Abstract—The Fast Field-Cycling Nuclear Magnetic Resonance is a widely used technique in the study, at nuclear level, of the structure of materials, and their behavior when exposed to varying radiation frequencies. One of the crucial processes of this technique is about polarizing the particles that constitute the sample, through the utilization of strong, and as stable as possible, magnetic fields. The detection of the signal emitted by the sample should be done with a Radiofrequency system, which should be tuned to a previously-set frequency. To obtain these kind of magnetic fields, which should have maximum precision, in terms of value, and also regarding the time duration of the polarization and detection pulses, the devices used are switching electronic systems, and a magnet, into which the sample is placed. The power source should provide the magnet with current values proportional to the desirable magnetic field density values at each phase of the NMR-FFC process. Modern devices are capable of inducing very quick transitions of the current level in the magnet. There are, however, some perturbations, such as parasitic currents, that interfere with the process, and accuracy of magnetic field variation.

It was developed, in this work, the project for a power source of a relaxometer, which makes possible to obtain accurate current values, very quick transitions between the possible current values, and the compensation of the so called parasitic currents, present in the magnet.

The method used is based on a power converter named Four-Quadrant Converter, which is a yet less used alternative, but with some advantages, than the utilization of auxiliary, field compensation, windings. The developed model was tested in computational simulation environment, and reveals promising results, fulfilling all the previously stated requirements and also achieving reasonably low power consumption.

I. INTRODUCTION

The study of materials at an atomic level is becoming more and more important, in the various fields of Physics and Chemistry, both for investigation and industrial purposes. Various industries have deep interests in the study of substances at a molecular level, including the Medical, Pharmaceutical, and Mining industries. These studies often include the observation of the variation of several properties of the materials when exposed to ambient abnormal characteristics such as temperature, pressure, acidic environments, among other possibly harmful variables.

The development of modern, sophisticated, and efficient technologies to help these scientific studies is of high importance. A relevant technique used nowadays by the laboratories that make these analyses is the Fast-Field Cycling Nuclear Magnetic Resonance.

1. Theoretical Background

The FFC-NMR is a technique based on the measurement of the spin-lattice relaxation times of the sampled substance. In nuclear physics, the relaxation is a set of processes in which the particles of the nuclei of a substance are subjected to a certain magnetization, that represents a forced, non-equilibrium state, and are then released of that magnetization force, and go back to the equilibrium state. When on the forced state, the particles of the nuclei often change their spin orientation. The relaxation time consists in the delay between the release of the magnetic excitation and the return of the nuclear particles to their original spin orientation.

The process of FFC-NMR is usually implemented by devices called relaxometers, and has several phases, during which the particles are submitted to repeated cycles of magnetic field exposure of different intensity. A regular cycle comprises three phases: The polarization phase, when a magnetic field with high field density (Bp) is applied in order to magnetize the particles and arrange them in a certain way. Then, the second step is the relaxation phase, where a very low field is applied during a certain interval, to let the particles rearrange to their equilibrium state, both magnetically, as well as thermal equilibrium. Finally, the detection phase, in which
a field of high flux density ($B_D$) is applied, preferably, during a short period of time.

Relaxometers are usually composed by a magnet, where the sample is put, an acquisition system, that is, most of the times a Radio-Frequency system, that sends periodic pulses, and the power supply, that should be able to provide the needed magnetic field, in the form of a current waveform. The main requirements of the power supply are: stability and accuracy values of magnetic field density on the magnet, at each phase of the cycle; speed on the transitions between the several possible levels of magnet current.

The main problem with the modern relaxometer power supplies has to do with particles that have low relaxation frequencies (~100Hz), which require the use of low fields. These fact leads to interference problems, namely with the Earth magnetic field, that causes the existence of additional, unwanted magnetic fields in the relaxometer magnet.

2. Motivation and Objectives

As stated above, one of the main challenges in this field nowadays is to build relaxometers that can detect a wider range of spin-lattice relaxation times. In the low frequencies operation, external magnetic fields, like the earth own magnetism, interfere with the technique operation, by leading to the appearance of parasitic currents on the magnet which, of course, lowers the accuracy of the induced magnetic field. The solution must be implemented on the power supply side, and may have different approaches.

The goal of this work is to project and size a power supply that implements a system of magnetic field compensation, to counter the effect of the parasitic currents. The methodology used is based on the utilization of a reversible current power supply, namely, a Four-Quadrant Converter. This converter should be able to perform the usual FFC-NMR operation, while compensating the parasitic currents by allowing the current to flow in both senses in the magnet.

A regular cycle of FFC-NMR operation comprises the three stages referenced above, plus the transition phases between the steady stages. The current on the magnet should vary between a certain maximum value, $I_{\text{max}}$, and a minimum value, $I_{\text{min}}$ (Fig.1).

![Figure 1 - Example of a cycle of current in a relaxometer.](image)

In this paper, it will be presented the solution for the relaxometer, with the topology of the converter, the design of the control system, the calculation of the power losses on the semiconductors, and their respective heat-sinks.

II. Operation Principles

The power supply complete circuit includes the converter, and the control system that is based on PI controller that receives the reference signal for the magnet current and the reading of the actual current on the magnet, by a Hall Sensor.

![Figure 2 - Global representation of the power source for the relaxometer.](image)

The objective of using the 4Q converter to solve the problem at hand is to provide the load with a range of current values of $[-I_o, I_{\text{max}}]$ in order to obtain a range of magnetic flux density values of $[0, B_{\text{max}}]$. To do that, the converter must supply current in the stages of positive values of B, and absorb the current at the stages of zero field, and at the fall stages.

The typical cycle of work used as reference here, also called Regular Cycle, consists in a polarization stage, of intermediate magnetic field value ($B_{\text{middle}}$), followed by the relaxation stage, with a, preferably, null magnetic field ($B_{\text{min}}$), and finally, the detection stage, with the maximum magnetic field value ($B_{\text{max}}$), as shown in Figure 1. Typically, the polarization and relaxation stages have equal duration, about 40% of the period each, and the detection stage is shorter, with about 20% of the period, T.

The 4Q converter chosen has 2 Diodes and 2 IGBTs (identified as $D_1$, $D_2$, $S_1$ and $S_2$), two DC voltage sources of 32V, and an RC filter, to help smooth the transitions. Additionally to the 4Q Chopper, there is a circuit, to help increasing the speed of the transitions, called boost circuit. This is achieved through a capacitor, connected to a high-voltage source ($V_B$), both put in parallel with the load. This circuit is only connected to the load during the Rise and Fall transitions, providing and absorbing current from the magnet, respectively. This means that this circuit should also have electronic switches (IGBTs and Diodes, identified as $D_3$, $D_4$, $S_3$ and $S_4$) in order to connect and disconnect it from the rest of the circuit. The rest of the converter should be disconnected during these stages, in order to maximize efficiency, and also to protect those semiconductors. In the branch of the RC filter there should be another switch, $S_5$, in order to isolate it from the boosting circuit in the rise stage, preventing the Boosting Capacitor from unnecessarily sending high amounts of current to the capacitor of the filter.

The switching of the various semiconductors, five IGBTs in this case, is done with a control system, that has both a PI controller, responsible for the linear control, of the main 4Q converter, and some logic control, to command the remaining semiconductors. The representation of the entire circuit is
presented above, and each part of it is explained in this section and in the following sections of the document.

To explain the operation of the circuit, it is useful to use the equivalent circuits for each stage of operation. First, as the IGBTs are treated as switches, they can be governed by a set of variables that define their binary state, called state variables, $\Gamma = \{\gamma_1, \gamma_2, \ldots, \gamma_n\}$. So, $\gamma_i$ is associated with the semiconductor $S_i$. The formal description is, for $\gamma_i$:

$$\gamma_i = \begin{cases} 0, & \text{if } S_i \text{ is OFF} \\ 1, & \text{if } S_i \text{ is ON} \end{cases} \quad \forall i \in \{1, 2, 3, 4, 5\}$$

(1)

Starting with the Middle and High Stages, which have the same behavior, in terms of active switches, the processes can be described as:

**Middle/High Stage**

In this stage, $S_1$ switch alternately, to maintain the magnet current at the desirable level, with the help of the capacitor $C_0$, which charges and discharges continuously. $I_U$ is the current provided by voltage source, $U$.

**Fall Transition**

During the transitions, only $S_3$ or $S_4$ are ON, while $S_1$ and $S_2$ should be OFF, to avoid interferences with the process of charging/discharging the magnet. $S_3$ can be ON during the Fall transition, and in that case, it also sends current back to the boosting capacitor. The boosting capacitor shall be bigger than needed, i.e. should have a higher capacitance value than the value needed for the boost current, so it should be able to take the recharging current from both the magnet and the RC filter.

**Low Stage**

In the low steady-state, both semiconductors of the main converter ($S_1$ and $S_2$) conduct alternately, to maintain the current level on the magnet at 0A value. Again, the capacitor $C_0$ charges and discharges continually, which is important in order to have a low variation of the desirable DC value. The capacitor supplies current to the load when $S_1$ is ON and collects current when $S_2$ is ON.

As for the voltage on the load, in this stage it only varies between the two power supplies, depending on which semiconductor is active.

$$I_M = I_{S1} + I_{S2} + I_c$$

(7)

$$\gamma_1 I_U - \gamma_2 I_U + (\gamma_2 - \gamma_1)I_c = 0$$

(8)

$$V_M = \begin{cases} U, & \gamma_1 = 1, \gamma_2 = 0 \\ -U, & \gamma_1 = 0, \gamma_2 = 1 \end{cases}$$

(9)
Rise Transition

At this transition, the behaviour of the converter is very similar to the Fall transition, but now it is $S_3$ that is active. Also, $S_5$ is now OFF, so the filter is detached to the rest of the circuit so that the boosting capacitor supplies only the load and not the filter’s capacitor.

Boosting capacitor sizing:

As for the boosting capacitor, the sizing must be based on the energy it should keep and provide. The equation of the electric energy on a capacitor is:

$$ W_C = \frac{1}{2} CV^2 $$

(13)

On the other hand, the energy it must provide in each discharge to the magnet is:

$$ W_C = \frac{P_{0\text{max}} \Delta t}{2} $$

(14)

Where $\Delta t$ is the time it must take to discharge or charge completely, which is also the rise or fall time for the current on the load, an maximum power on the load is given by:

$$ P_{0\text{max}} = V_{H}I_{m\text{ax}} $$

(15)

Thus, the boosting capacitor can be obtained by:

$$ C = \frac{2W_C}{V_{H}^2} $$

(16)

The calculation of the maximum energy that it should hold must be done carefully, and it should always be given some margin. It is concluded that it is during the fall stage where there is maximum flow of energy, as the filter is connected in parallel to the boosting capacitor and sends some amount of current to it, plus the current from the load, in a very small period. Also, the boosting capacitor has a small resistor (less than 1Ω) connected in series with it, to avoid sudden peaks.

About the semiconductors, all the Diodes and IGBTs should sustain a voltage higher than $V_H$, so they are all chosen to have 600V tolerance. Also all these devices should support a maximum current of 20A.

III. CONTROL SYSTEM

The control system is based on feedback loop, with the value of the current read from the magnet, and a PI Controller, custom designed for this project. Besides the usual proportional and integral gains ($K_p$ and $K_i$, respectively), this controller has an additional feedback loop with a limiter block, and a gain associated, $K_w$ or saturation gain. This extra loop is used in order to prevent the signal to even reach the limits of the saturation block. This way, it is assured the control signal is always bounded within a certain limit.

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The controller is obtained by the transfer function of the system. It is a second order system with a pole and a zero, each with a time constant associated. The expression of the simple controller is obtained independently of the extra loop, and is given by:

$$C(s) = \frac{1 + \frac{s T_e}{s T_p}}{s T_p}$$

(17)

The gains are directly related to the time constants by:

$$K_p = \frac{T_e}{T_p} = \frac{L}{4\xi^2 K_D k_i T_D}$$

$$K_i = \frac{1}{T_p} = \frac{R}{4\xi^2 K_D k_i T_D}$$

(18)

(19)

For the saturation gain, $K_w$, the choice for the value depends on a condition imposed by the controller's topology.

$$\frac{1}{K_p} \leq K_w \leq \frac{K_i}{K_p}$$

(20)

The linear controller is used for the main part of the converter, it commands the switching of semiconductors $S_1$ and $S_2$, and helps achieving the main current waveform on the magnet. For the remaining semiconductors it is necessary additional logic control circuits. Specifically, $S_3$ and $S_4$ do not need any linear control, as they only need to switch ON once or twice per period.

For the command of $S_1$ and $S_2$ it is used a Flip-Flop for each of the semiconductors. One of the Flip-Flops shall enable $S_3$ to be ON when a rising edge is detected in the reference signal, and disable it when the error signal reaches zero value (Magnet current equals reference current). A parallel reasoning is used for the linear control of $S_4$, the semiconductor responsible for the dissipation of the magnet current. It should be ON when a falling edge is detected in the reference, and then disabled also when the error signal equals or passes zero.

As semiconductors $S_1$ and $S_2$ should not be conducting when $S_1$ or $S_4$ are ON, it should also be used some additional logic control for these devices. So, besides the control signal that commands the switching of these semiconductors during the steady stages, it must be assured that neither $S_1$ nor $S_2$ are ON. This is done with AND logic circuits and the inputs should be the negated outputs of the aforementioned Flip-Flops.

IV. SIMULATION OF THE NON-LINEARITY

The circuit in the Simulink should have not only the converter, but also some mechanism that simulates the effect of the magnetic saturation. In practice, the simplest way to look at this is as a logical problem. The affected stage of the cycle is the Down stage, where the current cannot reach the 0A value. The operation consists in injecting an extra signal in the retroactive loop of the control system, specifically in the measurement of the current in the magnet. The signal alternates between 0 and -2 and is constantly added to the measurement of the current. The choice of these alternating values is done with a Switch block, which is activated through a Flip-flop. This Flip-flop chooses the value 0 when a rise edge on the reference signal is detected. It stays like this until a falling edge on the reference and a zero value reference signal are detected. It is necessary to wait to both these two conditions to guarantee that the residual current is only added in the down stage and not during the fall transitions or the middle phase. Once the flip-flop is turned off, the switch chooses the second input which is the constant to be added. In the simulation the value chosen for the residual current is 2A but this is only exemplificative, and other values, or even other perturbation signals might be used and tested in the Simulink simulation.

V. POWER SOURCES

For the 32V sources of the main circuit there shall be only one transformer, with the secondary divided in two sub windings, using a middle tape that allows the use of two identical voltage sources, through the same power supply. The middle tape is connected to the ground reference, so this is ideal for a 4-Q converter such as this, as one of the windings on the secondary has the negative pole connected to the ground and the other winding has the positive pole connected to the ground. Naturally, two identical rectifiers are used, one for each of the voltage sources. Also, each of the sub-windings of the secondary operate separately and, according to the 4-Q converter’s operation principles, the two windings of the secondary are never used at the same time. The power of transformer is obtained by:

$$S_T = I_{\text{max}} U_{\text{max}}$$

(21)

For the Transformer, it should always be given some margin in the maximum power. For the capacitor, the reasoning used is based in the equation of a current that passes through a capacitor:

$$i_c = C_a \frac{du_c}{dt}$$

(22)

This capacitor must be sized in order to obtain the desired value of DC voltage to supply the circuit. If the maximum average value of the current is used, the capacitance can be given by:

$$C_a = \frac{I_{\text{max}} \Delta t}{\Delta U}$$

(23)

The maximum average current is observed through the
computational simulation, as the average current supplied by one of the 30V sources in a work-cycle where one of the power sources supplies the most current. Also, the period used in this case is in respect to the power source, which is the 50Hz grid. But, as the rectifier used is a full-bridge type, the period of the rectified wave is half of the wave from the power grid, 10ms.

For the high-voltage power source, it is also used a transformer with a different topology. Here there is only one source, so a regular transformer is used with a rectifier bridge and an auxiliary capacitor. The secondary of the transformer must be connected to the ground reference so that when the boosting circuit is activated, the magnet is connected to the same reference point as the transformer. Also, the rectifier bridge should only enable the flow of current in one sense, i.e. from the transformer to the boosting capacitor, and isolate the feeding circuit from the capacitor in the stages where the magnet current falls. This should be done so that the boosting capacitor can recharge itself from the load.

In this case, the power is obtained in a different way, as the main objective is to supply a certain amount of energy, to the boost-capacitor, in a short period of time. Once again, it should chosen a work cycle that represents the worst case scenario, in this case, in terms of minimum time to recharge the boosting capacitor. This may be a cycle with two rises of magnet current, with an interval of about 30ms between them.

The power of the transformer is then defined by the relation of the energy that should be stored in the boosting capacitor and the interval in which the capacitor should be recharged.

\[ \Delta W = V_H I_T \Delta t_1 \]  

(24)

\[ S_{T2} = V_H I_T \]  

(25)

The energy the capacitor receives from the transformer has to be equal (or greater) to the energy delivered from the capacitor to the load. This quantity is known, so the computation of the current can be done. From there, it is possible to calculate the power of the transformer.

The sizing of the capacitors for the rectifier follows the same principle and the same equations of the previous example. The voltage supplied should be as close to 500V as possible, so the rectifier’s capacitor should be sized in order to obtain a DC voltage as close to that value as possible.

VI. PROTECTION SNUBBERS

Snubbers are a part of the power electronics design, specifically part of the semiconductors protection equipment. In this case, the use of snubbers is mainly to protect the semiconductors of the sudden peaks of voltage and current, making them more progressive and reducing their value, because when the IGBTs turn on and off, there may occur peaks, which may affect the converter’s performance, increase the switching losses, and may even damage the devices. The semiconductors that need the inclusion of the snubbers are the IGBTs (S₁ to S₃) and the diodes (D₁ to D₄). The chosen snubber type for these semiconductors will be the simplest type, RC snubber, as it is sufficient to limit the quick rise of voltage levels, when the semiconductors turn off, and the rise of currents, when they turn on. Additionally, there may be added switching-aiding circuits, such as Zener diodes in parallel with the IGBTs. For sizing the RC snubber, the following equations are used:

\[ C_s = \frac{t_{Cap} t_f}{2U} \]  

(26)

\[ R_s = \frac{t_{ONmin} t_r}{3C_s} \]  

(27)

The parameters used are the voltage applied upon the switching off, the average current across the semiconductors, the rise and fall times, and the minimum ON time interval during a period of commutation (t_{ONmin}). This time is, naturally, not equal for all the semiconductors in the circuit. The resulting values for the capacitors and resistors, are, therefore, all different from each other, as the different semiconductors deal with different peaks of voltage and different durations of the stimulus.

<table>
<thead>
<tr>
<th>Component</th>
<th>Cₛ(pF)</th>
<th>Rₛ(kΩ)</th>
</tr>
</thead>
<tbody>
<tr>
<td>S₁</td>
<td>1000</td>
<td>5.03</td>
</tr>
<tr>
<td>S₂</td>
<td>75</td>
<td>56</td>
</tr>
<tr>
<td>S₃</td>
<td>15</td>
<td>45.6x10⁴</td>
</tr>
<tr>
<td>S₄</td>
<td>25</td>
<td>27.2x10²</td>
</tr>
<tr>
<td>S₅</td>
<td>1000</td>
<td>7.06</td>
</tr>
<tr>
<td>D₁</td>
<td>2.0</td>
<td>6.6</td>
</tr>
<tr>
<td>D₂</td>
<td>75</td>
<td>56</td>
</tr>
<tr>
<td>D₃</td>
<td>15</td>
<td>45.6x10⁴</td>
</tr>
<tr>
<td>D₄</td>
<td>25</td>
<td>27.2x10²</td>
</tr>
</tbody>
</table>

VII. SIMULATION RESULTS

The results shown come from a simulation with a regular cycle of operation with two Falls of the current and one Rise, and three steady stages of current (I_{high}, I_{middle} and I_{low}), with a period of T =100ms. It is interesting to observe the speed of the transitions of the current level, as well as other details, like the ripple of the current in the steady stages and the stabilization time of the current on those stages. As it can be seen, the transition from low to high stage is just above 4ms.

These current waveforms are proportional to the waveform of the magnetic field density on the magnets, which is not visible in the simulation, but is obtained experimentally, through the Hall Sensors.
The fall transition from the high level to the middle level is identical to the other fall transition, in terms of speed, and stabilization time, so it is no represented.

And a detailed view on each of the steady stages of the current shows that the variation from the DC point is almost irrelevant, as it is less than 1mA in the high and low stages and about 3.5mA in the middle stage. This is an important result as it should mean that the magnetic field has the same precision.

VIII. POWER LOSSES AND HEAT DISSIPATION

Power converters that deal with values of voltage and currents such as this, usually have a considerable amount of power losses. The accumulated heat will cause overheating, which can lead to serious damage to the circuit’s components, so it is of high importance to deal with this problem and incorporate components in the circuit that help the convection of heat. The power losses that need to be studied, for heat dissipation control purposes, are the losses of the semiconductor, in this case: diodes and IGBTs. The different types of losses in these devices are: conduction losses, commutation or switching losses, and Joule losses. These power losses depend on a variety of factors and variables such as: the work cycle of the current on the magnet; the duration of the cycle; the filter; the boosting capacitor; the specifications of the semiconductors. To calculate the losses, it will be used the worst case scenario, i.e. it will be used the kind of operation that leads to the highest losses on each of the semiconductors.

So, each of the devices should be studied individually to conclude which mode of operation is responsible for the maximum losses with each semiconductor. The kind of operation cycles defined and considered for this study are:

- Regular Cycle, current level and fraction of the period: Polarization – 5A, 0.4T; Relaxation – 0A, 0.4T; Detection – 10A, 0.2T;
- Inverted Cycle: Polarization – 10A, 0.35T; Relaxation – 0A, 0.35T; Detection – 5A, 0.3T;
- Maximum Cycle: Polarization – 10A, 0.4T; Relaxation – 5A, 0.2T; Detection – 10A, 0.4T;
- Minimum Cycle: Polarization/Detection – 10A, 0.2T; Relaxation – 0A, 0.8T;
The max cycle is used to calculate the losses of the switches \( S_1, S_3 \) and \( D_1 \), as it should be the worst case possible for these switches. The min cycle is for the calculation of the losses of \( S_2 \) and \( D_2 \). The regular sequence and the reverse sequence should represent the highest losses for the switches of the boosting circuit (\( S_4, D_4 \), and \( S_5 \) and \( D_5 \), respectively).

1. **Semiconductor Losses**

About the kinds of power losses, for this power supply, the only semiconductor losses relevant are the switching losses for the IGBTs and the conduction losses for all of the semiconductors. This means that Joule losses and switching losses on the Diodes are not calculated as they are minimal. In the case of the Joule losses, it has to do with the size of the semiconductors, which have equivalent ON resistances of \( m\Omega \).

The general equation for conduction losses of the semiconductors is:

\[
P_{CI} = V_{CEsat}I_{Siav}
\]

For the \( i \)-th IGBT, and:

\[
P_{CI} = V_{RI}I_{Piav}
\]

For the \( i \)-th Diode.

And for the IGBT’s switching losses, the equation used is:

\[
P_{Si} = V_{CEoM}I_{SiM}\left(\frac{t_r + t_f}{2T}\right)
\]

For the \( i \)-th IGBT.

So, the calculation of the losses for the semiconductors need a lot of parameters, some that are constant, established, and accessible on the datasheets, like the saturation voltage of the IGBTs, and rise and fall times of the current. But it is still needed to obtain some others like the average currents on the semiconductors, and the maximum currents and voltages applied to them. These values are observed in the simulations, or calculated based on them, and must be carefully handled as for example, \( S_1, S_2 \) and \( S_3 \) do not conduct continuously so, the average current is not given directly.

2. **Joule Losses**

Besides the losses on the semiconductors it should be taken into account, for the total losses, the Joule losses on the magnet. As it will be observed, these actually represent the major part of the power losses, as there is current flowing in and out of the magnet during most of the period. The equation for the Joule losses is:

\[
P_{MJ} = R_M I_M^{RMS}^2
\]

The magnet resistance has the value of \( R_M = 3\Omega \), and the root-mean square value of the current is obtained in auxiliary calculations and has the value of 5.52A. The value of the Joule losses for a regular cycle is 91.51W. The maximum possible for this quantity occurs with the max cycle, where there is always current in the magnet during the entire period.

It is now presented some statistics regarding the total power losses, which show the distribution of the losses per semiconductor, or per phase of the period, in different situations.
3. Heat Sinks

The inclusion of heat sinks in a power electronics system is of the highest importance, as it is vital to drain the heat out of the semiconductors case in order to maintain the device in its safe operating area and preventing it from overheating and subsequent harmful consequences. Each of the semiconductors of the converter should then have attached a heat sink designed specifically for it.

The design of the sink consists in determining the thermal resistance it must have in order to dissipate the necessary heat. The model used for sizing the sink is based on heat conduction and convection, where each space of conduction is represented by its thermal resistance.

Generally it is considered the thermal resistances of the following parts: junction to case (jc); case to sink (cs); sink to ambient (sa). The model is a heat circuit, equivalent to an electric circuit, that treats the thermal resistances as resistors, and their junction points as nodes.

![Figure 20 - Representation of the thermal circuit for a semiconductor, and its heat sink.](image)

Figure 20 - Representation of the thermal circuit for a semiconductor, and its heat sink. Tj – Temperature of the junction; Tc – Temperature of the case; Ts – Temperature of the sink; Ta – Ambient temperature; Rth(jc) – Junction-to-case thermal resistance; Rth(cs).

To calculate the thermal resistances it is applied a relation between it and the dissipated power, an equation that resembles Ohm law, for heat propagation:

\[ P_d = \frac{T_j - T_a}{R_{th(ja)}} \]  

(32)

Each semiconductor has its own established parameters, of maximum junction temperature, and internal junction-to-case thermal resistance \( R_{th(jc)} \), accessible on their datasheets. So, the point is to calculate the remaining thermal resistances, for a certain power dissipation. These resistances, \( R_{th(cs)} \) and \( R_{th(sa)} \) refer to the heat sink, so this are the parameters that need to be calculated to size the sink. As the relation between the two of them is not known, it is assumed that the thermal resistance of case-to-sink and sink-to-ambient are equal, and from now on defined as only one resistance, \( R_{th(ca)} \).

Therefore, the equation for the sink of the \( i \)-th semiconductor is:

\[ R_{th(ca)} = \frac{T_{j_{max}} - T_a - P_{d_i}R_{th(jc)_i}}{P_{d_i}} \]  

(33)

As for the temperatures, it should be used the maximum junction temperature referred for each device on its datasheet, and Ta is the ambient temperature, defined as 25°C. The total power losses of the worst case scenario for each semiconductor, thermal characteristics and respective heat sink for each semiconductor are listed above.

### Table 2 - Total Losses, Thermal Characteristics, and Heat Sinks.

<table>
<thead>
<tr>
<th>Device</th>
<th>Conduction Losses (W)</th>
<th>Switching Losses (W)</th>
<th>Total Losses (W)</th>
<th>( T_{j_{max}} ) (°C)</th>
<th>( R_{th(jc)} ) (°C/W)</th>
<th>( R_{th(ca)} ) (°C/W)</th>
</tr>
</thead>
<tbody>
<tr>
<td>( S_1 )</td>
<td>14.25</td>
<td>10.4</td>
<td>24.64</td>
<td>150</td>
<td>0.42</td>
<td>4.7</td>
</tr>
<tr>
<td>( S_2 )</td>
<td>0.23</td>
<td>0.107</td>
<td>0.337</td>
<td>150</td>
<td>0.42</td>
<td>373</td>
</tr>
<tr>
<td>( S_3 )</td>
<td>0.35</td>
<td>0.015</td>
<td>0.365</td>
<td>150</td>
<td>0.42</td>
<td>343</td>
</tr>
<tr>
<td>( S_4 )</td>
<td>0.58</td>
<td>0.023</td>
<td>0.603</td>
<td>150</td>
<td>0.42</td>
<td>205</td>
</tr>
<tr>
<td>( S_5 )</td>
<td>8.36</td>
<td>0.016</td>
<td>8.376</td>
<td>150</td>
<td>0.42</td>
<td>14.5</td>
</tr>
<tr>
<td>( D_1 )</td>
<td>24.04</td>
<td>-</td>
<td>24.04</td>
<td>150</td>
<td>0.8</td>
<td>4.4</td>
</tr>
<tr>
<td>( D_2 )</td>
<td>0.38</td>
<td>-</td>
<td>0.38</td>
<td>150</td>
<td>0.8</td>
<td>325</td>
</tr>
<tr>
<td>( D_3 )</td>
<td>0.59</td>
<td>-</td>
<td>0.59</td>
<td>150</td>
<td>0.8</td>
<td>212</td>
</tr>
<tr>
<td>( D_4 )</td>
<td>0.99</td>
<td>-</td>
<td>0.99</td>
<td>150</td>
<td>0.8</td>
<td>126</td>
</tr>
</tbody>
</table>

As expected, the magnet has far superior power loss values, as there is current on it for almost the whole period of operation, unlike the semiconductors. The semiconductors are grouped in pairs \( S/D \), because they both conduct at the same time, and both conduction losses are accounted. In the last graph, the losses accounted are in respect to the worst case for each of the semiconductors, and the magnet. It also includes the switching losses of the IGBT, although they represent a very insignificant part of the semiconductors losses, and even less in the total.

### IX. Conclusions

#### 1. General Considerations

This model allowed for a simulation that showed that all the normal requisites of a FFC-NMR relaxometer were achieved, particularly, the main simulation used a magnet current \( i_m = \{5, -2, 10\} \) [A], with \( T=100\)ms, and the rise time of the current from -2A to 10A equal to 4.2ms. Other parameters of this simulation are the current ripple, which reaches the maximum value of 3.5mA in the middle stage of steady current \( (i_{r_m} = 5A) \); and the maximum overshoot of the current is equal to 40mA, in the same stage.

The problem of the parasitic currents was successfully beaten, proving the validity of the solution of the reversible current power supply. In comparison with the existing solution in IST, that uses the technology of extra windings to compensate the parasitic currents, the IST prototype takes advantage in the speed criterion but with the extra winding it deals with more losses, more heating, and extra power sources. Besides, in the 4-Q the control is more direct, in opposition to the prototype that needs to have a control system over the main converter and over the current supplied to the compensation winding.

Through the simulation and further calculations it was possible to have some good perspectives on the converter’s overall efficiency. It demonstrated relatively low power...
consumption, and low power losses by the semiconductors. Such fact is due to the converter’s topology that allows each semiconductor to work only a fraction of the period, as they almost always conduct alternately. This might be relevant not only for the heat dissipation issue, but for the conservation of the devices, and the whole power supply lifetime.

2. Applicability and Future Work

The promising results obtained in the simulation environment help the idea that the application of this model to a practical power source could very well be implemented. This thesis was done with the objective of projecting the model and simulating it on a virtual environment but always thinking in the practical side of it, in order to leave it ready to be implemented. In that way some questions of practical interest and highest relevance are also addressed, such as the design of the heat sinks, the feeding of the converter, and the description of the custom projected PI controller, so that in the near future, the concretization of this project can be achieved. The question of the recharging of the boosting capacitor is left open, as it may or may not be implemented that way, but it can always be put in a way that receives current from the transformer and sends back some to the grid.

As this topology is not used either in IST or by a company that has commercial models for a NMR power supply, it might turn to be an innovative device, as it effectively compensates the parasitic currents, with precise control over the converter, and shows low power losses and low power consumption overall.

X. REFERENCES


ANNEX

Constants/Parameters:
Period of a Cycle: $T = 100\text{ms}$;
Main Voltage Source: $U = 32\text{V}$;
High Voltage Source: $V_H = 500\text{V}$;
Magnet: $L_M = 200\text{mH}$; $R_M = 3\text{Ω}$;
Ambient Temperature: $T_a = 25^\circ\text{C}$;
Current Levels: $I_{\text{max}} = 10\text{A}$; $I_{\text{middle}} = 5\text{A}$; $I_{\text{min}} = -2\text{A}$