10,000 RPM. During operation at reduced voltage, the torque of the machine was limited at 1,250 RPM. Therefore, the torque of the machine is only shown for a maximum speed of 5,000 RPM respectively 1,250 RPM. This also applies for the mechanical power, the DC link current, the losses and the efficiency of the machines that are shown. All figures will be discussed.

The machines have been tested on three different voltage levels. The nominal voltage at which the test of the 48 V RSG was performed, was at 40 V. The RSG was tested at both 13.5 V, when the machine was connected to a voltage supply, and around 11 to 12.7 V, when the machine was connected to a lead acid battery. The battery was an AGM Varta 7P0 915 105 D battery, with a capacity of 68 Ah. Please note that lead acid batteries are characterized by a behaviour where the voltage of the battery drops as more current is demanded. In addition, the machine operation at 13.5 V is not as interesting as the machine operation when it is powered by a real battery. Therefore, the performance of the machine under operation at 13.5 V is not discussed here.

During the measurements, at some point the maximum torque capability of the machine at a specific speed has been reached. This is noticed when even though higher values of torque are asked, the machine was delivering the same value of maximum output torque. Then, the value of the previous measured quantities has been copied for higher values of asked torque that were not demanded yet. This has been done in order to save a significant amounts of time between measurements, as then it was not required to charge the capacitor parallel to the machine and voltage supply. The supercapacitor was required as the voltage supply was not rated for such a high power output.
Torque-speed characteristic

The machines show unexpected behaviour, at an operating voltage of 40 V as well as when the BSG is powered by a battery. The torque-speed characteristics are shown in figure 6-2 and 6-3 respectively. The behaviour of the machine at both operating voltages will be discussed separately.

At 40 V, the maximum torque of the machine below 1,000 RPM appears to be around 55 Nm, as the machine was unable to deliver more torque when demanded. In addition, the maximum torque of the machine decreases after around 800 to 1,000 RPM. The maximum torque of the machine is limited by the magnetic saturation of the iron inside the machine, and the heating limits of the machine.

The iron in the machine saturates as more current is flowing inside the machine, as shown in figure 5-14. After a certain point has been reached, more current will lead to no or only a slight increase in flux, and thus torque output. Further increasing the current contributes to much higher losses without significantly increasing the torque of the machine. Therefore, the control of the machine limits the torque output of the machine so as to prevent too high currents and losses when its not necessary.

In addition, the torque of the machine is limited by its stator heating limit. The machine is not able to output 55 Nm torque on a continuous basis. After a couple of seconds, the currents in the machine are reduced so as not to overheat the machine, and therefore its output torque is reduced. This can also be observed in figures 5-7 to 5-10, where the output torque indeed decreases when the currents in the machine go from the maximum current limit circle to the continuous current limit circle.

The field winding voltage was also measured, both for 48 V and when the machine was supplied by the battery. It appeared that the machine always regulates its field winding voltage to the maximum possible, as can be seen in figures C-2 and C-1. In other words, the machine keeps the field winding voltage constant and effectively operates as a permanent magnet synchronous machine. Furthermore, at 2,500 RPM or higher the inverters of the machine use square wave pulse width modulation. This was found out by measuring the gate voltage of the inverters during the measurements, which is shown in figures C-3 and C-4 in appendix C.

For speeds above 500 RPM, there is an offset between demanded and delivered torque. The difference between the demanded torque and the delivered torque is due to the control of the machine. Below 2,500 RPM the machine delivers less torque than demanded, while at speeds higher than 2,500 RPM the machine delivers a higher torque than demanded. The machine is controlled by its power electronics by referencing its stator current with a lookup table with the required values. The value for the current in the lookup table appears to be incorrect. The torque of the machine has been investigated for different values of terminal voltage and temperature of the coolant of the machine. These results are shown in appendix C, in figure C-5. From these figures it follows that the torque output of the machine, and its related offset, do not change for different temperatures or voltages at the terminals, as long as the machine is operating in sinusoidal PWM, i.e. below 2,500 RPM. The change from a negative offset to positive offset is due to the change of sinusoidal to square wave pulse width modulation.

At 12 V, the performance of the machine is much worse than expected. As previously discussed, it would be expected that the torque is almost as high as during normal operation at 40 V, until the machine operates at base speed. Obviously the base speed will be reached
The torque-speed characteristic of the machine, operated at 40 V. 

much quicker than during operation at 40 V due to the reduced voltage, as can be observed from 5-22. For operation at 12 V, the maximum torque is around 30 Nm. Furthermore, the machine stops operating as a motor after 1,250 RPM.

The worse than expected performance of the machine is due to the control of the machine. From measurement logs, it can be seen that the maximum dynamic torque that the machine is allowed to output is restricted. In figure 6-4 an example plot of the control signals of the machine is shown. From this plot, it can be concluded that the maximum torque of the machine when supplied by a battery is indeed restricted by the software.

The figure shows several signals of the machine, that the indicate values according the machine. These values differ from the real values due to control and inaccuracy. In the figure, the red line indicates the RPM of the machine which stays constant at 200 RPM. The orange line indicates the maximum torque that the machine is allowed to output, while the blue line indicates the maximum negative allowed torque that the machine is allowed to output. At this speed, the machine is not allowed to function as a generator and therefore, this value is zero. The yellow line indicates the demanded torque during the measurement, while the green line indicates the delivered torque of the machine. Please note that the axis of the green line has been shifted down by 5 Nm torque on purpose, as it would have not have been visible otherwise.

The restriction of the maximum torque also occurs at 1,500 RPM, at which the maximum dynamic torque of the machine is set to zero. Therefore, at this point the machine stops operating as a motor.
Figure 6-3: The torque-speed characteristic of the machine, powered by the battery.

**Mechanical power and dc link current**

The behaviour of the machine at 40 V is also visible in the graphs showing the mechanical power and current, figure 6-5a and 6-5c respectively. The maximum mechanical power and dc link current at 1,500, 1,800 and 2,000 RPM are indeed lower, as expected based on our knowledge of the torque of the machine. This is due to the relationship $P_{mech} = \omega_{mech} T$ between power and torque, and current and torque, given by equations 5-6.

The behaviour of the machine at 40 V up to the base speed is as expected. As the torque of the machine remains constant and maximum until the base speed is reached, the mechanical power increases proportional to the speed. This can be observed for the machine at 40 V up to 1,000 RPM.

The machine with a 12 V supply shows a similar continuous increasing mechanical power. The mechanical power and current for their respective demanded torque increase as the RPMs of the machine increase. In addition, the maximum power output of the machine appears to be around 1,500 W. This makes sense, because if the efficiency and the dc link current are assumed to be constant at the same RPM, then the decrease in the voltage drop will lead to a similar decrease in mechanical power output. Indeed, 12 V divided by 40 V multiplied with the around 5,000 maximum mechanical power output is around 1,500 W.
Figure 6-4: The control signals of the machine, that indicate that the performance of the machine is constrained by software. In the figure, the red line indicates the RPM, the orange line indicates the maximum torque for motor operation, the blue line indicates the maximum allowed torque for generator operation, the yellow line indicates the demanded torque during the measurement, while the green line indicates the delivered torque of the machine. Please note that the green line has been shifted down by 5 Nm on purpose to make it visible.
Measurements of real machine performance

(a) Machine mechanical power output, at 40 V.

(b) Machine mechanical power output, supplied by a battery.

(c) DC link current, at 40 V.

(d) DC link current, supplied by a battery.

Figure 6-5: Overview of mechanical power and dc link current of the machine at 40 V and when supplied by a battery.
Power loss and efficiency

The efficiency of the machine has been calculated using the following formula:

$$\eta = \frac{P_{\text{mech}}}{P_{\text{elec}}} \cdot 100\% \quad (6-1)$$

where $P_{\text{elec}}$ is calculated by multiplying $V_{\text{dc}}$ and $I_{\text{dc}}$, while the mechanical power was measured by the test set-up. The test set-up measured both the RPM of the shaft and the torque on the shaft. Therefore, the efficiency includes all losses of the machine, namely the mechanical losses, copper losses, iron losses and the losses of the power electronics.

The copper losses are the major part of the losses of the machine at low speeds. While the losses may seem too high, calculations show that such high losses can be possible. An analysis of the losses of the machine in appendix B shows that calculated losses resemble the losses from the measurements. For a stator current of 90 A and a stator winding temperature of 190°C, calculated losses are approximately 3,800 W. These losses include the stator copper losses, the rotor field winding losses, and the switching and conducting losses of the MOSFETs in the power electronics, although the stator copper losses are the major part of the losses.

Furthermore, it can be observed that there is a non-linear relation between the difference in losses between 50 and 55 Nm torque and the difference in losses between 10 Nm and 15 Nm. This is because the losses in the stator are given by $P = 3I_{\text{ph}}^2R_{\text{ph}}$, and saturation occurs, so that a larger increase in current is necessary to realize the same increase in torque. One might expect the losses to be low as the dc link is low. However, due to equation 5-32 there is no direct relation between the dc link current and the phase currents.

At 1,000 RPM, the maximum torque of the machine decreases. This is due to the voltage limit ellipse that limits the maximum currents on the machine, as given in figure 5-9. Due to the limit of this ellipse, the currents $I_d$ and $I_q$ are restricted to a lower value, and therefore the copper losses decrease. However, for higher speeds the total losses of the machine still increase due to the iron losses and the mechanical losses.

The efficiency of the machine operating at 40 V shows interesting behaviour. As mechanical power output at $\omega_m = 0$ is zero, the efficiency is zero. Furthermore, the efficiency at low torque values between 500 and 2,500 RPM is significantly lower. Sometimes, the machine actually delivered negative mechanical output. As higher torques were demanded, the phenomenon occurred less. In addition, the maximum efficiency of this prototype machine increases as demanded load increases, to a maximum of around 75%.

The losses and efficiency of the machine at 12 V over different speeds is more in line with expectations. It is worth noting that the efficiency of the machine at low RPMs is higher for operation at 12 V than at 40 for the same torque output. This is a consequence of the reduced torque output capability of the machine, reducing the current required, and therefore reducing the losses in the machine, and increasing the efficiency.

Now that the real behaviour of the machine has been discussed, both when supplied with a 40 V voltage source and with a 12 V lead acid battery, the practical application of the machine when supplied with 12 V is investigated. First, the possible performance is clarified using the model of section 5-3. Then, the starting of the motor is investigated, both for the current machine as for the modelled machine.
Losses at different speeds, with 40 V source

(a) Power loss, at 40 V.

Losses at different speeds, battery powered

(b) Power loss, supplied by a battery.

Efficiency at different speeds, with 40 V source

(c) Machine efficiency, at 40 V.

Efficiency at different speeds, battery powered

(d) Machine efficiency, supplied by a battery.

Figure 6-6: Overview of power loss and efficiency of the machine at 40 V and when supplied by a battery.
6-3 Interpreting the measurement results

6-3-1 Comparison of model and results

In order to make concluding remarks about the performance of the machine in a real application, it is necessary to look at the non-ideal performance of the machine. The non-ideal performance of the machine has been approximated by combining the efficiency of the machine with the constructed model of the previous chapter. This is plotted for different saturation levels in the same figure as the maximum torque of the machine from the measurements described in this chapter. The result is plotted in figure 6-7, for the case that the inverters of the machine are operating in square wave pulse width modulation. In this model the saturation has been approximated linearly, while in practice saturation is highly non-linear.

The model shows that the maximum torque of the machine when it is powered by a battery can be higher than the performance that was measured. In theory, even nearly the same torque of the machine should be possible at low speed. This is due to the fact that the field winding saturates relatively quickly. In practice however, the performance of the machine will be limited due to the maximum power output of the battery and the losses of the machine. The battery can only provide approximately 3 kW, as the current of the machine is limited at 250 A. Therefore, the mechanical power and the losses of the machine cannot exceed 3 kW.

In addition, the calculated maximum speed of the machine for which maximum torque is available is much higher in the model. Obviously, this is not the case in practice when losses of the machine are also considered.

The effect of saturation is also very visible, as a higher saturation increases the base speed of the machine and decreases the maximum torque. Indeed, from equation 5-6 and 5-22 it follows that a saturated inductance leads to a lower maximum torque and a higher base speed of the machine.

Now that both the real performance and a possible theoretical performance of the machine have been discussed, the practical application of the starting of the motor is investigated.

6-3-2 Starting the Internal Combustion Engine

The starting of an ICE engine is a somewhat complicated process, as the torque of the machine, the ramp of the torque of the machine and the speed of the ICE and machine are relevant. There is currently no model for the starting of an engine. The assessment of whether a BSG is able to start an ICE is always tested at a setup in practice, at different motor temperatures.

It was possible to estimate the required torque to start an ICE, based on data from a measurement that was carried out with a different BSG. In this measurement, the starting capability of a BSG was assessed for different motor temperatures, different gear ratios between the BSG and ICE, and different torque ramps of the BSG. Exactly at the moment the speed of the BSG starts to increase, the mechanical torque of the engine has been overcome. With this data, it was possible to estimate the required torque to start the engine. This data is summarized in table 6-3. An example of the results from testing the starting with a BSG is shown in figure 6-8.

While an ICE requires around 550 RPM for a smooth start, the engine starts igniting
Figure 6-7: Comparison of the ideal and real torque speed characteristics. It is assumed that the inverters of the machine are operating in square wave pulse width modulation. The dashed lines correspond to the modelled torque speed characteristic of the machine when it is powered by the battery. Only the saturated 12 V model had a maximum speed < 6000 RPM, after which field weakening could be applied to increase the operating range. This has not been included in this graph.
and accelerating by itself at around 80 to 120 RPM. As the BSG is connected through a belt with engine with a gear ratio of 2.8 to 3.4, the BSG requires 1540 to 1870 RPM for a smooth start, and 200 to 400 RPM to bring the ICE to a speed by which it ignites and accelerates by itself.

The BSG seems to be able to start an ICE while being powered with a 12 V battery. The machine can deliver 30 Nm torque while being supplied with a 12 V battery, and the ICE requires between 25 to 30 Nm torque to start turning. Furthermore, the BSG is able to deliver 30 Nm torque up to speeds higher than 300 RPM. At this point, the ICE should be able to accelerate by itself. In any case, the BSG will not provide the smooth starting capability when supplied with the 12 V battery as when under normal operation.

The starting capability of the machine can be improved if the control of the machine is adapted, as shown by the model. With adapted control, the machine is able to provide a higher output torque, which leads to a higher torque to accelerate the ICE. Therefore, the starting capability of the BSG can be improved with adapted control.

Once the BSG has started the ICE, the BSG can start generating energy. The BSG can generate up to 3,000 to 3,500 W, depending on the RPM of the ICE. This can be seen in figure C-6. Currently the steering and braking systems combined consume 3,000 to 3,500 W at peak demands, and therefore the BSG should be able to generate enough energy required for the safety critical systems in the car, assuming their power usage does not change.

However, in order to be certain about the starting capability of the machine, a test in praxis will have to be carried out. While at first sight it may appear that the BSG is able to start an ICE, the speed at which the ICE accelerates also plays an important role in whether the ICE can be started by the BSG. Therefore, the only way to be certain about the starting capability of the BSG is to assess it in practice.

### 6-3-3 Steering and braking

Based on the investigated BSG, it is not possible to determine with certainty whether the 48 V steering and braking systems supplied with 12 V can provide enough performance. It was not possible to investigate the difference in performance between operation at normal and reduced voltage. Specifically, it would have been interesting to assess whether the performance of these safety-critical systems at 12 V would have sufficed. This was not possible due to the fact that the steering and braking systems for 48 V were still in a pre-development phase, and therefore no prototypes were available.
The operation of prototype safety-critical system should be investigated at a later stage in the development of these systems for 48 V, when they are transferred to the 48 V side. Only then it is possible to accurately determine whether the performance of the system at 12 V meets the minimum requirements.

Now that the possibility of starting an ICE with a BSG and the backup function for steering or breaking at reduced voltage has been investigated, the connection and switching of the 12 V network to the BSG will be investigated. This will be discussed in the next chapter.
Figure 6-8: An example of the results from the starting test involving the starting of an ICE with a translation of 2.8 (a) and 3.4 (b).

In the upper part of the graph, the torque of the machine is visible. In the lower part, the RPM of the BSG is visible. The solid red (a) and blue (b) line and the time indicate the minimum required RPM for starting the ICE smoothly and after how many seconds this speed has been reached.

The red line indicates a flank, the green line indicates a 250 ms ramp, the blue line a 500 ms ramp, and the orange line a 2 s ramp. The 250 ms ramp was not tested in the case of starting with a gear ratio of 3.4.
Implementing and testing the switching topology

The objective of this chapter is to validate the proposed solution of connecting the 48 V and 12 V network through switching modules. First, the set-up that was built to test the topology is described in section 7-1. Then, the results are discussed in section 7-2.

7-1 The set-up used for testing

The set-up that was built to test the switching between 48 V and 12 V is schematically shown in figure 7-1. The set-up consists of a variable voltage source, a lead acid battery, a capacitor, an electronic load, the Q-Diodes, and a relay that can be controlled manually. This relay is used to simulate the disconnect of the power sources to the load in the car, or a lack of generated power in the circuit. It was chosen to use a relay to simulate the turning off, because otherwise the output capacitor of the voltage source would have influenced the switching. The components used for switching are summarized in table 7-1.

As there was no electrical machine available that could function on both 48 V and 12 V, an electronic load was used to simulate a machine. The electronic load had a negligible capacitance of 2 µF. The machine was simulated with the electronic load set on constant power mode. Furthermore, measurements were also done for the case that the machine would behave as a constant current load.

Due to the very high currents that were expected to flow through the circuit, the connecting cables were selected such that the voltage drop over the cables was minimized. Note that a resistance of 0.01 Ω leads to a voltage drop of 2 V at a current of 200 A, which is a large voltage drop compared to the 12 V battery voltage. With a 100 A current to the load, the voltage drop over the cables was approximately 1.5 V. Therefore, the resistance of the cables was around 15 mΩ at room temperature.
The tests were done for different values of capacitance parallel to the load, so that the influence of the value of the capacitors could be investigated. This would provide information on the design considerations of the capacitance parallel to the load, if the solution would indeed be implemented without a change to the switching application.

The values of capacitance for which the switching has been tested were 9.4 mF, 4.7 mF, 0.33 mF, and without a capacitance parallel to the load. One module of the electronic roll stabilization system had a relatively high capacitance compared to other systems in the 48 V iHEV topology, and it has a capacitance of 8.8 mF. Furthermore, a 0.33 mF capacitance is planned for the BSG, so both 0.33 and 8.8 mF are interesting, and a measurement point between 0.33 mF and 8.8 mF provides extra information. In addition, it is interesting to be able to compare these switching results with switching without a capacitance parallel to the load. Therefore, these values were chosen. As these capacitors were rated for 35 V, the voltage never exceeded 30 V during the tests. This is sufficient for testing, as in practice the undervoltage range starts below 24 V, and below 20 V the systems shut down, according to internal AUDI documentation [19].

The testing was done for different constant power loads of 500 W, 1 kW, and 1.5 kW. Higher constant power loads proved to be not suitable for testing. This was due to the maximum current rating of the electronic load, which was 300 A, and the behaviour of the lead acid battery when a high current flows. At a voltage of 12 V, a 2,0 kW constant power load requires more than 150 A of current. This current induces a voltage drop of more than 2 V due to the resistance of the wiring, which had a resistance of approximately 15 mΩ. In addition, the terminal voltage of the battery decreases by approximately 1 V. Now, the load sees a voltage of around 9 V and demands around 225 A of current. This leads to another decrease in voltage due to the ohmic voltage drop. Both this effect and the switching transients caused currents of 300 A to flow for a short time at 1,500 W. Therefore, a constant power load of 2,0 kW was not tested.

A small circuit was made to detect a voltage drop at the load, and to send a high signal to the Q-Diodes to switch when the voltage dropped too low. A detailed description of the circuit and its design is given in Appendix D. The circuit was designed such if the voltage at the load was less than approximately 2 V higher than the battery voltage, the Q-Diode would receive a high signal. A high signal for the Q-Diode is defined as a voltage higher than 9 V at the enable pin of the Q-Diode.

As there was no data available on the switching times of the Q-Diode, it was difficult to decide for the value of the switching threshold value. Ideally, the threshold value is chosen such that the switching transient is minimum. Switching too slow leads to a high inrush current from the battery to the load and its parallel capacitors. If the voltage difference between the load and the capacitors is high, and the 12 V is connected too soon, a high inrush current to the battery is generated. This phenomenon has to be prevented also.

For the second generation CAN-controlled Q-Diodes, the internal MOSFETs were known. These AUIRFS8409-7P MOSFETs from International Rectifier had a switch on and switch off speed of 180 ns respectively 256 ns. The switching speeds specified in a datasheet are for ideal situations, and therefore it was expected that even though the switching time would likely be higher, it would be definitely lower than 0.1 ms. Then, the switching topology has been built in SimPowerSystems in MATLAB and was then simulated. The constant power load of 1.5 kW was approximated with a controlled current source. Furthermore, a value of 8,8 mF was chosen for the capacitor, which is the same capacitance of one module of the
Table 7-1: Overview of different components used for the test set-up.

<table>
<thead>
<tr>
<th>Device</th>
<th>Type number</th>
</tr>
</thead>
<tbody>
<tr>
<td>Variable voltage source</td>
<td>Heiden Electronics DC Power Supply 1150 – 042, 42V/100A</td>
</tr>
<tr>
<td>Lead acid battery</td>
<td>VARTA 12V/75Ah, 420A peak</td>
</tr>
<tr>
<td>Electronic load</td>
<td>Heiden Electronics DS 3606, 60V/300A</td>
</tr>
<tr>
<td>Oscilloscope</td>
<td>LeCroy WaveJet 324A oscilloscope</td>
</tr>
<tr>
<td>Current probe</td>
<td>PR1001 from LEM, ±1% at 25°C, up to 10 kHz</td>
</tr>
<tr>
<td></td>
<td>CP 330 GOSSAN METRAWATT ±1% at 25°C, up to 20 kHz</td>
</tr>
</tbody>
</table>

Figure 7-1: A schematic overview of the setup used for testing.

electronic roll stabilisation system. In this case, a threshold voltage of 2 V led to a relatively small switching transient with no dead time at the load, and it was on the safe side regarding non-ideal capacitors and unexpected lower switching times. Therefore, a threshold value of 2 V was chosen.

For the investigation of the switching, the voltage at the load, the voltage of the enable signal of the Q-Diode, the current to the load, and the current from the battery were measured. These signals were measured with a LeCroy WaveJet 324A oscilloscope, and the oscilloscope was set to trigger on the event that the voltage at the load was < 19.5 V. Unfortunately, the current probes available could only measure with a frequency of up to 20 kHz. This is very low compared to the given MOSFET switching speed. There were however no measuring devices available that had a higher measuring frequency, so these were used for the measurements.

The difference in time between the enable signal and the current from the battery can be used to determine the switching time of the load. The results will be discussed in the next section.

7-2 Results

The switching of the Q-Diode takes approximately 3 ms, as can be observed in figure 7-2. During this measurement, the load had a capacitance of 4.7 mF parallel. For the first couple of milliseconds, the current is flowing through the body diode of the Q-Diode. After around 3 ms, the voltage of the load increases and there is no longer a voltage drop across the body.
Implementing and testing the switching topology

While the Q-Diode conducts quickly, the switching itself takes long when compared with the switching speeds of the MOSFETs given by the datasheet. The Q-Diode is designed such that the switching speed is independent on the current of the enable signal of the Q-Diode. Instead, it charges the internal capacitances through its terminals when the voltage of the enable signal $> 9$ V. Therefore, the only cause for the longer than expected switching time is the embedded circuit that drives the MOSFETs in the Q-Diode.

In order to prevent dead time of the load, it was decided to turn on the first Q-Diode by standard, so that current can flow through the body diodes of the MOSFETs in the second Q-Diode for a short time. The Q-Diode is able to tolerate high currents through the body diode for a short time, and with this adaptation of the circuit the tests have been carried out.

The results obtained from the oscilloscope have been analysed using MATLAB. The results are plotted for a constant power load with different capacitances, for a continuous current load with different capacitances, and different constant power loads with a constant capacitance of $9,4 \text{ mF}$. The results are shown in graphs, where they have been aligned on the moment the voltage of the load reaches a value lower than $19,5 \text{ V}$.

It can be concluded that in principle, the switching concept functions. The results in the figures 7-3, 7-4 and 7-5 show that it is possible to switch a $48 \text{ V}$ network load to the $12 \text{ V}$ network in a short time frame, while supplying the load with power during the switching. This is due to the fact that current is flowing through the body diode of the Q-Diode for around $3 \text{ ms}$ after which the Q-Diode switches, as is observed in figure 7-2. For each of the different load types, the switching will now be analysed.

Figure 7-2: Switching waveform of the Q-Diode of a $100 \text{ A}$ constant current load with a $4.7 \text{ mF}$ capacitance parallel to the load.
Constant current load

In the case of a constant current load there are less switching transients, as can be observed from figure 7-3. The current at the load stays stable during the switching period, although there is a transient the moment the power supply is manually disconnected from the load. The Q-diode enable signal is then high for a very short time also.

As a constant current is demanded the load voltage drops linearly, as can also be observed well for switching with a 9.4 mF capacitance parallel to the load. Furthermore, the voltage decreases slower and the enable signal becomes high later with increasing capacitance parallel to the load, as expected.

Different constant power loads with same capacitance parallel

The switching of different constant power loads, while a capacitance of 9.4 mF was connected in parallel with the load, was also investigated. This is shown in figure 7-4. The figure clearly shows that in the case of a higher constant power load, the voltage indeed drops more rapidly for the same capacitance in the network. Furthermore, the short high overshoot of the Q-diode enable signal can be observed more clearly. This is due the action of disconnecting the power supply manually from the load, as directly after the enable signal again drops to a low voltage.

In the case of switching the 1500 W load, the voltage drops below the steady state voltage for a short time. Here it can be observed that the battery then provides a higher current than absorbed by the load for a short time. This current flows to the capacitor to charge it to the battery voltage.

Constant power load with different capacitances

Figure 7-5 shows that increasing the capacitance parallel to the load leads to a smoother switching transient for a constant power load, as long as the voltage of the load does not drop below the voltage of the battery. A higher capacitance increases the energy available during the switching period, and therefore the voltage at the load decreases more slowly. Indeed, switching a 1.5 kW load with a 9.4 mF capacitance parallel leads to a smooth switch to the battery supply.

While it appears that the switching transients are caused by the fact that the load is switched with no or low capacitance, it is actually the regulation of the load that causes the overshooting and undershooting. In figure 7-3 no overshoot or undershoot can be observed, even though the voltage drops significantly right after the load has been disconnected from the voltage supply.

The electronic load regulates the current based on the voltage at the load in order to create a constant power load, according to the datasheet [59]. This can also be observed from the figure. In other words, the current to the electronic load always lags the voltage. Due to this inductive behaviour of the electronic load, while there is also a capacitance in parallel with the load, effectively an RLC circuit has been constructed. This induces the swing as if the circuit is underdamped. Therefore, it is important that if an electrical machine does indeed behave as such, either the switching or the conduction of the body diode happens rapidly. Otherwise, a very high current will be demanded.
Figure 7-3: Switching characteristics for a constant current load of 100 A at different capacitances.
Figure 7-4: Switching characteristics of different constant power loads at a constant capacitance of 9.4 mF.
Implementing and testing the switching topology

The switching time of the Q-Diode that was controlled via the CAN-bus was also investigated. It was not possible to test this on the measurement set-up, due to the fact that the Q-Diode required a controller to function, and it was not possible to obtain and program such a device. However, measurement results of the switching of the CAN-bus controlled Q-Diode inside a car have been obtained and these are shown in figure 7-6. According to the description of the requirements for this Q-Diode, the switching time itself is very low and should be < 100 µs. However, there is a delay between the moment the switching signal becomes high and the initiation of the switching. This CAN-bus controlled Q-Diode switches faster, but closing still takes approximately 2 ms. There is an approximate delay of 3,5 ms between the moment the enable signal asks for opening, and that the current decreased to 10% of its initial value.

In practice, it is possible to achieve a higher switching speed when a switch is specifically designed for the application. The Q-Diodes were designed to have a higher switch off than switch on speed, as their function is to disconnect a 12 V battery used for starting from the rest of the 12 V network. Due to the Q-Diode, the voltage drop due to the starting of the engine does not propagate to the rest of the network.

If the switching can be done quicker, the capacitance can also be reduced, which leads to a cost, size and weight reduction. A lower capacitance also decreases possible inrush currents after switch on. Furthermore, if the machine behaves as the electronic load, lower peak currents will occur as well if switching speed is reduced.

Furthermore, a protection against a short circuit at the load or in the 48 V network should be built. This is to prevent a short circuit in the 48 V network passing on to the 12 V network. The protection can be implemented by switching off the Q-Diode that is standard in conduction mode, at a specific voltage level of the 48 V network. This can be done with a similar circuit as that has been used for switching on the Q-Diode. However, in this case the enable signal of the Q-Diode has to be made low when the voltage at the 48 V side of the network drops below, for example, 9 V. This voltage level has to be chosen such as to provide sufficient safety.

**Added value of the proposed solution**

The added functional safety of the proposed solutions is difficult to determine. Whether the safety-critical systems are powered directly or indirectly with 12 V, their performance when supplied with 12 V has to be investigated in order determine with certainty whether it makes sense to have a redundant 12 V supply for the 48 V network, and whether the proposed solution adds functional safety. However, it was not possible to test this due to unavailability of these systems.

On the other hand, the solution has other interesting applications. First of all, it may be possible that later on, the starting of an ICE may become a safety-relevant requirement. This is to prevent the situation in where a driver has an unexpected loss of acceleration, which could lead to dangerous situations when the driver wants to overtake another driver.

Furthermore, the application could be interesting to implement in the current iHEV topology. If in this topology the 48 V network breaks down, the customer is obliged to drive his car to a dealer within a short time frame. This is due to the fact that energy is consumed, while no energy is generated, as the BSG in the 48 V has been shut off. If the 12 V lead acid battery is depleted, the steering and braking systems will also degrade in performance. The proposed solution of supplying the BSG with 12 V to start the ICE and generate energy
Figure 7-5: Switching characteristics for a constant power load of 1.5 kW at different capacitances.

Furthermore, it is likely more interesting to adapt the design of the DC/DC converter instead of adding an extra connection between the 48 V and 12 V network. On one side, the extra connection can cause extra faults and requires extra safety measures. Furthermore, a Q-Diode is relatively costly, and a new type has to be developed that can block the voltage between 12 V and 48 V. Therefore, an adaptation of the DC/DC converter will probably be much more interesting, both on a cost level and on a safety level.

To conclude, if the BSG is indeed able to start an ICE when supplied by a 12 V battery through a Q-Diode to the BSG, then the concept as a whole is expected to function. Then, the proposed solution does indeed provide a backup system for the fault that is likely to occur most frequently according to the analysis. However, the added functional safety of the proposed solution for the iHEV topology with a breaking and steering system on the 48 V network side is very dependent on the performance of the safety-critical systems at reduced voltage.
Figure 7-6: The switching characteristics of a CAN-bus controlled Q-Diode. Switch on is shown in figure (a) and switch off in figure (b).
Conclusions and recommendations

In this chapter, the conclusions of this thesis are discussed. First, a short summary including the results will be discussed. Then, the contributions of this thesis are discussed. Last, a number of recommendations for further work is given.

8-1 Conclusions

The research objective of this thesis is to make an overview of possible failures of safety-critical systems in the 48 V network and investigate possibilities to operate these safety-critical systems in case of a failure so that the systems can still carry out the minimum required level of performance. If a higher level of functional safety is achieved, the 48 V network is more robust and it is more likely that safety critical systems can be moved to the 48 V network. In order to achieve this, multiple steps have been taken.

First, the safety-critical systems have been identified. The steering and braking systems inside the car are safety-critical, while the active roll stabilization and electronic and the electronic turbo are safety relevant. It is required that the driver can always switch two lanes before putting the car safely at the side of the road.

All possible failures in the 48 V network have been investigated as accurately as possible using a Fault Tree Analysis (FTA). The most likely fault to occur is the failure of the battery. When there is no energy storage present in the 48 V network, the BSG has to be shut off, and thus there is a power failure of the 48 V network.

The most promising solution to this fault is a redundant connection of the safety critical systems at the 48 V network with the 12 V network. One concept is to provide these systems with power from the 12 V network directly, for a limited time. The other concept is to start the ICE by operating the BSG at 12 V, after which the BSG can generate power to provide the safety critical loads with power for a longer duration.

The switch consisted of back-to-back parallel MOSFETs. MOSFETs are required for the high switching speed, and the MOSFETs have to be connected in parallel due to the high currents. At Audi a switching module called Q-Diode was available, that consisted of
parallel MOSFETs that could cope with high currents and serve as a switch. Other solutions did not add extra reliability, were more expensive, or added only a little extra reliability at a significant increase of costs.

The theoretical and practical performance of a Belt-driven Starter Generator (BSG) at reduced voltage was then investigated. Only this machine could be analysed due to availability constraints. First, a model of the theoretical performance was created with the measured parameters of the machine. Then, the real performance of the machine at reduced voltage was measured. It proved that the performance of this machine at reduced voltage is worse than expected, due to the fact that the machine is constrained by its control.

The BSG already appears to be able to start a motor. Adapting the control of the machine would increase the starting capability of the BSG. To determine whether the machine is indeed able to start an ICE however, a test in practice has to be carried out.

The added value of the proposed for the iHEV topology with a steering and braking system on the 48 V network is highly dependent on the performance of these systems at 12 V. If the performance of these systems suffices at 12 V, then the proposed solution provides extra functional safety. Otherwise, the proposed solution makes little sense to implement.

However, the concept of starting the ICE with a 48 V BSG supplied with 12 V can be interesting for other applications. In the case of a safety-requrement on the starting of an ICE, for example, the proposed solution can provide extra functional safety. Furthermore, in the case of the iHEV topology, the proposed solution can provide the driver the possibility to keep driving when a fault occurs in the 48 V network. In addition, it is likely more interesting on both a cost and safety level to adapt the design of the DC/DC converter, than to add an extra connection between 48 V and 12 V.

### 8-2 Contributions of work

The following contributions have been made to answer the research questions of chapter 1:

- An overview of the possible faults safety-critical systems in the 48 V network has been determined through a Fault Tree Analysis (FTA), and possible solutions to these problems have been proposed.

- A model of the performance of a dual three-phase claw-pole hybrid excited synchronous machine with combined delta-wye windings, that includes the voltage and current operating limits of the machine. The model can also be used for regular three-phase hybrid excited synchronous machine.

- The performance of the 48 V BSG prototype in practice, during normal operation and at reduced voltage operation, has been determined, and it has been tried to assess the starting capability of the machine.

- The feasibility of increasing the functional safety of a vehicle by a 12 V redundant supply of the 48 V network by a connection through a switching module, has been investigated.
Based on the results obtained in this thesis, the following recommendations for further work are given:

- The model of the dual-three phase hybrid excited claw-pole machine can be improved by taking into account saturation and losses in the prediction of the performance. Furthermore, the model should also include the limit of the total available power when the machine operates at reduced voltage. This would lead to a much more accurate model of the performance of a claw-pole synchronous machine.

- The starting of an engine with the BSG should be tested in practice, in order to conclude with absolute certainty whether the BSG is able to start the ICE. The test has to be performed when the machine is powered by a battery supply with a Q-Diode in series, and with testing conditions that are as close to the real case as possible. In addition, the software of the BSG could be altered to assess the starting capability of the machine, when the control of the machine has been optimized.

- The added value of the concept has to be evaluated. It should be investigated whether the performance of a 48 V steering or braking system can deliver a sufficient performance at 12 V. Furthermore, the failure rate of the proposed concept has to be investigated. In any case, the voltage blocking capability of the switching application has to be improved and a short circuit protection has to be implemented in the case of implementation. In addition, the value of the capacitance in parallel to the load has to be chosen such that the switching transients are minimized.
The objective of this appendix is to construct a model for the dual three-phase machine that was investigated. The modelling of a dual three-phase machine is similar to the modelling of a classical three-phase machine. There are two approaches that can be taken [28, 54–56], where one approach leads to a model without magnetic coupling between the dual three-phase windings. The other approach results in a model with magnetic coupling, which is more difficult to analyse.

The approach followed considers the machine as a polyphase machine with equally distributed phases. With vector space decomposition, a model can be constructed with orthogonal subspaces, and thus without magnetic coupling between the dual three-phase winding sets. Prior to the construction of the model, the following is assumed:

- The two three-phase winding sets in the stator are identical and shifted by an angle $\gamma = 30^\circ$
- The magnetomotive forces (MMF) are sinusoidally distributed
- Only the first space harmonics are considered, and the mutual inductances are characterized by their fundamental component alone
- Magnetic saturation is not regarded

The machine has two three-phase winding sets, that are $30^\circ$ shifted electrically. Furthermore, the rotor has a field winding and no damper windings. As the flux of the permanent magnet is in the same direction as the field winding, the flux of the permanent magnet and field winding combined are expressed as an equivalent rotor field winding with a mutual stator-field coupling
The rotor position independent part is given by the matrix
\[
\begin{bmatrix}
R_a & 0 & 0 & 0 & 0 & 0 \\
0 & R_a & 0 & 0 & 0 & 0 \\
0 & 0 & R_a & 0 & 0 & 0 \\
0 & 0 & 0 & R_a & 0 & 0 \\
0 & 0 & 0 & 0 & R_a & 0 \\
0 & 0 & 0 & 0 & 0 & R_f \\
\end{bmatrix}
\]
where the matrix \([L]\) is given by
\[
[L] = \begin{bmatrix}
[L_{s1}] & [M_{s1s2}] & [M_{s1R}] \\
[M_{s1s2}]^T & [L_s] & [M_{s2R}] \\
[M_{s1R}]^T & [M_{s2R}]^T & L_f \\
\end{bmatrix}
\]

Here, the matrices \([L_{s1}]\) and \([L_{s2}]\) represent the inductances of the first and second phase set. The matrices \([M_{s1s2}]\) represent the magnetic coupling between the two phase winding sets, while \([M_{s1R}]\) describes the coupling between the rotor and the stator. The matrices \([L_{s1}]\) and \([L_{s2}]\) can be broken down into the sum of a constant and rotor position dependent part:
\[
[L_{sx}] = [L_{sx0}] + [L_{sx}(\beta_x)]
\]
The rotor position independent part is given by the matrix \([L_{sx0}]\),
\[
[L_{sx0}] = \begin{bmatrix}
L_s & M_{ss} \cos(\frac{2\pi}{3}) & M_{ss} \cos(\frac{\pi}{3}) \\
M_{ss} \cos(\frac{2\pi}{3}) & L_s & M_{ss} \cos(\frac{2\pi}{3}) \\
M_{ss} \cos(\frac{\pi}{3}) & M_{ss} \cos(\frac{2\pi}{3}) & L_s \\
\end{bmatrix}
\]

where the self inductance \(L_s\) consists of a leakage inductance \(L_a\) and stator mutual inductance \(M_{ss}\), thus \(L_s = L_a + M_{ss}\), while the rotor position dependent part is given by
\[
[L_{sx}(\beta_x)] = M_{sfm} \begin{bmatrix}
\cos(2\beta_x) & \cos(2\beta_x - \frac{\pi}{3}) & \cos(2\beta_x + \frac{\pi}{3}) \\
\cos(2\beta_x - \frac{\pi}{3}) & \cos(2\beta_x + \frac{\pi}{3}) & \cos(2\beta_x) \\
\cos(2\beta_x + \frac{\pi}{3}) & \cos(2\beta_x) & \cos(2\beta_x - \frac{\pi}{3}) \\
\end{bmatrix}
\]
The angle \(\beta_x\) is given by \(\beta_x = \theta - (x - 1)\frac{\pi}{6}\) and \(x = 1 \text{ or } x = 2\). Due to the shift of the second three-phase winding set, the coupling between these two different three-phase winding sets is shifted as well. The magnetic coupling between the two phase windings is given by
\[
[M_{s1s2}] = [M_{ss0}] + [M_{ss}(\theta)]
\]
The matrix \([M_{ss0}]\) is only dependent on the shift in electrical angle between the two phases, and is given by
\[
[M_{ss0}] = M_{ss} \begin{bmatrix}
\cos(\frac{\pi}{6}) & \cos(\frac{5\pi}{6}) & \cos(\frac{2\pi}{3}) \\
\cos(\frac{5\pi}{6}) & \cos(\frac{2\pi}{3}) & \cos(\frac{\pi}{6}) \\
\cos(\frac{2\pi}{3}) & \cos(\frac{\pi}{6}) & \cos(\frac{\pi}{6}) \\
\end{bmatrix}
\]
The matrix \([M_{ss}(\theta)]\) is given by
\[
[M_{ss}(\theta)] = M_{sfm} \begin{bmatrix}
\cos(2\theta - \frac{\pi}{6}) & \cos(2\theta - \frac{5\pi}{6}) & \cos(2\theta + \frac{\pi}{6}) \\
\cos(2\theta - \frac{5\pi}{6}) & \cos(2\theta + \frac{\pi}{6}) & \cos(2\theta - \frac{5\pi}{6}) \\
\cos(2\theta + \frac{\pi}{6}) & \cos(2\theta - \frac{5\pi}{6}) & \cos(2\theta - \frac{5\pi}{6}) \\
\end{bmatrix}
\]
The matrix describing the coupling between the stator and the rotor, \([M_{szR}]\), is given by

\[
[M_{szR}] = [M_{zR}(\beta_x)] = M'_sF \begin{bmatrix}
\cos(\beta_x) \\
\cos(\beta_x - \frac{\pi}{2}) \\
\cos(\beta_x + \frac{\pi}{2})
\end{bmatrix}
\]

Also for this matrix, \(\beta_x = \theta - (x - 1)\frac{\pi}{6}\) and \(x = 1\) or \(x = 2\).

The voltage equations previously defined remain the same, but the equations are rearranged so that the equations can be decoupled. Now, the voltage equations are written as

\[
\begin{bmatrix}
v_{a1} \\
v_{a2} \\
v_{b1} \\
v_{b2} \\
v_{c1} \\
v_{c2} \\
v_f
\end{bmatrix} =
\begin{bmatrix}
R_a & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & R_a & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & R_a & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & R_a & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & R_a & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & R_a & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & R_f
\end{bmatrix}
\begin{bmatrix}
i_{a1} \\
i_{a2} \\
i_{b1} \\
i_{b2} \\
i_{c1} \\
i_{c2} \\
i_f
\end{bmatrix} + [L] \cdot \frac{d}{dt} \begin{bmatrix}
i_{a1} \\
i_{a2} \\
i_{b1} \\
i_{b2} \\
i_{c1} \\
i_{c2} \\
i_f
\end{bmatrix}
\]

where in this case the matrix \([L]\) is written in the form

\[
[L] = \begin{bmatrix}
[L_{ss}] & [M_{zR}] \\
[M_{zR}]^T & L_f
\end{bmatrix}
\]

The matrix \([L_{ss}]\) comprises of a leakage flux \(L_a\), the magnetic coupling between the two winding sets \(M_{ss}\), which is dependent on the phase shift between the phases, and \(M_{sfm}\), which depends on the electric position:

\[
[L_{ss}] = L_a[6 \times 6] + M_{ss} \cdot [M_0] + M_{sfm} \cdot [M_2]
\]

The matrix \([M_0]\) is then given by

\[
[M_0] =
\begin{bmatrix}
\cos(0) & \cos(\frac{\pi}{6}) & \cos(\frac{4\pi}{6}) & \cos(\frac{5\pi}{6}) & \cos(\frac{8\pi}{6}) & \cos(\frac{9\pi}{6}) \\
\cos(\frac{\pi}{6}) & \cos(0) & \cos(\frac{4\pi}{6}) & \cos(\frac{5\pi}{6}) & \cos(\frac{8\pi}{6}) & \cos(\frac{9\pi}{6}) \\
\cos(\frac{4\pi}{6}) & \cos(\frac{5\pi}{6}) & \cos(0) & \cos(\frac{9\pi}{6}) & \cos(\frac{1\pi}{6}) & \cos(\frac{8\pi}{6}) \\
\cos(\frac{5\pi}{6}) & \cos(\frac{4\pi}{6}) & \cos(\frac{8\pi}{6}) & \cos(\frac{9\pi}{6}) & \cos(\frac{1\pi}{6}) & \cos(\frac{8\pi}{6}) \\
\cos(\frac{8\pi}{6}) & \cos(\frac{5\pi}{6}) & \cos(\frac{1\pi}{6}) & \cos(\frac{8\pi}{6}) & \cos(\frac{9\pi}{6}) & \cos(\frac{2\pi}{6}) \\
\cos(\frac{9\pi}{6}) & \cos(\frac{1\pi}{6}) & \cos(\frac{8\pi}{6}) & \cos(\frac{9\pi}{6}) & \cos(\frac{2\pi}{6}) & \cos(\frac{8\pi}{6}) \\
\end{bmatrix}
\]

Furthermore, the matrix \([M_2]\) is then given by

\[
[M_2] =
\begin{bmatrix}
\cos(2\theta) & \cos(2\theta - \frac{\pi}{6}) & \cos(2\theta - \frac{4\pi}{6}) & \cos(2\theta - \frac{5\pi}{6}) & \cos(2\theta - \frac{8\pi}{6}) & \cos(2\theta + \frac{\pi}{6}) & \cos(2\theta + \frac{4\pi}{6}) & \cos(2\theta + \frac{5\pi}{6}) \\
\cos(2\theta - \frac{\pi}{6}) & \cos(2\theta - \frac{4\pi}{6}) & \cos(2\theta - \frac{5\pi}{6}) & \cos(2\theta - \frac{8\pi}{6}) & \cos(2\theta + \frac{\pi}{6}) & \cos(2\theta + \frac{4\pi}{6}) & \cos(2\theta + \frac{5\pi}{6}) & \cos(2\theta + \frac{8\pi}{6}) \\
\cos(2\theta - \frac{4\pi}{6}) & \cos(2\theta - \frac{5\pi}{6}) & \cos(2\theta - \frac{8\pi}{6}) & \cos(2\theta + \frac{\pi}{6}) & \cos(2\theta + \frac{4\pi}{6}) & \cos(2\theta + \frac{5\pi}{6}) & \cos(2\theta + \frac{8\pi}{6}) & \cos(2\theta + \frac{11\pi}{6}) \\
\cos(2\theta - \frac{5\pi}{6}) & \cos(2\theta - \frac{8\pi}{6}) & \cos(2\theta + \frac{\pi}{6}) & \cos(2\theta + \frac{4\pi}{6}) & \cos(2\theta + \frac{5\pi}{6}) & \cos(2\theta + \frac{8\pi}{6}) & \cos(2\theta + \frac{11\pi}{6}) & \cos(2\theta + \frac{14\pi}{6}) \\
\cos(2\theta - \frac{8\pi}{6}) & \cos(2\theta + \frac{\pi}{6}) & \cos(2\theta + \frac{4\pi}{6}) & \cos(2\theta + \frac{5\pi}{6}) & \cos(2\theta + \frac{8\pi}{6}) & \cos(2\theta + \frac{11\pi}{6}) & \cos(2\theta + \frac{14\pi}{6}) & \cos(2\theta + \frac{17\pi}{6}) \\
\cos(2\theta + \frac{\pi}{6}) & \cos(2\theta + \frac{4\pi}{6}) & \cos(2\theta + \frac{5\pi}{6}) & \cos(2\theta + \frac{8\pi}{6}) & \cos(2\theta + \frac{11\pi}{6}) & \cos(2\theta + \frac{14\pi}{6}) & \cos(2\theta + \frac{17\pi}{6}) & \cos(2\theta + \frac{20\pi}{6}) \\
\cos(2\theta + \frac{4\pi}{6}) & \cos(2\theta + \frac{5\pi}{6}) & \cos(2\theta + \frac{8\pi}{6}) & \cos(2\theta + \frac{11\pi}{6}) & \cos(2\theta + \frac{14\pi}{6}) & \cos(2\theta + \frac{17\pi}{6}) & \cos(2\theta + \frac{20\pi}{6}) & \cos(2\theta + \frac{23\pi}{6}) \\
\cos(2\theta + \frac{5\pi}{6}) & \cos(2\theta + \frac{8\pi}{6}) & \cos(2\theta + \frac{11\pi}{6}) & \cos(2\theta + \frac{14\pi}{6}) & \cos(2\theta + \frac{17\pi}{6}) & \cos(2\theta + \frac{20\pi}{6}) & \cos(2\theta + \frac{23\pi}{6}) & \cos(2\theta + \frac{26\pi}{6}) \\
\cos(2\theta + \frac{8\pi}{6}) & \cos(2\theta + \frac{11\pi}{6}) & \cos(2\theta + \frac{14\pi}{6}) & \cos(2\theta + \frac{17\pi}{6}) & \cos(2\theta + \frac{20\pi}{6}) & \cos(2\theta + \frac{23\pi}{6}) & \cos(2\theta + \frac{26\pi}{6}) & \cos(2\theta + \frac{29\pi}{6}) \\
\end{bmatrix}
\]

Last, the matrix \([M_{zR}]\) is given by

\[
[M_{zR}] = M'_sF \cdot
\begin{bmatrix}
\cos \theta \\
\cos(\theta - \frac{\pi}{6}) \\
\cos(\theta - \frac{2\pi}{6}) \\
\cos(\theta - \frac{3\pi}{6}) \\
\cos(\theta - \frac{4\pi}{6}) \\
\cos(\theta - \frac{5\pi}{6}) \\
\end{bmatrix}
\]
The voltage equations defined in A-10 have to be transformed into an orthonormal basis, so that there is no magnetic coupling. A transformation matrix has been obtained in [28, 56], and is repeated here:

$$[T] = \sqrt{\frac{3}{6}} \begin{bmatrix}
1 & \cos\left(\frac{\pi}{6}\right) & \cos\left(\frac{\pi}{3}\right) & \cos\left(\frac{2\pi}{3}\right) & \cos\left(\frac{5\pi}{6}\right) \\
0 & \sin\left(\frac{\pi}{6}\right) & \sin\left(\frac{\pi}{3}\right) & \sin\left(\frac{2\pi}{3}\right) & \sin\left(\frac{5\pi}{6}\right) \\
1 & \cos\left(\frac{\pi}{6}\right) & \cos\left(\frac{\pi}{3}\right) & \cos\left(\frac{2\pi}{3}\right) & \cos\left(\frac{5\pi}{6}\right) \\
0 & \sin\left(\frac{\pi}{6}\right) & \sin\left(\frac{\pi}{3}\right) & \sin\left(\frac{2\pi}{3}\right) & \sin\left(\frac{5\pi}{6}\right) \\
0 & 0 & 1 & 0 & 1
\end{bmatrix}$$  \hspace{1cm} (A-16)

This transformation matrix is applied to the voltage equations defined in A-10 as follows:

$$[T] \cdot [V] = [T] \cdot [R_s] \cdot [T^{-1}] \cdot [T] \cdot [I] + \frac{d}{dt} [T] \cdot [L] \cdot [T^{-1}] \cdot [T] \cdot [I]$$  \hspace{1cm} (A-17)

This transforms the voltage and currents vectors to the new \( \alpha\beta\zeta_2\zeta_3\zeta_4 \) reference frame, and its voltage equations are given by:

$$\begin{bmatrix}
v_{\alpha} \\
v_{\beta} \\
v_{\zeta_1} \\
v_{\zeta_2} \\
v_{\zeta_3} \\
v_{\zeta_4}
\end{bmatrix} = R_a \begin{bmatrix}
i_{\alpha} \\
i_{\beta} \\
i_{\zeta_1} \\
i_{\zeta_2} \\
i_{\zeta_3} \\
i_{\zeta_4}
\end{bmatrix} + [L] \frac{d}{dt} \begin{bmatrix}
i_{\alpha} \\
i_{\beta} \\
i_{\zeta_1} \\
i_{\zeta_2} \\
i_{\zeta_3} \\
i_{\zeta_4}
\end{bmatrix} + [M_{sfm}] \frac{d}{dt} \begin{bmatrix}
i_{\alpha} \\
i_{\beta} \\
i_{\zeta_1} \\
i_{\zeta_2} \\
i_{\zeta_3} \\
i_{\zeta_4}
\end{bmatrix} + \sqrt{3} M'_{sF} \begin{bmatrix}
cos(\theta) \\
\sin(\theta)
\end{bmatrix}$$  \hspace{1cm} (A-18)

where \([L]\) is given by

$$[L] = \begin{bmatrix}
L_\sigma + 3M_{ss} & 0 & 0 & 0 & 0 & 0 \\
0 & L_\sigma + 3M_{ss} & 0 & 0 & 0 & 0 \\
0 & 0 & L_\sigma & 0 & 0 & 0 \\
0 & 0 & 0 & L_\sigma & 0 & 0 \\
0 & 0 & 0 & 0 & L_\sigma & 0 \\
0 & 0 & 0 & 0 & 0 & L_\sigma
\end{bmatrix}$$  \hspace{1cm} (A-19)

and \([M_{sfm}]\) is given by

$$[M_{sfm}] = M_{sfm} \begin{bmatrix}
3\cos(2\theta) & 3\sin(2\theta) & 0 & 0 & 0 & 0 \\
3\sin(2\theta) & -3\cos(2\theta) & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0
\end{bmatrix}$$  \hspace{1cm} (A-20)

The voltage equations of the dual three-phase windings in the reference frame \( \alpha\beta\zeta_1\zeta_2\zeta_3\zeta_4 \) are the same as a three-phase machine in reference frame \( \alpha\beta\zeta \). By applying Park’s rotation matrix the following equations are obtained:

$$\begin{bmatrix}
v_d \\
v_q
\end{bmatrix} = R_a \begin{bmatrix}
i_d \\
i_q
\end{bmatrix} + \left[\begin{array}{cc}
L_d & 0 \\
0 & L_q
\end{array}\right] \omega \begin{bmatrix}
i_d \\
i_q
\end{bmatrix} + \omega \begin{bmatrix}
0 & -L_d \\
L_d & 0
\end{bmatrix} \begin{bmatrix}
i_d \\
i_q
\end{bmatrix} + \sqrt{3} \omega' \begin{bmatrix}
0 & M'_{sF} \\
M'_{sF} & 0
\end{bmatrix} + \sqrt{3} \begin{bmatrix}
M'_{sF} \\
0
\end{bmatrix} \frac{d}{dt} \omega'$$  \hspace{1cm} (A-21)
And the equations for the subspace $z_1, z_2, z_3, z_4$ are

$$\begin{bmatrix} v_{z1} \\ v_{z2} \\ v_{z3} \\ v_{z4} \end{bmatrix} = R_a \begin{bmatrix} i_{z1} \\ i_{z2} \\ i_{z3} \\ i_{z4} \end{bmatrix} + \begin{bmatrix} L_\sigma & 0 & 0 & 0 \\ 0 & L_\sigma & 0 & 0 \\ 0 & 0 & L_\sigma & 0 \\ 0 & 0 & 0 & L_\sigma \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{z1} \\ i_{z2} \\ i_{z3} \\ i_{z4} \end{bmatrix}$$

(A-22)

If the rotor field winding flux linkage is again expressed as the flux due to the field winding and the permanent magnets, the equations can be rearranged to

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = R_a \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \lambda_d \\ \lambda_q \end{bmatrix} + \begin{bmatrix} 0 & -\omega_e \\ \omega_e & 0 \end{bmatrix} \begin{bmatrix} \lambda_d \\ \lambda_q \end{bmatrix}$$

(A-23)

$$\begin{bmatrix} v_{z1} \\ v_{z2} \\ v_{z3} \\ v_{z4} \end{bmatrix} = R_a \begin{bmatrix} i_{z1} \\ i_{z2} \\ i_{z3} \\ i_{z4} \end{bmatrix} + L_\sigma \frac{d}{dt} \begin{bmatrix} i_{z1} \\ i_{z2} \\ i_{z3} \\ i_{z4} \end{bmatrix}$$

(A-24)

$$v_f = R_f i_f + \frac{d}{dt} \lambda_f$$

(A-25)

where the equations for the flux linkages are

$$\begin{bmatrix} \lambda_d \\ \lambda_q \\ \lambda_f \end{bmatrix} = \begin{bmatrix} L_d & 0 & \sqrt{3} M_{sF} \\ 0 & L_q & 0 \\ \sqrt{3} M_{sF} & 0 & L_f \end{bmatrix} \begin{bmatrix} i_d \\ i_q \\ i_f \end{bmatrix} + \begin{bmatrix} \sqrt{3} \lambda_{pm} \\ 0 \\ 0 \end{bmatrix}$$

(A-26)

and indeed, these equations align with the normal expressions for a hybrid excited synchronous machine [60]. Here, $L_\sigma$ is the leakage inductance, and $L_d$ and $L_q$ are given by

$$L_d = L_\sigma + 3 M_{ss} + 3 M_{sfm}$$

$$L_q = L_\sigma + 3 M_{ss} - 3 M_{sfm}$$

(A-27)

The torque equation is similar to the normal equation for the torque in the $d,q$ reference frame, and is given by:

$$T = p \left( \sqrt{3} i_q (M_{sf} i_f + \lambda_{pm}) + (L_d - L_q) i_d i_q \right)$$

(A-28)
Appendix B

Analysis of machine losses

The objective of this chapter is to provide some background information on the electrical losses of the machine for the interested reader.

The losses of the machine can be categorized in electrical and mechanical losses. The electrical losses consist of the losses in the power electronics, the stator copper losses, the iron losses and the rotor field winding losses. The iron losses consist of eddy current losses and hysteresis losses, and the mechanical losses consist of the windage and friction losses.

The electrical losses of the machine will now be analysed, starting with the iron losses of the machine. The iron losses consist of the eddy current losses and the hysteresis losses. The hysteresis losses are the losses due to the hysteresis effect of the core. From [58], the hysteresis losses can be approximated by:

\[ P_h = K_h B_{\text{max}}^n f \]  

where \( B_{\text{max}} \) is the maximum flux density, \( K_h \) is an empirically determined constant for a ferromagnetic material and volume of the core and \( f \) is the frequency of the phase current. The value for \( n \) is also empirically determined, and varies between 1.5 and 2.5. The eddy current losses are given by [58]:

\[ P_e = K_e B_{\text{max}}^2 f^2 \]  

where \( K_e \) is a constant that is dependent on the type of material and the lamination thickness. The remaining electrical losses of the machine are the losses in the power electronics, the losses in the stator, and the losses in the field of the machine.

The copper losses of the stator are given by:

\[ P_{\text{copper}} = 3 I_{\text{ph}}^2 R_{\text{ph}} \]  

where \( I_{\text{ph}} \) and \( R_{\text{ph}} \) is the phase current respectively phase resistance. The losses of the field winding are given by:

\[ P_{\text{field}} = I_f^2 R_f \]  

where \( I_f \) and \( R_f \) is the field winding current respectively field winding resistance. Last, there are also losses in the power electronics attached to the machine. The losses in the
power electronics consist of conduction losses and switching losses. The gate drive losses are neglected for simplicity, as it is not expected that these losses are a high proportion of the losses. The instantaneous on-state conduction losses of the MOSFETs are given by [57]:

\[ P_{\text{conduction}} = i^2 R_{DS(on)} \]  \hspace{1cm} (B-5)

where \( R_{DS(on)} \) is the drain–source on-state resistance of the MOSFET. Furthermore, the switching losses can be approximated by:

\[ P_{\text{switching}} = V_{DS,off} I_{DS,on} \cdot (t_{on} + t_{off}) \]  \hspace{1cm} (B-6)

where \( t_{on} \) and \( t_{off} \) are the rise and fall time of the switch.

Now that the relevant formula’s have been mentioned, the losses of the machine are calculated. As seen in chapter 6, the losses of the machine are especially high at low speed. Therefore, this region is investigated. As the machine is operating at very low speed, the frequency of the phase current is relatively low, and the hysteresis and eddy current losses as given by equations B-1 and B-2 can be conveniently neglected.

From [61], the values required to calculate the losses from the power electronics are taken. The value for \( R_{DS(on)} \) is taken for an operating temperature of \( T = 150^\circ \text{C} \). In the machine two MOSFETs are placed in parallel, and each phase has two pairs of MOSFETs in series. Therefore, the net resistance is equal to the resistance of one MOSFET. The values for the switching times are also taken from this data sheet.

The field winding losses can be calculated from the open circuit measurement from chapter 5. The machine was then operating while the coolant had a temperature of 90\(^\circ\)C. Furthermore, the voltage over and current through the field winding were measured during this test, and these values are used for the calculation of the losses in the field winding.

The resistance of the phases has been determined at room temperature, and the temperature of the wire is much higher under normal operation. Therefore, the resistance of the phase resistance is approximated by:

\[ R_{\text{high temp}} = R_{\text{room}} \cdot [1 + \alpha \cdot (T - T_{\text{room}})] \]  \hspace{1cm} (B-7)

where \( \alpha = 4 \cdot 10^{-3} \) is the temperature coefficient of copper. It was assumed that the room temperature \( T_{\text{room}} = 23^\circ \text{C} \). It is not known at which temperature the stator windings are operating, but the manufacturer claimed it was 190\(^\circ\)C during steady state operation. The dc link voltage and current is known, and the inverter of the machine operates in normal PWM. Therefore, the phase currents can be calculated with equation 5-29 and 5-32. It is assumed that the current divides itself perfectly between the two three-phase winding sets.

The machine has a combined delta-star winding. It is assumed that each resistance in the machine is of equal value. The value of each resistance can then be calculated from the measured value for the resistance. The losses in the star are then given by:

\[ P_{\text{Star}} = 3I_{ph}^2 \cdot R_{ph} \]  \hspace{1cm} (B-8)

The losses in the delta windings are then given by:

\[ P_{\Delta} = 3 \left( \frac{I_{ph}}{\sqrt{3}} \right)^2 \cdot R_{ph} \]  \hspace{1cm} (B-9)
Table B-1: Calculated losses for the machine for a stator current of $I = 70$ A.

<table>
<thead>
<tr>
<th>Stator winding temperature [$°\text{C}$]</th>
<th>120° C</th>
<th>190° C</th>
</tr>
</thead>
<tbody>
<tr>
<td>Losses field [W]</td>
<td>$\approx 100$ W</td>
<td>$\approx 100$ W</td>
</tr>
<tr>
<td>Losses power electronics [W]</td>
<td>$\approx 40$ W</td>
<td>$\approx 65$ W</td>
</tr>
<tr>
<td>Copper losses [W]</td>
<td>$\approx 1800$ W</td>
<td>$\approx 2200$ W</td>
</tr>
<tr>
<td>Sum [W]</td>
<td>$\approx 1950$ W</td>
<td>$\approx 2350$ W</td>
</tr>
</tbody>
</table>

Table B-2: Calculated losses for the machine for a stator current of $I = 90$ A.

<table>
<thead>
<tr>
<th>Stator winding temperature [$°\text{C}$]</th>
<th>120° C</th>
<th>190° C</th>
</tr>
</thead>
<tbody>
<tr>
<td>Losses field [W]</td>
<td>$\approx 100$ W</td>
<td>$\approx 100$ W</td>
</tr>
<tr>
<td>Losses power electronics [W]</td>
<td>$\approx 65$ W</td>
<td>$\approx 65$ W</td>
</tr>
<tr>
<td>Copper losses [W]</td>
<td>$\approx 3000$ W</td>
<td>$\approx 3650$ W</td>
</tr>
<tr>
<td>Sum [W]</td>
<td>$\approx 3100$ W</td>
<td>$\approx 3800$ W</td>
</tr>
</tbody>
</table>

If the required phase current is calculated based on the torque output, there is no saturation, and we assume that the temperature of the stator winding is 120° C and 190° C, then we calculate the results as shown in table B-1. In this case, the stator current is 70 A.

For an assumed stator current of 90 A, the losses are given in table B-2. There was a relative long time between measurements, and therefore a 120° C stator temperature could be realistic for some data points. The calculated losses for $I = 90$ A are close to the measured values for the losses, as in figure 6-6.
Field voltage at different speeds, battery powered

**Figure C.1**: The voltage of the field winding of the machine, when powered by a battery. The field voltage is always regulated to its maximum value. The field winding voltage is low when the machine is demanding high currents. Not only does this decrease the terminal voltage of the battery, but it also creates a voltage drop over the wiring feeding the machine.
Field voltage at different speeds, with 40 V source

Figure C-2: The voltage of the field winding of the machine, when powered by a 40 V voltage supply. Also in this, the field winding voltage is lower when the machine is demanding high currents. Not only does this decrease the voltage of the supercap, but the high currents also create a voltage drop over the wiring feeding the machine. This voltage decreases rapidly, and therefore the reader is advised not to attach much importance to specific values. The measurements were carried out by hand, and therefore the speed at which the measurement data was captured influenced the value of the field winding voltage.

MOSFET gate voltage during normal PWM

Figure C-3: A capture of the waveform of the gate voltage of a MOSFET pair for sinusoidal pulse width modulation. The capture was made at 2,490 RPM.
Figure C-4: A capture of the waveform of the gate voltage of a MOSFET pair for square wave pulse width modulation. The capture was made at 2,500 RPM.
Figure C-5: Overview of the torque-speed characteristic for different voltages and temperatures.
Generated power at different speeds, with 13.5 V clamped load

Figure C-6: The electrical energy that the 48 V BSG can generate at 12 V. The machine does not generate at speeds below 1,500 RPM.
Appendix D

Q-Diode driver circuit

The objective of this chapter is to provide information about the design choices behind the driver for the Q-Diodes. Please note that the idea behind the driver circuit was to test a concept, and that a minimum viable circuit was sufficient.

The Q-Diode switches when the enable signal becomes larger than 9 V. In other words, when the voltage is lower than the threshold voltage, the Q-Diodes have to receive a high signal. A differential amplifier with a low-side transistor is an easy solution to implement this functionality, as illustrated in figure D-1. If the difference between both input voltages becomes too low, the NPN transistor is turned off, and the Q-Diodes receive a high enable signal. The NPN transistor is conducting when the difference between the two voltage is high enough, and so pulls the enable signal for the Q-Diode to ground.

The next step in the design process was to determine the available transistors and operational amplifiers. It was assumed that the Q-Diode had a switching speed of around 4 ms.

The transistor was chosen as follows. First, the Q-Diodes have an approximate impedance of 1.8 kΩ. Therefore, the voltage drop across Rl should not be too much, when the transistor is turned off. On the other hand, a higher resistance leads to a lower power consumption when the transistor is turned on. A value of 200 Ω was chosen, so that the voltage drop across the resistor is minimal.

As Rl is 200 Ω, the transistor has to be able to carry a maximum current of 12 V / 200 Ω, which is 6 mA. In practice, it could be slightly higher if the battery is fully charged. Therefore, 10 mA is taken for safety.

The 2N3904 NPN transistor was chosen of all available transistors. The 2N3904 transistor was one of the few transistors that specified its turn off speed of 250 ns in its datasheet. Furthermore, it has a maximum $V_{CE}$ of 40 V and a continuous $I_C$ of 200 mA. These ratings provide enough room for safety, and all other transistors had even higher ratings, and therefore would likely be more expensive. While the difference is not large for a prototype, there could be a larger difference in the case of mass-production.

The worst case forward current gain $h_{fe}$ of the transistor is 30. With a resistance value of 2.2 kΩ for Rb, enough current can be supplied even if the transistor is about to be turned off at 0.65 V.
<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>R1</td>
<td>51 kΩ</td>
</tr>
<tr>
<td>R2</td>
<td>20 kΩ</td>
</tr>
<tr>
<td>R3</td>
<td>15 kΩ</td>
</tr>
<tr>
<td>R4</td>
<td>13.15 kΩ</td>
</tr>
<tr>
<td>Rb</td>
<td>2.2 kΩ</td>
</tr>
<tr>
<td>Ri</td>
<td>200 Ω</td>
</tr>
<tr>
<td>Transistor</td>
<td>2N3904</td>
</tr>
<tr>
<td>OpAmp</td>
<td>LM 324</td>
</tr>
</tbody>
</table>

Now, also an operational amplifier has to be selected. Due to the voltage followers, a quad amplifier was chosen. Of the available ICs, all except the LM324 were designed for a dual voltage supply. Furthermore, these amplifiers were unable to output rail to rail. As they would have been supplied with 12 V on their plus side, and connected to ground on their minus side, they would be unable to turn off the transistor. The LM324 however, is able to provide a 0 V to $V^+ - 1.5$ V output voltage, and is therefore able to turn off the transistor.

The other characteristics of the LM324 were also sufficient enough for the application. With a slew rate of 0.5 V per μs, dropping from 10.5 V to the base-emitter saturation voltage of 0.65 V would take 19.7 μs, which negligible compared to the assumed switching time of the Q-Diode.

The differential input voltage of LM324 should not exceed the input voltage. Therefore, the voltages have to be divided. The maximum allowed input voltage can be calculated as follows:

$$V_{allowed} = \frac{12}{V_{max}} \cdot V_{in}$$  \hspace{1cm} (D-1)

The variable voltage source used in the test set-up, had a maximum voltage of 42 V. Therefore, $V_{max}$ is 42 V. This comes down to divider with a relation of $\frac{12}{42}$, which is $\frac{2}{7}$, in this case. Therefore, resistances with the values of 20 kΩ and 51 kΩ were selected.

For the gain of the differential operational amplifier, we can write:

$$V_o = \frac{R4}{R3} (V'_{48V} - V'_{12V})$$  \hspace{1cm} (D-2)

where $V'_{48V}$ and $V'_{12V}$ are the voltages at the output of the voltage followers that serve as input for the differential amplifier. The output voltage has to be 0.65 V when the difference between the input voltages is 2 V. In other words, we could write for the gain:

$$G = \frac{V_{BE\text{(sat)}}}{\Delta V} \cdot \frac{V_{max}}{V_{bat}} = \frac{0.65}{2} \cdot \frac{42}{12} = 1.1375$$  \hspace{1cm} (D-3)

To approximate this, resistors with the values of 15 kΩ and 13.15 kΩ were used. The latter resistance was obtained with two resistors in series.
Figure D-1: The circuit used to drive the Q-Diode.


# Glossary

## List of Acronyms

<table>
<thead>
<tr>
<th>Acronym</th>
<th>Definition</th>
</tr>
</thead>
<tbody>
<tr>
<td>ASIL</td>
<td>Automotive Safety Integrity Level</td>
</tr>
<tr>
<td>BSG</td>
<td>Belt-driven Starter Generator</td>
</tr>
<tr>
<td>EPS</td>
<td>Electrical Power Steering</td>
</tr>
<tr>
<td>ETA</td>
<td>Event Tree Analysis</td>
</tr>
<tr>
<td>EV</td>
<td>Electric Vehicle</td>
</tr>
<tr>
<td>FCV</td>
<td>Fuel Cell Electric Vehicle</td>
</tr>
<tr>
<td>FO</td>
<td>Fail-operational</td>
</tr>
<tr>
<td>FMEA</td>
<td>Failure Mode and Effects Analysis</td>
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<tr>
<td>FS</td>
<td>Fail-safe</td>
</tr>
<tr>
<td>FSIL</td>
<td>Fail-silent</td>
</tr>
<tr>
<td>FTA</td>
<td>Fault Tree Analysis</td>
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<tr>
<td>HA</td>
<td>Hazard Analysis</td>
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<tr>
<td>ICE</td>
<td>Internal Combustion Engine</td>
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<td>iHEV</td>
<td>intelligent Hybrid Electric Vehicle</td>
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<tr>
<td>MTTF</td>
<td>Mean Time to Failure</td>
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<tr>
<td>PHEV</td>
<td>Plug-in Hybrid Electric Vehicle</td>
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<tr>
<td>PWM</td>
<td>Pulse Width Modulation</td>
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<tr>
<td>RPM</td>
<td>revolutions per minute</td>
</tr>
<tr>
<td>SSFR</td>
<td>Standstill Frequency Reponse</td>
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