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Investigation of current sensing techniques for a high bandwidth application
Focusing on Hall-Current Sensing

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Abstract

With the ongoing surge in electrification within the automotive industry, the demand for high-power inverters is steadily on the rise. Furthermore, these inverters are now expected to handle increased power loads while maintaining enhanced efficiency. Fulfilling these demanding prerequisites, places significant importance on the current sensors that meticulously monitor each phase of the inverter.

This thesis tackles the deep study of a current sensor that has been chosen by Scania to be introduced in the new generation of inverters. The chosen sensor is based on the Integrated Magnetic Concentrator (IMC)-Hall technology. Given its innovative nature within the company’s scope, there exists a keen interest in investigating its performance. The principal parameters that will be assessed during this thesis are accuracy, temperature dependencies, bandwidth and noise immunity.

The evaluation of these properties will be done though the creation of a test rig capable of producing current pulses, spanning the range necessary for the sensor evaluation. A high-end current sensor based on the fluxgate technology will be used as a reference during the assessment. Across the course of the research, diverse experiments and modifications to the testing setup will be undertaken to accommodate the assessment of each individual parameter. The acquired data will be summarized and presented with simplified figures.

The results of this study show that the accuracy of the sensor highly depends on a proper end of line calibration and a mechanical assembly, in other words, the sensor must be perfectly centered within the IMC. Moreover, the study identifies linear relationships between temperature and accuracy, while exponential correlations are found between the impact of an external magnetic field and its distance.

Keywords

Current sensing, High bandwidth, Inverter, Hall-sensors

Sammanfattning

Med den pågående ökningen av elektrifiering inom fordonsindustrin ökar efterfrågan på högeffektsväxelriktare stadigt. Dessutom förväntas dessa växelriktare nu hantera ökad effektbelastning samtidigt som de bibehåller förbättrad effektivitet.
Att uppfylla dessa krävande förutsättningar lägger stor vikt vid strömsensorerna som noggrant övervakar varje fas i omriktaren.


Resultaten av denna studie visar att sensorns noggrannhet i hög grad beror på korrekt ”end of line” kalibrering samt mekanisk montering. Med andra ord måste sensorn vara perfekt centrerad inom IMC. Dessutom identifierar studien linjära samband mellan temperatur och noggrannhet, medan exponentiella korrelationer finns mellan påverkan av ett externt magnetfält och dess avstånd.

**Nyckelord**

Strömavkänning, Hög bandbredd, Inverter, Hall-sensorer

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Chapter 1

Introduction

An area that is growing exponentially nowadays is electrification, which intends to replace applications that use fossil fuels with technologies that use electricity as a source of energy. For example, the European Green Deal [1] aims to reduce 90% of transport emissions. When it comes to the vehicle’s industry, the objective is to replace the Internal Combustion Engines Vehicles (ICEV) with Electric Vehicles (EV) or Hybrid Electric Vehicles (HEV). The international Energy Agency (IEA) foresees a growth in the market of electric vehicles from 5 million in 2018 to 130-250 million by 2030 [2]. Given that the electric motors that are used for such applications are powered with alternating current (AC), a high power inverter is necessary to convert the direct current (DC) from the batteries. Such inverters, need high performance and reliable current measurements in order to protect and control the system, at the same time, these sensors need to prove a high resistance to the electromagnetic interferences (EMI), given that they are in really noisy environments. Therefore, there is a growing interest in investigating new current sensing techniques which can improve the overall performance and robustness of those inverters. With that in mind, the goals of this thesis are to perform a background study of different current sensing techniques. This study will encompass an analysis of their individual merits and limitations. Later, the new current sensors that Scania plans to use in the next generation of inverters will be further evaluated. A thorough testing of them will be made in order to determine if they can reach the expected performance for the application.
1.1 Problem statement

When it comes to current measurement for high power inverters, the most used sensors are the shunt resistors. These sensors are quite big given that they have to measure high currents, and they have to dissipate some power, which makes them difficult to integrate within the inverter. Therefore, it is necessary to look into other techniques that can keep the same performance but with a better form factor. On the other hand, for a better control and protection of the inverters, it is necessary to have a higher bandwidth on the current measurement. Usually, an improvement of bandwidth comes with the trade-off of a decreased accuracy of itself, thus, it is necessary to have a sensor that has both, good accuracy and bandwidth in order to fulfill the requirements of the application. Finally, due to the noisy environment where the inverters are situated, it is necessary that this current sensors have a good robustness against EMI.

1.2 Research questions and goal

Is it possible to use highly integrable current sensors and increase its bandwidth without losing accuracy or noise immunity?

1.3 Purpose

From an industrial point of view the purpose of this thesis is to investigate new current sensing techniques for a new generation inverter, and prove whether it fulfills the strict requirements on this new component demand, in terms of accuracy, bandwidth and noise immunity. From an academic point of view, the outcomes of this thesis will give not only specific results of new current sensing technologies, but also knowledge on its testing method and simulations of high power circuits.

1.4 Research Methods

Firstly, a deep analysis of different current sensing techniques will be done. This will give a base to support further discussions later in the thesis. Secondly, a test rig has to be designed. This needs to accomplished both, the safety of working with high power in the lab, and enable a thorough testing of the sensor under evaluation. To really understand how well the sensor works,
tests regarding its accuracy, bandwidth, noise immunity and temperature drift will be performed. All the data will be compared with a reference sensor.

### 1.5 Delimitations

Even though an extended background is presented with different current sensing techniques, only one kind of sensor will be studied deeply. This sensor has been chosen previously by Scania. Therefore, it does not exist a requirement specification presented in this thesis that dictates which is the most appropriate sensor for this application.

As working with high power is of high risk, it was mandatory to work with low voltage during this thesis. This will limit the capabilities to provide specific current values necessary for an extended assessment of the sensor under evaluation.
Introduction
Chapter 2

Background

2.1 Sensors characteristics

When a new system design, which requires a sensor, is proposed, it is important to define the requirements of the application and therefore choose the sensor accordingly. If a too good sensor is chosen it will be an "overkill" and increase the price of the system, whereas if the sensor is too bad the requirements of the application will not be met, and the system will not work properly[3]. The most important characteristics of a sensor are described in the following sections.

2.1.1 Transfer function

From the input stimuli of a sensor (temperature, pressure, current, voltage, etc.), the output will be a function of it. We can consider a sensor as a black box, that performs several conversion steps before it outputs a signal, analog or digital, that is readable for a Data Acquisition Device (DAQ). Hence, the transfer function of a sensor is defined as the relation between its unknown input stimuli and its output[4]. For example, if we’re talking about a current sensor that turns current into voltage, the transfer function might look something like:

\[ V_{out} = f(I) \]  

(2.1)

As mentioned above, a sensor is usually attached to a measuring system. One of the tasks of the measuring systems is to infer the unknown value of \( I \) from the measured value of \( V_{out} \). therefore, the measurement system has to employ an inverse transfer function to obtain the unknown value:
It is also desirable to obtain a transfer function of not only the sensor alone, but also its interface circuit, such as amplification of filtering stages.

The transfer function of a sensor can have different shapes or complexities. Those functions are usually given in the datasheet of the sensor, but in case that one is designing the sensor from scratch, a process called curve fitting often comes into play. Within this procedure, empirically measured data points are systematically linked to input stimuli that are acquired by an alternate instrument or sensor (commonly regarded as the reference or true value). As per now onward, the output of the sensor will be named $E$, while the input or stimuli will be names $s$. The most common transfer function types are reviewed below:

- **Linear:** A linear transfer function is described with the following expression:

$$E = A + Bs$$  \hspace{1cm} (2.3)

Where $A$ is the value of the output in absence of input stimuli, often called offset. And $B$ is the rate of change of the output in relation to the input. It is usually called sensitivity.

The linear transfer functions are usually the most sought-after among the system designers, given that it is really easy to compute its inverse.

- **Logarithmic:** The logarithmic approximation transfer function is described as:

$$E = A + B\ln(s)$$  \hspace{1cm} (2.4)

Where $A$ and $B$ are fixed parameters.

- **Exponential:** The logarithmic approximation transfer function is described as:

$$E = Ae^{ks} + B$$  \hspace{1cm} (2.5)

Where $A$, $k$ and $B$ are fixed parameters.
Polynomial: We might encounter the case that a sensor cannot be approximated to any of the basic functions. In such instances, a viable approach involves employing a polynomial approximation. In most cases a sensor’s response can be approximated by the second (see equation 2.6) or third (see equation 2.7) degree polynomials.

\[ E = a_2s^2 + a_1s + a_0 \]  \hspace{1cm} (2.6)

\[ E = a_3s^3 + a_2s^2 + a_1s + a_0 \]  \hspace{1cm} (2.7)

### 2.1.2 Accuracy

The accuracy of a sensor is described as the difference between the value that is computed at the sensor output and the true value that it is measuring. We have to accept the true value with some kind of uncertainty since it is not possible to be sure what the true value is. The accuracy of a sensor can be represented in different ways:

- As a range of the physical value that is being measured, e.g. ±1°C.
- As a percentage of the full scale (FS) of the sensor, e.g. ±0.5%FS.

### 2.1.3 Precision

The precision of a sensor is described as how close are the measurements of a sensor to each other under unchanged conditions. It can also be related to its repeatability or reproducibility. The precision can be quantified by calculating the difference of two output readings under the same conditions, and it is usually represented as a percentage of the full scale:

\[ \delta_r = \frac{\Delta}{FS} \times 100\% \]  \hspace{1cm} (2.8)

The precision can also be quantified by plotting a histogram or a probability density function of a measurement under unchanged conditions. The number of measurements that are outside a certain number of standard deviations will tell us some information regarding the precision.
2.1.4 Resolution

When a sensor system maps a continuous real world signal (pressure, temperature, etc.) into discrete values described as quantization. Given that it’s mapped into a discrete set of value there will always be some information lost from the continuous signal. Then, we define resolution as the smallest change in the physical quantity that will produce a response in the measurement, in other words, the smallest change of the physical quantity that the sensor will be able to detect.

2.1.5 Bandwidth

For a slow changing input stimulus, the transfer function of a sensor may be described as non time-dependent. Whereas, if the input signal changes really fast it will be noticeable that the sensor cannot follow it with the same speed. Therefore the transfer function of the sensor can be described as time-dependent characteristic which is called dynamic. The bandwidth of a sensor is described as how fast the response of a sensor can follow the variance of a physical magnitude. The bandwidth of a sensor is expressed in Hz and it is measured at the -3dB point, which is the point where the sensor output value differs -3dB(-29.3%) from the physical magnitude.

2.2 Current sensing techniques

In this section an overlook of the most used current sensing techniques will be done. Additionally the performance and limitations of each technique will be mentioned.

2.2.1 Resistive-based

Resistive-based current sensing techniques sense the voltage drop across a resistor, which is called the sensing resistor. By applying the well-known Ohm’s Law and a constant sensing resistor value the current can be calculated as:

\[ I = \frac{\Delta V}{R_{\text{sense}}} \]  \hspace{1cm} (2.9)

Deduced from the above equation, it can be seen that the current is proportional to the voltage drop across the resistor, where the proportional
constant is \( \frac{1}{R_{\text{sense}}} \). Therefore, we can consider this sensors as having a linear transfer function. Resistive-based current sensing techniques are usually considered to have a lower price compared to other techniques, as well as good accuracy and a simple principle of working. On the other hand, its main disadvantage is the power loss on the \( R_{\text{sense}} \), and the non-isolated voltage measurements, which might be a problem if there are transient voltages in the load. Its bandwidth is relatively high compared with other techniques, being in the \( MHz \) range. However, given that usually the values of \( R_{\text{sense}} \) used are quite low some amplification and signal filtering needs to be included in the design, which will affect the overall bandwidth of the system.

Some resistive-based current sensing techniques are presented below:

- **Shunt resistors:** Are the most common within the resistive-based techniques. They are used on a wide range of applications but when it is necessary to measure large current for long duration of time they are limited by its power dissipation. They can be modeled by its nominal resistance \( R \), a parasitic inductance in series \( L_s \) and a parasitic resistance is series \( R_s \), usually due to the skin effect. The parasitic impedance of the resistor start to be noticeable at frequencies of 1MHz. Shunt resistors are usually build with materials that exhibit a very low temperature coefficient such as manganese-copper or nickel-chrome alloys, this makes that the temperature dependency or thermal drift of this resistors is very low.

- **Voltage drop on a MOSFET:** When MOSFET's are switched on, they exhibit a resistance (\( R_{DS-\text{on}} \)) between their drain and source terminals, which depends on the voltage between its gate and source terminals (\( V_{GS} \)). Given that the resistance of the MOSFET is known, either from the datasheet of the component or by calculating it:

\[
R_{DS-\text{on}} = \frac{l}{W \mu C_{ox}(V_{GS}) - V_T}
\]  

- \( l \): channel length
- \( W \): channel width
- \( \mu \): carrier mobility
- \( C_{ox} \): gate oxide capacitance
- \( V_T \): threshold voltage
By calculating the voltage drop between its drain and source terminals ($V_{DS}$) we are able to calculate the current flowing through it:

$$I_D = \frac{V_{DS}}{R_{DS - on}}$$  \hspace{1cm} (2.11)

The use of this technique has the advantage that it is a lossless technique, given that no extra $R_{sense}$ need to be introduced in the system. It is a good technique when measuring low-voltage and high-current, however, it loses accuracy when the switching voltage or the switching frequency increases.

- **Voltage sense of a copper trace:** When the voltage drop across a copper trace is sensed, there is no introduction of additional power losses into the system compared to using a shunt resistor. Nevertheless, the implementation of this technique may encounter challenges due to the thermal drift of the copper material and the manufacturing tolerances associated with copper traces or busbars. On the other hand, induced current to the sensing circuit and the skin effect might cause several deviation while measuring AC currents. Due to the novelty and inherent challenges associated with this technique, there is limited existing research in this field. However, [5] presents a novel solution where they use a temperature sensor to compensate the temperature drift of the copper resistance. Additionally, a compensation network is utilized to address the skin effect and mitigate induced voltages in the sense wires.

### 2.2.2 Sensors based on Faraday’s law of Induction

The Faraday’s law of induction is a physics principle that relates the change of a magnetic field and the induction of an electromotive force (EMF), also presented as $\varepsilon$, in a single conductive loop. Faraday’s law is expressed as [6]:

$$\varepsilon = -\frac{d\Phi}{dt}$$ \hspace{1cm} (2.12)

Where $\Phi$ is the magnetic flux, and it is described for a surface $\Sigma$ whose boundary is a wire loop as:

$$\Phi_B = \int_{\Sigma(t)} B(t)dA$$ \hspace{1cm} (2.13)

Emil Lenz later formulated that the direction of the electromotive force induced in a conductor by a changing magnetic field is such that the magnetic
field created by the induced current will oppose the initial magnetic field. Therefore, the negative sign on equation 2.12 appears. This statement is named as Lenz’s law.

Given the nature of this law, only current changes over time can be measured, therefore the measurement of DC currents is a limitation for sensors that are based on this law.

**Current transformers**

A current transformer (CT) consists of a single primary winding and a secondary winding of several turns which is wrapped around the primary conductor, usually in a toroidal form. This gives a proportional relation between the primary current \( i_p \) and the secondary current \( i_s \) expressed as:

\[
i_s = \frac{i_p}{N}
\]  

(2.14)

Where \( N \) is the number of turns on the secondary winding. The current transformers are usually loaded with a sense resistor \( R_s \), which makes it easier to sense the primary current with a proportional voltage in the secondary:

\[
V_{sense} = \frac{I_p}{N} R_s
\]  

(2.15)

It must be noted that this is a very simple version of a CT and it is only valid for low frequency models. On higher frequencies, there appears the interwinding capacitance \( C_w \), which will limit the bandwidth of the system. As pointed in the previous section the CTs are only able to measure AC currents. However, some advantages of using CTs are; it provides an isolated measurement, low losses, simple working principle and a proportional output voltage from the input current, without any need of further amplification.

In contrast, the design of a current transformer presents certain engineering obstacles such as those related to core hysteresis and saturation, which appear when the peak magnetization current saturates the transformer core. Additionally, it needs to be taken into account the core losses resulting from hysteresis and eddy currents.

**Rogowski coil**

A Rogowski coil is an air core coil with many turns, wrapped in a toroidal form, it is used for measuring alternating currents and high speed pulses [7]. It is based on Ampere’s and Faraday’s laws.
First, Ampere’s law relates the magnetic field inside the Rogowski coil with the current $i_C$ that is flowing across the enclosed area of the coil:

$$ \int_C \vec{B} \cdot d\vec{l} = \mu_0 i_C $$

(2.16)

If we assume that the cross section radius of the coil is much smaller than its radius $r$, then the magnetic field $B$ is simplified as:

$$ B = \frac{\mu_0 i_C}{2\pi r} $$

(2.17)

Later, by applying Faraday’s law of induction, we can determine the voltage induced in the coil. Equations 2.12 and 2.17 are merged:

$$ v = -N \frac{d\Phi}{dt} = -NA \frac{dB}{dt} = -\frac{NA\mu_0}{2\pi r} \frac{di_C}{dt} $$

(2.18)

Theoretically, equation 2.18 should be valid for any shape of the Rogowski coil and for any position of the primary current inside of the loop formed by the Rogowski coil. However, it has been demonstrated that the imperfections on the winding density affect the accuracy of the system, therefore, if the primary current conductor is not centered with the coil, the current measurement accuracy will be decreased, as a consequence, if the main current conductor is situated next to the clip of the coil, this inaccuracy will be even more noticeable, due to the fact that in that point it is impossible to ensure a constant density in the winding of the coil [8].

As it can be seen from equation 2.18, the nature of the Rogowski coil requires an integrator at its output in order to relate its output voltage $V_{sense}$ with the current flowing through the sensed conductor $i_C$:

$$ V_{sense} = -k \frac{NA\mu_0}{2\pi r} i_C + v_{out}(0) $$

(2.19)

Where:

- $k$: Integrator gain.
- $v_{out}(0)$: Output voltage when zero current is applied at the primary.

When it comes to implement an integrator, the more common to use are the RC and the Op-amp integrators. The RC integrator is appropriate for frequencies higher than 100MHz, whereas the Op-Amp integrator is appropriate for frequencies lower than 100MHz as it reaches its usual bandwidth figures. The use of an integrator will reduce the noise thanks to its low-pass filtering characteristic, but it will limit the bandwidth of the system.
The principal drawback of the Rogowski coils is that it cannot measure DC currents, but on the other hand, it has several advantages respect other devices that can only measure AC currents, such as, it does not experiment saturation, it is flexible, it has a large bandwidth and it has a non-intrusive nature among others.

### 2.2.3 Magnetic field sensors

Magnetic field sensors can sense both, static and dynamic fields, thus it is a perfect solution for applications that need to sense not only AC currents but also DC currents, which results in an advantage compared with the sensors based on the Faraday’s law of induction.

In order to understand further working principle of these sensors it is important to review the Lorentz force law. In physics, it is described as the combination of electric and magnetic force in a point of charge due to electromagnetic fields. When a charged particle, such as an electron or a proton, moves through a region where there are both electric and magnetic fields, it experiences a force known as the Lorentz force. This force is given by the following equation:

\[
\vec{F} = q(\vec{E} + \vec{v} \times \vec{B})
\]  

Where:

- \(\vec{F}\): Lorentz force experienced by the charged particle in N.
- q: Charged particle in C.
- \(\vec{E}\): Electric field in V/m.
- \(\vec{v}\): velocity vector of the charged particle in m/s.
- \(\vec{B}\): Magnetic field vector in T.

**Hall effect**

Hall Effect sensors are utilized for the quantification of direct current (DC) and alternating current (AC) while maintaining electrical isolation. These sensors operate based on the fundamental principle of the Lorentz force, wherein they detect the magnetic field \(B\) generated in the vicinity of the current-carrying conductor intended for measurement.
They are formed by a transducer, usually called hall plate, and an analog front end (AFE). The hall plate is usually a rectangular piece of semiconductor material with two pair of terminals placed in a orthogonal manner, see figure 2.1. In one pair, the bias electric field $E_{bias}$ will be applied, whereas the other pair will be used to sense the Hall voltage $V_H$, generated by the Lorentz force. The hall plate will also be placed orthogonal to the magnetic field generated in the vicinity of the current-carrying conductor [9].

![Figure 2.1: Hall plate scheme.](image)

If we describe the bias current as $I_{bias}$, the magnetic flux density created by this conductor as $B$, the constant that depends on the hall plate as $K$, the thickness of the hall plate as $d$ and the offset voltage of the plate in absence of magnetic field as $V_{OH}$, the Hall voltage $V_H$ is given by:

$$V_H = \frac{K}{d}BI_{bias} + V_{OH}$$  \hspace{1cm} (2.21)

The constant $K$ can be further separated out as, the Hall coefficient $R_H$ times the Hall geometrical factor $G_H$. For further information on $G_H$ please refer to [9]. On the other hand, $R_H$ is expressed as:

$$R_H = \frac{r_H}{nq}$$  \hspace{1cm} (2.22)

Where:

- $r_H$: Hall factor, dependent on the plate´s material. It has a value of 1 for the majority of standard materials.
- $n$: Carrier density.
- $q$: Charge of the current carrier.
As it can be seen in equation 2.21, the Hall voltage is directly proportional to the magnetic field that it is sensing, which makes this sensor really attractive due to its inherent linearity. However, the sensitivity of this sensors is quite low, which can be fixed with a higher gain of the AFE amplifier, consequently increasing the noise level of the sensor. Alternatively, an additional approach to enhance sensitivity involves employing magnetic concentrators, resembling yokes, see figure 2.2, or similar structures, which serve to concentrate the ambient magnetic field in the proximity of the sensor’s location.

With this arrangement, the magnetic field as a function of the sensed current $I_{\text{sense}}$ can be expressed as:

$$B = \frac{\mu_0 \mu_r I}{2\pi r - d + d\mu_r}$$

(2.23)

Where $d$ is the thickness of the air gap where the Hall sensor is placed and $r$ can be approximated to the radius of the yoke. A good design that satisfies $d\mu_r >> 2\pi r$ will allow to simplify the previous expression to:

$$B = \frac{\mu_0 I}{d}$$

(2.24)

With the use of yokes, relations as high as 1mT/A can be accomplished.
as well as robustness against EMI. However, this will limit the bandwidth of the sensor and it might require demagnetization methods, also called degaussing, in case of large over-currents that might magnetize the yoke. It is crucial to consider that any slight misalignment or off-center placement of the Hall sensor within the air gap of the yoke may lead to inaccurate readings of the sensor. Hence, the implementation of a robust production and assembly process becomes of utmost importance to ensure precise and reliable measurements.

An alternative to the use of yokes as concentrators is to install the Hall sensors in the PCB as SMD packages, in this way they can be placed directly on top of a copper trace for relatively low currents, or on top of bus-bars for larger currents. However, using this solution will also give rather low sensitivities of the sensor, thus the installation of magnetic concentrators is also used. In this case, the concentrators have a U-shape and they can be better integrated than the yokes into the whole design, see image 2.3.

![Figure 2.3: Magnetic concentrator as U shape](image)

Finally, another Hall effect configuration that is used to increase the performance of the sensor is called the closed loop. The closed loop configuration uses a feedback or compensation circuit based on a winding wrapped around the high permeability core of magnetic concentrator. This winding creates a flux that counteracts the flux generated by the primary current $I_{\text{sense}}$. This kind of configuration is known also as compensated or zero-flux given that the feedback loop created a flux equal in amplitude but opposite in direction to that produced by $I_{\text{sense}}$. By operating within the zero-flux region, issues such as hysteresis errors and linearity dependencies are effectively mitigated.
Magnetoresistance

Magnetoresistance is the property of a material to change its electric resistance under an externally-applied magnetic field. Those materials are usually permalloys or ferromagnetic. A Magnetoresistor (MR) is a two terminal device that changes its resistance parabolically with the magnetic field, this phenomena is called anisotropic magnetoresistive effect. Due to the natural characteristics of this effect, the transfer function of the resistance with the magnetic field is not linear and it has very low sensitivity. In order to improve this problem, a configuration called barberpole is used [8]. In this configuration, aluminum bars are wrapped around a permalloy trip in a fashion that resembles a barberpole, by doing this, the current through the MR is forced to flow at 45º angles to the magnetic field. At such angles, the transfer function is linear and also the sensitivity is higher.

The Giant Magnetoresistance Effect (GMR) is another effect associated with magnetoresistance. The GMR consists on stacking two ferromagnetic layers separated by a thin non-magnetic layer. With this structure, the resistance of the two ferromagnetic layer will vary as a function of the angle of magnetization of these layers, as a consequence, changes of up to 50% resistance are possible [10]. Drawbacks of the GMR are nonlinearity, a distinct thermal drift and notable hysteresis.

There are other MR sensors such the Tunnel Magnetoresistance (TMR). The TMR sensor bears resemblance to the GMR architecture; however, its operation is founded on the principle of quantum mechanical electron tunneling traversing from one ferromagnetic layer to the other. Notably, this phenomenon operates in accordance with the principles of quantum mechanics, which raises its complexity. We can also distinguish the Colossal Magnetoresistance (CMR), which is a property of manganese-based oxides of changing its resistance by orders of magnitude under a magnetic field. However, this last technique is still under development.
Background
Chapter 3

Practical case study: Hall effect sensors

In this chapter, the practical case study will be presented. The sensor that has been chosen in this thesis is a hall effect sensor. This particular choice has been driven by the strategic considerations of the sponsoring entity, Scania. The rationale for this selection stems from the sensor’s notable attributes, including its commendable integrability, a favorable trade-off between cost and performance, and its inherent electrical insulative properties. Furthermore, the sensors under evaluation are from the Belgian company Melexis [11], a global supplier of semiconductor devices, specially sensors. This company has offered Scania several advise, as well as, an evaluation kit with multiple prototypes. These prototypes serve the purpose of evaluating the efficacy of their Hall effect current sensors.

The application of the sensors is a three phase 2-level inverter. In this configuration, there are several MOSFET’s arranged in parallel to handle the switching of both the high and low sides of the load. This parallel configuration is aimed at distributing the current load among several MOSFETs, effectively minimizing current flow through each one. By doing so, power losses during the switching process are reduced. Two distinct driver stages are engaged to ensure synchronized control of all the high and low side MOSFETs. This synchronized operation is crucial for the proper functioning of the inverter. To achieve a precise sinusoidal output waveform, the MOSFETs are driven through Pulse Width Modulation (PWM) techniques. The control strategy is finely tuned in a real time manner thanks to a current sensor at the output of each phase, as well as a rotatory encoder a the output of the electric machine. Moreover, an additional current sensor will be positioned at the DC
link for additional protection or diagnosis. As dictated by the specifications of the inverter, the sensors need to be rated for 1200A of current.

According to the organization purposes, the sensors are anticipated to exhibit operational bandwidths spanning up to 400kHz. This broad frequency range serves the dual purpose of quantifying the switching noise attributed to the MOSFETs and enhancing the efficacy of the control strategy employed. It is important that these sensors demonstrate minimal thermal drift to ensure accurate measurements, due to their proximity to a high temperature zone within the inverter.

Finally, it is important that the sensors are not susceptible to external or adjacent magnetic fields, given that the bus-bars of each phase are placed quite close to each other.

3.1 Experimental circuit

Given that the inverter works at high power and it is restricted to perform tests under such high voltages, an alternative test setup needs to be developed. The main requirements of this experimental circuit is that it should operate under 60V which is considered low voltage at the organization. On the other hand, the experimental circuit needs to give currents of at least 1200A in order to test the whole dynamic range of the sensors under evaluation. Last but not least, the $\frac{di}{dt}$ needs to be sufficiently large in order to evaluate the bandwidth of the sensor.

In order to reach high levels of current with low voltage, the chosen solution consists on the circuit presented at Image 3.1. A power supply is connected between the terminals VDC and GND. The power supply will charge a capacitor (C1) to a voltage between 0 and 60V. An inductance (L1) will be connected to the capacitor in parallel, but separated by a switch (SW1A). At the moment that the switch is closed, all the energy that is stored in the capacitor will be transferred to the inductance in form of current.
Figure 3.1: Test circuit diagram

Considering a circuit consisting of a capacitor and an inductor, with very minimal resistance (mostly the cables, the capacitor’s Equivalent Series Resistance (ESR) and the inductor itself), its inherent reaction result is a damped oscillation. To prevent harmful negative voltage and sudden surges of current from affecting the power supply and capacitor, a pair of diodes (D1 and D2) are connected in parallel. These diodes serve to confine the oscillations between the inductance and themselves until the energy is dissipated in the inductance’s resistance, in other words, they have a freewheeling finality.

The current sensors denoted as $MES_1$ and $MES_2$ will be the current under evaluation and the reference sensors, while the voltage sensor denoted as $MES_3$ will be measuring the voltage at the capacitor, to control its discharge. In further chapters they will be explained.

To determine the values of the components, some iterations were carried out. First, if we consider the circuit to zero resistance, we can make some initial approximations. From equation 3.1, we know how much energy will be stored in a capacitor for a given potential. Additionally, equation 3.2, gives us insight into the required inductance value, under the assumption that all the capacitor’s energy transfers to the inductance. However, the resistance of the cables, the capacitor, and the inductance ($R_S$) will dissipate energy according to equation 3.3. This last term is time dependent and will complicate the calculation of the circuit parameters.

$$Q_C = 0.5 \times C \times U_{max}^2$$  \hspace{1cm} (3.1)

$$Q_L = 0.5 \times L \times I_{max}^2$$  \hspace{1cm} (3.2)
\[ E_R = P \ast t = R \ast I^2 \ast t \]  \hspace{1cm} (3.3)

Table 3.1: Capacitor characteristics.

<table>
<thead>
<tr>
<th>Brand</th>
<th>Kemet</th>
</tr>
</thead>
<tbody>
<tr>
<td>Model</td>
<td>ALS31</td>
</tr>
<tr>
<td>Voltage</td>
<td>63V</td>
</tr>
<tr>
<td>Capacitance</td>
<td>0.1F ± 20%</td>
</tr>
<tr>
<td>ESR</td>
<td>6mΩ</td>
</tr>
</tbody>
</table>

In order to proceed, some simulations using Simulink were conducted. A fixed capacitor value of 100mF is chosen, due to its availability and its price. Some important properties of the capacitor can be seen in Table 3.1. Thereafter, some iterations are done in order to find a proper value for the inductance that satisfies most of the requirements mentioned at the beginning of this section. Thanks to the Simscape engine within Simulink, it is possible to introduce non-idealities of the components, such resistance of the Inductance or the Capacitor. It is also possible to introduce initial state of components before the simulation, for example the initial voltage of the capacitor. After some simulations it has been decided that an inductance value of around 10uH should be used. This inductance is handmade. 10mm of cable have been wound around a plastic pipe in order to create a air core and avoid its saturation. Given that the physical size of the inductance was of low importance, this solution was taken. In order to dimension the inductance, equation 3.4 is used.

\[ L = \frac{10\pi\mu_0 N^2 r^2}{9r + 10l} \]  \hspace{1cm} (3.4)

Where:
- L: inductance in H.
- \(\mu_0\): vacuum permeability in H/m.
- N: number of turns of wire.
- r: radius of the core in m.
- l: length of the inductance in m.
In order to decrease the total length of the inductance, a set of windings has been stacked on top of the other, hence, after some calculation it is obtained that 11 turns of wire around the plastic pipe are necessary.

When addressing the diodes, heavy duty ones are selected given that they need to withstand high current surges. To address this requirement, a pair of diodes operate in parallel due to their $I^2t$ constraint. The main diode characteristics are presented on Table 3.2 [12].

Table 3.2: Main diode characteristics.

<table>
<thead>
<tr>
<th>Brand</th>
<th>Vishay</th>
</tr>
</thead>
<tbody>
<tr>
<td>Type</td>
<td>Schottky</td>
</tr>
<tr>
<td>Model</td>
<td>VS240U</td>
</tr>
<tr>
<td>$I_{F,max}$</td>
<td>4500 A</td>
</tr>
<tr>
<td>$I^2t$</td>
<td>92 kA²s</td>
</tr>
<tr>
<td>$V_F$</td>
<td>0.83 V</td>
</tr>
</tbody>
</table>

3.2 Sensor under evaluation

As previously mentioned, the evaluated sensors are Hall-effect based sensors from the company Melexis. These sensors utilizes the technology IMC-Hall, which stands for Integrated Magnetic Concentrator, illustrated in Figure 2.3. By using this technology, the magnetic field is concentrated around the sensor, effectively enhancing its sensitivity. Furthermore, the integrated magnetic concentrators serve as a shield against electromagnetic interference (EMI).

From Melexis’ extensive selection of current sensors, the chosen ones for evaluation in this thesis are the MLX91216 and MLX91218 models. The rationale behind this selection includes their high integrability, a crucial factor in the new generation of inverters. Additionally, these models represent the latest and most advanced options available to date.

Melexis provided dedicated evaluation kits for these sensors. Those evaluation kits include copper bars, U-shaped magnetic concentrators in three varying sizes, separate PCBs tailored for the MLX91216 and 91218 respectively, and finally a plastic holder which allows to mount all the parts together. Further information regarding the evaluation kits can be found here [13].

Each sensor model has an option code which specifies its sensitivity or supply voltage among other parameters. In this thesis, the sensor models that have been evaluated are MLX91216-LDC-ACV-001 and MLX91218LDC-
ARX-300. The most important specifications of these sensors are listed in Table 3.3. Notably, these sensors are capable of measuring both positive and negative currents, resulting in their output for zero current aligning with the midpoint of their output voltage range, 2.5V for the MLX91216 and 1.65V for the MLX91218. Finally, the output of the sensor is trimmed/saturated at 90% and 10% of supply voltage for the high and low full scale limits respectively.

<table>
<thead>
<tr>
<th>Model</th>
<th>MLX91216</th>
<th>MLX91218</th>
</tr>
</thead>
<tbody>
<tr>
<td>Option code</td>
<td>LCD-ACV-001</td>
<td>LCD-ARX-300</td>
</tr>
<tr>
<td>Sensitivity</td>
<td>60mV/mT</td>
<td>14mV/mT</td>
</tr>
<tr>
<td>Supply voltage</td>
<td>5V</td>
<td>3.3V</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>$250kHz$</td>
<td>$400kHz$</td>
</tr>
<tr>
<td>Step response time</td>
<td>$2\mu s$</td>
<td>$2\mu s$</td>
</tr>
<tr>
<td>Operational Magnetic Field Range</td>
<td>$\pm 60mT$</td>
<td>$\pm 100mT$</td>
</tr>
<tr>
<td>Linearity Error(Magnetic)</td>
<td>$\pm 0.5%FS$</td>
<td>$\pm 0.5%FS$</td>
</tr>
<tr>
<td>Noise$^a$</td>
<td>$6.5mV_{RMS}$</td>
<td>$175nT/\sqrt{Hz}$</td>
</tr>
<tr>
<td>Thermal Sensitivity Drift</td>
<td>$\pm 1.0%S$</td>
<td>$\pm 1.5%S$</td>
</tr>
</tbody>
</table>

$^a$ Note that noise is expressed as RMS for MLX91216 and as PSD for MLX91218

On the other hand, three different U-shaped laminated shields are included in the evaluation kit. For specific dimensions, please consult Table 3.4 and Image 3.2. The most important dimension parameter is the width of the shield, which would be represented as the magnetic field lines in Image 2.3 as a reference. The width of the shield will be the parameter that relates the magnetic field with the current carrying the bus-bar, as denoted by Equation 3.5.

$$B[mT] = 1.25 \frac{I[A]}{W[mm]}$$  \hspace{1cm} (3.5)
Table 3.4: U-shaped shields dimensions.

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Short name</td>
<td>LU15</td>
<td>LU20</td>
<td>LU25</td>
</tr>
<tr>
<td>Width</td>
<td>15</td>
<td>20</td>
<td>25</td>
</tr>
<tr>
<td>High</td>
<td>15</td>
<td>15</td>
<td>18</td>
</tr>
<tr>
<td>Length</td>
<td>13</td>
<td>13</td>
<td>13</td>
</tr>
<tr>
<td>Thickness</td>
<td>3</td>
<td>3</td>
<td>3</td>
</tr>
</tbody>
</table>

Figure 3.2: Magnetic concentrator dimensions

With the use of equation 3.5 and the sensitivity parameter of each sensor listed in table 3.3, the transfer function of each sensor can be extracted. See table below:

Table 3.5: Transfer functions for each sensor and shield.

<table>
<thead>
<tr>
<th></th>
<th>LU15</th>
<th>LU20</th>
<th>LU25</th>
</tr>
</thead>
<tbody>
<tr>
<td>MLX91216</td>
<td>$V_{OUT}[V] = 0.005 * I[A] + 2.5$</td>
<td>$V_{OUT}[V] = 0.00375 * I[A] + 2.5$</td>
<td>$V_{OUT}[V] = 0.003 * I[A] + 2.5$</td>
</tr>
<tr>
<td>MLX91218</td>
<td>$V_{OUT}[V] = 0.001167 * I[A] + 1.25$</td>
<td>$V_{OUT}[V] = 0.000875 * I[A] + 1.25$</td>
<td>$V_{OUT}[V] = 0.0007 * I[A] + 1.25$</td>
</tr>
</tbody>
</table>

However, table 3.5 shows the ideal transfer function of those sensors. As introduced in Section 2.1 there are certain non-idealities such as; production
tolerances, temperature, external noise or assembly imperfections that will cause the sensor to be inaccurate. Hereafter, those will be introduced. The real transfer function is defined as:

\[ V_{out} = S_I * I + S_B * SF + V_{oq} \]  

(3.6)

Where:

- \( S_B \): Magnetic sensitivity of the sensor, subjected to datasheet error.
- \( S_I = S_B * FF \): Current sensitivity.
- \( FF \): Field factor, subjected to mechanical assembly errors.
- \( I \): The input current.
- \( SF \): The stray field, considered as an unwanted input.
- \( V_{oq} \): The output voltage at 0A of input current.

Then, the output total error is given by:

\[ \epsilon_{V_{out}} = \sqrt{\epsilon_{S_I}^2 * I^2 + \epsilon_{S_B}^2 * SF^2 + \epsilon_{V_{oq}}^2 + \epsilon_{NL}^2} \]  

(3.7)

Where each of the individual errors is broken down as:

\[ \epsilon_{S_I} = \sqrt{\epsilon_{S_T}^2 + \epsilon_{S_a}^2 + \epsilon_{S_R}^2 + \epsilon_{S_L}^2 + \epsilon_{S_{Tol}}^2 + \epsilon_{vib}^2} \]  

(3.8)

- \( \epsilon_{S_T} \): Sensitivity thermal drift.
- \( \epsilon_{S_a} \): Sensitivity accuracy (Can be compensated with End-Of-Line(EOL) calibration).
- \( \epsilon_{S_L} \): Sensitivity lifetime drift.
- \( \epsilon_{S_R} \): Sensitivity ratiometry error.
- \( \epsilon_{T_{Tol}} \): Mechanical assembly fixed tolerances (Can be compensated with EOL calibration).
- \( \epsilon_{vib} \): Mechanical vibrations of the assembly.

\[ \epsilon_{S_B} = \sqrt{\epsilon_{S_T}^2 + \epsilon_{S_a}^2 + \epsilon_{S_R}^2 + \epsilon_{S_L}^2} \]  

(3.9)
• $\epsilon_{ST}$: Sensitivity thermal drift.

• $\epsilon_{Sa}$: Sensitivity accuracy (Can be compensated with EOL calibration).

• $\epsilon_{SL}$: Sensitivity lifetime drift.

• $\epsilon_{SR}$: Sensitivity ratiometry error.

\[
\epsilon_{V_{eq}} = \sqrt{\epsilon_{oT}^2 + \epsilon_{oa}^2 + \epsilon_{oR}^2 + \epsilon_{oL}^2 + \epsilon_{oH}^2} \tag{3.10}
\]

• $\epsilon_{oT}$: Offset thermal drift.

• $\epsilon_{oa}$: Offset accuracy (Can be compensated with EOL calibration).

• $\epsilon_{oL}$: Offset lifetime drift.

• $\epsilon_{oR}$: Offset ratiometry error.

• $\epsilon_{oH}$: IMC hysteresis.

• $\epsilon_{oN}$: Offset noise error.

\[
\epsilon_{NL} = \epsilon_{L} \cdot I[\%FS] = \epsilon_{L} \cdot \frac{I}{I_{FS}} \tag{3.11}
\]

Where $\epsilon_{L}$ is the linearity error expressed in the datasheet, and $I_{FS}$ is the full scale current of the sensor.

Even though equation 3.7 seems quite complex, it can be simplified. In the majority of measurements, the influence of stray fields can be disregarded, leading to the elimination of the equation’s second term. Moreover, errors such as thermal drift, lifetime drift and mechanical vibrations can also be neglected given the nature of the tests performed.

Sensitivity accuracy and mechanical assembly tolerances can be compensated through EOL calibration, this will be later explained on Section 4.1. In the same manner, offset error will be calibrated out.

However, the non-linearity error is immune to calibration. This particular error is predominantly attributed to the Magnetic Concentrator when its saturation limits are reached. It is worth noting that this error becomes evident as the sensor approaches its Full Scale (FS) limits. For an illustration of this behavior with different LU20 shields, refer to Figure 3.3.
3.2.1 Noise improvement

During the initial series of experiments, it was noted that the sensors used in the evaluation kit exhibited a notable level of noise, surpassing that of the reference sensor. In light of this observation, a decision was made to enhance the system’s performance.

To address this issue, the first approach involved employing twisted pairs of cables for both the power line and sensor output. However, this solution barely made any difference in the results.

Later, a new PCB design was undertaken. This design incorporated the inclusion of a Low-Dropout regulator (LDO) to mitigate power supply noise. Additionally, an RF connector was used into the design to facilitate the connection of a coaxial cable to the measurement system. Additionally, a rectangular cut on the ground plane of the board was added aiming to minimize the eddy currents under the sensor position as recommended in the Melexis design guide.

Figure 3.3: Linearity representation for some LU20 shields
Figure 3.4 shows the 3D model of the new PCB. The width of the PCB is maintained, ensuring compatibility with the aforementioned plastic holder. However, the length of the PCB was slightly increased given that two new components were introduced.

3.3 Reference sensors

The main current transducer that is used the most as a reference sensor is the IN 1000-S from the supplier LEM. Operating on the principle of LEM’s patented fluxgate technology, this sensor finds prominent use in laboratory and industrial settings due to its exceptional performance. Selected specifications for this sensor are outlined at Table 3.6 [14].
## Table 3.6: LEM sensor characteristics.

<table>
<thead>
<tr>
<th>Model</th>
<th>In 1000-S</th>
</tr>
</thead>
<tbody>
<tr>
<td>Supply voltage</td>
<td>±15V</td>
</tr>
<tr>
<td>Nominal primary current</td>
<td>1000A</td>
</tr>
<tr>
<td>Maximum measuring range(^a)</td>
<td>1500A</td>
</tr>
<tr>
<td>Peak primary current</td>
<td>5000A</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>440kHz</td>
</tr>
<tr>
<td>Step response time</td>
<td>1µs</td>
</tr>
<tr>
<td>Linearity Error</td>
<td>±3ppm</td>
</tr>
<tr>
<td>Noise</td>
<td>34ppm</td>
</tr>
<tr>
<td>Temperature coefficient</td>
<td>±0.3ppm</td>
</tr>
</tbody>
</table>

\(^a\) Single pulse only. The transducer may require a few seconds to return to normal operation when autoreset system is running.

Subsequently, a determination was made to incorporate a faster sensor for the bandwidth tests. This decision arose from the similarity in bandwidth specifications between the LEM sensor and the Melexis sensors. As a result, the inclusion of a Rogowsky coil was deemed necessary. Despite the trade-off involving the loss of DC current measurement due to this sensor technology, its impact on the test was considered negligible given the experimental circuit’s current pulse, which lacked a relevant DC component.

With this context in mind, the selected sensor for this purpose is the PEM-CWT6, with its specifications listed in Table 3.7.

## Table 3.7: Rogowsky coil characteristics.

<table>
<thead>
<tr>
<th>Supplier</th>
<th>PEM</th>
</tr>
</thead>
<tbody>
<tr>
<td>Model</td>
<td>CWT6</td>
</tr>
<tr>
<td>Peak current</td>
<td>1200A</td>
</tr>
<tr>
<td>Sensitivity</td>
<td>5mV/A</td>
</tr>
<tr>
<td>Accuracy</td>
<td>±2% of reading</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>30MHz</td>
</tr>
<tr>
<td>Noise</td>
<td>mV/p – p</td>
</tr>
</tbody>
</table>

### 3.4 Data acquisition

The data acquisition is made using Dewesoft, a high end data acquisition system, with compatibility for a wide range of sensors and configurations.
It also allows sampling rates of up to 15MHz with an Analog to Digital Converter (ADC) resolution of 24bit.

The module used in this thesis is a Sirius XHS [15] which has 4 High Voltage inputs and 4 Low Voltage inputs. One of the high voltage inputs will be used to measure the voltage at the capacitor while the low voltage ones will be used for measuring the sensor under evaluation and the reference sensors.

Thanks to its flexible User Interface it is quite easy to set up new sensors. For example, the LEM sensor is automatically recognized by the software and its transfer function and measuring range are set. On the other hand, the Melexis sensor is easily set by introducing its sensitivity and offset levels in accordance with Table 3.5.

3.5 Experiments

Under this section a description of the experiments carried out during the thesis is done. First, an insight of how the test is set together with the experimental circuit is given. Later, it is reasoned which data is acquired in order to evaluate and characterize the sensors under evaluation.

3.5.1 Main sensor characteristics

This test is the most extensive of them all. It consists on using the circuit explained in Figure 3.1, and the LEM as a reference sensor. Each combination of sensor model and magnetic concentrator is tested in order to evaluate every sensor transfer function.

For each combination of sensor model and magnetic concentrator, a series of tests are done using a different starting potential on the capacitor. Once the sensor has reached its saturation limit, the next sensor-magnetic concentrator pair will be tested. All those tests are conducted under steady temperature and vibrations. The assembly of the sensor and magnetic concentrator is always done using the plastic holder provided by Melexis in the evaluation kit.

With this test, the accuracy and the noise of the sensors will be evaluated.

3.5.2 Temperature drift

The aim of this test is to evaluate how sensitive the sensor is to a high temperature environment. As expected in a real case scenario, this sensor will be placed inside the housing of the inverter, which due to the switching of the high power MOSFETs will experience a high temperature increase.
It’s essential to acknowledge that both the sensor and the ferromagnetic concentrator will encounter temperature increases. These components are both influenced by thermal drift; the sensor due to its electronic elements, and the ferromagnetic concentrator due to alterations in its magnetic properties (Curie Temperature) and its physical expansion.

In order to proceed with this test, the sensor will be encapsulated in a plastic box. Thereafter a thermoelement will regulate the temperature inside the whole environment. In this way, both the sensor and the ferromagnetic concentrator will reach the desired temperature. As a result, it will be possible to evaluate the temperature dependencies of the whole system. The rest of the circuit will be kept the same as mentioned in previous sections. Current pulses are generated in order to validate the Melexis sensor against the LEM sensor.

Figure 3.5 shows the test setup. The thermoelement adheres to the red plate at the base of the plastic enclosure, while the adjacent orange rectangular component functions as a temperature sensor, providing feedback control. The temperature regulation system employed is a MINCO CT425, which, in tandem with computer interfacing, permits the specification of temperature values and control strategies. During the test, the enclosure remains sealed to ensure the entire environment maintains the intended temperature conditions.
3.5.3 Stray field susceptibility

The aim of this test is to evaluate how susceptible the sensor is together with the magnetic concentrator against external magnetic fields. Since the bus-bars are routed in parallel inside the inverter housing and within a close distance it is highly probable that the magnetic field produced by one phase of the inverter disturbs the other phases. As hinted previously, one of the functions of the magnetic concentrators is to shield the sensors against other magnetic fields that might disturb the sensor’s signal. However, a test is necessary to verify how good the shield work against those interferences.

The test setup is shown in Figure 3.6, the upper bus-bar is the one that contains the sensor under evaluation. As it can be seen, the bus-bar is not connected to any current, as a result its output should be zero ideally. On the other hand, the lower bus-bar is connected to the circuit explained in Section 3.1. This one will carry the current pulse which might disturb the sensor under evaluation. In order to acquire the main current pulse, the LEM sensor is added to the main current path, in this way, we are able to synchronize the moment that the main current pulse is produced and the interference is acquired.

This test is repeated for different distances between the bus-bars as well as for two current peaks.

Figure 3.6: Noise pickup setup
3.5.4 Dynamic characteristics

The aim of this test is not only to test the bandwidth of the sensor but also other timing characteristic such as the step response time. To achieve this, fast current pulses are necessary. Therefore, the creation of two new inductors of $1\mu\text{H}$ and $5\mu\text{H}$ was undertaken. Unfortunately, it came to light that even with these smaller inductors, the produced current pulses remained slow. This slowness in rise time could be attributed to the inherent wire resistance and the Equivalent Series Resistance (ESR) of the capacitor, collectively dampening the anticipated rapidity of the current pulse rise.

As a solution, it was proposed to insert the evaluation sensor kit into a prototype inverter, which was being tested during that time at the company. Even though it was a really noisy environment and the generated currents were not really high, some interesting results were extracted from that test and will be exposed in the next section.

When testing the dynamic characteristics of the sensor, the Rogowsi coil introduced in Section 3.3 was used as reference sensor, as it exhibit an extraordinary high bandwidth.
Chapter 4

Results

4.1 Sensor calibration

As outlined in Section 3.4, the standard Melexis sensor transfer curve, extracted from its datasheet, is initially incorporated into the data acquisition system. However, as introduced in the last part of Section 3.2, there are several factors that might affect the sensor’s error and need to be calibrated out.

To begin with, addressing the offset error involves calibration at the start of each experiment. To do so, there exists the possibility to zero the output of the sensor by just clicking a button at the User Interface (UI) of the data acquisition system. Therefore, at the start of each experiment, when there is no current flowing, the acquisition system will store the output voltage of the sensor. That term will be used as the offset for the transfer function of the sensor during that experiment. This practice is commonly referred to as "start-up zeroing."

Continuing with the current sensitivity error, the best way to calibrate it out is by performing an EOL calibration. Unfortunately, this kind of calibration was not possible to be performed at the company due to the lack of instrumentation capable of producing a wide range of constant currents according to the FS of the sensor. Nevertheless, a Software calibration was proposed, based on the assumption that the LEM sensor provide us a ‘true value’ of the measured current.

This Software calibration is executed using Dewesoft, as a post processing of the acquired data. By dividing the data of the LEM sensor and the Melexis sensor, a gain factor is obtained. This gain factor will later be multiplied with the Melexis time-series curve, thereby producing the calibrated curve. For instance, in the lower display of the Figure 4.1 we can see the LEM sensor (marine blue) and the Melexis sensor (cyan blue). Evidently, a sensitivity
error is present as the difference between the curves widens with the measured current. However, the upper display shows the LEM sensor (marine blue) and the calibrated Melexis sensor (pink) with a gain factor of 1.12. Worth noting is the near-invisibility of the LEM sensor line, as the evaluated sensor closely tracks its performance.

In principle this calibration seems to work quite well. However, the amplification of the noise seen in the Melexis sensor is its main drawback. This is due to the fact that the multiplying factor is usually larger than 1. The amplification gains employed for individual sensor and magnetic concentrator combinations are detailed in Table 4.1. Surprisingly, for the pair MLX91218-LU15 no additional gain is applied as it turns out to be perfectly calibrated. This will be discussed later.

Table 4.1: Gain applied to each sensor after calibration.

<table>
<thead>
<tr>
<th>Sensor</th>
<th>LU15</th>
<th>LU20</th>
<th>LU25</th>
</tr>
</thead>
<tbody>
<tr>
<td>MLX91216</td>
<td>1.12</td>
<td>1.1</td>
<td>1.11</td>
</tr>
<tr>
<td>MLX91218</td>
<td>1</td>
<td>1.1</td>
<td>1.1</td>
</tr>
</tbody>
</table>
4.2 Accuracy assessment

In order to assess the accuracy of the Melexis sensor we start from the assumption that the LEM sensor provides a reference ‘true value’. Assessing the accuracy of the Melexis sensors was not an easy task to do. Most of the time the observed errors could be attributed to a poor dynamic performance of the sensor. However, given that both sensors share a similar bandwidth and the generated current pulses comfortably stay within the sensors’ bandwidth limits, these discrepancies could potentially stem from non-linearity, deficient calibration, or stray magnetic fields.

By using the Dewesoft UI, further processing of the acquired data is done in order to obtain some accuracy metrics. After the calibration of the sensor detailed in Section 4.1 the time-series data of the Melexis sensor is differentially compared to that of the LEM sensor. This yields a time-series curve depicting the magnitude difference between the two sensors at each time point. Since the Melexis sensor signals exhibit notable noise characteristics (see Section 4.3), a low-pass filter with a cutoff frequency of 10kHz is employed to attenuate the error signal’s noise. For instance, Figure 4.2 shows the LEM and recalibrated Melexis sensor on the top, while the yellow line on the bottom represents the error and the red line is the filtered error.

![Figure 4.2: Accuracy assessment procedure](image)

In order to present the results, the maximum value of the filtered curve will be taken and referred to as the maximum absolute error, for each experiment. For each pair sensor-shield several experiments are made, depending on the
initial charge of the capacitor, larger current pulses will be generated, referred as full scale current pulse. As expected, the larger the generated current pulses are, the larger the error becomes. Therefore, it was decided to present the results as percentage of the full scale of the measurement. This percentage is calculated through the subsequent equation:

\[
\text{Abs.error}^{[\%FS]} = \frac{\text{maximum absolute error}}{\text{full scale current pulse}} * 100 \quad (4.1)
\]

It must be taken into account that the larger the magnetic concentrator is, the larger the full scale of the sensor becomes. In a similar manner, the MLX91218 possesses a capacity to capture larger magnetic fields than the MLX91216. Consequently, the evaluation of accuracy will be contingent upon the full scale range of each sensor-magnetic concentrator pair. Figure 4.3a plots the absolute error of the MLX91216 sensor for each magnetic concentrator, while Figure 4.3b plots the absolute error of the MLX91216 sensor for each magnetic concentrator. It is worth noting that relative to the full scale of each measurement the error remains constant for each sensor-magnetic concentrator pair. It is also interesting to point out that the highest error is found at the point where the current is decreasing. This observation could potentially suggest that some dynamic characteristics of the sensor are causing this error.

![MLX91216 absolute error.](image1)

![MLX91218 absolute error.](image2)

**Figure 4.3**: Accuracy assessment of the sensors under evaluation.

### 4.3 Noise

As the first measurements started to take place, noise quantization became more important. It was observed that the noise acquired in the Melexis sensor
was way larger than the LEM sensor. This can be seen in Figure 4.1 as the Melexis sensor line is thicker than the LEM sensor.

In order to properly quantize this noise levels, we will zoom in to the beginning of each experiment, where there is no current flowing through the bus-bar and the sensor output should be 0A. For instance, Figure 4.4 shows the noise level of the Melexis sensor (Yellow) and the LEM sensor (Blue). The first one presents a peak to peak noise level of 16A while the second one is 3A.

Figure 4.4: Noise comparison between Melexis and LEM

It is important to note that the larger the Magnetic Concentrator, the larger the peak to peak noise that we observe. That is explained by referencing to Equation 3.5, where the sensitivity of the magnetic concentrator is inversely proportional to its size. Therefore, for the same sensor a smaller sensitivity will provide a larger measuring range. In order to disregard the noise amplification due to the shield size, the results will be given in $\text{mV}$ as it is the output of the sensor. Finally, according to the sensor’s datasheet quantization, a Root Mean Square (RMS) calculation over 10,000 data points windows will be done in order minimize the noise randomness over frequency.

Table 4.2 presents the noise quantization for each pair of sensor and magnetic concentrator. It is worth noting that the noise present on each sensor remain consistent regardless of the specific magnetic concentrator in use. For the MLX91216 sensor, the noise level is approximately 7.05mV, aligning closely with the specifications outlined in Table 3.3.

On the other hand, the supplier of the sensor had provided an equation in order to approximate the translation between the PSD and RMS quantization for the noise of the MLX91218 sensor. This relationship is expressed in Equation 4.2. When this equation is applied to the MLX91218 specifications across its entire bandwidth of 400kHz, it yields an RMS noise value of
1.55mV. Remarkably, this value closely corresponds with the noise data presented in Table 4.2.

\[
\text{Noise}_{(\text{RMS})} = \text{Noise}_{(\text{PSD})} \times \sqrt{BW} \times \text{Sensitivity}
\]

(4.2)

Where:

- \(\text{Noise}_{(\text{RMS})}\) is given in mV.
- \(\text{Noise}_{(\text{PSD})}\) is given in \(\frac{nT}{\sqrt{Hz}}\).
- BW is given in Hz.
- Sensitivity is given in \(\frac{mV}{nT}\).

Table 4.2: Noise quantization.

<table>
<thead>
<tr>
<th></th>
<th>MLX91216</th>
<th>MLX91218</th>
</tr>
</thead>
<tbody>
<tr>
<td>Noise (mV_{p-p})</td>
<td>50</td>
<td>60</td>
</tr>
<tr>
<td>Noise (mV_{RMS})</td>
<td>7.1</td>
<td>7.05</td>
</tr>
</tbody>
</table>

Concerning the new PCB that was designed to improve the noise of the system, the results deviated from initial expectations. Contrary to anticipations, the noise level exhibited no reduction when contrasted with prior experiments. Consequently, the data garnered through the utilization of these new PCBs will be omitted from this report. Nevertheless, this outcome offers valuable insight, indicating that the observed noise likely does not originate from the power input or the measurement cables. This revelation provides a promising starting point for the process of noise elimination.

4.4 Temperature dependence

The temperature dependence test has only been done for the MLX91218 sensor model and the LU15 shield given the difficulties and hazards of heating the enclosure (as depicted in Fig. 3.5) to high temperatures. As explained in Section 4.1 this combination of sensor and Magnetic concentrator does not require of any additional calibration, thus the data acquired in this experiment is processed directly.
The experiment has been executed with three test temperatures, 60°C, 80°C and 100°C. Upon reaching the target temperature within the enclosure, an identical current pulse peaking at 604A was applied to the circuit. It’s worth noting that the LEM sensor remained unaffected by these temperature variations, consistently capturing the same peak current throughout the testing process, thereby validating the integrity of the experimental setup.

In order to present the results of the experiment, the same procedure as in accuracy assessment (Section 4.2) is applied. As illustrated in Figure 4.5, the data reveals a nearly linear relationship between the absolute error, expressed as a percentage of the full scale, and the temperature. This relationship can be estimated as $0.037\%$/°C. If this relationship is extended up until 120°C, we get a error of 4.4% of the full scale of the sensor, which is almost three times larger than the error expressed in the datasheet.

![MLX 91218 Temperature dependency](image)

Figure 4.5: Temperature dependency of MLX91218

### 4.5 Stray field dependence

The assessments of stray field dependence were conducted using the setup introduced in Section 3.5.3. This experiment was only executed with the MLX91218 and LU20 magnetic concentrator. The focus lays primarily on understanding the correlation between the sensor’s distance from the stray field source.
Tests are done using two different current pulses of 1460A and 2550A, at varying distances between the Melexis sensor and the stray field generator. Those high currents were chosen as it was realized that lower current pulses would barely affect the Melexis sensors in terms of noise pickup.

Figure 4.6 in the upper panel displays the current pulse generated in the main circuit and captured by the LEM sensor (depicted as the marine blue line). In contrast, the lower panel exhibits the MLX91218 (represented as the cyan blue line), which is not connected to any current, but only affected by the stray field. The red line represents the filter signal of the Melexis sensor, processed through a low-pass filter with a cut-off frequency of 10kHz.

![Figure 4.6: Stray field dependency test results](image)

In a similar manner as before, the results will be presented as a maximum value derived from the filtered curve of the Melexis sensor, in order to avoid its random noise error. Figure 4.7 illustrates the impact of the stray field on the Melexis sensor concerning the distance at which the stray field is produced. Notably, the connection between the pickup current and the distance from the stray field source follows an exponential pattern for a constant current pulse.
4.6 Dynamic characteristics

As previously introduced, employing smaller inductors within the primary circuit failed to yield current pulses of sufficient rise time to effectively test the sensor’s bandwidth. The data garnered from these two inductor experiments closely mirrors the previously presented data, prompting us to disregard these results.

Nevertheless, when the Melexis sensor is connected to the output of the previously mentioned inverter, some interesting dynamic characteristics can be seen. Figure 4.8 shows one output phase of the inverter captured by the MLX91218 (depicted in blue), with a Rogowski Coil (shown in yellow) acting as the reference sensor. In this scenario, the output phase of the inverter is oscillating at 50kHz and it has a peak to peak current of 100A. It is clearly visible that the MLX91218 curve is delayed with respect to the Rogowski coil, a distinction verified by the two cursors superimposed on the graph. The temporal disparity between these cursors, noted at $2.1\,\mu s$ and detailed in Table 3.3 under the step response time parameter, is discernible in the lower right corner of the display.

Regarding the bandwidth limitation it is difficult to distinguish since the slope of both sensors seem quite similar and there is a quite noisy environment. However, during the switching instants, it can be discerned that the sharp edges
that the Rogoski coil captures, seem to be round shaped by the Melexis sensor.

Figure 4.8: Output current of the inverter
Chapter 5

Discussion

5.1 Accuracy

Initially, it’s essential to note that the accuracy outcomes have been derived using two methods that, while not entirely free of errors, have been chosen for their simplicity. Those methods are: the use of the LEM sensor as a reference value, and the use of a low pass filter in order to attenuate the noise of the sensor under evaluation. While the former has some inaccuracies in the ppm range over its full scale, the latter might filter out some noise components that might be relevant in the error evaluation.

Nonetheless, the extracted results of accuracy are summarized in Figure 4.3. It is interesting to point out that after scaling the results to the full scale of the sensor, each sensor-magnetic concentrator pair maintains a consistent error value across its entire dynamic range. This observation implies that the observed error is independent of the measured current’s magnitude.

Furthermore, it is evident that both sensors exhibit similar errors when paired with the LU20 and LU25 shields, registering values of 2.25% and 4.5%, respectively. In contrast, the LU15 shield displays an error of 0.5% for the MLX91216 and 1% for the MLX91218. This discrepancy could potentially be attributed to the absence of calibration for the MLX91218 when utilized alongside the LU15 magnetic concentrator. On the other hand, the errors presented by the LU20 and LU25 are most probably due to mechanical tolerances during assembly, as previously highlighted by a Melexis expert during a meeting. It is important to mention that the shape of the plastic holder in the evaluation kit, which consistently places the sensor in a non-centralized position relative to the magnetic concentrator, thereby generating an inherent error. Moreover, the uniform error displayed by both sensors
in conjunction with LU20 and LU25 magnetic concentrators suggests that assembly’s mechanical tolerances represent the primary contributor to this error. The fact that the assessment was conducted under steady temperature conditions and without exposure to vibrations lends further credence to this interpretation.

5.2 Noise

Noise analysis will be addressed separately for each sensor due to its distinct presentation in each sensor datasheet.

For the MLX91216 sensor it is observed that the peak to peak noise spans between 50mV and 60mV, which is translated to values of up to ±10A after employing the transfer function. At a first glance, these values seem notably elevated. However, after applying the ‘root mean square’ the 7mV obtained, precisely match with the value indicated in the sensor’s datasheet.

In contrast, for the MLX91218 sensor, the observed peak-to-peak noise ranges from 22mV to 26mV, approximately half of the noise level measured in the MLX91216. Nevertheless, considering this sensor’s broader dynamic range, this translates to a peak to peak noise of up to ±20A after applying its transfer function. In a similar manner as before, when translating its specification from PSD to RMS, we notice a remarkably similar profile. The measured noise stands at 2.4mV, while the calculated value is 1.5mV.

When it comes to the PCB manufactured for an improved noise immunity, it was observed that there was no improvement from the previous used PCB’s. This suggests us that the noise observed in the sensor is not originated at the power supply or the measuring cables. Instead, it is proposed that all this noise is generated in the Integrated Circuit itself.

5.3 Temperature dependence

While the dataset for the temperature test is limited, it provides sufficient information to initiate the observation of a linear correlation between the error and temperature rise. This error, presented as a percentage of the full scale in relation to temperature, is quantified at 0.037%/°C. This error is nearly three times greater than what is indicated in the datasheet. This can potentially be attributed to the omission of the error generated in the magnetic concentrator due to the alteration of its magnetic properties.
Given this particular relationship, rectifying this error through software-based post-processing seems straightforward. Essentially, the sensor output could be effectively corrected by employing a multiplication factor. A thorough study of temperature dependency will be later presented as future work.

5.4 Stray field dependence

Similarly to the previously discussed point, the dataset here is limited, having been tested solely with a single sensor-magnetic concentrator pair, and under conditions of exceptionally high current peaks. Nevertheless, preliminary insights can be drawn from the acquired results, suggesting an exponential correlation between the impact of stray fields and the distance between the sensor and the stray field generator.

With this results we could consider that the errors are still low in contrast with the dynamic range of the sensor. For instance 0.3% for a 10cm distance or a 1.7% for a distance of 4cm. Moreover, the high current peak, as depicted in Figure 4.7, should only be considered during over-current events. Luckily, this events should be detected by the inverter’s processor.

5.5 Dynamic characteristics

The evaluation of the dynamic characteristics of the sensor became a difficult task given that the current pulses that the main circuit was able to produce were too slow. Even though the use of smaller inductances, it was seen that the current pulse was too slow. This can probably be explained with the fact that the thick cables used in the setup were introducing some inductance to the circuit.

Another interesting behavior that might be related to the dynamic characteristics of the sensor has been observed during the evaluation of the sensor’s main characteristic experiments. Specifically, it becomes apparent that the most significant absolute error occurs when the current decreases. Notably, the Melexis sensors consistently register lower current levels compared to the LEM sensor at that point. While this discrepancy likely stems from some dynamic aspects of the magnetic concentrator, pinpointing the precise underlying cause remains challenging, even after consulting with Melexis application engineers.

However, upon integrating the sensor into the inverter’s output, a delay of
2\textmu s between the Melexis sensor and the Rogowski coil became apparent (refer to Figure 4.8), aligning precisely with the information provided in the sensor’s datasheet as step response time. Although examining the slope of each sensor in the same figure revealed minimal disparity, indicating that the bandwidth limit had yet to be reached, a notable limitation appeared in capturing the edges or switching points by the Melexis sensor. This limitation could potentially reduce its applicability, particularly if capturing the switching noise of the MOSFETs constituted a primary objective.
Chapter 6

Conclusions and future work

6.1 Conclusions

In this thesis the deep study of a novel current sensor has been undertaken. This current sensor has been chosen by Scania for a prototype inverter as it has a favorable balance between cost and quality, together with its high integrability.

The study has been carried out by evaluating the main parameters of interest of the sensor such accuracy, noise, temperature drift, immunity to external magnetic fields and bandwidth. All those parameters are evaluated by comparing the output of this sensor with a high-end reference current sensor.

The assessment of accuracy underscores that the evaluated sensors perform as anticipated when paired with the LU15 magnetic concentrator. However, a significant decline in accuracy is evident when the same sensors are employed with the other two magnetic concentrators. This is attributed to the mechanical tolerances introduced by the evaluation kit, and it is expected to improve if the sensor is properly installed in the real-world application.

The observed noise influencing the sensor surpasses that of the reference sensor, yet it remains within the specifications outlined in the datasheet. Despite efforts of enhancing it by building an improved PCB for the sensors, it was not accomplished. Consequently, the source of this noise may be entirely attributed to the sensor itself.

The sensor’s temperature dependency has been demonstrated to increase linearly with the temperature. However, no thermal drift offset has been observed as the sensor was zeroed before each experiment through the data acquisition system.

The sensor’s bandwidth could not be fully tested due to a lack of specific
equipment. However, the step response time was successfully assessed and found to align with the values outlined in the datasheet.

In conclusion, we are now ready to address the research question. Our comprehensive sensor evaluation has consistently demonstrated the ability to attain high bandwidths without significant compromises in accuracy or noise immunity, particularly in the case of the smaller magnetic concentrators that were subject to testing. However, when measuring larger currents, the necessity arises to employ larger magnetic concentrators. It is important to note that this comes with a trade-off, as larger concentrators can lead to a reduction in both accuracy and noise immunity.

As a result, we can deduce that the performance of the entire system is constrained by the characteristics of the magnetic concentrator. Nevertheless, it remains an indispensable component for enhancing noise immunity and sensitivity, ensuring the overall effectiveness of the system.

6.2 Future work

For a future work the first proposal is to design a better holder that can hold the LU20 and LU25 magnetic concentrators and the sensor in a well centered position. As it has been observed during the thesis, the positioning of both is of extreme importance for a high accuracy of the system.

Moreover, it is essential to expand the examinations on temperature dependence and stray field dependence for every sensor-magnetic field combination. Although this research has provided valuable insights into these dependencies, additional investigations are necessary. These may include exploring the linear relationship between temperature and system sensitivity or the quadratic correlation between stray field susceptibility and the distance between bus-bars. A more comprehensive evaluation is required to validate these preliminary findings.

Finally, the sensor should be installed to the output of a series inverter and make it function at full power. This approach would facilitate a comprehensive assessment of the sensor’s dynamic characteristics, potentially unraveling new insights into its operational behavior.
References


REFERENCES


