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Electromechanics & Power Electronics

Master of Science Thesis

High Frequency
Galvanic Insulated
Shore Power Connections

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The department Electrical Engineering
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 faculteit elektrotechniek
Never try to snub a technician’s mind.

Positive words are much more adequate to prevent sparks.
Summary

Grid power connections between ships and docks always should be galvanically isolated. In this thesis, first the necessity of galvanic insulation will be explained in chapter one. Also, the general differences between the standard 50Hz laminated core and the presented High Frequency Switch Mode transformer will be shown. As some of the main advantages of the HF transformer, the reduction in size and weight and the increased capabilities for monitoring and output voltage control are explained.

The Dual Active Bridge topology will be discussed as implementation of a HF Switch Mode transformer. In order to gain insight in the working principle and possibilities of the DAB, chapter two will discuss the theoretical background. Subsequently, a design evaluation of the Mass GI, the first 230V, 3.5kVA HF galvanic insulator for marine applications will be made in chapter three. Improvements to the design will be presented, together with their effect on the circuit efficiency or robustness.

Based on these DAB schematics, also a 120V, 30A push pull layout will be presented in chapter four. The most important component and sub circuit design processes will be discussed. A working model has been built to perform tests on efficiency, thermal energy flow and robustness. The possibility of a 3.5kVA, 120/230V in, 120V/230V out HF galvanic insulation transformer will be presented.

In chapter five and six, two technologies will be discussed to enable a variable ratio between the input and output voltage. First of all, the Mass GI 230V Dual Active Bridge converter will be redesigned. A DSP will now control the converter and implement a phase shift between the primary and secondary H-bridge. For the current design constraints, this solution however will not work for high current values. Together with the theoretical background of chapter two, it is recommended not to use this technology for a controllable HF galvanic insulator.

Secondly, a theoretical model of the 230V Dual Active Bridge design with primary boost inverter circuit will be presented in chapter six. Two possibilities for practical implementation are given, together with their advantages and drawbacks. In spite of the increased number of components, in relation to the phase shift controlled DAB, the eventual solution will prove to be the better option to implement a variable ratio between the primary and secondary voltage levels.

Eventually, in chapter seven a number of conclusions and topics for further discussion and research will be presented.
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1 Introduction

In order to obtain the title of Master of Science in Electrical Engineering at the Eindhoven University of Technology, a final thesis assignment of nine months has to be fulfilled. Since I wanted to do my thesis at a company and in Power Electronics, Arno van Zwam from the company Mastervolt and I exchanged the first emails November 2005.

This eventually led to an assignment, based on the Mass GI, a high frequency galvanically insulated transformer for shore power connections. It had entered the market in the spring of 2006 and was the first one in its range. My task eventually was to investigate the possibilities of controllability of the output voltage and the connection to various grid voltages. In order to do so, I first would have to analyze the current product, evaluate the options for both input and output voltage scalability and then implement the necessary changes.

1.1 Mastervolt

In order to see the place, which the Mass GI is taking in in the product line of Mastervolt, first a general overview will be given on the company and its products.

Founded in 1991, Mastervolt is a world leader in the supply of electrical power solutions to a wide variety of markets. Its mission is to make dependable AC or DC available to all those who require power at locations where no public utility is available.

Mastervolt offers a wide range of products to customers in the global marine, mobile and renewable energy markets. On the shore, houses and buildings are powered by Mastervolt solar inverters in countries around the world. Countless boats sail the oceans, sea, lakes and rivers fitted with Mastervolt systems. For mobile appliances, Mastervolt can supply auxiliary power supplies and inverters.

The Mastervolt name is synonymous with independent power. The smart solutions are a key element in getting the most out of 230V or 120V AC power and 12, 24 or 48V DC appliances, for example by providing DC back-up sources and power conversion products that convert DC power into a clean and stable non-stop AC power or advanced switch mode battery chargers for various mobile and marine propulsion devices. Due to the efficient design, energy loss is set at a minimum level.

1.2 Galvanic insulation of shore power connections

In general, the task of an insulation transformer is to eliminate any electrical ground current between AC shore power and the boat. To gain insight in the properties of galvanic insulation, the two main reasons for implementation in boats are now explained.

1.2.1 Safety

Looking at standard on-shore installations, safety of the electrical installation is guaranteed by fuses and Ground Fault Current Interrupters, GFCI in short. A GFCI will measure the currents flowing through the line and neutral wires. If these currents are not equal, for example in case of a short circuit or current leakage to ground, the GFCI will trip, as the resulting current is flowing through another current path via a ground connection.
A common place to find galvanic insulated transformers is the bathroom: many homes are equipped with small galvanic isolated transformers in order to prevent potentially harmful currents and decrease the risk of shock in case of an equipment failure, for example due to the high humidity. This, because of the fact that now both outputs need to be touched, in order to get electrified, instead of only the phase wire in normal grid connections.

Since the electrical equipment on a boat however is also surrounded by water, which can be a good conductor, especially sea water, extra safety is also needed in this case.

Incorrect grounding without galvanic insulation can then be very dangerous. In this case, a short circuit to ground can lead to an alternative path for the current flow to earth ground through the water. This can threaten swimmers, or possibly through persons stepping on the boat from a metal dock or reaching out to touch a metal piling from the boat. If the ground connection happens to be broken or disconnected somewhere between the boat and the grounding rod on shore, a harmful current, due to a line to ground fault, can flow directly into the water.

The danger to swimmers in salt water is less hazardous because the salt water, being an excellent conductor, will likely pass the current directly down to earth ground under the boat. The danger to swimmers in freshwater is considerably greater. Because fresh water is a poor conductor, the voltage potential of the electric field will extend much further out from the boat as the current seeks a better path to ground, which a swimmer with outstretched arms and legs can easily provide through their body. This can cause muscle seizure and possibly drowning.

1.2.2 Galvanic Corrosion
Boats can be made of various materials. The hulls of light, more expensive boats are made of aluminum, but most recreational boat hulls are made out of Fiberglass reinforced plastic (FRP), more commonly just called fiberglass. One of the main advantages of fiberglass boats is the fact that if properly built, they are extremely strong, do not rust, corrode or rot and are relatively easy to maintain. The downside however is the lifetime of fiberglass hulls. Besides these materials, also still quite a lot of recreational boats are made of steel. Disadvantage however is the high weight of this kind of boats. Also other materials are used, for example most propellers are made of bronze, brass or steel. These metal parts usually are connected to the ground connection of the electrical circuit.

Normally, the usage of different materials will not have a great influence on the boat. However, one of the characteristic features of metals and alloys is their ability to form a galvanic cell when being placed in an electrolyte, in our case the surrounding water with ionized minerals. Seawater is a perfect electrolyte for metals.

Looking at the chemical reaction in a galvanic cell, it is clear that a redox reaction is taking place. The term redox comes from the two concepts of reduction and oxidation. It can be explained in simple terms:

- Oxidation describes the loss of an electron by a molecule, atom or ion
- Reduction describes the gain of an electron by a molecule, atom or ion

The anode is defined as the electrode at which electrons come up from the cell and oxidation occurs, and the cathode is defined as the electrode at which electrons enter the cell and reduction occurs. Each electrode may become either the anode or the cathode depending on the voltage applied to the cell. However, as we will see, connection of the ground terminal of boats to a common shore connection will usually define the anode.
High frequency galvanic insulated shore power connections and cathode, due to the electrochemical potentials of the used metals and alloys. For a number of materials, this characteristic value is mentioned in addendum C2. To make this table more useful, figure 1.1 contains a compact overview of commonly used metals and alloys in the offshore industry.

![Electrochemical Potential Chart](image)

**Fig. 1.1. Electrochemical potential values for different materials used in offshore technologies [27].**

Bringing this theory into practice, it is clear that as soon as the harbor is reached and a shore power connection is made, problems may appear. Due to the usage of varying materials, a difference in potential will appear between electrically connected parts of different metals, which will cause a current to start flowing and the metal parts to ionize.

Most obvious examples are the currents between the aluminum hull or the bronze propeller of a boat and grounded metal shore walls, shown in figure 1.2 as I₁ and I₂. First of all, ions are needed in the water. The simplest way to provide these is the ionization of salt:

\[
NaCl_{(s)} \rightarrow Na^+_{(aq)} + Cl^-_{(aq)} \quad (1.1)
\]

The electro-chemical reaction between the aluminum and the steel is then as follows:

\[
Fe^{2+}_{(aq)} + 2e^- \leftrightarrow Fe_{(s)} \quad E_0 = -0.44V \quad (1.2)
\]

\[
Al^{3+}_{(aq)} + 3e^- \leftrightarrow Al_{(s)} \quad E_0 = -1.66V \quad (1.3)
\]

The electrochemical potential difference between the steel wall and the aluminum hull theoretically would then be 1.22 V. Due to the fact that the conditions are not standard, the Nernst equations should be used for a good estimation of this voltage. However, since the actual value of the potential difference is not important in this case, no further investigation will be done in this field. The most important conclusion is the fact that the less noble material, in this case the aluminum, will ionize.
Another possibility is the difference in potential between two electrically boats, where one has an aluminum hull and the other one a bronze screw. The resulting current flow is shown in figure 1.2 as $I_1$.

A less common, but even better way to boost this process is an incorrect connection of batteries. When for example the ground connection of the shore is connected to the positive battery terminal and the boat hull is connected to the negative terminal, a constant potential difference of 12V or 24V is present between the shore and the boat. This will boost the process even more.

Due to these potential differences, a current will start flowing, shown with the arrows in figure 1.2, thus causing the metals to ionize. This is called galvanic corrosion and can cause severe damage in a very short period of time.

A standard solution against corrosion is to protect boats with a zinc anode, which will dissolve first, due to the high electrochemical potential, as shown in figure 1.1. Since this zinc anode actually is being "sacrificed" in order to protect other metal parts, it is important to regularly check the status of this anode. Another problem is the fact that when a boat is protected by a zinc anode, but a boat nearby has no zinc, the protected boat's zinc will corrode rapidly while protecting both boats. After the zinc has dissolved, the least noble metal on either boat will begin to corrode. This however will only protect against small potential differences, not against potential voltages of several volts between connections of for example shore grounds and ship hulls.

A better solution is galvanic isolation of the shore power connection. When a isolation transformer is installed, the electrical connection of the ship's hull with the earth connection is no longer there, so no current will flow here.

### 1.3 Possible technologies

As mentioned before, the main task of the Galvanic Insulator is to transfer power from the shore to the ship without an electrical connection between these two terminals. For many years already, this task has been done by standard iron laminated LF transformers.

The size and weight of a transformer is primarily a function of the saturation flux density of the core material and maximum allowable core and winding temperature rise.
power throughput density is inversely proportional to the operating frequency, therefore increasing the frequency allows higher utilization of the magnetic core and a reduction in the transformer size.

The subject of this final thesis therefore is the first HF galvanic insulated transformer, available for various applications. In order to gain better insight in the positive and negative points, a short overview will be given on the characteristics of both transformers. Eventually, also a device, also used for prevention of corrosion and also called “galvanic isolator” will be mentioned in order to complete the total image on Galvanic Insulation.

1.3.1 50Hz Galvanic Insulation Transformers

The history of the classic transformer goes back to the 29th of August, 1831, when Michael Faraday invented the “induction ring”, shown in figure 1.3.a, when he wound several different wires near each other on a wooden spool. The induction ring was therefore the basis for the transformer, which is used today in electric power systems to step up and step down voltages to different levels.

The transformer however also can have another task: galvanic insulation of power connections. When looking at the Mastervolt Galvanic Insulation transformers, it is clear that they are specially designed for applications on board ships and for heavy industrial applications. They are commonly used to avoid corrosion on seagoing vessels by galvanic insulation of the shore ground and ships ground.

The step up or step down of the incoming voltage is possible, as this kind of transformers can be produced as a multi-tap transformer. In this case, the primary and secondary windings are divided into two or more winding sets, which can be connected in series or parallel. A 120V input can therefore be connected to a 230V output by simply changing the transformer connections. A 3.5kVA model is shown in figure 1.3.b, further data on weight and size can be found in table 1-2.

A big disadvantage, especially when reaching higher power demands, is the weight of the transformer: due to the relatively low frequency of 50Hz, the transformer cores have to be quite big. This results again in large winding distances. All together, this leads to high volumes and high weights. In on-shore applications, this might not be a big problem, but in confined spaces as on boats, these factors surely are a big issue.

![Fig. 1.3. Transformers through the ages. a) First transformer, of Michael Faraday, b) standard 3.5kVA 50Hz transformer, c) 3.5kVA HF transformer Mass GI [27].](image-url)
1.3.2 Switch mode galvanic Insulation Transformers

The High Frequency switch mode transformer principle is quite common in Power Electronics. Also for shore connections, HF transformers are used, but mainly in battery chargers. The advantages can easily be seen: due to the high switching frequency, in this case 25kHz, the internal main transformer can be much smaller. As a result, the total product will not only reduce in volume, but also in weight, both very important on smaller boats. Another advantage is the possibility to control the output voltage and power, as this final thesis will show.

Of course, this transformer also has disadvantages. The efficiency is lower, in the case of the 230V Mass GI, shown in figure 1.3.c about 93%. Furthermore, due to the implementation of power electronic circuits, the robustness of the transformer is lower than the classic 50Hz transformer. High harmonics and spikes are potentially more harmful for this type of transformer. However, looking at the positive aspects, they can easily outnumber the current disadvantages of HF galvanic insulators.

When a market review on HF Galvanic Insulation transformers is done, only one company appears to fill this upcoming niche: Mastervolt. Other manufacturers of classic 50Hz transformers so far haven't filled in this gap, in spite of the possible advantages.

1.3.3 Galvanic isolators

A possibility which only will handle the corrosion problem is the electronic galvanic isolator, often installed in yachts. The only task of this device is to interrupt the DC path from boat to dock, or to other boats. In order to achieve this goal, schematics as shown in figure 1.4 are applied.

Low voltages will not make the diodes conduct. This way, the current flow for potentially harmful DC currents, causing corrosion, is disabled. The most efficient isolators have a capacitor, which allows low levels of only AC current to immediately bypass the diodes to the shore ground.

Safety circuits, using the ground wire for safety will now still work, as any short circuit between line and ground will lead to a potential difference between shore ground and boat ground, which is higher than the threshold voltage, caused by the diodes.

This device however does not totally galvanically isolate the shore and boat side power connection, but will only block low voltage differences, under a few volts, between the shore and the hull. Since this type of equipment can only solve one part of the problem, this type of galvanic insulator is not further investigated.

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Fig. 1.4. Corrosion protection by a) transformers and b) electronic galvanic insulators.
1.4 Technology review by product demands

Looking at the current market for Galvanic Insulated transformers, it is clear that the product itself is good, but old-fashioned. Due to an increasing demand for this type of transformers, it is time to review the current specifications, in order to gain a leading position in the market for Galvanic Insulation transformers. A number of specifications and possibilities of both standard 50Hz and the new HF transformers are given in table 1-1. As can be seen, the HF transformer is an interesting product to be investigated.

<table>
<thead>
<tr>
<th>50Hz transformers</th>
<th>HF transformers</th>
</tr>
</thead>
<tbody>
<tr>
<td>+ Robust</td>
<td>+ Light, small</td>
</tr>
<tr>
<td>+ Switchable output voltage</td>
<td>+ Continuous controllable output voltage and current</td>
</tr>
<tr>
<td>+ Capable of handling high peak power</td>
<td>+ Implementation of active filter</td>
</tr>
<tr>
<td>+ High efficiency: 97+%</td>
<td>+ Silent</td>
</tr>
<tr>
<td>- Heavy, voluminous</td>
<td>- Able to handle limited high peak currents.</td>
</tr>
<tr>
<td>- Not controllable</td>
<td>- Lower efficiency: 93%</td>
</tr>
<tr>
<td>- 50Hz humming sound</td>
<td></td>
</tr>
</tbody>
</table>

Looking at the customers this product is intended for, it is clear that equipment, bought in one country, will probably also be used in another country. This can cause several problems, as can be seen in the World Power Guide [2]. In order to be sure that the HF transformer will meet the customers’ demands, the most important customer issues will now be discussed.

1.4.1 Robustness and reliability

Two of the most important factors for these applications are robustness and reliability. A power connection should never fail, in order to provide maximum comfort and safety to the customer. Looking at these demands, the 50Hz transformer would win at first sight. Due to the simple layout, there are only few elements that can break down.

An electronic HF transformer however has much more components, possibly leading to a higher error rate. However, due to thorough design, robustness can also be assured for this type of transformer, thus leading to a comparable Mean Time To Failure. Also, the usage of electronic safety circuits can detect over voltages, over currents and over temperatures, thus leading to a safer product. A standard 50Hz transformer usually only has a over temperature safety circuit and can possibly lead to higher damage to the connected equipment or to itself in case of over voltages or over currents. Therefore, external safety circuits are necessary when using 50Hz transformers.

1.4.2 Grid voltage Controllability

The grid voltage level can have a different value than the customer is used to. In the Netherlands, for example, customers are used to an RMS AC voltage level of 230V. However, once traveling abroad, the grid voltage can differ from 220V in Portugal and Scotland to 240V in Malta and parts of the UK. Certain parts of Spain even still are electrified with a 120V AC grid. Going outside of Europe, this last grid voltage is quite common, next to grid voltages of 110V and 127V. Quite dangerous for electrical equipment however are single phase grid voltages of 250V, for example to be found in certain parts of Africa and South-America.
The mentioned values are average grid voltage values. They can differ in value, which can even increase the amplitude difference between the given grid voltage and the preferred voltage level. Also low AC grid values, for example caused by long, inadequate wiring, can cause this problem. A controllable output voltage is therefore a valuable product feature.

1.4.3 Grid frequency controllability

As can be seen in the World Power Guide, [2], not only the grid voltage can differ between various countries. Also the grid frequency can vary, as both 50Hz and 60Hz are common for AC power lines. In Europe, only 50Hz grids are present, but many other countries, like the US, use 60Hz. Another reason to control the grid frequency is the usage of a generator of poor quality. Since the power control in these cases often is controlled by the rotational speed of the generator, the output frequency can also differ, for example when heavy loads are connected.

Classical 50Hz transformers are unable to handle this problem. Due to the working principle, it is not possible to change the output frequency, as the output voltage is being generated by the primary induced field. Since this is varying with 50Hz, the output voltage will also vary with this value.

HF transformers however can be able to fix this problem. The only problem is the fact that for this case, extra power electronic circuits are necessary, as this is only possible by making a constant DC bus between the AC input and AC output. To ensure a bidirectional power flow, for example first a bidirectional AC-DC converter has to be made, then a bidirectional DC-DC converter, followed by a bidirectional DC-AC converter. This way, the output voltage can still be brought to high levels, even though the input voltage is lower due to the frequency difference.

This way of transformation however is much more complicated than the current HF Galvanic Insulation transformer principle, therefore variation of output frequency will not be taken into account in the following designs.

1.4.4 Product weight and size

The most important factor of the new HF transformer can be found in its weight and size. As mentioned before, the size of a transformer is related to the transferable power and the frequency, at which this power is transferred. The higher the power level, the bigger the transformer, but the higher the switching frequency, the smaller the transformer can be.

When making a comparison between the several types of encapsulated 230V 3,5kVA Galvanic Insulation transformers, it is clear that the classical 50Hz transformer does not stand a chance, as is shown in table 1-2. The Mass GI relatively saves up to 81% in weight and up to 51% in volume and therefore is of great value in confined spaces, such as marine applications.
High frequency galvanic insulated shore power connections

Table 1-2. Comparison of various Galvanic Insulation transformers.

<table>
<thead>
<tr>
<th>Manufacturer</th>
<th>Type</th>
<th>Weight</th>
<th>dimensions</th>
<th>Volume</th>
</tr>
</thead>
<tbody>
<tr>
<td>Mastervolt</td>
<td>Mass GI, HF transformer, 3.5kVA</td>
<td>6 kg</td>
<td>340 x 261 x 144 mm</td>
<td>12.8dm³</td>
</tr>
<tr>
<td>Mastervolt</td>
<td>Ivet D, torroidal core, 3.5kVA</td>
<td>23 kg</td>
<td>407 x 295 x 220 mm</td>
<td>26.4dm³</td>
</tr>
<tr>
<td>Victron Energy</td>
<td>Isolation Transformer, 3.6kVA</td>
<td>23 kg</td>
<td>362 x 258 x 218 mm</td>
<td>20.4dm³</td>
</tr>
<tr>
<td>ProMariner</td>
<td>Prosafe Isolation Transformer 3.6kVA</td>
<td>29 kg</td>
<td>178 x 191 x 394 mm</td>
<td>13.4dm³</td>
</tr>
<tr>
<td>Energy Solutions</td>
<td>Hull Isolation Transformer 3.6kVA</td>
<td>28 kg</td>
<td>362 x 258 x 218 mm</td>
<td>20.4dm³</td>
</tr>
<tr>
<td>Charles Industries</td>
<td>Iso Guard Isolation Transformer 3.8kVA</td>
<td>35 kg</td>
<td>245 x 267 x 223 mm</td>
<td>14.6dm³</td>
</tr>
</tbody>
</table>

1.5 Switch mode converter overview

As has been shown in the previous paragraph, it is obvious that switch mode converters also can offer many interesting possibilities for the market of shore power connections. Due to the fact that the total converter size should be as small as possible, all converters with a DC link are not interesting for this application. Therefore, only a primary conversion of LF, 50-60 Hz, AC to HF AC and a secondary conversion of HF AC to LF AC need to be made.

As a result, the Dual Active Bridge topology is chosen for the current Mass GI design. However, this is not the only topology, which will be investigated in this report. For the 120V version, the Push Pull topology will be investigated. Also the possibilities for controllability of the output voltage will be discussed for both topologies.

1.5.1 Dual Active Bridge

The dual active bridge topology, which will be discussed in detail later on, is based on the fact that the 50Hz grid voltage can be switched at a high frequency. This results in a decrease of the energy storage which needs to be processed in the main transformer each cycle. In order to do so, both on the primary and the secondary side of the main HF transformer, an H-bridge is used to switch the polarity of the voltage over the main transformer input and output.

1.5.2 Push Pull topology

Another way to implement a switch mode converter is the push pull technology. Using this technology, the main transformer has two primary and two secondary coils, of which per half period only one out of two is powered. The core magnetic field polarity will be alternated at a high frequency again. This enables again a compact main transformer design. However, due to the fact that always only two of the four coils will be used for power transfer, the main transformer design will be more voluminous, compared to the DAB main transformer design.
1.5.3 Output voltage controllability
In order to meet demands regarding output voltage controllability, several techniques will be investigated. Most techniques will only be tested on the original 230V version, but are also applicable to a future 120V version. The techniques, which will be discussed, are:

- DAB phase shift control
- Boost converter control

The first option is the most innovative, but also the most difficult one for physical implementation. The second option might lead to high losses at high power. The third option is simple, robust, but not linear controllable, due to the fact that discrete voltage steps will be made.

1.6 Design, simulation and measurement processes
As shown in the previous paragraphs, this report will present a variety of designs for a HF galvanic insulator of 3.5kVA. The current design of the Mass GI is given, but can be improved. In order to design and test these ideas, in the following chapters first the general DAB theory will be discussed, followed by several applications.

Due to the interesting broad scope of this assignment, taking inverter designs from theory to working products, also a wide range of activities have been performed. These activities imply literature research, calculations, circuit design, simulations and measurements. This report will only summarize the results of these processes.

In order to maintain the overview in this report, most of the calculations and measurement results are presented in the addendum. Also control schemes, component data and supplementary circuit designs can be found there. All supporting simulation programs, together with an overview of the current power electronic simulation software packages, are mentioned in addenda C3 and C4. The used measurement setups are presented in addenda A7, B1 and B6.
2 Dual active Bridge topology

For the past years, High Frequency (HF) link power converters have been used more and more in a wide range of applications. Not only low voltage, low power inverters for consumer electronics, but also high power inverters for industrial applications, such as motor drives or active filters for Power Quality are using HF links for power conversion. The main advantage of an increase in applied switching frequency is the size reduction of magnetic components, due to the lower amount of energy, which needs to be stored in the core.

HF link converters were first used for DC-DC conversions [7, 8, 9, 30, 32, 42, 46]. This is caused by the fact that this setup will simplify the switching scheme, as will be shown in a later paragraph. Also DC-AC converters, [1, 16], and AC-DC converters, [29], were developed. For AC-AC conversion the switches will have to be bidirectional for both voltage and current [13, 14, 21, 43].

One of the options for a HF link is the usage of a Dual Active Bridge converter [4, 5, 6, 12, 15, 23, 25, 26, 34, 37], as shown in figure 2-1. As can be seen in this figure, the topology consists of a primary and a secondary H-bridge. These H-bridges are alternating the polarity of the voltage over the main transformer T1 at frequencies of a few kilohertz, up to hundreds of kilohertz. The HF terminals of both bridges are connected by a HF transformer, enabling a theoretical power flow in both directions.

Usually, the main transformer T1 is designed in such a way that the leakage inductor L1 is as low as possible. When there is no phase shift between the primary and secondary bridge and the switches and main transformer would be ideal, then the output voltage would be equal to the input voltage. A lower grid voltage would therefore also lead to a lower load voltage.

![Fig. 2.1. Standard scheme of a Dual Active Bridge converter.](image)

By adding a calculated amount of leakage inductance into the main transformer, phase shift control between the primary and secondary H-bridge could control the power flow between the primary and secondary side. As a result, the DAB topology can even ideally be described as an inductor, powered by two controlled square wave voltage sources, $V_{\text{grid}}$ and $V_{\text{load}}$.

The shown primary and secondary snubbers are shown for all applications, as the following paragraph will show that they are unmissable in AC conversion systems. Therefore, they are also already present in this model.
The now called DAB topology first showed up in literature in the 1970's and was first used for DC-DC conversion, but later on also for DC-AC and AC-AC conversion. However, an efficient implementation of this switching scheme was only possible with semiconductor switches, which could operate at higher frequencies. As the demand for these high frequency components, as MOSFETs and IGBTs grew and the technology became more and more familiar, the components improved and therefore also the inverter characteristics.

The following paragraphs will now explain the working principle of the DAB and the belonging control schemes. Also the task of the main components, needed for a correct implementation, will be discussed.

### 2.1 Switching schemes

In order to gain more insight in the general switching principle, the general switching schemes of the DAB have been given in figure 2.2. The mathematical model will be given in the next paragraphs.

Four steady state switching states of the DAB are shown, together with two extra states: the one in which all switches are off, figure 2.2.e and the one in which all switches are on, figure 2.2.f. It is clear that the switching scheme in figure 2.2.f should never happen, as this will lead to a primary and secondary short circuit.

Since a minimum switching time is required for deactivation of the active switches and activation of the deactivated switches, a certain dead time will have to be present, as shown in figure 2.2.e. For completeness of the switching schemes, also this state is mentioned, but will not be further taken into account in this chapter. The practical consequences will be shown in the next chapter.
2.2 **DC DAB versus AC DAB**

Most DAB circuits, which are presented in the literature, are focused on DC-DC conversion. As will be shown, the changes needed for AC-AC conversion present quite a challenge.

One of the important features of the DC Dual Active Bridge is the fact that the switches contain anti parallel diodes, which can act as a current path for freewheeling inductor currents, when the switches are turned off. At this moment, the voltage over the leakage inductance of the main transformer will change its polarity, in order to maintain the current flow. The absence of anti parallel diodes in a switch would then lead to a drastic increase in the voltage over the switches and main transformer.

An AC DAB will not have an anti-parallel path, as will be shown in paragraph 5 of this chapter. Therefore, an adequate snubber circuit is unmissable during the short times, when all switches are turned off in order to change the current path. These snubber circuits are already shown in the figures 2.1 and 2.2, and will be further explained in paragraph seven of this chapter.

For small signal calculation purposes however, the usage of AC signals at the input and output can be simplified. The assumption for this simplification is the fact that the value of this AC grid frequency is significantly lower than the value of the switching frequency, used to control the H-bridges.

This assumption is very important, as it enables us to split up the behavior into two parts: the LF grid and load voltage behavior and the HF transformer voltage behavior, at which the 50Hz grid voltage can be estimated as a constant value.

2.3 **Assumptions for an ideal AC DAB model**

In order to understand the working principle of the Dual Active Bridge topology, the formulas and switching schemes, describing an ideal DAB with phase shift control, will be explained in order to gain insight in the control principle behind the generation of the output voltages, as shown in figure 2.3. In the DAB system model, shown in figure 2.1, the following assumptions are valid:

- All semiconductors are ideal and can therefore be replaced by switches
- System losses, for example in control, snubber and magnetic circuits are neglected
- No grid filter present, an ideal filter is only used on the load side
- The load is resistive
- The switching frequency is significantly higher than the input and output voltage frequency, thus enabling a LF and a HF model.

Using these simplifications, a standard model can be designed. This model will be able to give general predictions, but will not give any information yet on factors like short circuit behavior, inductive or capacitive loads or safety circuits.
2.4 Ideal AC DAB model without phase shift

Due to the assumptions, just made, the HF DAB converter can now be modeled, regarding the slow alternating AC input voltage as a "DC" source. This can be seen in the simplified graph in figure 2.3. In this figure, a 50Hz sinusoidal wave, switched at 1kHz already can be estimated as a square wave. This can also be seen in the formula for the transformer voltage. The standard expression is given in equation 2.1

\[ u_{\text{transformer}}(t) = m(t) \cdot \hat{V}_{\text{grid}} \cdot \sin(\omega_{\text{grid}} \cdot t) \] (2.1)

The amplitude factor \( m(t) \) can be described with equation 2.2.

\[ m(t) = \begin{cases} 1 & nT_s < t < \left( n + \frac{1}{2} \right) T_s \\ -1 & \left( n + \frac{1}{2} \right) T_s < t < (n+1)T_s \end{cases} , n = \text{floor} \left( \frac{t}{T_s} \right) \] (2.2)

In the basic situation, when there is no phase shift between the primary and secondary bridge, this signal will be "unwrapped" again, resulting in:

\[ v_{\text{load}}(t) = m(t) \cdot u_{\text{transformer}}(t) \] (2.3)

\[ v_{\text{load}}(t) = m(t) \cdot m(t) \cdot \hat{V}_{\text{grid}} \cdot \sin(\omega_{\text{grid}} \cdot t) \] (2.4)

\[ v_{\text{load}}(t) = \hat{V}_{\text{grid}} \cdot \sin(\omega_{\text{grid}} \cdot t) \] (2.5)

As shown before, the output voltage would then ideally be equal to the input voltage. The resulting voltage over the transformer is now given in formula 2.6.
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\[
\begin{align*}
    u_{\text{transformer}}(t) &= \begin{cases}
        \dot{V}_{\text{grid}} \cdot \sin(\omega_{\text{grid}} \cdot t) & nT_s < t < \left( n + \frac{1}{2} \right) T_s \\
        -\dot{V}_{\text{grid}} \cdot \sin(\omega_{\text{grid}} \cdot t) & \left( n + \frac{1}{2} \right) T_s < t < (n+1)T_s
    \end{cases}, \quad n = \text{floor}\left( \frac{t}{T_s} \right) \tag{2.6}
\end{align*}
\]

In this formula, the slow varying sinusoidal grid voltage has become the amplitude factor. As can be seen, the low frequency and high frequency signals both have their own influence on the DAB behavior. Therefore, the LF and HF behavior will first be calculated separately.

Ideally, the transformer leakage inductance for a phase shift of zero rad, is equal to zero Henry. However, since a transformer will always have a leakage inductance, and regarding the fact that, for a controllable DAB model, a leakage inductor is one of the basic components, this will now also be taken into account.

Now, the switch control will be further explained. In order to avoid short circuits in the primary or secondary bridge and to prevent disconnections of the main inductor, the assumptions in equations 2.7 to 2.10 are valid for the switches:

\[
\begin{align*}
    S_{1p}(t) + S_{1n}(t) &= 1; & S_{1p}(t), S_{1n}(t) &\in \{0,1\} \tag{2.7} \\
    S_{2p}(t) + S_{2n}(t) &= 1; & S_{2p}(t), S_{2n}(t) &\in \{0,1\} \tag{2.8} \\
    S_{3p}(t) + S_{3n}(t) &= 1; & S_{3p}(t), S_{3n}(t) &\in \{0,1\} \tag{2.9} \\
    S_{4p}(t) + S_{4n}(t) &= 1; & S_{4p}(t), S_{4n}(t) &\in \{0,1\} \tag{2.10}
\end{align*}
\]

It is important to give the high frequency square wave control signal a 50% duty cycle, in order to equally distribute the power over the switching components. Also will this be of importance when implementing the correct gating circuit. When the duty cycle would not be equal to 50%, the implementation of gating transformers might give problems, as they would go into saturation.

Switches with the value “1” are closed and therefore conducting, switches with the value “0” are opened. The phase shift between the legs of each H-bridge is set at the standard 180°, adding the following rules:

\[
\begin{align*}
    S_{1p}(t) + S_{2p}(t) &= 1 \tag{2.11} \\
    S_{3p}(t) + S_{4p}(t) &= 1 \tag{2.12}
\end{align*}
\]

Since the phase shift between the primary and secondary bridge is set at zero, also the following rules are valid:

\[
\begin{align*}
    S_{1p}(t) &= S_{3p}(t) \tag{2.13} \\
    S_{2p}(t) &= S_{4p}(t) \tag{2.14}
\end{align*}
\]

This will result in a simultaneous switching pattern for the primary and secondary H-bridge, as also shown in fig. 2.3.

\section{Ideal AC DAB model with phase shift}

There are several options to make the DAB controllable. The introduction of a phase shift between the primary and secondary bridge is a possibility, but also a phase silt between the gating signals in both legs of a single H-bridge can influence the output power.
Therefore, now the DAB behavior at varying input and output voltages and at a varying load will be regarded. In general, four situations can be described, each varying in the following variables:
- Effective amplitude difference between $V_{\text{grid}}$ and $V_{\text{load}}$
- Sign of the implemented phase shift $\varphi$

### 2.5.1 General switching behavior

For a general understanding of the working principle, a few situations are shown in figure 2.4. In all situations, there is a difference between the effective value of the input and the output voltage. The situations in figure 2.4a and 2.4b show the desired wave forms.

![Diagram](image-url)

**Fig. 2.4. Switching behavior at various in- and output voltages and various phase shifts.**
The graphs in 2.4.c and 2.4.d show the effect, when the phase shift duration does not have the correct value. Graph 2.4.e and 2.4.f show what happens if the phase is implemented at the incorrect flank, thus leading to even more spikes.

The voltages $u_p(t)$ and $u_s(t)$ are idealized. Due to this idealization, also $I_p(t)$ and $I_s(t)$ can easily be drawn. It is clear that the current trough the inductor has two derivative values per half period, a low and a high one. These derivative values are caused by the voltage difference between $u_p(t)$ and $u_s(t)$, as shown in equation 2.15.

$$u_L(t) = u_p(t) - u_s(t) = L \cdot \frac{di_L(t)}{dt} \quad (2.15)$$

As can be seen in figure 2.4, a big problem in phase shifted DAB can at the switching points, when the timing is not fully correct. Also a practical problem will arise, when the added leakage inductor is disconnected, due to the switching of the H-bridge. Due to the placement of the added inductor at the primary side of the main transformer, the primary snubber circuit will also receive the most energetic switching peaks. However, due to the fact that the secondary bridge is switching at another time than the primary bridge, the voltage peaks, due to the energy in the inductor, also need to be snubbed in the secondary snubber.

As a result, not only will the snubber circuits have to be activated twice as much as in the DAB without a phase shift. The energy, which needs to be snubbed, will also increase, due to the larger amount of energy in the leakage inductor.

### 2.5.2 Switching behavior calculations

Figure 2.4 showed the general switching pattern, together with the locations, where spikes are expected. Now, in this paragraph a mathematical model will be presented, which will help in understanding the small signal behavior.

In this project, only a phase shift $\delta$ between the primary and the secondary bridge will be implemented. Therefore, the equations 2.13 and 2.14 will be replaced by respectively 2.16 and 2.17.

$$S1p(t) = \begin{cases} 1 - S3p(t) & 0 < t < \delta \\ S3p(t) & \delta < t < T/2 \end{cases} \quad (2.16)$$

$$S2p(t) = \begin{cases} 1 - S4p(t) & 0 < t < \delta \\ S4p(t) & \delta < t < T/2 \end{cases} \quad (2.17)$$

Now, also the grid, load and transformer voltages can be calculated. The voltage over the primary terminals of the main transformer is given in formula 2.18 and can be rewritten to 2.19.

$$u_p(t) = (S1p(t) - S2p(t)) \cdot v_{grid}(t) \quad (2.18)$$

$$u_p(t) = (2 \cdot S1p(t) - 1) \cdot v_{grid}(t) \quad (2.19)$$

Due to the fact that the general layout is mirrored on both sides of the main transformer, the following equations are also valid:
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\[ u_1(t) = (S3 \cdot p(t) - S4 \cdot p(t)) \cdot v_{load}(t) = (2 \cdot S3 \cdot p(t) - 1) \cdot v_{load}(t) \] (2.20)

\[ v_{load}(t) = (2 \cdot S3 \cdot p(t) - 1) \cdot u_i(t) \] (2.21)

The current in the inductor is equal to:

\[ L \cdot \frac{d}{dt} i_L(t) = u_p(t) - u_i(t) \] (2.22)

The output current of the secondary bridge can then be calculated by:

\[ i_s(t) = (2 \cdot S3 \cdot p(t) - 1) \cdot i_L(t) \] (2.23)

The current in the load is calculated by:

\[ i_{load}(t) = \frac{v_{load}(t)}{R_{load}} \] (2.24)

With these currents, the capacitor current, and therefore the output voltage can be calculated with formula 2.25.

\[ C \cdot \frac{d}{dt} v_{load}(t) = i_c(t) = i_s(t) - i_{load}(t) \] (2.25)

This can be rewritten as one of the state equations, in equation 2.27.

\[ \frac{d}{dt} v_{load}(t) = \frac{1}{C} \left( 2 \cdot S3 \cdot p(t) - 1 \right) \cdot i_L(t) - \frac{v_s(t)}{R_{load} \cdot C} \] (2.26)

\[ \frac{d}{dt} v_{load}(t) = \begin{cases} - \frac{1}{C} \cdot i_L(t) - \frac{v_{load}(t)}{R_{load} \cdot C} & \text{if } S3 \cdot p(t) = 0 \\ \frac{1}{C} \cdot i_L(t) - \frac{v_{load}(t)}{R_{load} \cdot C} & \text{if } S3 \cdot p(t) = 1 \end{cases} \] (2.27)

Bringing the previous formulas together will result in the other final state equation, 2.28.

\[ \frac{d}{dt} i_L(t) = \frac{1}{L} \left( 2 \cdot S1 \cdot p(t) - 1 \right) \cdot v_{grid}(t) - \frac{1}{L} \left( 2 \cdot S3 \cdot p(t) - 1 \right) \cdot v_{load}(t) \] (2.28)

The voltage over the inductor can be written as formula 2.29:

\[ u_L(t) = L \cdot \frac{di_L(t)}{dt} = \begin{cases} v_{load}(t) - v_{grid}(t) & S1 \cdot p(t) = 0 \land S3 \cdot p(t) = 0 \\ - (v_{grid}(t) + v_{load}(t)) & S1 \cdot p(t) = 0 \land S3 \cdot p(t) = 1 \\ v_{grid}(t) + v_{load}(t) & S1 \cdot p(t) = 1 \land S3 \cdot p(t) = 0 \\ v_{grid}(t) - v_{load}(t) & S1 \cdot p(t) = 1 \land S3 \cdot p(t) = 1 \end{cases} \] (2.29)

Interesting about these relations is the fact that it can be rewritten into equation 2.30:
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This is the basic formula for the phase shift DAB topology. It shows that the output voltage will decrease if a phase shift is implemented in an ideal circuit without leakage inductance. It also shows the working principle of the boost functionality in the DAB: addition of an inductor $L$ can not only lead to decrease, but also to an increase of the output voltage, depending on the applied phase shift.

Another conclusion, which can be drawn, is the fact that always a phase shift needs to be present between the primary and secondary voltages, in order to have a current flow through the main transformer with leakage inductor.

2.5.3 Influence of the output capacitor

Since in a practical implementation the output capacitor will have a finite value, this influence will now also be taken into account, using the previous formulas:

$$i_L(t) = \begin{cases} 
-C \left( \frac{d}{dt} v_{\text{load}}(t) + \frac{v_{\text{load}}(t)}{R_{\text{load}} \cdot C} \right) & S3p(t) = 0 \\
-C \left( \frac{d}{dt} v_{\text{load}}(t) + \frac{v_{\text{load}}(t)}{R_{\text{load}} \cdot C} \right) & S3p(t) = 1 
\end{cases}$$

(2.31)

and:

$$v_{\text{load}}(t) = \frac{(2 \cdot S1p(t) - 1)}{(2 \cdot S3p(t) - 1)} \cdot v_{\text{grid}} - \frac{L}{(2 \cdot S3p(t) - 1)} \cdot \frac{d}{dt} i_L$$

(2.32)

Combined together, the result is:

$$v_{\text{load}}(t) = \begin{cases} 
-(2 \cdot S1p - 1) \cdot v_{\text{grid}}(t) + LC \cdot \frac{d}{dt} v_{\text{load}}(t) + \frac{v_{\text{load}}(t)}{R_{\text{load}} \cdot C} & S3p = 0 \\
(2 \cdot S1p - 1) \cdot v_{\text{grid}}(t) + LC \cdot \frac{d}{dt} v_{\text{load}}(t) + \frac{v_{\text{load}}(t)}{R_{\text{load}} \cdot C} & S3p = 1 
\end{cases}$$

(2.33)

In order to provide the output signal with a smaller voltage ripple, the effect of several values for the output capacitor bank has been simulated. The results, when applied to a switching frequency of 1kHz and a load of 220Ω, are shown in figure 2.5.
This is logical, as the whole circuit can be simplified to the schematics, shown in figure 2.6. Here, it is clear that the output voltage is maximized, due to the RLC filter value.

![Figure 2.6. Simplification of the DAB](image)

### 2.5.4 Calculation of the average output current

Once a delay is applied, the effective output current at a certain interval $t_4 - t_5$, called $I_{load}$, can be calculated, using the drawings in figure 2.7. The phase shift of $\varphi$ rad per half period is shown in figure 2.7.a and 2.7.b as a time delay between $t_1$ and $t_2$. The graph in 2.7.a shows the situation, when the phase shift is still being set at the correct value. The graph in figure 2.7.b shows the steady state situation. In order to make a model, which is applicable in both situations, the situation of graph 2.7.a will first be investigated. Subsequently, the current value $i_1$ can be set equal to $i_2$ and $-i_1$ to $i_4$, in order to get the situation of figure 2.7.b.

![Figure 2.7. Average current calculation per half period](image)

Looking at the graph in figure 2.7.a, the average output current can be calculated with equation 2.32, which can be rewritten to 2.35.

- **For the DAB**
Looking at figure 2.7, it is clear that from $t_1$ to $t_3$ the values of $i_1$ and $i_4$ are correlated with the voltage difference between the slow varying input and output voltage. This relation can be described with equations 2.36 and 2.37.

\[
\begin{align*}
 v_{\text{grid}}(t) - v_{\text{load}}(t) &= L \cdot \frac{di_1(t)}{dt} = L \cdot \frac{\Delta i_1}{\Delta t} & t_1 < t < t_3 \\
 v_{\text{grid}}(t) + v_{\text{load}}(t) &= L \cdot \frac{di_4(t)}{dt} = L \cdot \frac{\Delta i_4}{\Delta t} & t_3 < t < t_5
\end{align*}
\]

Therefore also equations 2.38 to 2.41 are valid.

\[
\begin{align*}
 V_{\text{grid}} - V_{\text{load}} &= L \cdot \frac{i_2 - i_1}{\left(1 - \frac{\varphi}{\pi}\right) \cdot t_S} & t_1 < t < t_3 \\
i_2 &= i_1 + \frac{1}{L} \cdot (V_{\text{grid}} - V_{\text{load}}) \cdot \left(1 - \frac{\varphi}{\pi}\right) \cdot t_S & t_1 < t < t_3 \\
 V_{\text{grid}} + V_{\text{load}} &= L \cdot \frac{i_3 - i_4}{\frac{\varphi}{\pi} \cdot t_S} & t_3 < t < t_5 \\
i_3 &= i_4 + \frac{1}{L} \cdot (V_{\text{grid}} + V_{\text{load}}) \cdot \frac{\varphi}{\pi} \cdot t_S & t_3 < t < t_5
\end{align*}
\]

Combining the previous formulas will result in:

\[
\begin{align*}
 I_{\text{load}} &= \frac{\pi - \varphi}{\pi} \cdot \frac{i_1 + i_2 + \frac{\varphi}{\pi} \cdot i_3 + i_4}{2} \\
 I_{\text{load}} &= \frac{\pi - \varphi}{\pi} \cdot \frac{i_1 + \frac{1}{L} \cdot (V_{\text{grid}} - V_{\text{load}}) \cdot \left(\frac{\pi - \varphi}{\pi}\right) \cdot t_S}{2} \\
 I_{\text{load}} &= \frac{\pi - \varphi}{\pi} \cdot \frac{i_4 + \frac{1}{L} \cdot (V_{\text{grid}} + V_{\text{load}}) \cdot \frac{\varphi}{\pi} \cdot t_S + i_4}{2} \\
 I_{\text{load}} &= \frac{\pi - \varphi}{\pi} \cdot \frac{i_1 + \frac{\varphi}{\pi} \cdot i_4}{2} \\
 &+ \frac{t_S}{2 \cdot L} \cdot \left(\frac{\pi - \varphi}{\pi}\right)^2 \cdot (V_{\text{grid}} - V_{\text{load}}) + \left(\frac{\varphi}{\pi}\right)^2 \cdot (V_{\text{grid}} + V_{\text{load}})
\end{align*}
\]
Due to the fact that a further relation between these variables cannot be derived from this theory, the exact amplitude of the spikes, as shown in figure 2.4 with the red lines, cannot fully be predetermined with this model.

However, once the steady state of figure 2.7.b has been reached, \( i_1 \) can now be made equal to \(-i_4\), thus leading to equation 2.45.

\[
i_1 = -I_{\text{load}} + \frac{t_s}{2 \cdot L} \cdot \left( \left( \frac{\pi - \varphi}{\pi} \right)^2 \cdot (V_{\text{grid}} - V_{\text{load}}) + \left( \frac{\varphi}{\pi} \right)^2 \cdot (V_{\text{grid}} + V_{\text{load}}) \right) \tag{2.45}
\]

Once \( U_{\text{grid}} \), \( U_{\text{load}} \), \( t_s \) and \( I_{\text{load}} \) are known, now also \( \Delta I_A \) and \( \Delta I_B \), as shown in figure 2.7.b, can be calculated with the equations 2.46 and 2.47.

\[
\Delta I_A = \frac{\pi - \varphi}{\pi} \cdot t_s \cdot \frac{V_{\text{grid}} - V_{\text{load}}}{L} \tag{2.46}
\]

\[
\Delta I_B = \frac{\varphi}{\pi} \cdot t_s \cdot \frac{V_{\text{grid}} + V_{\text{load}}}{L} = 2 \cdot I_{\text{load}} \tag{2.47}
\]

Equation 2.47 can now be used to calculate the phase shift, which needs to be applied. The result is presented in equation 2.48.

\[
\varphi = \frac{2 \cdot \pi \cdot L \cdot I_{\text{load}}}{(V_{\text{grid}} + V_{\text{load}}) \cdot t_s} \tag{2.48}
\]

This equation can now be used for the calculations in the control unit of the phase shifted DAB.

### 2.5.5 Restrictions for the leakage inductance

The added leakage inductance should be as small as possible, in order to reduce costs, size and power loss. Also, the value of the leakage inductance, together with the switching frequency, will limit the maximum output power. Therefore, the effect of a serious decrease of the leakage inductance will be shown. This could happen, if an external coil with ferrite core would be set into saturation, thus drastically decreasing the inductance to the value of the same inductor with air core.

However, the coefficient for the \( \text{di/dt} \) would become so high, that serious damage to the whole circuit would be inevitable, as shown in equation 2.49.

\[
\frac{dI_L}{dt} = \frac{1}{L} \cdot U_L \tag{2.49}
\]

These high derivative values could seriously damage switching components and could cause HF noise, possibly affecting internal measurement and control signals or even external equipment.

Therefore, the main inductor should always have an adequate value. This value can be calculated by calculating the maximum delay time and the maximum voltage, at which the inductor will be used. First of all, a maximum \( \Delta I \) can be set, due to hardware restrictions. Together with the maximum delay time, a maximum value for the \( \text{di/dt} \) can
be calculated. The maximum voltage can follow from the maximum difference between the input and output peak level. Together, this will result in a minimum value for the leakage inductance.

### 2.5.6 Phasor diagram of circuit voltages and currents

In order to get an overview on the general working principle, the average values for various input voltages, the steady output voltage and the belonging various values for the average voltage over the inductor have been set into a phasor image, shown in figure 2.8. The simplified scheme of figure 2.6 can be used again. The only difference is the fact that the voltage source is now giving a 50 Hz sine wave output. Four different situations have been given, to show the result of varying load impedances.

As can be seen, the DAB requires a minimum capacitor size in order to be able to operate well. In the picture, the phasor $I_C$, added to $I_{load}$, causes the inductor current to rotate towards the imaginary axis. This way, the output voltage can be controlled, using a varying phase shift. This phase shift between the primary and secondary bridge will then influence the average voltage over the inductor, which enables a voltage difference between the grid voltage and the load voltage. It will also introduce a phase shift between the grid voltage and the load voltage. This is necessary, in order to maintain a voltage over the leakage inductor, necessary for the transfer of power through this inductor.

![Phasor diagram](image)

**Fig. 2.8. DAB voltage and currents for various loads and input voltages.**

Figure 2.8.b shows the phasor image for heavy loads. As can be seen, the influence of $I_C$ is relatively smaller, resulting in larger values for the average inductor voltage. This requires higher phase shifts again. In theory, it is still possible to boost the grid voltage, but this is becoming more and more difficult.

The effect of adding an inductive or capacitive component to the load is shown in figures 2.8.c and 2.8.d. As can be seen in figure 2.8.c, adding inductive loads can destabilize the control aspect of the DAB, and also can cause the control to turn around. Adding a capacitive load however will increase the effect of a phase shift, resulting in possible over- and under voltages.
2.5.7 Implementation of concrete design parameters into a DAB model

In order to see the effect of a practical implementation of a certain leakage inductance value, a further simplified basic scheme as in figure 2.9.a can be made, based on the given design parameters. As can be seen, the current, which needs to be switched, has to feed a resistive load. Due to the fact that the secondary side does not have a driving voltage, the di/dt in the inductor can reach high values every switching period, thus enabling a fast increase of the secondary voltage.

![DAB model](image)

Fig. 2.9. 25kHz DAB model. a) circuit, b) zoom in on half period, with RMS value of $U_p$ and $U_s$ for two inductor values, c) two full periods.

The blue line in figure 2.9.b and 2.9.c is the resulting output voltage, when a leakage inductance of 10uH and a desired output current of 16A at 100V are applied. As will be shown in chapter 4, this is a standard value for the leakage inductance of the main transformer of the current Mass GI design. The red line shows the output, when an inductor of 100uH is applied.

It is clear that the voltage on the secondary side can't reach the desired level, due to the slow di/dt. However, the inductor value also can't be further reduced, as was shown in paragraph 2.5. Therefore, it is expected that the output voltage undesirably will drop for high currents.

2.6 AC Switch layout

As has been shown in the previous paragraph, the implementation of DAB for AC grids also affects the choice of switches. In order to make a good choice for the switch implementation, first of all the layout possibilities will be shown. Subsequently, the used semiconductor components will be compared. Eventually, the final choice of layout and components will be explained.

2.6.1 Topology overview

The main task of the switches is to enable and disable a current path between the grid and the main transformer. As has been shown previously, this current is an AC current, therefore flowing into both directions. As a result, a single IGBT or MOSFET can’t be used, as this device will only switch a current flow in a single direction. Therefore, a combination of two components is needed. Also, diodes are needed to prevent a opposite current from flowing into the wrong device.

In general, several topologies can be used. Figure 2.10 shows five of them. As can be seen, circuits A and C, as well as topology B and D, are based on the same topology. Therefore, only the circuits A, B and E will be discussed.
High frequency galvanic insulated shore power connections

Looking at topology A in figure 2.10.a, it is clear that currents either flow through D1a and are switched by T1b or flow through D1b and then are switched by T1a. Realization will not be a problem, as there are many IGBTs and MOSFETs with built in anti-parallel diode available, thus enabling a compact PCB design. The voltage drop will be equal to the voltage drop over an IGBT or MOSFET and the voltage drop over a power diode. Gating circuits can be built up relatively easy. Since both IGBTs have a common emitter, only one gating circuit is needed per AC-switch.

Fig. 2.10. Switch topology layouts with IGBTs and MOSFETs.

In topology B, shown in figure 2.10.b, currents either flow through D2a and are switched by T2a or flow through D2b and then are switched by T1b. A practical implementation however would be more difficult, as this requires either external diodes, or IGBTs and MOSFETs with built in series diode. The voltage drop will be also equal to the voltage drop over an IGBT or MOSFET and the voltage drop over a power diode. Interesting about this topology is the fact that certain IGBTs are specifically designed with a reverse blocking capability [39], thus resulting in a lower total voltage drop. Due to the fact that the two IGBTs have separately connected emitters, also two gating circuits will have to be made to build up the correct $V_{GE}$ for each IGBT.

Topology E, in figure 2.10.e, finally, has the advantage of only one switch, thus decreasing cost for the needed gating circuit. As a disadvantage, this is the topology with the highest number of components. The AC current is rectified by power diodes D5a, D5b, D5c and D5d. These can be implemented as a single bridge unit, but also as individual components. The last way of construction however would demand a lot of space on heat sinks, as all components have to be attached to a heat sink, due to the high loss. The voltage drop over this AC switch will be equal to the voltage drop over the MOSFET, together with the voltage drop over two power diodes. Clearly, since only one MOSFET is used, also only one gating circuit is necessary.

Comparing all aspects, switch topology A and C are chosen because of the simplicity in gating networks and the low PCB surface, needed for the gating circuits, switching components and anti parallel diodes.

### 2.6.2 Comparison between MOSFETs and IGBTs

The bipolar transistor was the only "real" power transistor until the MOSFET came along in the 1970's. The bipolar transistor requires a high base current to turn on, has relatively slow turn-off characteristics (known as current tail), and is liable for thermal runaway due to a negative temperature coefficient. In addition, the lowest attainable on-state voltage or conduction loss is governed by the collector-emitter saturation voltage $V_{CE(SAT)}$.

The MOSFET, in comparance, is a device that is voltage- and not current-controlled. MOSFETs have a positive temperature coefficient and are therefore protected for a
possible thermal runaway. The on-state-resistance $R_{DS(on)}$ has no theoretical limit; hence on-state losses can be low. The MOSFET also has a body-drain diode, which is particularly useful in dealing with limited free wheeling currents.

All these advantages and the comparative elimination of the current tail soon meant that the MOSFET became the device of choice for power switch designs.

Then in the 1980s the IGBT came along. The IGBT is a cross between the bipolar and MOSFET transistors. The IGBT has the output switching and conduction characteristics of a bipolar power transistor but is voltage-controlled like a MOSFET. In general, this means it has the advantages of high-current handling capability of a bipolar transistor at a fixed voltage drop in the on-state, with the ease of control of a MOSFET.

The major problem of IGBT for operation at high frequency is the "current tailing". At turn-off the device current does not fall rapidly but a considerable portion of the current lingers or tails for a longer time. The co-existence of tail current and high collector-to-emitter voltage of IGBT cause high turn-off switching losses. This sets the upper limit on the switching frequency of an IGBT.

Early versions of the IGBT were known to latch up, but nowadays, this problem is mostly eliminated. Another potential problem with several IGBT types is the negative temperature co-efficient, which could lead to thermal runaway and makes the paralleling of devices hard to effectively achieve. This problem is now being addressed in the latest generations of IGBTs, based on "non-punch through" (NPT) technology.

Table 2.1. MOSFET versus IGBT in characteristics and applications

<table>
<thead>
<tr>
<th>MOSFET properties and applications</th>
<th>IGBT properties and applications</th>
</tr>
</thead>
<tbody>
<tr>
<td>• High frequency applications (&gt;200kHz)</td>
<td>• Low frequency applications (&lt;100kHz)</td>
</tr>
<tr>
<td>• Low-voltage applications (&lt;250V)</td>
<td>• High-voltage applications (&gt;1000V)</td>
</tr>
<tr>
<td>• &lt; 5kW output power</td>
<td>• &gt;5kW output power</td>
</tr>
<tr>
<td>• Low switching losses</td>
<td>• Higher switching losses</td>
</tr>
<tr>
<td>• Conduction losses linear with $I_{DS}$ due to $R_{DS(on)}$</td>
<td>• Conduction losses controlled by $V_{CEO}$</td>
</tr>
<tr>
<td>• Wide line or load variations</td>
<td>• Narrow or small line or load variations</td>
</tr>
<tr>
<td>• Switch mode power supplies (SMPS): Hard switching above 200kHz</td>
<td>• Motor control: short circuit/in-rush limit protection</td>
</tr>
<tr>
<td>• Switch mode power supplies (SMPS): ZVS below 1000 watts</td>
<td>• Uninterruptible power supply (UPS): Constant load.</td>
</tr>
<tr>
<td>• Battery charging</td>
<td>• Welding: High average current, ZVS circuitry</td>
</tr>
<tr>
<td></td>
<td>• Low-power lighting: Low frequency (&lt;100kHz)</td>
</tr>
</tbody>
</table>

When comparing the layouts, the MOSFET and IGBT structures look very similar, as can be seen in addendum E1. The basic difference is the addition of a p substrate beneath the n substrate. The IGBT technology is certainly chosen in situations, in need of breakdown voltages above 1200V, while the MOSFET is better applicable for device breakdown voltages below 250V.
Between 250 to 1200V, there are many technical papers available from manufacturers of these devices, some preferring MOSFETs, some IGBTs. However, choosing between IGBTs and MOSFETs is very application-specific and cost, size, performance and speed and thermal requirements should all be considered.

Due to technical constraints, high voltage MOSFETs will have a higher $R_{DS(on)}$, than the low voltage types. This can cause problems, for example in the case in which the heat sink temperature has increased due to maximum operating conditions. When a high load current is applied, the conduction losses of the MOSFET would be higher than the switching losses of the IGBT, thus leaving the IGBT in a better position, as it can operate longer.

When the average power consumption during an entire day is considered, the maximum output power is used only 10-20% of the time and the average output power is only 25% of the rated output power. Under conditions of low average load, a MOSFET might provide a higher efficiency. At high output currents, the GI would be bothered with a high voltage drop over the MOSFETs, and therefore also a high loss within the GI. As a result, the heat development at high load will increase and will therefore demand more cooling, leading to bigger fans or even a bigger case for the GI.

In this specific application, therefore the IGBT appears to be the device of choice, leading to the switch design in figure 2.10.a as the final AC switch topology.

2.7 Snubber design
Snubber circuits are usually used to protect fast switching IGBTs from turn-on and turn-off voltage transients [28, 41]. However, in the MASS GI there’s also another reason to use snubbers. Since a short dead time is implemented between deactivation and activation of the complementary switches in the DAB, the main transformer will also be disconnected during this time.

Due to the fact that no freewheeling diodes are present, as in DC-DC converters, this will result in a discontinuation in the transformer current path, thus creating a voltage spike. Snubbers provide a bypass current path to ensure a safe and successful commutation of the switching devices. Snubbers are available in various configurations and a clear understanding of their operation is necessary to make the appropriate selection. Therefore, several types of snubber circuits will now be discussed.

2.7.1 Non-dissipative snubbers
Non-dissipative snubbers, also called regenerative snubbers, are the class of snubbers which recycle the energy they collect [34, 38]. Recycling of energy can be done in three ways:

- Transfer of energy to the input side of the converter
- Transfer of energy to the output side of the converter
- Preparation of energy for the next cycle.

Even though this is energy wise the most efficient way to reduce spikes, the practical implementation would mean introduction of many extra components for storage and transportation of energy. Some designs even implement a transformer to transfer the energy, which makes the circuit expensive and large.
Due to the increased complexity, this kind of snubbers will therefore not further be investigated, as long as the power, which needs to be snubbed, will not get the overhand in the circuit losses.

2.7.2 Dissipative snubbers
The standard implementation of IGBT snubbers is shown in figure 2.11.a. As can be seen, it only consists of a resistor, a capacitor and a diode. The capacitor will suppress any spikes over the switch.

This will not work for the problems, as shown in the previous paragraph, due to the fact that AC signals will have to be snubbed. An opposite current will therefore lead to a direct current through the resistor, thus leading to increased losses.

Therefore, another way has to be found to prevent voltage spikes on the input and output of the main transformer. First of all, it is possible to convert the DC snubber design to a AC design. A schematic drawing is shown in figure 2.11.b. One of the problems however is not solved yet: energy should flow in and out the capacitor each time the IGBT switches on and off.

As a result, the schematics of figure 2.11.c can provide a solution. The capacitor is set behind the diode bridge, which reduces switching losses caused by the snubber capacitor, as it will only be filled on peak values and its voltage does not have to switch between high and low values or change polarity. The used diode bridge should be very fast, in order to obtain a quick snubber response time.

Some intelligence can now be added to the dissipative element, in order to only dissipate spikes in the snubber dissipation circuit. Implementation of only adding a resistor as dissipative element will also lead to losses, when a normal switching voltage without spikes is applied.

Looking at the DAB design, this would mean that for each bridge leg one snubber would be necessary. However, the spikes are mainly caused by the leakage in the main transformer. Therefore, the snubber will now not be placed over the individual AC switches, but over the primary and secondary windings of the main transformer, as shown in figure 2.11.d. This way, four AC switches and a transformer coil are protected while using only a single snubber.
2.8 Conclusion

The Dual Active Bridge converter, as presented in this chapter, has a number of advantages, but also various drawbacks, when compared to other switch mode converter topologies.

The main advantage is the fact that no storage device in the form of a DC bus is necessary, when the DAB converter is used for AC-AC conversion. The disadvantage of this switching scheme is the fact that in a single phase AC-AC DAB converter therefore the difference between the input and the output AC grid signal cannot be controlled. In other words: what comes in, will also come out again.

A solution can be found in the phase shift controlled DAB converter. This way, the ratio between the effective input and the effective output voltage can be controlled by controlling the phase shift between the HF gating signals for the primary and secondary H-bridge. Due to the current design constraints however, the added leakage inductance for the main transformer should have to become relatively high. Due to this high inductance value, it is expected that a phase shift controlled DAB will not be able to handle the full power range at the desired range for the ratio between input and output voltage.

Therefore, the inductance of the main transformer should be as small as possible, thus enabling a compact HF AC-AC converter, capable of handling high output currents. A phase shift controlled DAB converter however is expected not to meet the required standards.
3 Design evaluation of AC grid HF DAB Galvanic Insulator

Now the basic principle of a Dual Active Bridge High Frequency AC Galvanic Insulator has been shown, it is time for a practical implementation of the theory into a working setup. The design specifications for this device can be found in addendum CI. Since the product is already on the market, the actual version has several safety and measurement circuits. Not all of these circuits will be explained, since one of the main goals of this final thesis is the realization of a DAB structure, not the design of a complete product.

In order to provide a good overview, first an overview of all the final schematics of the non scalable GI will be presented. Following, the function, layout and realization of several components and sub circuits will be explained. In this top-down approach the specific choice in component values will also be explained.

3.1 General sub circuit overview

Since a good overview on the total schematics is necessary to see the connections between the different sub circuits, a general functional model of the MASS GI is presented in figure 3.1. This scheme does not only contain the ideal DAB circuit, but also the auxiliary power supply, filters, voltage and current measurements, control block and gating circuitry. In the following paragraphs, the function and layout of the most important sub circuits and components will now further be explained. All other, supplementary circuits are further explained in addendum A4. A detailed schematic overview of all schematics can be found in addendum A1.

![Fig. 3.1. General sub circuit overview of the Mass GI based on DAB technology.](image)

3.2 IGBT comparison and choice

A very important component in the circuitry is the switch, consisting of two series connected IGBTs with an integrated anti-parallel diode. In order to be able to make the right choice, addendum E1, "Implementation of IGBTs as switches", shows the general specifications of the IGBTs.
This addendum shows that switching losses and conduction losses mostly determine the efficiency of a specific IGBT. Therefore, addendum E3 displays the result of a component search for 20A and 40A 600V IGBTs.

From the overview in addendum E3, three IGBT types have been chosen for further tests. The choice for these components was not only lead by the specific characteristic values, but also by market availability of the component. The measurement results of this component comparison are presented in addendum E4, “IGBT test definition and results”. Eventually, the already implemented HGTG20N60B3D of Fairchild appeared to be the best option. More information on this IGBT can be found in addenda C5 and C6.

During my final thesis, also a very interesting problem rose: the deliverance of fake IGBTs, instead of the genuine product. These batches of IGBTs mostly came from undefined Asian suppliers, delivered via brokers. After a number of breakdowns of devices, using these IGBTs, the test was also used for checking the genuinity of the product. Addendum E5 therefore contains a number of measurements on batches of fake IGBTs.

### 3.3 Gate driver circuit

As previously shown, the IGBT has a MOS gate structure. Therefore it is necessary to provide not only a gate voltage, but also a good gate current path to charge and discharge this gate when switching. The standard solution for gate drives for IGBTs nowadays is mostly done with pure resistive control, meaning that a gate resistance is added to control the turn on and turn off characteristics. However, this is not the only important factor: also the applied gate voltage $V_{GE}$ has a great influence. A general overview of these factors on several IGBT characteristics is mentioned in table 3-1. The given relations are further explained in addendum A3, “Calculations for the gating and dead time circuits”.

<table>
<thead>
<tr>
<th>Main characteristics</th>
<th>increase of $V_{GE}$+</th>
<th>increase of $V_{GE}$-</th>
<th>increase of $R_G$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{CE(sat)}$</td>
<td>decrease</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>$T_{on, E_on}$</td>
<td>decrease</td>
<td>-</td>
<td>increase</td>
</tr>
<tr>
<td>$T_{off, E_off}$</td>
<td></td>
<td>decrease</td>
<td>increase</td>
</tr>
<tr>
<td>Turn-on surge voltage</td>
<td>increase</td>
<td>-</td>
<td>decrease</td>
</tr>
<tr>
<td>Turn off surge voltage</td>
<td></td>
<td>increase</td>
<td>decrease</td>
</tr>
<tr>
<td>$dV_{CE}/dt$ malfunction</td>
<td>increase</td>
<td>decrease</td>
<td>decrease</td>
</tr>
<tr>
<td>Current limit value</td>
<td>increase</td>
<td>-</td>
<td>decrease</td>
</tr>
<tr>
<td>Short circuit withstand capability</td>
<td>decrease</td>
<td>-</td>
<td>increase</td>
</tr>
<tr>
<td>Radiational EMI noise</td>
<td>increase</td>
<td>-</td>
<td>decrease</td>
</tr>
</tbody>
</table>

The end result of the gate driver calculation is also given in addendum A3. As can be seen, the gating circuit consists of a $R_{on}$ and a $R_{off}$, of which the first one is equipped with an anti-parallel connected fast diode.

### 3.4 Additional dead time control

Dependable on the implemented IGBTs and gating circuit, a short period can still be discovered, in which both paths of each H-Bridge of the DAB are conducting. This implies that the IGBTs, which were off, are already starting to conduct, when the IGBTs,
which are turning off, are still conducting due to tail currents. This is called shoot-through and is present in the current Mass GI design.

In order to reduce the losses, caused by this temporary short circuit, there are two options: further increase of the turn on resistor in all gating circuits or implementation of a dead time in the gating signal.

The first option, increasing all turn on resistors in the gating circuits, has several major drawbacks. First of all, there is a practical reason: variation of the turn off resistors in the 8 gating circuits is only possible by replacing all resistors. A bigger problem however is the increase of switching losses, when increasing the turn on resistors, as shown before. This is caused by the fact that the transition between the off and the on state of the IGBT is slowed down, leading to a longer time in which a current can flow through the IGBT, while \( V_{CE} \) is not at a minimum level yet.

The second option is based on the fact that the 25kHz square wave gating signal can also be altered and can contain a dead time delay. The dead time can then be controlled with a single circuit for all gating circuits and therefore will be easily adjustable. Since the optimal amount of dead time will vary for different IGBTs and maybe even for different loads, this last option has been chosen.

In order to implement an additional dead time, a safety circuit, already present in the Mass GI is used. Normally, this circuit pulls down the 12V block signals, going to the gate driver, for example in case of over current or in order to enable Master-slave applications. Now, this circuit will be used to add a dead time to the gating signals every half period. Schematics, calculations and simulations regarding the dead time control circuit can be found in addendum A3. The effect of this circuit on the efficiency of the total circuitry of the Mass GI can be found in addendum B2, "Measurement results non scalable 230V DAB converter". As can be seen, a decrease in power loss up to 12W can be reached, using this additional circuit.

### 3.5 Main transformer design

The ideal model of a transformer, shown in figure 3.2.a, will not satisfy in this application. Therefore, the T-model is chosen. The T-model of a transformer, convenient because of its close connection with the physical structure, is expressed in terms of the permeabilities associated with magnetic paths. Until now, in the theoretical approaches of the Dual Active Bridge converter, always a two-winding transformer has always been implemented, as shown in figure 3.2.b. In the practical implementation however, a three winding transformer has been chosen, as shown in figure 3.2.c. This model also shows the internal losses, which will be discussed later on.

![Fig. 3.2. Main transformer models. a) Ideal model, b) addition of leakage inductance, c) addition of other parasitic components.](image)
Several predictions of the model important for practical designs are experimentally confirmed: the leakage inductance of one winding of a three-winding transformer is increased by closer coupling between the other two windings; the leakage inductances are essentially independent of air gap, although the coupling coefficient decreases with increasing air gap length.

To investigate the effect of several transformer parameters to the efficiency of the total transformer, influences of both material and layout properties are presented in addendum Fi, “Main transformer design for the 230V DAB converter”. Also a number of literature sources have been used for this investigation [10, 18, 24, 35].

Once these parameters are set, two very important calculations of the main transformer can be made: the number of windings, in relation to the core size and the winding thickness, in relation to fill factor of the winding package. Once the core and the copper specifications are chosen, further analysis of the DC and AC resistance of the coil and the core losses can approve or disapprove a certain design.

The square-wave voltage at the input of the transformer causes a triangular shaped magnetizing current $I_m$, which is almost independent of the secondary current. The magnetizing current is approximately proportional to the magnetic flux $\Phi$ i.e. to the magnetic flux density $B$. The input voltage $V_1$ determines the magnetic flux in the transformer core corresponding to Faraday’s Law:

$$\dot{U}_{\text{prim}} = N_{\text{prim}} \frac{d(\Phi)}{dt}$$  \hspace{1cm} (3.1)

The magnetic flux density $B$ is determined by the core material and shape. A good estimation of the practical usage of Faraday’s Law can be given by:

$$\Delta B = \frac{\dot{U}_{\text{prim}} \cdot T_s / 2}{N_{\text{prim}} \cdot A_{\text{core}}}$$  \hspace{1cm} (3.2)

With:

- $\Delta B = 2 \cdot B_{\text{max}} = \text{max imum flux density swing}$
- $\dot{U}_{\text{prim}} = \text{peak input voltage}$
- $T_s = \text{switching time}$
- $N_{\text{prim}} = \text{primary number of windings}$
- $A_{\text{core}} = \text{core surface}$

For the main transformer of the Mass GI the material N87 has been used. One of the characteristics of this material is the high performance factor, as shown in addendum C7, given by N87 manufacturer EPCOS, formerly Siemens.

As can be seen in addendum Fi, for the 230V DAB a transformer with a 19:20 turn ratio, built around two E65 core halves is the best option. The secondary coil is split up into two coils, thus increasing the coupling between the primary and secondary winding. The copper foil diameter is set at $30\text{mm} \times 0.15\text{mm}$, well below the thickness at which skin effect could influence the AC resistance. Winding instructions can be found in addendum F5.
As is shown in addendum Fr, the average DC resistance of the E65 transformer design at 80°C is calculated at 26.0Ω and the average AC resistance at this temperature is calculated at 40.2Ω. In these calculations, the temperature difference between the inner layers and the outside of the core is not taken into account.

### 3.6 Dissipative snubber design

As has been shown in the previous paragraphs, the snubber should limit the voltage over the IGBTs and the main transformer. In order to achieve this goal and to be able to convert enough power, a snubber circuit has been designed, which uses a MOSFET for dissipation of the snubber energy. The Mass GI contains two snubbers: one over each winding of the main transformer. Figure 3.3 shows the schematic of the MOSFET snubber, together with the test circuit: a series connection of an inductor and resistor, which is being switched off for 1% of the time, while being fed by a 230 V AC grid the other 99%. These values are calculated by the fact that the average switching period is 40μs, while the required minimum dead time is 200ns per half period, thus leading to a maximum freewheeling time of the main transformer of 1%. Combined with the maximum output power of 3500W, this will result in a maximum average snubber power of 35W.

In the snubber circuit, the main transformer voltage first is being rectified by four hyperfast diodes. The output voltage is fed to a capacitor and to the actual dissipation circuit. Since the snubber only has to start dissipating power at high voltages, the circuit only activates the gate of the MOSFET at voltages above 400 V. At this voltage, the two 200V series connected zener diodes D1 and D2 start conducting. This will result in a rising voltage over D3. As soon as the threshold value \( V_t \) of the gate voltage, \( V_{GD} \), is reached, the MOSFET will start conducting, thus dissipating power.

#### 3.6.1 Limitation of maximum snubber power

At this point, the current through the MOSFET theoretically can keep on rising. However, in order to limit the power dissipation in this device, resistor R5 is present. When the emitter current rises, also the voltage over R5 rises. Since the gate voltage is limited with D3, a 18V zener diode and VGD always should reach \( V_t \), the voltage over this resistor cannot exceed 14V. Since the average voltage over the MOSFET is equal to 400V and its maximum thermal power dissipation is 60W, the maximum emitter current can now be calculated with equation 3.3.

\[
I_{e(max)} = \frac{P_{max}}{U_{snubber}} = \frac{60}{400} = 0.15A
\]
At this current, the voltage over $R_5$ should be $14\, \text{V}$, therefore the minimum value for $R_5$ is $93.3\, \Omega$. For practical implementation, $R_5$ is assigned the value of $100\, \Omega$.

Simulation results of the snubber circuit of figure 3.3 are shown in figure 3.4. As shown, the voltage peaks, caused by the disconnection of the inductive load, are limited. Also the voltage over the capacitor is limited, but is not limited at $404\, \text{V}$, which would be the result when the limiting resistor would not have been implemented.

![Simulation results of power limited snubber design](image)

**Fig. 3.4. Simulation results of power limited snubber design**

### 3.7 Temperature, current and voltage control

As can be seen, a general control block is also shown in the overview of sub circuits. The following processes should take place in this control unit:

- Startup control
- Master / Slave behavior and communication:
  - Error signaling from and to connected master and one or more slave units
  - Startup signal control between master and one or more slaves.
- Over current protection
- Over voltage protection
- Over temperature protection and fan control
- User interface control:
  - Output power indication to user
  - Error signaling to user in case of over current, over voltage or over temperature.

At this moment, a PIC, type $18F2521$-SO, is used for these tasks. Since the software still belongs to the external designer of the Mass GI and therefore only can be read in assembly in the PIC. The current software is not complete and needs to be updated with a number of tasks, for example the voltage control loop.

Therefore, addendum D1 contains a new design of the full control scheme, in flow charts for both the main routine as all sub routines. Using this control scheme, one of the software engineers will soon be writing the C-program. This code could then not only be applied in the current PIC, but also in new control units, such as DSPs. This is an interesting issue, as future developments, such as the ideas presented in this final thesis,
High frequency galvanic insulated shore power connections will need more processor power and peripherals. Regarding the possibilities of the current PIC, these demands would automatically call for a new type of processing unit.

3.8 Practical implementation and measurements

The implementation of most of the presented circuits was very easy, as they were already present in the current design of the Mass GI, as given in addendum A1. Some circuits have been redesigned, such as the snubber circuits. Other circuits have been added, such as the dead time control circuit. In order to see the influence of these circuit variations, numerous simulations and measurements have been done. Simulations mostly are presented in the addenda, in which they were presented. Measurement results are given in addendum B2, "Measurement results non scalable 230V DAB design". A number of the proposed improvements are needed to reach reinforced isolation between the primary and secondary circuit, as shown in addendum A2, A6 and B7.

Very interesting are the results of the measurements with a variable dead time applied to the gating circuit. As can be seen, these measurements show that the currently applied dead time is not high enough, as a certain amount of shoot-through is still present. A dead time of 500ns would simply solve these problems. The dead time circuit is presented in addendum A3.

Also interesting are the results on the IGBT measurements, as shown in addendum E4. As can be seen, the IHW40N60T can't measure up with the chosen HGTG20N60B3D. Eventually, also the measurements on the fake batches of this IGBT show that correct component modeling can save money, if only for the fact that fake batches now can easily be recognized.

Eventually, the efficiency measurements show that the device will never be able to reach the high efficiency levels at high outputs of a 50Hz transformer. Due to the fact that the 50Hz device has a relatively high magnetizing current, at very low loads, for example when the ship is on the dock in winter times, the Mass GI can reach a higher efficiency.

3.8.1 Electromagnetic Compatibility

One of the negative aspects of switching AC grid voltages at 25kHz is the generation of electromagnetic pulses, as the operation of switching devices can generate both of $\frac{dv}{dt}$ and $\frac{di}{dt}$. It is clarified that these phenomena can cause of the Electro Magnetic Interference, EMI, emission which can affect systems standing near the GI.

Therefore standards have been made regarding the electromagnetic compatibility of the equipment [3, 19, 20]. Examples of these standards are the IEC. These standards not only set restrictions to the generation of electromagnetic waves, but also demand a certain robustness of the equipment when electromagnetic pulses are injected to in- or outputs of the system.

3.9 Conclusion

Looking at the current state of the non scalable 230V DAB design, it is clear that this product is applicable in much more areas than the current niche, in which it is operating. The design is robust; a number of control circuits will ensure that the device will stop in time. Measurement results show that the efficiency is lower than the efficiency of a 50Hz transformer, but at the same time, the Mass GI is smaller in size and weight.
During this project, already a number of improvements to the product have been made. Examples are the adjusted winding instructions for the main transformer, as shown in addendum F4, after the discovery of malicious transformers, shown in addendum F3, and the adjusted snubber design, as shown in addendum A5. Table 3-2 contains an overview on the most important current and possibilities for future improvements.

Table 3-2. Overview of current properties, achieved improvements and suggestions for future improvements.

<table>
<thead>
<tr>
<th>Circuit / component</th>
<th>Present situation</th>
<th>Achieved improvements</th>
<th>Suggestions for future improvements</th>
</tr>
</thead>
<tbody>
<tr>
<td>Switch design</td>
<td>Series connected IGBTs with internal anti-parallel diode.</td>
<td>Comparison of various switch topologies. No improvement implemented.</td>
<td>Replacement of IGBTs by series connected MOSFETs with internal anti-parallel diode, newest type COOLMOS</td>
</tr>
<tr>
<td>Choice of IGBT</td>
<td>Now used: HTG20N60B3D</td>
<td>IGBT evaluation gave second best replacement: IRG4B40UD. Evaluation also valid for recognizing fake batches of current IGBT type.</td>
<td>Continuous search for better IGBT types.</td>
</tr>
<tr>
<td>Snubber circuits</td>
<td>Unlimited snubber circuit</td>
<td>Limitation of snubber power. Circuit change implemented in current production line.</td>
<td>Investigation for regenerative snubber design for further increase of efficiency</td>
</tr>
<tr>
<td>Dead time</td>
<td>Determined by ( R_{on_a} ) and ( R_{off} )</td>
<td>Addition of external adjustable dead time circuit, leading to an increase of efficiency.</td>
<td>Implementation of dead time circuit on PCB or PWM signal generation by controllable device</td>
</tr>
<tr>
<td>Isolation between IGBTs/MOSFETs and heat sinks</td>
<td>Al. ox. substrate</td>
<td>Investigation of effect of silicone tube over all switching devices in order to meet minimum creeping distance requirements</td>
<td>Further investigation of other possibilities to meet creeping distance requirements.</td>
</tr>
<tr>
<td>Main transformer design</td>
<td>Winding instructions for 19:20 E65 foil transformer</td>
<td>Adjustment of winding instructions after transformer fall out.</td>
<td>-</td>
</tr>
<tr>
<td>Voltage / current measurement circuits</td>
<td>Input voltage measurement, dual output current measurement</td>
<td>Suggestion for new current transformers, in order to meet creeping distance requirements.</td>
<td>Implementation of over voltage protection in controller software.</td>
</tr>
<tr>
<td>12V power supply</td>
<td>Standard application of Switch Mode power supply</td>
<td>Suggestions for further surge protection of power supply circuit</td>
<td>Implementation of suggestions in current design</td>
</tr>
<tr>
<td>Input voltage and output current control</td>
<td>Control implemented in PIC. Software changes are difficult.</td>
<td>New control scheme in block diagram.</td>
<td>Conversion of new block diagram into software for GI</td>
</tr>
</tbody>
</table>

For the near future, a number of problems however should still be tackled. First of all, the mentioned new control scheme should be implemented as soon as possible, in order to protect the device from over voltages. Further more, the 12V power supply should be protected from surges and tested again with surges of 1kV and 2kV, as shown in addendum B2.
Eventually, also a number of changes have to be made in order to pass the tests from Kema, as shown in addendum A6. A number of these changes are mentioned in table 3-2. Once these changes have been made, the Mass GI can also be sold as an official certified device.
4 Push Pull technology for 120V HF Galvanic Insulators

In order to make the non-controllable Mass GI also applicable for the American market, a 120V-120V version had to be developed. In order to save costs, the 120V GI should be based upon the current 230V schematics, while meeting the design specifications in addendum C1.

Generally, preparing the GI for the American market means that the average voltage level will decrease to 120V, half of the previous standard value. However, the maximum average current will almost double to 30A. As a result, conduction losses might increase. Therefore, the topology, which will be used, has to be changed, as power loss has to be reduced to a minimum level.

In this chapter, first of all, the electrical characteristics will be specified. Subsequently, the choice of topology for the 120V GI will be explained. Then, the effect of decreasing voltage levels and increasing current values on several components will be examined. Therefore, the two parts, expecting to cause the highest losses, will be zoomed in upon: the AC switches and the main transformer.

Also the option of a switchable GI, which handles both 120V as well as 230V at the grid side and the load side should be taken into account.

4.1 Implementation of DAB technology at 120V input voltage

Using the current Dual Active Bridge schematics of the 230V Mass GI for the 120V version has several drawbacks. First of all, the main transformer should be changed, due to the applied higher currents and lower voltages. This will also mean that the AC switches have to be able to handle 30A. This implies that gating circuits and the snubbers also would have to change.

Secondly, in the 230V version, the difference in the number of primary and secondary windings now compensates the voltage drop over the primary and secondary IGBTs. Even though the average input voltage decreases to half of the original value, the absolute value of the voltage drop over the IGBTs will stay the same. This is obvious, as this value is dependant on the number of implemented switches, which is still the same as in the 230V DAB. Since the losses in the current 230V DAB design are concentrated in the IGBTs, these losses will therefore increase furthermore when applying a lower input voltage and higher currents.

Combining the presented facts, it is clear that a 120V DAB design would have a relatively higher voltage drop over the switches, where as these switches also need to be redesigned, due to the doubled maximum nominal current. Therefore, the currently available Dual Active Bridge design is not the best option for HF galvanic insulation within a 120V AC grid.

4.2 Transformation of the DAB circuit into a Push-Pull circuit

One of the challenges during this final thesis was to redesign the current 230V Mass GI for 120V grids, without many major changes to the current circuit. Fortunately, in the original design, already a number of components had been designed in such a way, that implementation of a new switching scheme was possible, without having to make a lot of changes to the PCB.
The new switching principle, which had to be implemented, was the Push-Pull circuit, of which the basics are shown in figure 4.1.

As can be seen, the physical layout of the H-Bridge has been used to implement a push-pull circuit around a new main transformer, TI. New about this transformer is the division of the primary set of windings into 2 sets of 9 windings each. Instead of using all windings continuously, each half of a switching period only one set of primary and one set of secondary windings is used.

The basic principle however is still the same as in the DAB topology: the incoming grid voltage is switched at a frequency of 25kHz by switches S1-S8. Due to the fact that the main transformer sees a frequency of 25kHz instead of 50Hz, the size of this magnetic component can drastically be reduced. A further explanation of the working principle will be given in the following paragraph.

As a result it is clear that the only major drawback is the fact that the main transformer is used less optimally, due to the presence of 2 coils primary and 2 coils secondary. As one of the many advantages, the fact can be mentioned, that the previously used IGBTs can still be used, due to the fact that the main current is divided between two parallel connected AC switches. Further more, the ability of using the same PCB as the 230V DAB design, as will be shown later, will provide many advantages.

4.2.1 Switching schemes of the Push-Pull converter
To clarify the working principle, figure 4.2 contains the possible switching schemes. Image 4.2.a shows a situation where S1 and S4 are conducting. A positive current is flowing through the primary winding set Pr of TI. This induces a positive magnetic field, which again not only sets a voltage on the secondary windings, but also on the windings of P2. As a result, the average voltage over the Snubber has the double value of the input voltage. On the secondary side, the current is transferred to the load and the ripple capacitor through S5 and S8.

Image 4.2.b shows a situation where S2 and S3 are conducting. A negative current is flowing through the primary winding set P2 of TI. This induces a negative magnetic field, which again not only sets a voltage on the secondary windings, but also on the windings of Pr. Again, the average voltage over the snubber has the double value of the input voltage. On the secondary side, the current is transferred to the load and the ripple capacitor through S6 and S7.

The transformation between these states can be done in 2 ways. The first, most logical way is turning off all switches for a short moment, as shown in figure 4.2.c. At this
moment, the tail currents of the IGBTs, which are turning off, as well as the transient caused by the leakage inductance of the windings will be transferred to the snubber circuits. During the turn off time, the voltage over the windings can decrease and enable a softer switch, resulting in lower switching losses.

Another possible way of switching between states is turning on all switches, as shown in figure 4.2.d. Where it was forbidden in the DAB to turn on all switches, here it can't destroy the converter. The only thing, which will happen is a hard switch of the switch, which has to be turned on, as the voltage level it has to switch down is two times the input voltage. At this moment, a small current might flow through this switch, but soon the magnetic fields of both primary windings will cancel each other out. The only current, flowing now into the primary windings, is the current, caused by the internal resistance.

Since the last mentioned pattern might lead to extra losses, turning off all switches before turning on two new switches is the preferred way of commutating between switching states.

4.2.2 Voltage stress over switches and snubbers

An interesting fact can be seen when the expected main transformer voltage pattern is drawn, as shown in figure 4.3. Due to the lower grid voltage, one might also expect a lower voltage stress over the switches and snubbers. However, due to the choice for the push-pull configuration, the voltage over these devices will even increase, compared to the 230V DAB topology.

Even though every time only one of the primary coils is activated, the changing magnetic flux will also induce a voltage on the passive primary coil. The amplitude of this voltage however is opposite to the 120V grid voltage, due to the winding polarity. As a result, both the non-activated switches as well as the primary snubber are connected on one side to $V_{\text{grid}}$ and on the other side to $-V_{\text{grid}}$. As a result, the voltage over these switches is equal to $2 \times V_{\text{grid}}$, thus 240V RMS, 340V peak.

For the secondary side, the same circuit is present, causing the same or even slightly higher voltage levels over the secondary switches and snubber as in the previously mentioned components in the 230V DAB topology.
A positive aspect of the snubbers in the push pull converter is the fact that the leakage energy, still present in the coil, which is turned off, can be removed from the inductor in the full 20μs, and not in the much shorter dead time, which is the case in the DAB design.

![Switching pattern with belonging coil voltages and currents for the push pull design](image)

### 4.3 Switch design

One of the most important parts in the current schematics are the AC switches, consisting of a series connection of 2 mirrored IGBT-diode pairs. In the 230V DAB topology, for these IGBT-diode pairs the Fairchild HGTG20N60B3D is used, as presented in addendum C5. This 600V-type can handle high peak currents and has a maximum continuous collector current of 20A at 110°C.

As has been shown in the previous paragraph, the maximum value for $V_{CE}$ can't be reduced, due to the chosen topology in which the switches will still endure the same voltage levels. Furthermore, since 600V is a standard value for the maximum value of $V_{CE}$ of IGBTs, lowering this value will not severely decrease costs. Also, since the voltage drop over an IGBT is dependant on the typical value for $V_{CE}$, which is not dependant of $V_{CE,max}$, a reduction of $V_{CE,max}$ will not increase efficiency.

In order to support the increased current level, there are several options, again also dependant on the choice of topology. In general, two paths can be chosen:

- Insertion of an IGBT with a higher current rating.
- Parallel connection of IGBTs with a lower current rating.

The first solution would be the simplest way to support the increased current demands. However, IGBTs with higher collector current ratings often also have higher values for $V_{CE}$, which will increase the conduction losses in the switches. Furthermore, the switching losses also increase drastically, as these high current types often are designed for lower switching frequencies. Thermal issues might also show up, due to the fact that the new IGBT will have to switch double currents and therefore also will dissipate more power, which needs to be transferred to the heat sink.

Parallel connection of IGBTs will also add a number of possible technical problems, which have to be taken into account. These issues are presented in addendum E2, “Parallel connection of IGBTs”. Besides these issues, there is also the drawback of higher
component numbers. As a drawback, this might not only increase costs, but also decrease efficiency, as lower losses per IGBT might be overshadowed by the double number of components, where these losses will occur.

Very important however also is the fact that the new design should be multifunctional. As mentioned before, a converter, switchable between 120V and 230V grid voltage would be very interesting. Therefore a switch design of parallel connected IGBT switches, presented in the previous chapter, is chosen.

### 4.4 Main Transformer design

Since the main transformer will also have to handle the same power flow, but at a higher current and lower voltage value, the thermal management will have to be thought through. Various options can now be chosen to adjust the 230V DAB main transformer design into a 120V Push-Pull main transformer design. Two things are important: the design of the windings and the design of the core. Both designs are interconnected and therefore will need to be calculated simultaneously. Eventually, also the possibility of implementation of a decent voltage boost needs to be investigated.

In order to reduce the losses in the transformer, a lower number of windings can be chosen. However, a reduction in the number of turns $N_1$ will cause an increase in $\Delta B$ and a quadratic increase of hysteresis losses. Logically, also the surface of the core $A_{core}$ is important.

![Table](Image)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Single E65 core</th>
<th>Dual E65 core</th>
<th>Single E80 core</th>
<th>2 separate E65 cores</th>
<th>2 separate E55 cores</th>
<th>Single E71 core</th>
</tr>
</thead>
<tbody>
<tr>
<td>Core diameter</td>
<td>20x27 mm$^2$</td>
<td>20x54 mm$^2$</td>
<td>20x20 mm$^2$</td>
<td>2x separate 20x27 mm$^2$</td>
<td>2x separate 17x25 mm$^2$</td>
<td>22x32 mm$^2$</td>
</tr>
<tr>
<td>Core volume</td>
<td>79.4 cm$^3$</td>
<td>138.8 cm$^3$</td>
<td>72.1 cm$^3$</td>
<td>2x separate 79.4 cm$^3$</td>
<td>2x separate 51.4 cm$^3$</td>
<td>102 cm$^3$</td>
</tr>
<tr>
<td>Total transformer weight</td>
<td>0.70 kg</td>
<td>1.25 kg</td>
<td>1.10 kg</td>
<td>2x separate 0.70 kg</td>
<td>unknown</td>
<td>unknown</td>
</tr>
<tr>
<td>Winding ratio</td>
<td>9:10</td>
<td>6:6</td>
<td>14:15</td>
<td>2x separate 9:10</td>
<td>14:15</td>
<td>8:8</td>
</tr>
<tr>
<td>Foil diameter</td>
<td>30x0.15 mm$^2$</td>
<td>30x0.23 mm$^2$</td>
<td>40x0.20 mm$^2$</td>
<td>30x0.15 mm$^2$</td>
<td>25x0.12 mm$^2$</td>
<td>30x0.18 mm$^2$</td>
</tr>
<tr>
<td>Average winding length</td>
<td>149mm</td>
<td>203mm</td>
<td>172mm</td>
<td>149mm</td>
<td>123mm</td>
<td>163mm</td>
</tr>
<tr>
<td>DC Resistance</td>
<td>12.6mΩ</td>
<td>6.73mΩ</td>
<td>12.5mΩ</td>
<td>12.6mΩ/2 = 6.30mΩ</td>
<td>23.8mΩ/2 = 11.9mΩ</td>
<td>9.7mΩ</td>
</tr>
<tr>
<td>coil losses at 30AZc</td>
<td>11.4W</td>
<td>6.06W</td>
<td>11.3W</td>
<td>5.7W</td>
<td>10.7W</td>
<td>8.8W</td>
</tr>
<tr>
<td>Voltage boost @ 120Vin</td>
<td>13.3V</td>
<td>0V</td>
<td>8.6V</td>
<td>13.3V</td>
<td>8.6V</td>
<td>0V</td>
</tr>
</tbody>
</table>

Since the main calculations already have been explained in the previous chapter, a selection of different magnetic topologies is discussed in addendum F2. For all
High frequency galvanic insulated shore power connections
topologies, the core size, number of windings, winding diameter and the transformer losses are discussed.

The results are shown in table 4.1. In the situations, in which two separate cores are used, all corresponding coils are connected in parallel, both primary and secondary.

Literature study has been done on the possibility of implementing half windings, but the increased leakage inductance, caused by these windings, can gain too high values for this application [11, 45].

Combining these results with the test results in addendum B3, it is clear that the single E6S, single E80 and two separate E55 core designs will not work, as the dissipated heat inside the coils is too high.

The design, which is using two parallel connected E6S cores, has the ability of a proper voltage boost, but has a high chance of failure, as the division of the total current over the transformers is not controlled. Since transformers might show variations in internal impedance, when coming from the factory, this might lead to a possible overheating of one of them.

As a result of this transformer comparison, the dual core E6S has been used for implementation and various tests at 120V, as shown in addendum B3. Winding instructions can be found in addendum F5.

4.5 Voltage drop compensation with an external E42 transformer

The main problem of the dual core E6S design is the inability of the design to implement a proper discrete voltage boost in the design. Changing the ratio from 6:6 to 6:7 would lead to a voltage boost of 20V RMS, which is too much for a 120V AC grid. Therefore, a small transformer could be added, as shown in figure 4.4.

This transformer, with a winding ratio of 20:1, will fully be placed at the secondary side of the main transformer. This way, the demands for winding isolation and creeping distances are more convenient, as now only standard isolation, instead of reinforced isolation needs to be implemented. It will be fed by the same 25kHz block voltage and will therefore also boost the voltage to the secondary H-bridge with 1/20th of the input voltage. Since the RMS input voltage is equal to 120V, the voltage boost would be equal to 6V. At low loads, with an internal voltage drop of ~3V over the converter, this would result in an output voltage of 123V. At full load, with an internal voltage drop of 9V, an average output voltage of 117V is reached, thus always leaving the difference between the input and output voltage less than 3%.
In order to minimize the size, an E42 core is used. Due to the low core diameter, the minimum number of windings will also be quite high, in order to keep the core out of saturation. Therefore, the transformer will have to be implemented with 40 litz windings primary and two times two foil windings with a diameter of 20·0.4mm.

When applying this solution, the main transformer will also have to deliver 1/20th more power. However, due to the current thermal behavior of the 6:6 design, no problems are expected.

4.6 Transformer optimization at higher switching frequencies

Unfortunately, the presented single E65 and the single E80 core transformer resulted in higher transformer power losses, than the given design thermally could handle. The general possibilities of these transformer types seem quite interesting, especially because of the fact that they are standard components, whereas the E65 dual core design would have to be specially built. The problem, which caused the overheating of the core, could be found in the fact that the winding length was too long, in relation to the foil thickness. Therefore, the internal resistance was too high.

A reduction of the number of windings was therefore a viable possibility. Of course, this would increase the total converter switching losses. However, the heat sinks are able to dissipate more power, as the maximum average current has decreased from 16A to 15A per IGBT. Therefore, a number of calculations have been made on various configurations. The results for the single E65 core configurations are shown in table 4-2.

<table>
<thead>
<tr>
<th>Switching frequency</th>
<th>Winding ratio</th>
<th>Voltage boost</th>
<th>Foil thickness</th>
<th>Conduction losses</th>
<th>Core losses</th>
</tr>
</thead>
<tbody>
<tr>
<td>25kHz</td>
<td>9:10</td>
<td>13.3V</td>
<td>0.15mm</td>
<td>11.4W</td>
<td>100%</td>
</tr>
<tr>
<td>30kHz</td>
<td>8:8 (8:9)</td>
<td>0V (15V)</td>
<td>0.18mm</td>
<td>8.0W</td>
<td>120%</td>
</tr>
<tr>
<td>35kHz</td>
<td>7:7 (7:8)</td>
<td>0V (17.1V)</td>
<td>0.20mm</td>
<td>6.3W</td>
<td>140%</td>
</tr>
</tbody>
</table>

The results show that the frequency has to be increased to 35kHz, in order to be able to reduce the conduction losses in the coils to an acceptable level. For the past calculations, no core losses have been calculated, as the transformers were only bothered by overheating of the windings, not due to over temperature of the core.

Due to the low number of windings at 30kHz and 35kHz, an extra winding for the secondary coils will result in too high values for the voltage boost. Therefore, an increase of switching frequency would reduce losses, but would also take away the possibility for voltage boost, resulting in the same problem the dual E65 core configuration is facing.

<table>
<thead>
<tr>
<th>Switching frequency</th>
<th>Winding ratio</th>
<th>Voltage boost</th>
<th>Foil thickness</th>
<th>Conduction losses</th>
<th>Core losses</th>
</tr>
</thead>
<tbody>
<tr>
<td>25kHz</td>
<td>14:15</td>
<td>8.6V</td>
<td>0.2mm</td>
<td>11.3W</td>
<td>100%</td>
</tr>
<tr>
<td>30kHz</td>
<td>12:13</td>
<td>10V</td>
<td>0.23mm</td>
<td>8.5W</td>
<td>120%</td>
</tr>
<tr>
<td>35kHz</td>
<td>10:11</td>
<td>12V</td>
<td>0.27mm</td>
<td>5.9W</td>
<td>140%</td>
</tr>
</tbody>
</table>
Table 4-3 contains the test results for the E80 core. For all frequencies, the voltage boost can be maintained, up to 12V at 35kHz. This is quite high, but will still keep the voltage boost on the output less than 10% in all situations.

4.7 Adjustments for a 120V/230V in, 120V/230V out system

Now the push pull technology has proven to work, an interesting option can be explored: a single converter, of which both the input and the output can be configured between 120V push pull configuration and 230V H-bridge configuration.

As shown, the converter works for grid and load voltages of both 120V and 230V. The only difference is the fact that a primary or secondary IGBT bridge should be switched as a push pull converter for 120V grid or load voltages, and as a H-bridge for 230V grid or load voltages. The circuits and switching schemes for a 120V-230V converter and a 230V-120V converter are shown in figure 4.5. As can be seen, only minor changes in the hardware are necessary.

This way, it is possible to apply different grid- and load voltages. The primary bridge can be switched as push pull converter and the secondary bridge as H-bridge, in order to convert a 115V grid voltage into a 230V, or vice versa. Now, another function of 50Hz galvanic isolation transformers can also be performed with the HF galvanic insulator: power transfer from different types of AC grids. One only has to realize that the load frequency will still be equal to the grid frequency: a 60Hz 115V grid voltage will also result in a 60Hz load voltage. Certain domestical appliances might not be able to handle this change of grid frequency.

The control could be done automatically, as the grid voltage is already an input of the central control unit. Together with an extra input, determining the required value of the
load voltage, the control unit could then simply decide how to switch the primary and secondary IGBT bridges.

The only sub circuits, which should be switched, are the power connections to the main transformer and the gating signal to half of the IGBTs. This could both be done with a relay. In the future, gate signal generation could also be made part of the central control unit. In that case, inverting the gate signal can also be done within the software. As a result, only a 30A 230V AC power relay for switching the main current would be necessary.

4.8 Practical implementation and measurements
Since it was the main goal to implement the 120V push pull converter on the basis of the current 230V DAB design, numerous additional circuits were already present. In order to be able to detect any HF noise as good as possible, the input and output common mode filters had been disabled. The common mode coils have been short circuited. The gating circuit, current and temperature control and 12V power supply could directly be used.

The only possible problem, which might occur, is the fact that the 12V power supply will shut down when the RMS grid voltage is lower than 80V. This might occur in case of a voltage dip. However, due to the fact that most 120V equipment will also not be able to operate at 80V anymore, this is not regarded as a problem.

4.8.1 Enlargement of the primary and secondary capacitor banks
The original primary capacitor bank of 6μF needs to be increased, in order to be able to compensate the fact that the RMS voltage has decreased and, as a result, also the energy in the capacitors. The switching frequency still is 25kHz and the current ripple therefore is related to the boost factor and the output voltage. For the primary capacitor bank, this would result in a theoretical value of $6 \times \frac{230}{120} = 11.5 \mu F$.

For the moment, the primary capacitor bank is therefore replaced by the secondary capacitor bank of another unit. However, these capacitors are no longer series connected, but in parallel, resulting in a value of $8 \times 3.3 \mu F = 26.4 \mu F$. This is possible, due to the fact that the grid and load voltage has decreased, therefore decreasing the average capacitor voltage. For current tests, the capacitor bank can keep this high value. For future use, however, a reduction of the number or value of these capacitors is a logical option.

The secondary capacitor bank, originally consisting of two series connected sets of 4 capacitors of 3.3μF each, is also reconnected as a parallel connection of 8 capacitors with a total value of 26.4μF, due to the lower output voltage.

4.8.2 Standards for 120V systems
The design specifications in addendum C1 do not yet contain all demands, the eventual product should meet. Before the product can be placed on the market, it should first meet the international standards regarding this type of equipment. In the marine industry, the criteria for galvanic isolators are defined by the American Boat and Yacht Council (ABYC) recommended standard A-28. This standard is currently in the process of revision. Due to recommendations to the A-28 galvanic isolator standards committee regarding the need to retain safety grounding under all conditions, proposed revisions to this standard now make provision for fail-safe galvanic isolators.
To be considered “fail-safe,” an independent laboratory, like KEMA, must confirm that the isolator will either remain fully functional or remain a permanent, effective grounding path if it fails when subject to the maximum current capability of the grounding conductor for a given galvanic isolator current rating. For the GI, the last option is chosen. Since the current 230V version is already being tested by KEMA, these results can be taken along in the eventual design of the 120V version.

4.8.3 Test results
Addendum B3, “Measurement results 120V push pull design”, contains a number of graphs on efficiency, voltage drop and other characteristic details. It is clear that the efficiency is slightly lower than it is for the 230V DAB version. This is mostly caused by a relative increase in the voltage drop over the converter. This can be found in the PCB layout, due to the increased currents. Also other components, such as the increased capacitor bank, demand a higher current and therefore also cause higher conduction losses. Interesting is the fact that the power dissipation at no load is actually lower.

Further, the thermal measurements on the various transformers are interesting, as they validate the calculated data, presented in table 4-1.

4.9 Conclusion
An important conclusion of the presented theory and measurements is the fact that it appeared to be possible to connect the previously used IGBT based AC switches in parallel, in order to use them in a push pull configuration at an output power of 3.5kW, at 120V.

The main transformer design process delivered a working prototype, which can transport RMS currents of 30A without overheating and RMS currents up to 38A for some minutes, before the thermal control circuit will turn off the device. The only problem of the main transformer is the lack of voltage boost, necessary due to the voltage drop over the IGBTs. A few solutions have been presented, such as the external voltage boost transformer. Also the effect of an increase of switching frequency on the transformer design has been shown.

For future applications, the input and output filter should still be implemented, as well as a new set of grid and load connectors, as they are not rated for 30A. Further more, the PCB needs to be rerouted and the capacitor banks should be remodeled. The space, which can be won with this, can then be used for a bigger main transformer. Also, the individual switches could be thermally isolated by placing them on separate heat sinks. This way, the effect of different IGBTs, for example due to batch changes, can be investigated.

The 120V push pull converter is an interesting project, especially because of the fact that, in combination with the circuitry of the 230V DAB design, a universal HF transformer can be constructed. 120V or 230V grids can be mixed, as both the input and the output can be switched between push pull and H-bridge behavior, resulting in the belonging input or output voltage. The HF transformer can now not only galvanically isolated connect a grid and load, but also connect grids and loads, when each specific voltage is unequal. Therefore, this application should further be investigated, especially regarding the voltage drop over the Mass GI and the number of windings on the main transformer.
5 Output scalability by DAB phase shift control

The main problem of the current circuit designs is the fact that they sometimes will provide an over, but more often an under voltage to the user. If a high load is used, while the Mass GI is connected to an unsolid grid with a low voltage, for example due to long wiring, the output voltage of the 230V Mass GI can easily drop beneath 210V. Therefore, a variable ratio between the input and the output voltage can be very useful. All possible solutions preferably should be designed at a switching frequency of 25kHz, in order to fit into the current range of GI products.

The first technology, which has been applied in a standard model of the Mass GI, is the Dual Active Bridge with phase shift control. This chapter will show the changes in control, circuit layout and component design, necessary for implementation of a phase shift between the primary and secondary H-bridge of a DAB converter. Also, the results of measurements will be discussed.

5.1 Changes in gate signal control

In the standard 230V layout, the gating signal is produced by an astable multivibrator and, in case of errors or in the start up sequence, pulled down by the processing unit. This unit gathers information on temperature, output current and input voltage. For the phase shifted DAB however, this processing unit will be disabled.

All control should now eventually be done by the new control unit, the DSP board, as presented in addendum D2. The current main task of this DSP board, capable of supplying six different PWM signal pairs, is to provide the gating signals for the primary and secondary H-bridge. These PWM signals are fed to two standard gate driver circuits, resulting in a +/-12V gating signal. The gate driver circuits are also shown in addendum D2.

In order to be able to react to over heating or over currents, a number of input signals can be defined. Since all signals are already present within the current circuits, they only have to be adjusted to the maximum ratings of the input channels of the DSP board. The belonging circuits are presented in addendum D2, but are not used for this application yet.

5.2 Calculation of added inductance

In chapter 3, the reason was shown why an inductor should be inserted between the primary and secondary H-bridge: during the phase shift, the di/dt in the inductor should not rise too fast, thus creating high peak currents in the switches. The switching frequency is 25kHz, thus giving the phase shift control a frequency of 50kHz, as the phase shift is applied twice per switching period. The maximum value for the phase shift is set at 51\(\mu\)s, the maximum allowable current value is 40A. This results in the following minimum value for the added inductance

\[
L_{\text{DAB}} \geq \frac{V_{\text{load}} + V_{\text{load}}}{I_{\text{ripple}}} \cdot t_{\text{phase shift}} = 181\mu\text{H}
\]  

(5.1)

For the tests, lower voltages have been used, so an inductor of 100\(\mu\)H was used.
5.2.1 Integration of Leakage Inductance in the main transformer

The transformer and the additional leakage inductor can be integrated in a cost and space-efficient design through the simultaneous control of the leakage inductance and the magnetizing inductance of the assembly.

The primary and secondary foil windings of the transformer have to be wound on a closed-path ferromagnetic core, for example the current E65 core. The desired magnetizing inductance is achieved by controlling an air gap disposed in the magnetic flux path of the core, while the desired leakage inductance is achieved by winding a certain portion of the secondary turns on a loosely coupled leg of the core. A non-magnetic housing and a spring clamp can then hold the assembly and maintain the air gap constant. Figure 5.1.a shows a practical implementation of additional leakage inductance in the main transformer by adding an air gap in the bobbin.

Figure 5.1.b shows a possible solution, using an extra core. This design will give a higher leakage inductance value at a lower volume and will also have less influence on the surrounding circuits due to the fact that the stray flux is centered in core2, thus decreasing the potentially harmful electromagnetic fields in the surroundings of the transformer.

![Figure 5.1. Possibilities for integration of leakage inductance in the main transformer. a) Air hole for adding leakage, b) more compact design, with additive core for less stray flux](image)

The increased proportion of leakage inductance can also be achieved by introducing an air gap in the core design, thus reducing the permeability of the core and therefore the value of primary inductance. The ratio of flux that does not link the primary winding to the secondary winding will therefore increase relative to the flux that links both windings.

Due to the fact that, for the current design, a standard main transformer was used, the extra inductor has been added as supplementary leakage inductance in this device. The inductor has been externally added, as can be seen in addendum D2.

5.3 Practical implementation and measurements

The implementation of the phase shifted DAB started with getting to know the possibilities of the DSP board and the belonging software, as shown in addendum D2 and [22, 40]. Various tests have been done to get to know the performance of the cascade switched circuits of the DAB and the gate driver.
Since the PCB of a standard Mass GI was used as a base for the phase shift controlled DAB, also the current IGBT switches, gate resistors, filter circuits, main transformer and component cooling were taken over.

The next step was the implementation of the circuits in a Mass GI. The measurement results are shown in addendum B4. It is clear that the phase shift between the primary and secondary bridge causes high values for the di/dt, leading to a high current stress on the switches. Also, the MOSFETs of the primary snubber circuit showed a high dissipation of power, due to the large spikes, caused by the added inductor.

At high loads, a boost of the output voltage was impossible, as expected. Only for low loads, both an increase as a decrease of the output voltage, in relation to the input voltage, was still possible.

5.4 Conclusion

The measurements proved that a realistic DAB for high currents is not feasible at the current conditions. In order to be able to reverse the polarity of the current through the leakage inductor fast enough, either the applied voltage should increase or the inductor value should decrease. The first option is not applicable: grid voltage values are a given fact. At zero crossings of the grid voltage, the voltage difference should therefore fully be made by the voltage of the capacitor, while also supplying the output load. This would lead to high capacitor values, but that would decrease the output voltage again. The second option will also not work, due to the too high di/dt values, which can damage the switching devices.

A solution could be found in an increase in switching frequency. With a reduction of the switching period, also the maximum phase shift in seconds per period is smaller. This could also lead to a reduction of the added leakage inductor value, to less energy in the switching spikes. The number of spikes however would increase. The output capacitor could then be increased, which might enable higher current flows. However, due to all current theoretical and practical constraints, this option will not further be investigated.
6 Output scalability by primary boost converter control

Due to the fact that the phase shifted DAB gave too many problems at the given constraints, another way had to be found to implement an output voltage control within the Mass GI.

A solution has been found in an added boost converter, in front of the primary H-bridge. A standard boost converter is shown in figure 6.1.a. The positive aspects of the design is the fact that it can be implemented within the current design, as it will be inserted between the current incoming grid power line and the primary H-Bridge. A simple connector on the PCB could even be used to choose between cheap versions, without the boost circuit, and the full versions, with the boost circuit. Therefore, also this design will have to work again at a switching frequency of 25kHz.

6.1 General working principle of a boost converter

The AC-AC boost converter is not a common converter, but has already been investigated [17, 33]. Also interesting for this subject were a number of papers on AC-AC buck-boost converters, [31, 44], as these authors had a number of comparable problems, which had to be overcome. In order to gain more insight in the characteristics of a boost converter, first an ideal DC boost converter will be presented. Subsequently, the changes, necessary for implementation in AC applications will be discussed.

6.1.1 Ideal DC boost converter design

Generally, boost converters are implemented for DC sources. The general layout is shown in figure 6.1.a.

In stage A, shown in figure 6.1.b, the switch $S_{\text{boost}}$ directly connects the boost inductor over the voltage source. At the same time, the load is fed by the energy in the output capacitor, leading to a decrease of the output voltage.

In stage B, shown in figure 6.1.c, the switch $S_{\text{boost}}$ is opened again, connecting the boost inductor to the load, thus reversing the polarity of the voltage over the inductor. Now, a current can flow through the diode to the output capacitor and the load. As a result, the average output voltage will be higher than the input voltage.

![Diagram of Boost Converter](image)

**Fig. 6.1. Boost converter topology. a.) general circuit design, b.) stage A, c.) stage B, d.) AC boost design**

6.1.2 AC boost converter design

In DC boost converters, the switch is usually implemented as a single MOSFET, due to the small switching losses. For implementation in an AC grid, this switching device however also should be able to handle AC currents. Also, the diode will have to be replaced by an AC switch, as shown in figure 6.1.d.

The main problem, which now might occur, is the presence of a possible dead time, during which both switches are turned off. In this case, the current through the boost
inductor could not find a path any more, leading to high peak voltages, until one of the switches is conducting again. A snubber circuit should therefore be implemented over switch $S_{\text{boost}}$ in order to limit these voltage peaks.

At the same time the switches $S_{\text{boost}}$ and $S_d$ should also not be activated simultaneously: if, for example the switch $S_a$ would already be activated, while $S_d$ is still conducting, a short circuit will be made on the secondary side, leading to peak currents, flowing from the output inductor through both switches.

### 6.2 Required boost ratings

Before starting the design, a number of specific values need to be set. First of all, the possible grid voltage range will be set at 196V (-15%) to 241V (+5%), as these ranges are common limits for grid voltages.

A 19:20 transformer will still be implemented, but this could be changed into a 19:19 or 20:19 transformer, in order to be able to connect the device to 240V grids, as in Great Britain. With the current 19:20 transformer, this would create a maximum output voltage of 253V at no load, regarding to the measurements in addendum B2.

Given a certain switching period $T_s$ and an on time of the boost switch $t_{\text{on}}$, resulting in a duty cycle $D$, the average ideal ratio between the input and the output voltage can be calculated with equation 6.1. This results in a ratio between the input and output current as shown in equation 6.2.

$$V_o T_s = \frac{1}{V_i (1 - t_{\text{on}})}$$  \hspace{1cm} (6.1)

$$I_o = \frac{V_o}{V_i} = 1 - D$$  \hspace{1cm} (6.2)

The boost function should, for the current 19:20 transformer, need to be able to deliver 230V load voltage at full load and 196V grid voltage. The full load would cause a voltage drop of ~15V, according to the measurements in addendum B2. A boost from 181V to 230V would therefore be necessary, equal to a duty cycle of 27%.

Due to the fact that the input power should be equal to the output power, the maximum RMS input current would have to increase from 16A to 21.9A, in order to maintain the same output power. However, since the most grid power connections are thermally or electronically fused at 16A, this would be impossible. The maximum output power would therefore be reduced to 2.56kW.

Offering a grid voltage, higher than 230V will, at the same time, not enable a higher output power, as the cascading DAB converter is limited at 16A.

### 6.3 Allocation of the boost circuit in a DAB converter

The boost converter could be placed on the primary and on the secondary side. However, in order to keep the total topology as robust as possible, the boost trap is placed on the primary side. This way, any losses, caused by this circuit, will not affect the DAB converter efficiency.
As presented in the previous paragraph, in the normal boost converter design, a diode ensures a power flow to only one side. As has already been shown before, this diode is not possible in the new design, because of the fact that AC signals are switched. If the boost circuit therefore would be placed on the secondary side, another AC switch, replacing the standard diode, would have to be added in order to ensure a correct power flow.

6.3.1 Primary H-bridge switched as boost switch

In theory, the primary bridge, could act as boost switch, simply by activating one or both legs of the primary H-bridge. Due to the cross connected gating signals of the IGBT gating circuits, a phase shift between the gating signals to the primary bridge would already give the desired effect. Removing the cross connection of the gating signals of the secondary H-bridge, thus using each gating signal to control both switches of a phase leg, would result in the desired function of $S_b$ in the ideal switching scheme of figure 6.3.

As can be seen, the primary and secondary H-bridge of the DAB converter are switched differently and have become part of the boost converter. This solution however will not be chosen, as it will lead to higher conduction losses on the moment, that the primary bridge needs to act as boost switch. The total voltage drop would then be equal to the voltage drop over two IGBTs and two diodes. Implementing the boost switch as a separate switch, as shown in the original boost scheme in figure 6.2, will lead to a voltage drop of only one IGBT and one diode.

The increased voltage drop over the boost switch could be decreased by activating two of these series connected AC switch pairs could be connected in parallel, and therefore activating all switches in the primary H-bridge. This would then lead to duty cycles, which are higher than 50%, which would lead to a redesign of the gating circuits, as the standard gating transformers can't be used in that case any more.
Implementing a separate boost switch will also solve another problem. As has been shown before, a minimum amount of dead time is necessary when switching a H-bridge. In the previous paragraph, the necessity of a snubber circuit in order to constrain the boost inductor voltage was also shown. Switching the primary H-bridge in the time that the boost switch is activated, will solve this problem: as soon as the boost switch is turned off, the primary H-bridge is conducting and provides a path to the primary and secondary snubber circuits.

The disadvantage of adding an AC switch with the belonging gating circuits will therefore easily be overcome by the stability, the extension will bring along.

6.4 Application of a boost circuit in a Mass GI converter
Taking all constraints, as presented in the previous paragraphs, into account, a standard circuit of a Mass GI DAB converter with a primary boost circuit can be designed. The result is shown in figure 6.2.

The boost circuit includes a separate switch $S_{boost}$, periodically short circuiting the primary bridge. The switching sequence, belonging to the circuit in figure 6.2 is shown in addendum D3.

Due to the fact that the secondary bridge turns off before the primary boost switch turns on and that the secondary bridge only turns on again a short period after the boost switch has turned off again, there will never be a conflict. In this short dead time, the snubbers will be needed to limit the transformer voltage.

The primary bridge will change it's polarity during the middle of the boost pulse, thus always being turned on when the boost switch turns on or off. This way, the primary boost switch and therefore also the boost inductor are always connected to both the primary as the secondary snubber on the moments that the current through the boost inductor can't flow through the boost switch or the secondary H-bridge, as mentioned above.

6.5 Calculation of boost inductance
The value of the boost inductance is very important, as this will limit the $di/dt$ during the on time of the boost switch. Limitation is necessary, in order to protect the switching components of the boost and the DAB converter, as well as the main transformer. High peak currents will also lead to higher losses, as they go up with the quadratic value of the current.
Now the following formula can be used for calculation of the current ripple and the minimum boost inductor value, as shown in equation 6.3 and 6.4.

\[
I_{\text{ripple}} = \frac{V_{\text{grid}}}{L_{\text{boost}}} \cdot D \cdot T_s = \frac{\dot{V}_{\text{load}}}{L_{\text{boost}}} \cdot D(1-D) \cdot T_s \quad (6.3)
\]

\[
L_{\text{boost}} > \frac{\dot{V}_{\text{load}}}{I_{\text{ripple}}} \cdot D(1-D) \cdot T_s \quad (6.4)
\]

The maximum switchable average current of the IGBTs is 40A. The maximum value of the 50Hz current through the IGBTs will be 22.6A, leaving room for a maximum current ripple of 17.4A. At a peak load voltage of 325V and a maximum duty cycle of 27%, this would lead to a theoretical minimum boost inductor value of 741μH. However, in order to reduce the maximum ripple to less than 10% of the average current value, an inductor value of 567μH would be necessary.

High values of the current ripple will lead to higher RMS current values, thus leading to higher converter losses. Therefore, this ripple should be as small as economically and technically possible.

At the value of 567μH and at a maximum resistive load, the output voltage of course will drop, due to the impedance of the inductor and the 50Hz grid voltage. The voltage drop however will mostly be caused by the internal copper resistance of the coil, as shown in equation 6.5.

\[
V_{\text{load}} = V_{\text{grid}} \cdot \frac{R_{\text{load}}}{\sqrt{(R_{\text{load}})^2 + (2 \cdot \pi \cdot f_{\text{grid}} \cdot L_{\text{boost}})^2}}
\]

\[
V_{\text{load}} = 230 \cdot \frac{14.4}{\sqrt{(14.4)^2 + (0.178)^2}} = 230V
\]

It is clear that the boost inductance will not affect the impedance of the output voltage. The phase shift between the primary and secondary voltage, due to the boost inductor is equal to 0.710° and is therefore neglectable.

### 6.6 Output capacitor calculation

The absolute and relative output voltage ripple, and the belonging output capacitor value, can be calculated using formulas 6.6 and 6.7.

\[
\Delta V_o = \frac{\Delta Q}{C} = \frac{I_o \cdot D \cdot T_s}{C} = \frac{V_o}{Z_{\text{load}}} \cdot D \cdot T_s \quad (6.6)
\]

\[
\Delta V_o = \frac{D \cdot T_s}{V_o \cdot Z_{\text{load}} \cdot C} \quad (6.7)
\]

\[
C_{\text{min}} > \frac{D_{\text{max}} \cdot T_s \cdot I_{\text{o max}}}{\Delta V_{\text{o max}}} \quad (6.8)
\]
The maximum voltage ripple will be reached at the maximum on time of the boost switch, being 27%. When the relative voltage ripple is given a maximum of 1%, the corresponding minimum capacitor value would be 53\mu F. This capacitor would then have to be placed at the secondary side of the DAB boost converter.

### 6.7 Switch design

Due to the fact that the boost switch will never have to switch any over currents, in case of short circuits, the switch design can be relatively simple and can even be made with two series connected MOSFETs with built in anti parallel diode, as presented in chapter 2.

For a first implementation, the standard AC switch design, based on two IGBTs, type HGTG20N60B3D, can also be chosen. Switching losses now will be higher, but they will be able to handle the current easily and they have proven to be robust, in case of voltage spikes.

The switch should eventually be equipped with a current sensor, in order to protect the switch and other components in case of over currents or short circuits. In case of saturation of the boost inductor core, the di/dt would be much higher, possibly damaging components. A current sensing circuit could therefore also detect any possible saturation of this inductor core.

### 6.8 Practical implementation and measurements

For the practical implementation of the boost switch, a primary bridge of a standard Mass GI has been cut out and used, as now also the gating transformer, gate resistors and heat sink can be used. The boost switch has been mounted on the back side of another standard Mass GI. This way, the connection between the switch and the primary H-bridge of the Mass GI converter is as short as possible, thus minimizing the inductance of the connection. The eZDSP board has also been mounted on the back of the Mass GI converter and controls four gate drivers, as presented in addendum D2. Control of the eZDSP PWM outputs however gave problems, possibly due to problems in the memory initialization. The intended output signals and a number of pictures of the converter can be found in addendum B5.

### 6.9 Conclusion

Performed simulations and the belonging switching schemes show that connecting a boost converter in front of the standard, non phase shifted DAB topology of the Mass GI can give good results.

The primary bridge of the Mass GI could be used as switch, but this might lead to a number of problems. A big disadvantage of this usage of the primary bridge is the fact that the boost inductor then will need an extra snubber, for the time the primary bridge is switching between it's functionality as boost switch and H-bridge switch. Looking at figure 6.2, this implementation of the boost switch would also lead to higher losses during the time it is acting as boost switch, as the total voltage drop is higher, due to the series connection of 2 AC switches.

Unfortunately, measurements could not be performed due to lack of time and initialization problems in the DSP board. The general layout however has theoretically proven to be quite interesting and should therefore be investigated more thoroughly in the near future.
7 Conclusions and discussions

For all presented circuits, the basic setup has been equal: The 50Hz sinusoidal grid voltage is being modulated with a 25kHz square wave signal. This means that the current, flowing through the main transformer is also a 25kHz signal. This results in a decrease of main transformer size, but in an increase in used components. This thesis presents two groups of converters: converters with a constant ratio between the input and output voltage and converters with a controllable output voltage.

The current Mastervolt Mass GI design has been evaluated. Both hardware and control improvements have been suggested. A number of hardware improvements are already implemented in the product. The presented hardware changes can improve not only the reliability of the design, but also the efficiency. The currently used control software has been evaluated and a new control scheme has been presented. Implementation of this new control scheme will further increase the reliability of the design.

In order to expand the product range, a 120V 30A push pull converter has also been developed. Various main transformer designs have been calculated and evaluated. Since the general hardware layout is based on the 230V DAB design of the Mass GI, setting up production of this new product can be relatively simple. A number of issues still have to be investigated. Input and output filter circuits still need to be added. Also, the effect of a slight increase of switching frequency should be investigated, in order to reduce the size of the main transformer.

The first converter with controllable output voltage, which has been investigated, is the phase shifted Dual Active Bridge. Given the current design constraints, it appeared not to be possible to reach the correct output voltage at high loads. Due to the necessary high value for the leakage inductor, the energy in this leakage is also higher than in the case of the non-phase shifted DAB. Therefore, the snubber circuits have to dissipate more power. Test results confirm these calculations and simulations.

The second, more promising design, also based on the Mass GI DAB layout, uses a primary boost converter. The presented circuits show several possibilities for implementation of a boost converter in a DAB design. However, in order to keep the converter circuit simple, an external boost switch has been added. Due to lack of time, this converter could not yet be tested. Due to the interesting possibilities, this design should further be investigated, as it has the potential to become an interesting product.

As a result, it is clear that the current Mass GI design is interesting and robust, but can be made even better. Various options for future improvements have been presented and can lead to a further increase of the Mastervolt HF galvanic Insulated Shore Power connection product range.
8 Literature overview


High frequency galvanic insulated shore power connections


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ADDENDUM

A

Circuit and component design
A.1 Schematics Mass GI V3

For more info regarding the schematics, contact Mastervolt:

www.mastervolt.com
For more info regarding the schematics, contact Mastervolt:

www.mastervolt.com
For more info regarding the schematics, contact Mastervolt:

www.mastervolt.com
Electrical insulation in the MASS GI

Low Voltage control circuits
A.3 Calculations for the gating and dead time circuits
This chapter will further explain the calculations and choices, as they have been made for the gating circuit.

A.3.1 Variation of $V_{GE}$
IGBTs are voltage controlled devices and therefore require a sufficient gate voltage to establish collector-to-emitter conduction. A $+15$ V positive gate drive is normally recommended to guarantee full saturation and limit short circuit current. A negative voltage bias is used to improve the IGBT immunity to collector-to-emitter dv/dt injected noise and reduce turn-off losses.

As shown in table 3.1, the amplitude of this gate voltage $V_{GE}$ can influence quite a lot of IGBT characteristics, such as switching losses. Since it is known that these losses for example are caused by the fact that internal capacitors need to be loaded, it is clear that higher voltages will faster load the gate bus to an appropriate voltage. At the same time, the maximum value for $V_{GE}$ certainly is limited, usually to $+/20$ V. The exact value can be found in the datasheets.

Secondly, the on state voltage $V_{GE+}$ can have different values than the off state voltage $V_{GE-}$. A recommended value for the on state voltage is $12-15$ V, for the off state voltage levels between $-5$ and $-15$ V usually are applied. However, as can be seen in the eventual gating circuit, due to usage of gating transformers it is usually best to choose a balanced amplitude of these signals at a constant level, in this case. $12$ V for $V_{GE+}$ and $-12$ V for $V_{GE-}$.

A.3.2 Calculation of the gate resistance
The value of the gate resistance, $R_c$, connected to the gate drive output via the gating transformers is determined based on the peak currents $I_{ON}$ and $I_{OFF}$, which charge and discharge the gate terminals of the IGBTs. These maximum currents which can electrically charge and discharge the gate oxide between the gate and the emitter are determined based on the maximum current of the gate drive circuit. The minimum gate resistance value, $R_c$, is, in turn, determined based on these determined maximum charge and discharge currents.

Usually, data sheets contain standard values for the gate resistance, when supplying values for the switching times. The greater the value of the gate resistance $R_c$, the longer the switching time and the greater the switching losses will be. However, as $R_c$ increases, the surge voltage during switching becomes smaller, thus decreasing the chance for a dv/dt shoot through current. In addendum C4, it can be seen that for the measurements in the datasheet of the HGTG20N60B3D, a gate resistor of $10\Omega$ has been used, at a gate voltage value of $V_{GS}=15$ V.

However, in this case, the situation is different from the situation shown in the data sheet. When the H-bridge switches, all previously conducting IGBTs should be off, before the new IGBTs can be turned on. The current of the newly activated IGBT should not start rising, before the current in the series connected deactivated IGBT has stopped. Therefore, a certain dead time should be built in by using the right value for $R_{on}$. A higher value will decrease the load current for the internal capacitor, leading to a slower rise of $V_{GE}$, and therefore to a longer delay until the new IGBT is turned on.

This implies that three IGBT parameters are important: for the IGBT, which is turning off, the Current Turn-Off delay time $t_{g(OFF)}$ and the current fall time $t_f$ are important. For
the IGBTs, which are turning on, the current turn-on delay time $t_{d(ON)}$ is important. All these times can be found in the datasheet.

The total minimum delay, which should be met in order to stay safe, is now:

$$t_{dead\ time} = t_{d(OFF)} + t_f - t_{d(ON)}$$

$$t_{dead\ time} = 220 \cdot 10^{-9} + 140 \cdot 10^{-9} - 20 \cdot 10^{-9} = 340\text{ns}$$

Another important factor for determination of the gate resistance is the gate voltage. In addendum C4, figure 3, the following value for the input capacitance $C_{IES(max)}$ of the HGTG20N60B3D can be found: $C_{IES(max)}=4.5\text{nF}$. The voltage, coming from the gating circuit is switching between -12 and 12V.

Knowing that the current IGBT typically will start conducting as from $V_{GE}=5\text{V}$, as shown in the electrical specifications, the desired value for the gate resistance will then be:

$$V_{GE(min)} = \left( V_{G+} - V_{G-} \right) \cdot \left( 1 - e^{-\frac{t_{dead\ time}}{R_{gate} \cdot C_{IES}}} \right)$$

$$R_{gate} = \frac{-t_{dead\ time}}{\ln \left( 1 - \frac{V_{G+} - V_{G-}}{V_{G+} - V_{G-}} \right) \cdot C_{IES}}$$

$$R_{gate} = \frac{-340 \cdot 10^{-9}}{\ln \left( 1 - \frac{5 - 12}{12 - 12} \right) \cdot 4.5 \cdot 10^{-9}} = 61.3\Omega$$

When a small overlap is permitted, in which both the deactivating and the activating IGBT are turned on, this value can be reduced.

**A.3.3 Division of the gate resistance into $R_{on}$ and $R_{off}$**

So far, the gate resistance has had only one value. However, when developing gate circuits, it is often advantageous to divide the gate resistance into two values, one for turning the IGBT on and one for turning the IGBT off. The turn-on behavior, for example, can be affected by the gate resistance specified in the data sheet, whereas turn-off can be slowed somewhat with an increased gate resistance value to reduce switching over voltage spikes and in this case, also to build in a delay to prevent short circuits in the H-Bridge.

Splitting up the gate resistance can be done by placing a second resistor and a series diode in parallel to the actual gate resistor. The turn-on path therefore has higher resistance than the turn-off path. The resulting schematics are shown in figure A1. As can be seen, a 25kHz signal is generated, boosted and sent into a transformer with one primary and four secondary windings, all with the same number of windings. All four secondary windings then are connected to IGBT pairs, the switches of the H-bridges.
The IGBTs' rate of current change can be reduced somewhat by a moderate increase in gate resistance. However, care must be taken not to make gate resistance so high that it acts together with the IGBTs’ parasitic components to produce gate voltage oscillations. However, increased gate resistance will also result in a substantial increase of turn-on/turn-off power losses in comparison to those presented in the data sheets. Therefore, the previously value of 61.3Ω will be used.

Looking at Roff, it is clear that this value should be as low as possible, therefore unloading the input capacitance as fast as possible, as only then the IGBT will turn off. The main limiting factor in this setup is the maximum peak current, which can flow through Roff and the series connected diode. First of all, the average power in the resistor should not exceed 0.2W. At a voltage difference of 24V, this would mean a maximum current of 8.3mA. However, this current is not continuously flowing, but maximally for estimated 100ns every 40us. This enables the maximum peak current to be 33.3A. At this value, for this time, the average power will not exceed the 0.2W.

The diode is different: in diode datasheets, usually the non-repetitive peak forward current is mentioned. In this case, the used BAV70 dual diode can deal 4A per diode for 1μs. Since both internal diodes are connected in parallel, the maximum current through these devices is 8A. Setting this as the maximum peak value for the turn off current, the minimum value for Roff can now be calculated:

\[
R_{off(\text{min})} = \frac{V_{G+} - V_{G-}}{I_{G(\text{max})}}\\
R_{off(\text{min})} = \frac{12 - 12}{8} = 3\Omega
\]  

Due to the fact that in the practical implementation, Ron and Roff are connected in series when a positive voltage is applied to the gating circuit, Ron actually can be 61.3-3=58.3Ω.

A simulation with a varying value of Ron is shown in figure A2. It is clear that for high values of Ron the current peak when switching on is decreased. For lower values of Ron, it is clear that the peak level increases, but that, at the same time, the duration slightly decreases.
A.3.4 Dead time circuit design

As mentioned in chapter 4, there are two options for reducing the losses, caused by this shoot through. First of all, the value of the turn on resistor in all gating circuits can be increased. Another, more efficient option is the implementation of a dead time in the gating signal.

In order to implement an additional dead time, a safety circuit, already present in the Mass GI is used. Normally, this circuit pulls down the 12V block signals, going to the gate driver, for example in case of over current or in order to enable Master-slave applications. Now, this circuit will be used to add a dead time to the gating signals every half period.

A.3.4.1 Schematics

As can be seen in fig. A3, the dead time control is relatively simple. First of all, it is necessary to see how the 25kHz square wave signal is generated. As can be seen in the general schematics in addendum A1, this signal is generated by the IC HEF4047BT, which is an astable multivibrator. This device uses an external resistor and capacitor to control the output frequency, as shown before. The drivers, providing the supply voltage to these components, switch at a frequency, which is twice as high as the output frequency. Therefore, this voltage can be used to implement a dead time signal for each half period, thus for both elements of the H-Bridges.

The way this is done, is shown in figure A3. The schematics show the HF square wave source as the voltage sources V1 and V2. They are actually implemented in the IC, responsible for making the 25kHz block signal. The signal, going to the output frequency determining capacitor, is now also lead to a RC circuit, which is providing one of the inputs of a comparator with an e-curved voltage. The dead time capacitor is smaller than the capacitor, used for determination of the output frequency. In order to enable a wide range of dead time settings, the resistor is set at 20kΩ.
The resistor voltage now can be calculated with the following formulas:

\[ V_{R_d}(t) = 12 \cdot \left(1 - e^{-\frac{t_{d,t}}{\tau_{d,t}}}ight) \]  

(A3.5)

\[ t_{d,t} = t - n \cdot 20 \mu s \quad n \in \mathbb{R} \]  

(A3.6)

\[ \tau_{dead time} = R_{d,t} \cdot C_{d,t} = 20 \cdot 10^3 \cdot 220 \cdot 10^{-12} = 4.4 \mu s \]  

(A3.7)

The resulting voltage over the resistor is then compared with the voltage, set with the variable resistor \( VR_1 \).

A.3.4.2 Simulation results

In order to see if the circuit is designed properly, a simulation model of the circuit has been built. As can be seen in figure A.4, the voltage on the \( V^- \) connection of the comparator rises up to a level over 11V and then exponentially decreases. The green line represents the voltage on the \( V^+ \) connection of the comparator. Once the voltage on \( V^- \) is higher than on \( V^+ \), the internal emitter driver in the comparator pulls down the gating signals via D3 and D4. These diodes are already present in the 230V and are part of the currently present safety circuit.

Fig. A.4. Voltage levels after implementation of the dead time circuit.
The eventual output signals are also shown in figure A4 with the blue and red line. As can be seen, the output signals are low for 50% of the time. However, on the moment that the output of the square wave generator turns high, the dead time circuit pulls down both the high and low gating signals. This situation lasts as long as the voltage over the 20kΩ resistor R2 is higher than the voltage, set with the variable resistor. As soon as this point is reached, the internal emitter driver at the output of the comparator seizes to conduct, enabling the high gating signal to go to the gating driver.
A.4 Auxiliary circuits for the 230V Mass GI

The Mass GI clearly cannot operate without a number of auxiliary circuits, such as filters, a 12V power source and measurement circuits. Since they were no specific part of the main project, they have been explained in this addendum.

A.4.1 Input and output filters

Since the GI operates at a frequency of 25kHz, a high load of HF noise is produced within the GI. This noise has to be filtered out by a filter at the input and output side. However, these filters also are used in another way, as they also filter out the HF noise, which is present in the grid and the HF noise, produced by the equipment, connected to the GI.

The filtering is done by a Common Mode filter, an inductor, a number of capacitors to reduce the voltage ripple and a number of RC filters, to reduce the HF noise. Provided, that the total power in the capacitors should not exceed 5% of the maximum output power, the following calculation can be made for the maximum total filter capacity:

\[
C_f \leq \frac{5\% \cdot P_{o,max}}{U_o^{2} \cdot 2\pi f_o} \quad \text{(A.4.1)}
\]

\[
C_f \leq \frac{5\% \cdot 3500}{(230)^2 \cdot 2\pi \cdot 50} = 10.5 \mu F
\]

Since the input and output voltages are equal in the MASS GI, this capacitor value can equally be divided into a primary and a secondary bank.

A.4.1.1 Input capacitor bank

In the MASS GI, the input voltage is filtered twice. First of all, a 1\mu F MKP capacitor filters out disturbances directly at the input. MKP capacitors use a metallized polypropylene film and have extremely low losses due to a polypropylene dielectric. After the common mode filter, a capacitor bank of four parallel MKT capacitors of 1.5\mu F. MKT capacitors use a metallized polyester film. The goal of all capacitors is to reduce the voltage ripple. The total input capacitance therefore is 7\mu F.

A.4.1.2 Output capacitor bank

The output voltage ripple is first filtered out, by series connecting two groups of four parallel connected MKP capacitors of 3.3\mu F each. At both sides of the common mode coils a 1\mu F MKP capacitor is present. The total output capacitance is equal to 8.6\mu F. This does implement that the total capacitance of the input and output banks is 12.6\mu F, slightly higher than previously calculated. The total capacity power now is 6% of the maximum output power, which will cause relatively high capacitive currents. Even when no load is connected, a reactive current of 0.96A will be drawn from the grid, thus causing conduction losses.

A.4.2 Measurement circuits

In order to protect the GI from unwanted situations, such as under and over voltages and high output currents, the central controller constantly needs to be aware of the momentaneous input voltage and output current. Therefore, the GI is equipped with a
number of circuits, which accurately provide this data. The belonging schematics will now be discussed.

A.4.2.1 Input Voltage measurement circuit

For monitoring the incoming AC grid voltage, a compact voltage measurement circuit is implemented. The circuit is shown in figure A5. The 1.2MΩ resistors minimize the maximum current, which can flow from the grid into the control circuit. These resistors have to be special types, as there are several demands for AC measurement resistors. A thermal or over voltage breakdown should never lead to a short circuit within the resistor.

![Input voltage measurement circuit](image)

Fig. A5. Input voltage measurement circuit

These special resistors are part of a resistive division, which downscales and filters the AC grid voltage. Eventually, an output voltage between 0V and 2.5V is transferred to the central controller.

A.4.2.2 Output current measurement circuit

An important factor of the Mass GI is the current limiting circuit. This circuit has two tasks. First of all, it should pull down the gating signals in case of a secondary short circuit, leading to high current peaks. Secondly, the circuit should provide information on the momentaneous current to the central processor. This device then can also pull down the gating signals, in case of over current.

When a current protection is implemented into a circuit, one of the first considerations is the technique to detect the over current condition. Popular techniques involve current sense transformers, Hall effect current sensors, current sensing resistors, and even de-sat sensing on power transistors. For each method, tradeoffs have to be made between cost, size, simplicity, isolation, noise immunity, power dissipation and robustness. For this application, current transformers have been chosen, because of the robustness and simplicity.

In order to implement both mentioned tasks into one circuit, two 1:100 current transformers have been chosen, as shown in figure A6. In order to be able to also detect short circuit faults in the separate secondary H-bridge legs, each output of the secondary AC switches is being fed through a current transformer, where the two upper respectively two lower paths are flowing through the same current transformer.
The output current, $\frac{1}{1000}$th of the original value, is then fed into a $0.5\Omega$ resistor, causing a voltage. In order to pull down the gating signal in case of a short circuit or severe over current, a minimum voltage over this resistor of $0.6V$ will activate the pull down transistor. Therefore, for currents above $120A$ the gating signal will directly be disabled, without the intercourse of any other external control system.

However, the load current also needs to be limited for lower values. One of the design demands is the fact that the insulation transformer should simulate B-characteristic fuse behavior as a reaction on over currents. Therefore, the voltage over the current transformers is also fed to two amplifiers, each with a multiplying factor of $10$. As a result, a RMS current of $16A$, $22.3A$ peak, will lead to a rectified current transformer output voltage with a peak value of $0.113V$, resulting in a peak value of $1.13V$ on the input port of the PIC.

Taking into account that the maximum input voltage on the PIC is equal to $5V$, the maximum detectable current is equal to: $98.6A$. Any instantaneous currents higher than this value will be presented with the maximum $5V$ level.

**A.4.3 Auxiliary DC power supply**

Since all control and gating circuits operate at $12V$ or $5V$, a DC power supply has to be implemented in the MASS GI. This has been done around a standard power supply IC, the TOP244Y. The minimum RMS input voltage should be $80V$, which is enough to also survive voltage drops at both $230V$ and $120V$ in. Since this schematic is a standard circuit, given by the manufacturer of the IC, it will not be discussed further more.
A.4.4 Thermal power loss management

Now the general working principle, together with the belonging circuit power losses, is discussed, an overview can be made. This will show if the design is practically implementable. Therefore, first of all, a sum of all individual losses has to be made. Subsequently, removal of this thermal energy will be discussed.

A.4.4.1 Power loss overview

As we have seen, the various circuits each cause their own power loss. Summarized, these losses form the following table:

- IGBT switching loss
- IGBT conduction loss
- Transformer winding loss
- Transformer core loss
- Snubber loss
- Control circuit loss

The specific values for these losses can be found in the corresponding addenda.

A.4.4.2 Thermal energy flow

As shown in the previous paragraphs, power loss is taking place in several components. Since this power loss is transferred into thermal energy, the heat flows in the GI have to be carefully modeled, in order to prevent component overheating. In order to gain more insight in the thermal energy flow in the MASS GI, certain aspects can to be taken into account:

- Airflow
- Heat flow from the electrical components to the heat sink
- Variation in the heat sink temperature
- Maximum device temperatures

The last aspect, maximum device temperatures, is further discussed in the addenda on component design. Therefore, now the various thermal energy transfer possibilities will shortly be discussed.

A.4.4.2.1 Radiative heat transfer

The first phenomenon, enabling heat transfer, is thermal radiation. Thermal radiation is defined as electromagnetic radiation in the wavelength range of 0.1 to 100 microns, therefore invisible. The amount of heat transferred into or out of an object by thermal radiation is a function of several components. These include its surface reflectivity, emissivity, surface area and distance to other objects. Also the temperature of both objects is important.

If an hot object is radiating energy to its cooler surroundings the net radiation heat loss rate can be expressed as follows.

\[ q = \varepsilon \sigma (T_{\text{body}}^4 - T_{\text{c}}^4) A_{\text{object}} \]  

(A4.2)

with

\[ \varepsilon = \text{emissivity of the object (one for a black body)} \]
\[ \sigma = 5.6703 \times 10^{-8} \text{ (W/m}^2\text{K}^4) \] - The Stefan-Boltzmann Constant

\[ T_{\text{body}} = \text{hot body absolute temperature (K)} \]
\[ T_c = \text{cold surroundings absolute temperature (K)} \]
\[ A_c = \text{surface of the object (m}^2\text{)} \]

For a semiconductor in a T0247 body with a total surface of 9.9cm\(^2\), seen as a black body with a temperature of \(T_{\text{body}}=353\)K and a surrounding temperature of \(T=293\)K, the maximum power loss would be 0.87W. This value is a theoretical maximum, as the body itself will not have the characteristics of a black body. Also the nearby objects might partially reflect the thermal radiation, due to the fact that also these components do not have the characteristics of a black body.

For the main transformer design, with a total estimated surface of 200cm\(^2\), seen as a black body with a temperature of \(T_{\text{body}}=353\)K and a surrounding temperature of \(T=293\)K, the maximum power loss would be 17.6W. This value is again a theoretical maximum, as the main transformer itself will not have the characteristics of a black body, and neither will the nearby objects.

### A.4.4.2.2 Convective heat transfer

A temperature difference between a thermal energy producing component and the surrounding air can induce an air motion. This is known as "natural convection" and it is a strong function of the temperature difference between the component and the surrounding air. Blowing air over the device by using external devices such as fans can also generate an air motion. This is known as "forced convection".

Convective heat of a cube with dimensions l by w by h, with no convection through the bottom and no forced convection, such as the main transformer, can be calculated with formula A4.3.

\[ P_{\text{conv,con}} = 10^{-4}(9.00 \cdot (l + w) \cdot h^{0.75} + 3.52 \cdot (l \cdot w)^{0.75} \cdot (l + w)^{0.25}) \cdot (T_{\text{obj}} - T_{\text{amb}})^{0.25} \]  \hspace{1cm} (A4.3)

A quick calculation for the original main transformer design of the 230V Mass GI, with \(l=b=6.5\)cm and \(h=5\)cm and at 80°C component temperature and 25°C ambient temperature, would give an estimation of 2W.

Since the IGBTs are placed between clips and the heat sink, thus reducing the surface, which can enable convective heat transfer, to a minimum, this mechanism will not be investigated for the IGBT heat transfer.

### A.4.4.2.3 Conductive heat transfer

As can be seen in the previous paragraphs, the main transformer can be cooled by simply using a fan. The IGBTs and snubber MOSFETs however will need extra cooling by means of a heat sink. The heat transfer between for example the IGBTs and the heat sink is called conductive heat transfer. Calculations can be made with the following formula:

\[ P_{\text{con,cond}} = \frac{(T_{\text{obj}} - T_{\text{amb}})}{R_{\text{th,cond}}} \]  \hspace{1cm} (A4.4)

with

\[ R_{\text{th,cond}} = \text{conductive thermal resistance from object to ambient (K/W)} \]
\[ T_{\text{obj}} = \text{object temperature (K)} \]
\[ T_{\text{amb}} = \text{ambient temperature (K)} \]
The thermal resistance, due to the usage of the ceramic isolators between the IGBTs and the heat sinks for example is 2.0K cm²/W. At an IGBT surface of app 1.3*1.6=2.1 cm², this will result in a thermal resistance of 0.95K/W. For a given IGBT power loss of 10W, this would mean that, thermal loss only transferred by conduction, would give a thermal difference of 10.5K between the IGBT and the heat sink.

A.4.4.3 Heat sinks
As can be seen in addendum E1, the losses in the IGBTs can reach up to 15W. The snubber MOSFET can even dissipate up to 60W. This energy has to be transferred to the atmosphere in order to prevent these components from overheating. Therefore, these components will be mounted on heat sinks, one for the primary circuit and one for the secondary circuit.

The eight IGBT switches per H-bridge should be reasonably spread out across the heat sink in order to distribute heat as evenly as possible. This arrangement will assure the best possible cooling effect – but it also might result into long commutation paths between individual IGBT switches and into disturbances due to the stray inductance which accompanies such a PCB arrangement. Therefore, the wiring between the IGBTs and the gating transformer should be as equal as possible.

A.4.4.4 Temperature measurements
In order to keep the temperature within certain limits, a number of NTCs, temperature variable resistors, will have to be placed in the design. As we have seen, the highest losses are reached in the H-bridges, the snubbers and the main transformer. Since the snubber MOSFETs are attached to the same heat sink as the IGBTs of the belonging H-bridge,

Therefore, NTCs will be attached to the primary heat sink, the secondary heat sink and on the windings of the main transformer. In order to keep the NTC current low, 10k types are used. The NTCs for the heat sink which can be screwed onto the heat sink, the NTC for the transformer will be attached onto the outer winding.

A.4.4.5 Forced cooling
The result of the temperature measurements on the heat sinks and the main transformer can not only used by the over temp protection circuit, it can also be used to control a fan. This fan provides a forced cooling and therefore enables higher power losses in the power circuitry and is chosen as large as mechanically possible: 90mm*90mm*25mm.
A.5 Change notification Snubber design

Version change effects:
Mass GI: 88000350

Reason of Change:
The primary and secondary snubber circuits dissipate too much energy when extreme over voltages are applied to the input. Adding a resistor between the gate of the snubber MOSFETs and the original gate connection of the snubber circuits will limit the snubber currents.

To be implemented as from when and where:
Date of release: : July 24th, 2006
Production Company : All new products
MV Ware House : All stocked products
MV Production Company : All stocked products
MV Service : All incoming products

Version numbers:
Mass GI: 88000350 Insert R200 and R201 between the source of respectively T14 and T15 and their original connection to the snubber circuit.
Changes Software:

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<th>Action Production</th>
</tr>
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<tbody>
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<td></td>
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Changes Hardware:

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<th>Changes</th>
<th>Reference</th>
<th>Action Production</th>
</tr>
</thead>
<tbody>
<tr>
<td>Addition of R200 en R201</td>
<td>R200, R201: 100Ω PRO2 resistor. Farnell art.nr.: 9484746</td>
<td>Change in connection of T14, T15 to the PCB.</td>
</tr>
</tbody>
</table>

Service modification description:

Insert R200 and R201 between the source of respectively T14 and T15 and their original connection to the negative side of the snubber circuits. Use the following addendum as a guideline.
Before connecting the two cooling frames with the IGBTs and the MOSFETs, the following adjustment has to be made:

**Modification of the resistor, as shown in figure 1**

![Figure 1. Left: Bending scheme of the 100 ΩRO2 resistor. Right: wire cut off point of resistor](image)

1. Start with a 100Ω PRO2 resistor.
2. Bend one of the pins 180 degrees. This wire will now be the upper side of the resistor.
3. Bend the other pin 90 degrees towards the down side of the resistor.
4. Cut off the upper pin at the end of the resistor. Don’t let it hang over the vertical lower pin!

**Modification of the MOSFET, as shown in figure 2**

![Figure 2. Bending scheme of the MOSFET IRFP460](image)

1. If already soldered on the PCB, remove MOSFETs T14 and T15, type IRFP460.
2. Holding the metal side of MOSFET T14 and T15 to the back, bend the right pin 90 degrees to the front.
Connection of the resistor to the MOSFET, as shown in figure 3

![Connection of resistor R200, R201 to resp. MOSFETs T14 and T15](image)

Connect the upper pin of the resistor to the bended pin of the MOSFET by means of soldering. Make sure the long pin of the resistor ends up at the same place as where the bended pin of the MOSFET used to be, in row with the other two MOSFET pins.

Connection of the MOSFET-resistor combination to the cooling frame

Make sure the MOSFET-resistor combination is mounted 2-3 mm higher than the IGBTs, to ensure that no mechanical pressure will be opposed to the MOSFET-resistor combination, while screwing the cooling frame to the PCB. The down side of the resistor has to be on the same level or higher than the downside of the cooling frame.

Kit addition, as shown in figure 4

![Kit addition, view after application of kit to T14, R200 (left photo) and T15, R201 (right photo)](image)

Freeze the added resistors to the PCB by putting a drop of kit between the end of the resistor and the PCB, as shown in figure 4.
A.6 Circuit modification for Kema approval

In order to get the Mass GI approved by the Kema, a number of adjustments still have to be made. Some of them are related to the electrical circuit, others to the terminals or case. For all tests, first of all the classification of this transformer needs to be clear. Therefore, these aspects will all shortly be discussed in this chapter.

A.6.1 Transformer classification

According to IEC standards, transformers can be divided into several classes. These classes differ in restrictions regarding electrical and mechanical layout, utilization and safety. The GI will be handled as a class 1 transformer, because of the presence of the primary and secondary Protective Earthing connectors, “PE GND input” and “PE GND output”.

The power electronic circuits however will be tested as class 2 equipment. Therefore, the design will have to accommodate reinforced isolation between the primary and secondary side. The placement of reinforced isolation in the GI is shown in the figure of addendum A2 with the blue line. Classification of the GI as a class 2 transformer would also require reinforced isolation at the secondary side, besides the absence of PE terminals.

A.6.2 Adjustment of current measurement transformers

At this moment, these transformers are constructed on a layer of epoxy and shielded with kit and plastic tubes. The minimum creeping distance of 4mm between the low voltage circuit and the output voltage can not be guaranteed. Therefore, a isolating support case, as shown in figure A7, has to be used instead of the currently used epoxy plate transformer, . In this new design, the secondary grid current will be flowing through the isolated red wires, which will guarantee a safe creeping distance. A possible replacement is the Elytone ET72502.

Another option is the usage of other current measurement transformers. A possible replacement is the Elytone ET72007, a fully isolated current transformer which is also used in the Mastervolt Combi. The output current can flow through an external wire, inserted through the isolated hole of the current transformer, thus enabling a safe creeping distance and clearance.
A.6.3 Adjustment of safety capacitors

The capacitors C48 and C49 are both type Y2, and therefore are valid for reinforced isolation. The capacitors C42, C43 between the line and neutral input and PE_Ship and C51 over TR4 have to be changed from one Y2 into 2 series connected Y2 or Y1.

Capacitors C31, C53 and C66 can remain type Y2, since no reinforced isolation is needed here.

A.6.3.1 Class X Capacitors

An "X" capacitor is used in Across-the-Line applications. Typically between the Hot and the Neutral power lines in North American applications and between L-1 and L-2 in European and other power line applications. In this position, a capacitor failure should not cause any electrical shock hazards, rather, a capacitor failure "between-the-lines" would usually cause a fuse or circuit breaker to open. The definition for "X" capacitor application is: "where damage to the capacitor will not lead to the danger of electrical shock". The X classification is divided into three sub-classifications, X1, X2, X3 as defined by IEC664 categories I, II III. Most commonly used are X1 (impulse tested to 4000 Volts) and X2 (tested to 2500 V).

A.6.3.2 Class Y Capacitors

A "Y" capacitor is used in Line-Bypass applications. Typically between the Hot line and Ground or Neutral and Ground in North American applications and between L-1 or L-2 and Ground in European and other power line applications. The definition for "Y" capacitor application is: "where damage to the capacitor may involve the danger of electrical shock".

The failure of a “line-to-ground” capacitor would not open any safety fuse. In other words, the failure of a line bypass capacitor could create a 230 volt “hot” chassis. The "Y" classification is divided into four sub-classifications, Y1, Y2, Y3, Y4 as defined by EN132400. Most commonly used are Y1 (tested to 8000 V) and Y2 (tested to 5000 V).

A.6.4 Adjustment of Optocoupler OC1

To ensure a safe creeping distance, optocoupler OC1 has to be changed into a long body type. The current type has a creepage distance of 6.4mm. The enlarged version of the TLP621 would be a proper option, as shown in table A-1.

<table>
<thead>
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<th>TLP621 type</th>
<th>7.62 mm pitch standard type</th>
<th>10.16 mm pitch (LF2) type</th>
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<tbody>
<tr>
<td>Creepage distance</td>
<td>6.4 mm (min.)</td>
<td>8.0 mm (min.)</td>
</tr>
<tr>
<td>Clearance</td>
<td>6.4 mm (min.)</td>
<td>8.0 mm (min.)</td>
</tr>
<tr>
<td>Insulation thickness</td>
<td>0.4 mm (min.)</td>
<td>0.4 mm (min.)</td>
</tr>
</tbody>
</table>

A.6.5 Heat sink and thermal conductive materials

For the primary side, silicone tubes will be used to electrically isolate the IGBTs and the MOSFET. Since the MOSFET placed close to one of the bolts, connecting the heat sink to
the case, the pins of this device have been replaced as shown in the drawing on page 5. Also, the heat sink will be placed on the power-PCB using 8mm instead of 4mm spacers. Drawings of these adjustments can be found at the end of this chapter. More information on the measurement results on thermal conductive materials can be found in addendum B6.

A.6.6 Case
Several general remarks have been made on the case:
- The fuse is not replaceable, therefore the identification can remain as it is
- The case bolt “PE” has to be removed
- The connections “PE GND” at both the input and the output side do not have to be renamed and can keep their yellow green terminal.

These remarks can easily be implemented.

A.6.7 Touch current
Due to the presence of capacitive coupling between the incoming grid voltage and the case, a small current could flow to the user if the equipment is touched. In order to stay well below safety standards, this touch current should always be less than 3,4mA. At a maximum input voltage of 325V, a maximum for the capacity between the grid and the case can be calculated:

\[
I_{\text{touch}}(t) = C_{\text{capacitive coupling}} \cdot \frac{dV_{\text{grid}}(t)}{dt}
\]

\[
I_{\text{touch max}} = 2 \cdot \pi \cdot f_{\text{grid}} \cdot C_{\text{capacitive coupling}} \cdot V_{\text{grid}}
\]

\[
C_{\text{capacitive coupling}} < \frac{I_{\text{touch}}}{2 \cdot \pi \cdot f_{\text{grid}} \cdot V_{\text{grid}}}
\]

As can be seen, any capacitive coupling should stay well below 33nF. This is important, when applying capacitors in order to reduce HF noise.

A.6.8 Inflammability requirements
The Mass G1 should be able to withstand internal fires, for example caused by short circuits or malicious components. Therefore, the following aspects should be taken into account:
- Officially, case holes are not allowed to exceed a diameter of 5mm. Sleeves can have an infinite length, as long as their width is smaller than 1mm.
- The front now contains a hole, through which the front PCB is connected. This hole is not shielded with 5V material. The case however is made of Vo material, the hole itself is covered by the front PCB, and between the electronical parts and the front also the power PCB and a layer of mylar foil is present. A possible solution might be the installation of a metal grating, with holes smaller than 5mm, which will fully cover the hole behind the front PCB.
- In order to protect the surroundings from dripping components, the ventilation sleeves on the left and right side of the case have to be inaccessible for material,
falling under a vertical angle of 0-5°. This means that the current ventilation sleeves officially will not satisfy. This however is not regarded as a big problem.

- The fan is regarded as a closure, according to Kema. Does this only count for rotating devices? The fan material is not regarded. The fan however now is possibly blowing flames and residue out of the case.

A.6.9 PE connection on the shore

In case of disconnection of the ship ground, for example caused by lifting the ship out of the water, the connection PE_Ship has to be connected to a shore Protective Earthing. This remark has to be implemented in the manual and in the report.

A.6.10 Identification and manual

Several additions have to be made to the product identification tag and the manual. The exact division of information between these two still has to be made. Changes and additions contain:

- Transformer logo: 

- Power factor: \( \cos(\phi) \) should be mentioned, according to Kema. This is not possible and dependable on the applied load.
- Power at output.
- IP21
- “GI3.5” in stead of “GI isolation transformer” on label
- Correct names of input and output on label: should be “PRIM” AND “SEC”
- Grid voltage symbol: 
- Maximum ambient temperature: \( T_a = 40°C \)

A.6.11 Redundancy of processors

At this moment, the internal processor has to be switched off during tests, due to safety reasons. Addition of a backup circuit with a second processor could solve this issue, since only 1 of the two processors has to be switched off for the test. Adequately implemented redundancy circuits could then take over the temperature, voltage and current controls. This option however is quite drastical, especially because of the fact that there is no official guideline yet on the usagae of software enabled safety circuits.
A.6.12 Creeping distances for the new heat sink design
A.7 Design and application of the Thermal Comparator

In order to be able to accurately measure losses in switching devices, a Thermal Comparator has been developed. The goal of the Thermal Comparator is to measure and compare the temperature in two equivalent setups of a heat sink and one or two semiconductors. On one heat sink, the switching devices, such as a GI switch, consisting of two IGBTs, or a GI snubber MOSFET, will be connected. Since these semiconductors show HF behavior, accurate power loss measurements sometimes are quite difficult.

Therefore, on the reference heat sink, one or two MOSFETs are connected, which are series connected and conducting a DC-current. Of each MOSFET, the drain and the gate are connected. As a result, a current source can be connected to the MOSFETs, which will cause the MOSFETs to heat up. In order to get the same heat dissipation in both heat sinks, the reference MOSFETs are chosen to have the same casing as the semiconductors, which will be tested: TO247.

Main task of the Thermal Comparator now is to measure both heat sink temperatures and to signal in case of temperature difference and over heating. Temperature differences will be caused by a difference in power loss between the measurement setup and the reference setup. Therefore, the accurateness of the thermal comparator will have to be very high. However, a certain “equality-state” should also be present. This can mean that, in case of a temperature difference of 0.5°C or less, the Thermal Comparator will give a “Temp Equal” sign.

An overheating protection has to be implemented in order to protect the semiconductors at both the measurement as the reference heat sink and should give an alarm at 90°C. Both heat sinks should be thermally protected separately.

In order to also be able to handle to high power losses, up to 100W per heat sink, forced cooling is necessary. Therefore, a fan is connected to both heat sinks. The average heat sink temperature, at which the fan will turn on, can be controlled by a potentiometer. This temperature has to be as high as possible, in order to have a high temperature difference with the surrounding air. The higher this difference, the more accurate the measurement can be.

The total schematics are given in figure A8. As can be seen, the thermal measurements are done by NTCs, calibrated by VR1 and VR2. The comparators IC1a and IC1b compare the output voltages of the NTC and activate the LEDs “TEMP2>TEMP1” and “TEMP1>TEMP2”. When both activated, however, the circuit around IC1d and Q4 will turn off these LEDs and will activate the LED “TEMP EQUAL”.

The individual over heating protection is done by the comparators IC 2a and IC2b, which activate LEDs “OVERTEMP 1” and “OVERTEMP 2”. The temperature, for which this warning will turn on is controlled by VR3. Since this value is chosen equally for both heat sinks, the output of VR3 can be used for both overheating comparators.

The fan control is done by comparator IC1c and can be controlled by turning VR4. Since this is a variable value, this variable resistor is implemented as a potentiometer with axis, in order to easily adjust the threshold value.
Fig. A.8: Circuit design of the Thermal comparator.

Addendum

Mastervolt
Technische Universiteit Eindhoven
Title: Thermal Comparator
Version: 7
Date: 8/14/2006
Drawn By: Elanne Thewissen
Furthermore, the input voltage is maintained at 5V by voltage regulator IC3. Not only will this ensure an accurate output for all voltage divisions, it also enables the input voltage to vary between 6V and 30V. This is very important, since this way, the 24V fan can be connected at the same input voltage. The fan speed can then be controlled by controlling the input voltage, thus still enabling a high temperature difference when turned on, resulting in a high accuracy.

A realization of the Thermal Comparator is shown in the pictures in figure A9. The signal LEDs are clearly visible, as also the heat sinks and fan. In order to save space, a sandwich construction has been used, as shown in figure A9.c. The lower PCB contains the control circuit; the upper PCB mostly contains the LEDs and a number of transistor drivers.

All variable resistors have been placed in such a way, that they are easily accessible with a screwdriver, which makes them easily adjustable.

The semiconductors, which are tested, can be connected by clamps onto the heat sink, as shown in figure A9.b. They are isolated electrically from the heat sink by a ceramic heat conducting layer. The reference MOSFETs are connected to the reference heat sink in the same way.
ADDENDUM

B

Measurement results
B.1 Measurement equipment and setups

One of the best things of power electronics is not only making a design, but also testing it. Working on the edge of technological possibilities means that theoretical designs will not always be practically applicable, for example due to unexpected thermal constrains or generation of HF noise.

Therefore, a wide range of tests have been performed, not only on the non-scalable 230V and 120V versions and the scalable DAB and boost converters, but also on several components and sub circuits.

This chapter will first present the used equipment and measurement setups and will subsequently show the test results of the different project details.

B.1.1 Variable input voltage setup

For the setups, at which the direct grid connection could not deliver the necessary voltage, several types of setups have been used. First of all, a 10A variac of FILEC with a variable output voltage of 0-260V has been used for variable grid voltage and 120V grid voltage measurements at low load currents.

In order to test the 120V Mass GI at full load, two external switchable 50Hz isolation transformers, types Mastervolt IVET 3.5 and IVET D 3.5kVA/16A, have been used. This way, with one of them, the 230V grid voltage could be transformed into a 115V grid voltage, capable of delivering 32A. At the output of the 120V Mass GI, an isolation transformer had then been switched 115:230 again, resulting in a normal supply voltage for the 230V loads again.

The isolation transformers could also be used as a 1:1 transformer, thus electrically isolating the Mass GI. This was useful for a number of measurements with normal instead of differential oscilloscope probes, as normal probes have a ground connection, which could cause a short circuit when applied wrongly.

For exact measurements at high load currents, also a California Instruments 5kVA AC power source, model 5001 ix was available. This power source could make any AC or DC voltage between 0 and 420V, with frequencies up to 1kHz. Since this power source could not only control the output power, but also the output current, tests were simple with this device. Simply choosing a low value for the maximum current effectively has saved several setups, which actually would have broken down, in case of a direct connection to the grid.

Further more, also several Delta power supplies have been used, when a certain DC voltage or current was needed. For general test setups, a SM7020-D, 70V and 20A max, was used, as this was installed on my work bench. For high voltage DC measurements in for example the IGBT measurements, a SM300-10-D, 320V and 10A max, was used. For the measurements regarding power loss and thermal behavior of the 120V main transformers, a SM15-200-D, 15V and 200A max, has been used.

B.1.2 Variable load measurement setup

In order to be able to test the GI under different load conditions, the following test setup has been used, as shown in figure B1.b. The used laboratory setup is equipped with a switchable load, shown in figure B1.a. This load consists of several light bulbs: 1×40W,
Addendum

**1.60W, 1*100W, 1*200W, 1*500W, 1*1kW and 6*2kW.** The light bulbs are situated on the roof of one of the test rooms, so the cables to these loads are quite long. Also, the grid voltage drops at high loads, due to the same problem of cable length. Therefore, the theoretical load values will normally not fully be met. However, a creative use of the endless combination possibilities will eventually always give a correct load current.

---

![Measurement setups](image)

**Fig. B.1. Measurement setups. a) Switchable variable load, b) Efficiency measurement setup**

### B.1.3 LF measurements

For RMS measurements of grid or load currents or grid or load voltages, several types of multimeters have been used. Since the internal shunt of the used multimeters only allows a maximum RMS current of 10A, in a number of situations an external shunt of 0.4mΩ has been used. Since the measured voltages were so low, for example 12mV at 30A, these current measurements could only serve as a guide line, not as an official measurement.

Also diode tests, for example when receiving a broken Mass GI, which needed to be checked for further tests, have been performed with these devices.

### B.1.4 HF measurements

For the HF measurements on the electronic circuits, a four channel digital oscilloscope of Yokogawa, type DL1640, has been used, together with standard 1:1 or 1:10 probes and Le Croy differential probes, switchable between 1:10 and 1:100.

For current measurements, two different kinds of current probes have been used. First of all, a current probe together with the Tektronix amplifier, type TCPA 4000, has been used for measurements from DC up to 17MHz. Secondly,

### B.1.5 Efficiency measurement setup

In order to be able to measure the efficiency of a certain Mass GI setup, a Yokogawa PZ4000 Power Analyzer was used. The device is capable of calculating amplitudes and phase angles of not only voltage and current, but also of real, reactive and total power. Further it can calculate efficiencies, a very useful function for the current setup.

The power analyzer can measure input voltages up to 600V DC and has two internal shunts per channel, capable of a maximum DC current of respectively 5A and 20A. In order to measure higher currents in the 120V version, 5:1 current transformers, with a maximum rating of 60A are used.
For the wires of the current measurements, a thickness of \(4\text{mm}^2\) is chosen. This way, a minimum voltage drop over the wires is realized. The wiring for the voltage measurements will not see any high currents and can therefore be much thinner. However, since it should still meet isolation standards, \(0.75\text{mm}^2\) wires are used. In order to disable any measurement errors, due to wiring resistance, the voltage measurement wires should directly be connected to the Mass GI, not to any of the connections of the current measurement of the power analyzer.

**B.1.6 Thermal measurements**

For temperature measurements, several devices have been used. First of all, thermocouples have been used, together with a Fluke 51 K/J thermometer. Thermocouples are a widely used type of temperature sensor and can also be used as a means to convert a thermal potential difference into an electric potential difference. Each combination of two different metals or alloys will result in a different potential difference. Standard thermocouples are J-type, made of iron and constantan, and K-type, made of chromel and alumel. One of the differences between these types is the temperature range for which they can be used: J-type thermocouples theoretically can be used between \(-40^\circ\text{C}\) and \(750^\circ\text{C}\), whereas K-type thermocouples can be used between \(-200^\circ\text{C}\) and \(1300^\circ\text{C}\). Therefore, K-type thermocouples are more common and are chosen for all “on the spot” thermal measurements.

For thermal management of complete circuit designs and transformer measurements, also a thermal camera of FLIR, type ThermaCAM™ E45, has been used. With this camera, with a sensor of 320 by 240 pixels, one can easily find thermal hotspots on PCBs and in components. This way, design flaws can be found before a circuit is destroyed by overheating.
B.2 Measurement results non scalable 230V DAB design

In order to get a correct image of the working principle of the non scalable 230V Mass GI, several tests have been performed. Not only load currents, but also input voltages have been varied, in order to compare the calculated results of the loss mechanisms with actual measurements.

B.2.1 Gating circuit

As figure B2.a shows, the current gating signal has a slow rise time, with a maximum rise time of 6e7 V/s. While the diode conducts, the IGBT module acts as a capacitor, which is loaded. During this time, the IGBT impedance increases or decreases, depending on the applied gating signal.

![Gate signal CH1 and CH2 and resulting collector-emitter voltages CH3 and CH4](image1)

![Input snubber current CH1 (0.1A/div) and snubber voltage CH2 (100V/div)](image2)

Fig. B.2. a): Gate signal CH1 and CH2 and resulting collector-emitter voltages CH3 and CH4(a). b): Input snubber current CH1 (0.1A/div) and snubber voltage CH2 (100V/div).

B.2.2 Snubber activity

As mentioned previously, the new snubber circuit contains a power limiter. To limit the current, flowing through the MOSFET, a 100Ω resistor has been put in series with the source in both snubber circuits, as presented in chapter 4. In order to test the new setup, a 420V pulsating signal, produced by the AC power source, has been fed into a 220:240 transformer, further increasing the peak level to 458V.

Measurement results on the working aspect of this adjustment are given in figure B2.b. The green signal shows the voltage over the snubber capacitor. The yellow signal shows the voltage over the added source resistor. Every time when it reaches 14V, the current through the MOSFET is being limited to the maximum value of 160mA.

B.2.3 Variation of input voltage and output current

In order to see the influence of varying input voltages on the efficiency of the Mass GI, two voltages have been supplied to the GI, besides the standard 230V: 210V, which is almost 9% lower than average, and 240V, with an over voltage of 4.3%.

As can be seen in figure B3.a, the voltage drop between input and output voltage is relatively equal. Therefore, the efficiency at low input voltages automatically is lower. Interesting is the fact that the efficiency for 240V also is lower than the 230V graph, shown in graph B3.a. This however is caused by the fact that at this time, switching losses increase and therefore put a mark on the overall efficiency.
B.2.4 Total power loss

When measuring input and output voltages and currents, also a total power loss over the GI can be calculated. This is done again for various voltages, as shown in figure B4. However, not only the total power loss is shown: In order to get an image on the power loss, not caused by resistive losses, the total loss has also been compensated. This has been done by subtracting the product of the voltage drop and the output current from the total power loss. Now, this voltage drop is compensated and only the general losses, caused by the 12V and 5V supply, the filter capacitors and part of the switching losses.

Looking at the data in figure B4, this supplementary loss is only low, with an average of 50-60W at full load, which is less than 2%. Reduction of conduction loss by choosing other IGBTs with lower losses would therefore drastically decrease the total power loss.

B.2.5 Implementation of dead time

Due to the fact that small peak currents could be measured in each leg of the primary H-bridge at the switching moments, there had to be shoot through, leading to a temporary short circuit in the H-bridge. This will not only lead to high component stress, but will also cause HF noise. Therefore, a dead time has been implemented in one of the setups of the Mass GI. In order to see the difference, the delay has been altered for various loads, as shown in figure B5.
Applying a loss calculation to the various measurements, it was also possible to determine the decrease in power loss, caused by the dead time circuit. Since it is a decrease in loss power, the values in figure B5 are negative. As can be seen, an up to almost 12W can be saved with the appropriate dead time level. However, it can also be seen that a further increase in dead time will lead to increased losses. This is due to the fact that the IGBTs and main transformer will have to transfer the same amount of power in less time, therefore causing higher conduction losses in these devices due to the higher RMS current value.

![Graph showing efficiency and reduction in power loss due to implementation of dead time.](image)

In order to keep the total loss reduction equal over the whole load area, this graph shows that a delay of 0.6μs would have to be implemented.

### B.2.6 Implementation of IGBT type IHW40N60T

After the evaluation of various IGBT types, as shown in addendum E4, the IHW40N60T also seemed to have quite interesting characteristics. Theoretically, this IGBT could have good characteristics, especially due to the low conduction losses of the device.

Looking at the test results in addendum B2, it is clear that an additional dead time circuit is necessary, in order to prevent shoot through. This could be caused by the fact that the turn off delay time and the fall time at high loads are much higher for the IHW40N60T than for the HGTG20N60B3D.

The effect of these differences in characteristics can be seen in the graphs of figure B6, where, after addition of a relatively large dead time of 1.2ns, a variation is visible in the behavior of both IGBT types. The voltage over the main transformer shows only a small difference over the output power range, when the HGTG20N60B3D is used. Implementation of the IHW40N60T however leads, at high output power levels, to an increased time on which the main transformer is still being fed by the IGBTs. Only after 0.6μs, this voltage slowly drops, meaning that the main transformer is not powered by the turned off IGBTs any more. Only then, a safe transition between the H-bridge switches can be ensured.
Fig. B.6. Influence of dead time while using the HGTG20N60B3D (up) and the IHW40N60 (down) at various loads. Left: 1kW, middle: 2kW, right: 3kW. CH1 = $V_{GE}$ (2V/div), CH2 = $V_{P\text{main transformer}}$ (50V/div)

### B.2.7 Surge test results

In order to see if the current design was able to handle surges on the incoming grid, a number of tests have been done on the 230V Mass GI.

<table>
<thead>
<tr>
<th>Test no.</th>
<th>$V_{surge}$</th>
<th>Test model</th>
<th>Result</th>
</tr>
</thead>
<tbody>
<tr>
<td>Test 1</td>
<td>1kV</td>
<td>Standard model</td>
<td>Short circuit in device, caused by punch through diode bridge of the 12V power supply. No damage to Power Electronics or gating circuits.</td>
</tr>
<tr>
<td>Test 2</td>
<td>1kV</td>
<td>Standard model</td>
<td>Short circuit in device, caused by punch through diode bridge of the 12V power supply. No damage to Power Electronics or gating circuits.</td>
</tr>
<tr>
<td>Test 3</td>
<td>1kV</td>
<td>Adjusted model: diode bridge replaced by four 1N4007 diodes</td>
<td>No damage</td>
</tr>
<tr>
<td>Test 4</td>
<td>1kV</td>
<td>Adjusted model: 12V power supply disabled, replaced by external 12V power supply</td>
<td>No damage</td>
</tr>
<tr>
<td>Test 5</td>
<td>1kV</td>
<td>See previous test</td>
<td>No damage</td>
</tr>
<tr>
<td>Test 6</td>
<td>2kV</td>
<td>See previous test</td>
<td>No damage</td>
</tr>
<tr>
<td>Test 7</td>
<td>2kV</td>
<td>See previous test</td>
<td>No damage</td>
</tr>
<tr>
<td>Test 8</td>
<td>2kV</td>
<td>See previous test</td>
<td>No damage</td>
</tr>
<tr>
<td>Test 9</td>
<td>2kV</td>
<td>See previous test</td>
<td>No damage</td>
</tr>
</tbody>
</table>

The tests have been performed on the 21st of July 2006, at Betronic, the company, which manufactures the Mass GI, as the equipment for this kind of tests is not standard. The surge pulses, applied to the Mass GI were in confirmation with the standard shape for surges, according to IEC 61000-4-5. The surges had peak values of 1kV and 2kV, a rise time of 1.2μs and a fall time of 50μs.
The results are shown in table B-I. It is clear that the 12V power supply causes problems several times. At first, the diode bridge was thought to be the problem, as this could be not able to transfer peak currents. After insertion of new diodes, the problem however stayed, thus placing the problem in the 12V power supply IC.

After removal of power to the 12V power supply and feeding the 12V circuits with an external 12V power supply, the Mass GI remained working, as expected, even when several 2kV tests were performed with only 30seconds in between. Therefore, no problems are now expected for any future surge tests, besides the 12V power supply.

While performing surge tests at 1kV, the 12V power supply short circuited, causing a breakdown of two diodes of the diode bridge. To prevent such damage in the future, several options have been investigated. The following paragraphs present three possible solutions to this problem

B.2.7.1 Insertion of a filter inductor
It is possible to filter out voltage peaks by inserting a filter inductor between the DC bridge and the circuit around IC7, the TOP244Y. Also, a small capacitor has to be put directly over the output of the bridge. Small peaks thus will be damped.

B.2.7.2 Insertion of a series resistor
Another possibility is the implementation of a small resistor, for example 47Ω, in series with the AC side of the bridge. This will effectively lower the peak current at a surge. During normal conduction mode, it will cause a voltage drop of 4.7V at an AC current of 0.1A, thus causing an extra power loss of 0.47W and increasing the minimum AC input voltage from 80 to 85V. In order to be able to withstand not only the continuous, but also the peak power, a 1W resistor is recommended.

B.2.7.3 Tuning of the over voltage protection
Another option of controlling over voltages implies tuning the input resistors on the line sense pin. This pin measures the input voltage by measuring the current, flowing into pin 2, the Line Sense pin, of this IC. Under voltage is sensed at a current of 50µA, Over voltage at a current of 225µA, flowing through this pin. This current is equal to the DC voltage, divided by the total value of resistors R99 and R100. Currently, this value is 2MΩ. Therefore, the TOP244Y will only activate the 12V circuit at an input voltage of 80V and will turn it off at a DC input voltage of 450V. To lower this voltage to 400V, the resistors will have to be given a value of 1.778MΩ. Possible implementations are a parallel connection of two resistors, of 3MΩ and 3MΩ, resulting in 1.7875MΩ, or a parallel connection of two resistors of 3MΩ and 3MΩ, put in series with a resistor of 82kΩ, resulting in a resistance of 1.7777MΩ. Of course, high precision resistors have to be used to obtain this precision.
B.3 Measurement results 120V push pull design

Also the 120V Mass GI has been tested on for example efficiency and input voltage dependency. Since the gating circuits and therefore the gating signals did not drastically change, the measurements on these signals have not been added.

B.3.1 Gating circuit

The first measurements on the 120V push pull converter design resulted in a high HF disturbance, or in other words: the laboratory radio suddenly only produced white noise instead of music. This meant that the switches of the converter were switching on too fast, leaving the IGBTs, which were switching off, too little time to actually switch off. Therefore, a dead time of 0.5us was implemented.

Since the gating has been adjusted, precise and equal timing is very important. Otherwise, one half of the IGBTs might switch on first, resulting in the fact that it has to take the full 30A by itself. However, due to the relative symmetrical layout of the PCB, symmetry appeared not a big issue.

B.3.2 Thermal behavior of various transformer designs.

Since in chapter three, several transformer designs have been calculated, it was very interesting to also build them. In order to test the transformers in the same conditions, a wind tunnel was built, in which the three series connected transformers were cooled by three transformers, as shown in figure B7.a. Of each transformer, all coils had also been connected in series, in order to get a realistic power loss and maximum voltage drop per transformer.

![Measurement setup and loss results of various 120V push pull transformer designs.](image)

The measured power losses, as shown in figure B7.b, are slightly higher than expected. This is mainly caused by the connecting wires. Due to the fact that the dual E65 core windings have been connected with four 1mm² wires per connection, the differences between the theoretical and the measured power loss is only small. The single E80 core, connected with litz wires, shows a high difference, caused by the high loss in the connecting wires.

Also interesting is the difference between internal and surface temperature between the various transformer designs at 40A DC. These measurements clearly show the fact that the current E80 core design will not be usable. The measured values are shown in table B-2.
Table B-2. Internal and surface temperatures of various transformer designs at 40A DC.

<table>
<thead>
<tr>
<th></th>
<th>E65</th>
<th>Dual E65</th>
<th>E80</th>
</tr>
</thead>
<tbody>
<tr>
<td>Internal temp.</td>
<td>158.5°C</td>
<td>89.2°C</td>
<td>160.4°C</td>
</tr>
<tr>
<td>Surface temp.</td>
<td>103.8°C</td>
<td>51.1°C</td>
<td>81.3°C</td>
</tr>
</tbody>
</table>

When looking at the results, it is clear that only the dual E65 core can remain within limits for both the internal and the surface temperature. The single E65 core, shown in will overheat, but will also trip the over temperature alarm. An important conclusion, when looking at these measurement results, is the fact that the E80 however will not trip the over temperature alarm, as this will only measure the surface temperature. The internal temperature will at the same time exceed maximum limits, for example the limit of 125°C for the tape. This can lead to dangerous situations.

These results can also be seen in the thermal images in figure B8. Here, a 16A load has been fed for 1 minute, in order to get the highest temperature differences between the heat producing and the heat conducting components. The main transformer does not yet reach over temperatures, but already shows the highest component temperature.

Fig. B.8. ThermaCAM images of the 120V Mass GI. a) Sideview, b) Upper view.

A number of other transformer configurations also have been tested. Pictures of a number of these setups are shown in figure B9.a, B9.b and B9.c. The first and last picture are taken, when the converter was actually converting power. Photo B9.c even shows the surface temperature of the dual E65 core transformer at full load, being less than 80°C.

Figure B9.b also shows the dead time circuit, hanging in front of the converter. The blue block is the potentiometer, with which the applied dead time can be controlled.

Fig. B.9. Various test setups of the 120V converter. a) E80 core configuration, b) two E65 configuration, c) Measurement setup with dual E65 core configuration.
As can be seen in the presented graphs in figure B.10, the single E65 and single E80 core designs are not capable of handling the loss power for high load currents. Therefore, the dual E65 core design has been used for further tests.

**B.3.3 Variation of input voltage and load with a dual E65 core**

Interesting is the voltage drop for various output currents. As is shown in figure B.11, this is equal for various values of the grid voltage. This is logical, as this parameter is caused by the main transformer and the IGBT and diode conduction losses. These are for a great deal only dependant on the output current.

As also can be seen in figure B.10, the efficiency is lower for the 120V push pull design, compared to the 230V DAB topology. However, this efficiency still increases for higher values of $V_{in}$.

**B.3.4 Total power loss with a dual E65 core**

Also for the 120V, dual E65 core Mass GI the total power loss and supplementary power loss can be calculated after efficiency measurements, as explained in paragraph B.2.4. This is done again for various voltages, as shown in figure B.12.
Fig. B.12. Total loss power and supplementary loss power at various load currents and input voltages for the 120V dual E65 core push pull converter

Compared to the results of the 230V DAB topology, the total power loss is equal. However, the supplementary power loss is lower. This can be explained by the facts that for example capacitor banks are loaded to a lower voltage and the 12V power supply now also can use a higher, more efficient conduction time for the internal MOSFET.
B.4 Measurement results 230V DAB layout with phase shift

The 230 DAB layout, using phase shift control, has been built and tested in various setups. At first, the DSP control board has been programmed and tested. Subsequently, the DSP has been connected to an adjusted Mass GI model. Since the setup was very unstable, creating high amounts of HF noise, the number of tests was limited.

B.4.1 Gate signal generation

As a first step, the gating transformers, already present in the MASS GI have been connected to the gating drivers, driven by the DSP. This way, DSP control of the MASS GI can be tested under standard conditions, at a transfer rate of 1:1. Furthermore, also a phase shift between the primary and secondary bridge can be implemented.

Since the driver does not only contain a MOSFET driver, but also two in cascade connected MOSFETs, a delay between the input and the output is expected and present, as shown in figure B13.

Since the gating drivers will be implemented in all four legs, the delay will affect all gating signals equally. Only possible problem might be a difference in delay between the rising and the descending flanks. If a difference is present, this will affect the current dead time settings.

As can be seen in the left two images of figure B13, the rising and the falling edges of the DSP outputs 1 and 2 are quite similar.

The result of a negative delay of only 1 step is shown in the right two graphs of image B14. Clearly visible is a delay between the rising and falling edges of PWM output 1 and PWM output 2 of 10ns. This is logical, as this is caused by the 100MHz clock frequency.
B.4.2 Variation of phase shift

The result of varying loads at varying phase shifts is shown in figure B15. As can be seen, the voltage and current signals partially show the expected behavior. However, the currents at a positive phase shift have a too high dv/dt. This is caused by a high voltage drop over the inductor. As a result, the output voltage cannot reach high values.

As a result, it is clear that the phase shift controlled DAB is only implementable for low output currents. High load values will automatically result in a decrease of the output voltage. The theory, as presented in chapter 2, can be found back in the graphs of figure B15, thus disabling the phase shifted DAB for the current application.
B.5 Measurement results 230V DAB boost converter

Also the DAB boost converter has been built, as shown in figure B16. The DSP is mounted on top of the device. The output signals of four PWM channels of the DSP are fed to four gate drivers, mounted in a 2 by 2 matrix between the DSP and the gating transformers. This way, the path between the DSP and the gating circuits is kept as short as possible.

The boost switch is mounted on the added heat sink and the boost inductor can be placed next to the heat sink of the boost switch.

Fig. B.16. Side view shots of the 230V DAB boost converter.

The output of the DSP, once it was working, is shown in figure B17. The switching patterns in the switching scheme of addendum D3 can also be found in the measured signals. A delay of 300ns is built in between deactivation of the secondary bridge and the activation of the boost switch. This will prevent a possible short circuit of the secondary side.

These first gating signals looked promising, and also the converter hardware was ready to be used. Due to indefinable problems with the initialization of the DSP, the gating signals, as shown above, could not be instigated any more. Therefore, due to lack of time to find out where the exact problem was, the boost converter could unfortunately not further be tested.
B.6 Thermal loss measurements

Component losses can not only electrically be measured; they can also be measured by building a reference inverter. To be able to simulate the same thermal conditions in this heat model, a PCB of the GI has been adjusted. In this inverter, the components, responsible for the main power loss, will be replaced by other components or connected differently, in order to get the same heat flow as in the original model. In this case, these components are the IGBTs and the main transformer. Also, a thermal comparator will be used to measure individual losses of IGBT pairs or snubber MOSFETs.

B.6.1 IGBT bridge dissipation model

In order to be able to accurately estimate the biggest losses, which are in the IGBT bridges, the GI heat model is equipped with a double MOSFET bridge as replacement for the 2 x 8 IGBTs. Connecting the Gate to the Drain of each MOSFET ensures that it will operate in saturation mode, as a MOSFET goes into saturation mode for $V_{GS} > V_{DS} - V_T$, with $V_T$ as the threshold voltage of the device. Since in this case $V_{GS} = V_{DS}$, the MOSFET is always in saturation mode.

All MOSFETs are then connected in series and connected to a current source, as shown in figure B15.a. Since the current source will raise the voltage, to enable the set current to flow through the MOSFETs, each individual MOSFET value of $V_{GS}$ will rise until it reaches the value, for which the set current can flow. This results in a predicted value of the voltage over the series connected MOSFETs of $16 \times 4V/\text{cell} = 64V$.

One of the requirements, set to the MOSFET characteristics, is the package: this has to be a TO247 package. Another demand regards the maximum current: in theory, one would want to be able to have a maximum current flow of 55A, in order to simulate full dissipation ($16 \times 4V \times 55A = 3520W$). However, this situation will never be reached; therefore a total dissipation of 80W per MOSFET is more than enough. This is also the maximum average power dissipation, a TO247 package can handle. This means that the MOSFETs will have to be able to handle a maximum current of 20A.

Furthermore, this also sets demands to the power supply source. Taking conduction losses and differences in characteristics into account, the maximum output voltage will have to be 80V, at a current of 20A, in case of a series connection of all MOSFETs. The
Addendum

voltage level can be reduced to 50V, when the MOSFETs are divided into two parallel branches of 8 series connected devices, as shown in figure B18.a. The maximum current value however then will double.

Taking all considerations into account, the type IRFP460 has been chosen. This device is not only chosen because of the TO247 body, but also because of the fact that for future use, it is possible to connect two mounted heat sinks in parallel, due to the positive temperature coefficient of the IRFP460. As a result, the total incoming current will be equally divided over both heat sinks, thus creating an equal spread of the dissipated power.

A simulation of the circuit of figure B18.a is shown in the graph of figure B18.b. In figure B15.b, simulation results with the voltage on the Gate-Drain node of each MOSFET is shown. As expected, the supply voltage is equally shared over the eight MOSFETs, e.g. a total input value of 32V results in a voltage of 4V over each device.

The continuous current, flowing through the IRFP460, is limited at 20A at 25°C, and 13A at 100°C, according to the datasheet of International Rectifier. Combining the Drain current and the Source Drain voltage gives the power per MOSFET. This is also shown in figure B17.b.

According to the model, the value of 13.0A is reached at a voltage of 4.90V per device, thus a total voltage of 39.2V. As a result, the maximum power dissipation per device at 100°C is equal to 64W. The value of 20.0A is reached at an input voltage of 5.30V per device, thus a total voltage of 42.4V. At 25°C the maximum power dissipation per device is equal to 106W.

The maximum current is not the only limiting factor of the MOSFETs. Looking at the maximum power rating for the IRFP460, the datasheet provides a maximum value of 280W at 25°C and a derating factor of 2.2W/°C. This results in a maximum power rating at 100°C of 280-2.2*75=115W. As a result, this parameter will not be taken into account, as it exceeds the calculated maximum values, calculated above.

As a result, the maximum power, which can be fed to each heat sink is equal to 8*80W=640W at 100°C and 8*106=848W at 25°C. At both temperatures the maximum possible power dissipation is high enough to ensure an adequate simulation of the thermal loss in the IGBT bridges. It is clear that a power load of 80W per device is very well obtainable. An implementation is shown in figure B19.a. Here, the eight MOSFETs are clearly shown, attached to the heat sink with clamps via a ceramic layer. Figure B19.b shows the yellow isolation tape, which is electrically isolating the MOSFETs from the PCB.
B.6.2 Transformer dissipation model

Also the main transformer is put into the main heat model in order to get an accurate image of the losses. As calculated before, the main transformer has a DC resistance of 0.22Ω. The maximum dissipation in this device is estimated at 20W, therefore the applicable current will be estimated at 9.5A. A basic DC electrical circuit of the main transformer is given in figure B20.a, the simulation results are given in figure B20.b.

![Transformer dissipation model](image)

As can be seen, the currents through the inductor are quite high. In order to have low losses outside the transformer, the wiring therefore needs to be made of thick and short copper wires.

An implementation is shown in figure B19.c, where the terminals from the transformer are connected via 6mm² wires to a Delta current source.
B.7 Variation of heat sink insulation material

In the standard model of the GI aluminum oxide plates are used to transfer the heat between the IGBTs and MOSFETs and the heat sink and, at the same time, electrically isolate these devices. In order to meet certain requirements regarding isolation distances, also other materials have been looked at. Eventually, thermally conductive silicone caps have been chosen as new insulation material. However, this material differs in several ways from the aluminum oxide plates.

Most important factor of the thermally conductive material is the thermal conductance. This variable can be calculated with formula B6.1:

$$\Delta T = \frac{P_{\text{heat}} \cdot d}{A \cdot G_{th}} = \frac{W}{A \cdot m \cdot ^\circ K}$$  \hspace{1cm} (B6.1)

With $P_{\text{heat}}$ as the total thermal power, to be transferred, $d$ as the thickness of the thermally conductive material, $A$ as the surface, over which the heat transfer is taking place and $\Delta T$ as the temperature difference between the surfaces, which are transferring the heat from and to external devices. The lower this temperature difference, the better it is for the circuit. Not only will this often improve the semiconductor losses, it also enables the semiconductor to be used at higher power ratings.

B.7.1 Aluminum Oxide thick film substrates

The thermal conductor, mostly used in power electronics, is Aluminum Oxide, also referred to as ceramic layer. This material is preferred because of its high value for thermal conductance, which results in a low temperature difference between semiconductor and heat sink. Several types of aluminum oxide thick film are available on the market, often differing in alumina mass content and thickness. The higher the first alumina content, the higher the thermal conductance value. For electrical insulation requirements, often a minimum thickness is required. Typical values for a number of characteristics are given in table B-3.

One of the negative aspects of ceramics is the practical implementation. The material is not easily shapeable and therefore less well suited for complicated semiconductor or heat sink surfaces. Bending the surface will cause a break. Also, a heat conducting paste is necessary for proper heat transfer between the ceramic plate and the semiconductor or heat sink.

<table>
<thead>
<tr>
<th>Material constant</th>
<th>Al.Oxide ADS-90R</th>
<th>Al.Oxide ADS-96R</th>
<th>Silicone Cap SC-08</th>
<th>Silicone tube</th>
</tr>
</thead>
<tbody>
<tr>
<td>Alumina content (mass %)</td>
<td>91</td>
<td>96</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Thickness (mm)</td>
<td>0.6</td>
<td>0.6</td>
<td>0.45</td>
<td>0.5</td>
</tr>
<tr>
<td>Thermal conductivity (W/mK)</td>
<td>12</td>
<td>20</td>
<td>0.8</td>
<td>0.7</td>
</tr>
<tr>
<td>Rel. thermal resistance ($K \cdot cm^2/W$)</td>
<td>0.5</td>
<td>0.3</td>
<td>5.6</td>
<td>7.1</td>
</tr>
<tr>
<td>Breakdown voltage (kV)</td>
<td>13.4</td>
<td>14.8</td>
<td>5.0</td>
<td>8.0</td>
</tr>
</tbody>
</table>
**B.7.2 Thermally conductive silicone caps and tubes**

Since in the current application the demands for creeping distances couldn't be met, new materials were searched for. It was found in thermally conductive caps and tubes, made from silicone rubber.

Silicones are a chemical substance which can vary in consistency from liquid to gel to rubber to hard plastic. Advantages of silicones are the fact that they are flexible, water resistant, chemical resistant, oxidation resistant, have a high dielectric strength and stable at high and low temperatures. This means that they will not melt or change shape easily at high temperatures or crack at low temperatures. Due to the rubber softness, it is even ideal as an interface between uneven surfaces, as long as a certain amount of compression forces are set on the connection. This even reduces the need of silicone grease.

Disadvantages of silicones include poor bonding, combustibility, which means that the material, once ignited, is hard to extinguish and significant smoke development. However, the biggest disadvantage is the fact that the thermal conductivity is relatively low. Compared to the two ceramic substances, the temperature difference at the same thermal energy flow will theoretically be 10 to 25 times be as high. This causes the semiconductors to heat up even more and possibly even reach critical junction and die temperature. Also, the efficiency of the IGBTs and MOSFETs will drop down further more.

In order to see the difference, a number of measurements have been performed for both silicone tubes as the standard aluminum Oxide substrates. The thermal model of figure B19.a has been used to perform the tests. The results are shown in figure B21. As can be seen, the temperature difference, when using silicone tubes for the TO247 MOSFETs, is higher an therefore can result in higher losses for the IGBT.

![Fig. B.21. Temperature difference between heat sink and IGBT case for electrical insulation by silicone tubes and aluminum oxide thick film substrates.](image)

A practical implementation of the tubes is shown in figure B22.a. As can be seen, the commonly used Aluminum Oxide substrate has been replaced by individual silicone packages for all semiconductors.

The creepage distance has increased, as shown in figure B22.b. The distance between the semiconductors heed to have a minimum value of two times the wall thickness of the tubes, thus further disabling any convective cooling of the device.
As a result, it is clear that these tubes should only be used when it's necessary, in order to meet isolation requirements. Therefore, in the Mass GI the tubes should only be used in the primary bridge, as this bridge needs to have reinforced isolation. The isolation requirements for the secondary bridge are already met.
ADDENDUM

C

Datasheets and standards
**C.1 Design Specifications Mass GI 230V / 120V**

The table below contains an updated overview of the current design specifications of the 230V Mass GI and the desired design specifications of the 120V Mass GI design. The original overview has been manufactured by Marc Persoon in November 2006.

<table>
<thead>
<tr>
<th>Model (status)</th>
<th>Mass GI 3.5 (established design specs)</th>
<th>Mass GI 120V (desired design specs)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Article number</td>
<td>8800350</td>
<td>8800120</td>
</tr>
</tbody>
</table>

**General specifications**

<table>
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<tr>
<th></th>
<th>Mass GI 3.5</th>
<th>Mass GI 120V</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal power</td>
<td>3500VA</td>
<td>3600VA</td>
</tr>
<tr>
<td>Input Voltage</td>
<td>90-255Vac</td>
<td>90-130Vac</td>
</tr>
<tr>
<td>Input Current</td>
<td>16A cont @ 40°C</td>
<td>30A cont @ 40°C</td>
</tr>
<tr>
<td>Input frequency</td>
<td>45 - 65 Hz</td>
<td>45 - 65 Hz</td>
</tr>
<tr>
<td>Output voltage</td>
<td>Same as input voltage ±5%</td>
<td>Same as input voltage ±5%</td>
</tr>
<tr>
<td>Output frequency</td>
<td>Same as input frequency</td>
<td>Same as input frequency</td>
</tr>
</tbody>
</table>

**Environment**

<table>
<thead>
<tr>
<th></th>
<th>Mass GI 3.5</th>
<th>Mass GI 120V</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dimensions</td>
<td>340x261x144 mm / 13.4x10.3x5.7inch</td>
<td></td>
</tr>
<tr>
<td>Weight</td>
<td>5.5kg</td>
<td></td>
</tr>
<tr>
<td>Specified operation temperature: Full specifications from 0°C/32°F to 40°C/104°F. Derating with 5%/°C or 2.8%/°F at ambient temperatures from 40°C/104°F to 60°C/140°F. Shutdown at 80°C/176°F heat sink temperature.</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Non operating temperature: Ambient temperature -50°C/-58°F to 100°C/212°F</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Relative humidity: Protected against humidity and condensing air by conformal coating on both sides of all PCB's. Max 95% relative humidity, non condensing.</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

**Options**

<table>
<thead>
<tr>
<th></th>
<th>Mass GI 3.5</th>
<th>Mass GI 120V</th>
</tr>
</thead>
<tbody>
<tr>
<td>Alarm contact: Yes, potential free, triggered by over and undervoltage protection, short circuit and temperature</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Remote panel: Yes, Masterlink ACM article nummer 70403220</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Parallel: Up to four modes parallel</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
## C.2 Electrolyte potentials

<table>
<thead>
<tr>
<th>Half-reaction</th>
<th>( E^\circ ) (V)</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \text{Li}^+(\text{aq}) + e^- \rightarrow \text{Li}(s) )</td>
<td>-3.05</td>
</tr>
<tr>
<td>( \text{Al}^3+(\text{aq}) + 3e^- \rightarrow \text{Al}(s) )</td>
<td>-1.68</td>
</tr>
<tr>
<td>( \text{Mn}^{2+}(\text{aq}) + 2e^- \rightarrow \text{Mn}(s) )</td>
<td>-1.18</td>
</tr>
<tr>
<td>( 2 \text{H}_2\text{O}(l) + 2e^- \rightarrow \text{H}_2(g) + 2 \text{OH}^-(\text{aq}) )</td>
<td>-0.83</td>
</tr>
<tr>
<td>( \text{Zn}^{2+}(\text{aq}) + 2e^- \rightarrow \text{Zn}(s) )</td>
<td>-0.76</td>
</tr>
<tr>
<td>( \text{Cr}^{3+}(\text{aq}) + 3e^- \rightarrow \text{Cr}(s) )</td>
<td>-0.74</td>
</tr>
<tr>
<td>( 2\text{TiO}_2(s) + 2\text{H}^+ + 2e^- \rightarrow \text{Ti}_3\text{O}_5(s) + \text{H}_2\text{O} )</td>
<td>-0.56</td>
</tr>
<tr>
<td>( \text{Fe}^{2+}(\text{aq}) + 2e^- \rightarrow \text{Fe}(s) )</td>
<td>-0.44</td>
</tr>
<tr>
<td>( \text{Cr}^{3+}(\text{aq}) + e^- \rightarrow \text{Cr}^{2+}(\text{aq}) )</td>
<td>-0.42</td>
</tr>
<tr>
<td>( \text{Ni}^{2+}(\text{aq}) + 2e^- \rightarrow \text{Ni}(s) )</td>
<td>-0.26</td>
</tr>
<tr>
<td>( \text{Sn}^{2+}(\text{aq}) + 2e^- \rightarrow \text{Sn}(s) )</td>
<td>-0.13</td>
</tr>
<tr>
<td>( \text{O}_2(g) + \text{H}^+ + e^- \rightarrow \text{HO}_2^-(\text{aq}) )</td>
<td>-0.13</td>
</tr>
<tr>
<td>( \text{Pb}^{2+}(\text{aq}) + 2e^- \rightarrow \text{Pb}(s) )</td>
<td>-0.13</td>
</tr>
<tr>
<td>( 2\text{H}^+(\text{aq}) + 2e^- \rightarrow \text{H}_2(g) )</td>
<td>0.00</td>
</tr>
<tr>
<td>( \text{Cu}^{2+}(\text{aq}) + 2e^- \rightarrow \text{Cu}(s) )</td>
<td>+0.34</td>
</tr>
<tr>
<td>( \text{Cu}^+(\text{aq}) + e^- \rightarrow \text{Cu}(s) )</td>
<td>+0.52</td>
</tr>
<tr>
<td>( \text{I}_2(s) + 2e^- \rightarrow 2\text{I}^-(\text{aq}) )</td>
<td>+0.54</td>
</tr>
<tr>
<td>( \text{I}_3^-(\text{aq}) + 2e^- \rightarrow 3\text{I}^-(\text{aq}) )</td>
<td>+0.54</td>
</tr>
<tr>
<td>( \text{Fe}^{3+}(\text{aq}) + e^- \rightarrow \text{Fe}^{2+}(\text{aq}) )</td>
<td>+0.77</td>
</tr>
<tr>
<td>( \text{Hg}_2^{2+}(\text{aq}) + 2e^- \rightarrow 2\text{Hg}(l) )</td>
<td>+0.80</td>
</tr>
<tr>
<td>( \text{Ag}^+(\text{aq}) + e^- \rightarrow \text{Ag}(s) )</td>
<td>+0.80</td>
</tr>
<tr>
<td>( \text{Ag}_2\text{O}(s) + 2\text{H}^+ + 2e^- \rightarrow 2\text{Ag}(s) )</td>
<td>+1.17</td>
</tr>
<tr>
<td>( \text{Cl}_2(g) + 2e^- \rightarrow 2\text{Cl}^-(\text{aq}) )</td>
<td>+1.36</td>
</tr>
<tr>
<td>( \text{HO}_2^- + \text{H}^+ + e^- \rightarrow \text{H}_2\text{O}_2(\text{aq}) )</td>
<td>+1.51</td>
</tr>
<tr>
<td>( \text{Au}^{3+}(\text{aq}) + 3e^- \rightarrow \text{Au}(s) )</td>
<td>+1.52</td>
</tr>
<tr>
<td>( \text{Au}^+(\text{aq}) + e^- \rightarrow \text{Au}(s) )</td>
<td>+1.83</td>
</tr>
</tbody>
</table>
C.3 Choice of simulation program

A thorough investigation of simulation programs has been made in the beginning of this final thesis. Since already quite a number of software packages were known, it was easy to try to build up the schematics in both mathematical as power electronics simulation programs. Throughout this report therefore various simulations can be found.

C.3.1 Switch layout in simulations

For the switches in the implemented switch mode converters, two kinds of layouts have been used. For a single H-bridge, real models of IGBTs and diodes could be used. However, when simulating a complete DAB structure with phase shifts between the primary and secondary gating circuits, the IGBTs with anti-parallel diodes had to be replaced by ideal switches.

This simplification had as a positive point the fact that the ideal behavior of the Dual Active Bridge could be investigated, at different switching frequencies and for different values for the leakage inductance and the output capacitance. This choice also brought up several restrictions to the simulation output. First of all the steady state behavior was incorrect, as the current and temperature dependant voltage drop over the switches could not be modeled. Therefore, a good model for calculating the efficiency was out of reach. Secondly, also the dynamical behavior of the simulation refrained from certain parasitic effects, as for example the Miller and input capacitance and the diode reverse recovery, affecting the switching losses of the series connected IGBT.

C.3.2 Mathematical simulation programs

In order to gain insight in the control of the DAB technology, building a model in MATLAB Simulink was very useful. A big disadvantage of this package however is the simulation time, needed when using switching devices in a simulation. Simulations with ideal switches instead of IGBTs took minutes. More realistic components, using semiconductor models, however could take an hour, making them inapplicable for dynamic modeling.

C.3.3 Power Electronics simulation programs

In the beginning of my thesis, only student and evaluation versions of power electronics simulation programs were available at Mastervolt. Since there was no overview yet on the available software packages and their price, an overview on the electrical schematics simulation programs, suitable for Power Electronic circuits, is put in addendum C3.

Finally, the program SIMetrix has been chosen for most of the performed simulations. Even though only the evaluation version of this Circuit Simulation software package was available, the belonging numerical restrictions weren't limiting the size of the simulations too much. Therefore most circuits, even up to the DAB itself, could be simulated, when only a smart usage of components was implemented.

Concluding, even though a licensed version of SIMetrix of course would have produced more specific results, the current simulations were able to give more than enough insight information on the results, which could be expected when transferring the electrical circuit from SIMetrix into a real life PCB.
C.4 Power Electronics Simulators

Since this project contains many electronic circuits, a good Power Electronics simulator was necessary for further development. Since quite some electronic simulators are available at this moment, I decided to make a short overview on the current simulators, dedicated to simulation of Power Electronics.

Most of the simulators are based on SPICE (Simulation Program with Integrated Circuits Emphasis), which is a general-purpose circuit simulation program for nonlinear dc, nonlinear transient, and linear ac analyses. SPICE was originally developed at the Electronics Research Laboratory of the University of California, Berkeley in 1975 by Larry Nagel and Donald Pederson. Many commercial versions of SPICE have later replaced Berkeley SPICE as the industry standard.

While many are still compatible with the original Berkeley syntax, commercial vendors added proprietary extensions that limit the portability of circuit descriptions and models between different programs. For digital circuits (e.g., digital driver control), dedicated simulators exist that run orders of magnitude faster than the traditional Spice tools.

Typical power electronics system analyses consist of many aspects and are multidisciplinary. These are, for example, parameters of magnetic actuators, parameters and precise models of semiconductor switches, electrical machine parameters, thermal effects, different control issues, packaging and parasitic effects as a result of different layout. The power electronics simulator can take a central position in this case. In this project however, only the electronic circuits will be simulated.

All mentioned prices do not include VAT.

C.4.1 PSCAD

Website: www.pscad.com

PSCAD (Power Systems Computer-Aided Design) is developed at the Manitoba HVDC Research Centre, a division of Manitoba Hydro, a large Canadian electrical utility.

PSCAD is a fast, accurate, and easy-to-use power system simulator for the design and verification of all types of power systems. PSCAD acts as a powerful and flexible graphical user interface to the EMTDC™ transients and POWER FLOW simulation engines. The world-renowned Fortran based EMTDC solution engine is most suitable for simulating time domain instantaneous responses (also popularly known as electromagnetic transients or instantaneous solutions), in both electrical and control systems. The POWER FLOW engine provides a fast and accurate load flow solution.

PSCAD is used for the modeling and simulation of the following applications:

- Wind energy system design and integration to the electrical grid
- Power quality studies showing the full frequency response of the system including harmonics
- Power electronic design for high voltage applications including HVDC, facts
- Control systems design and optimization
- Protection system validation studies

The graphical interface is very user friendly. Since I've been using this simulator for my internship at ABB Corporate Research in Västerås, Sweden, I have had the chance to get thoroughly get to know the program. The simple implementation of switches and
variables, for example to change component values during a simulation, make the program very well suited for dynamical analysis and design. Also the help desk is very fast and willing to help. A bug in the software, which I accidentally had found during my time at ABB Corporate Research, was quickly acknowledged. Unfortunately there was no solution for the problem at that time, but they did accurately help me to solve the problem on another way.

A single, non-floating license will cost 16,660,-. This does not include an optional Intel Visual Fortran V9 & Visual C++.net software package, which will cost another 1190,-. A floating license for 2 simultaneous users will cost 39,270,-. As can be seen, this is by far the most expensive simulation software.

C.4.2 PLECS

Website: www.plexim.com

PLECS is developed by Plexim GmbH, founded as a spin-off company from the Swiss Federal Institute of Technology ETH Zurich in June 2002. PLECS is a software tool for the fast simulation of electrical circuits within the Matlab Simulink environment. Although specially designed for power electronics systems it is a versatile program for any combined simulation of electrical circuits and controls.

Each PLECS circuit is represented in a Simulink model as an individual block. Signals can be fed into these blocks in order to control electrical sources or switch devices. Measurements taken in the circuit are accessible at the block outputs. The measurements can be viewed in a Simulink scope, post-processed in MATLAB, or used for controlling a system.

On the website, a trial version is available. The commercial version of PLECS is available in two different shapes: via individual licenses and via concurrent licenses. Individual licenses have more options, but cannot be shared by multiple users. Current costs are €10,000,- per individual licenses. Concurrent licenses can be used as floating license, therefore usable by all engineers at Mastervolt. A set price for two floating licenses is €10,000,-. The more floating licenses are bought, the lower the individual price will be.

C.4.3 PSpice

Website: www.cadence.com/orcad

PSpice, originally from MicroSim, then OrCAD is now part of the product range of Cadence Design Systems. PSpice A/D is a full-featured, native mixed-signal simulator, capable of integrating analog and event-driven digital simulations to improve speed without loss of accuracy. PSpice AA (Advanced Analysis) incorporates four extra capabilities—sensitivity analysis, optimization, Smoke (stress analysis), and Monte Carlo (yield analysis). Used in conjunction with PSpice A/D, these capabilities enable engineers to create virtual prototypes of designs and maximize circuit performance automatically.

Support for multiple simulation profiles enables users to recall and run different simulations on the same schematic. Simulation bias results can be viewed directly on the schematic including node voltages, device power calculations, and pin and sub circuit current.

PSpice and Simulink, a platform for multi-domain simulation and model-based design of dynamic systems, can be used together. The combination of these two programs provides the ability to perform system-level simulations that include realistic models of electrical components.
**C.4.4 SwitcherCad**  
*Website: http://www.linear.com/company/software.jsp*  
One of the freeware simulators is SwitcherCad, formally known as LTSpice. It is developed by Linear Technology and therefore intended for simulation the performance of LinearTech's switch mode power supply controllers.

SwitcherCAD III is a high performance Spice III simulator, schematic capture and waveform viewer with enhancements and models for easing the simulation of switching regulators. The Linear Technology enhancements to Spice have made simulating switching regulators faster compared to normal Spice simulators, allowing the user to view waveforms for most switching regulators in just a few minutes.

Even though the schematic capture front-end is more difficult to use than the other simulation programs, the SPICE simulator and the many methods for analysis are of a good quality. Addition of many models of components however is a necessity, since the program is dedicated to switch mode power supplies and therefore only has certain libraries.

**C.4.5 Caspoc**  
*Website: www.simulation-research.com*  
The simulation package CASPOC is especially designed for simulation of Power Electronics and Electrical Drives. With CASPOC power electronics, electrical machines, load and control can be implemented into one multilevel model.

This multilevel model includes a circuit level for the modeling of Switched Mode Power Supplies, a component level for the modeling of electrical machines / loads and a system level for the modeling of control algorithms. Because CASPOC is especially designed for the simulation of power electronics and electrical drives, there are no convergence problems and the simulation runs 10 to 100 times faster than any general electronics simulation program.

Also, Caspoc allows that the circuit is animated during the simulation. This gives insight in the behavior of the circuit during its operation. The user can therefore interact with the simulation/animation by changing parameters during the simulation/animation and immediately see their influence on the circuit behavior.

**C.4.6 Simplorer**  
*Website: www.simplorer.com*  
Simplorer is a multi-domain, system simulation software package for the design of high-performance electromechanical systems commonly found in the automotive, aerospace/defense, and industrial automation industries. Simplorer includes a fast and numerically stable circuit simulator, a block-diagram system simulator for signal analysis and control design, and an event-driven state-machine simulator for discontinuous processes.

Simplorer incorporates VHDL-AMS, the IEEE industry-standard modeling language for analog, digital, mixed-signal, and multi-domain systems. Simplorer Pro also supports third-party mathematical codes of the following programs: MATLAB, Simulink, MathCAD, ADVISOR, and C/C++ Programming.
Simplorer is developed by Ansoft and is part of their range in simulators for electromechanical systems.

**C.4.7 PSIM**

*Website: www.powersimtech.com*

PSIM is developed by Powersim. Powersim develops and markets leading simulation and design tools for research and product development in power supplies, motor drives, and power conversion and control systems.

Powersim's PSIM is a simulation software tool, specifically designed for power electronics and motor control. PSIM's simulation engine uses efficient algorithms that overcome problems of convergence failure and long simulation times that exist in many other simulation software packages. Its fast simulation allows repetitive simulation runs and significantly shortens the design cycle. PSIM also features an intuitive graphic user interface and extensive on-line help. External DLL blocks are provided that allow users to write custom C/C++ code, compile it into DLL, and dynamically link to PSIM. This significantly expands PSIM's flexibility as it allows users to implement virtually any device models or control circuitry in powerful programming languages.

Users can change parameter values and view voltages/currents in the middle of a simulation. This makes the PSIM environment interactive, and users can easily fine tune parameters until desired performance is achieved. Also Magnetics Modeling Blocks are available. These are basic building blocks, provided to model the magnetic equivalent circuit based on the flux flow path. These blocks are very flexible and can model various types of magnetic devices.

After requesting quotations on the costs of the software, a single version appeared to cost €5,170,- and a floating license with two simultaneous work spaces is going to cost €10,990,-

**C.4.8 Power Supply Designer**

*Website: www.intusoft.com*

Power Supply Designer integrates Intusoft's "SpiceNet" schematic entry and design management system, with the proven "lsSpice4" analog and mixed-signal simulation kernel. The package also features "IntuScope," ICAP/4's waveform viewing and signal processing tool. Other features are the "Magnetics Designer" for the design and synthesis of transformers and inductors and model development and library management tools. Power Supply Designer also contains newly enhanced analyses for advanced design verification, like design pass/fail measurement capability and automated component stress alarms.

IntuScope features more than 100 waveform calculator functions plus a powerful scripting language for creating special waveform operations. Further more, the "press button to sweep component values" while simultaneously viewing real-time waveforms makes debugging and redesigning easier. Magnetics Designer provides complete transformer and inductor synthesis based upon electrical specifications. It includes a winding sheet report and SPICE-compatible models with parasitics.

A single node locked license of Magnetics Designer and ICAP/4 Windows Power Deluxe can be bought through the Dutch company Tech 5 B.V., www.tech5.nl, and will cost €4232,-. A set of two floating licenses of the mentioned software package would cost €13,248,-.
**C.4.9 SIMetrix/SIMPLIS**

Website: www.catena.uk.com

SIMetrix is a mixed-mode circuit simulation package designed for professional electronics engineers. SIMetrix comprises a substantially enhanced SPICE simulator, schematic editor and waveform viewer in a unified environment.

SIMetrix/SIMPLIS is an advanced software tool that enables efficient design of power electronics circuits. SIMetrix/SIMPLIS combines accuracy and convergence in a unique and powerful design environment, enabling 10-50x faster simulation for power supply designs. When it senses trouble converging, it automatically invokes a program called "Pseudo Transient Analysis", which places large capacitances at all nodes and then finds the correct biasing points. The program stores the circuit state at any point during a transient run. This result can be used to initialize an AC analysis and is also useful for investigating the stability of circuits in operating conditions that cannot be obtained using a DC analysis. Measurement points can be added after the simulation, which is very convenient while debugging a circuit.

SIMetrix Intro and SIMetrix/SIMPLIS Intro are free products available by download. SIMetrix Intro has virtually all the features of SIMetrix Micron AD but is subject to circuit size restrictions. SIMetrix/SIMPLIS Intro is the same but also includes the SIMPLIS simulator.

The component restriction has limited the size of the simulations I have made. However, by replacing IGBTs by ideal switches or by simulating only a part of the total circuit, these limitations have never been a real problem.

Since the free intro version is used, there is a limit to the size of circuit that can be simulated. The exact limits are:

- 120 analog nodes
- 36 digital nodes
- 72 digital ports
- 24 digital components
- 36 digital outputs

Although the number of allowed digital nodes is small, there is no limit to the size of definition for an arbitrary logic block. This in practice means that quite large digital circuits can be simulated.

<table>
<thead>
<tr>
<th>Component</th>
<th>Component score</th>
</tr>
</thead>
<tbody>
<tr>
<td>BJT</td>
<td>21</td>
</tr>
<tr>
<td>Diode</td>
<td>5</td>
</tr>
<tr>
<td>JFET</td>
<td>13</td>
</tr>
<tr>
<td>GaasFET</td>
<td>13</td>
</tr>
<tr>
<td>MOS devices</td>
<td>17</td>
</tr>
<tr>
<td>Capacitor</td>
<td>2</td>
</tr>
<tr>
<td>Inductor</td>
<td>2</td>
</tr>
<tr>
<td>BSIM3</td>
<td>18</td>
</tr>
<tr>
<td>Laplace</td>
<td>$2 + 2 \times$ (order of denominator)</td>
</tr>
<tr>
<td>All other devices</td>
<td>0</td>
</tr>
</tbody>
</table>

There is also a limit to the number of several analog components. The limit can be calculated by multiplying the number of each type of component by the score, given in
the table C-1, and finally summing all the results. The total must be 384 or less for the circuit to be accepted.

Since I've been using several different IGBT models, which are not present in the common library, I had to load several component models. The simulator has accepted all component models that I have tried. Generally, SIMetrix is not only more user-friendly than other programs I have used, but it is also fast and has capabilities that others don't have, like the many different methods for nested analysis.

A single version, which is using a USB-dongle, will cost €4995,-, a network version with 2 floating licenses will cost €11,350.-
C.5 Datasheet HGTG20N60B3D

40A, 600V, UFS Series N-Channel IGBT with Anti-Parallel Hyperfast Diode

The HGTG20N60B3D is a MOS gated high voltage switching device combining the best features of MOSFETs and bipolar transistors. The device has the high input impedance of a MOSFET and the low on-state conduction loss of a bipolar transistor. The much lower on-state voltage drop varies only moderately between 25°C and 150°C. The diode used in anti-parallel with the IGBT is the RH3P060.

The IGBT is ideal for many high voltage switching applications operating at moderate frequencies where low conduction losses are essential.

Formerly developmental type TA49016.

Ordering Information

<table>
<thead>
<tr>
<th>PART NUMBER</th>
<th>PACKAGE</th>
<th>BRAND</th>
</tr>
</thead>
<tbody>
<tr>
<td>HGTG20N60B3D</td>
<td>TO-247</td>
<td>G20N60B3D</td>
</tr>
</tbody>
</table>

NOTE When ordering, use the entire part number

Features

- 40A, 600V at T_C = 25°C
- Typical Fall Time: 140ns at 150°C
- Short Circuit Rated
- Low Conduction Loss
- Hyperfast Anti-Parallel Diode

Packaging

JEDEC STYLE TO-247

Symbol

The Fairchild Semiconductor IGBT product is covered by one or more of the following U.S. patents:

4,364,073 4,417,385 4,430,792 4,443,931 4,466,176 4,516,143 4,532,534 4,567,713
4,596,461 4,605,948 4,631,564 4,639,754 4,639,782 4,641,162 4,644,037
4,662,105 4,694,313 4,717,579 4,743,952 4,763,590 4,794,432 4,801,986
4,803,533 4,809,045 4,822,047 4,837,506 4,838,507 4,860,080 4,883,767
4,888,627 4,890,143 4,901,127 4,904,609 4,933,740 4,963,951 4,969,027
## Addendum

**HGTG20N60B3D**

### Absolute Maximum Ratings

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Test Conditions</th>
<th>Min</th>
<th>Typ</th>
<th>Max</th>
</tr>
</thead>
<tbody>
<tr>
<td>Collector to Emitter Voltage</td>
<td>BV_CES</td>
<td>TC = 25°C, VGE = 5V</td>
<td>600</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Collector to Gate Voltage, RGE = 1M</td>
<td>BV_CGR</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>V</td>
</tr>
<tr>
<td>Collector Current Continuous</td>
<td>I_CES</td>
<td>TC = 150°C</td>
<td>-</td>
<td>20</td>
<td>A</td>
</tr>
<tr>
<td>Average Diode Forward Current</td>
<td>I(D)</td>
<td>V_CES = 0.5BVCES</td>
<td>2A</td>
<td>1.9A</td>
<td></td>
</tr>
<tr>
<td>Collector Current Pulled (Note 1)</td>
<td>I(CM)</td>
<td>TC = 150°C</td>
<td>160</td>
<td>A</td>
<td></td>
</tr>
<tr>
<td>Gate to Emitter Voltage Continuous</td>
<td>VGES</td>
<td>TC = 25°C</td>
<td>-</td>
<td>-</td>
<td>20V</td>
</tr>
<tr>
<td>Gate to Emitter Voltage Pulsed</td>
<td>VDEM</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>V</td>
</tr>
<tr>
<td>Collector Current Continuous</td>
<td>ICM</td>
<td>TC = 150°C</td>
<td>-</td>
<td>20</td>
<td>A</td>
</tr>
<tr>
<td>Power Dissipation Total at TC = 150°C</td>
<td>PD</td>
<td>-</td>
<td>165</td>
<td>W</td>
<td></td>
</tr>
<tr>
<td>Power Dissipation Derating</td>
<td>T_C &gt; 25°C</td>
<td>-</td>
<td>-</td>
<td>0.68W/°C</td>
<td></td>
</tr>
<tr>
<td>Operating and Storage Temperature Range</td>
<td>TJ, TSOT</td>
<td>-40 to 150°C</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Maximum Lead Temperature for Soldering</td>
<td>TL</td>
<td>260°C</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Short Circuit Standby Current (Note 2)</td>
<td>IL</td>
<td>-</td>
<td>30</td>
<td>A</td>
<td></td>
</tr>
<tr>
<td>Short Circuit Standby Current (Note 2)</td>
<td>V_CES</td>
<td>-</td>
<td>45</td>
<td>V</td>
<td></td>
</tr>
<tr>
<td>Collector to Emitter Breakdown Voltage</td>
<td>BVCES</td>
<td>TC = 250°C, VGE = 0V</td>
<td>600</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Collector to Emitter Leakage Current</td>
<td>ICES</td>
<td>V_CES = BV_CES</td>
<td>-</td>
<td>-</td>
<td>250μA</td>
</tr>
<tr>
<td>Collector to Emitter Saturation Voltage</td>
<td>V_CES(SAT)</td>
<td>TC = 25°C</td>
<td>-</td>
<td>18</td>
<td>20V</td>
</tr>
<tr>
<td>Gate to Emitter Leakage Current</td>
<td>IGES</td>
<td>VGE = -20V</td>
<td>-</td>
<td>-</td>
<td>100nA</td>
</tr>
<tr>
<td>Switching SOA</td>
<td>SSOA</td>
<td>V_CES = 480V</td>
<td>100</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Gate to Emitter Plateau Voltage</td>
<td>V_GEP</td>
<td>TC = 150°C, VGE = 0V</td>
<td>80</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>On-State Gate Charge</td>
<td>Q_D(on)</td>
<td>TC = 0.5BVCES</td>
<td>80</td>
<td>105</td>
<td>nC</td>
</tr>
<tr>
<td>Current Turn-On Delay Time</td>
<td>t_D(on)</td>
<td>TC = 150°C</td>
<td>-</td>
<td>25</td>
<td>ns</td>
</tr>
<tr>
<td>Current Turn-Off Delay Time</td>
<td>t_D(off)</td>
<td>TC = 0.8BVCES</td>
<td>-</td>
<td>275</td>
<td>ns</td>
</tr>
<tr>
<td>Current Fall Time</td>
<td>t_F</td>
<td>VGE = 15V</td>
<td>-</td>
<td>175</td>
<td>ns</td>
</tr>
<tr>
<td>Turn-On Energy</td>
<td>E_ON</td>
<td>L = 100μH</td>
<td>-</td>
<td>475</td>
<td>μJ</td>
</tr>
<tr>
<td>Turn-Off Energy (Note 3)</td>
<td>E_OFF</td>
<td>-</td>
<td>1050</td>
<td>-</td>
<td></td>
</tr>
<tr>
<td>Diode Forward Voltage</td>
<td>V_D</td>
<td>VGE = 20A</td>
<td>-</td>
<td>1.5</td>
<td>1.9V</td>
</tr>
<tr>
<td>Diode Reverse Recovery Time</td>
<td>t_R</td>
<td>RC = 20A, dQ/dt = 100μA/μs</td>
<td>-</td>
<td>55</td>
<td>ns</td>
</tr>
<tr>
<td>Thermal Resistance</td>
<td>R_JC</td>
<td>-</td>
<td>0.078</td>
<td>°C/W</td>
<td></td>
</tr>
</tbody>
</table>

**CAUTION:** Stresses above those listed in "Absolute Maximum Ratings" may cause permanent damage to the device. This is a stress only rating and operation of the device at these or any other conditions above those indicated in the operational sections of this specification is not implied.

**NOTES:**

1. Repetitive Rating Pulse width limited by maximum junction temperature.

### Electrical Specifications

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Test Conditions</th>
<th>Min</th>
<th>Typ</th>
<th>Max</th>
</tr>
</thead>
<tbody>
<tr>
<td>Collector to Emitter Breakdown Voltage</td>
<td>BV_CES</td>
<td>TC = 250°C, VGE = 0V</td>
<td>600</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Collector to Emitter Leakage Current</td>
<td>ICES</td>
<td>V_CES = BV_CES</td>
<td>-</td>
<td>-</td>
<td>250μA</td>
</tr>
<tr>
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<td>V_CES(SAT)</td>
<td>TC = 25°C</td>
<td>-</td>
<td>18</td>
<td>20V</td>
</tr>
<tr>
<td>Gate to Emitter Leakage Current</td>
<td>IGES</td>
<td>VGE = -20V</td>
<td>-</td>
<td>-</td>
<td>100nA</td>
</tr>
<tr>
<td>Switching SOA</td>
<td>SSOA</td>
<td>V_CES = 480V</td>
<td>100</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Gate to Emitter Plateau Voltage</td>
<td>V_GEP</td>
<td>TC = 150°C, VGE = 0V</td>
<td>80</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>On-State Gate Charge</td>
<td>Q_D(on)</td>
<td>TC = 0.5BVCES</td>
<td>80</td>
<td>105</td>
<td>nC</td>
</tr>
<tr>
<td>Current Turn-On Delay Time</td>
<td>t_D(on)</td>
<td>TC = 150°C</td>
<td>-</td>
<td>25</td>
<td>ns</td>
</tr>
<tr>
<td>Current Turn-Off Delay Time</td>
<td>t_D(off)</td>
<td>TC = 0.8BVCES</td>
<td>-</td>
<td>275</td>
<td>ns</td>
</tr>
<tr>
<td>Current Fall Time</td>
<td>t_F</td>
<td>VGE = 15V</td>
<td>-</td>
<td>175</td>
<td>ns</td>
</tr>
<tr>
<td>Turn-On Energy</td>
<td>E_ON</td>
<td>L = 100μH</td>
<td>-</td>
<td>475</td>
<td>μJ</td>
</tr>
<tr>
<td>Turn-Off Energy (Note 3)</td>
<td>E_OFF</td>
<td>-</td>
<td>1050</td>
<td>-</td>
<td></td>
</tr>
<tr>
<td>Diode Forward Voltage</td>
<td>V_D</td>
<td>VGE = 20A</td>
<td>-</td>
<td>1.5</td>
<td>1.9V</td>
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<tr>
<td>Diode Reverse Recovery Time</td>
<td>t_R</td>
<td>RC = 20A, dQ/dt = 100μA/μs</td>
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<td>55</td>
<td>ns</td>
</tr>
<tr>
<td>Thermal Resistance</td>
<td>R_JC</td>
<td>-</td>
<td>0.078</td>
<td>°C/W</td>
<td></td>
</tr>
</tbody>
</table>

**NOTE:** Turn-Off Energy Loss (E_OFF) is defined as the integral of instantaneous power loss starting at the trailing edge of the input pulse and ending at the point where the collector current equals zero (I_CES = 0A). The HGTG20N60B3D was tested per JEDEC standard No. 24-1 Method for Measurement of Power Device Turn-Off Switching Loss. This test method produces the true total Turn-Off Energy Loss. Turn-On losses include diode losses.
Typical Performance Curves

**FIGURE 1. TRANSFER CHARACTERISTICS**

**FIGURE 2. SATURATION CHARACTERISTICS**

**FIGURE 3. DC COLLECTOR CURRENT vs CASE TEMPERATURE**

**FIGURE 4. COLLECTOR TO EMITTER ON-STATE VOLTAGE**

**FIGURE 5. CAPACITANCE vs COLLECTOR TO EMITTER VOLTAGE**

**FIGURE 6. GATE CHARGE WAVEFORMS**
Typical Performance Curves (Continued)

**FIGURE 7. TURN-ON DELAY TIME vs COLLECTOR TO EMITTER CURRENT**

**FIGURE 8. TURN-OFF DELAY TIME vs COLLECTOR TO EMITTER CURRENT**

**FIGURE 9. TURN-ON RISE TIME vs COLLECTOR TO EMITTER CURRENT**

**FIGURE 10. TURN-OFF FALL TIME vs COLLECTOR TO EMITTER CURRENT**

**FIGURE 11. TURN-ON ENERGY LOSS vs COLLECTOR TO EMITTER CURRENT**

**FIGURE 12. TURN-OFF ENERGY LOSS vs COLLECTOR TO EMITTER CURRENT**
Typical Performance Curves (Continued)

\[ T_J = 150^\circ C, \quad T_C = 75^\circ C, \quad V_{CE} = 15 V \]
\[ \frac{R}{L} = 10, \quad \frac{L}{100 m} \]
\[ V_{CE} = 400 V \]

\[ f_{MAX} = 0.05 \left( V_{DSS} + V_{PD} \right) \]
\[ P_D = \text{ALLOWABLE DISSIPATION} \]
\[ P_C = \text{CONDUCTION DISSIPATION} \]
\[ (\text{DUTY FACTOR} = 60\%) \]
\[ R_{JC} = 0.78^\circ C/W \]

**FIGURE 13. OPERATING FREQUENCY vs COLLECTOR TO EMMITTER CURRENT**

**FIGURE 14. SWITCHING SAFE OPERATING AREA**

**FIGURE 15. IGBT NORMALIZED TRANSIENT THERMAL RESPONSE, JUNCTION TO CASE**

**FIGURE 16. DIODE FORWARD CURRENT vs FORWARD VOLTAGE DROP**

**FIGURE 17. RECOVERY TIMES vs FORWARD CURRENT**
Handling Precautions for IGBTs
Insulated Gate Bipolar Transistors are susceptible to gate-insulation damage by the electrostatic discharge of energy through the devices. When handling these devices, care should be exercised to assure that the static charge built in the handler's body capacitance is not discharged through the device. With proper handling and discharge procedures, however, IGBTs are currently being extensively used in production by numerous equipment manufacturers in military, industrial and consumer applications, with virtually no damage problems due to electrostatic discharge. IGBTs can be handled safely if the following basic precautions are taken:

1. Prior to assembly into a circuit, all leads should be kept shorted together either by the use of metal shorting springs or by the insertion into conductive material such as "ECCOSORB LD26" or equivalent.
2. When devices are removed by hand from their carriers, the hand being used should be grounded by any suitable means — for example, with a metallic wristband.
3. Tips of soldering irons should be grounded.
4. Devices should never be inserted into or removed from circuits with power on.
5. Gate Voltage Rating - Never exceed the gate-voltage rating of V_{GEM}. Exceeding the rated V_{GEM} can result in permanent damage to the oxide layer in the gate region.
6. Gate Termination - The gates of these devices are essentially capacitors. Circuits that leave the gate open-circuited or floating should be avoided. These conditions can result in turn-on of the device due to voltage buildup on the input capacitor due to leakage currents or pickup.
7. Gate Protection - These devices do not have an internal monolithic zener diode from gate to emitter. If gate protection is required an external zener is recommended.

Operating Frequency Information
Operating frequency information for a typical device (Figure 13) is presented as a guide for estimating device performance for a specific application. Other typical frequency vs collector current (I_{CE}) plots are possible using the information shown for a typical unit in Figures 4, 7, 8, 11 and 12. The operating frequency plot (Figure 13) of a typical device shows f_{MAX1} or f_{MAX2} whichever is smaller at each point. The information is based on measurements of a typical device and is bounded by the maximum rated junction temperature.

- f_{MAX1} is defined by f_{MAX1} = 0.05(t_{d(OFF)} + t_{d(ON)})/t_{d(ON)}.
- Deadtime (the denominator) has been arbitrarily held to 10% of the on-state time for a 50% duty factor. Other definitions are possible. t_{d(OFF)} and t_{d(ON)} are defined in Figure 19.

Device turn-off delay can establish an additional frequency limiting condition for an application other than T_{JMAX}. t_{d(OFF)} is important when controlling output ripple under a lightly loaded condition.

- f_{MAX2} is defined by f_{MAX2} = (P_D - P_C)/(E_{OFF} + E_{ON}).
- The allowable dissipation (P_D) is defined by P_D = (T_{JMAX} - T_Jc)/R_{JC}.
- The sum of device switching and conduction losses must not exceed P_D. A 50% duty factor was used (Figure 13) and the conduction losses (P_C) are approximated by P_C = (V_{CE} x I_{CE})/2.
- E_{ON} and E_{OFF} are defined in the switching waveforms shown in Figure 19. E_{ON} is the integral of the instantaneous power loss (I_{CE} x V_{CE}) during turn-on and E_{OFF} is the integral of the instantaneous power loss during turn-off. All tail losses are included in the calculation for E_{OFF}; i.e., the collector current equals zero (I_{CE} = 0).
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FRFET™
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HiSeC™
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LittleFET™
MicroFET™
MicroPak™
MICROWIRE™
OPTOLOGIC™
OPTOPLANAR™
PACMAN™
POP™
Power247™
PowerTrench™
QS™
QT Optoelectronics™
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PRODUCT STATUS DEFINITIONS

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<th>Product Status</th>
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<td>Advance Information</td>
<td>Formative or In Design</td>
<td>This datasheet contains the design specifications for product development. Specifications may change in any manner without notice.</td>
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<tr>
<td>Preliminary</td>
<td>First Production</td>
<td>This datasheet contains preliminary data, and supplementary data will be published at a later date. Fairchild Semiconductor reserves the right to make changes at any time without notice in order to improve design.</td>
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<tr>
<td>No Identification Needed</td>
<td>Full Production</td>
<td>This datasheet contains final specifications. Fairchild Semiconductor reserves the right to make changes at any time without notice in order to improve design.</td>
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<td>Not In Production</td>
<td>This datasheet contains specifications on a product that has been discontinued by Fairchild Semiconductor. The datasheet is printed for reference information only.</td>
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C.6 Package marking conventions for the HGTG20N60B3D

The case of an IGBT usually gives not only more information on the IGBT type, but also on the location of the assembly plant and the date of manufacturing. Sometimes, this way also a fake badge of IGBTs can be recognized, as the code gives an invalid combination, as shown in addendum E5.

<table>
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<tr>
<th>Line 1:</th>
<th>Example $Y&amp;Z&amp;2&amp;T</th>
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<tr>
<td>&amp;Z = Assembly Plant Code</td>
<td></td>
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<tr>
<td>&amp;2 = Date Code</td>
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<tr>
<td>&amp;T = Die Trace Code</td>
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<tr>
<td>&amp;P = Marketing Status Code</td>
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<table>
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</thead>
<tbody>
<tr>
<td>Line 3:</td>
<td>Part designator continued</td>
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</tbody>
</table>

**Fairchild logo code**

$Y = \mathcal{F}$

**Assembly plant code**

<table>
<thead>
<tr>
<th>A = China (Subcontractor)</th>
<th>J = Japan (Subcontractor)</th>
<th>S = Singapore (NSSG, Subcontractor)</th>
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<tbody>
<tr>
<td>B = Bangkok, Thailand (Subcontractor)</td>
<td>K = Korea</td>
<td>T = Taiwan (Subcontractor)</td>
</tr>
<tr>
<td>C = Singapore (Subcontractor)</td>
<td>L = Salt Lake City, UT</td>
<td>V = Malaysia (Carsem, Subcontractor)</td>
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<tr>
<td>D = Cebu, Philippines</td>
<td>M = Malacca, Malaysia (Subcontractor)</td>
<td>X = USA (Subcontractor)</td>
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<td>E = Korea (Subcontractor)</td>
<td>N = (ChipPAC, Subcontractor)</td>
<td>Y = Malaysia (Unisem, Subcontractor)</td>
</tr>
<tr>
<td>H = Philippines (Subcontractor)</td>
<td>P = Penang, Malaysia</td>
<td>Z = South Portland, ME</td>
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</table>

**Date code**

This code identifies the calendar year and work week the product was manufactured.

2 Character Format:

&2 = XY

X = Last Digit of Calendar Year
Y = One Digit Work Week Code

3 Character Format:

&3 = XYY

X = Last Digit of Calendar Year
YY = Two Digit Work Week Code

4 Character Format:

&4 = XXYY

XX = Last Two Digits of Calendar Year
YY = Two Digit Work Week Code

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# C.7 Datasheet core material N87

23.02.2001

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**Material data sheet**

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<th>Initial permeability $100^\circ C$</th>
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<th>N87 old</th>
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<td>[mT]</td>
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<td>480</td>
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<td>[mT]</td>
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<td>[A/m]</td>
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<td>Resistivity $\rho$</td>
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<td>8</td>
</tr>
</tbody>
</table>

| Relative core losses $P_v$ (typical values) | $25^\circ C$ | 25 kHz, 200 mT | [kW/m²] | 115 | 115 |
|                                          | $100^\circ C$ | [kW/m²] | 55 | 55 |
|                                          | 20 kHz, 200 mT | [kW/m²] | 620 | 670 |
|                                          | 100 kHz, 200 mT | [kW/m²] | 375 | 385 |
|                                          | 25 kHz, 100 mT | [kW/m²] | 590 | 620 |
|                                          | 100 kHz, 100 mT | [kW/m²] | 390 | 410 |
|                                          | 25 kHz, 50 kHz | [kW/m²] | 255 | 260 |
|                                          | 100 kHz, 50 kHz | [kW/m²] | 215 | 235 |

Table 1: Material data sheet

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![Graph](image)

**Fig. 1:** Relative core losses at $f = 500$ kHz / $B = 50$ mT versus temperature (measured at R29 ring cores)

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ADDENDUM

D

Software and control
D.1 Control schemes for the Mass GI

For communication and measurement and control of various parameters, a PIC is implemented in the circuits of the Mass GI. Due to a possible future change of processor, all routines in the Mass GI will now be specified. Implementations of these routines can be found in the flow chart diagrams at the end of this chapter.

As mentioned in chapter 4, the following processes are taking place in this control unit:

- Startup control
- Master / Slave behavior and communication:
  - Error signaling from and to connected master and one or more slave units
  - Startup signal control between master and one or more slaves.
- Over current protection
- Over voltage protection
- Over temperature protection and fan control
- User interface control:
  - Output power indication to user
  - Error signaling to user in case of over current, over voltage or over temperature.

The information for the mentioned control circuits is provided by the following input signals:

- Heat sink temperature
- Main transformer temperature
- Input voltage
- Output current
- Position dipswitch Master / Slave
- Error status of connected master and / or one or more slave units
- Startup signal, generated by connected Master device

Most of the control circuits are clear: the temperature of the heat sink for example, should never exceed a certain temperature, set by hardware restrictions, thus enabling a simple temperature control loop. Also peak input voltage and peak output current control can be implemented. Again, the corresponding maximum values are controlled by hardware restrictions, for example by the maximum peak current the IGBTs can handle.

More interesting are the circuits, responsible for start up control and output current B curve control. Also, the main routine shows a number of specific details. Therefore, these three routines will now be explained more thorough.

D.1.1 Main routine

Logically, the main routine starts with the detection of incoming power. Once a supply voltage is fed to the controller, the main routine can start. The routine, handling this startup is called “initialization” and is explained more thoroughly in the next paragraph.

In order to be able to adequately respond to all situations, the main routine is divided into two loops: a fast and a slow loop. The fast loop, implementable with interrupts, checks if the input voltage and the output current stay under predefined values and stores them in a register for later use.
The slow loop, only called once per 10ms, checks if the RMS values of the input voltage and output current and temperatures of the two heat sinks and main transformers do not exceed the preset limits.

At the end of both loops, the status of parallel connected master or slave devices is checked. Eventually, the cycle is started again.

**D.1.2 Initialization routine**

To insure a safe start up sequence, a proper initialization routine is unmissable. First of all, all inputs and outputs are defined and are given an initialization value. Since the main gate signals are generated by an external circuit, it is important to switch off the gate signals as soon as possible when initializing the device. This way, active power transfer cannot be started, even though the control routines are not started yet.

Subsequently, the “Slave” input is checked to know if the device should act as a master or a slave unit, as the slave units will refrain from starting up as long as the master unit does not give a start signal. In order to give the slave devices enough time to power up and finish the initialization routine, the master will then wait for 200ms until it will activate itself and the connected slave devices.

The gate signals will now still be disabled, until a zero crossing is detected in the input voltage. At that moment, both the master and the slave devices will individually activate the gate signals.

**D.1.3 Output current B curve control routine**

The output current is guarded in two ways. As mentioned before, high peak values will turn off the device immediately. One of the demands of the original design however was the fact that the Mass GI should be able to emulate a B curve fuse control. In general, this means that low over currents can be allowed for a longer time than high over current values.

For a correct control, the RMS value of the output current is calculated and used as the input of this routine. Furthermore, the past values should be taken into account when judging the current situation. This is done by the introduction of a B curve limiter, using an over current history variable “Iav”.

Therefore, a counter system has been designed, which relates the RMS value of the output current to the maximum nominal current, in this case 16A. If the resulting value is higher than one, a situation of over current is present. As a result, Iav is increased with a number, related to the matter of overshoot. As the power, which is fed to the heat sink, is linear with the output current, the mentioned number should minimally relate linearly to the measured RMS value of this current.

When looking at the B curve characteristics, the relation between current and time is not linear, but quadratic. Therefore, the current spectrum between 1 and 2 times the maximum average output current is divided into a number of linear areas, each resulting in a specific increment for the over current history variable “Iav”.

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D.1.4 Temperature influence on the output current protection

A problem, which is not solved in this control however is the temperature dependency of the maximum current value of the IGBTs and diodes. As can be seen in addendum C4, containing the datasheet of the HGTG20N60B3D, the maximum current drops from 40A at room temperature to 20A at 125°C. This value of course is also dependent on the on-time of the component, being a constant 50% for the Mass GI. A possibility for the future could therefore be the implementation of a temperature related decrement factor, which could realistically estimate the effect of an over current on the behavior of the mass GI.

The cooling process of the heat sink and the main transformer is related to the difference in temperature, thus enabling a higher loss power flow at higher differences between air flow temperature and heat sink temperature. Knowledge of the ambient temperature could therefore also lead to higher allowable momentaneous currents at low ambient temperatures.

Besides the ambient temperature, also the current heat sink and main transformer temperature are important in this calculation. When the heat sink for example already has a high temperature, a lower amount of over current is allowable than in case of start up, when the components are still at ambient temperature.
D.2 DAB PWM control with the eZDSP F2808 DSP board.

The Mass GI obtains its gate signal from a relatively simple circuit around a standard CMOS IC, as previously shown. Of course, this task can also be done by a more sophisticated DSP. Implementation of a dead time, for example, can then simply happen by some lines of C-code instead of by an extra dead time circuit, as previously shown. This way, the DSP can continuously control aspects like dead time and phase shift between the various PWM signals.

For this final thesis, the eZDSP board with the TI TMS320F2808 processor was going to be the central control unit.

D.2.1 General F2808 eZDSP characteristics

The F2808 eZDSP comes standard with a wide range of i/o terminals and internal software and hardware configurations. Some general hardware features are:

- TMS320F2808 @ 100MHz, socketed
- 18K on-chip zero wait-state SARAM
- 256K bit 12C EEPROM
- 20MHz input clock to TMS320F2808
- On-board embedded JTAG emulation with USB host connection
- Support for external emulator via standard JTAG header
- 2 SCI ports with on-board transceivers. One port pinned out to standard 9 pin DSUB.
- 2 Enhanced CAN ports with on-board transceivers. One port pinned out to standard 9 pin DSUB.
- Boot mode selection switches
- 6 PWM pair outputs
- 16 analog inputs.

Some of these characteristics will be used, but overall, the possibilities exceed the minimum requirements. The main reason to choose this platform is the fact that 6 PWM outputs can be controlled simultaneously, with a high resolution in the time domain.

D.2.1.1 eZDSP software overview

For programming the chosen hardware platform, several software packages are available. First of all, there is VisSim/Embedded Controls Developer. For this application however, the by Texas Instruments supported Code Composer Studio, version 3.1, has been used. This program even supports Matlab models via “Embedded target for TI C2000 DSP”.

D.2.2 Gate signal generation with ePWM

Programming gate signals for the DSP is relatively simple. As mentioned before, the DSP contains six PWM modules. Each PWM module delivers two output signals, which can be controlled individually. The PWM modules can be connected, thus enabling a Master-(multiple)Slave configuration. An example is shown in figure D1.a, [36]. Between these modules, it is possible to implement a phase shift.

For the programmed applications, a triangular timing signal has been used, as shown in figure D1.b, [36]
Further details on programming the DSP can be found in [36] and other application manuals, which are placed on the added CD-ROM, as presented in addendum G.

### D.2.3 Gate driver

In order to be able to properly control the IGBTs, the PWM signals, with an amplitude of 3V, have to be fed to a gate driver. The boosted signal can then directly feed semiconductor switches or control them via gating transformers. The used gate drivers are standard gate drivers from the Mastervolt Mass Combi. This gate driver is also used for the IGBT test circuit, presented in addendum E4.

The circuit is shown in figure D2. The incoming gating signals Gate_A_in and Gate_B_in are each fed to a gate driver, type MIC4422. The resulting output signal is then fed to a half bridge, made of a P-channel MOSFET and an N-channel MOSFET. The gating circuit of these MOSFETs is divided into a $R_{on}$ and a $R_{off}$, as is explained in Addendum E4. The capacitors $C_1$ and $C_2$ are meant to filter out any HF disturbance, caused by the switching pattern.

In order to limit the peak current through the gate driver circuit, $R_{11}$ and $R_{12}$ are connected in series with the driver circuit power supply line. Capacitor $C_3$ reduces the ripple on the voltage, fed to the MOSFETs.

Four gate drivers have been mounted on a strip of epoxy. To this strip, also a metal L-profile is mounted, thus enabling a quick and solid mechanical connection to the converter, which needs to be controlled.
A practical implantation of the DSP with four driver circuits is shown in figure D3.a. Here, no gating transformers are connected yet, but LED loads are used to test the working principle. Figure D3.b shows the setup, used for the first measurements on the phase shifted DAB. The drivers are mounted on the converter PCB, but the DSP board is still lying on the table. Figure D3.c shows the setup, used for the first measurements on the phase shifted DAB. The drivers are mounted on the converter PCB, but the DSP board is still lying on the table. Figure D3.d shows the final setup, with the DSP attached to an isolated plastic plate, which is mechanically connected to the converter PCB with spacers with tapped ends.

Fig. D.3. Various test setups for implementation of DAB PWM control

D.2.4 Addition of voltage and current feedback

In order to be able to correctly control the DSP, a number of voltage and current values can be measured and used for adjustment of control parameters or for activation of safety circuits.

Especially for correct control of the phase shifted DAB, measurements of the output current and input and output voltage are important. Since these signals can already be monitored via the communication bus, the only task is to adjust the amplitude of these signals. The DSP ADC input range is 0-3V, input signals with an amplitude over 3V will destroy the internal ADC converter.

Fig. D.4. Input signal measurement. a) Circuit for 5V max input signals, b) circuit for 12V max input signals.

Figure D4 shows two simple circuits for adjustment of 0-5V and 0-12V input signals to 0-2.5V DSP input signals. The zener diode of 2.7V will protect the DSP input for unexpected over voltages.
D.3 Switching sequence of the DAB with primary boost

[Diagram showing the switching sequence with labels: $S_{\text{boost}}$, $S_{1p, S_{2n}}$, $S_{1n, S_{2p}}$, $S_{3p}$, $S_{3n}$, $S_{4p}$]
ADDENDUM

E

The IGBT: working principle, applications and measurement
E.1 Implementation of IGBTs as switching devices

In chapter three, the choice for IGBTs as the switch components has been explained. The IGBT is not a perfect switch; therefore the gating circuit will have to be adjusted to the specific properties of this device. This means that the output from the gate driver can not be connected directly to the gate of the IGBTs. To be aware of the problems, that have to be overcome, first the IGBT model will be discussed. Later on, the implementation of a gating circuit will be shown.

E.1.1 Physical IGBT model

A general description of an IGBT is a power MOSFET constructed on a p-type substrate, as illustrated by the generic IGBT cross section in figure E.1. This creates a parasitic transistor driven by the MOSFET and permits increased current flow in the same die area. The sacrifice is an additional diode drop due to the extra junction and turn-off delays while carriers are swept out of this junction.

![Physical model of an IGBT](image)

A positive voltage applied from the emitter to the gate terminal causes electrons to be drawn toward the gate in the body region. If the gate-emitter voltage reaches the threshold voltage, enough electrons are drawn toward the gate to form a conductive channel across the body region. This enables a current to flow from the collector to the emitter. This flow of electrons draws positive ions, or holes, from the p-type substrate into the drift region toward the emitter.

The drawback of the IGBT becomes visible during the switching sequences. During turn-off the electron flow can be stopped rather abruptly, just as in a power MOSFET, by reducing the gate-emitter voltage below the threshold voltage. However, as mentioned, the current flow also uses holes, which at turn off are left in the drift region. Since they can only be removed by voltage gradient and recombination, the IGBT suffers from a tail current during turn-off until all the holes are either recombined or swept out.

The rate of recombination however can be controlled, which is the purpose of the extra n+ buffer layer, shown in figure E.1. This buffer layer quickly absorbs trapped holes during the IGBT turn-off. Not all IGBT types have an n+ buffer layer implemented; those that have are called punch-through (PT), those that do not are called non punch-through (NPT). PT IGBTs are sometimes referred to as asymmetrical, and NPT as symmetrical.
For a given switching speed, NPT technology generally has a higher $V_{CE(on)}$ than PT technology. This difference is magnified further by fact that $V_{CE(on)}$ increases with temperature for NPT (positive temperature coefficient), whereas $V_{CE(on)}$ decreases with temperature for PT (negative temperature coefficient). A positive temperature coefficient is desirable for paralleling devices because a hot device will conduct less current than a cooler device, so all the parallel devices tend to naturally share current. In order to enable this current share in future parallel connection of IGBTs, the NPT type IGBT will now be chosen for implementation in the GI.

E.1.2 Statical IGBT model

After the physical model, now the electrical model of the IGBT, shown in figure E2.a, can be made. Figure E2.b shows a simplified schematic of an IGBT. An interesting aspect is the fact that what is called the “collector” is actually the emitter of the internal PNP. Therefore, the IGBT is comparable with a MOSFET driving an emitter follower. Although this model is capable of producing the basic function of an IGBT, refinements are required for more accurate modeling and to emulate the non-linear capacitance and breakdown effects.

A more adequate model, in figure E2.c, shows one of the drawbacks of the IGBT. A parasitic NPN bipolar transistor exists within all N-channel power MOSFETS and consequently all N-channel IGBTs. The base of this transistor is the body region, which is shorted to the emitter to prevent it from turning on. However, the body region has a certain resistance, called body region spreading resistance, as shown in Figure 3. The P-type substrate and drift and body regions form the PNP portion of the IGBT. The PNPN structure thus forms a parasitic thyristor.

If the parasitic NPN transistor ever turns on and the sum of the gains of the NPN and PNP transistors are greater than one, latchup occurs. This means that the parasitic thyristor turns on, and the IGBT cannot be turned off by the gate and may be destroyed due to over-current heating. This is not the only way to activate this parasitic thyristor, as high $dv/dt$ during turn-off combined with excessive collector current can also turn on the parasitic NPN transistor. This is called dynamic latchup, which is actually what limits the safe operating area since it can happen at a much lower collector current than the previously mentioned static latchup, as it depends on the turn-off $dv/dt$.

---

Fig. E.2. Model overview of the IGBT. a) general symbol, b) working principle, c) inclusion of parasitic components, d) inclusion of parasitic capacitances.
**E.1.3 Dynamical IGBT model**

Since the IGBT losses are severely affected by switching losses, the dynamical behavior of the IGBT also has to be modeled, to calculate losses. Figure E2.d shows the equivalent IBGT model that includes the capacitances between the terminals. Input, output, and reverse transfer capacitances are combinations of these capacitances. Each of these capacitors has its own influence on the switching behavior and often is known under different names than named here. Datasheets often use other names and test procedures. Specific test conditions to measure these capacitances usually are specified in the datasheet. For some general parasitic components, the effect on the dynamical behavior is now investigated.

**E.1.3.1 Miller Capacitance**

The Miller Capacitance, also known as the reverse transfer capacitance, is measured between the collector and gate terminals with the emitter connected to ground. The reverse transfer capacitance is equal to the gate to collector capacitance.

\[ C_{GC} = C_{res} = C_{Miller} \quad (E1.1) \]

Since the current, flowing through the gate usually is relatively low, the Miller capacitance is one of the major parameters affecting voltage rise and fall times during switching. As will be shown later, a low value for \( R_{off} \) in the gating network will minimize these effects.

**E.1.3.2 Input Capacitance**

The input capacitance \( C_{ies} \) is the capacitance measured between the gate and emitter terminals with the collector shorted to the emitter for AC signals. \( C_{ies} \) is made up of the gate to collector capacitance, \( C_{GC} \), in parallel with the gate to emitter capacitance, \( C_{GE} \):

\[ C_{ies} = C_{GE} + C_{GC} \quad (E1.2) \]

The input capacitance must be charged to the threshold voltage before the device begins to turn on, and discharged to the plateau voltage before the device begins to turn off. Therefore, the sum impedance of the drive circuitry and \( C_{ies} \) has a direct relationship to the turn on and turn off delays.

In figure 5 of addendum c4, the datasheet of the HGTG20N60B3D, the maximum value for the input capacitance \( C_{ies} \) is equal to: 4.5nF. Increasing the switch current will decrease this value.

**E.1.3.3 Output Capacitance**

The output capacitance \( C_{oes} \) is measured between the collector and emitter with the gate shorted to the emitter for AC voltages. \( C_{oes} \) is made up of the collector to emitter capacitance, \( C_{CE} \), in parallel with the gate to collector capacitance, \( C_{GC} \):

\[ C_{oes} = C_{CE} + C_{GC} \quad (E1.3) \]

For soft switching applications, \( C_{oes} \) is important, as it can affect the resonance of the circuit. In the current circuit, controlled soft switching however is not yet an issue.

In figure 5 of addendum c4, the datasheet of the HGTG20N60B3D, also the maximum value for the output capacitance \( C_{oes} \) can be found: \( C_{oes}=2.3nF \). Increasing the switch current will again decrease this value.
E.1.4 IGBT switching losses

Now both the IGBT model and the gating circuit have been explained, it is time to start looking at the switching losses. As these losses can be calculated in several ways, both the analytical and the datasheet calculation will be presented.

E.1.4.1 Turn on and turn off switching loss

The turn-on switching transients of an IGBT, switching an inductive load as in figure E3.a are shown in figure E3.b. With a gate voltage applied across the gate to emitter terminals of the IGBT, the gate to emitter voltage rises up in an exponential fashion from zero to $V_{GE(th)}$ due to the circuit gate resistance $R_G$ and the gate to emitter capacitance $C_{GE}$. The Miller capacitance $C_{GC}$ effect can be divided into two stages. Directly after setting a positive value on the gate, the voltage $V_{GC}$ will drop to a lower level, but will not directly fully reduce. This will only happen once the current through the IGBT has gained its full value. Due to the large total voltage drop, $V_{GC}$ has to make, this capacitance will be important for the gate circuit design when high collector to emitter voltages are applied.

Beyond $V_{GE(th)}$, the gate to emitter voltage continues to rise as before and the drain current begins to increase linearly as shown in figure E3.b and E3.d. Due to the clamp diode, the collector to emitter voltage remains at $V_{AC}$ as the IGBT current is less than $I_0$. Once the IGBT is carrying the full load current but is still in the active region, the gate to emitter voltage becomes temporarily clamped to $V_{GE,Io}$ which is the voltage required to maintain the IGBT current at $I_0$. At this stage, the collector to emitter voltage starts decreasing in two distinctive intervals $t_{tr}$ and $t_{tf}$. The first time interval corresponds to the traverse through the active region while the second time interval corresponds to the completion of the transient in the ohmic region. During these intervals, the Miller capacitance becomes significant where it discharges to maintain the gate to source voltage constant. When the Miller capacitance is fully discharged, the gate to emitter voltage is allowed to charge up to $V_G$ and the IGBT goes into deep saturation. The resultant turn on switching losses are also shown in figure E3.b.

The turn-off switching transients of an IGBT with an inductive load are shown in figure E3.c and E3.d. When a negative gate signal is applied across the gate to emitter junction, the gate to emitter voltage starts decreasing in a linear fashion. Once the gate to emitter voltage drops below the threshold voltage $V_{GE(th)}$, the collector to emitter voltage starts increasing linearly. The IGBT current remains constant during this mode since the clamp diode is off. When the collector to emitter voltage reaches the dc input voltage, the clamp diode starts conducting and the IGBT current falls down linearly. The rapid drop in the IGBT current occurs during the time interval $t_{tf}$ which corresponds to the turn-off of the MOSFET part of the IGBT.
The tailing of the collector current during the second interval $t_{f2}$ is due to the stored charge in the n-drift region of the device. This is due to the fact that the MOSFET is off and there is no reverse voltage applied to the IGBT terminals that could generate a negative drain current so as to remove the stored charge. The only way for stored charge removal is by recombination within the n-drift region. Since it is desirable that the excess carriers’ lifetime be large so as to reduce the on-state voltage drop, the duration of the tail current becomes long. This will result in additional switching losses within the device. This time increases also with temperature similar to the tailing effect in BJTs. Hence, a trade off between the on-state voltage drop and faster turn-off times must be made.

E.1.4.2 Switching losses calculated with datasheets

IGBT datasheets also usually give a standard value for turn on and turn off losses $E_{on}$ and $E_{off}$. $E_{on}$ and $E_{off}$ can be calculated as the instantaneous power loss $P_{loss} = I_{CE} \times V_{CE}$ during turn on and turn off. All tail losses have to be included in the calculation of $E_{off}$, therefore this calculation only ends once the collector current reaches zero. According to the datasheet of the HGTG20N60B3D, which can be found in addendum C4, these variables are respectively $475 \mu J$ and $1050 \mu J$. These losses are only applicable when the IGBT is hard switched at a collector emitter voltage of 480V.

However, looking further in the datasheet, two graphs can be found, showing the relation between the output current and the switching losses. Since in this application, the IGBTs will be used at a maximum RMS current of 16A, leading to a peak of 22.6A, also only for this area the switching losses have to be calculated. Due to the fact that the switching losses show a relatively linear behavior in relation to the collector to emitter current, the maximum average switching losses can be found in the graph, while looking at the maximum RMS current value. Now, the values for turn on and turn off losses $E_{on}$ and $E_{off}$ appear to be relatively $350 \mu J$ and $750 \mu J$. The linear relationship between $I_{CE}$ and $E_{on}$ and $E_{off}$ ensures a lower loss at lower RMS output currents.

The maximum total switching losses should therefore be:

$$ P_{\text{switching loss}} = f_s \cdot (E_{on} + E_{off}) $$

$$ P_{\text{switching loss}} = 25 \cdot 10^{-3} \left( 350 \cdot 10^{-6} + 750 \cdot 10^{-6} \right) $$

$$ P_{\text{switching loss}} = 27.5W $$

It is clear however that these losses cannot be that high, as this would already mean a total switching loss of $16 \times 38.1 = 610W$, meaning a power loss of 20%.

This is caused by the fact that the total voltage, which needs to be switched, is lower than the value in the datasheet. Not only is this caused by the lower average voltage, there is also a more important reason: when turning off one set of switches, the leakage inductance of the main transformer will change the polarization of the transformer voltage. This enables a form of soft switching, as the voltage over the IGBT, which needs to be switched on a fragment later, is already drastically reduced.

Since these switching losses are caused by the fact that a current starts flowing through the IGBT, while $V_{CE}$ is not low yet, a reduction of $V_{CE}$ before turning it on will therefore also drastically reduce these switching losses.

Of course, the IGBT switching loss can also be calculated by hand. However, due to the fact that the datasheets do not supply enough information regarding the delay between...
the ascending $V_{CE}$ and descending $I_{CE}$, during turn off or descending $V_{CE}$ and ascending $I_{CE}$, during turn on, this is only calculatabl after performing measurements on the turn on and turn off moments.

**E.1.5 IGBT and diode conduction losses**

The model of figure E2.a shows that in on-state, the IGBT will conduct as a transistor and thus will cause a voltage drop. The advantage of this voltage drop however is the fact that it is relatively constant for varying loads. Other losses, occurring in the circuit are caused by the voltage drop over the diode. These conduction losses are

Looking at the datasheet of the HGTG 20N60B3D in addendum C4, the value for $V_{CE(sat)}$ gives an indication on these losses for the IGBT. However, a more accurate estimation can be made by looking at figure 2 in this datasheet. Due to the applied value for $V_{GE}$ of 12V and the maximum average current of 16A, the maximum value for $V_{CE}$ can be estimated at 1.6V. Therefore, the conduction loss in the IGBTs can estimated at full power at a value of $16A \times 1.6V = 25.6W$.

Since in series with each IGBT an internal diode of another IGBT package is conducting, the voltage drop over this diode has a typical value of 1.3V at working temperature. The loss can be estimated at $1.3V \times 16A = 20.8W$. This loss is calculated at working temperature, due to the negative temperature coefficient of the internal diode, this loss will be higher at lower temperatures.

Due to the symmetrical, 50Hz grid voltage, each conducting IGBT module will be used 50% of the time as IGBT and 50% of the time as diode. Due to the switching principle, each IGBT is only conducting 50% of the time. The average loss per IGBT is therefore equal to:

$$P_{\text{IGBT loss av}} = 50\% \times \frac{P_{\text{IGBT}} + P_{\text{diode}}}{2}$$

$$P_{\text{IGBT loss av}} = 50\% \times \frac{25.6 + 20.8}{2} = 11.6W$$

Since the maximum power dissipation per module is equal to 60W, this loss power per device is well below the maximum ratings.

In conduction mode, each bridge contains two active IGBTs and two active diodes. For the Dual Active Bridge concept, two H-bridges are used. Therefore, the theoretical maximum total conduction losses, caused by the IGBTs and diodes, can be calculated with equation E1.6.

$$P_{\text{conduction loss}} = 2 \times 2 \times V_{CE(sat)} \times I_{N_{max}} + 2 \times 2 \times V_{AK} \times I_{N_{max}}$$

$$P_{\text{conduction loss}} = 2 \times 2 \times 1.6 \times 16 + 2 \times 2 \times 1.3 \times 16$$

$$P_{\text{conduction loss}} = 185.6W$$

This theoretical switch based power loss is quite accurate, as will be shown in the measurement results, as presented in addendum B2.
E.2 Parallel connection of IGBTs

When connecting IGBT switches in parallel, it is necessary to properly manage certain element characteristics. Otherwise, a current sharing imbalance may occur, depending on the characteristic distribution between the parallel connected modules.

E.2.1 Current imbalance due to IGBT characteristics

IGBT $V_{CE(sat)}$ and diode $V_{forward}$ distribution. A difference in the output characteristics of two IGBT modules connected in parallel can cause a current imbalance, as can be seen in figure E4. Therefore, both IGBTs have to be equal.

In general, the voltage over both parallel switches $S_1$ and $S_2$ can be calculated with the equations E2.1 and E2.2.

$$V_{S1} = V_{forward1} + V_{CE1}$$  \hspace{1cm} (E2.1)
$$V_{S2} = V_{forward2} + V_{CE2}$$  \hspace{1cm} (E2.2)

Since the sum of the forward voltage drop over the diode and the IGBT can be estimated by the graph in figure E4, the formulas E2.3-E2.5 can then be used to estimate the voltage over the switches:

$$R_{S1} = \frac{V_a}{(I_{C2} - I_{C1})}$$  \hspace{1cm} (E2.3)
$$R_{S2} = \frac{V_b}{(I_{C2} - I_{C1})}$$  \hspace{1cm} (E2.4)

$$V_{S1} = V_a + R_{S1} \cdot I_{C1}$$  \hspace{1cm} (E2.5)
$$V_{S2} = V_a + R_{S2} \cdot I_{C2}$$  \hspace{1cm} (E2.6)

Based on these formulas, the current distribution over both switches can be calculated, using the equations E2.7 and E2.8.
As can be seen, differences in the values of $V_{CE(sat)}$ have a major influence on the current balance. Therefore, always the same type of IGBT should be used, preferably from the same production batch to minimize differences in characteristics.

### E.2.2 Current imbalance due to distribution of wiring resistance.

If the PCB connection of both modules differs in length, width or thickness, a difference in path resistance might cause different values for the voltage over the series connections of IGBTs and diodes. This will then influence the current balance, as shown in figure E.4. The resistance of the IGBT connections and copper layers will lessen the parity of the slope of the IGBT output characteristics, which will cause the collector current to decrease. Therefore, if $R_1 > R_2$, the slope of the $S_1$ output characteristics will decrease, resulting in $I_{C1} < I_{C2}$ and therefore in a current sharing imbalance.

In order to reduce this imbalance, it is necessary to design the wiring as short, wide, thick and uniform as possible.

### E.2.3 Current imbalance due to temperature differences

Another important factor, influencing the current balance, is the thermal behavior of the IGBT-diode modules. If a module has a higher temperature than the parallel connected module, the current through the colder component should increase, thus preventing a thermal run-away from the hotter module.

This characteristic property is called the thermal coefficient and is shown in figure E.5. If it is positive, the value for $V_{CE(sat)}$ at a certain collector current will increase when the die temperature increases. This automatically implements a proper current division between the parallel connected modules.

If a IGBT-module has a negative thermal coefficient, $V_{CE(sat)}$ will decrease when the die temperature increases. As a result, the collector current will increase, thus heating up the device even more and possibly leading to destruction of the module.

\[
I_{C1} = \frac{V_2 - V_1 + R_{S2} \cdot I_{SUM}}{R_{S1} + R_{S2}} \quad (E.2.7)
\]

\[
I_{C2} = \frac{V_1 - V_2 + R_{S1} \cdot I_{SUM}}{R_{S1} + R_{S2}} \quad (E.2.8)
\]
As a result, the temperature difference between the parallel connected IGBT packages should be as small as possible. This goal can be reached by taking care of three factors:

- Proper heat transfer between the module and the heat sink
- Implementation of a heat sink with a low thermal resistance
- Thermal connection of parallel connected devices on the same heat sink, at a short distance of each other.

**E.2.4 Oscillation of parallel connected IGBTs**

When IGBTs are connected in parallel, the uniformity of the wiring is very important. Otherwise, concentrated transient currents could occur on the device having a shorter wiring path during the switching time. This can cause device destruction or degrade the long-term IGBT reliability.

Also, unbalanced wiring can also bring the overall circuit inductance out of balance between the devices. As a result, a high $di/dt$ during switching moments can cause different voltages over the devices. This can lead again to an abnormal oscillating current, also leading to possible device destruction.
E.3 Overview IGBT characteristics

The following IGBT overview, in table E-1 is based on the corresponding datasheets, available on the manufacturers’ websites. The diode currents are only mentioned, if they differed from the IGBT specifications, thus possibly creating a problem.

<table>
<thead>
<tr>
<th>Type</th>
<th>Manufacturer</th>
<th>Vces (V)</th>
<th>Ic (A)</th>
<th>Ifw diode (A)</th>
<th>Vce(sat) (V)</th>
<th>Eon (mJ)</th>
<th>Eoff (mJ)</th>
<th>Esw (mJ)</th>
</tr>
</thead>
<tbody>
<tr>
<td>APT30GP60BDFl</td>
<td>Advanced Power Technologies</td>
<td>600</td>
<td>49 @110°C</td>
<td>2.2</td>
<td>0.26</td>
<td>1.0</td>
<td>1.3</td>
<td></td>
</tr>
<tr>
<td>APT30GP60BDQ1G</td>
<td>Advanced Power Technologies</td>
<td>600</td>
<td>49 @110°C</td>
<td>2.2</td>
<td>0.26</td>
<td>0.56</td>
<td>0.82</td>
<td></td>
</tr>
<tr>
<td>APT40GP60B2DF2</td>
<td>Advanced Power Technologies</td>
<td>600</td>
<td>62 @110°C</td>
<td>2.2</td>
<td>0.39</td>
<td>1.6</td>
<td>2.0</td>
<td></td>
</tr>
<tr>
<td>APT50GT60BRDQ1G</td>
<td>Advanced Power Technologies</td>
<td>600</td>
<td>52 @110°C</td>
<td>2.0</td>
<td>1.0</td>
<td>3.2</td>
<td>4.2</td>
<td></td>
</tr>
<tr>
<td>IXDH35N60BD1</td>
<td>IXYS</td>
<td>600</td>
<td>35 @90°C</td>
<td>25 @90°C</td>
<td>2.2</td>
<td>1.6</td>
<td>0.8</td>
<td>2.4</td>
</tr>
<tr>
<td>IXGH40N60C2D1</td>
<td>IXYS</td>
<td>600</td>
<td>40 @110°C</td>
<td>2.0</td>
<td>0.60</td>
<td>0.50</td>
<td>1.1</td>
<td></td>
</tr>
<tr>
<td>IXRH40N120</td>
<td>IXYS</td>
<td>1200</td>
<td>35 @90°C</td>
<td></td>
<td>2.3</td>
<td>19</td>
<td>1.0</td>
<td>20</td>
</tr>
<tr>
<td>IHW40N60T</td>
<td>Infineon</td>
<td>600</td>
<td>40 @100°C</td>
<td>20 @100°C</td>
<td>1.55</td>
<td>1.4</td>
<td>1.4</td>
<td>2.8</td>
</tr>
<tr>
<td>IKW40T120</td>
<td>Infineon</td>
<td>1200</td>
<td>40 @100°C</td>
<td></td>
<td>2.1</td>
<td>3.3</td>
<td>3.2</td>
<td>6.5</td>
</tr>
<tr>
<td>IKW50N60T</td>
<td>Infineon</td>
<td>600</td>
<td>50 @100°C</td>
<td></td>
<td>1.5</td>
<td>1.8</td>
<td>1.8</td>
<td>3.6</td>
</tr>
<tr>
<td>IKW75N60T</td>
<td>Infineon</td>
<td>600</td>
<td>75 @100°C</td>
<td></td>
<td>1.5</td>
<td>2.9</td>
<td>2.9</td>
<td>5.8</td>
</tr>
<tr>
<td>IRGP35B60PD</td>
<td>International Rectifier</td>
<td>600</td>
<td>34 @100°C</td>
<td>15 @100°C</td>
<td>1.85</td>
<td>0.33</td>
<td>0.41</td>
<td>0.71</td>
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<tr>
<td>IRG4PC40UD</td>
<td>Infineon</td>
<td>600</td>
<td>20 @100°C</td>
<td>15 @100°C</td>
<td>1.7</td>
<td>0.35</td>
<td>0.71</td>
<td>1.06</td>
</tr>
<tr>
<td>HGTG20N60A4D</td>
<td>Fairchild</td>
<td>600</td>
<td>40 @110°C</td>
<td></td>
<td>1.8</td>
<td>0.12</td>
<td>0.33</td>
<td>0.45</td>
</tr>
<tr>
<td>HGTG30N60A4D</td>
<td>Fairchild</td>
<td>600</td>
<td>60 @110°C</td>
<td></td>
<td>1.8</td>
<td>0.28</td>
<td>0.45</td>
<td>0.72</td>
</tr>
<tr>
<td>HTGT20N60B3D</td>
<td>Fairchild</td>
<td>600</td>
<td>20 @110°C</td>
<td>20 (RMS) @110°C</td>
<td>2.0</td>
<td>0.48</td>
<td>1.0</td>
<td>1.48</td>
</tr>
<tr>
<td>HTGT30N60B3D</td>
<td>Fairchild</td>
<td>600</td>
<td>30 @110°C</td>
<td>25 (RMS) @110°C</td>
<td>1.45</td>
<td>1.3</td>
<td>1.6</td>
<td>2.9</td>
</tr>
<tr>
<td>STGY40NC60VD</td>
<td>ST Microelectronics</td>
<td>600</td>
<td>50 @100°C</td>
<td>30 (RMS) @25°C</td>
<td>1.9</td>
<td>0.64</td>
<td>2.04</td>
<td>2.68</td>
</tr>
</tbody>
</table>
E.4 IGBT test definition and results

One of the most important components of the Mass GI are the IGBTs. Not only do they have to be fast and accurate, they also should be robust and reliable. Therefore, the following paragraphs will describe a test, with which it is easily possible to gain insight in a number of specific IGBT characteristics. Further more, several IGBTs are compared in order to find the correct IGBT for the Mass GI.

Eventually, a problem, which presented itself at Mastervolt in November 2006, will be explained: the rising market in fake, unspecified components. Also the test results of a number of batches of these fake batches will be presented.

E.4.1 IGBT component replacement

Due to changes in component portfolio of semiconductor manufacturers, changes in design specifications or the presence of unofficial, unspecified copies of the genuine product, it can be necessary to look for a replacement of the currently used component.

When looking for a replacement for an IGBT a number of parameters should be taken into account. A general checklist in order to sort out a number of candidates contains the following characteristics:

- The package usually should be equal.
- The IGBT should be able to handle the same or more current
- The internal diode, if present, should be able to handle the same or more current
- The maximum voltage ratings for $V_{CE}$ and $V_{GE}$ should be equal or higher.

However, more data is needed, in order to decide if a certain type of IGBT is applicable for the job. These characteristics mostly handle the dynamical behavior of the component. The most important dynamical characteristics are:

- Turn on delay time
- Turn off delay time
- Turn on rise time
- Turn off descend time

These characteristics, once the component is implemented in the circuit, can be influenced by a correct choice of the gating circuit.

E.4.2 Setting up a test procedure

In order to set up a test procedure which can simply characterize an IGBT or diode, first a number of general details must be checked. Secondly, some static component characteristics must be tested and compared with the datasheet results. A third test can then give more insight in the dynamical response of the IGBT. Combined, these tests can answer the question if the measurements can ensure the buyer that a product is genuine.

E.4.2.1 Visual check test

The main question, which has to be answered first, is: Do the samples look like the original component? Cheap copies have different dye sizes, can be mounted in inferior cases or can contain an error in the date and product code, stamped on the case.
Therefore, first of all, the stamp on the front of the component should be checked. Genuine examples will show a code, which usually tells more about the date and place of dye production and packaging. Component datasheets will usually inform customers on the structure of these codes.

When the component is suspected to be a copy of a genuine product, also the case should be checked. The length of the connectors can differ from the official specifications, or the case can be mounted differently, showing shape differences. The component code can be checked with the package marking conventions, in addendum C5.

Also the IGBT and diode dye should then be checked, as inferior copies often use dyes of different sizes than the original component. This can be checked by carefully suppressing a force on the sides from the case in a working bench. The case will crack, thus revealing the dyes.

E.4.2.2 Conduction loss test
In order to determine some general component aspects, first of all datasheet of the original component can be checked for all necessary data. The IGBT, needed in the Mass GI, for example the HGTG20N60B3D, will need to contain both an IGBT and an anti-parallel diode. Therefore, they will both need to be characterized.

E.4.2.2.1 Influence of temperature variation
The temperature coefficient of the IGBT and the diode will be of great importance on the component efficiency. As mentioned in the paragraph on parallel connection of IGBTs, this application prefers IGBTs with a positive temperature coefficient. This means that \( V_{CE(sat)} \) increases together with the temperature. Therefore, the following measurements of \( V_{CE(sat)} \) and \( V_{AK} \) will have to be done at heat sink temperatures of both \( 25^\circ C \) and \( 80^\circ C \), as this is the maximum operating temperature of the heat sink.

E.4.2.2.2 Measurement of \( V_{CE(sat)} \) of the internal IGBT
This test can simply determine one of the characteristics of an IGBT, which can also be found in every IGBT datasheet. \( V_{GE} \) must be set at a standard high gate voltage. In this case, a value of \( 12V \) is chosen. For \( I_E \), the following values have to be chosen, which represent the whole operating area. In this case, the following values are set: \( 1A, 2A, 5A, 10A, 15A \) and \( 20A \).

E.4.2.2.3 Measurement of \( V_{AK} \) of the internal diode.
The values from this test can normally also be compared with a graph in the datasheet. For \( I_A \), the following values have to be chosen, which represent the whole operating area. In this case, the following values are set: \( 1A, 2A, 5A, 10A, 15A \) and \( 20A \).

As shown in this overview, four sets of each 6 measurements should be done. After having done these measurements, a graphic overview of the test results can be compared with the values, found in the datasheet or the data of the replaceable component.

E.4.2.2.4 Conduction loss test setup
For the previously discussed tests, the test setup, shown in figure E6.a, will be used. For the source power, a Delta power supply is used as current source, with its output voltage limited at \( 12V \). Due to this implementation, the power source in figure E6.a also is shown as a current source. However, it should be mentioned that this source of course is limited in output voltage, thus permitting the series connected switch. The gate voltage of \( 12V \) is
Addendum

fed to the gate of the IGBT through a gate resistance of 10Ω, as this is a standard value, used in many datasheets for IGBT evaluation.

The diode and inductor are not necessary, but will be used in the switching loss test. In order to keep the setup as simple as possible, they are also already present in these schematics. Due to the fact that the voltages are measured, while applying a DC current, the inductor will not influence the measurement.

The switch S1 is added in order to be able to quickly turn on and off the current and therefore makes sure that the IGBT temperature, especially at high DC currents, will not rise too fast, in order to be able to do the 25°C measurements. Therefore, the IGBT will also be mounted on a small heat sink, equipped with a thermo couple, connected digital thermometer. Further more, a 12V fan ensures that the IGBT will keep it’s temperature.

The 80°C measurements are performed, after heating up the device by turning off the fan and turning on the 20A current through the device. This way, the heat sink will heat up to the desired temperature. In order to reach a stable situation at 80°C, the fan should be empowered with a few volts once the heat sink reaches a temperature of 70°C.

In order to keep the temperature of the IGBT on a constant level, it is mounted on a heat sink, as shown in figure E6.r.d. A practical implementation of the measurement setup is shown in figure E6.r.c.

E.4.2.3 Dynamic response test

Due to the fact that each IGBT will cause switching losses at turn on and turn off, the response to a compact repetitive pulse can give a quite accurate overview on the dynamical behavior of a device. Characteristics like tail currents and delay or rise times then can easily be shown. In order to get an accurate image, the frequency, at which the signal is repeated, should be taken the same as the standard operating frequency.

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Fig. E.6. IGBT test setups. a) conduction loss test, b) switching loss test, c) built up test circuit, d) IGBT sample on heat sink.
E.4.2.3.1 Influence of temperature variation

An important factor in also the dynamical response is the influence of the heat sink temperature. In order to gain insight in this relation, the pulse should be registered at 25°C and at 80°C. If the differences for the various delays and rise times are significant between these temperatures, they might cause gating problems for certain designs. An H-bridge at full load might for example at working temperature suddenly be bothered with shoot-through, due to an increased turn off delay. Due to this increased delay, the formerly turned on switches will not yet have turned off when the other switches already turn on.

E.4.2.3.2 Calculation of pulse width

The inductor energy, which is constantly refilled by switching on the IGBT, is dissipated in two ways. First of all, this loss is caused by the resistance of the inductor and the wiring. Secondly, the voltage drop over the diode results in energy loss. Having defined the main reasons for power loss, it is now possible to calculate the on-time per period for the IGBT.

Taking into account a DC voltage of 300V, due to the limitations of the applied DC-voltage source, an average current of 20A, an internal winding and wire resistance of 75mΩ and an average diode voltage drop of 1,5V, this would ideally mean that the maximum relative on-time should be 1/100th of a period. At a period of 40μs, this would then be 400ns. However, due to losses in wires and due to the fact that the voltage over the inductor will not directly be 300V, but will slowly rise to a value just below, a slightly higher on-time for the pulse could be necessary.

E.4.2.3.3 Dynamic response test setup

The general test setup is shown in figure E6.b. As can be seen, the test setup did not change. In order to be able to switch a high current, but at the same time reach a high voltage drop for VCE, the IGBT is connected in series with an inductor. In order to maintain a current path for the inductor, a diode is connected in reverse direction in parallel with the inductor.

This way, turning off the IGBT will not harm this device or the inductor, as the current will now flow in opposite direction through this diode. Due to the fact that this diode should have hyperfast response times, the RHRP3060 has been chosen. Not only does it have the right characteristics, it is also the internal diode of the HGTG20N60B3D, and therefore easily obtainable. In order to be able to use this internal diode, the gate of the IGBT has been connected to the emitter.

Now, it is only a task to shortly turn on the IGBT, to reload the inductor. This is done by a standard gate driver, as mentioned in addendum D2. The gate driver is fed by a multifunctional pulse source.

Again, the IGBT is still mounted on a small heat sink, equipped with a thermo couple, connected digital thermometer. Further more, a 12V fan ensures that the IGBT will keep its temperature at 25°C.

The 80°C measurements are performed, after heating up the device by turning off the fan and turning on the pulse for a while. This way, the heat sink will heat up to the desired temperature. In order to reach a stable situation at 80°C, the fan should be empowered with a few volts once the heat sink reaches a temperature of 70°C.
**E.4.3 Comparison of IGBTs for the Mass GI**

In order to make the right choice for the IGBTs, which will be applied in the Mass GI, a number of IGBTs will be tested. Of course, the number of IGBTs, which can be used in the schematics, is high, as shown in a component search based on datasheets, as shown in addendum E3. However, aspects as obtainability and costs are also of influence.

Therefore, the scope of candidates narrowed down to a smaller number, from which eventually four of them have been chosen for a comparative evaluation. The procedure of this evaluation is given in addendum E4.

**E.4.3.1 Conduction losses**

First of all, the relation of the IGBT between $I_{CE}$ and $V_{CE(sat)}$ at a heat sink temperature of 25°C, room temperature, and 80°C, the maximum operating temperature, have been tested. The results are given in the graphs in figure E7.

![Graph E7](image-url)

*Fig. E7. Variation of $I_{CE}$ for the internal IGBT of various IGBT packages at 25°C and 80°C.*

Secondly, the relation of the diode between $I_{AK}$ and $V_{AK}$ at a heat sink temperature of 25°C, room temperature, and 80°C, the maximum operating temperature, have been tested. The results are shown in figure E8.

![Graph E8](image-url)

*Fig. E8. Variation of $I_{AK}$ for the internal diode of various IGBT packages at 25°C and 80°C.*

As can be seen, the IGBT in the IRGP35B60PD has low losses for low emitter currents. However, for higher currents at higher temperatures, this device will have the highest conduction losses. Both the IGBT and the diode of the IHW40N60T clearly have the lowest conduction losses and therefore seem quite interesting.

The HGTG20N60B3D has a relatively good IGBT, and has a comparable diode. Due to the fact that the temperature in the Mass GI usually will be quite high, the lower efficiency at 25°C is not a big problem.
E.4.3.2 Switching losses

In order to gain insight in the switching losses, the response of the IGBTs to a short 20A pulse at a 300V supply voltage has been tested. Since certain characteristics can be temperature dependant, the devices have been tested at 25°C and 80°C heat sink temperature.

First of all, the HGTG20N60B3D has been tested. A number of parameters can be read from the graphs in figure E9. First of all, an important factor is the short turn on and turn off delay time, which are also relatively undependable of the heat sink temperature.

The dynamical response of the IRGP35B60PD is shown in figure E10. As can be seen, the test results show similar behavior as the HGTG20N60B3D, however the turn off delay is slightly higher.
Figure E11 shows the test results of the IRG4B40UD. As can be seen, the switching losses are slightly higher, compared to the IRGP35B60PD. Also, the effect of a rising temperature is visible in the picture. Turn off losses increase as soon as the dye heats up.

Finally, figure E12 shows the test results of the IHW40N60T. As can be seen, the turn on and turn off losses are higher. Also, the turn off delay is much higher. It is clear that this IGBT is meant for lower switching frequencies. The relatively high difference between turn on and turn off delay will result in a necessary dead time implementation when implemented in H-bridges, in order to prevent shoot through.

**E.4.4 Conclusion**

Combining all measurements, a few comments can be made. First of all, it is clear that one should never look at only one of the previously performed measurements. The IHW40N60T is a clear example: even though the conduction losses are low, the switching losses and the turn off delay are high. This will cause a shoot through in a H-bridge, if no extra dead time is implemented. If this problem is overcome, it might however be an interesting component.

Further more, the dynamic characteristics of the HGTG20N60B3D are similar to the the characteristics of the types of International Rectifier. However, when conducting high currents, the last two IGBTs will have a higher value for $V_{CE(sat)}$. The caused higher thermal losses might lead to over heating of the device. Therefore, the HGTG20N60B3D will be chosen to be the IGBT, used in the Mass GI.
E.5 Irregularities in IGBT supply chains

In order to be sure that IGBT packages contain the correct IGBT and diode dye, it appeared that only checking the date and product code on a package can not always be enough. At Mastervolt, this fact became reality, after a number of unexplainable breakdowns of apparatus during a final factory test. Circuit evaluation brought up that the fault could be found in the IGBT packages, which overheated and even exploded.

Fortunately, a meeting with an application engineer and a sales manager of Fairchild was planned for the same week the problems appeared, so Mastervolt was able to get a quick explanation for the risen problems. Due to the heavily increased market for power supplies, for example for LCD-TVs, a shortage on the market has developed of not only IGBT and diode dyes, but even also of TO247 packages, the standard package for many IGBTs and MOSFETs. Due to this shortage, delivery times have suddenly risen from 15 weeks to 45 weeks for companies like Mastervolt. Logically, this would result in a unworkable situation, in which production companies can not manufacture any products any more for months, seriously threatening the availability of several Mastervolt products.

Luckily, components like IGBTs can not only be bought through official dealers, which are directly supplied from the factories. It is also possible to search for component stacks through brokers. These stacks are built up at warehouses and factories, for example due to change in component choice or over dimension of annual production plannings of manufacturers.

However, in this path, a problem rises: due to the shortage, not only left stocks are offered through these brokers. Also, fake components, manufactured in mostly Asian factories, are offered. Since the casing and the stamped product information look the same, a visual check will not satisfy in order to gain assurance on the authenticity of the component. Therefore, the previously explained measurement procedure is used on these samples, in order to give an accurate answer on the question if a certain product is genuine or fake.

E.5.1 Component evaluation of Batch “N52AF”

After the discovery that the IGBT batch “N52AF” did not meet its specifications, the previously described IGBT test has been used to characterize these IGBTs. The test batch of these IGBTs contained three samples.

The IGBT case contained the Fairchild logo and the batch number “N52AF” on the first line and the standard inscription “G20N60B3D” on the second line. The batch number N52AF could mean that they are manufactured in April 2005, by ChipPAC, a subcontractor.

In the Mastervolt laboratory, these samples have been given the names FAKE1, FAKE2 and FAKE3.
Fig. E.13. Comparison of $V_{CE(tau)}$ at various currents and temperatures. $V_{CE} = 12V$.

The results of the IGBT measurements are shown in figure E13. In order to be able to compare these results with the general characteristics of the HGTG20N60B3D, the values of a recent (NG21AK) and an older (NG34AB) standard batch of this IGBT have been added to the graph.

The results of the diode measurements are shown in figure E14. Also here, the values of two batches of the genuine IGBT have been added to the graph.

Fig. E.14. Comparison of $V_{AK}$ at various currents and temperatures. $V_{CA} = 0V$

It is clear that all samples contain different IGBT dyes. The diode dye of FAKE2 also is different from the other two diode dye.

In order to see the difference in dynamical behavior, also the pulse patterns have been tested again. The results are shown in figure E15. For comparison, figure E16 shows the standard graphs of batches NG21AK and NG34AB.

The graphs show big differences in switching times between all samples. Sample FAKE1 would simply be destroyed in a very short time due to the very high turn off loss.

Fig. E.15. Dynamical behavior for pulsed gate signal of samples FAKE1, FAKE2 and FAKE3. CH1=$V_{CE}$ (2V/div) CH2=$V_{CE}$ (50V/div) CH3=$I_C$ (5A/div)
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Fig. E.16. Dynamical behavior for pulsed gate signal of standard batches NG21AK and NG34AB. 
CH1=$V_{GE}$ (2V/div) CH2=$V_{CE}$ (50V/div) CH3=$I_C$ (5A/div)

It can be concluded that this batch will either thermally break down due to high conduction losses, as shown for FAKE1 and FAKE2, or high switching losses, as shown for FAKE3. In either case, the diversity in both IGBT and diode dyes disables the usage for any application, as the characteristics can never be known.

E.5.2 Component evaluation of Batch “S0512”

Due to the component shortage in the regular trading routes, brokers were asked to send samples of presumed old stocks of the HGTG20N60B3D. Of this batch of IGBTs Mastervolt obtained ten samples. The case contained the Fairchild logo and the batch number “s0512” on the first line, together with the inscription “HGTG 20N60B3D” on the second line. The date code could mean that they have been manufactured in December 2005 or January 2006, in Singapore. However, the second line is different from the standard Fairchild package marking conventions.

These samples have been given the batch name A, together with an individual batch number. Subsequently, they have been tested in two ways. First of all, four of them have randomly been chosen for a thorough evaluation. The results of both the IGBT as the diode measurements are shown in figure E17 and E18.

In order to be able to compare these results with the general characteristics of the HGTG20N60B3D, the values of a recent (NG21AK) and an older (NG34AB) standard batch of this IGBT have been added to the graph.

Fig. E.17. Comparison of $V_{CE(sat)}$ at various currents and temperatures. $V_{GE} = 12V$.

As it is possible to use both different IGBT as diode dyes, both components have been tested.
Addendum

Fig. E.18. Comparison of $V_{AK}$ at various currents and temperatures. $V_{CA} = 0V$

It is clear that one of the samples, the sample A9, clearly contains both a different IGBT as a different diode dye. The other IGBT and diode measurements however also show different values.

In order to see the difference in dynamical behavior, also the pulse patterns have been tested again. The results are shown in figure E19. They can be compared with the graphs of batches NG21AK and NG34AB in figure E16.

Fig. E.19. Dynamical behavior for pulsed gate signal of samples A1, A6, A9 and A10. CH1=$V_{CE}$ (2V/div) CH2=$V_{CE}$ (50V/div) CH3=$I_C$ (5A/div)

The graphs show again differences between A9 on one hand and A1, A6 and A10 on the other. The graphs also show a drastic increase in switching losses at turn-off.

E.5.3 Sample evaluation of HGTG20N60B3D Batch “HO14”

Mastervolt also received samples of another batch of IGBTs. In order to confirm the authenticity of this batch of IGBTs, several tests have been done on a number of samples. The device ordering code on the IGBT started with the logo of Intersil, the previous manufacturer of the HGTG20N60B3D, before it was taken over by Fairchild. Furthermore, the case contained the code “HO14” on the first line, the code “G20N60B3D” on the second line and “AN9” on the third line. This is interesting, as the take over of Intersil by Fairchild took place in March 2001, while this date code shows a date in July 2001.

At Mastervolt, these samples have been given the batch name C, together with an individual sample number.

Subsequently, one of them has been opened, as shown in figure E20. Left, the opened sample of batch “HO14” is shown, right a sample of a standard batch of the HGTG20N60B3D can be found. When comparing the two components, it is clear that both the diode dye and the IGBT dye differ in size.
Due to the fact that not only different IGBT, but also different diode dyes are used for batch "H014", both components have been electrically tested. The values of a recent (NG21AK) and an older (NG34AB) standard batch of this IGBT have been added to the graph.

The diode test went well, as can be seen in figure E21. The diode steady state conduction losses clearly match quite well with the original values.

However, when testing the conduction losses in the IGBT, the standard test setup appeared not to work. Even though a gate voltage of 12V was present, the IGBT would not start conducting at a DC supply voltage of $V_S = 12V$.

As this might be caused by a broken down IGBT, this device was checked, but proven to be a working example. A pulse pattern was applied to the gate signal, together with an increase of $V_S$ to the standard 300V. Surprisingly, these results were quite well, for both 25°C and 80°C, as shown in figure E22.

Now also an estimation of the conduction losses could be made, simply by zooming in on $V_{CE}$ near 0V. The results are shown in fig. E23 for both a sample of the test batch and a
genuine IGBT. As can be seen, the voltage drop $V_{CE}$ at the end of the pulse is much higher for the sample from this batch, in comparison with the standard test results.

An interesting result from these measurements is the fact that the IGBT somehow only will conduct at higher voltages for $V_s$. In order to get more insight in the IGBT characteristics, the IGBT conduction threshold level had to be found. The left graph in figure E24 shows the graphs of $V_{CE}$, $V_{CE}$ and $I_E$ for a levels of $V_s$=28V, just below the threshold. The right graph in figure 24 shows the response for $V_s$=30V, just above the threshold at which the IGBT starts conducting.

For comparison, two comparative graphs of a standard HGTG20N60B3D are shown in figure 25. As can be seen, even at a DC supply voltage of 2V, the genuine IGBT already conducts when a gate voltage of 12V is applied.

The right graph in figure 25 shows the response at $V_s$=30V. A comparison with the same supply voltage in figure 24 shows that implementation of this batch in applications, which must be able to also handle low voltages, is not possible. For applications, which use this device for PWM applications at high DC bus voltages and low currents, this device might be interesting.
It is clear that also the batch "HOI4" therefore can not replace the real HGTG20N60B3D in Mastervolt applications.

**E.5.4 Component evaluation of batch "D"**

One of the traders, who was asked to send samples of presumed old stocks of the HGTG20N60B3D, was supplier "X". They had a number of old batches. Of these batches of IGBTs, Mastervolt obtained six samples.

The cases contained the Harris logo and the numbers "H9702", "H9801", "H9806" and "H9810" on the first line, together with the inscription "G20N60B3D" on the second line. Harris was the first manufacturer of the HGTG20N60B3D, before Intersil took over Harris in August 1999. The date code means that they have been manufactured in 1997 and 1998, in the Philippines. It is interesting to finish a MSc in Electrical Engineering while testing a component, which was produced before I even started the BSc program.

These samples have been given the batch name D, together with an individual batch number. The previously discussed tests have been performed. The results of both the IGBT as the diode measurements are shown in figure E26 and E27.

Again, the values of a recent (NG21AK) and an older (NG34AB) standard batch of this IGBT have been added to the graph.
As can be seen, the values do not differ significantly. This gives the first indication that these batches are genuine.

In order to see the dynamical behavior, also the pulse patterns have been tested again. The results at 25°C are shown in figure E28 a to c. Figure E28.d shows the increased delay of sample D3 at 80°C. As can be seen, the previously measured increase of turn off delay can also be found in these graphs.

Combining both test results, the conclusion can be drawn that these components probably are genuine. However, due to the fact that this sample batch contained several different IGBTs, it is wise to check the eventual delivery for any more batch numbers, in order to validate the product details.
ADDENDUM

F

HF transformer design
F.1 Main transformer design for the 230V DAB converter

The main transformer design for the 230V Mass G1 was already given. In order to see if any improvements could be made to this design, the transformer design has been investigated. The calculations and results are also used for transformer designs in other chapters of this addendum.

F.1.1 Core shape

The task of the core material and structure in most magnetic components is to concentrate the magnetic field generated by the primary coil and carry it in a circuit to magnetically connect both the primary and all secondary coils. Many different shapes and material types are available, each optimal for a different purpose.

Since we are looking for a core which has a good coupling between the primary and secondary coils and which is also easily woundable, the E-65 core has been chosen. This core type uses two E-shaped core elements, which are pressed together for maximum coupling. As we will see, when looking at the winding materials, this type of core supports both foil windings as litz wires. This in contrary for example to torroidal cores, which only support normal copper and litz wires and are less well applicable for copper foil windings.

F.1.2 Core material

For the core material, several materials can be chosen, as long as they are ferromagnetic. Small 50Hz transformers or relay switches often have an iron or iron alloy core. For higher efficiency, laminated iron or iron alloy cores are used. As we will see later, the eddy current losses will be reduced by using laminated iron.

For the frequency range this galvanic insulator is in, also iron powder cores can be used. This type of core consists of small parts of iron, approximately 50-100μm in cross section, which are isolated individually in order to reduce eddy currents. These isolating layers however have a negative effect on the magnetic permeability of the material, which decreases the coupling between the primary and secondary windings.

This application however uses high frequencies, at which also a high level of efficiency is needed. Due to these demands, high quality material has to be used for the core. Therefore, the core has to be made of ferrite. Ferrite is a ceramic-like, electrically non-conductive ferromagnetic material that is a mix of iron oxides such as Hematite, Fe₂O₃, or Magnetite, Fe₃O₄, plus other metal oxides with manganese, zinc and nickel. After mixing these materials, they are pressed and formed into numerous shapes, in this case E-type. The type, ratio of materials and size of particles determines the magnetic properties of each material grade. In comparison to iron powder cores, ferrites suffer from low maximum flux density, which emphasizes again the need for high switching frequencies.

Another important factor of the core material is the Curie temperature. This is the temperature of a magnetic material at which its non-ferromagnetic start rendering. In other words, the permeability, or magnetic gain of a core can transition abruptly from a high value below the Curie temperature to almost non-existent above the Curie temperature. For ferrite materials it can be from 170° to 350° C depending on the ferrite grade and formulation.

The performance factor is a measure of the maximum power which a ferrite can transmit, whereby it is generally assumed that the loss does not exceed 300 kW/m³. Heat
dissipation values of this order are usually assumed when designing small and medium-sized transformers. Increasing the performance factor can either enable an increase of the power that can be transformed by the current core design, or a reduction in core size if the transformed power is not increased.

As can be seen in figure F1, better materials are on the market, as well as materials with less good characteristics. N27 for example is usually recommended for power applications in the frequency range up to about 100 kHz, the now implemented material N87 for the frequency range up to 500 kHz. Even though the material N87 thus can be used for much higher frequencies than 25kHz, already at this value, a maximum value for the magnetic flux density can reached of $B_{max} = 0.36 \text{T}$ is reachable.

**F.1.3 Coil winding number calculations**

In order to design an efficient transformer, several calculations have to be made on the core design. Most important factors are the material specific maximum magnetic flux density $B$ and the core surface $A_{core}$. Other important factors are the possible implementation of an air gap and the length of the magnetic path. It is important that a core will never go into saturation due to a too high magnetic flux density, as this will affect the behavior of the transformer.

The square-wave voltage at the input of the transformer causes a triangular shaped magnetizing current $I_M$ which is almost independent of the secondary current (see also the equivalent circuit). The magnetizing current is approximately proportional to the magnetic flux $\Phi$ i.e. to the magnetic flux density $B$. The input voltage $V_I$ determines the magnetic flux in the transformer core corresponding to Faraday’s Law:

$$V_{prim} = N_{prim} \frac{d(\Phi)}{dt} \quad (F1.1)$$

The magnetic flux density $B$ is determined by the core material and shape. A good estimation of the practical usage of Faraday’s Law can be given by equation F1.2.

$$\Delta B = \frac{I_{prim} \cdot T_s / 2}{N_{prim} \cdot A_{core}} \quad (F1.2)$$
With:
\[ \Delta B = 2 \cdot B_{\text{max}} = \text{maximum flux density swing} \]
\[ \hat{V}_{\text{prim}} = \text{peak input voltage} \]
\[ T_s = \text{switching time} \]
\[ N_{\text{prim}} = \text{primary number of windings} \]
\[ A_{\text{core}} = \text{core surface} \]

Formula F1.2 can be rewritten into F1.3. This will be the standard formula, used in the upcoming transformer calculations.

\[ N_{\text{prim}} = \frac{\hat{V}_{\text{prim}} \cdot T_s}{2 \Delta B \cdot A_{\text{core}}} \]  \hspace{1cm} (F1.3)

For the main transformer of the Mass GI the material N87 has been used. One of the characteristics of this material is the high performance factor, as shown in addendum C6, given by N87 manufacturer EPCOS, formerly Siemens.

This means that, for a maximum 50Hz input voltage of 325V, a switching time of 40μs and a core surface of 21mm by 28mm, the theoretical minimum number of turns should can be calculated, using F1.3.

\[ N_{\text{prim}} = \frac{325 \cdot 40 \cdot 10^{-6}}{2 \cdot 0.36 \cdot 21 \cdot 28 \cdot 10^{-6}} = 15.4 \]

In order to prevent the core from going into saturation, the actual number of windings should be a whole number, greater than the value calculated here. This would give a minimum number of windings of 16. However, since the GI should also be able to handle a certain amount of harmonics and over voltages, the maximum input value will now be set at 400V. At this value, the snubbers will turn on and prevent the voltage from rising to higher levels, therefore this can be seen as a solid maximum value.

For a peak voltage of 400V, the minimum number of windings should now be:

\[ N_{\text{prim}} = \frac{400 \cdot 40 \cdot 10^{-6}}{2 \cdot 0.36 \cdot 21 \cdot 28 \cdot 10^{-6}} = 18.9 \]

As a result, the primary winding will now have 19 turns.

In order to reduce the effect of the voltage drop over the IGBTs and diodes, the secondary winding will have 20 turns. This way, the output voltage will be 1/19 higher than the input voltage, equal to estimated 12V, leading to a maximum output voltage of 242V without any voltage drop over switches or the main transformer.

As can be found in addendum B2, the voltage drop over the primary and secondary coils is neglectable, due to the low values of the internal resistance. When neglecting this
transformer voltage drop over the main transformer, caused by its internal resistance, the theoretical output voltage at an input voltage of 230V and at full load can be calculated with equation F.1.4.

\[ V_{\text{load}} = \left( V_{\text{grid}} - 2 \cdot V_{\text{IGBT}} - 2 \cdot V_{\text{DIODE}} - V_{\text{primary coil}} \right) \frac{20}{19} - 2 \cdot V_{\text{IGBT}} - 2 \cdot V_{\text{DIODE}} - V_{\text{secondary coil}} \]  

(\text{F.1.4})

\[ V_{\text{load}} = \left( 230 - 2 \cdot 1.6 - 2 \cdot 1.3 - 0 \right) \frac{20}{19} - 2 \cdot 1.6 - 2 \cdot 1.3 - 0 = 230.2V \]

As can be seen, a 19:20 winding ratio would be a satisfying solution for the voltage drop over the switches at full load.

**F.1.4 Winding isolation**

Another important factor of the copper windings is the used isolation and the belonging isolation voltage. This is the voltage requirement in which primary and secondary coils must withstand from each other. It is usually in terms of the application of the transformer or inductor. In other words, independent of the voltage applied to the primary and voltage transferred to the secondary, there is a voltage difference between the primary and secondary that must also be met.

In this case, the main transformer must be able to withstand a peak voltage of 2kV between the primary and secondary windings. This not only sets demands to the thickness of the used isolation, but also to the creeping distance. Therefore, in this case Mylar isolation foil is used to separate the primary and secondary winding packages. Mylar is a brand of boPET tape. Biaxially-oriented polyethylene terephthalate (boPET) polyester film is used for its high tensile strength, chemical and dimensional stability, transparency and electrical insulation.

To meet the requirements regarding the isolation voltage, the primary and secondary winding packages are separated by 3 windings of Mylar foil.

For electrical isolation, also 3M 135°F Tape is used. This is an electrical insulating polyester film tape which meets the flame retardancy requirements of UL 510. It consists of materials that are represented in many UL Recognized Insulation Systems and has a UL 130°C temperature rating, which makes it applicable for the designed HF transformers. 3M 135°F Tape is especially suited for applications such as wrapping coils, capacitors, wire harnesses, transformers and motors and is therefore ideal for winding isolation within the designed power transformer samples. The datasheet can be found on the included CD-ROM.

**F.1.5 Copper foil width and thickness**

As previously shown, the task of the windings is to provide a path for the electrical current. One of the goals of the winding design is to decrease both the DC as the AC resistance. Since high currents have to be fed through the transformer, the winding package should have as little loss as possible. Not only is this necessary for optimum efficiency, also the heat transfer, belonging to the losses, is an important problem. Thus, the used windings have to be as thick as possible.

To gain a maximum efficiency in usage of the core bobbin, copper foil layers with a thickness of 0.15mm have been chosen. This way, the fill factor is as high as possible. In practice, the actual maximum value is dependent upon the tightness of winding, variations in insulation thickness, and foil thickness. Consequently, the fill factor is
always less than the theoretical maximum. A typical working value for copper wire is 0.6, but with foil, fill factors up to 0.9 can easily be reached.

Another advantage of the chosen foil thickness is the fact that it stays well below the range, in which the skin effect shows up, as will be shown later.

**F.1.6 Leakage inductance**

A portion of the magnetic field in a core does not get converted to magnetic flux, and as such does not couple to the secondary winding. Similarly, some of the magnetic flux does not couple to the secondary either. All of these stray magnetic fields are called leakage inductance.

Usually, transformers used in power electronics are designed for minimum leakage inductance, since this can cause over voltages in power switches at switch turn-off. Leakage inductance can cause the following non-desirable situations:

- Generation of voltage spikes
- Generation of RFI energy
- Reflection of RFI energy to secondary side as switching spikes
- Excessive switching noise coupling into other circuits
- Excessive noise reflected back to source, leading to conducted EMI

For simplicity, in this report, all leakage inductance will be summarized into a single inductive component on the primary winding only and refer to that as the leakage inductance. Normally the leakage inductance in a transformer is minimized as much as possible, but at the expense of dielectric withstand capability. There is usually a trade-off between dielectric withstand (or isolation voltage) and leakage inductance.

**F.1.7 Winding methods**

The winding techniques of high frequency transformer are compared in three different types: conventional, sandwiched and interleaved winding, respectively shown in figures F2.a, F2.b and F2.c. Specific electrical characteristics such as power loss, efficiency, voltage stress across IGBTs and ringing frequency, but also other properties such as production costs and difficulties in practical implementation are important when choosing the right setup.

![Fig. F.2. Winding methods for the main transformer.](image)

In order to increase the magnetic coupling between the primary and the secondary windings, in the main transformer sandwiched windings have been chosen, as shown in figure F2.b. This interleaving increases the HF performance of the transformer as well as lowers the leakage inductance between windings. It will also decrease the influence of the proximity effect, as shown in figure F3.
The best way to reduce the influence of the proximity effect is the usage of interleaved windings, as shown in figure F3.b. For the Mass GI a sandwich construction has been chosen, as can be seen in figure F3.c. Since the calculated secondary number of turns is 20 and therefore even, the secondary winding can easily be split up into two windings of 10 turns each.

**F.1.8 Transformer losses**

In each transformer, losses are present. These transformer losses can be split up into several categories. For low frequencies, such as 50 Hz (European commercial power) and 60 Hz (American commercial power), copper loss is the main factor. Calculating transformer losses however gets more complicated as the frequency increases, for example in this case with a switching frequency of 25kHz. While the core is energized, even if there is no load on the secondary, power is being dissipated. Therefore, next to the effect of copper loss, also other loss mechanisms will have to be taken into account at higher frequencies, such as the proximity effect and skin effect. Therefore, these effects will now shortly be described.

**F.1.9 Copper Loss**

Copper loss, also referred to as $I^2R$ loss, is the power dissipated in the coil when a low frequency current is flowing through the windings. As the second term implies, this dissipated power is not only proportional to the square value of the coil current, but also to the DC resistance of the coil. This DC resistance is dependable of several factors, such as the size of the copper surface, the length of the windings and even the temperature. In order to be able to calculate the DC resistance, first of all the resistivity of copper has to be calculated. At 20°C, this is equal to $\rho_{20} = 1.68 \times 10^{-8} \Omega \cdot m$. However, formula F1.5 shows the influence of temperature:

$$\rho_t = \rho_{20} (1 + \alpha (T_{\text{coil}} - 293)) \quad (F1.5)$$

In this formula, $\alpha$ stands for the temperature coefficient and $T_{\text{coil}}$ is the temperature of the coil in degrees Kelvin.

Since copper losses are proportional to the square value of the coil current, this loss is mostly present at high currents. Therefore, also the average coil temperature at high currents, 16A, will be taken for the following calculations. The working temperature can be estimated at 70°C and the temperature coefficient is equal to $\alpha = 3.93 \times 10^{-3}$. Using equation F1.5, the copper resistivity will now be:

$$\rho_{70} = 16.8 \times 10^{-9} \cdot (1 + 3.93 \times 10^{-3} \cdot (343 - 293))$$
$$\rho_{70} = 20.1 \times 10^{-9} \Omega \cdot m$$

In order to calculate the DC resistance, now the size of the coil needs to be taken into account. The measurements of the first winding on the E65 bobbin are 24mm² 31mm.
The outside winding measurements are 42mm x 52mm. The average length per winding is therefore 149mm. Since the primary and secondary windings are wound symmetrically around the core, this average winding length is valid for both.

Since the primary windings have 19 turns, with an average length of 149mm and a surface of $30 \times 0.15\text{mm}^2$, equal to a copper surface of $4.5\text{mm}^2$, the resistance of the primary coil can be calculated with equation F1.6.

$$R_p = \frac{\rho_{70} \cdot n_p \cdot l_{winding}}{A_{winding}}$$

$$R_p = \frac{20.1 \cdot 10^{-9} \cdot 19 \cdot 149 \cdot 10^{-3}}{30 \cdot 10^{-3} \cdot 0.15 \cdot 10^{-3}}$$

$$R_p = 12.7\text{m}\Omega$$

The total resistance of the two series connected secondary coils, with 10 turns each, can be calculated with equation F1.7.

$$R_s = \frac{\rho_{70} \cdot n_s \cdot l_{winding}}{A_{winding}}$$

$$R_s = \frac{20.1 \cdot 10^{-9} \cdot 20 \cdot 149 \cdot 10^{-3}}{30 \cdot 10^{-3} \cdot 0.15 \cdot 10^{-3}}$$

$$R_s = 13.3\text{m}\Omega$$

The calculated resistance will cause a copper loss, as shown in equation F1.8.

$$P_{\text{Copper Loss}} = (I_{\text{grid}})^2 \cdot R_{\text{DC}} = (I_{\text{grid}})^2 \cdot (R_p + R_s)$$

At a nominal AC voltage of 230V and a RMS current value of 16A, this will lead to a maximum total copper loss equal to:

$$P_{\text{Copper Loss}} = 16^2 \cdot (12.66 \cdot 10^{-3} + 13.3 \cdot 10^{-3})$$

$$P_{\text{Copper Loss}} = 6.6W$$

As can be seen, these losses are still acceptable, as long as they can be transferred to the surroundings.

**F.1.9.1 Skin Effect**

Also other copper losses exist, which are above and beyond normal DC current losses. They are caused by high frequencies, tending to push the current towards the outside edge of a wire or foil. It is discussed in terms of skin depth, which refers to how deep in the wire or foil the current will be able to flow. The higher the frequency the less wire area is used to pass current. The result is high copper losses due to the fact that the current will only be traveling under the surface of the copper windings.

The copper depth at which the skin effect will show up, can be calculated with formula F1.9.
As calculated in the previous paragraph, the resistivity of copper will be taken at a temperature of 70°C. The switching frequency $f_s$ is equal to 25kHz. Using equation F1.9, the skin depth is equal to:

$$\delta = \sqrt{\frac{2}{\sigma \cdot \mu_0} \cdot \frac{\rho_{copper}}{\pi \cdot f_s \cdot \mu_0}}$$  \hfill (F1.9)

This could give a problem when winding with normal copper wires, as this would implement a maximum efficient thickness of 0.9mm, as the skin depth can be calculated at both sides of the wire. In order to solve this problem, two solutions can be chosen: usage of several smaller wires in parallel or usage of copper foil, with a thickness less than twice the skin depth. The ultimate in this approach is litz wire, which stands for woven wire. However, the voltage spread, as can be seen in the following paragraph on proximity effect, can be less efficient when using litz wire.

Since for the main transformer copper foil with a thickness of 0.15mm has been chosen, the skin effect however will not be present. A quick calculation with a skin depth of 0.075mm tells us that, using this foil, the skin effect will only turn up as from switching frequencies of 900kHz.

### F.1.9.2 Proximity Effect

When two conductors, $i_1$ and $i_2$, are in proximity and carry opposing currents, the high frequency current components spread across the surfaces facing each other in order to minimize magnetic field energy transfer. However, if currents are flowing through one or more other nearby conductors, such as within a closely wound coil, the distribution of current within the conductor will be constrained to only certain regions of the copper winding due to the eddy current effect. The resulting current crowding is termed proximity effect.

Just like the previously mentioned skin effect, the proximity effect significantly increases the AC resistance of the conductor when compared to its resistance to a DC current. At higher frequencies, the AC resistance of a conductor can easily exceed ten times its DC resistance.

Using the theory on the proximity effect, explained in [17], this additional AC resistance can be calculated, using equation F1.10.

$$R_{prox} = \frac{1}{\pi \cdot s^2 \cdot \sigma} \sum \left(\frac{d}{2s}\right)^{2(m-1)} \rho_{em} \left(\frac{d}{2\delta}\right)$$  \hfill (F1.10)

In this formula, $d$ is the thickness of the copper foil, $s$ the distance between two foil core centre points and $\rho_{em}$ is the relative resistivity, calculated with formula F1.11.
In this formula, \( m \) is the winding number and \( J \) is the current density in the windings.

Since in this case, the thickness of the foil is much smaller than the skin depth, the previous formulas can be simplified into formula F1.12.

\[
R_{\text{prox}} \approx \frac{1}{\pi \cdot s^2 \cdot \sigma} \rho_{\text{prox}} \left( \frac{d}{2\delta} \right) \approx \frac{1}{\pi \cdot s^2 \cdot \sigma} \cdot \frac{1}{2m^2(m+1)} \left( \frac{d}{2\delta} \right)^4
\]  
(F1.12)

With \( d=0.15\text{mm} \) and \( s=0.18\text{mm} \), this equation will result in the following estimation:

\[
R_{\text{prox}} \approx 14m\Omega
\]

As can be seen, the proximity effect can form a serious threat to the efficiency of the transformer.

**F.1.9.3 Calculation of the total winding AC resistance**

Another way to calculate the total AC resistance of the windings, due to both the skin effect as the proximity effect, is to look at the relation between \( R_{\text{AC}} \) and \( R_{\text{DC}} \). For transformers with multiple windings, the exact AC resistance of the \( m \)th layer for solid round wires can be calculated, using formula F1.13, according to “Accurate prediction of high-frequency power-transfer losses and temperature rise”, [34]:

\[
\frac{R_{\text{AC}}}{R_{\text{DC}}} = \frac{\zeta}{2} \cdot \frac{\sinh(\zeta) + \sin(\zeta)}{\cos(\zeta) - \cos(\zeta)} + (2m-1)^2 \cdot \frac{\sinh(\zeta) - \sin(\zeta)}{\cos(\zeta) + \cos(\zeta)}
\]  
(F1.13)

with:

\[
\zeta = \sqrt{\pi} \cdot \frac{d}{\delta}
\]

\( d = \text{wire diameter} \)

\( \delta = \text{skin depth} \)

For copper foil, this can be simplified to formula F1.14, according to [34]:

\[
\frac{R_{\text{AC}}}{R_{\text{DC}}} = y \cdot \left( M(y) + \frac{2}{3} \cdot (m^2 - 1) \cdot D(y) \right)
\]  
(F1.14)

with:

- \( y \)
- \( M(y) \)
- \( D(y) \)
\[ y = \frac{d_{foil}}{\delta} \]
\[ d_{foil} = \text{foil thickness (m)} \]
\[ \delta = \text{skin depth (m)} \]
\[ m = \text{number of layers in winding section} \]
\[ M(y) = \frac{\sinh(2y) + \sin(2y)}{\cosh(2y) - \cos(2y)} \]
\[ D(y) = \frac{\sinh(y) - \sin(y)}{\cosh(y) + \cos(y)} \]

Since, due to the sandwich construction, the winding section only has to contain 9 primary and 10 secondary windings have to be taken into account, \( m \) is now equal to 19. Filling in equation Fr. 14 in Matlab, will now give:

\[ \frac{R_{AC}}{R_{DC}} = 1.55 \]

In this case, a calculated DC resistance of 25 m\( \Omega \) will therefore lead to a total resistance at 25 kHz of

\[ R_{AC} = 1.55 \cdot R_{DC} = 1.55 \cdot 26.0 \cdot 10^{-3} = 40.2 m\Omega \]

This again comes close to the calculated value of the proximity resistance, added to the DC resistance.

**F.1.9.4 Hysteresis Loss within cores**

Hysteresis loss is the power loss, which occurs when the core material is continually being magnetized, demagnetized and re-magnetized in the opposite polarity, which happens in this application at a frequency of 25 kHz. When a core is magnetically energized, there is a small amount of energy required to do it. The more cycles that occur per second the more power is dissipated. Some core material is best suited to low frequency applications, such as silicon steel, while ferrite is well suited to higher frequencies due to lower hysteresis losses. Therefore, ferrites have been chosen as core material in this application.

**F.1.9.5 Eddy Current loss within cores**

A varying magnetic flux can induce a varying voltage, and therefore current in a good conductor, in any material even material that is not ferromagnetic. As shown before, it can happen in wire within a transformer, but mostly it happens within the core itself. To reduce this effect, and thus increase the efficiency, steel based cores are laminated so that each slice can be insulated from the next. This prevents any large eddy currents from forming.

In this case however, Eddy current loss does not have to be taken into account. Owing to the high specific resistance of ferrite materials, the eddy current losses in the frequency range common today (1 kHz - 2 MHz) may be practically disregarded except in the case of core shapes having a large cross-sectional area. Therefore, no Eddy Current losses will be calculated for the used E65 core.
**F.1.9.6 Calculation of total core losses**

The total core losses can be calculated, but an easier and more accurate way to determine these core losses is the usage of the datasheet, as shown in addendum C4. In this datasheet, a value for the relative core losses can be found. In order to find the total losses, first of all the total core volume has to be calculated with equation (F1.15).

\[
V_{\text{core}} = 2 \cdot V_{\text{E-body}} = 2 \cdot (V_{\text{center leg}} + 2 \cdot V_{\text{outer leg}} + V_{\text{lower leg}}) \quad \text{(F1.15)}
\]

\[
V_{\text{core}} = 2 \cdot (2.0 \cdot 2.7 \cdot 2.2 + 2 \cdot 1.0 \cdot 2.7 \cdot 2.2 + 6.5 \cdot 2.7 \cdot 10) = 82.6 \text{cm}^3
\]

The relative core loss of the material N87 is equal to \( C_{\text{loss}} = 115 \text{ kW/m}^3 \) at 25kHz and 25°C. As can be seen, these losses decrease when the core temperature rises, due to the movability of the core molecules. With this value, the maximum core losses will be equal to (F1.16).

\[
P_{\text{core}} = C_{\text{loss}} \cdot V_{\text{core}} \quad \text{(F1.16)}
\]

\[
P_{\text{core}} = 115 \cdot 10^3 \cdot 82.6 \cdot 10^{-6} = 9.5W
\]

However, since it is clear that these losses are dependable on the temperature, the actual value will be lower.

**F.1.10 Temperature measurement**

Electrical and magnetic losses in the core and in the winding package will be transformed into thermal energy. These losses will therefore lead to a higher temperature of the core and winding package. As shown before, this will lead to a lower efficiency of the core and therefore will affect the efficiency of the total transformer. More important however is the fact that overheating of the transformer might lead to destruction of isolation layers, which might eventually lead to mechanical deformation and an electrical short circuit in the transformer. Therefore, a thermal resistor, an NTC, is taped onto the winding package. In case of over-temperatures, this will signal the central control unit to turn off the GI.
F.2 Main Transformer design for the push pull converter

Since the main transformer will also have to handle the same power flow, but at a higher current and lower voltage value, the thermal management will have to be thought through. Various options can now be chosen to adjust the 230V DAB main transformer design into a 120V Push-Pull main transformer design. Two things are important: the design of the windings and the design of the core. As will be shown, both designs are interconnected and therefore will need to be calculated simultaneously.

In order to reduce the losses in the transformer, a lower number of windings can be chosen. However, a reduction in the number of turns \( N_1 \) will cause an increase in \( \Delta B \) and a quadratic increase of hysteresis losses. As can be seen, also the surface of the core \( A_{core} \) is important. Therefore, a selection of different magnetic topologies will also be discussed. For all topologies, the number of windings and the winding resistance will be discussed.

F.2.1 Single core E65 transformer

The easiest way to adjust the standard main transformer design to the new demands is the insertion of a center tap in the primary winding package. Since the voltage is lower, the number of windings can also be reduced, as was shown in the previous paragraph. The minimum number of windings can be calculated, using Eq. 3:

\[
N_{prim} = \frac{V_{prim} \cdot T_s}{2 \cdot \Delta B \cdot A_{core}} = \frac{120 \cdot \sqrt{2} \cdot 40 \cdot 10^{-6}}{2 \cdot 0.36 \cdot 20 \cdot 27 \cdot 10^{-6}} = 8.7
\]

Therefore, the primary coils, which used to have 19 windings, will now be wound at 2\(^x\)9 windings. The secondary side already was already divided into 2\(^x\)10 windings, so no redesign is needed here. As a remark, the fact should be mentioned that in this winding number, no over voltage is taken into account and therefore only valid for an ideal 120V sinusoidal signal.

The total number of windings, wound on the bobbin is equal to 38, which is almost equal to the previous 39. Since the fill factor, the used space within the bobbin, was already very high, the thickness of the copper foil cannot be increased to decrease the winding resistance.

The voluminal resistance of copper is 17\(\Omega \cdot \text{m} \). The inner measurements of the bobbin are 24mm\(\times\)31mm. The outside winding measurements are 42mm\(\times\)52mm. The average length per winding is therefore 149mm. Since the primary and secondary windings are wound symmetrically around the core, this average winding length is valid for both.

Since the primary coils have 9 turns and the secondary coils have 10 turns, with an average length of 149mm and a surface of 30\(\times\)0.15mm\(^2\), equal to a copper surface of 4.5mm\(^2\), the total transformer resistance is estimated at:

\[
R_{foil} = \frac{\rho \cdot (n_p + n_s) \cdot l_{winding}}{A_{winding}} \quad (F2.1)
\]

\[
R_{foil} = \frac{20.1 \cdot 10^{-9} \cdot (9 + 10) \cdot 149 \cdot 10^{-3}}{30 \cdot 10^{-3} \cdot 0.15 \cdot 10^{-3}}
\]
Addendum

\[ R_{foil} = 12.6m\Omega \]

Using equation F1.8, a nominal AC voltage of 120V will lead to a maximum total copper loss equal to:

\[
P_{Copper\ Loss} = (I_{\text{grid}})^2 \cdot R_{DC} = 30^2 \cdot 12.6 \cdot 10^{-3}
\]

\[ P_{Copper\ Loss} = 11.4W \]

This is the double value of the copper losses at 230V, due to the fact that these losses are proportional to the quadratic value of the RMS current. Therefore, the transformer will have to transfer more heat to the surrounding air flow. Since in the 230V design, already temperatures over 70°C are reached, the Mass GI will turn off due to the main transformer thermal safety circuit.

A positive aspect of this winding configuration is the fact that the 9:10 winding ratio permits an 11% higher output voltage, thus reducing the effect of the internal voltage drop over the IGBTs and the main transformer. This voltage, equal to 13.3V at a 120V grid voltage, will then, even at high loads, boost up the output voltage to 125V. This configuration however might be dangerous at high grid voltages: at no load, an input voltage of 125V can create an output voltage of 139V, thus resulting in a total voltage over the secondary windings of \(2 \times 139 = 278V\), with a peak value of 393V. A harmonic voltage with RMS value of 5V, only 4% of the 50Hz signal, can then easily trigger the secondary snubber circuit, as the internal MOSFET starts conducting at 404V.

\[ \text{F.2.2 Dual core E65 transformer} \]

As was shown in the previous paragraph, a single E65 core transformer will not be able to transform the power without overheating. The main goal therefore is to reduce the winding resistance. This effect can be reached in three ways:

- Reduction of the total foil length
- Increase of foil thickness
- Increase of foil widthness

Combining these first two aspects with the fact that an increase of core profile will lead to a decrease of the number of windings, it is quite logical to increase the core surface. This enables the core to handle higher B-fields before it will go into saturation. One way to do this, is to use two E65 cores next to each other, which doubles the core surface.

Using a setup with a dual E65 core however will increase the average winding length. An increase of core profile must therefore lead to a high reduction of the number of windings. The new number of windings, taking a 20% voltage overshoot into account, can be, using F1.3:

\[
N_{\text{prim}} = \frac{\dot{V}_{\text{prim}} \cdot T_s}{\Delta B \cdot A_{\text{core}}} = \frac{120 \cdot \sqrt{2} \cdot 120\% \cdot 40 \cdot 10^{-6}}{2 \cdot 0.36 \cdot 2 \cdot 20 \cdot 27 \cdot 10^{-6}} = 5.2
\]

The total number of windings reduces from two times 9 primary and two times 10 secondary to two times 6 windings primary and two times 6 windings secondary.
Due to this lower number of windings, the copper foil thickness can be increased. At the first configuration, copper foil of 0.15mm has been used. In order to maintain the current high value for the fill factor, the thickness can now maximally be:

\[
d_{foil} = d_{old} \cdot \frac{n_{old}}{n_{new}}
\]

\[
d_{foil} = 0.15 \cdot \frac{2.9 + 2.10}{4.6} = 0.24\text{mm}
\]

At this foil thickness, the internal resistance at working temperature will be, using F2.1:

\[
R_{foil} = \frac{20.1 \cdot 10^{-9} \cdot 12 \cdot (147 + 2.27) \cdot 10^{-3}}{30 \cdot 10^{-3} \cdot 0.24 \cdot 10^{-3}} = 6.73\text{m}\Omega
\]

At a nominal AC voltage of 120V and a RMS current value of 30A, this will lead to a maximum total copper loss equal to the following value, using formula F1.8.

\[
P_{Copper\ Loss} = (I_{grid})^2 \cdot R_{DC} = 30^2 \cdot 6.73 \cdot 10^{-3} = 6.06\text{W}
\]

This calculated copper loss is even lower than in the current 230V DAB version. Taking into account that the cooling surface also has increased, due to the larger construction, heat transfer will therefore not be a problem for the main transformer.

Of course, this implementation also has a number of drawbacks. First of all, the increased size of the main transformer might lead to problems within the current Mass G1, as it will take a higher percentage of the PCB surface.

Furthermore, the voltage drop over the total schematics cannot be compensated anymore. A winding configuration of 6:7 or even 5:6 will lead to output voltages of 140-144V at 120V RMS in, thus activating the secondary snubber. Therefore, this is no option.

An extra half turn for the secondary windings can be considered, but will give many problems. Since the winding profile however should be symmetrically, this could only be implemented by an electrical parallel connection of two half windings, in order to keep the B-field in the E-core equal. This could then theoretically lead to a 6:6.5 configuration, with a voltage boost of 9.2V. Since this however would lead to a higher secondary leakage inductance and 2*2 extra connections per secondary winding and therefore to 8 extra secondary connections in total, this option will not further be investigated.

\[\text{Fig. F.4. Thermal images of the new transformer designs. a) single E65 vs dual E65, b) power loss in the connections of the dual E65.}\]
Thermal measurements on the single and dual core E65 are shown in figure F.4. As can be seen, heat dissipation is much lower and the temperature is much lower at the dual core setup.

The thermal images above however show also another problem: due to the high currents, in this measurement up to 20A DC, the current through the connecting wires is also high. In the first design, all foil windings were connected with two copper wires of 1mm in diameter, soldered in parallel over the whole width of the foil. However, this transition of the averagely 7.2mm² thick foil to a surface of 1.6mm² of the wires resulted in a relatively high resistance of the wires. This can be seen in figure F.4.b, where the connecting wires are white, thus having the highest temperature. Therefore, the dual core transformer eventually has been unwound, equipped with four copper wires per connection and rewound.

F.2.3 Single core E80 transformer

One of the possibilities to reduce the copper losses is the usage of another core. An E80 core, for example, has much more space for windings. However, the core effective area is smaller, as an E65 core has a profile of 20*27mm², whereas an E80 core has a standard profile of 20*20mm². This reduction in core surface will have to lead to a higher number of windings, in order to prevent the core from saturation. The minimum number of windings can be calculated with equation F1.3, taking into account that the original transformer will not saturate at 230V over 19 windings:

$$N_{\text{prim}} = \frac{\dot{V}_{\text{prim}} \cdot T_e}{\Delta B \cdot A_{\text{core}}} = \frac{120 \cdot \sqrt{2} \cdot 120\% \cdot 40 \cdot 10^{-6}}{2 \cdot 0.36 \cdot 20 \cdot 20 \cdot 10^{-6}} = 14.0$$

A convenient factor of this relatively high number of windings, compared to the dual 65 core setup, is the fact that secondary now 15 windings can be chosen per coil. This implies an ideal voltage boost of 8.6V at a 120V input voltage, which can easily compensate the voltage drop due to conduction losses in the main transformer and the IGBTs. Only RMS grid values higher than 130V will now trigger the secondary snubber.

With the new bobbin dimensions and the calculated number of windings, an optimal foil thickness can be calculated. Again, the current high value for the fill factor should be maintained. The E65 bobbins support winding packages up to 10mm thick; the E80 bobbins can contain winding packages as thick as 17mm. Based on the same fill factor as in previous calculations and using formula F2.2, the following foil thickness for the E80 120V transformer can be calculated:

$$d_{\text{foil}} = d_{\text{foil, old}} \cdot \frac{n_{\text{old}}}{n_{\text{new}}} \cdot \frac{d_{\text{E80}}}{d_{\text{E65}}} = 0.15 \cdot \frac{19 + 20}{2 \cdot 14 + 2 \cdot 15} \cdot \frac{18 \cdot 10^{-3}}{10 \cdot 10^{-3}} = 0.181 \text{mm}$$

A thicker and wider copper foil will lead to a lower copper resistance. However, using an E80 core will also lead to a higher average winding length. The minimum winding length can be estimated by looking at the dimensions of the bobbin, on which the core will be wound:

$$l_{\text{wmin}} = 4 \cdot d_{\text{bobbin}} = 4 \cdot 25 = 100\text{mm}$$
The maximum winding length can be estimated by looking at the total width of the E-cores, which will be placed over the E-80 bobbin:

\[ l_{\text{max}} = 4 \cdot (w_\text{core} - d_\text{bobbin}) = 4 \cdot 61 = 244 \text{ mm} \]

This will therefore result in an average winding length of:

\[ l_{\text{winding}} = \frac{l_{\text{max}} + l_{\text{max}}}{2} = \frac{100 + 244}{2} = 172 \text{ mm} \]

Using formula F2.1, a foil thickness of 0.18mm will lead to the following internal resistance at working temperature:

\[ R_{\text{foil}} = \rho_{\text{foil}} \cdot \left(\frac{n_p + n_S}{A_{\text{winding}}}\right) \cdot l_{\text{winding}} \]

\[ R_{\text{foil}} = \frac{20 \cdot 10^{-9} \cdot (14 + 15) \cdot 172 \cdot 10^{-3}}{40 \cdot 10^{-3} \cdot 0.18 \cdot 10^{-3}} \]

\[ R_{\text{foil}} = 13.9 \text{ m}\Omega \]

At a nominal AC voltage of 120V and a RMS current value of 30A, this will lead to a maximum total copper loss equal to:

\[ P_{\text{Copper Loss}} = (I_{\text{grid}})^2 \cdot R_{\text{DC}} = 30^2 \cdot 13.9 \cdot 10^{-3} \]

\[ P_{\text{Copper Loss}} = 12.5 \text{ W} \]

A higher fill factor, using copper of 0.20mm thick, would then lead to a copper resistance of:

\[ R_{\text{foil}} = 12.5 \text{ m}\Omega \]

At a nominal AC voltage of 120V and a RMS current value of 30A, this will lead to a maximum total copper loss equal to:

\[ P_{\text{Copper Loss}} = (I_{\text{grid}})^2 \cdot R_{\text{DC}} = 30^2 \cdot 12.5 \cdot 10^{-3} \]

\[ P_{\text{Copper Loss}} = 11.3 \text{ W} \]

This calculated copper loss is equal or even higher than the loss in the current 230V E65 version. However, taking into account that the cooling surface also has increased, due to the larger construction, heat transfer can therefore more easily take place and might not be a problem for the main transformer.
F.3 Unwinding results of transformer M1551-1

The following pictures in figure F5 show the unwinding process of transformer type M1551-1, which is used as the main transformer in the Mass GI. This process has been done to discover the cause of the high frequency noise a number of transformers produced.

On the first photos, the complete transformer is shown. Already an extension can be seen on the place where the NTC is attached to the windings. After removal of the NTC, further removal of the primary and secondary windings did not deliver any production irregularities anymore. As a final step, a failure report containing this information was sent to Elytone, the manufacturer of the transformers.

![Fig. F.5. Unwinding steps of the main transformer, type M1551-1.](image)
F.4 Failure report transformer M1551-1

Author: Etienne Thewissen, Mastervolt, dep. Engineering
Date: 2006-06-06
Application: MASS GI
Part: M1551-1, main transformer Ti
Manufacturer: Elytone Electronic Co., Ltd
Error: hearable mechanical vibrations

In May 2006, a MASS GI was returned, because it was producing a high frequency audible noise. After some tests, the main transformer M1551-1, responsible for the power transfer from the primary to the secondary side of the MASS GI, appeared to cause the sound.

The transformer only produced the sound during the first 20-30 seconds after switching on the MASS GI. To reproduce the sound, the transformer had to be cooled down again. The noise however could be taken away by slightly pressing against one of the sides of the core.

Therefore, the transformer was removed from the main PCB. Now appeared, that one of the outer sides of the windings was touching the core, as can be seen in picture 1A. At the other side, the gap between the windings and the core was present as intended, shown in picture 1B.

After removal of the core and the first layers of isolation foil, the NTC appeared to have caused the problem: the NTC was mounted to the side of the windings, as shown in figure 2A. According to the winding instructions however, the NTC should be mounted to the down side of the windings, in front of the connectors, pin 15 and pin 16, as shown in figure 2B.

Fig. 1: Top views on transformer. A: Top view on gap at the side with NTC. B: Top view on gap at the correct side.
F.4.1 Summary

The dislocated NTC enabled the transformer to transport mechanical vibrations from the core to the winding package and vice versa and thus caused the hearable noise.
F.5 Winding instructions main transformers

F.5.1 Original winding instructions main transformer Mass 61

Addendum

ARTIKEL: TR1 (HF-TRAFO)
ontwikkeling en produktie van
powersupplies en converters
Datum : 25-08-2002
Blad : 1 van 2

WIKKELVOORSCHRIFT E65 TRAFO

Samenstelling:

<table>
<thead>
<tr>
<th>aantal</th>
<th>art.nummer</th>
<th>leverancier</th>
<th>omschrijving</th>
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<tbody>
<tr>
<td>1 st</td>
<td>Wz6959/1st</td>
<td>Weisser</td>
<td>Spoelkoker M65 1 kamer</td>
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<tr>
<td>16 st</td>
<td>pen 256/3</td>
<td>Weisser</td>
<td>Soldeerpen</td>
</tr>
<tr>
<td>2 st</td>
<td>B66387 GX187</td>
<td>Siemens</td>
<td>E65/28/21 kernhelft materiaal N87</td>
</tr>
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<td>431204594</td>
<td>Philips</td>
<td>E65/28/21 kernhelft materiaal 3F3 (altern)</td>
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<tr>
<td>...cm</td>
<td></td>
<td></td>
<td>Blank vertind koperdraad 1 mm</td>
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<tr>
<td>...cm</td>
<td></td>
<td></td>
<td>Isolatie folie 39X0,05mm</td>
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<td>V.Spijk</td>
<td>Koperfolie 30X0,15 mm met isol.39X0,036mm</td>
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<td>...cm</td>
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<td>Capable</td>
<td>Teflon isolatiekous inw. diam. 1,0mm</td>
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<tr>
<td>...cm</td>
<td></td>
<td>Capable</td>
<td>Teflon isolatiekous inw. diam. 1,2mm</td>
</tr>
</tbody>
</table>

Afwerking:

1: Kernen op elkaar lijmen met Ablebond 789-3 zonder luchtspleet
2: Aantal wdg, isolatie en zelfinductie testen ( 3,4 mH +/-25% tussen pen 2 en pen 7, met pen 4 en 5 doorverbonden)
3: 2 uur uitharden bij 130°C.
4: Impregneren
6: Isolatiespanningstest met 3750Vac tussen pen 2,3,6,7 en pen 11,14.

onderaanzicht:

```
+-------------------------------------*---*
| 16 o----+    +---|+| 1
| NTC
| 15 o----+    +---|+| 2
| sccl
| 14 o      +---|+| 3
| *-+-
| 13 o      +---|+| 4
| 12 o----+    +---|+| 5
| ----+      +---|+| 6
| 11 o----+    +---|+| 7
| prim     sec2
| 10 o----+    +---|+| 8
| ----+      *-+-|
| 9 o----+   *-+-|
```

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POWER RESEARCH ELECTRONICS
ontwikkeling en produktie van powersupplies en converters
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WIKKELVOORSCHRIFT E65 TRAFO

Wikkelmethode:

Opmerking: alle uitlopers dienen te worden voorzien van een dubbele teflon isolatiekous. De stroken koperfolie aan beide zijden voorzien van 2x blank vertind montagedraad. Extra isolatietape aanbrengen op de soldering.

sec.1: 10 wdg met een geïsoleerde strook koperfolie 30x0,15mm. Uitlopers over gehele lengte dubbel isoleren. Begin uitlopers op pen 3,4, eind uitlopers op pen 1,2.

isolatie: 4 lagen mylar isolatiefolie 39x0,036mm.

prim: 19 wdg met een geïsoleerde strook koperfolie 30x0,15mm. Uitlopers over gehele lengte dubbel isoleren. Begin uitlopers op pen 10,11 eind uitlopers op pen 13,14.

isolatie: 4 lagen mylar isolatiefolie 39x0,036mm.

sec.2: 10 wdg met een geïsoleerde strook koperfolie 30x0,15mm. Uitlopers over gehele lengte dubbel isoleren. Begin uitlopers op pen 7,8, eind uitlopers op pen 5,6.

isolatie: 2 lagen mylar isolatiefolie 39x0,036mm.
**F.5.2 Winding instructions 230V single E65 core transformer**

**Sample Manufacturer:** Elytone

**Component list**
- 1 Bobbin E65
- 2 E65 core halves, material N87
- Single side isolated copper foil, 30x0.15mm
- Mylar foil, 39x0.036mm
- tape, 25x0.02mm
- NTC, 10kΩ

**Winding instruction**

**sec.1:**
10 Windings isolated copper foil 30x0.15mm.
Isolate both ends over whole length.
Connect first end to pen 3,4, connect second end to pen 1,2.

**Isolation:**
4 Layers of Mylar isolation foil, 39x0.036mm.

**prim:**
19 Windings isolated copper foil 30x0.15mm.
Isolate both ends over whole length.
Connect first end to pen 9,10 connect second end to pen 11,12.

**Isolation:**
4 Layers of Mylar isolation foil, 39x0.036mm.

**sec.1:**
10 Windings isolated copper foil 30x0.15mm.
Isolate both ends over whole length.
Connect first end to pen 7,8, connect second end to pen 5,6.

**Isolation:**
2 Layers of Mylar isolation foil, 39x0.036mm.

**NTC:**
Connect NTC to pen 15,16.

**Isolation:**
2 Layers of tape
**F.5.3 Winding instructions 120V single E65 core transformer**

**Sample Manufacturer:** Elytone

**Component list**
- 1 Bobbin E65
- 2 E65 core halves, material N87
- single side isolated copper foil, 30x0.15mm
- Mylar foil, 39x0.036mm
- tape, 25x0.02mm

**Winding instruction**

**sec.1:**
10 Windings isolated copper foil 30x0.15mm. Isolate both ends over whole length. Connect first end to pen 3,4, connect second end to pen 1,2.

**Isolation:** 4 Layers of Mylar isolation foil, 39x0.036mm.

**Prim1:**
9 Windings isolated copper foil 30x0.15mm. Isolate both ends over whole length. Connect first end to pen 9,10 connect second end to pen 13,14.

**Isolation:** 4 Layers of Mylar isolation foil, 39x0.036mm.

**Prim2:**
9 Windings isolated copper foil 30x0.15mm. Isolate both ends over whole length. Connect first end to pen 13,14 connect second end to pen 11,12.

**Isolation:** 4 Layers of Mylar isolation foil, 39x0.036mm.

**sec.1:**
10 Windings isolated copper foil 30x0.15mm. Isolate both ends over whole length. Connect first end to pen 7,8, connect second end to pen 5,6.

**Isolation:** 2 Layers of Mylar isolation foil, 39x0.036mm, 2 Layers of tape
F.5.4 Winding instructions 120V dual E65 core transformer

Sample Manufacturer: E. Thewissen

Component list
- 2 Bobbins E65, standing
- 4 E65 core halves, N87
- single side isolated copper foil, 40x0.2mm
- Mylar foil, 39x0.036mm
- tape, 25x0.02mm

Bobbin preparation
Take two standing E65 core bobbin, shown in figure F6.a, and remove one side of each bobbin, as shown in figure F6.b. Place remaining bobbin parts against each other and mechanically connect them with tape.

Winding instruction

sec.1 : 6 Windings isolated copper foil 30x0.23mm. Isolate both ends over whole length, as shown in figure F6.c. Connect first end to pen 3,4, connect second end to pen 1,2. The

Isolation : 4 Layers of Mylar isolation foil, 39x0.036mm.

Prim1 : 6 Windings isolated copper foil 30x0.23mm. Isolate both ends over whole length, as shown in figure F6.c. Connect first end to pen 9,10 connect second end to pen 13,14.

Isolation : 4 Layers of Mylar isolation foil, 39x0.036mm.

Prim2 : 6 Windings isolated copper foil 30x0.23mm. Isolate both ends over whole length, as shown in figure F6.c. Connect first end to pen 13,14 connect second end to pen 11,12.

Isolation : 4 Layers of Mylar isolation foil, 39x0.036mm.

sec.2 : 6 Windings isolated copper foil 30x0.23mm. Isolate both ends over whole length, as shown in figure F6.c. Connect first end to pen 7,8, connect second end to pen 5,6.

Isolation : 2 Layers of Mylar isolation foil, 39x0.036mm, 2 Layers of tape

Fig. F.6. Winding instructions for a dual core E65 transformer. a) Connection of wires over whole foil width, b) original, single E65 bobbin, c) removal of bobbin side, d) dual E65 bobbin with first winding.
**F.5.5 Winding instructions 120V single E80 core transformer**

Sample Manufacturer: E. Thewissen

Component list
- 1 Bobbin E80
- 2 E80 core halfs, N87
- copper foil, 40x0.2mm
- tape, 25x0.02mm

**Bobbin preparation**
Remove two opposite inner sides of bobbin, as shown in figure F7.a. Isolate holes with two windings of tape.

**Copper foil preparation**
Due to the lack of single side isolated copper foil, non-isolated copper foil has been isolated with tape. Due to the fact that no applicable tape of 48-50 mm wide was available, two slightly overlapping layers of 25mm wide tape have been used, as shown in figure F6.b.

**Winding instruction**

| Sec. I | 15 Windings isolated copper foil 40x0.2mm. |
| Isolation | 4 fully covering layers of tape, 25x0.02mm. |
| Prim1 | 14 Windings isolated copper foil 40x0.2mm. |
| Isolation | 4 fully covering layers of tape, 25x0.02mm. |
| Prim2 | 14 Windings isolated copper foil 40x0.2mm. |
| Isolation | 4 fully covering layers of tape, 25x0.02mm. |
| Sec. I | 15 Windings isolated copper foil 40x0.2mm. |
| Isolation | 2 fully covering layers of tape, 25x0.02mm, as shown in figure F7.b |

*Fig. F.7. Winding instructions for a single E80 core transformer. a) manually isolated copper foil, b) adjusted bobbin, c) bobbin with first winding, c) fully wound transformer with first core half.*
ADDENDUM

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