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Cryogenic H-Bridge Converter for HTS Degaussing Application

Djurre Wikkerink, Mladen Gagić, Armando Rodrigo Mor, Henk Polinder, *Senior Member, IEEE*, and Robert Ross, *Senior Member, IEEE*

***Abstract*—A degaussing system can be used to reduce the detectability of the magnetic signature of a ship. Commonly, a degaussing system consists of a set of onboard copper coils that produce a magnetic field to compensate for the magnetic signature. High-temperature superconductive degaussing coils are considered an alternative to copper degaussing coils because of a reduction in energy losses, weight, volume, and costs. The losses of a high-temperature superconductor (HTS) degaussing system can be reduced even further by powering it with a cryocooled converter with parallel MOSFETs. A low-duty cycle and smaller current leads can be used. These solutions eliminate most of the power source losses. This article investigates such a cryocooled converter. The effect of the low switching frequency on the converter performance is tested. A prototype that can operate at cryogenic temperatures was built. The converter powers an HTS coil. It was found that a load current of 50 A can be achieved with a duty cycle of just 0.025 at an input voltage of 3.5 V while still meeting the requirement of a maximum current ripple of 0.5%. At a switching frequency higher than 100 Hz, the converter's performance deteriorates. Also, oscillations were observed in the circuit. This is a problem due to the low blocking voltage of the MOSFETs. The parasitic inductances in the circuit have a high impact on the performance because the resistance in the circuit is very low.**

***Index Terms*—Converter, cryocooled electronics, cryostat, degaussing, high-temperature superconductors (HTSs), magnetic signature, parallel MOSFETs, rare-earth barium copper oxide (ReBCO).**

I. Introduction

THE ferromagnetic hull of a ship locally distorts Earth's magnetic field [1]. The amount of distortion, or magnetic signature, can be detected by sea mines, which pose a potential danger to the ship [2]. Detection of the ship can be avoided by reducing the magnetic signature by means of degaussing [3]. A degaussing system commonly consists of a set of onboard copper coils that produce an opposing magnetic field to the magnetic signature.

In a degaussing system, a significant amount of Ohmic losses is dissipated in the copper coils. A high amount of ampere-turns is required and the coils are relatively long [4]. Furthermore, the degaussing coils are heavy and may take up valuable space onboard the ship [5]. A solution for the aforementioned problems is to use high-temperature superconductors (HTS) instead of copper degaussing coils. HTS have negligible conduction losses and the system can be made more compact. However, the system needs a cryostat to cool the HTS coils below the critical temperature. The advantage of using HTS with respect to low-temperature superconductors (LTS) is that the cryostat can operate using liquid nitrogen or air instead of liquid helium. This makes the cryostat design cheaper and much more efficient.

The feasibility of using HTS degaussing coils is currently under investigation. The US Navy tested a ship with a full-size HTS degaussing coil in 2011 [6] using a cryostat with gaseous helium [7] and connecting multiple cryostats to a single cooler [8]. In work preceding this article, a theoretical comparison between HTS and copper coils has been made [4], [5], a scaled demonstrator was built with HTS and copper degaussing coils [9], the current ripple in HTS degaussing coils was investigated [10], and a converter was designed for HTS degaussing coils [11], [12].

When using HTS, a higher current can be utilized than when using copper. The number of turns can, therefore, be limited, and less HTS material is needed. However, the power supply must supply a higher current than in the conventional case. This is expected to cause more losses in the current supply [5]. Also, with a higher current, larger current leads are needed [13]. Through the current leads, heat leaks into the cryostat, which needs to be cooled away. Cooling is an energy-inefficient process [14]. It was estimated that for every watt of dissipated power inside the cryostat, the cooler consumes 14 W of power. Dissipating heat in the cryostat should be avoided as much as possible.

A solution for reducing the extra losses in the current sources and the cryostat has been suggested in the previous work [12]. By using a converter with MOSFETs with low ON-state resistance, the duty cycle can be minimized [15], [16]. Because the HTS coils have no resistance but the inductance is in the range of tens

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Djurre Wikkerink is with High Voltage Technologies, Delft University of Technology, 2628 Delft, The Netherlands (e-mail: djurre.wikkerink@rhmarine.com).

Mladen Gagić is with DC Systems, Energy Conversion & Storage, Delft University of Technology, 2628 Delft, The Netherlands.

Armando Rodrigo Mor is with the Instituto de Tecnología Eléctrica, Universitat Politécnica de Valéncia, Camino de Vera, 46022 Valencia, Spain.

Henk Polinder is with Transport Engineering and Logistics Group, Delft University of Technology, 2628 Delft, The Netherlands (e-mail: d.p.wikkerink@tudelft.nl).

Robert Ross is with Transport Engineering and Logistics Group, Delft University of Technology, 2628 Delft, The Netherlands, and also with the Institute for Science and Development, 6711 Ede, The Netherlands (e-mail: r.ross@iwo.nl).

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of millihenry, the time constant of the system is large. During the relatively long freewheeling state, almost no energy is dissipated. Moreover, by placing the MOSFETs inside the cryostat, the rms current through the current leads is decreased and the heat leak into the cryostat is minimized. The cross-section of the current leads can be made more than a hundred times smaller if a sufficiently large smoothing capacitor is used in the cryostat [12]. As a side effect, the ON-state resistance of the MOSFETs will be decreased by cooling them down. The MOSFETs themselves do dissipate energy directly in the cryostat, however. For this reason, the switching frequency should be kept low to reduce the switching losses inside the cryostat. By placing multiple MOSFETs in parallel, the ON-state losses can be reduced even further, which, in turn, also reduces the duty cycle.

In this article, the results from the previous work will be experimentally tested. The aim is to investigate a cryocooled H-bridge converter with multiple MOSFETs in parallel that delivers a dc to an inductive HTS load. The focus is to obtain the lowest duty cycle and switching frequency while still meeting the requirements for a maximum current ripple of 0.5% and a current of 50 A. Also, different switching frequencies will be tested to investigate the effects of low switching frequencies on the converter's behavior. The measured results are then compared to simulations.

For the experiments, a converter is needed, which can operate at cryogenic temperatures. The series resistance in the HTS coil current path and the deadtime should be minimized so that the lowest possible duty cycle can be obtained.

The rest of the article is organized as follows. Section II describes the experimental test setup and provides the design requirements of the converter. This section also describes the parallel MOSFET gate-driver circuit and the HTS coil design. Section III analyzes the results of the system model simulation and laboratory prototype experiment. Section IV discusses the results and future work. Finally, Section V concludes this article.

II. Description of the Setup

This chapter describes the different electromechanical components of the test setup. The test setup consists of a dc voltage source, which interfaces the dc voltage bus of the full-bridge cryogenic power electronic converter. An HTS load connects between the two midpoints of the H-bridge. Independent gate drivers drive the MOSFET switches while the measurement setup collects the input and output currents and voltage values.

A. Converter

The system consists of a full-bridge topology that enables bidirectional dc load current flow. The degaussing coils need to be able to create a magnetic field in two directions, which is why the current has to be able to be both positive and negative. A simplified schematic of the design is shown in Fig. 1. The dc through the load is controlled by either alternately opening and closing Q_1 and Q_3 and keeping Q_4 closed or by alternately opening and closing Q_2 and Q_4 and keeping Q_3 closed. The discharge time is long compared to the charge time because of the high time constant of the load and low series resistance on the printed circuit board (PCB). For this reason, the only losses during the discharge state are in the MOSFETs and the copper traces of the PCB. The power source does not experience any conduction losses during the discharge period, which makes this design very energy efficient.

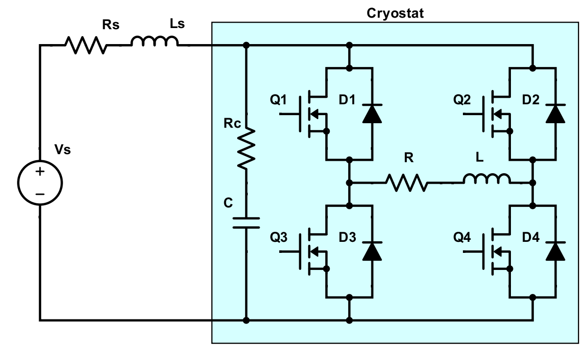

Fig. 1. Simplified schematic of the converter. During the short charge state, losses occur in the source equivalent resistance. During the much longer discharge state, there are almost no losses because of the low series resistance.

TABLE I
Converter Parameters

Symbol	Description	Value	Unit
I_r	Rated coil current	50	A
V_s	Supply voltage	3.5	V
R_s	Cable and supply resistance	29.3	mΩ
L_s	Cable and supply inductance	3.9	µH
R	Total resistance on PCB	102	µΩ
L	HTS coil inductance	1.46	mH
C	DC bus capacitance	21.7	µF
R_C	Capacitor ESR	14.9	mΩ
R_{DSon}	Parallel MOSFET on-state resistance	45	µΩ
N_{mos}	Number of parallel MOSFETs per switching arm	10	

The reason that the MOSFETs are placed inside the cryostat is that the current leads do not have to be dimensioned for the full-rated load current. The current leads only have to be able to handle the rms source current over the whole cycle, which significantly limits the heat leak into the cryostat [12].

The power input to the cryogenic converter connects to an external HP 6269A 60A dc voltage source by means of a cable. Table I contains the parameters of the system obtained from the measurements of the experimental setup. These are also the parameters, which are used in the simulations.

Each switching arm consists of ten parallel connected SiR178DP MOSFETs from Vishay. It was shown that it is not useful to place more than ten MOSFETs in parallel [12]. The individual MOSFET ON-state resistance was measured to be 611 µΩ at room temperature and 404 µΩ at 77 K. Parallel operation minimizes the losses and enables an even lower duty cycle. However, caution must be taken as these high-current MOSFETs have a relatively low blocking voltage; in this case, a datasheet listed a value of only 20 V.

The total resistance through the PCB and the MOSFETs between the connection points of the HTS coils was measured by injecting a known dc and measuring the voltage with a Keithley Nanovoltmeter Model 2182A. This measured resistance represents the sum-total resistance of the system discharge state and

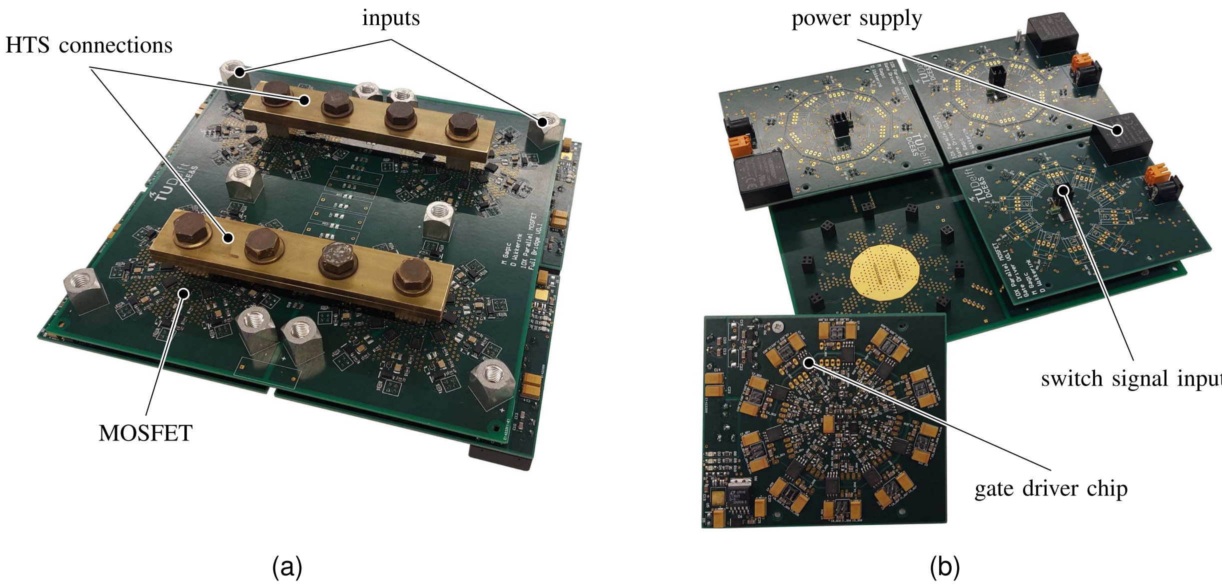

Fig. 2. Photos of the PCB, which was designed to operate in cryogenic temperatures. (a) Top view of the PCB. The ends of the HTS coil are clamped between two copper bars. The ten parallel MOSFETs per switch of the H-bridge configuration are placed as close as possible to the HTS connection points. (b) Bottom view of the PCB. The gate drivers are connected through headers, with each MOSFET having its own gate-driver circuit.

determines the limitations on the duty cycle. The chosen dc bus capacitance was shown to be sufficient by simulations in LTSpice XVII from Analog Devices.

Due to insufficient damping, high-frequency current transients across parasitic inductances increased the risk of drain–source voltage overshoot. To prevent damage or destruction of the low-voltage switches, an resistor capacitor (RC) snubber was later incorporated. However, this was still not sufficient to protect the MOSFETs. The voltage oscillations were found to be a factor of four higher than the input voltage, so the input voltage needed to be kept below 5 V.

Fig. 2(a) shows a photo of the converter. The parallel MOSFETs are placed as close as possible to the HTS connection points. The MOSFETs are placed so that the distances from each MOSFET to the HTS coil connection point are symmetrical and equal. This ensures an optimal load-current sharing in the MOSFETs. The HTS coil is clamped between two copper bars over a distance to minimize the contact resistance.

B. Gate Driver

Each switching arm is driven by a separate cryogenic-rated gate-driver board. The gate driver is cryogenic rated because each separate component is rated for cryogenic temperatures. The gate-driver boards are connected to the converter board through headers, ensuring an equal distance from the gate-driver chips to the MOSFETs. Each gate-driver board consists of the following:

1) a common isolated and several step-down linear power supply regulators;
2) a common input signal port;
3) 10 gate-driver ICs;
4) 10 identical gate driving circuits.

In this way, the driver can deliver sufficient gate current to switch the MOSFET as quickly as possible so that the deadtime can be kept as short as possible. Fig. 2(b) shows a photo of the gate drivers. Just like the MOSFETs, the gate drivers are placed circularly on the PCB. This is done to ensure equal propagation delays and gate-driving circuit impedance amongst all ten parallel MOSFETs.

The input signals for the gate drivers are provided by a function generator, which has an adjustable switching frequency, duty cycle, and deadtime between the two signals. The control of the switching is open loop.

C. HTS Load

For the superconductive coil, rare-earth barium copper oxide (ReBCO) HTS tape with a critical current of 105 A is used. The HTS tape is wound around an iron core to maximize the inductance and to simulate a ship's ferromagnetic hull. Iron does not perform well under cryogenic temperatures because it loses some of its permeability. However, at 77 K, it should still have at least 80% of the permeability at room temperature [17].

An amount of 30 m of HTS tape is used to create an as high as possible inductance. With all the tape used, the inductor has 88 turns. The iron core was taken from an old choke and we used less than half of the rated ampere-turns. In order for the iron core not to saturate, an air gap is installed that increases the reluctance of the core. The inductance of the coil, which is stated in Table I, is measured with an inductor capacitor resistor (LCR) meter.

The current in the superconductive coil cannot be measured with a shunt because the series resistance of the coil needs to be kept as low as possible. Therefore, the current is measured indirectly by measuring the magnetic flux density due to the current in the HTS coil. Hall sensors of different types are

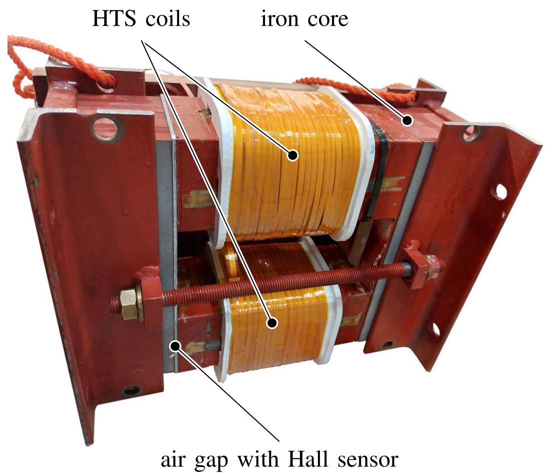

Fig. 3. HTS tape wound around an iron core. An air gap in the core prevents the core from saturating. Magnetic flux sensors are placed inside the air gap.

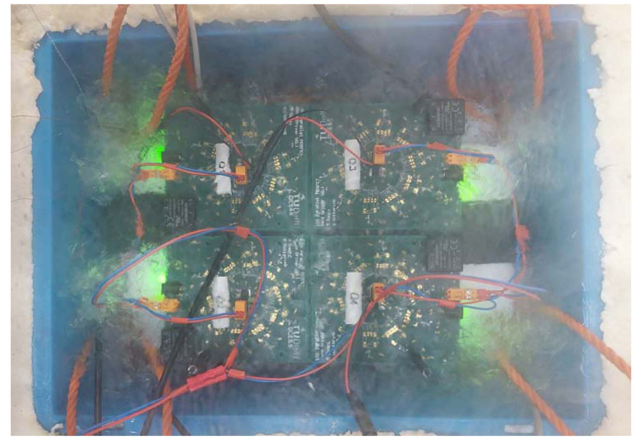

Fig. 4. PCB connected to the HTS coil. The setup is operating while submerged in liquid nitrogen.

installed in the air gap of the iron core. Fig. 3 shows a photo of the HTS load.

D. Measurement Setup

Since measuring current with an Ohmic shunt significantly influences the measurement, Hall sensors of the type HE244 by ASensor Technology AB are placed in the iron core of the load to measure the current in the coil indirectly. These sensors have a bandwidth of dc—100 kHz. However, there is always a conducting MOSFET in series with the current path. By measuring the voltage across this MOSFET as a shunt, the current can be measured directly as well. The voltage across the MOSFET was probed by soldering thin, copper, and shielded wires as close to the MOSFET drain and source as possible. This voltage was connected to a differential amplifier and then to an oscilloscope. The sensitivity of the ON-state resistance of the MOSFETs at different currents and at 77 K was investigated in previous work and showed consistent results [12]. The Hall sensors and the shunt measurements show an accurate reading of the current in the coil. Finally, the shunt measurements are used in the graphs of this article.

The current measurements were calibrated by injecting a range of known currents into the HTS coil and measuring the corresponding steady-state voltages from the current sensors. Linear amplification circuits then condition these voltage measurements, which are then logged. To measure the ripple at every switching frequency, a special amplifier was developed to be able to have ac coupling for the whole frequency range. The current ripple is obtained by postprocessing of the measured current waveforms.

During the measurements, the PCB and the HTS load are entirely submerged in liquid nitrogen. A photo of the test setup under operation is shown in Fig. 4.

III. Results

Fig. 5 shows the coil current of the HTS coil for different cases. The measured results are compared to simulations with the modeling software LTSpice XVII from Analog Devices. The output current was measured for all the combinations within a range of frequencies and duty cycles. Fig. 5(a) shows the output current as a function of frequency at two of the measured duty cycles. Fig. 5(b) shows the output current as a function of the duty cycle at two of the measured switching frequencies.

Fig. 6(a) shows the ripple in the coil current as a function of frequency for a duty cycle of 0.025. The requirement of a minimum ripple of 0.5% is met when the switching frequency is at least 21 Hz. The ripple was found not to be dependent on the duty cycle.

Fig. 6(b) shows the measured power losses of one gate driver as a function of switching frequency. During operation, two gate drivers are switching. One gate driver produces a dc signal to ensure that one leg is continually conducting. This gate driver consumes 1.5 W.

The effect of different dead times was investigated as well. A range of dead times from 100 ns to 1 µs was tested. However, there was no difference observed in the coil current.

IV. Discussion

While the source voltage is only 3.5 V, the required current of 50 A can be achieved with a duty cycle of just 0.025. In a regular converter, the low duty-cycle would be a problem because the ON-time might become too short for the switches. However, since the switching frequency in this converter can be kept very low compared to a regular converter, the low duty cycle is not a problem. The ON-time of the MOSFETs is still long enough.

As the switching frequency increases, the performance of the converter drops. The inductance of the cable that connects the source to the converter in combination with the capacitance on the PCB acts as a low-pass filter. The measured cable inductance in combination with the converter was simulated in LTSpice XVII from Analog Devices. Fig. 5(a) shows that the measured results correspond to the simulations. The drop in output current is significant because of the lack of damping in the circuit. However, for a cryocooled converter, a low switching frequency is wanted anyway. A lower switching frequency reduces switching losses, which otherwise would have been dissipated inside the

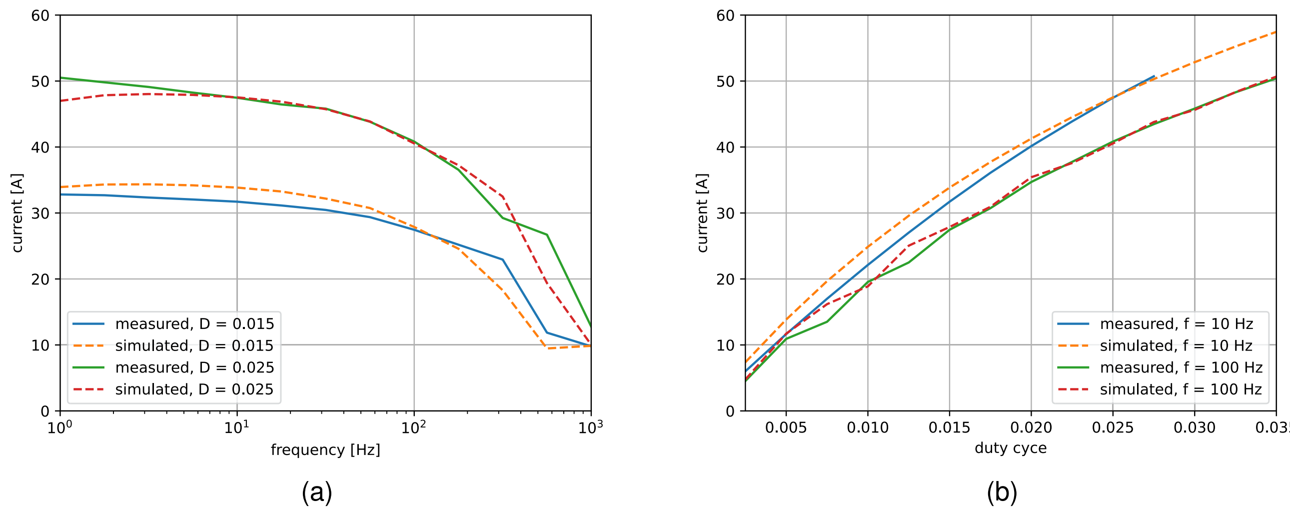

Fig. 5. Simulated and measured HTS coil inductor current under different circumstances. (a) Current as a function of frequency for two different duty cycles. (b) Inductor current as a function of a duty cycle for two frequencies.

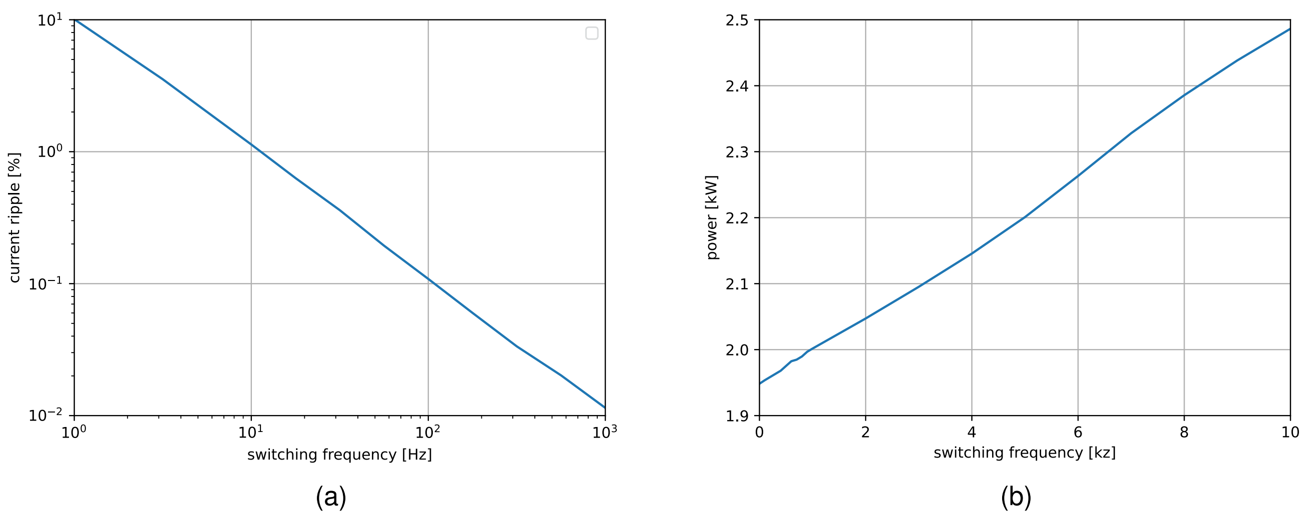

Fig. 6. (a) Measured ripple in the current as a function of frequency for a duty cycle of 0.025. (b) Measured gate-driver losses for one gate driver as a function of switching frequency.

cryostat. However, the current ripple requirements should be met to reduce the ripple in the magnetic signature. The current ripple can be reduced by using an interleaved forward converter topology [16], but this will increase the losses because there are more components inside the cryostat [12]. Another solution to reduce the current ripple is to use a switching frequency modulation scheme [10].

A low input voltage is necessary because MOSFETs with a low ON-state resistance usually have a low blocking voltage. Oscillations should be taken into consideration carefully. This problem may occur because of the lack of damping in the system. With almost no resistance, any kind of oscillation can be very high. Even when using a snubber circuit, the oscillations in this particular circuit are very high. Besides, the snubber circuit introduces extra losses inside the cryostat, which makes the system less efficient. In future designs, the oscillations should be controlled using a soft switching scheme. The extra losses and reduction of oscillations with a soft switching scheme need to be quantified in future work. However, the losses are expected not to have such a significant impact that the output current is affected. This is because, in this article, we have shown that with longer dead times, the output current is not affected. Another solution is to decrease the parasitic inductances on the PCB as much as possible.

By placing the MOSFETs inside the cryostat, the size of the current leads can be limited and, therefore, the heat leaks into the cryostat. However, the gate signal still has to be delivered to the MOSFETs by means of a physical connection. A potential solution for this problem is to use an optic fiber connection for the gate signal. Optic fiber cable has a much lower heat transfer coefficient than copper. The gate signal can be transported wirelessly by means of a transformer as well. In this case, no physical connection is needed at all. However, in both cases, some kind of active secondary circuit is required inside the cryostat.

By placing multiple MOSFETs in parallel, the resistance in the conduction path of the HTS coils can be reduced, and therefore, the time constant increases. At some point, adding more switches in parallel does not decrease the ON-state resistance any further because the resistance of the PCB is dominant. However, the switching losses increase by adding more parallel MOSFETs. An optimum amount of parallel MOSFETs must be found.

The deadtime does not have a significant influence on the performance of the converter. This might be because of the low switching frequency. The period of time is much longer than the deadtime. The gate drivers are designed in such a way that the MOSFETs can switch as fast as possible to reduce the deadtime losses. Because of this, the gate drivers have a large power consumption, which is dissipated in the cryostat. Since the deadtime does not influence the performance of the converter significantly, and soft switching can reduce the switching oscillations, the gate drivers can be less powerful in a future design, which reduces the losses.

In future designs, the series resistance of the PCB should be decreased even further. This can be done by using a thicker copper layer on the PCB or, since the PCB is inside the cryostat anyway, by using a superconductive layer on the PCB. Also, the distance between the switches can be made shorter.

V. Conclusion

When a cryocooled MOSFET-based H-bridge converter is used to power an HTS degaussing coil, losses can be decreased significantly. A condition for this converter is that it should operate with a very low duty cycle and switching frequency. A prototype with an HTS load was built successfully. It was found that a load current of 50 A can be achieved with a duty cycle of 0.025 and an input voltage of 3.5 V while still meeting the requirement of a maximum current ripple of 0.5%.

At switching frequencies higher than 100 Hz, the output current decreases. This is a consequence of the inductance in the cables and the lack of damping in the superconductor. This research shows that a cryogenic H-bridge converter for an HTS degaussing application is a feasible option to improve the performance of an HTS degaussing system.

MOSFETs with a low ON-state resistance usually have a low blocking voltage. The oscillations in this converter are high because of the lack of resistive damping. The oscillation voltage should be taken into account carefully when designing a converter like this.

The conducted experiments show that a cryogenic H-bridge converter for an HTS degaussing application is a viable approach to power energy-efficient HTS degaussing systems.

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Djurre Wikkerink received the M.Sc. degree in electrical engineering from the Delft University of Technology, Delft, The Netherlands, in 2017.

He was a Process Operator for Total E&P from 2012 to 2013, an Electrical Engineer with Teamwork Technology from 2016 to 2018, and is currently a Consultant with RH Marine and a Ph.D. Candidate with the Delft University of Technology, working on high-temperature superconductors for degaussing systems. His research interests include high-temperature superconductors, degaussing, converters, and electrical machines.

Mladen Gagić received the bachelor's and master's degrees in electrical engineering from the Faculty of Technical Sciences, University of Novi Sad, Novi Sad, Serbia, in 2014 and 2015, respectively.

Since 2015, he has been a Ph.D. Candidate with the Electrical Sustainable Energy Department, Delft University of Technology (TU Delft), Delft, The Netherlands. As of 2020, he was a Research-Technician with Electrical Sustainable Power Laboratory, TU Delft. His research interests include power electronics, modular multilevel systems, and multifrequency power conversion.

Armando Rodrigo Mor received the Ph.D. degree in electrical engineering from the Universitat Politécnica de Valéncia, Valencia, Spain, in 2012.

He joined and later led High Voltage Laboratory and Plasma Arc Laboratory, Instituto de Tecnológico de la Energía, Valencia. From 2013 to 2021, he was an Assistant Professor with Electrical Sustainable Energy Department, and in 2019, he became the Head of the High Voltage Team, Delft University of Technology, Delft, The Netherlands. Since 2021, he has been with the Instituto de Tecnología Eléctrica, Universitat Politécnica de Valéncia. His research interests include high-voltage technology, high-voltage testing, monitoring and diagnostic, sensors for high-voltage applications, space charge measurements, and HVdc.

Henk Polinder (Senior Member, IEEE) received the Ph.D. degree in electrical engineering from the Delft University of Technology, Delft, The Netherlands, in 1998.

Since 1996, he has been an Assistant/Associate Professor with the Delft University of Technology, working in the field of electrical machines and drives. He worked part-time in industries with wind turbine manufacturer Lagerwey, from 1998 to 1999; Philips CFT in 2001; and ABB Corporate Research, Västerås, in 2008. He was a Visiting Scholar with Newcastle University, Newcastle upon Tyne, in 2002; Laval University, Quebec, in 2004; The University of Edinburgh in 2006; and the University of Itajubá in 2014. He has authored or coauthored more than 300 publications. His main research interests include sustainable drive and energy systems for maritime applications.

Robert Ross (Senior Member, IEEE) received the Ph.D. degree in physics and chemistry from Utrecht University, Utrecht, The Netherlands, in 1990.

He was with Utrecht University, KEMA, National Institute of Materials & Chemical Research (Tsukuba, Japan), NL Defence Academy (Electrical Energy Systems), TenneT TSO (Asset Management), and HAN University of Applied Sciences. He is currently with IWO (Institute for Science & Development), Ede, The Netherlands, as a Senior Researcher and a Board Member, and teaches at the Delft University of Technology as a Retired Professor. His work led to four patent applications on materials or superconductivity. His book on reliability analysis is published with Wiley/IEEE (ISBN 9781119125174). His research interests include materials, reliability, and availability of systems and components, electrical energy and mobility, applied superconductivity, forensic investigations, scarce-data-based decision-making, diagnostic techniques, and sustainable energy.

Dr. Ross is a recipient of the 2004 SenterNovem Annual Award for Best Energy & Environment and was nominated by the World Technology Network (USA) for Best Researcher Energy in 2006.