

Wireless Power Transfer for Structural Health Monitoring of Composite Structures

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Abstract

We, here, lay down the base ground and framework for the design of the wireless power transfer unit of a sensor node for structural health monitoring of composite structures. Composite structures like aircraft, wind turbine blades or car are prone to hidden damages. Carbon fiber composite risk of failure also increases with lifetime due the environmental impacts. These factors are leading the industry to look toward condition-based maintenance techniques. This requires the development of structural health monitoring techniques assessing parameters like strain, temperature or humidity within the structure at regular time intervals. In this context, the need for insertion techniques and powering of electronics within the composite has arisen.

This work focuses on the wireless power transfer through resonant magnetic coupling for a receiver embedded into carbon fiber reinforced composites.

A complete analytic model of coupled coils separated by a conductive sheet was first developed. The equations obtained were used to numerically evaluating the impact of the composite plies on the self-inductance, quality factors, mutual inductance of the system and so on the power transfer efficiency.

Following that, a design flow using the 3D simulation software CST was put in place. The fiber composite was approximated as an homogeneous media of effective conductivity of 7500 S/m, approximation which was experimentally verified. This was done by implanting three PCB test coils in composite material. The results of measurements on the coils were then compared to simulation results obtained with CST in which similar 3D coils were inserted into the equivalent homogeneous material. To do so, a custom composite fabrication technique was put in place to embed pancake coils into composite samples. It was concluded that the combination of CST simulations and of the homogeneous material approximation could predict the coils self-inductances as accurately in CFRP as in air.

Finally, this process was applied to small pancake coils and following aircraft application constraints. They were inserted 1- and 2-ply deep into industrial pre-impregnated composite. With those specimens, it is proved that power can be transferred within the composite achieving 0.6% and 0.45% power transfer efficiencies for the 1- and 2-ply deep samples respectively.

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Abbreviations

CFRP, composite	Carbon Fiber-Reinforced Polymer
SHM	Structural Health Monitoring
WSN	Wireless Sensor Node
WPT	Wireless Power Transfer
σ	Electrical conductivity
Ply	One layer of carbon fiber reinforced composite (considered 300 μm thick)
Prepeg	Pre-impregnated carbon fiber composite
CST	The software <i>Computer Simulation Technology</i>
PTE	Power Transfer Efficiency

Introduction

Until now, the lack of knowledge in barely visible impact, fatigue and thermic history of carbon fiber reinforced composite structure have lead to two necessary economical non-senses in the design and maintenance of those structures. First, the structures thicknesses are typically oversized in order to reach high security margins which increases the weight and fabrication costs. Second, maintenance programs are currently preventive, making most of the servicing practically unnecessary. Beside the obvious economical impact, both of these factors also have consequences on performances and the ecological footprint. An alternative and replacement to these engineering techniques is the insertion of small electronic systems called sensor nodes within the structure to permanently assess its inner environment and integrity. There actually already exist many excellent designs of wireless sensor node like [1] or [2]. In the case of interest, the nodes are meant to be embedded inside a composite structure which brings many additional requirements that the existing designs do not fulfill. As it will be detailed later, the allowable size of the system is, for example, very limited due to manufacturing visible default norms. Most of all, due to this size limitation, the powering of the system has to be entirely re-thought. The limitations hardly allow autonomous energy harvesting and dictate a wireless power transfer alternative. This alternative itself is also very constrained due to the electrical conductivity of the carbon fiber. In the case of magnetic resonant coupling, the fibers net acts like a conductive sheet which results in a proximity and a screening effect on the power transfer coils and inductive link. This strongly impacts the power transfer efficiency according to pattern missing in the literature and which are not easily obtainable since the composite conductivity varies strongly with frequency and its fabrication process itself. Beside the original power transfer reduction, the proximity effect can impact the design of the coils matching network which leads, again, to additional requirements (see Chapter 2). Considering all of this, it appeared that the keystone in the design of a wireless sensor node to embed in a carbon fiber composite structure is the power unit. Taking the magnetic resonant coupling approach, it is this aspect that has been studied and which is reviewed in this document.

The original goal was to assess and demonstrate the feasibility of the system. After reviewing the literature and the state-of-the-art in this field, it appeared to be necessary to refocus on a study of the impact of the composite on the wireless power transmission. This happened following the decision to take the resonant magnetic coupling approach. The literature on WPT through or in conductive media simply does not exist. Considering the huge industrial potential, it seemed more than worth it to start researching it in order to later on be able to lay down a design framework for future work.

Before doing so and in order to understand the full picture, Chapter 1 goes through a short study of the carbon fiber composites themselves as well as of the main difficulties linked to such electronic integration. It also includes an overview of the overall context and the industrial application goals in more details.

Through Chapter 2, a complete analytic model is built in order to intuitively understand the relevant phenomena. This model allows to evaluate the inductive link properties when the primary and secondary coils are separated by a conductive sheet and was experimentally validated through simple experiments. Re-using the model results, a standard circuit analysis permits, then, to assess the wireless power transfer from primary to secondary. This was actually done for a simple case study and all equations were numerically evaluated using realistic coil size and composite material properties from the literature.

In chapter 3, a full 3D-simulation is done and a potential flow on how to design a more complete power transmission system is detailed. This includes the necessary steps to approximate carbon fiber composites as an homogeneous material in the software CST. The technique is experimental verified as well as its ability to make prediction on the impact of the composite on the self-inductance of an embedded receiving coil. This was only made possible by the implementation of a custom composite fabrication in order to include test coils between the plies. The process is also described there.

Finally, the last chapter aims to show that it is possible to transmit sufficient power to a node embedded into a composite structure in order to power a sensor unit and perform measurements. This is done by embedding and testing of-the-shelf small back-ferrite coils having realistic dimensions for such applications. Those coils were coupled with larger readers circuit and their mutual inductance measured. AC and DC power transfer measurements were conducted and the power transfer efficiency between a reader and a receiving coil embedded one and two composite plies deep into a structure is assessed there.

To conclude, a list a experiments is proposed for future continuation of this work.

Chapter 1

Overall context and literature review

This first chapter analyses the context in which this work takes place. The original goal was to study the design of a wireless sensor node for structural health monitoring of composite structure. As its name indicates, such a node senses its environment (strain, temperature, humidity..) and wirelessly transmits the data to a receiving system. In the framework of this work, it is also meant to be embedded within the composite structures. Before going further, this raised many questions and various subjects like the nature of the composite and its interactions with the system itself had to first be looked at. This includes other factors like the manufacturing requirements of composites and, most of all, the node powering options.

This is what this first chapter is about. It is meant to offer a general synthesis on the subject of interest and of the constraints raised by the nature of the composites themselves. It also shows the wide industrial application potential and the current holes in the related literature.

First, carbon fibers reinforced polymers are overviewed. This includes composition, fabrication processes and main industrial applications. Then, a section goes through the idea of structural health monitoring (SHM) and how and why it applies to composite structures. After that, the main research focus in SHM and technologies are described. The constraints due to the environment and the fabrication process in the case of a composite aircraft are laid down and the design impact on a wireless sensor node (WSN) detailed. Finally, a literature review on the impact of the insertion of small elements on the structural integrity of a carbon fiber reinforced composite (CFRP).

1.1 Carbon fiber reinforced composite materials

1.1.1 Overview

A composite is an artificially constructed material system consisting of multiple components designed to attain specific properties superior to those of the individual constituents [3].

Mainly, this means composite materials consist of a combination of multiple materials mixed together. The material resulting of the mix then offers better properties than its constitutive materials alone. In the case of fibers composite, the material is made of a succession of fibers elements embedded into a resin or epoxy matrix. The fibers can be weaved together according to various patterns and generally the material is laminated with fibers oriented in alternating directions to give better strength and stiffness. A schematic of such a laminated composite can be found in Figure 1.1.

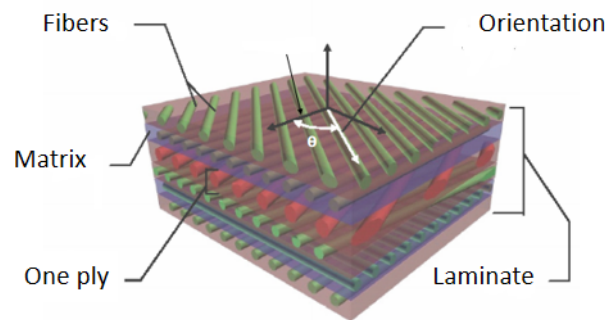


Figure 1.1: Fiber composite structure, reproduced from [4]

More accurately, carbon fibers weaves are made of fibers of about 5 to 12 μm diameter and are mostly composed of carbon atoms. The atoms are bonded together forming a crystal more or less aligned to the fiber axis. The crystal alignment gives the fibers their strength to volume ratio. Thousands of individual carbon fibers are bundled together in what is called a tow. Tows can then be either aligned to each other, forming a unidirectional ply, or woven together into a fabric (Figure 1.2). The tows are denominated 3K, 6K 12K,.. which represents the number of filaments per tow, 3K for 3000, 6K for 6000,.. There exist various different weaves offering different properties. Figure 1.3 shows some of them. The matrix has for role to support the fibers and to bond them together. There are three main matrix types: epoxy, phenolic and bismaleimide each having their own advantages and specificities.

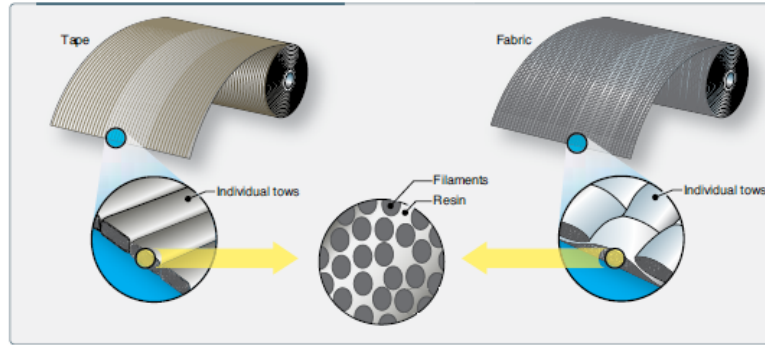


Figure 1.2: Tow structure for unidirectional (left) and woven (right) fiber plies, reproduced from [5]

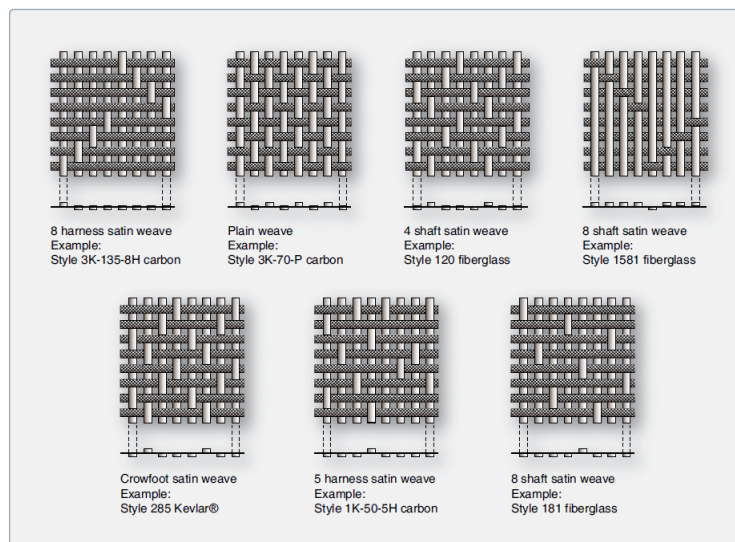


Figure 1.3: Different ply weaves, reproduced from [5]

As previously said, the fibers may be of various types. Today, the overall composites market are still largely dominated (85%) by glass fibers ([6]) but carbon fibers composites have been experiencing strong growth. Carbon fiber composites main advantages are their high tensile strength, high stiffness, relatively low weight, corrosion resistance, high chemical resistance and low thermal expansion. They have one main disadvantage: the cost. Unless the weight of the structure is exceptionally important, the use of carbon fiber is usually not worth the extra cost.

1.1.2 Applications

Aircraft structures are the primary field of application and current use of carbon fiber composites. Carbon fibers first appeared in the aviation sector in the early 80's. Today composite structures reinforce almost all aircraft primary parts, including wings and fuselage. To make an exhaustive list, within the aerospace industry, carbon fiber based composites can be found in:

- Flight control surfaces
- Landing gear doors
- Leading and trailing edge panels on the wing and stabilizer
- Interior components
- Floor beams and floor boards
- Vertical and horizontal stabilizer primary structures on large aircraft
- Primary wing and fuselage structures on new generation large aircraft
- Turbine engine fan blades
- Propellers

Since the 90's, carbon fibers have also gradually conquered the areas of various industries like automobiles, sports or wind turbines,... Today, the scope of carbon fiber on the market is incredibly vast and they appear in many diverse mass or niche applications. The main reasons of this expansion are on one hand that prices dropped down while new suppliers appeared. On the other hand, the high demand for lighter and so less power hungry products have lead manufacturers to look for alternative to metallic material and aluminum more specifically. On top of that, according to Airbus, the use of carbon fibers has allowed to improve performances and production cost while reducing life cycle cost and environmental impact. Currently, Airbus A380 rear fuselage, center wing box, wing ribs, ... are made of CFRP for a total of about 25% of the structure weight. This was even extended to 53% of structure weight with the A350XWB outer wings and full fuselage [7].

Looking at a different field, in 2050, there will be no less than 2.5 billion cars in the

world [6]. This represents an average production of about 100 000 cars per years until then. Europe has strong CO_2 emissions goals and desires to reach 95g/km by 2020 [6]. It is a known fact that car weight reduction will decrease fuel consumption with numbers going up to a 7% decrease in consumption for a 10% weight reduction. Considering that typical industrial carbon fiber composite weight around $1.6g/cm^3$ for $2.7g/cm^3$ for Aluminum, fabricating many parts like the main frame or doors with the latter could lead to major weight reduction. this was illustrated by car manufacturer BMW which launched in 2013 the i3 all-electric model mainly made of resin transfer molded carbon fiber composite structures.

Carbon fibers composite is also widely used in wind turbine manufacturing. Currently, carbon fiber is used primarily in the spar or structural elements of wind blades longer than 45m [6]. With a wind turbine sector under constant growth, the wind energy industry represents a significant share of the carbon fiber composite market.

1.1.3 Fabrication processes

There exist many different composites fabrication methods. Some were precisely developed to meet design or manufacturing challenges composite related, other were borrowed from processes based on other materials. They mostly fall into two categories, the open-molding processes and the closed molding processes [8] [5].

In an open-molding process, the fibers weave or mat is laid down on an open mold and the resin is applied on top. The overall is then cured in an oven or an autoclave at temperature going up to $180^\circ C$. The autoclave curing is used mainly in the aerospace industry where high production rate and cost is not an important consideration [8]. An autoclave is a pressure vessel where the temperature, pressure and vacuum can be accurately specified in order to follow any desired curing cycle. Usually, the mold is first put into a vacuum bag out of which the air has been pumped out, the resin is in that case directly injected in the bag.

A close molding process, as its name indicate, uses a closed mold. A good example is the Resin transfer molding or RTM. In this technique, resin is transferred over the already placed reinforcement. A mat reinforcement form (the fibers) is placed on the surface of lower half of the mold and the mold is properly closed and clamped. The resin is injected inside the mold under pressure and pushes the air out through vents, until the mold is filled. This process is well-suited for mass production of 100 to 10,000 units/year of high-

quality composite fiberglass or fiber-reinforced plastic parts [8] [5]. It is recommended for products that require high strength-to-weight requirements. The uniformity of resin flow can be enhanced by vacuum application. After curing, the mold is opened and composite product is taken out.

There exist many others techniques and, beside the economical aspect, each has its own advantages which will not be studied further here. For more complete information, the previously mentioned sources are a good reference as well as [9].

Prepeg technologies Pre-impregnated or prepeg fibers are fiber woven which are already impregnated with resin. This means that resin matrix material does not necessarily need to be added. The prepeg's offer an alternative to the classic fabrication process and can be cured in similar manners as explained here above. Currently, they are the ultimate carbon fiber composite materials and are mostly used in the aerospace industry or under severe mechanical and geometrical constraints. A good overview of the prepegs would be the document from manufacturer Hexcel [10], they will also be discussed in more details in Chapter 3.

1.1.4 Types of mechanical failure and hidden damages

On one hand, unlike metallic structures, composites structures are prone to hidden damages from low velocity impacts. These damages are called *barely visible impact damages* or BVID. In the case of aircrafts, they might come from events like bird strikes, hand tool drops during manufacturing or servicing or a simple runaway stones. They will grow over time until they may become critical for the structural integrity of the structure.

Good examples of damages are [11] [12]

- **Matrix damages** can take various forms counting cracks parallel to the fibers or debonding between fibers and matrix. Matrix damages are usually parallel to the fibers direction and can result from transverse (to the fiber direction) low-velocity impact.
- **Delamination** are cracks going through the resin-rich zone between the plies. It results from a bending stiffness mismatch between plies of different fibers orientation.
- **Fiber failure** is self explanatory. It occurs much later in the fracture process than delamination and matrix cracking.

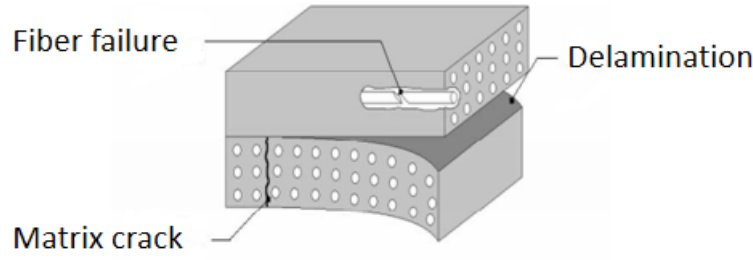


Figure 1.4: Illustration of different hidden damages, reproduced from [15]

On the other hand, it has been shown that temperature and humidity can foster these processes. In [13], epoxy carbon fiber-reinforced composite samples were aged under 177 °C for 5000 hours. It was shown that compression strenght can decline up to 30%. This is the effects of what is called *thermo-oxidative aging* on the thermal, fracture and compressive behavior of carbon composites. In [14], samples were conditioned in distilled water. Tests were conducted and lead to a relatively high saturation moisture content and to a "*significant reduction in interfacial bond strength after immersion in distilled water*". From these studies, we can conclude that the temperature and humidity have serious impacts on the mechanical properties of carbon fiber composite laminates. Historical knowledge of a given structure environment could so help to assess its structural integrity.

1.2 Carbon fiber reinforced composite aircraft damage detection techniques and structural health monitoring

Nowadays the main focus of damage detection technique of CFRP mostly applies to the aeronautic industry due to multiples factors. First, for obvious reasons safety checks and standards are much higher for airplanes than for cars or wind turbines, the aerospace industry would so really make use of SHM systems. Second, airplanes currently still represent the main part of of carbon fiber composite industry and market. Considering that we, here, focus on the composite aircraft context, current technologies and field of research.

First, the curves in Figure 1.5 represents the so called *safe life* of an aircraft versus

stress (the safe life is the lifetime of a given structure when under given stresses and for a 0% fracture probability). On the left, the continuous line represents the lifetime of the same structure as the dotted line but when crack propagation detection is taken into account in the design (meaning that the designer knows that crack detection will be applied and take that into account when designing). This qualitative illustration is very representative of the positive impact of structural health monitoring on structure lifetime which can double the operational life. [16]

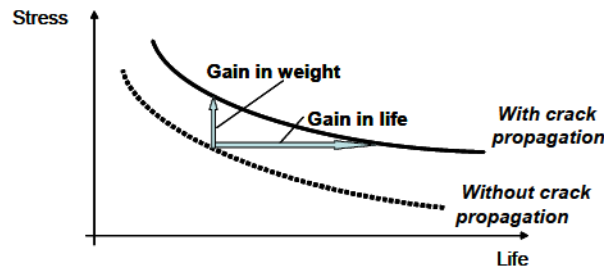


Figure 1.5: Influence of including damage propagation and detection in the design, reproduced from [16]

Second, dramatic incidents like the Aloha Airlines flight 231 of April 1988 have uncovered a phenomenon called Multi-Site Damage (MSD). This phenomenon, which can seem obvious to some extent is that the likelihood of cracking is increasing with an increasing age of an aircraft structure. This partially comes from the previously mentioned temperature and humidity cycles impacts the composite mechanical properties. It lead to regulation requiring enhanced aircraft inspection efforts for older plane, increasing the maintenance cost [16].

Many damage detection technics currently exist. They can mainly be divided into two categories [17] :

Active approaches

Active techniques measure the resulting responses of the monitored structure to actuation or stimulation. These methods require to immobilize the aircraft which is time and money consuming. Examples count for instance 3D laser vibrometer (see [18]) used to scan Lamb wave responses in order to reveal delamination in the composite as they go through the structures or ultrasonic inspection.

Passive approaches

Transducers are used to monitor perturbations directly caused by damages. The sensing and possibly actuation devices are attached on or integrated within the structure. The structure is "monitored" constantly or at defined time interval. This process is called *structural health monitoring*.

1.2.1 Structural health monitoring

SHM is the integration of sensing and possibly also actuation devices to allow the loading and damaging conditions of a structure to be recorded, analyzed, localized and predicted in a way that non-destructive testing becomes an integral part of the structure. [16]

Currently, the maintenance philosophies of most things are time based. Planes structural integrity are verified at regular time intervals or after a defined number of flights. Structural health monitoring or SHM aims to transform the current methodologies into condition-based maintenance philosophies. [18] SHM would so allow to first increase reliability and safety considering the structure could be monitored at all time. Second, runtime of aircrafts and lifetimes would also be condition-based and so would be significantly enhanced since the defects and cracks might be detected at the earliest possible state. Third, the structure operating and maintenance costs could be reduced to a minimum. A good general background on SHM as well as a general framework to select the right design in terms of environment, constraints, damages and economic aspect can be found in [18]. To summarize this in one sentence, structural health monitoring of composite aircraft might increase safety and lifetime while reducing overall cost.

In a different, but not-unrelated, sphere, many airframes and jet engines manufacturers would like to completely reshape their business model and go from a standard manufacturer-buyer model to a leasing-based model. They would so take on the responsibilities for the maintenance of the equipment. SHM has the potential to reduce the maintenance windows to the necessary only and, hence, to keep the equipment out in the field as much as possible to generate revenues for the owner [19] [16].

1.2.2 Main current techniques under study for composite structural health monitoring

Optical methods

Optical methods use optical fibers as sensor elements to detect variations in temperature,

strain, pressure or vibration. These methods have the great advantage of being immune to any electromagnetic interference considering they work on an optical basis. The simplest type are the F-P sensor. This method relies on variation of the refractive index of the fiber due to the environment (strain, temperature,...). The transmitted light intensity can then be measured and data retrieved. More up to date, the optical fibre Bragg grating is widely used. They are made of optical fiber of 100 μm diameter or less and can be either integrated or attached to the structure (for composite or metal respectively). Figure 1.6 shows a schematic of such a fiber. Series of lines are patterned in their core leading to regions of lower or higher refractive index. The interference between the light reflected by each line pattern results in part of the input spectrum being reflected back. The section of the spectrum being reflected directly depends on the spacing of the line patterned. A change in the reflected spectrum can then be directly correlated with a given strain [19] (see the strain induced shift in Figure 1.6). Fiber Bragg grating have been extensively tested and show robustness with regard to *in service application* [18] [20]. It is also worth mentioning the effort described in [21] [22] [23] [24] where an RFID based interfaced is designed to be embedded with the fibers to transmit the information (see Figure 1.7). Even though meant for structural health monitoring, such a system does not fit some standard requirements of many composite structure. It features as size of 3 to 4 mm thick and a 2.5-square-centimeter space. As it is explained later, maximum inclusion in the carbon composite structure for aircraft are much smaller. In this approach, they perform the energy transfer at 125 kHz and the data transmission at 13.56 MHz. Two coils seem to be necessary which lead to a larger system (this is just implied, it is not clearly stated in the publications). As it will be described later, in our approach we chose to perform both data and power transfer with a single coil.

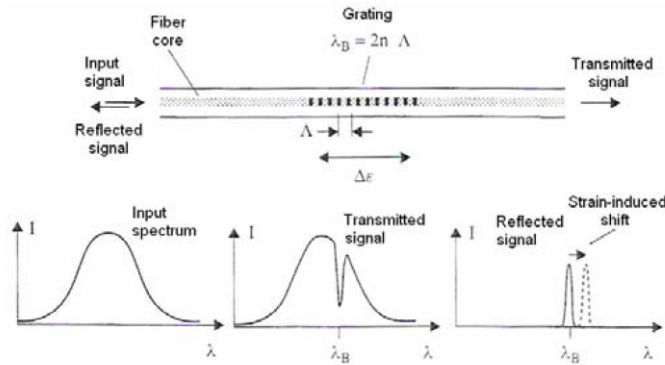


Figure 1.6: Sensing principle of Fiber Bragg grating, reproduced from [25]

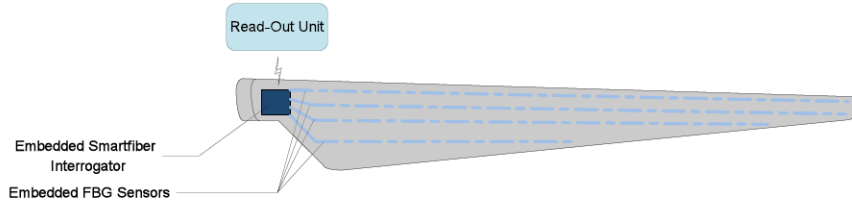


Figure 1.7: Sample application for SmartFiber system, reproduced from [21]

Piezoelectric sensors

Another state of the art damage detection techniques are piezoelectric transducers. They can be displaced as a sensing array and so are very suitable for SHM of laminated material in which they can be inserted between two composite sheets. Many prototypes have been designed out of which the Stanford Multi-Actuator Receiver Transduction layer (SMART layer) is used as a reference in the field (see Figure 1.8). Other worth mentioning efforts like the PWAS (Piezoelectric wafer active sensors) technology in University of South Carolina are detailed in [26]. Figure 1.8 also shows a picture of the SMART layer developed by NASA. It is very visible that the connections are meant to be done outside the composite structure. This means that the power transfer and consumption are of no concern.

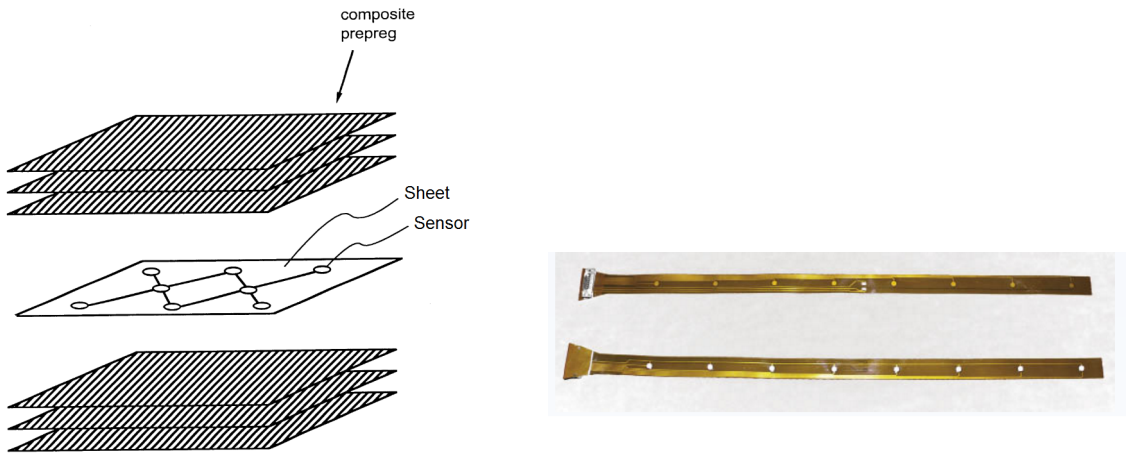


Figure 1.8: Diagnostic layer and method for detecting structural integrity of composite and metallic materials, reproduced from [27] (left). Picture of Acellent's SMART Layers with 8 sensors, reproduced from [28]

MEMS based

MEMS damage detection systems can integrate mechanical parts, sensors, actuators and

electronic parts on the same die. They virtually allow to make any kind of measurement but potentially require additional circuitry, data transmission and powering. They have the advantages of a small size, small weight, high reliability and low cost. A good example is the SOI CMOS smart strain sensor from Université Catholique de Louvain. Figure 1.9 shows the working principle of the sensor which uses the piezoresistivity of PMOS channels to obtain the strain of the die. This particular sensor has the advantage of being extremely low power ($10\mu W$ power consumption) and so fits well to wireless applications.

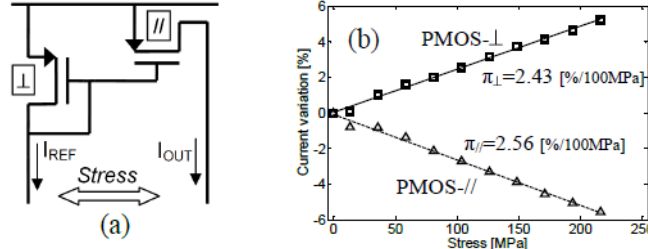


Figure 1.9: Circuit and working principle of the SOI CMOS strain sensor, reproduced from [29]

Other detection techniques count acoustic emission, ultrasonic inspection, laser vibrometry, comparative vacuum monitoring, electromagnetic foils and many more. Review of some of them can be found in [20] [16].

All of these damage detection techniques raises some challenges. First, implementing such systems requires a sophisticated monitoring hardware to be deployed and an advanced data analysis procedure in order to retrieve the measured data. Second, and in the case of an embedded system, before deploying sensing elements through the structure, the impact on the structural integrity must be assessed. Third, it must also be demonstrated that the system itself will not be damaged when deployed in the field.

In the techniques mentioned above, the sensing element might require additional circuitry for powering, actuation, data transmission or signal processing. Such a system is called a *sensor node* and its architecture and challenges in the context of structural health monitoring of composite structure is over-viewed here under.

1.3 Wireless sensor node

1.3.1 General architecture

Figure 1.10 shows a basic overview of the different blocks present in most wireless sensor node.

A wireless sensor node is an autonomous unit, capable of sensing its environment and to communicate with or being part of a larger network. The node is generally built around a controlling unit taking care of managing the sensor unit(s) and the memory. A communication system is used to transmit the information sensed. All blocks are powered through a power supply unit which can be of many type.

For a good survey and implication of choices in wireless sensor node design, [30] should be consulted.

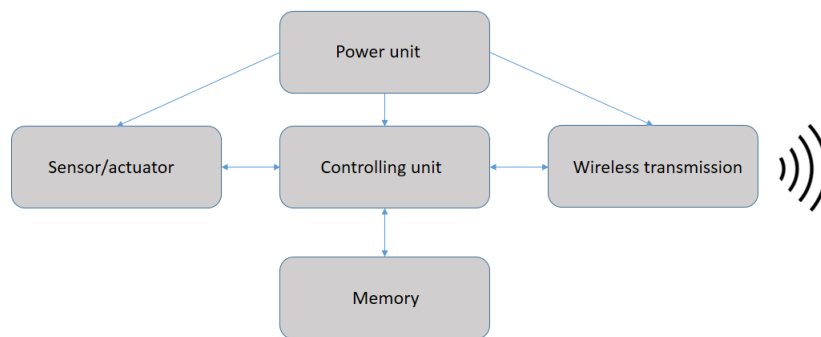


Figure 1.10: Basic architecture of a typical sensor node

There exist many good examples of WSN for structural health monitoring based on this architecture. The information in [31] gives the overall framework for a WSN network for SHM of aircraft wings. They design a miniature, high-precision, and shock-proof wireless sensor node for multi-channel strain gauge signal conditioning and monitoring. The network approach with the node being part of a cluster is also analysed. A second approach using components-of-the-shelf can be found in [32] and in [33] PZT patches are embedded within glass fiber composite and externally wired to the circuitry.. In [34], they also develop a wireless ultrasonic structural health monitoring (SHM) system for aircraft wing inspection using embedded piezoelectric transducers. All of these studies focus on designing a node capable of sensing its environment and to transmit the information to a parent device. At the opposite, none of them takes any manufacturing or environment requirement into account for the case of embedded nodes in CFRP. This means that on one hand they have dimensions way too large to be embedded in the composite. On the

other hand, the miniaturization of the power unit is so of no concern and the approach is incompatible with ours. Also, they all transmit data through a typical Radio transceiver (2.4 GHz) and usually with ZigBee protocol while it is most likely impossible to transmit anything through a conductive sheet at such high frequencies because of the skin effect (carbon fiber are conductive, this will be studied and detailed further in the next Chapters).

1.3.2 Constraints

Contacts were taken with aerospace company SABCA which manufactures various Airbus aircrafts parts in order to obtain first hand approximation for test nodes. According to their standards, the following requirements would apply to embedding a wireless sensor node into an actual composite aircraft structure part:

- The node has to be able to endure at least 3 hours at 180 °C
- Acceptable defects sizes on structure are 10mm X 10mm in general and 6mm X 6mm in critical zones
- Acceptable defect thickness is less then 8% of the part size with a maximum of one ply thickness ($300\mu\text{m}$)

Minimum requirements for embedded temperature and pressure sensors are a precision of 2 degrees and 0.1 bar respectively.

1.3.3 Power harvesting vs. Wireless power transfer

The starting point of the feasibility assessment of such a system for embedding into carbon fiber composite is the powering unit.

Considering sensors like the CMOS sensor from UCL [29], it was decided to target a received DC power of around $150\mu\text{W}$ for 1 to 1.5 Volts and 100 to $150\mu\text{A}$. Similarly, it was decided to limit the transmitter input power to 1 W. Figure 1.11 shows a basic circuit model of the load. If we count on a 75% power efficiency for the AC/DC conversion, the desired received voltage and current are 1 V_{rms} and $200\mu\text{A}$. From these values, it is possible to calculate the equivalent load

$$Z_{AC} = \frac{1}{200 \cdot 10^{-6}} = 5k\Omega$$

$$Z_{DC} = \frac{1.5}{150 \cdot 10^{-6}} = 10k\Omega$$

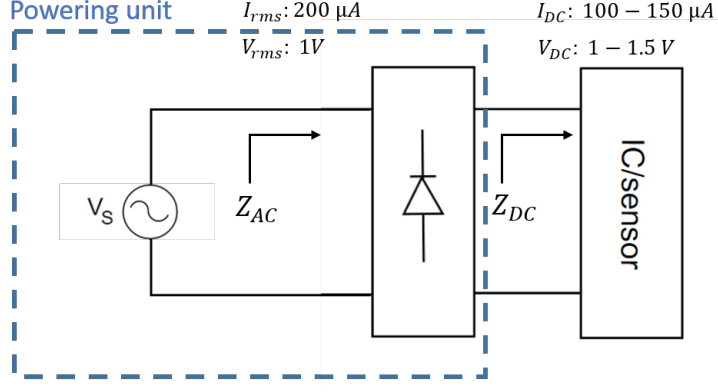


Figure 1.11: Basic circuit model of the node.

There are two possible approaches as a wireless sensor node. Either the power is wirelessly transferred to the node from a controllable source (through inductive coupling, magnetic resonant, acoustic waves, light , ..) or the node is able to harvest the energy needed to properly function from a non-controllable source in its environment (heat, vibration, EM,...). In both cases, two different architectures are possible: *Use* and *Store-Use*. As their name indicates it, the difference is the second option features an energy storing device such as a battery. The interest of doing so is the ability on one hand to power the node when desired when the battery is charged. On the other hand if the energy is harvested, this architecture allows the node to function in different environments where the energy source might not be as strong. In that case, when enough charges have been collected and stored, the system switches to an ON state and performs one cycle Sense $\rightarrow$ Process $\rightarrow$ Transmit operation. The measurement period so depends on the amount of power available.

[35] gives a good overview and comparison of the harvesting options precisely for aircraft structural health monitoring. Further information on thermoelectric generator harvesting will be found in [36] [37] and on vibration energy harvesting in [38] (again precisely for SHM of aircraft). After reviewing them, harvesting seemed here to be hardly applicable for embedded node. This is due, on one hand, to the large size of the harvesting unit. On the second hand, harvesting more often requires a *Store-Use* architecture which also increases the size. Table 1.1 shows the power density production of a TEG and a vibration generator. Both these devices were precisely assessed for the powering of a wireless sensor node for structural health monitoring of aircraft in [35]. Their densities are expressed in W/kg and it is hard to find data related to size. If here we make a best scenario approximation that their weight-to-volume ratio is really high and of $10000 kg/m^3$

Table 1.1: Power densities of different energy harvester described in [35].

Generator	Power density	Approximated power production
• Vibration generator VEH 360 (60Hz, $a=0.24 \text{ m/s}^2$)	42mW/kg	12.6 μ W
• Thermoelectric generator TEG-9.1-9.9-0.8/200 ($\Delta T = 10\text{K}$)	34mW/kg	10.2 μ W

(slightly more than iron), it is possible to get an approximate power production value for a volume of 1mmx1mmx0.3mm (corresponding to the maximum inclusion size, see the constraints above). The values obtained can be found in the third colon of the table and show that those powering methods will not reach the 200 μ W desired while meeting the size constraints.

Also, batteries could not function properly after undergoing the high temperature curing cycle of the composite. Following this, it was also concluded that a *Store-Use* architecture was not feasible either.

Considering this, it was decided to focus on a RFID-tag-like approach and to power the node through magnetic resonant coupling. This allows to reduce the size to a minimum, considering that both the data and the power can be transmitted through the same channel.

1.4 Impact of small inclusions on the structural integrity of the host laminate

Another important point that needs to be discussed is the impact of small inclusions of the mechanical properties and structural integrity of the composite structure. This section overviews the literature on the subject without going further. From it, it can be concluded that it is possible to embed a node within a carbon composite structure without affecting noticeably the integrity or the fatigue lifetime. In all cases this is if following some requirements which will strongly depend on the physical structure, the fabrication process and the integration process itself.

Over the last 30 years, many thorough studies have been conducted in order to determine the impact of embedded elements on the structural integrity of composite laminates. In the early 90's, Warkentin and Crawley [39] conducted a feasibility study on the inte-

gration of silicon chips into graphite/epoxy composite laminates for aeronautics. They developed an embedding procedure which addresses electrical, mechanical, and chemical compatibility issues. They also investigated the effects of the embedment under static mechanical loading. Their work concluded that *“physically embedding electronic components for the control of intelligent structures is both advantageous and feasible”* (from a mechanical point of view).

- In 1999 Paget and Levin [40] studied the strength reduction due to the embedment of a piezoelectric ceramic transducer in a pre-impregnated composite. They put the specimens under tensile and compressive static loading and found that the embedded piezoelectric ceramic element along with its subsequent interconnectors did not reduce the strength of the host composite material.

For a complete finite element method simulation and analysis result over a three-dimensional stress and displacement fields around a rectangular embedded element in a graphite/epoxy laminate Chow and Graves (1992, [41]) should be consulted. Their results showed the effect of various layed up orientation on the stresses near the corner of the rectangular inclusion in an 8-ply symmetric laminates.

For additional model in the case of unidirectional graphite/epoxy composite laminate Singh and Vizzini [42] should be consulted.

More recently, Varadan and Varadan [43] studied the requirements applying to MEMS to allow their integration into fibers reinforced composites. In this context they addressed some of the relevant issues related to the incorporation of non-structural elements into fiber reinforced composites. In the case of the previously mentioned SMART Layer in which Lin and Chang [27] developed a manufacturing method for the integration of a thin flexible layer with a network of sensors between plies of the composite structure. Various tests were conducted on the samples and showed that the SMART layer did not have any significant impact on the integrity of the host laminate. Considering that here the sheet cover the entire surface of the composite and acts more like an additional ply of different structure, the results do not really apply to our case. This kind of approach might at the opposite lead to different structural integrity issue considering the thermal expansion coefficient of the flexible layer is most likely different than the coefficient of the CFRP. Exposure to high temperature can so lead to delamination. No information on this subject were found in the literature.

In [3], the integration of a high density of small sensors into a laminated composite keeping the structural integrity of the host is studied.

In the document, various insertion method and element types were studied and tested. Figure 4.8 shows a schematics illustrating some of the insertion options. For each case,

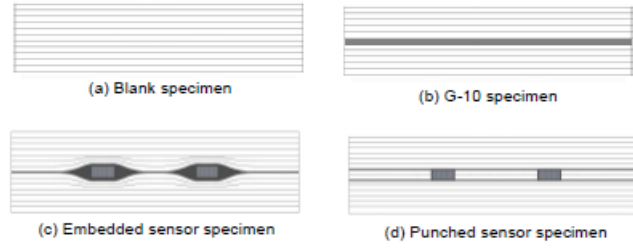


Figure 1.12: Four integration possibilities studied in [3]

an average fatigue life under three-point bending was inferred and compared.

To do so, the samples were *cyclically loaded between 222 N and 2224 N using a sinusoidal input wave function at a frequency of 10 Hz. The cyclic loads correspond to 6% and 60% of the short beam shear strength of the blank specimens, respectively.* Figure 1.13 summarizes the result for the cases illustrated in Figure 1.12 (reproduced from [3]) The average fatigue life mentioned in the figure is the number loading cycles the samples endured before failure. The normalized average fatigue life are the average fatigue life normalized by the average fatigue life of the blank specimen. All of those values were obtained by taking the mean of 12 different measurements for each configuration [[3].

Table 4.15: Summary of fatigue three-point bending (short-beam shear) test results including average fatigue life and normalized fatigue life.

Specimen Configuration	Specimen Diagram	Average Fatigue Life	Normalized Average Fatigue Life
Blank specimen		1.71E+06	1.00
G-10 specimen		0.76E+06	0.45
Embedded sensor specimen		0.53E+06	0.31
Punched sensor specimen		1.44E+06	0.85

Figure 1.13: Some results from [3]

Based upon these fatigue three-points bending (short-beam shear) test results it was concluded by the author that “ *it is evident that the type of embedment and the method in which the sensor is integrated into the host composite material have profound effects on the fatigue life*” [3].

Table 1.2: Summary of the system specifications at this point.

Specifications	
Dimensions	$\leq 10\text{mm} \times 10\text{mm}$ in general $\leq 6\text{mm} \times 6\text{mm}$ in critical zones
Thickness	$\leq 300 \mu\text{m}$
Temperature tolerance	$\geq 180^\circ\text{C}$
Powering	Magnetic coupling
Maximum input power	1 W
Received AC Power	$200 \mu\text{W}$
Equivalent load	10 k Ω

1.5 Focus of this work

Table 1.2 outline the node specifications that have been outlined so far.

The rest of this work mostly focuses on the wireless power transfer aspect and more precisely on magnetic resonant coupling with an embedded receiver.

Through literature review, it has been found that there is a great deal of wireless sensor node designs for SHM, a lot of documents and studies on wireless power transfer as well as on structural health monitoring of composite structure. At the opposite, the information and studies on the powering and integration of electronics within a composite structure or close to conductive material is pretty much nonexistent. In this context, the following chapters aim to show that it is feasible to wirelessly transmit enough power to power a WSN through and within composite plies.

Chapter 2

Analytic model

2.1 Wireless power transfer

Wireless power transfer is a topic that has attracted a lot of interest in research as well as in the industry in the recent years. Looking back in history, power transfer is nothing new. In the early 20th century, Nicola Tesla built the Wardencllyffe tower in Shoreham, Long Island, New York, for transatlantic communication [44] (Figure 2.1). Tesla first intended it to transmit messages to England and to ships at sea but eventually aimed to demonstrate wireless power transfer through magnetic resonance and magnetic field coupling. This particular design was never finished but many later researches were based on Tesla original assumptions.

Another famous experiment taking the electromagnetic radiation approach took place in 1964. William C. Brown demonstrated a powered rectenna element receiver helicopter [45]. The helicopter received the power needed for flight entirely from a ground based microwave beam (Figure 2.1).

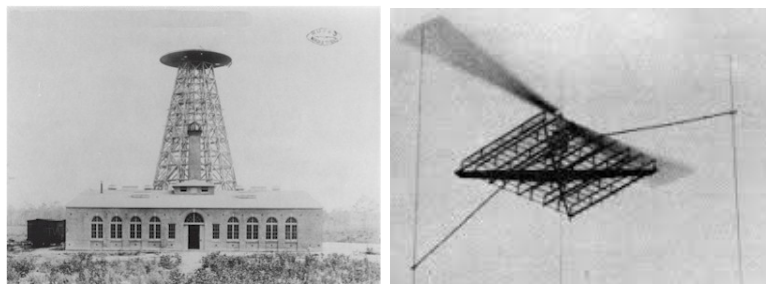


Figure 2.1: WardenCliffe tower (left) [44]. Brown's helicopter (right) [45]

With the proliferation of mobile device like mobile phones or electric toothbrush, they are now back at the heart of the matter.

They are mainly three types of WPT: *magnetic resonant coupling*, *radiative field approach* and the newer *highly-resonant wireless power transfer(HR-WPT)*. The radiative field approach and the HR-WPT will not be discussed here, more information may be found in [45] [46] and in [47] [48] respectively. For a nice introduction and survey of the basic wireless power transfer techniques reference [49] should be consulted.

In the overall, the main research themes about WPT technologies focus on improving the transmission efficiency and distance. At the opposite, there is very little literature about power transmission in conductive environment and close to nothing about transmission through a conductive media. There exist some studies like [50] in which the wireless power transfer through metal at low frequency (100Hz to 1kHz) is studied. Those frequencies require larger coils and will not fulfill the dimension specifications. There is also a general analysis of inductive coupling in the case of implantable medical devices in [51]. As it will be shown later, such analyses do not apply here due to frequency and size restrictions.

A few studies precisely on inductive coupling through CFRP have also been conducted. At Bremen University they studied the impact of the CFRP on the inductive link and concluded that it was possible to transmit enough power for a Piezo-Wafer-Active-Sensor system to be used in the structural health monitoring of fiber composites and embedded into the composite [52] [53]. They used a receiving 8cm diameter receiving coil while information on the sensor dimensions and the power transferred were not available.

As mentioned before, the most notable effort in embedding electronics into carbon composite was done by the Fraunhofer Institute for Integrated Circuits IIS in Nuremberg, Germany [21] [22] [23] [24]. There, they developed a RFID tag to embed in composite materials for storing manufacturing information and to interface a fiber Bragg grating network. In the articles, they target an output power of 1.5W for a maximum transmission distance of 10cm and a maximum system size of 10 x 10 x 3 mm (without the coil which seems to have a diameter of at least 4-5 cm in the pictures in [22]). Although in this case the approach is more similar to ours but for a much higher power, the target size of the system is again much larger.

Due to the lack of model for WPT through/in inductive media in the literature, it was decided to build our own. The next section first goes through a circuit analysis of magnetic resonant wireless power transfer. The basic equations are explained and the reflected load impedance theory is detailed. Following that, it is shown that a single element matching network cannot be sufficient for loads of the order of the kOhm for systems of these dimensions. The circuit model relies on the knowledge about the inductances, quality factors and coupling coefficient of the transmitter and receiver coils. Conductive

material affect those parameters and the power transfer. Because of that, an analytic model of the self and mutual inductance of circular coils in the presence of a conducting sheet is derived. This model is based on Dood and Deeds' original derivation of the vector potential in such a case [54].

It allows to evaluate the circuit parameters and the power transfer in presence of a conductive sheet. To concludes this chapter, this has been done through a case study for which all equations were numerically solved.

2.2 Resonant magnetic coupling

2.2.1 Reflected load circuit analysis

The circuit model starts with the inductive link between the transmitter and receiver coils can be schematized as in Figure 2.2. Assuming that the circuit is excited by a sinusoidal source, the voltages and currents can be expressed in phasor form as $\mathbf{V}_1 e^{j\omega t}$, $\mathbf{V}_2 e^{j\omega t}$, $\mathbf{I}_1 e^{j\omega t}$ and $\mathbf{I}_2 e^{j\omega t}$. As a consequence, the circuit equations are

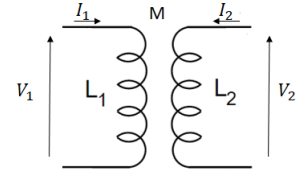


Figure 2.2: Coupled coils circuit.

$$\mathbf{V}_1 = j\omega L_1 \mathbf{I}_1 + j\omega M \mathbf{I}_2$$

$$\mathbf{V}_2 = j\omega L_2 \mathbf{I}_2 + j\omega M \mathbf{I}_1$$

M is the mutual inductance and generally depends on coil geometry, relative position as well as their surrounding environment. M is usually expressed in term of the coupling coefficient k:

$$0 \leq k \leq 1$$

$$M = k \sqrt{L_1 L_2}$$

In general an inductively coupled wireless sensor node power transfer chain can be represented as in Figure 2.3. In order to minimize the primary and secondary impedance magnitudes, the inductances L_1 and L_2 are matched with capacitors C_1 and C_2 according to various architectures. The series resistors R_{s1} , R_{s2} have also been added. After going through the RLC, the picked-up signal is transformed by an AC/DC converter in order to deliver DC power to the IC components and to drive the sensors. The overall chain is schematized in Figure 2.3 for an series-parallel architecture.

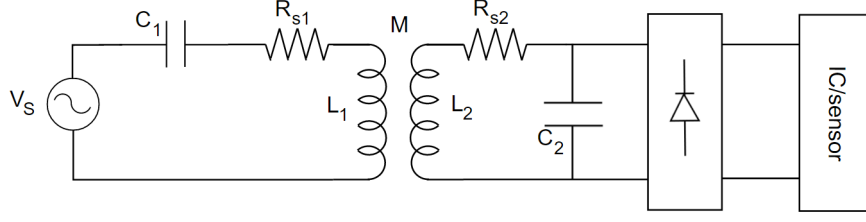


Figure 2.3: Circuit model of the full wireless power transfer.

The input impedance seen at the AC/DC converter can be replaced by an equivalent resistive load R_L (ignoring its parallel capacitance for the sake of a primary analysis considering $C_L \ll C_2$). This leads to the circuit of Figure 2.4.

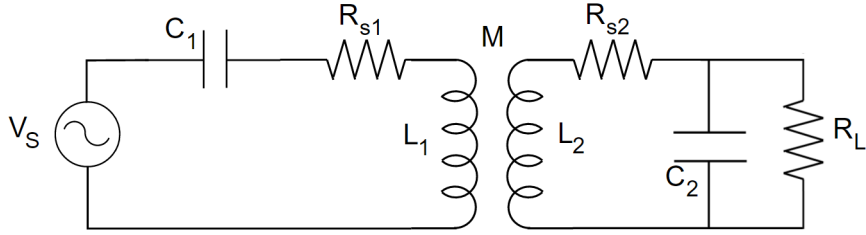


Figure 2.4: Circuit model of the wireless power transfer with an equivalent load.

Applying Kirckoff voltage law to the primary and secondary loops leads to

$$\begin{bmatrix} \mathbf{V}_s \\ 0 \end{bmatrix} = \begin{bmatrix} Z_1 & j\omega M \\ j\omega M & Z_2 \end{bmatrix} \cdot \begin{bmatrix} \mathbf{I}_1 \\ \mathbf{I}_2 \end{bmatrix} \quad (2.1)$$

Z_1 and Z_2 are the equivalent impedances of the RLC series and parallel at the primary and secondary sides respectively. Taking the inverse matrix and multiplying on both sides:

$$\begin{bmatrix} \mathbf{I}_1 \\ \mathbf{I}_2 \end{bmatrix} = \frac{1}{Z_1 Z_2 + \omega^2 M^2} \begin{bmatrix} Z_2 & -j\omega M \\ -j\omega M & Z_1 \end{bmatrix} \cdot \begin{bmatrix} \mathbf{V}_s \\ 0 \end{bmatrix}$$

$$Z_{tot} = \frac{\mathbf{V}_s}{\mathbf{I}_1} = Z_1 + \frac{\omega^2 M^2}{Z_2}$$

This expression represents the total impedance seen by the voltage source. The second term is the reflection of the receiver circuit on the transmitter circuit. The model in Figure 2.5 can then be used to study the power transfer.

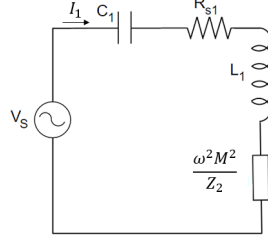


Figure 2.5: Circuit model of the secondary reflection at the primary.

The power dissipated over the term $\omega^2 M^2 / Z_2$ represents the power transmitted to the receiver circuit. Looking at it as a simple voltage divider the input and output powers are

$$P_{in} = \text{Re}\left(\frac{Z_2(\omega) \mathbf{V}_s^2}{Z_1(\omega) Z_2(\omega) + \omega^2 M^2}\right) \quad (2.2)$$

$$P_{load} = I_1^2 R = \left| \frac{\omega M \mathbf{V}_s}{Z_1(\omega) Z_2(\omega) + \omega^2 M^2} \right|^2 \text{Re}(Z_2 - R_{s2}) = \left| \frac{j\omega M \mathbf{I}_1}{Z_2(\omega)} \right|^2 * \text{Re}(Z_2 - R_{s2}) \quad (2.3)$$

The power transfer efficiency is

$$\frac{P_{load}}{P_{in}} = \frac{\text{Re}\left(\frac{Z_2(\omega)}{Z_1(\omega) Z_2(\omega) + \omega^2 M^2}\right)}{\left| \frac{\omega M}{Z_1(\omega) Z_2(\omega) + \omega^2 M^2} \right|^2 R_L} \quad (2.4)$$

If both the transmitter and receiver are matched according to the resonant frequency $\omega_0 = \frac{1}{\sqrt{LC}}$ equation 2.4 reduces to

$$\frac{P_{load}}{P_{in}} = \frac{\omega_0^2 M^2}{R_{s1}(R_{s2} + R_L) + \omega_0^2 M^2} \frac{R_L}{R_L + R_{s2}} \quad (2.5)$$

In this expression, the first term represents the efficiency of power transfer from the transmitter to the receiver while the second term represents the fraction of the received power which reaches the load. This shows that when the circuit resonates, the efficiency will depend only on the frequency and the mutual inductance. If the coils are in air, the mutual inductance is a purely geometrical factor and the efficiency will only vary with the frequency. This expression can be analyzed in term of the quality factor of the coils

$$Q_1 = \frac{\omega L_1}{R_{s1}} \quad Q_2 = \frac{\omega L_2}{R_{s2}}$$

Figure 2.6 shows plots of equation 2.5 in term of the coupling coefficient variations for 5 different coil quality factors and 4 different frequencies. It is very clear that the quality factor and so the parasitic serie resistance have a major impact on the power transfer. First, this shows that at low and high frequency, when the quality factor decreases, the

Table 2.1: Realistic coils properties

Properties	Transmitter	Receiver
R	3 cm	0.5 cm
a	1 mm	300 μm
N	6	8

inductive link efficiency decreases significantly. Second, in the case of our system, it is expected that the carbon fiber composite will seriously affect the quality factor of the coils which will so lead to a similar impact on the efficiency. This is because the simulation was run for a transmitter and a receiver of inductances of $5\mu\text{H}$ $1\mu\text{H}$. These values are fairly realistic for this kind of system and were obtained using realistic geometric parameter in the basic equation 2.6 for a circular loop of radius R and wire radius a (see Table 2.1).

$$L = N^2 \mu R \left(\ln \frac{8R}{a} - 2 \right) \quad (2.6)$$

Higher frequencies allow to achieve much better transfer efficiency for these low values of Q. Achieving high power transfer efficiency with coils of the microHenry order is not possible for load larger than a certain value: the denominator of the first term of equation 2.4 becomes too small because $\omega^2 M^2 \ll R_{s1}(R_{s2} + R_L)$ and the efficiency drops down. From this, it can be concluded that in a case of a load of the order of $10k\Omega$, like our, if the transmitter and receiving coils are of the order of the microHenry, an operating frequency of a minimum of 10MHz will more easily allow to obtain a power transfer efficiency of a few percents.

2.2.2 Maximum power transfer efficiency

For a given system of matched coupled coils, expression 2.5 can be maximised with respect to the load R_L . Differentiating with respect to R_L and equating to zero leads to

$$\frac{\partial \eta}{\partial R_L} = 0 \quad \rightarrow \quad \left(\frac{P_{out}}{P_{in}} \right)_{max} = \frac{\sqrt{1 + k^2 Q_1 Q_2} - 1}{\sqrt{1 + k^2 Q_1 Q_2} + 1}$$

for an optimum load of

$$R_{Lopt} = R_{2s} \sqrt{1 + k^2 Q_1 Q_2} \quad (2.7)$$

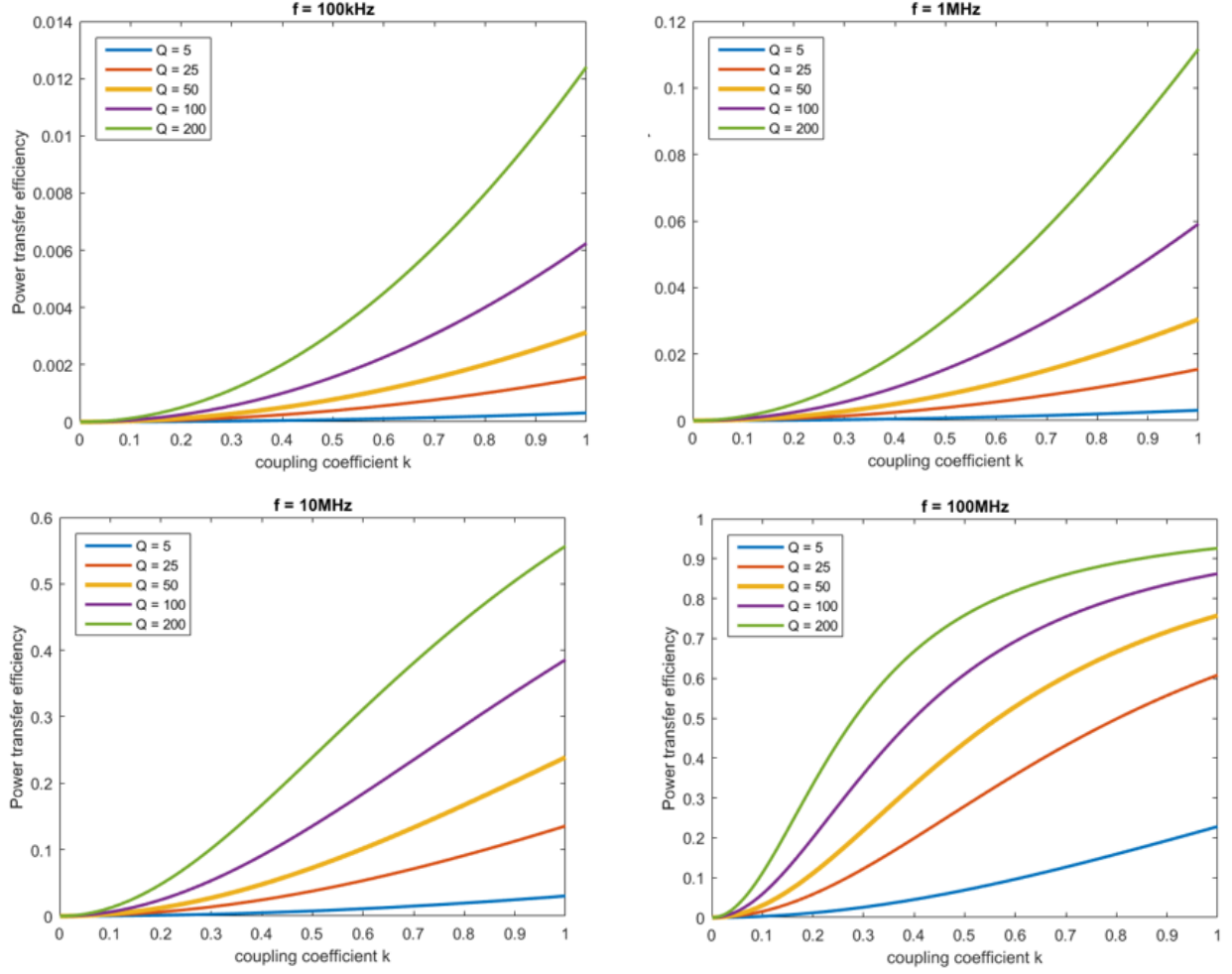


Figure 2.6: Variation of the power transfer efficiency with the coupling factor for different coils Quality factor $Q_1 = Q_2 = Q$ and four different frequencies. The inductances L_1 and L_2 were set to 1 and $5 \mu H$

These values give an upper limit for the power transfer efficiency of a given wireless power transfer system. This efficiency cannot be improved by the addition of any matching network or resonant system. Figure 2.7 shows the evolution of the maximum power transfer efficiency with the factor $k^2 Q_1 Q_2$. We can immediately realize that for our system, $R_L = 10k\Omega$ will be really far from the optimal value.

Following this, we can summarize the power transfer efficiency by saying that it will be maximum if $Im(Z_2) = 0$ (matching) and if equation 2.7 is respected. This can be developed as

$$\begin{cases} C_2 = \frac{1}{\omega^2 L_2} \\ Re(C_2 // R_L) = \frac{R_L}{1 + \omega^2 R_L^2 C^2} = R_{2s} \sqrt{1 + k^2 Q_1 Q_2} \end{cases} \quad (2.8)$$

This set of two conditions cannot be simultaneously respected which means that in

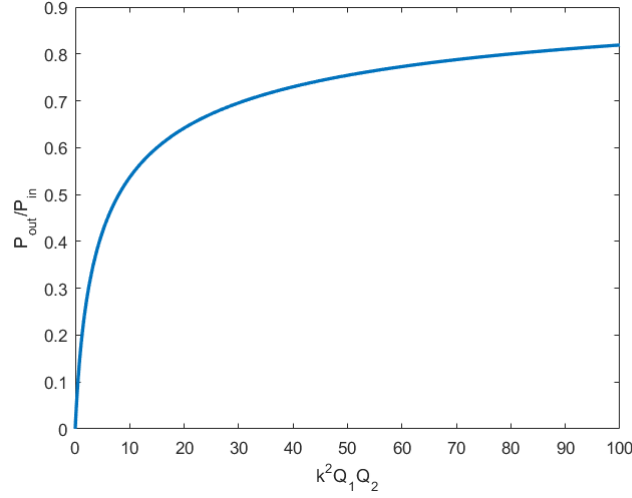


Figure 2.7: Impact of $k^2 Q_1 Q_2$ on the maximum power transfer efficiency.

order to achieve maximum PTE, the secondary matching network should not be implemented with a single element C_2 .

2.2.3 2-Elements matching network

In order to meet the requirement set by equations 2.8, a new degree of freedom is necessary. To do so, a second matching element is added. This element can be a capacitor or an inductor and be placed in series or in parallel (it obviously also works with 3 or more elements, this will not be studied here). Figure 2.8 shows the receiver circuit for a parallel-series capacitor matching network

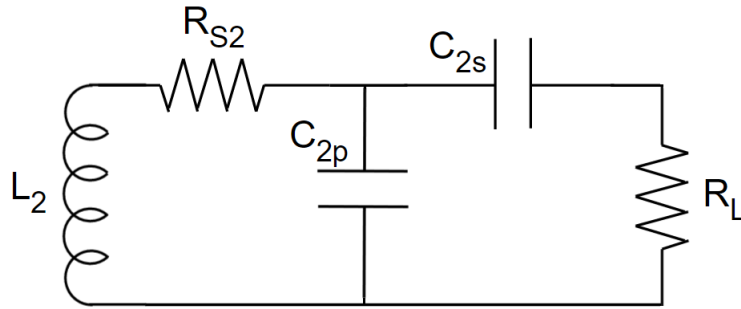


Figure 2.8: Receiver with 2-elements parallel-series capacitive matching network.

The conditions 2.8 becomes

$$\begin{cases} C_{2p} + C_{2s} = \frac{1}{\omega^2 L_2} \\ \frac{C_{2s}^2 R_L}{(C_{2p} + C_{2s})^2} \approx R_{2s} \sqrt{1 + k^2 Q_1 Q_2} \end{cases} \quad (2.9)$$

For which the solutions are¹

$$\begin{cases} C_{2p} = \frac{1}{\omega^2 L_2} \left(1 - \sqrt{\frac{R_{2s} \sqrt{1 + k^2 Q_1 Q_2}}{R_L}} \right) \\ C_{2s} = \frac{1}{\omega^2 L_2} \sqrt{\frac{R_{2s} \sqrt{1 + k^2 Q_1 Q_2}}{R_L}} \end{cases} \quad (2.10)$$

We can see that if $R_L = R_{Loptm}$ then $C_{2p} = 0$ and $C_{2s} = 1/\omega^2 L_2$ which corresponds to a standard series capacitor matching.

2.2.4 Eddy Current and impact of a conductive material on an inductive link.

As previously mentioned, the presence of the carbon composite will have an impact on the inductive link due to the fibers' electrical conductivity. The precise behavior of the composite laminate when inserted in a magnetic field is currently unknown. What is known is that the carbon fibers' plies act like metallic sheets and are subject to eddy currents. This will affect the self inductance of the coils as well as their quality factors. This is known as the proximity effect. Intuitively this makes sense, the eddy current will lead to Joule losses which means that powering the coils will dissipate more power and so decrease their ability to "store energy". The penetration of the magnetic field in the conductive material will also be subject to the skin effect and so to a shielding effect. This means that if there is conductive material between the primary and secondary coil, it will diminish the field generated by one coil reaching the other coil and so the mutual inductance will strongly decrease. Figure 2.9 qualitatively illustrates the impact of a strongly conductive sheet on the magnetic field generated by a circular coil. In this case, there is virtually no field reaching the other side of the sheet.

2.2.5 Impact of coils' quality factor decrease on the output power

Figure 2.10 shows the output power obtained with equation 2.3 for a frequency range going from 8 to 12 MHz. The source voltage V_s is set to 1V and the primary and the secondary coils are matched in order to have $\omega_0 = \frac{1}{\sqrt{LC}} = 10 \text{ MHz}$. The transmitter and receiver inductances are chosen to be respectively 5 and 1 μH . The coupling coefficient

¹The full derivation can be found in [55]

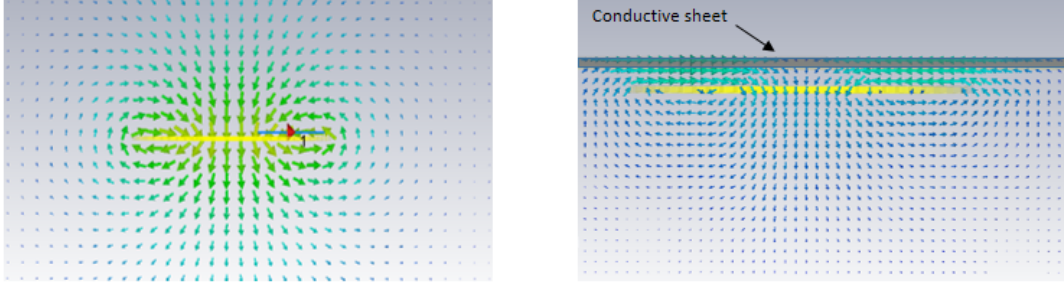


Figure 2.9: Representation of the magnetic field generated by a circular coil (left), with a conducting sheet next to it (right).

k is fixed at 0.1 and 0.3². It will be shown later that those values were too large and extremely optimistic. and the load is 10k Ω . The output power is plotted for multiple Q-factor values. The power drastically decreases with the Q factor. This phenomena is expected to appear with the future system considering that carbon fibers are strongly conductive and since the proximity of a conducting sheet simultaneously increase the serie resistance and decrease the self inductance of the coils.

Decreasing the self-inductances of the coil also has two other impacts on the wireless power transfer. First, it translates the resonant frequency to a different value. This effect can be seen on the left of Figure 2.11. Here a 10% decrease in the self-inductances values has been plugged in the expression of Z_1 and Z_2 for equation 2.3. the matching network is not appropriate anymore to resonate at 10MHz and the peaks are shifted to the right. In such a scenario, if no adaptation is done and that the system still operates at 10MHz, the power transferred is divided by 100 for the highest Q-factor curves of Figure 2.10. Second, a decrease in the self-inductances values has an impact on the mutual inductance $M = k * \sqrt{L_1 L_2}$ and so on the power transfer. This happens even if the network is re-matched to the desired operating frequency. Figure 2.11 (on the right) shows the output power for a 50% self-inductance variation but for which the capacitor values have been changed and rematched to 10MHz. We see on the Figure that for $Q = 200$, this has virtually no effect. For the lower Q-factor, a significant reduction of the output power is observed at the resonant frequency. Already for $Q = 100$, the power is reduced by 35%. For the lowest Q value, the output power is reduced to up to 60%.

Figure 2.12 shows the output power for the same case as Figure 2.10 (left) but with a 2-elements matching network at the receiver. One can see the power is about 5 times

²Those values were chosen arbitrarily and based on various experiment results found in the related literature [22] [52]

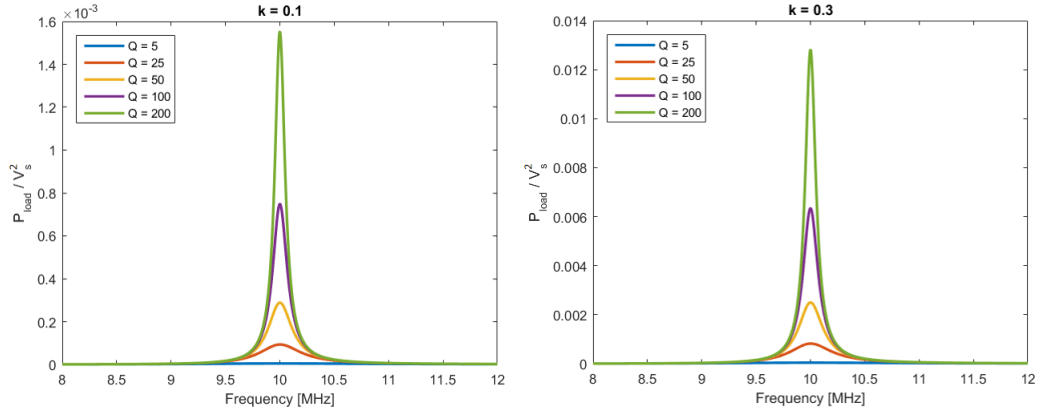


Figure 2.10: Output power in function of frequency for coupling factor $k=0.1$ (left) and $k=0.3$ (right) for inductances of value 1 and $5 \mu H$ and with 1-element matching network.

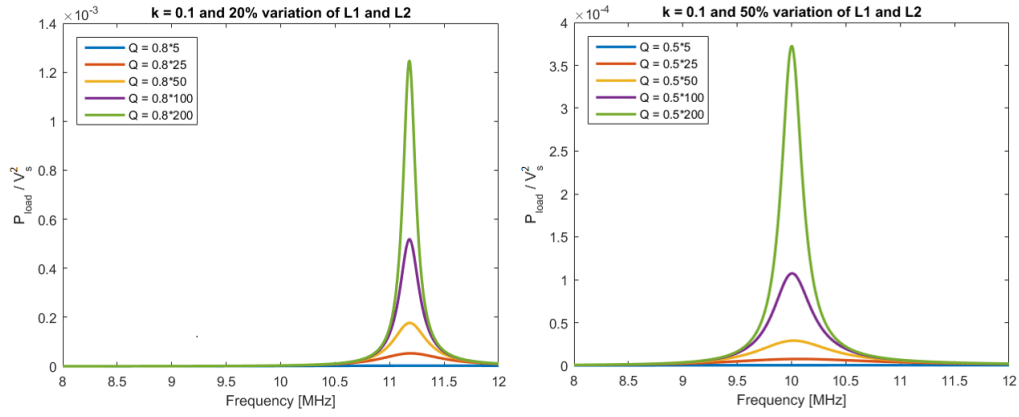


Figure 2.11: Output power in function of frequency for a 20% variation of the self inductance values at the transmitter and at the receiver (left). Output power for a similar 50% variation but with re-matching at both the transmitter and the receiver.

higher than with in the single parallel capacitor case. This shows that using a 2-elements matching network significantly improves the power transmission.

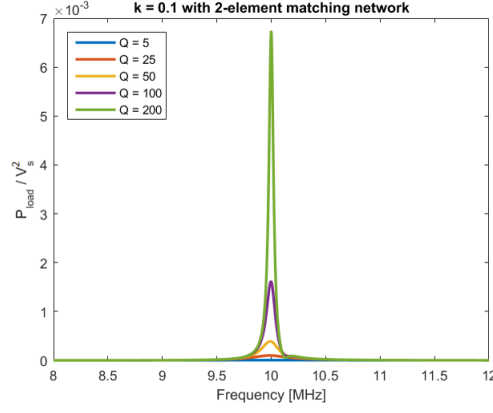


Figure 2.12: Output power in function of frequency for $k = 0.1$ and a 2-elements matching network.

2.3 Full analytical model

The goal of this section is to explain and develop a model to evaluate the self and mutual inductance of circular coils in the surroundings of a conductive sheet. Those results will allow to evaluate the power transfer through a conductive sheet by using the circuit equations of the previous section.

The idea behind the model is to compute the vector potential $\mathbf{A}$ generated by a current loop in all regions of space of Figure 2.14 and to determine the voltage (and so the impedance) through the induced voltage in a length of wire as

$$V = j\omega \int \mathbf{A} \cdot d\mathbf{l} \quad (2.11)$$

2.3.1 Dodd and Deeds' analytic solution

In 1968, Dodd and Deeds' [54] derived the analytic solution for the potential vector due to a circular coil above a conducting sheet. In an homogeneous, linear and isotropic medium of properties μ and ϵ (CFRP will be brought back to that later), the differential vector potential equation due to an applied current density i is

$$\nabla^2 A = -\mu i + \mu\sigma \frac{\partial A}{\partial t} + \mu\epsilon \frac{\partial^2 A}{\partial t^2} + \mu \nabla(1/\mu) \times (\nabla \times A) \quad (2.12)$$

- The term $\mu\epsilon \frac{\partial^2 A}{\partial t^2} \ll \mu\sigma \frac{\partial A}{\partial t}$ can be neglected to frequencies up to the order of 10MHz.
- Axial symmetry can be used since the potential will be symmetric around the coil axis.

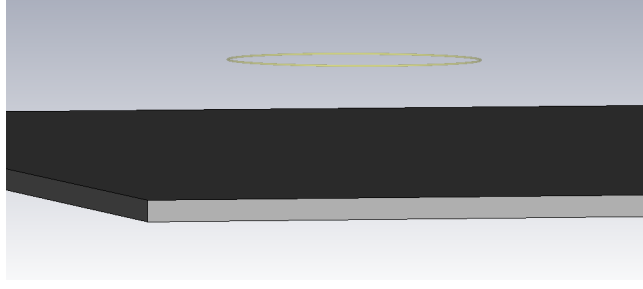


Figure 2.13: Delta function coil above the conducting sheet

- The coil driving current is the sinusoidal function of time $i(t) = i_0 e^{j\omega t}$ and the vector potential can be expressed

$$A(t) = A e^{j\omega t + \phi} = A' e^{j\omega t}$$

Equation 2.12 then becomes

$$\frac{\partial^2 A}{\partial r^2} + \frac{1}{r} \frac{\partial A}{\partial r} + \frac{\partial^2 A}{\partial z^2} - \frac{A}{r^2} - j\omega\mu\sigma A + \frac{\mu I}{\delta(r - r_0)\delta(z - z_0)}$$

where I is the current going through a delta-function coil of radius r_0 and at a height z_0 above the conducting sheet (see Figure 2.13). It is here worth noting that a mistake in the current term in Dood and Deeds' original article has been corrected.

This equation is solved applying the separation of variable $A(r, z) = R(r)Z(z)$. Applying the proper set of boundary conditions between the 4 space regions then leads to a set of 4 vector potential expressions. These expressions only stand for delta-function coils. Assuming a circular coil of arbitrary dimensions is the superposition of delta-function coils, the generated vector potential expressions previously found can be integrated over a "real coil" cross section $\mathbf{S}$ through

$$A_{tot}(r, z) = \int_{\mathbf{S}} A(r', z') d\mathbf{S}$$

Detailed calculations of these steps can be found in [54]. A schematic of the geometry of a real coil is pictured in Figure 2.14 and the model parameters can be found in Table 2.2. Doing this for the vector potential in the different space regions and using

$$\alpha_1 = \sqrt{\alpha^2 + j\omega\mu_1\sigma_1}$$

$$I(r_1, r_2) = \int_{\alpha r_1}^{\alpha r_2} x J_1(x) dx$$

with α a integration variable and $J_1(x)$ the Bessel function of the first kind leads to the following set of expressions:

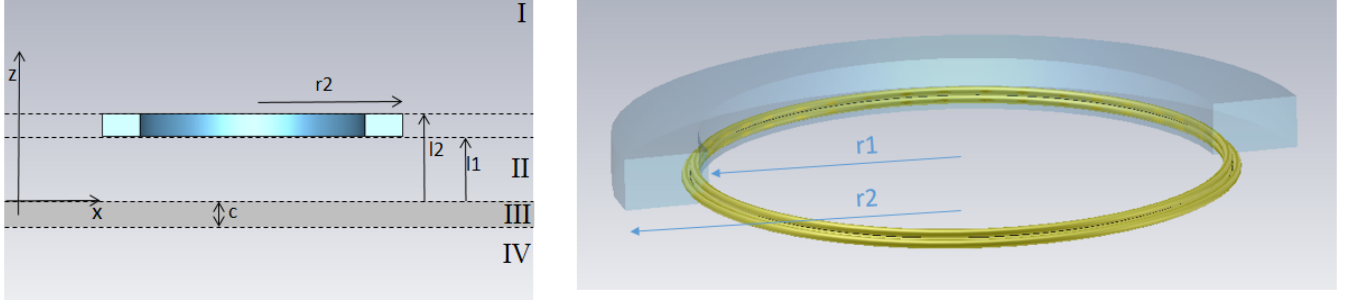


Figure 2.14: Geometry of the problem (left), rectangular cross section coil mad of many delta-function coils.

Table 2.2: Summary of the geometrical parameters and conductive sheet properties of the analytic model represented on Figure 2.14.

Properties	
r_1	inner radius
r_2	outer radius
l_1	distance from bottom of the coil to top of the sheet
l_2	distance from top of the coil to top of the sheet
c	sheet thickness
σ_1	sheet conductivity
μ_1	sheet permeability
Region I	region of space above the coil
Region II	region of space between the coil and the sheet
Region III	region of space within the sheet
Region IV	region of space under the sheet

$$A_I(r, z) = \frac{\mu i_0}{2} \int_0^\infty \frac{1}{\alpha^3} I(r_2, r_1) J_1(\alpha r) \times e^{-\alpha z} \left\{ e^{\alpha l_2} - e^{\alpha l_1} - (e^{-\alpha l_2} - e^{-\alpha l_1}) \left[\frac{(\alpha + \alpha_1)(\alpha_1 - \alpha)(1 - e^{2\alpha_1 c})}{-(\alpha - \alpha_1)^2 + (\alpha_1 + \alpha)^2 e^{2\alpha_1 c}} \right] \right\} d\alpha \quad (2.13)$$

$$A_{II}(r, z) = \frac{\mu i_0}{2} \int_0^\infty \frac{1}{\alpha^3} I(r_2, r_1) J_1(\alpha r) \times [e^{-\alpha l_1} - e^{-\alpha l_2}] \left\{ e^{\alpha z} + e^{\alpha z} \left[\frac{(\alpha + \alpha_1)(\alpha_1 - \alpha)(1 - e^{2\alpha_1 c})}{-(\alpha - \alpha_1)^2 + (\alpha_1 + \alpha)^2 e^{2\alpha_1 c}} \right] \right\} d\alpha \quad (2.14)$$

$$A_{III}(r, z) = \mu i_0 \int_0^\infty \frac{1}{\alpha^3} I(r_2, r_1) J_1(\alpha r) \times [e^{-\alpha l_1} - e^{-\alpha l_2}] \left[\frac{\alpha(\alpha + \alpha_1)e^{2\alpha_1 c} + \alpha(\alpha_1 - \alpha)e^{-\alpha_1 c}}{-(\alpha - \alpha_1)^2 + (\alpha_1 + \alpha)^2 e^{2\alpha_1 c}} \right] d\alpha \quad (2.15)$$

$$A_{IV}(r, z) = \mu i_0 \int_0^\infty \frac{1}{\alpha^3} I(r_2, r_1) J_1(\alpha r) \times [e^{-\alpha l_1} - e^{-\alpha l_2}] \left[\frac{2\alpha\alpha_1 e^{c(\alpha + \alpha_1)} e^{\alpha z}}{-(\alpha - \alpha_1)^2 + (\alpha_1 + \alpha)^2 e^{2\alpha_1 c}} \right] d\alpha \quad (2.16)$$

An extra expression is necessary to characterize the region within the coil winding (between regions I and II)

$$A_{I,II}(r, z) = \frac{\mu i_0}{2} \int_0^\infty \frac{1}{\alpha^3} I(r_2, r_1) J_1(\alpha r) \times \left\{ 2 - e^{\alpha(z-l_2)} - e^{-\alpha(z-l_1)} + e^{-\alpha z} [e^{\alpha l_1} - e^{-\alpha l_2}] \left[\frac{(\alpha + \alpha_1)(\alpha_1 - \alpha)(1 - e^{2\alpha_1 c})}{-(\alpha - \alpha_1)^2 + (\alpha_1 + \alpha)^2 e^{2\alpha_1 c}} \right] \right\} d\alpha \quad (2.17)$$

These expressions represent the vector potential generated by a circular coil of rectangular cross section $(l_1 - l_2)(r_2 - r_1)$ and excited by a sinusoidal current of amplitude

i_0 . This coil is at a distance l_1 from a infinite conductive sheet of conductivity σ and thickness c . Now that the vector potential is analytically expressed in all regions, the self and mutual inductances of any circular coil can easily be obtained by using equation 2.11 and integrating the above expressions.

Self inductance computation

Expression 2.11 allows to obtain the self inductance of the source coil itself in presence of a conductive sheet. Again, if the coil chosen has a rectangular cross section made of N turns of circular wire, the intercepted potential is considered as being equivalent to the potential intercepted by a single loop of wire of rectangular cross section $(l_1 - l_2)(r_2 - r_1)$ multiplied by N . The vector potential intercepted by such a loop is obtained by integrating the corresponding vector potential in which the pick-up coil lays over its cross section. Doing this analytically for the source coil in region I and using $\mathbf{Z} = \mathbf{V}/\mathbf{I}$

$$\begin{aligned} \mathbf{Z} = R + j\omega L = & \frac{j\omega\pi\mu n^2}{(l_2 - l_1)^2(r_2 - r_1)^2} \int_0^\infty \frac{1}{\alpha^5} I^2(r_2, r_1) \left(2(l_2 - l_1) + \frac{1}{\alpha} \right. \\ & \times \left. \left\{ 2e^{-\alpha(l_2 - l_1)} - 2 + e^{-2\alpha l_2} + e^{-2\alpha l_1} - 2e^{-\alpha(l_2 + l_1)} \left[\frac{(\alpha + \alpha_1)(\alpha_1 - \alpha)(1 - e^{2\alpha_1 c})}{-(\alpha - \alpha_1)^2 + (\alpha_1 \alpha)^2 e^{2\alpha_1 c}} \right] \right\} \right) d\alpha \end{aligned} \quad (2.18)$$

where $\mathbf{Z}$ is the impedance of the coil. R represents the losses due to eddy currents in the conductive sheet and does not take the wire resistance into account. If α_1 is replaced by α , $R = 0$ and this expression give the inductance of the circular coil in air.

Mutual Inductance computation

A similar expression can be obtained for a pick up-coil of dimensions l'_1, l'_2, r'_1, r'_2 located in region IV

$$\mathbf{V}_2 = \frac{j\omega 2\pi n}{\text{coil cross section}} \int_{l'_1}^{l'_2} \int_{r'_1}^{r'_2} r \cdot A_{IV}(r, z) dr dz = j\omega \mathbf{M} I_1$$

with $I = \frac{I_0(l_1 - l_2)(r_2 - r_1)}{N}$

Defining

$$K = \frac{2\pi n N}{(l_1 - l_2)(r_2 - r_1)(l'_1 - l'_2)(r'_2 - r'_1)}$$

$$\mathbf{M} = K \frac{1}{i_0} \int_{l'_1}^{l'_2} \int_{r'_1}^{r'_2} r A(r, z) dr dz$$

Using the expression of A in zone IV

$$\begin{aligned} \mathbf{M} = K \frac{1}{i_0} \int_{l'_2}^{l'_1} \int_{r'_1}^{r'_2} \int_0^\infty \frac{1}{\alpha} r r' J_1(\alpha r) J_1(\alpha r') \\ \times [e^{-\alpha l_1} - e^{-\alpha l_2}] \frac{2\alpha \alpha_1 e^{c\alpha + c\alpha_1} e^{\alpha z}}{-(\alpha - \alpha_1)^2 + (\alpha + \alpha_1)^2 e^{2\alpha_1 c}} dr' d\alpha dz \end{aligned} \quad (2.19)$$

using $\int_{x_1}^{x_2} x J_1(\alpha x) dx = \frac{1}{\alpha^2} I(a, b)$ and integrating over z

$$\begin{aligned} \mathbf{M} = K \mu \int_0^\infty \frac{1}{\alpha^5} I(r'_1, r'_2) I(r_1, r_2) \\ \times [e^{-\alpha l_1} - e^{-\alpha l_2}] [e^{\alpha l'_1} - e^{\alpha l'_2}] \frac{2\alpha \alpha_1 e^{c\alpha + c\alpha_1}}{-(\alpha - \alpha_1)^2 + (\alpha + \alpha_1)^2 e^{2\alpha_1 c}} d\alpha \end{aligned} \quad (2.20)$$

Here $\mathbf{M}$ is complex. The real part represents losses and does not have any direct meaning as in the case of the self-inductance. The mutual inductance between the source coil located in region IV and the pick up coil located in region IV is

$$M = \text{Im}(\mathbf{M})$$

Again, if the sheet conductivity is zero, α_1 can be replaced by α and this expression reduces to its imaginary part.

2.4 Verification of the analytical model

In order to verify the accuracy of the computation, three circular coils were hand made and compared to simulation results. They were built with a 1.26mm diameter wire. Figure 2.15 shows the pictures of the actual coils. Coil1 and coil2 are 4-turns solenoids and coil3 is 3-turns pancake coil.



Figure 2.15: Coil 1, 2 and 3 from left to right

	<i>Coil 1</i>	<i>Coil 2</i>	<i>Coil 3</i>
Inner radius	1.8 cm	1.8 cm	2 cm
Outer radius	1.92 cm	1.92 cm	2.378 cm
Turns	4	4	3
LCR meter (L , σ_L)	809 nH, 2.95	839 nH, 28.94	803nH, 6.59
Basic equation 2.6	826.6 nH	826.6 nH	771nH
Analytic equation 2.18	844.1 nH	844.1 nH	769nH
Error $(L_{LCR} - L_{DoddS})/L_{LCR}$	4.34%	0.61%	4.23%

Table 2.3: Analytic model test coils measurement and computation results comparison at 10MHz,

First, their self-inductance values were measured at 10MHz with an HP 4275A LCR meter (actually measured 5 times and the mean was taken, in Table 2.3 σ_L is the standard deviation over the measurements). Equation 2.18 was approximated with Matlab for similar geometries and a third set of values was obtained with equation 2.6. All sets of values were compared and can be found in Table 2.3. The analytic computation shows fair correlation with the LCR result. A sub 5% error was observed and can be accounted to the inaccuracy of hand fabrication process as well as the rectangular cross section approximation. There exist simpler and more precise analytic expressions for computing coils of these precise geometries but these methods cannot take the presence of conducting material into account. The frequency of 10MHz was choosen due to equipment limitations. The original design goal was a standard operating frequency of 13.56MHz but the LCR meters available do not offer measurement frequency above 10MHz. It is assumed that any of the results here also hold at 13.56MHz. It is worth noting that the numerical evaluation of 2.18 is extremely slow. The evaluation of the integral over α from 0 to infinity can be reduced to $[0,2000]$ in order to get a maximum error of 1% and to $[0,1000]$ for 5% error (error compared to the result obtained when taking $[0, \infty]$).

Second, a similar assessment of equation 2.20 was done. The coupling coefficients between coils 1 and 2 and between coils 1 and 3 were measured and equation 2.20 numerically evaluated for primary to secondary spacings going from 5mm to 5cm. To do so, a wave generator was connected to the primary to feed sinusoidal signals of 10Vpp and 20Vpp. The open voltage was simultaneously measured at the secondary to recover the mutual inductance through $M = L_1 * V_2 / V_1$. The results are plotted in Figure 2.16. Measurements

were made for two different input voltages in order to verify that the set-up was correct and the results repeatable. The error in Figure 2.16 come from similar reasons as for the self inductance calculation. The simulations mainly fit the measurements but present a positive offset. This offset may come from various sources. A potential explanation could be the AC resistance of the wire which are not taken into account in the analytic model. Considering both the primary and secondary coils are identical, they show the same series resistance. Looking back at the reflected load impedance circuit model (Figure 2.5), from the source point of view the series resistances of the primary appears larger than the secondary reflected resistance. This leads to a decrease of V_2/V_1 if those resistance are taken into account which is the case when measuring real coils.

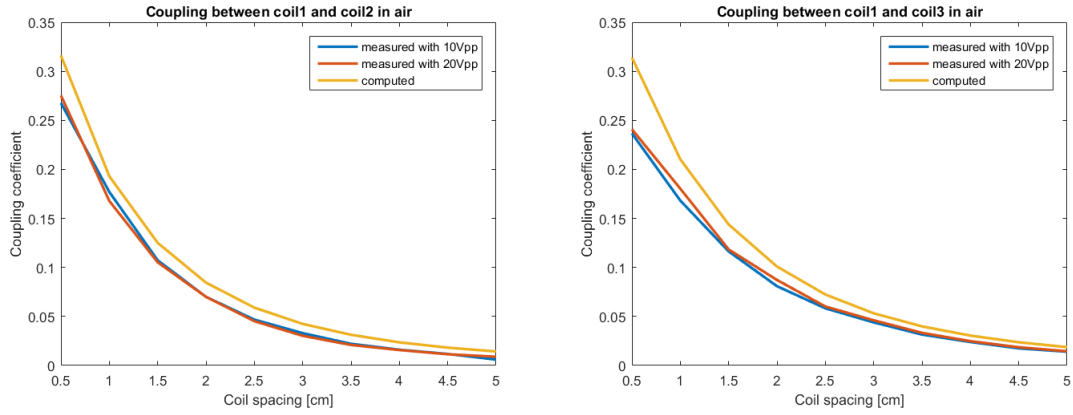


Figure 2.16: Comparison of measurements and simulations of the coupling coefficient between coils 1-2 and coils 1-3

Third, the calculation was verified in the surroundings of a conductive sheet. To do so, preliminary measurements of the inductances of coils 2 and 3 were done with the LCR when nearby a $440\text{ }\mu\text{m}$ copper sheet and a $30\text{ }\mu\text{m}$ aluminum sheet respectively. These values were compared with the imaginary part of the simulation results using the standard conductivities $\sigma_{Cu} = 5.96 \cdot 10^7\text{ S/m}$ and $\sigma_{Al} = 3.5 \cdot 10^7\text{ S/m}$. Figures 2.17 and 2.18 show the results. The tables are the measured and calculated inductance values. The plots represents the normalized inductance value L/L_{air} . Dividing by the value in air allows to minimize the impact of the original error between the measured values and the calculated values of coil2 and coil3. The correlation between the data sets is easy to see. Even if both curves are not identical, the overall aspects are the same. Considering this test and the previous ones, it is concluded that the analytic model evaluated and expresses fairly the impact of a conducting sheet on self inductance.

		Copper sheet 440 μm	
	Spacing	Measured	Calculated
Coil 2	5mm	654.4 nH	546.0 nH
	10mm	772.7 nH	773.2 nH
	15mm	809.6 nH	810.4 nH
	20mm	828.1 nH	825.5 nH

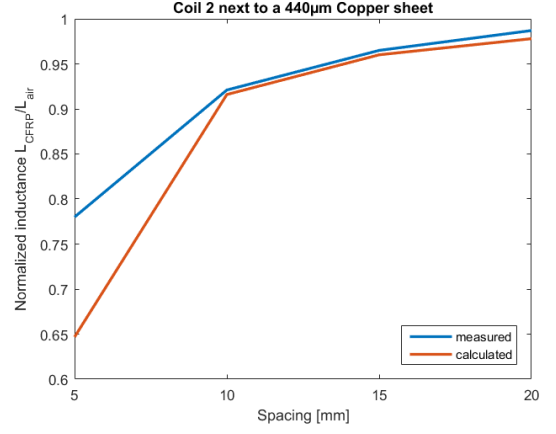


Figure 2.17: Summary of the self-inductance measurements and simulations (equation 2.18) results for coil 2 next to a 440 μm Copper sheet.

		Aluminum sheet 30 μm	
	Spacing	Measured	Calculated
Coil 3	5mm	540.0 nH	546.0 nH
	10mm	679.9 nH	646.0 nH
	15mm	747.4 nH	696.0 nH
	20mm	775.2 nH	723.6 nH

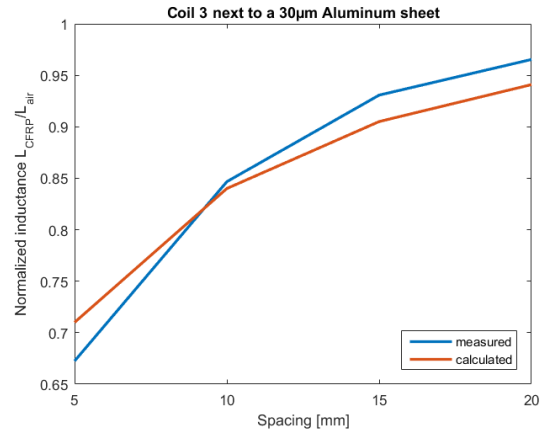


Figure 2.18: Summary of the self-inductance measurements and simulations (equation 2.18) results for coil 3 next to a 30 μm Aluminum sheet.

Table 2.4: Case study parameters summary.

Parameter	Value
Coil mean radius	5, 25 and 50 mm
Sheet conductivity	$50 * 10^3$ S/m
Sheet thickness	$300\mu m$ (1ply)
Wire cross-section	$1mm \times 1mm$
Frequency	10 MHz

2.5 Preliminary case study

Now that the analytic computation has been showed to produce fairly accurate results, it is possible to evaluate the impact of a conductive sheet on the self-inductance and quality factor of coils. This has been done for three circular loops of different radius and which are different then the previous test coils. Here, more realistic dimensions for a RFID system were chosen (readers of 2.5 and 5 cm radius and receiver of 5 mm radius). The simulation was done for coil-to-sample spacings going from 0 to 12 cm and for a conductive sheet of thickness similar to one ply of carbon fiber composite. The conductivity was approximated by $50 * 10^3$ S/m based on values found in the literature [52] [56] [57]. Table 2.4 summarizes the simulation parameters.

Figure 2.19 (on the left) shows a plot of the normalized self inductance of the loops. The first thing to notice is that for small distances of less than 1cm, the self-inductance is strongly reduced in all cases. At a distance of 1 mm, the self-inductance of the coil of radius 0.5 cm is reduced to 65% of its original value. For larger coils, in our case their self-inductance is reduced down to 35% of the original value for that distance. For a given system, such variations will strongly affect the resonant frequency of the LC tank. The transmitter and receiver will not be tuned anymore which will drastically reduce the output power.

On the second plot of Figure 2.19 one can see the impact of the same conductive sheet on the Q factors of the coils. Those values were obtained modeling the series resistance of the coil as the summation of the real part of equation 2.3 with the AC resistance obtained with equation 2.21. The AC serie resistance of a wire of length L and cross section πr^2 is then obtained using

$$R_{AC} = \frac{L}{\sigma A_{eff}} \quad (2.21)$$

with $\delta = \sqrt{\frac{1}{\pi f \sigma \mu}}$ and $A_{eff} = \pi r^2 - \pi(r - \delta)^2$

The same phenomena as for the self inductance is observed. For short distances (less than 1cm) the quality factor are strongly impacted by the sheet. A smaller coil also seems to perform better than a larger coil even if the larger coil originally features a much better Q-factor. As shown previously in this chapter, such decrease of the Q-factor of the transmitting or receiving coil strongly affect the power transfer efficiency.

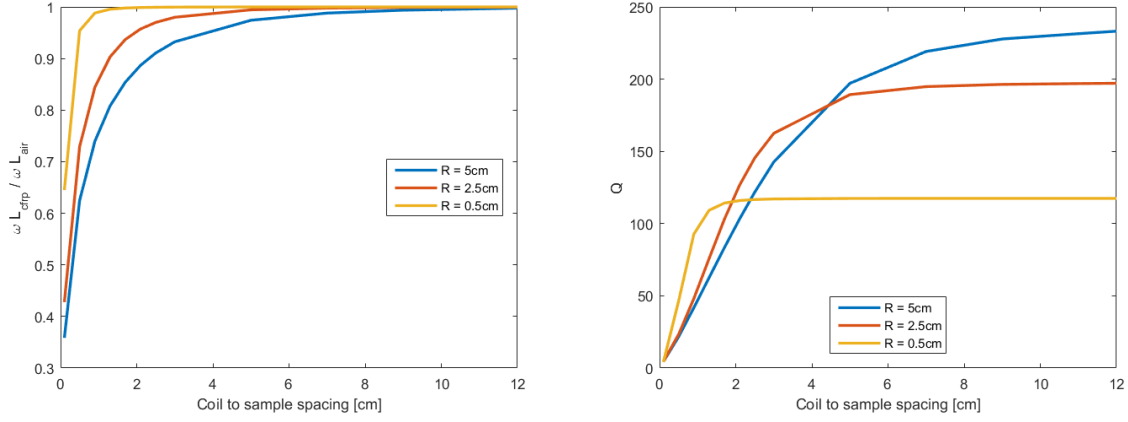


Figure 2.19: Normalized self inductance and Quality factor of the test coils next to a conductive sheet.

The mutual inductance between these coils can also be computed with equation 2.20. This was done for a first scenario of two 2.5 cm radius coils and for a second scenario of a 2.5 cm and a 0.5 cm radius coil. Total primary to secondary spacings were set going from 1 to 25 mm. The transmitting and receiving coils were first separated only by air then a 300 μm sheet was inserted in between. When computed with a inserted sheet, the secondary coil was placed at a fixed distance of 300 μm under the sheet and only the primary coil position was incremented. Figure 2.20 shows a schematic of the case study.

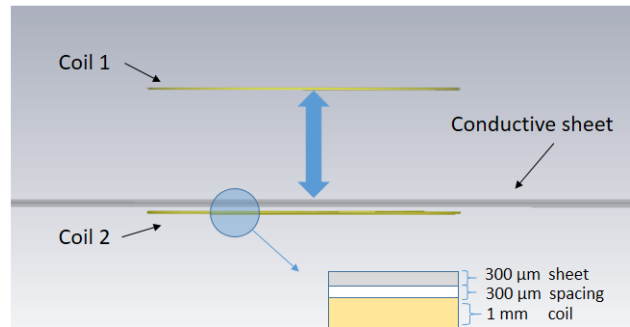


Figure 2.20: Graphic representation of the case study.

Figure 2.21 shows the results obtained. In air and for the 2.5 cm coils, the coupling is excellent. With the conductive sheet inserted, k still reaches 9% for a 1 mm spacing. When using the 0.5 cm coils, the coupling coefficient drop down to around 6% in the best case. The difference in order of magnitude between the two different receiving coils in air makes intuitive sense considering the magnetic flux intercepted is a function of the coil surface and that the surfaces ratio is $\pi r_2^2 / \pi r_1^2 = 0.5^2 / 2.5^2 = 1/25$. We can also see that while in the first scenario the decrease of k with and without the sheet reaches up to 85%, in the second scenario the decrease is about 50% (from 6 % to 3% at 1 mm spacing).

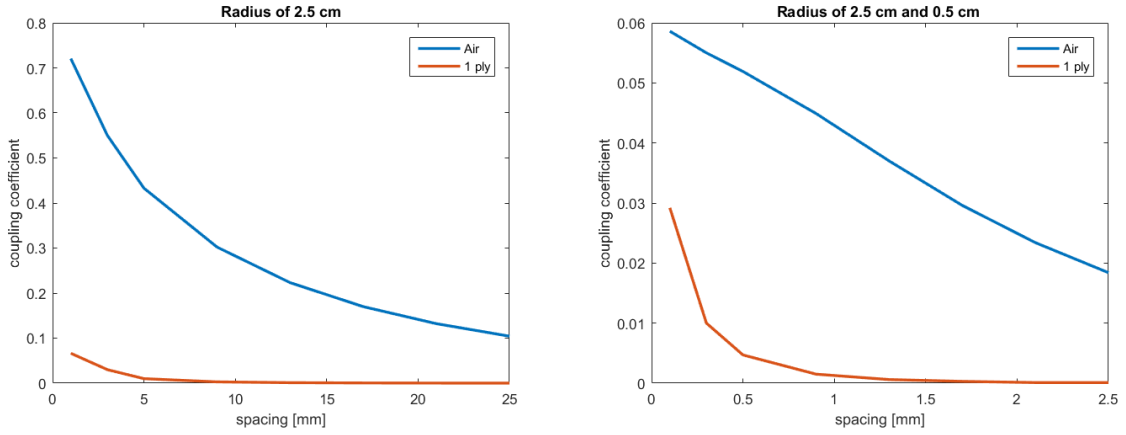


Figure 2.21: Coupling coefficient between two circular loop of radius 2.5 cm (left) and or radius 2.5 cm and 0.5 cm (right).

With data on the self inductance, quality factor and coupling coefficient it is now possible to evaluate the power transfer efficiency in these conditions with the circuit model developed earlier. To do so, the previously computed data set were plugged into equation 2.4. Both primary and secondary were matched according to the 2-elements matching network and at all distances from the conductive sheet. This means that the matching network capacitors values are recomputed and different for each spacing. Figure 2.22 shows the results for scenario 1 and Figure 2.23 for scenario 2. The efficiency obtained with the CFRP are really bad. For scenario 1, it reaches half a percent for 1 mm distance which is really low considering the 35% obtained without the sheet. For scenario 2, the efficiency drops from roughly a few percent to around $10^{-3}\%$. This means that if 1W of power is emitted, only $10 \mu\text{W}$ will reach the load.

Considering the result of this case study, it can be concluded that if CFRP can effectively be modeled as an homogeneous conductive sheet, its impact on the power transmission will be important. The power transfer must be improved one way or another in order to make applications such as embedded wireless sensor nodes possible.

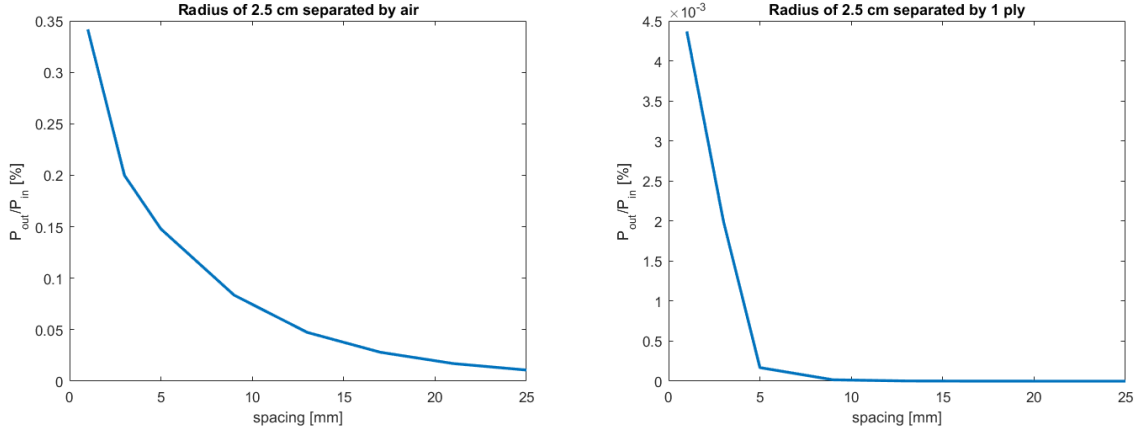


Figure 2.22: Power transfer efficiency between two circular loop of radius 2.5cm separated by air (left) and separate by a conductive sheet equivalent to one CFRP ply (right).

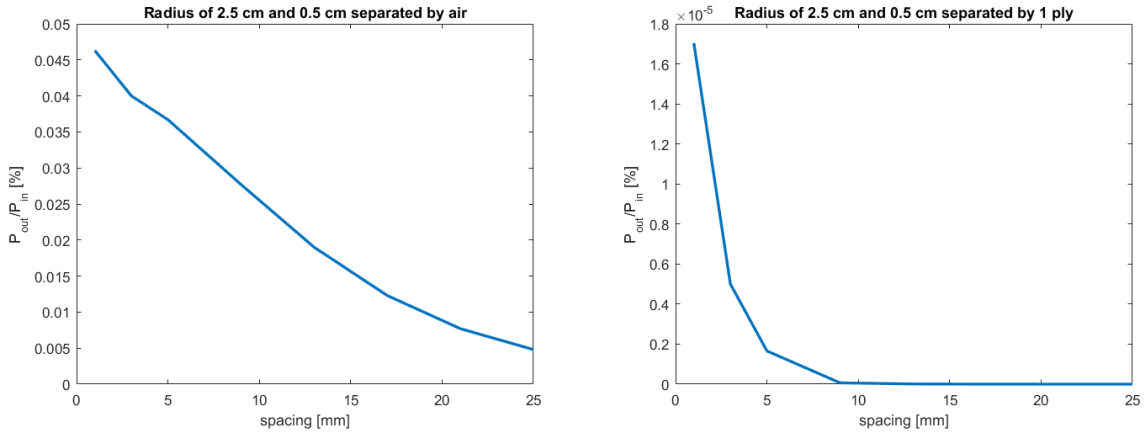


Figure 2.23: Power transfer efficiency between two circular loops of radius 2.5 cm and 0.5 cm separated by air (left) and separated by a conductive sheet equivalent to one CFRP ply (right).

2.6 Conclusion

In this chapter, a complete power transfer analytic model was built to evaluate the impact on a conducting sheet on the power transmission. The self and mutual inductance as well as the quality factors of the primary and secondary coils are calculated through a complete electromagnetic model. With those data, the power transfer between a transmitter and receiver can be evaluated at various frequencies and for different matching networks with a reflected load circuit model of the inductive link. It was also shown that the power transfer efficiency is load dependent and that if the load, the operating frequency and the primary and secondary coils are fixed, using a 2-elements matching network might multiply the output power by a factor of 5.

The model was also experimentally validated for different test coils in air and in the surroundings of conductive sheet. Finally, all the model equations were numerically evaluated for various geometries and it was shown that the sheet strongly affects the self-inductance, the quality factor and the coupling coefficient of the coils.

This overall model has one big limitation considering it only take into account the conductive sheet separating the two coils (and not the conductive material behind and on the side of the receiver). In order to accurately evaluate the carbon-fiber composite impact on the wireless power transmission to a receiver embedded in it, a more accurate model and methodology is necessary. This model will allow to make predictions and so to take the carbon fiber composite into account when designing the inductive link. Considering this, the next chapter covers a simulation based model.

Chapter 3

Design flow and effective properties approximation

3.1 Enhanced design flow

In the previous chapter, it was shown that CFRP has a profound impact on the wireless power transfer due to screening and shielding effects. In order to efficiently design an inductive link using an embedded receiver, a more accurate model of the situation is necessary. The model will permit to predict accurately these effects in order to take them into account in the design. In order to efficiently transmit power by inductive coupling, the primary and secondary networks need to resonate at the same frequency. The presence of conducting material nearby either the transmitter or the receiver will affect its coil inductance and so the resonant frequency. In order to evaluate and predict that change in resonant frequency, the surrounding material conductivity has to be known. Measuring conductivity usually is decently straightforward but not in the case studied here. Carbon fibers reinforced polymer conductivity significantly varies with fabrication process, frequency and knit architecture.

Considering the complex geometry of the full problem, it is not possible to build a complete analytic model of CFRP. At the opposite, it is possible to build to simulate any 3D design. Such simulations allow to predict the behavior of a chosen design when inserted into any material as long as the material properties are known. From the 3D simulation results, one can then extract equivalent lump parameters. Finally, doing so and using the proper tools allows to take them into account in the design of the matching network.

Carbon composites have a structure too complicated to be reproduced in EM simulation

software, modeling either the fibers themselves or the tow and weave would also lead to long and slow run times. In order to simulate any coil embedded into a composite material, an homogeneous model of the material is so also necessary. Considering the fluctuation of the conductivity with various factors, this model also has to be correlated with experimental data before making any prediction. Following this, we here propose the following design flow:

- Approximation of CFRP by an homogeneous medium through effective properties.
- 3D design and simulation with a first receiver embedded in this equivalent homogeneous media.
- Fabrication and measurement of coils identical to the coils simulated.
- Calibration of conductivity of the CFRP based on a comparison between the experiment and the simulation measurement.
- Equivalent lumped parameters extraction from the 3D simulation results and construction of a circuit based model of the inductive link.
- The power transfer is assessed using the lump parameters and the circuit model of chapter 1.
- Modification of the receiver coil design in order to obtain the desired power transfer efficiency., the behavior of the new designs should now be accurately predicted.

This chapter goes through the design flows steps that currently have not been overviewed. First the lump parameter extraction technique with CST is detailed. Second, effective properties of the composite are obtained. A simple composite fabrication process is than explained. Finally, three embedded test coils are embedded into composite sheets and the same coils are modeled in CST. The accuracy of the homogeneous material model is assessed and the both result sets compared.

3.1.1 Lumped parameters extraction to build an accurate circuit model.

The lumped parameters extraction was applied to a case study similar. First, a 3D design of inductively coupled coils was built in CST microwave studio. Figure 3.1 shows the schematic and Table 3.1 summarizes the geometrical parameters used. The system was simulated using the frequency domain solver for a frequency range going from 8MHz

to 12MHz. Considering the low frequency and huge wavelength, a time analysis is too computation demanding. Solver boundaries were set as open and the domain boundaries set at 150mm.

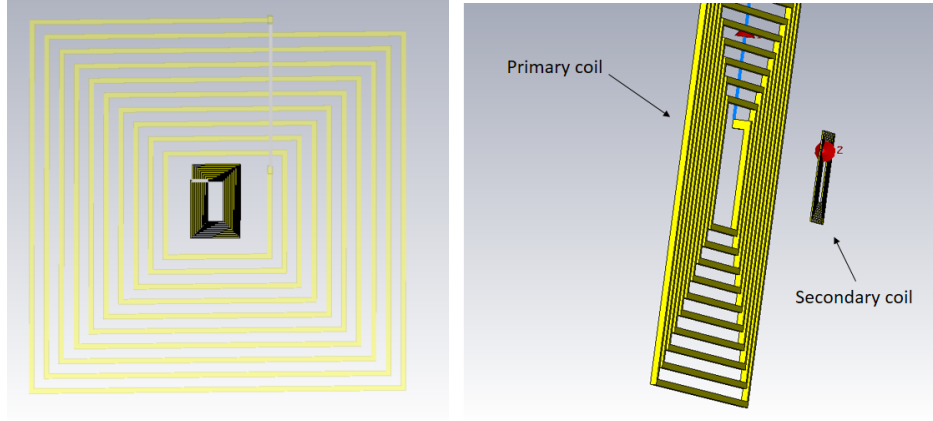


Figure 3.1: 3D design of the coupled reader and receiver coils

Table 3.1: 3D case study parameters summary.

Physical properties	Primary	Secondary
Outer dimensions	76mmX76mm	15mmX10mm
Inner dimensions	20mmX20mm	8.2mmX3.2mm
Trace thickness	1mm	0.2 mm
Trace width	1mm	0.2 mm
Spacing	2 mm	0.2 mm
Number of turns	10	10
Material	Pure Copper	Pure Copper

After, the self and mutual inductance of the discrete ports were obtained using the post processing template tools. These values and full 3D simulation results were exported in *CST Design Studio* where a simple SPICE circuit was built. Figure 3.2 shows the display in *Design Studio*: the top part represents the full 3D design simulation and its access port (numbered 1 and 2), the bottom part represents an equivalent circuit made of two perfect coupled inductors. The circuit inductors self and mutual inductances are initially set to the values obtained from the 3D simulations. A SPICE simulation can then be run for the circuit for the same frequency range and the results compared with the results of the full 3D simulation. Figure 3.3 shows a plot of the receiver side S parameters for both cases

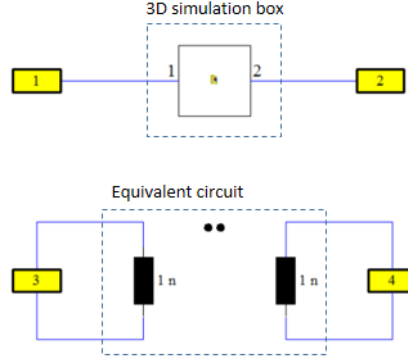


Figure 3.2: Primary circuit representation of the 3D model

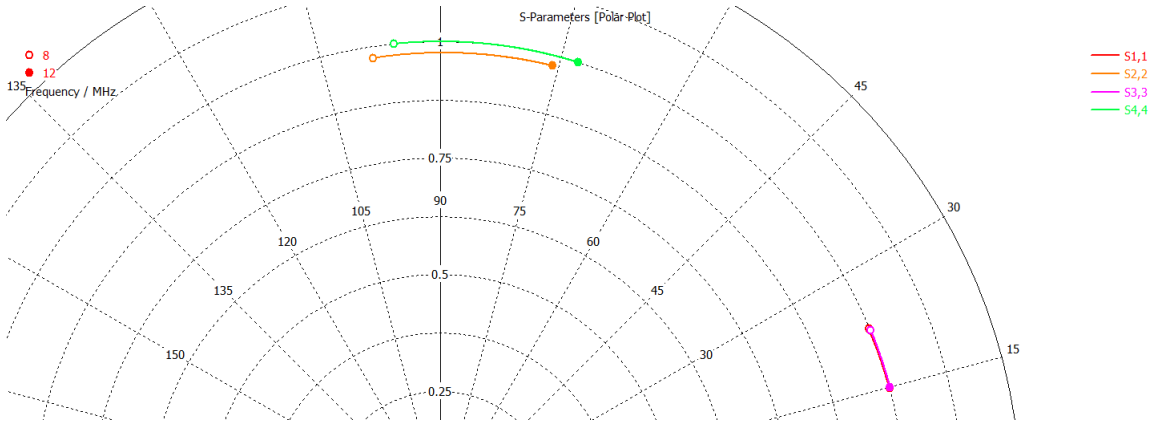


Figure 3.3: Comparison of S parameters between the 3D (S11, S22) and circuit simulations(S33, S44)

over the frequency range [8MHz,12MHz]. On the plot, S22 is the reflection coefficient at the receiver side for the 3D simulation and S44 for the circuit simulation. There is a significant deviation between the two curves on the plot. **This means is that the coupled inductances do not accurately represent the 3D design. The circuit model is actually incomplete and needs to be enhanced by lump elements.**

Full circuit model

Figure 3.4 shows a more complete circuit parameter model. Through the use of a built-in optimizer, it is possible to match the S-parameter results of the 3D simulation with a equivalent circuit model. This means we can ask the software to find values of the lump element parameters listed in the table in Figure 3.4 that will fit the values of S44 the closest to the values of S22. Doing so we can built a circuit model more accurately rep-

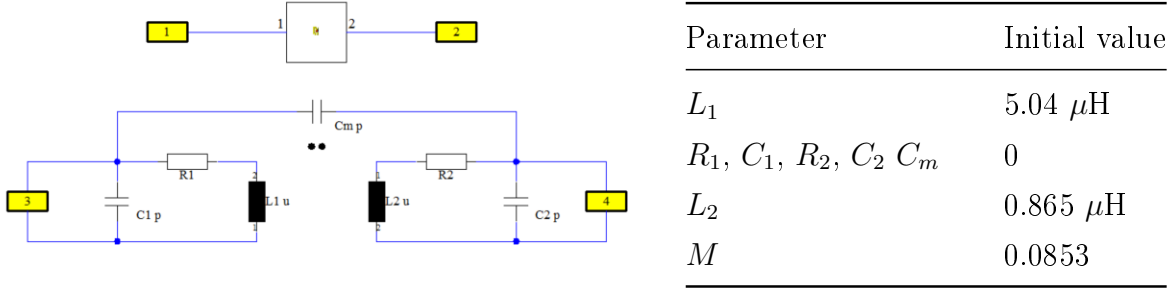


Figure 3.4: Full circuit representation of the 3D model and initial parameter set.

representing the 3D model which can be very useful in assessing the behavior of the system in a given environment.

Important point

In this case, considering there are many parameters to vary, the receiver and transmitter coils were first optimized alone (one coil at the time) in order to get first order values for the series resistance and parallel capacitance. This allowed to restrain their variation domains in the full optimization and so to significantly reduce the computation time. Here the *Trust region framework optimizer* was used and with an accuracy of 0.0001. It was given a single goal point at 10 MHz. It is very important to specify goals and for the amplitude and for the phase otherwise the optimizer does not reach the correct parameter set. Figure 3.5 shows the S parameters comparison for the receiver coil only. Figure 3.6 shows the S parameters comparison after amplitude optimization and Figure 3.7 after amplitude and phase optimization. In the later case, as predicted, the two curves overlap. The optimized circuit parameter obtained with this precise simulation can be found in Table 3.2.

Table 3.2: 3D case study receiver parameters.

Parameter	Before	magn. otpimization	magn. and phase otpimization
L	$0.865 \mu\text{H}$	$0.8666 \mu\text{H}$	$0.8407\mu\text{H}$
R_s	0	0.2104Ω	0.22Ω
C_p	0	4.678 pF	8.569 pF

So as previously said, this analysis can be done for the transmitter and receiver coil

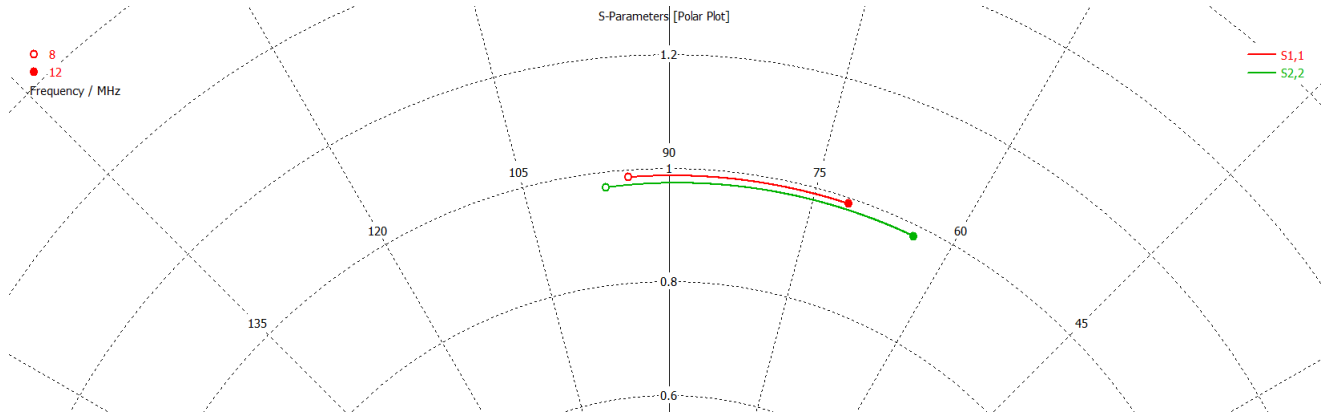


Figure 3.5: Case study S11 parameter before optimization

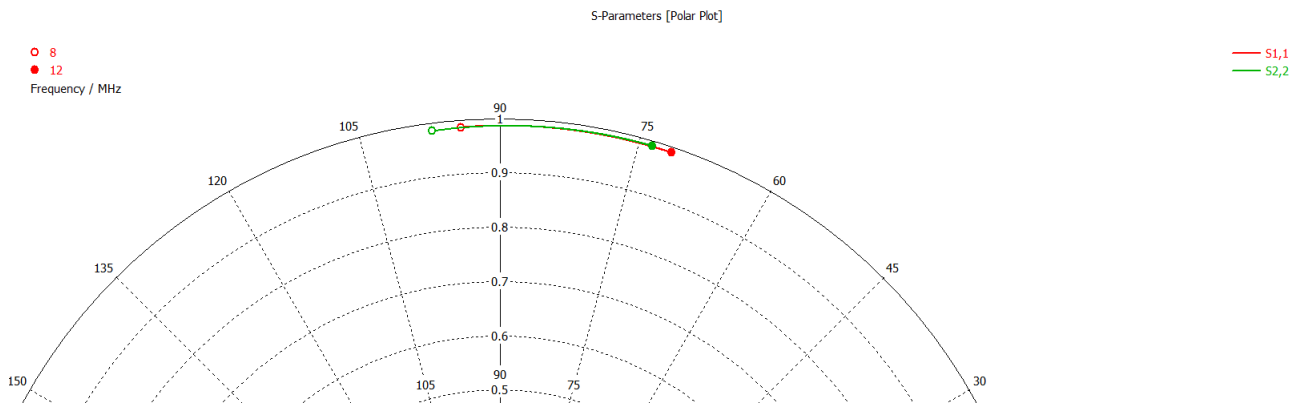


Figure 3.6: Case study S11 parameters after amplitude optimization

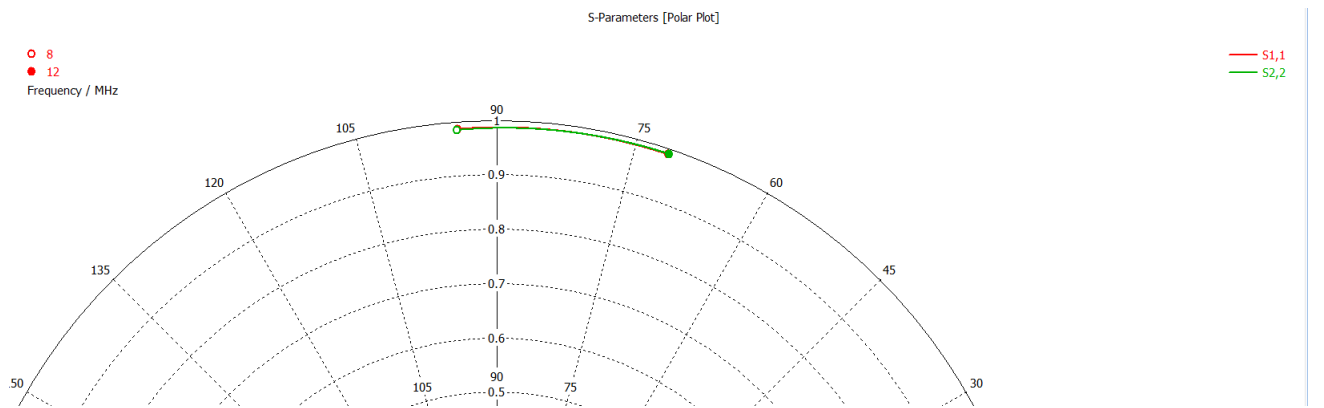


Figure 3.7: Case study S11 parameter after amplitude and phase optimization (right)

independently. Then, using the lump parameter values obtained as initial value, a similar optimization can be done for the two inductively coupled coils together . This was done here and Table 3.3 synthesizes the results obtained for this case study.

This analysis can be performed either with a conductive sheet between the two inductors or embedding the receiver in the conductive sheet. In those cases, the impact on the self-inductances, the quality factors and the mutual inductance can be studied and predicted in the surroundings of an homogeneous conductive sheet

Table 3.3: 3D case study parameters.

Parameter	Before optimization	After optimization
L_1	5.04 μH	4.8364 μH
R_1	0	0.0103 Ω
C_1	0	0.0724 pF
L_2	0.865 μH	0.8422 μH
R_2	0	0.2107 Ω
C_2	0	7.983 pF
C_m	0	0.0005
M	0.0853	0.0871

3.2 Effective properties of the composite ply

In order to simulate the wireless power transfer for an embedded receiver, CFRP needs to be approximated as being homogeneous. Implementing and simulating the full material structure to extract effective properties would be extremely computing and time demanding, so an alternative needed to be explored. For the permittivity and the permeability, two approximation models are tried: the 1-layer model and the 3-layer model. For the conductivity, a measurement based method comparing a numerical evaluation of equation 2.18 with experimental results was used.

3.2.1 CFRP electrical conductivity

The electrical conductivity measurements of carbon fiber reinforced composite is a subject that does not make unanimity in the scientific world.

Over the years, some studies have claimed that the conductivity of the carbon fiber composite was frequency independent. Others have precisely shown the opposite and accurately measured it for wide frequency ranges. This fracture between studies comes from two different origins. First, all general properties like density, tensile strength or heat conductivity of CFRP strongly depends on the fabrication processes. Considering this, it

makes sense that different studies would measure different electrical conductivities since they mostly used sample produced differently. Second the methods in themselves might lead to different results with. People have often been trying to reuse standard method which are not applicable to composite laminates. For instance, the use of isotropic probes to measure the conductivity of unidirectional ply which is highly anisotropic [58].

It is also basically impossible to create an analytic model for the electromagnetic behavior of the laminate and to make prediction with it. The reason for this is the highly complex geometrical nature of the material composition, a woven of tows themselves made of individual extremely conductive fibers embedded in a non-conductive resin.

Intuitively, it makes sense that the conductivity of carbon fiber reinforced polymer changes with frequency. At DC, there is no electrical connection between the fibers of parallel plies and the conductivity perpendicular to the fibers direction is very low. Under AC stimulation, the fibers can be assimilated to parallel wires and the layers act like a capacitance. For a unit length of parallel wires, this capacitance is equal to

$$C = \frac{\pi \epsilon_r \epsilon_0}{\text{arcosh}(D/d)}$$

where D is the distance between the wires axis and d is the wires diameter. The impedance magnitude of a capacitor is expressed as $|Z| = 1/\omega C$ which decreases with frequency. From this, it can be understood that a very low conductivity CFRP ply at low frequencies may become really conductive at high frequencies.

3.2.2 Electrical conductivity measurements

As mentioned before, measuring conductivity usually is decently straightforward but not in the case studied here. Carbon fibers reinforced polymer conductivity significantly varies with fabrication process, frequency and knit architecture. Multiple methods were so tried in this work in order to determine an effective conductivity for the CFRP used later on. First, measurements using a test fixture HP16541B read by a LCR HP4284A and automated through LabView were conducted for frequency going from 1kHz to 1MHz. The results were not repeatable and it was concluded that the method was not appropriate. The surface of the composite test sample used for the measurements do not have a perfectly smooth surface. Air bubbles are present between the test fixture electrodes and the sample according to a non predictable pattern. Similarly, the random presence of direct contact between carbon fibers and the electrodes disregarded this measurement method. Second, an experimental method measuring S-parameter between coplanar transmission lines was tried. The lines were mounted on a small glass die itself put on top of the

composite sheet. This method aims to shortcut the step of deposition of the lines directly on the sample. This method did not give realistic results either. Considering it is still under ongoing study it has not been investigated further in the framework of this work.

Following these failed tests, the most efficient method seemed to be to directly correlate the impact of the composite on coils inductance to its conductivity. Expressions 2.18 and 2.20 express a coil inductance value in term of its geometry, the coil-to-sample spacing, the sample (the composite) thickness and its conductivity. If the coil inductance is known in the surroundings of the composite sheet, an effective conductivity value can then be retrieved. This is the method that was applied. The inductance values of coils were experimentally measured next to composite sheets of known thickness and used for comparison in expressions 2.18 and 2.20. It is actually not possible to inverse those relations and solving for σ is too computer demanding. The value of sigma in the different cases need to be recovered by trial and error on σ values. Coil3 (from Chapter 2) was first simulated at distances 2mm, 5mm and 10mm of two sheets of thickness $300\mu m$ and $900\mu m$ for various conductivities ($300\mu m$ and $900\mu m$ correspond to 1 and 3 plies of cfrp respectively). This was done in order to ensure that the coil was sensitive enough to differentiate conductivities of various orders of magnitude.

Figure 3.8 shows the normalized impedance of the simulation results for coil3. Coil3 was chosen because its was tested to be more sensitive than the other test coils. Again, beside being more intuitive, it is the normalized impedance which is looked at in order to minimize the impact of the difference in the measured and simulated self inductance values. The vertical lines on the x axis represent 5 different conductivity values found in

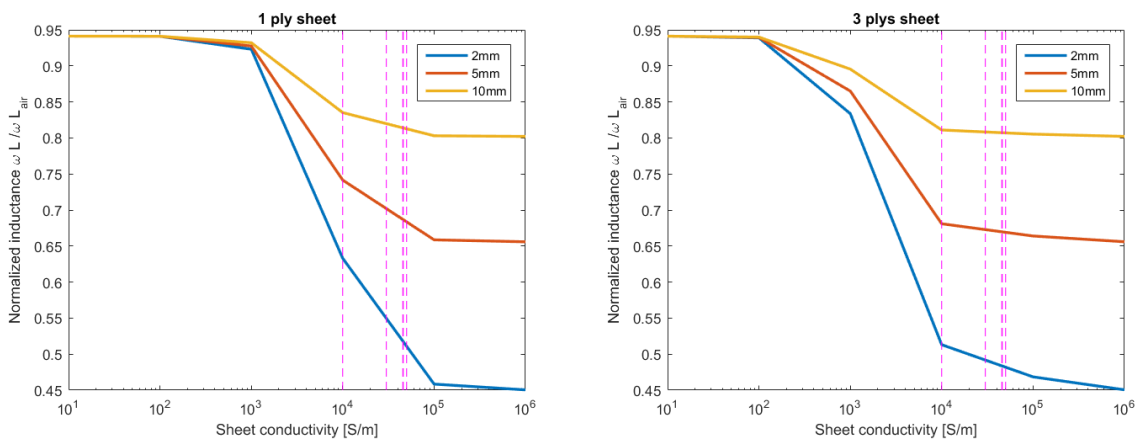


Figure 3.8: Coil 3 sensitivity to conductivity variations

the literature [52] [59] [56] [57]. It is expected that in the conductivity of the prepreg used

in this work will fall in the same range. It is visible that coil3 shows its best sensitivity in that range for the thinner sheet and closer to the sample. Generally, small and large conductivities cannot be differentiated easily while the methods shows good sensitivity for conductivity going from 10^3 S/m to 10^5 S/m. At a distance of 2mm to 1 ply of CFRP, the inductance of coil3 decreases from 93% to 46% over that conductivity range. This allows to say that correlating the hand made coil inductance at 2mm from a 1 ply CFRP sheet will allow to retrieve a fairly accurate conductivity value. Small coil-to-sample spacings could be obtained with great accuracy using the set-up described in chapter 4.

Measurements with 1 ply sheet next coil 3 gave an normalized impedance of 65% (mean of 5 measurements). A sheet of the same thickness and conductivity 10^4 S/m gives a normalized impedance of 63%. From this, it can be concluded that the electrical conductivity of the CFRP used here is slightly less than 10^4 S/m. This conductivity measurement procedure is applicable to any material sufficiently conductive. Depending on the material conductivity a different coil geometry might be necessary in order to make an accurate measurement. This has not been studied further here. From the previous data, it was decided to use $\sigma_{CFRP} = 10^4$ S/m and to see later on if a more accurate coil happened to be necessary.

3.2.3 Permittivity and permeability

Many other model evaluating some kind of average of the properties over the various material constituting different geometric cells are possible and can be more accurate. Considering the literature, it was assumed that 1- and 3- layer models [60] [61] were going to be accurate enough. In the following equations, the $\epsilon_{x,y,z}$ and $\mu_{x,y,z}$ are de relative permittivity and permeability in the different orientations of the composites plies (see Figure 3.9 for xyz axis orientations).

It has been shown that if the thickness of a unhomogeneous medium is very small compared to the wavelength of a transverse wave, the waves perceive the material as having effective properties. These properties are set by the material geometry and the proportion and properties of each of its constituents. It is then possible to write Maxwell equation in term of the zeroth order average fields in an anisotropic homogeneous medium [60] [61]:

$$\begin{aligned}\nabla \times (\bar{E}^0)_{avg} &= -j\omega[\mu^h] \cdot (\bar{H}^0)_{avg} \\ \nabla \times (\bar{H}^0)_{avg} &= j\omega[\epsilon^h] \cdot (\bar{E}^0)_{avg}\end{aligned}$$

where $[\mu^h] = \begin{bmatrix} \mu_x & 0 & 0 \\ 0 & \mu_y & 0 \\ 0 & 0 & \mu_z \end{bmatrix}$ and $[\epsilon^h] = \begin{bmatrix} \epsilon_x & 0 & 0 \\ 0 & \epsilon_y & 0 \\ 0 & 0 & \epsilon_z \end{bmatrix}$ are the homogenized permeability and permittivity.

1-Layer model The effective properties of an inhomogeneous layer of material can be approximated by the following expressions:

$$\epsilon_x^{-1} = (1 - g)\epsilon_{res}^{-1} + g\epsilon_{fib}^{-1}$$

$$\epsilon_y = \epsilon_z(1 - g)\epsilon_{res} + g\epsilon_{fib}$$

$$\mu_x^{-1} = (1 - g)\mu_{res}^{-1} + g\mu_{fib}^{-1}$$

$$\mu_y = \mu_z = (1 - g)\mu_{res} + g\mu_{fib}$$

where g is the called the "fill factor". This factor is simply the volume proportion of carbon fiber in a unit cell of CFRP. It can be expressed in term of the parameter shown in Figure 3.9.

$$g = \frac{(\pi D/2)^2}{LW}$$

The unit cell of the 1-layer model is pictured in Figure 3.9

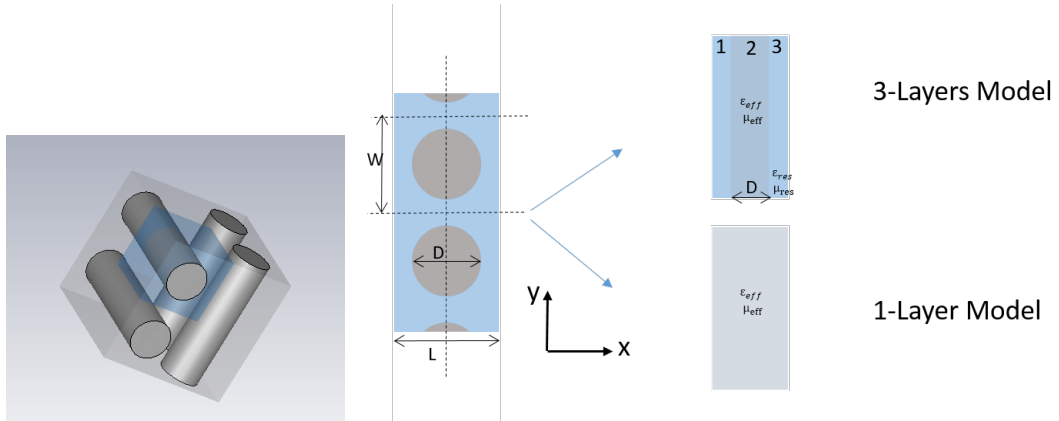


Figure 3.9: Fiber composite unit cell (in blue on the right), 1-Layer and 3-Layers models representations (on the left).

3-Layers model

The 3-layers model is similar to the 1-layer model but the composite ply is first divided into a 3 sheets laminate (Figure 3.9). The external sheets are homogeneous resin sheets while the center sheet is unhomogeneous and needs to be treated as in the 1-layer model. The fill factor for sheet number 2 is

$$g = \frac{(\pi D/2)^2}{DW}$$

The 3-Layers method is mostly applicable to non-prepeg composite. In those cases, the initial thickness of the woven alone is known and an chosen amount of composite is injected. If measuring the thickness of the sheet after curing, it is possible to define the thickness of the 3-layers schematized in Figure 3.9. In the case of prepeg, it is hard to figure it out and it has here been decided to keep up with the 1-layer model.

3.2.4 New material in CST

The expressions developed in the 1- and 3-layers models are supposedly applicable to unidirectional composite plies which is technically not the case of the Hexcel 8552. For the one layer model, it was decided to take $\epsilon_y = \epsilon_z$ as the "in plane" permittivity and ϵ_x as the transverse permittivity.

The issue here was the lack of database on the properties of the constituents of the composite themselves.

First, the information on the permeability of carbon fiber is pretty scarce. Considering the values used in [60] for the fibers permittivity and a fill ratio of 50% given in the datasheet [10], we have considered that $\epsilon_{epoxy} \approx [2; 4]$ and $\epsilon_{carbon} \approx 2$. This gives $\epsilon_y = \epsilon_z \in [2, 3]$ and $\epsilon_x \in [2, 8/3]$.

Second, for the permeability, due to the lack of information on μ_{res} and μ_{fib} , it was decided to use $\mu_x = \mu_y = \mu_z \approx 1,005$ as in [24]

With these equivalent homogeneous material properties, it is possible to define a new material in CST. General standard values for cured prepeg can be taken for other properties like the density or the thermal conductivity since those factors have very little impact on the wireless power transfer. This material is expected to have overall similar effects on the self and mutual inductances of the coils of an inductive link. The simulation will then allow to predict those effects and to take them into account when designing the full system.

It is now necessary to validate the effective properties model. To do so, experimental values need to be compared with simulation results. To easily obtain embedded coil test

samples, a fabrication technique was put in place. The next section describes this custom composite fabrication process.

3.3 Samples fabrication

3.3.1 Prepeg fabrication process

For tests, the HexPly 8552 pre-impregnated carbon fibers composite or *prepeg* was used (see Table 3.4 for detailed characteristic). HexPly 8552 is a high performance tough epoxy matrix for use in primary aerospace structures. A prepeg is made of similar plies of carbon fibers as other fabrication process but where a matrix material, such as epoxy, is already present. This means there is no resin injection during the fabrication process. (The resin that impregnates the raw fibers sheet melt at room temperature so the sheets need to be stored under -18°C in a freezer). In order for the polymerization to happen, the plies need to endure a precise curing cycle under high pressure. Figure 3.11 schematizes the curing cycle and Figure 3.10 shows the advised processing techniques from the Hexply 8552 datasheet [10]. In order to easily obtain test sample, a simple oven process was put in place and a mold fabricated. The mold was fabricated out of two metal sheets of 10cmX10cm. The prepeg plies are positioned between the metal sheets. The later are then clamped with standard bolts (Figure 3.12).

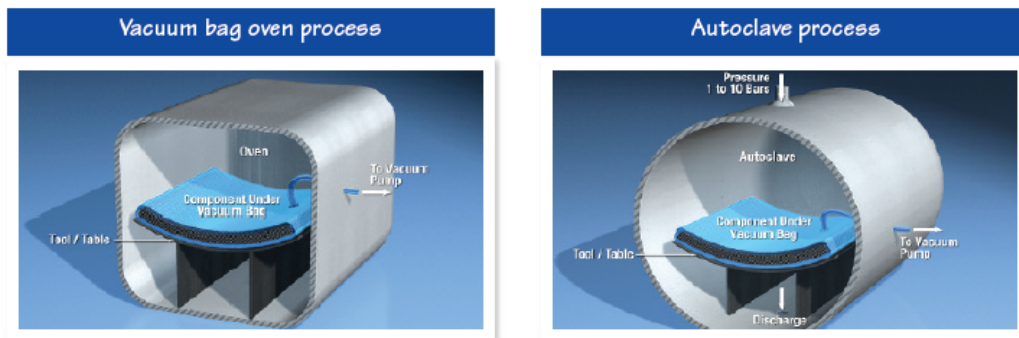


Figure 3.10: Processing techniques advised by HEXCEL [10]

The important things to be aware of when doing such a process is first that the inside of the mold needs to be sprayed with a release agent in order to avoid fibers sticking to the metal. Second, during the cycle, the resin also becomes liquid and needs to be contained in the mold if it is not completely airtight. To do so, a thick layer of silicone was spread around the composite plies stack. After those steps, the mold is ready to put into oven.

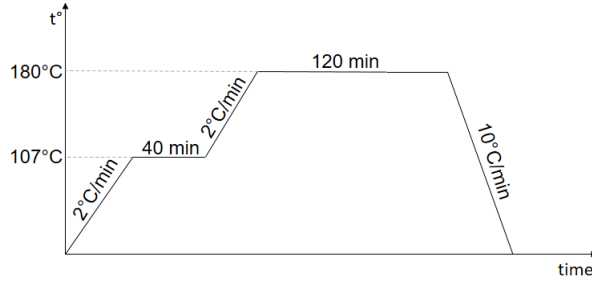


Figure 3.11: Curing cycle used

Here we used an Heareus W6100 Heating and Drying Oven (Figure 3.12) which allows to program precised curing cycle. Before embedding anything between the plies, three test sheets of thickness of 1, 3 and 5 plies were fabricated. This was done on one hand to verify the fabrication process. On the other hand, these sheet were used to derive an approximate electrical conductivity value for this precise process. It is here worth noticing that due to the unorthodox process, the thickness of the 1-ply sheet was larger than the $300\mu m$ expected. Attention to the effective properties so need to be given to it when comparing with simulation result. The 3 and 5 plies sheets were respectively of $960\mu m$ and $1062\mu m$ which mainly agrees with the $300\mu m/\text{per ply}$ standard for Hexply 8552.

Table 3.4: Properties of the Hexply 8552 used.

Physical properties	
Weave	5HS
Nominal cured ply thickness	0.289 mm
Mass	$286\text{ g}/m^3$
Fiber type	AS4 3K
Matrix type	Epoxy
Weight ratio	50:50
Tow cross section	0.12 mm^2
Filament diameter	$7.1\mu m$
Fiber electrical resistivity	$1.7 \times 10^{-3}\text{ }\Omega\text{-cm}$

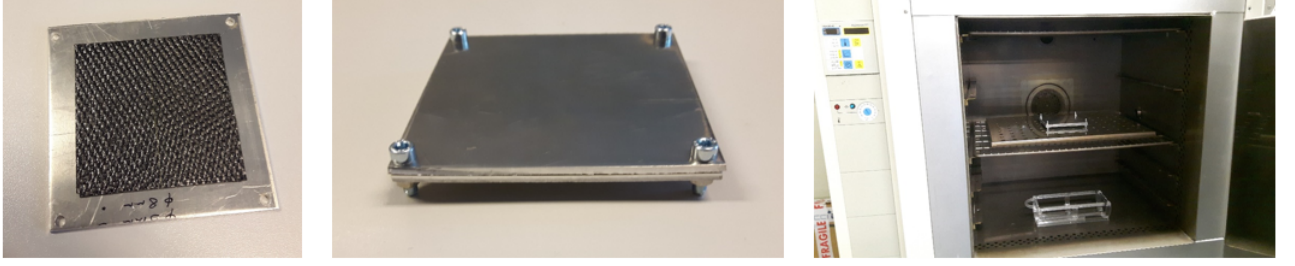


Figure 3.12: Stack of 3 square composite plies on the bottom part of the mold (left), mold closed (center), mold inside the Heraus oven (right).

3.4 Simulation and effective properties model validation

In order to verify that CFRP can be electrically modeled as a homogeneous material of effective conductivity, permeability and permittivity, three test coils were designed in order to compare simulation and experimental results. Table 3.5 summarizes their geometric properties. All the measurements and simulations here are done at the operating frequency of 10 MHz.

Table 3.5: Properties of the validation coils

Physical properties	C1	C2	C3
Outer dimensions	15mm	30 mm	20 mm
Inner dimensions	7.4 mm	12 mm	7.2 mm
Trace thickness	0.035 mm	0.035 mm	0.035 mm
Trace width	0.2 mm	1 mm	0.4 mm
Turn spacing	0.2 mm	1 mm	0.8 mm
Number of turns	10	5	6

The self inductance value of single layer planar spiral coil were first analytically calculated. Here the *modified Wheeler formula* was used. For a square inductor of inner diameter d_{in} and outer diameter d_{ou} it is expressed as

$$L_{mw} = 2.34\mu_0 \frac{n^2 d_{avg}}{1 + 2.75\rho}$$

where $\rho = (d_{out} - d_{in})/(d_{out} + d_{in})$ is called the *fill ratio* and $d_{avg} = (d_{in} + d_{out})/2$

Detailed information on the modified Wheeler formula as well as alternative calculation

methods called *Current sheet approximation* and *Data fitted monomial expression* will be found in [62].

The coils were fabricated and modeled in CST Microwave studio in order to compare measurement and simulation results. First, this was done in air (Figure 3.15). Then, a second set of identical PCB's were embedded between CFRP plies (Figure 3.16) according to the pattern shown on Figure 3.13. Three plies were stacked under and above the PCB's. Windows were cut away in four additional layers which were laid down around the inclusion. The fabrication process and curing cycles are the same as previously described and details in the insertion method will be given in the next chapter. Figure 3.14 also shows screenshots of the 3D-simulation designs.

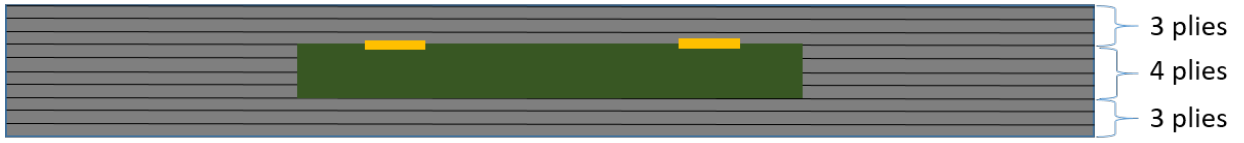


Figure 3.13: 2D schematic of the embedded validation coils.

The theoretical values as well as the measurements and simulation results in air can be found in Table 3.6. Considering the mismatch $(L_{LCR} - L_{CST})/L_{LCR}$ between the experimental and simulation results in air, better accuracy between the simulation and experiments in CFRP cannot be expected. Because of this, simulation results intervals were defined in which simulation results of the validations coils self-inductances in CFRP are assumed valid.

The target interval are based on the mismatch $(L_{LCR} - L_{CST})/L_{LCR}$. The upper limit of the intervals is simply the LCR measurement in CFRP. Considering that in air the simulation results were 11.9%, 7.52% and 9.43% lower than the measured values, the lower limit of the target interval were also set at 11.9%, 7.52% and 9.43% lower than the measured values in CFRP.

In CST, the composite material was modeled as a uniform material. Each ply was also given the same thickness of $300\mu\text{m}$. Figure 3.14 shows screenshots of the 3D designs. The simulation were first conducted for a composite electrical conductivity of 10^4 S/m (previously evaluated in Section 3.2.2 with the experimental/analytic model comparison). We can see in Table 3.7 that for this conductivity, none of the values are within the target intervals. To obtain the right value, the effective conductivity was iterated until all coils self-inductances values were in the intervals. All the result are found in Table 3.7.

One can see that for an electrical conductivity of around 7500 S/m, the correlation be-

Table 3.6: Calculated, simulated and measured values of the validation coils self-inductance.

	C1	C2	C3
Analytic value in air	1.704 μH	0.709 μH	0.627 μH
CST simulation in air	1.577 μH	0.701 μH	0.615 μH
LCR measurement in air	1.790 μH	0.758 μH	0.679 μH
$(L_{LCR} - L_{CST})/L_{LCR}$	11.90%	7.52%	9.43%
LCR measurement in CFRP	1.008 μH	0.228 μH	0.321 μH

Table 3.7: Iteration on the conductivity to reach the targeted self-inductance values intervals.

Physical properties	C1 [μH]	C2 [μH]	C3 [μH]
Target	[0.888, 1.008]	[0.211, 0.228]	[0.291, 0.321]
$\sigma = 10000 \text{ S/m}$	0.851 ✗	0.197 ✗	0.253 ✗
$\sigma = 9000 \text{ S/m}$	0.871 ✗	0.205 ✗	0.271 ✗
$\sigma = 8000 \text{ S/m}$	0.895 ✓	0.215 ✓	0.287 ✗
$\sigma = \mathbf{7500 \text{ S/m}}$	0.905 ✓	0.224 ✓	0.301 ✓
$\sigma = 7000 \text{ S/m}$	0.976 ✓	0.231 ✗	0.312 ✓
$\sigma = 6000 \text{ S/m}$	1.051 ✗	0.254 ✗	0.332 ✗

tween the three validation coils simulation and LCR measurements is really good. Reaching values within the target intervals for the three coils at the same time is the best accuracy that could be expected here. Being in the intervals means that the accuracy of the simulation in CFRP is basically as accurate as the simulation in air (considering the intervals are built on the mismatch in air).

It is expected that if someone was to embedded another coil with the same pattern, such simulations will be able to predict the coil self-inductance in CFRP as accurately as to predict it in air.

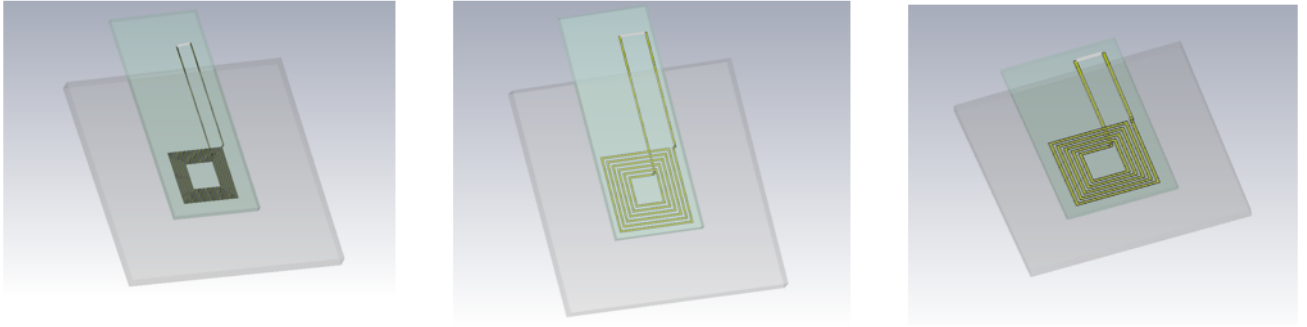


Figure 3.14: 3D design of the three validation coils.

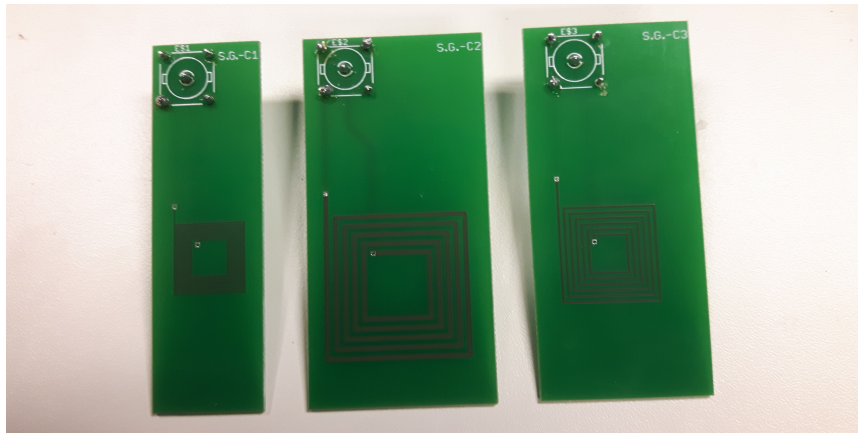


Figure 3.15: Pictures of the three validation coils.

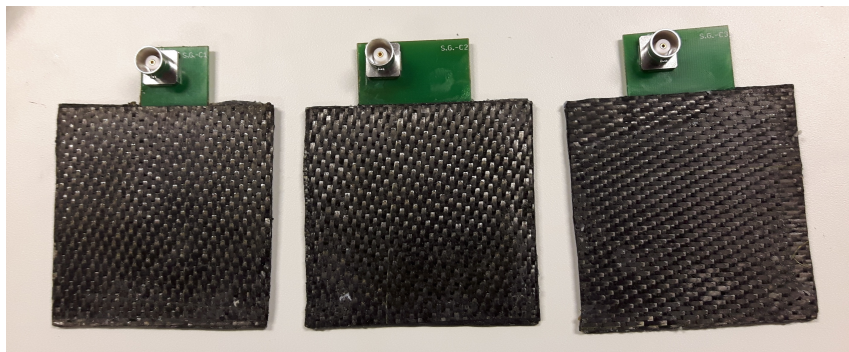


Figure 3.16: Picture of the three embedded validation coils.

3.5 Conclusion

There are many things to conclude here. On one hand a complete design flow was put in place and the following steps were studied through the chapter:

- CFRP was accurately approximated by an homogeneous medium using effective properties. The approximation was proved to be able to predict an embedded coil self-inductance in CFRP with an accuracy similar to the accuracy in air.
- A system of inductively coupled coil were 3D designed and simulated with the receiver embedded in this homogeneous medium simulated.
- A technique to extract the equivalent lumped parameters from the 3D simulation results was put in place.

The complexity of measuring the electrical conductivity of carbon fiber composite due to frequency dependency and variation with fabrication process was also outlined. Finally, a simple CFRP fabrication process was put in place. Using this process, three test coils were inserted into rectangular composite sheets and LCR measurements results were successfully matched with CST simulations using the homogeneous medium approximation.

Chapter 4

Wireless power transmission

The goal of this chapter is to assess the feasibility of an inductive link meeting the specifications laid down in Chapter 1. To do so it was decided to choose a receiving coil having the proper dimensions (10mmX10mmX0.3mm), to embed it into carbon fiber composite and to attempt to transmit a minimum of $200\mu\text{W}$ to it.

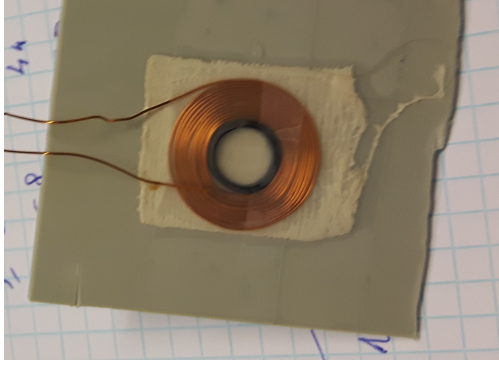
First and before selecting the coils, preliminary measurements were conducted in order to show the positive effect of a ferrite sheet.. Second, The transmitter and receiver coils are selected on those measurements results. Their self-inductance and quality factor are measured at various distances from the carbon fibers composite sheet fabricated in chapter 2. Third, receiver coil were embedded into CFRP. Last, the inductive link between reader and receiver is studied. The coupling coefficient is measured and compared with two readers. The AC received power is measured and DC power values also obtained using an ACDC converter.

4.1 Preliminary process

4.1.1 Preliminary test

In order to evaluate the impact of carbon fiber composite on coils of dimensions following the specifications, preliminary tests were conducted. A small pancake coil was hand crafted and studied within the proximity of the composite. Figure 4.1 shows a picture and the properties of the coil.

The coil self inductance and Q factor were measured with an LCR at distances 0 to 20 mm from a 5-ply thick carbon fiber composite sheet. Figure 4.2 shows plots of the result. Both properties are strongly affected by the composite. The self inductance decreases down to 30% of its value and the Quality factor from over 70 to under 10 for



Physical properties	
Inner diameter	22mm
Outer diameter	11mm
Wire diameter	0.3 mm
Number of turns	12
Material type	Copper

Figure 4.1: Picture and properties of the preliminary test coil.

sub 5 mm coil-to-sample spacing. While the self inductance decrease could be taken into account in the design of the matching network, the impact on Q factor cannot be compensated. When directly put on CFRP, this coil show a Q-factor of 4. Looking back in Chapter 2 at the maximum power transfer efficiency and taking an optimistic coupling factor of 0.01¹ and transmitter quality factor of 50 leads to $k^2 Q_1 Q_2 = 0.01^2 * 4 * 50 = 0.04$. This is represent a maximum power transfer efficiency of

$$\left(\frac{P_{out}}{P_{in}}\right)_{max} = \frac{\sqrt{1 + k^2 Q_1 Q_2} - 1}{\sqrt{1 + k^2 Q_1 Q_2} + 1} = 0.005$$

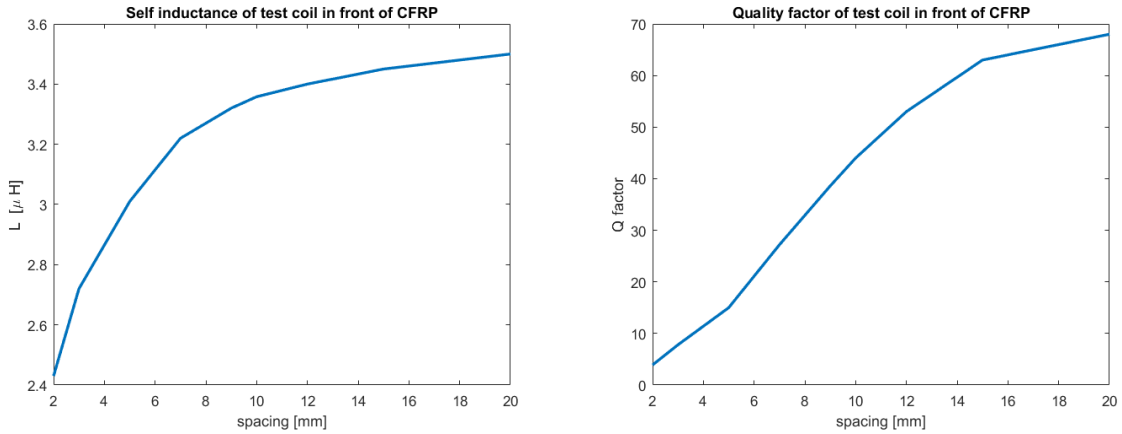


Figure 4.2: Behavior of the preliminary test coil close to a 5 plies CFRP sheet.

¹This value is based on the result obtained with analytic case study in Chapter 1

4.1.2 Ferrite shielding

Shielding the coils with ferrite materials can improve the efficiency of the power transfer. Strategic placement of ferrite sheets can reduce the eddy current and so the self and mutual inductances decrease caused by CFRP ply. Ferrite are high permeability materials which are defined by their complex permeability

$$\mu_r = \mu' - j\mu''$$

where μ' is the ideal part and μ'' represent the losses. It is important to remember that both vary with frequency. Depending on the application, high μ' or high μ'' ferrite will be used. In our case, we want to maximise the magnetic flux control while minimizing the losses so a high μ' is preferred. When selecting the right ferrite, strong care should be given at frequency dependency of the selected material. The characteristic of many ferrite sheets advertised for shielding start decreasing for frequencies under 10MHz. Figure 4.3 shows a plot of the real and complex permeabilities of the ferrite used. These sheets are $90\mu m$ thick so they fulfill the thickness requirement raised in chapter 1.

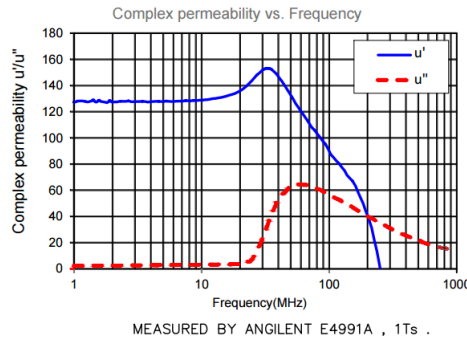


Figure 4.3: Permeability of the ferrite MHL6060-300, reproduced from [63].

Similar tests with the preliminary coil were conducted in order to quantify the impact of the ferrite sheet on the self-inductance and quality factor. Figure 4.4 shows the results in relation with the previous plots. Both curves follow similar variations. The inductance behavior does not really matter in this case. The inductance is stronger in presence of the CFRP but also in air due to the ferrite. What is worth looking at is the quality factor. The curve has been shifted slightly higher and for a small spacing of 2mm the factor goes from 4 to 11. This can be seen as irrelevant considering both values are still really small but has a significant impact on the output power. Going again back to the maximum

power transfer, this gives

$$\left(\frac{P_{out}}{P_{in}}\right)_{max} = \frac{\sqrt{1 + k^2 Q_1 Q_2} - 1}{\sqrt{1 + k^2 Q_1 Q_2} + 1} = 0.0134$$

This means that the maximum power transfer efficiency has more than doubled with the use of the ferrite sheet.

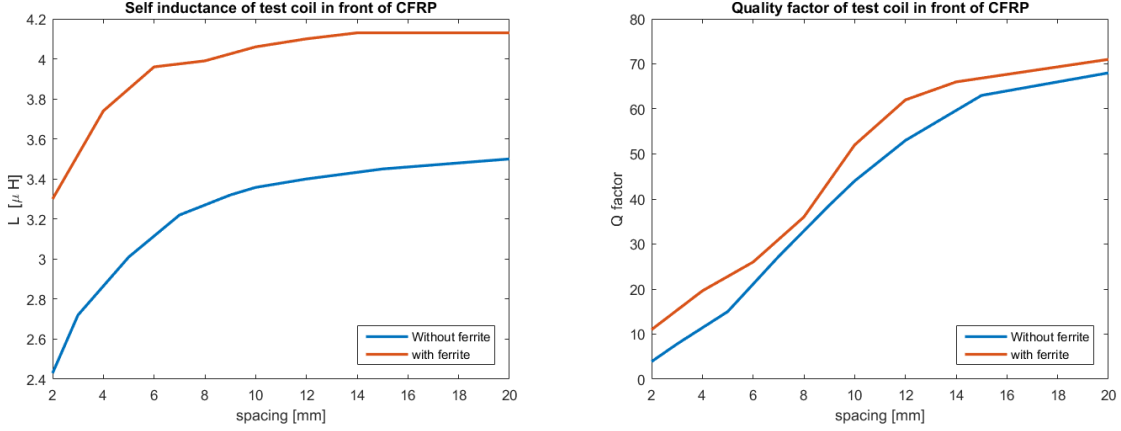


Figure 4.4: Comparison of the inductance and quality factor variations with and without the ferrite sheet.

Figure 4.5: Comparison of the Q factor variations with and without the ferrite sheet.

Table 4.1: Properties of the coil in and out of CFRP at 10MHz.

Antenna	L [μH]	Q
Without ferrite	3.52	70
With ferrite	4.53	79
Without ferrite between CFRP	2.2	3.4
With ferrite between CFRP	3	11

Components selection

Receiver The preliminary test results led to conclude that ferrite backed small rectangular coplanar coil would give good results. Any component selected also must be

able to endure a full curing cycle. Components-of-the-shelf are usually designed to resist maximum temperature of around 90°C and it was not possible to find high temperature components respecting the specifications. Considering that coils are very simple element made of materials that are suppose to resist higher temperature those 90 °C, the choice was made to first try standart component anyway at higher temperature than advised. Figure 4.6 shows a picture of the receiver coil selected and of the coil soldered to a U.fl cable (U. fl is small coaxial cable type.). The selected coil is an ultra thin flexible antenna structure (220μ) from manufacturer Abracon ². The full size of the die is 15mmX10mm which is larger than the specifications but the coil size is about 7x12mm so it is close enough to be extrapolated as a sub 10mmX10mm coil. In order to make sure later on to differentiate the composite effect from component failures due to the curing cycle, samples were initially analyzed before and after they endured a full cycle.

Antennas impedance and the U.fl cable were measured at various frequencies after curing to ensure their integrity. No variations were observed and it was concluded they could handle the curing cycle depicted in Figure 3.11. ³

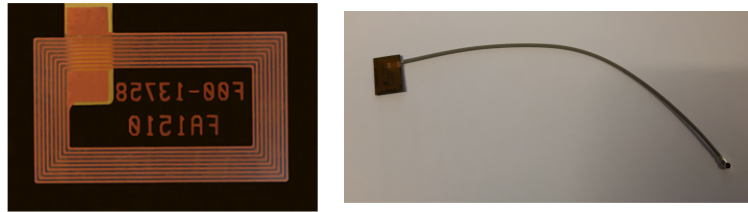


Figure 4.6: Receiver coil ANFCA A02 (left), soldered to an U.fl cable (right).

Readers Two primary coils were selected for the later power transfer testing. Figure 4.7 displays pictures of both readers and Table ?? their properties.

Insertion of antennas

The antennas were first soldered to U.fl coaxial cables. Electrical insulation around the connection (between the antenna and cable) is necessary considering the carbon fibers are conductive and might create short circuits. To do so, tests were made using an silicon tropicalisation varnish (Electrolube DCA silicon based conformal coating (SCC3)). The

²Component reference ANFCA-1510-A02

³A more thorough study of the cable before and after the cycle deriving the transmission line parameters from S parameters measured with a vector network analyzer was originally intended. Unfortunately, the equipment was in maintenance during the realization of this work and such tests were dropped.

Table 4.2: Properties of the reader coil

Physical properties	Reader 1	Reader 2
Self-inductance	$3.65 \mu H$	$5.3 \mu H$
Q-factor	255	100
Outer dimensions	3 cm	$\varnothing$ 43 mm
Inner dimensions	1.5 cm	$\varnothing$ 25 mm
wire radius	$300 \mu m$	2.4 mm
Number of turns	10	8
Manufacturer	Hand made	Würth Elektronik



Figure 4.7: Reader 1 (left), reader 2 (right).

varnish is firstly meant to prevent humidity but also offers electrical insulation. A $24 \mu m$ thick layer was sprayed on three test antenna and dried according to the manufacturer advices for intense thermal cycle application (12h drying at room temperature and 24h at 90 celcius degrees). After that, the insulation was tested using the logic function of a regular multimeter. The insulation was not efficient. A possible explanation was a nonuniform enough layer on the connection but a quick check with UV light rejected that. A second layer and drying cycle was done and the results were the same. As an alternative, a $25 \mu m$ thick Kapton film was applied over the antenna/cable junction (Figure 4.8). As for the validation coils of Chapter 3, in order to minimize the thickness difference within the composite stack, a window is cut off in the central plies. (Figure 4.9)

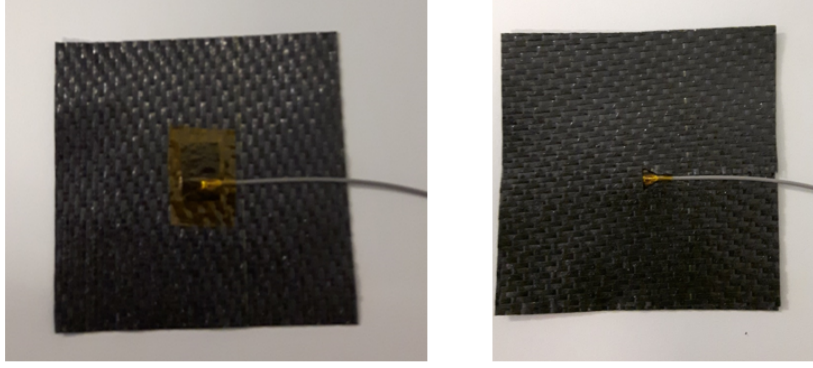


Figure 4.8: Coil and cable stuck on a cfrp plies before curing (left). Same coil and cable but covered by another CFRP ply (right).

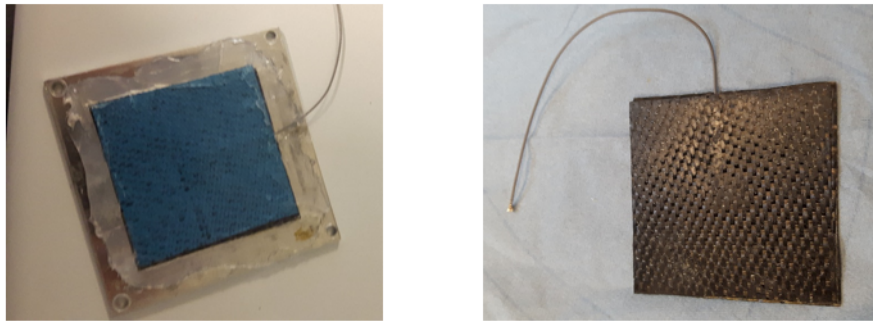


Figure 4.9: Open mold with the plies stack and silicon before curing (left). Composite sheet with embedded coil after curing (right).

4.2 Coils characterization

4.2.1 Set-up

Figure 4.10 shows the set-up used to measure the self inductances and quality factor of the coils in proximity of the composite plies. To set precise spacings between the coils and the composite, a *Melles griot* rack was used. The rack is made of metal and was affecting the results. To solve that problem, additional elements were fabricated in order to place the coils at a sufficient height above the rack at which the rack itself had no longer any measurable effect. Those elements were 3D printed and are 20cm height (white "L" shape pieces in Figure 4.10). It was verified that they also did not affect the power transmission in any way. The data were obtained with an HP 4275A LCR meter by directly connecting it to the coils and using it in L-R series mode at 10MHz.

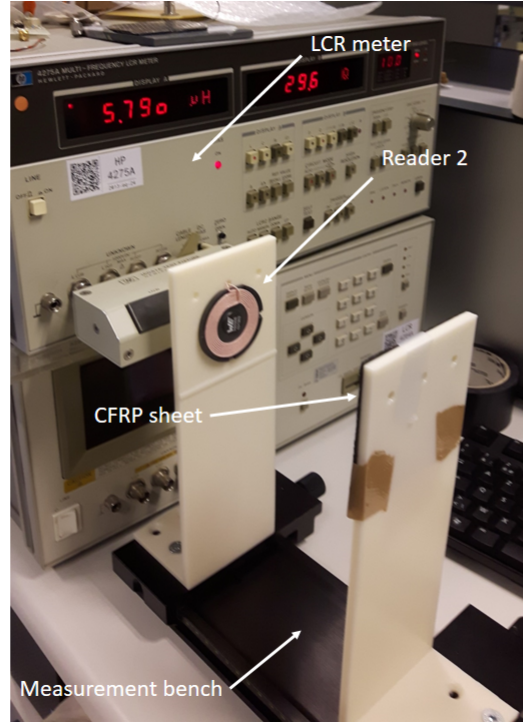


Figure 4.10: Measurement set-up for the self-inductances and quality factors.

4.2.2 Transmitter coils

Figure 4.11 shows the self inductance variation of the two reader coils next to a 3-ply sheet (on the left). One can see that reader 1 is more significantly impacted by the carbon composite than reader two. This phenomena is also extremely visible on the second plot which shows the variation of the Q factor of the reader in the surroundings of the conductive sheet. At a distance of 1.5 cm, the composite has virtually no impact on reader 1 while the impact on reader 2 is still a Q factor divided by two and a 10% decrease in self-inductance compared to its properties without CFRP.

4.2.3 Receiver coils

Table 4.3 shows the self-inductance and the Quality factor of the coil ANFCA A02 in air and inside a composite sheets (1 and 2 plies deep). The self-inductance varies very little in and out of the composite which is mostly due to the positive impact of the backed ferrite and to the coil small dimensions. At the opposite, the quality factor is strongly affected by the CFRP. While the self inductance drops by less than 5%, the Q-factor is reduced by over 75%.

Similar observations as for the transmitter coils were also made for the receiver coil

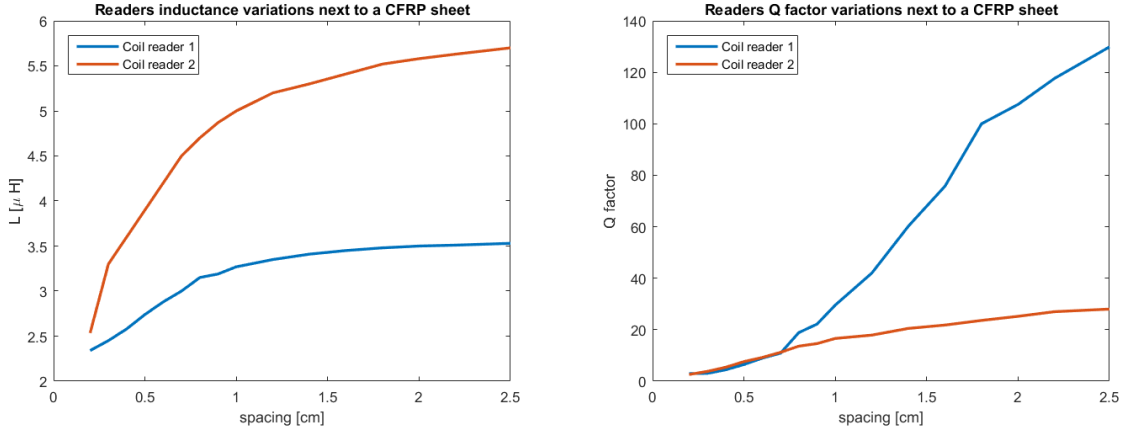


Figure 4.11: Self-inductance and quality factor variations of the two reader coils in the proximity of a 3 plies CFRP sheet.

Table 4.3: Properties of the receiver coil in and out of CFRP at 10MHz.

Antenna	L	Q
ANFCA A02 in air	2.131 μH	20.7
ANFCA A02 1 ply deep	2.071 μH	5.2
ANFCA A02 2 ply deep	1.984 μH	3.7

ANFCA A02 and for ANFCA A01 (a similar coil from the same manufacturer but bigger, the reference manufacturer is ANFCA-4545-A01). The results of the measurements of self inductance and Quality factor are displayed on Figure 4.12. The impact on A01 is significantly stronger than on A01 which is due to its larger dimensions.

4.3 Inductive link characterization

4.3.1 Set-up

Here, the same measurement bench as for the coils characterization was used. A picture of the measurement set-up can be found in Figure 4.13. The readers were stimulated with an Agilent 33250A wave generator. The Coupling coefficient and DC power were measured on a Infiniium MSO8104A oscilloscope while the AC power measurements were conducted with a Anritsu MS2713E Compact Handheld Spectrum Analyzer.

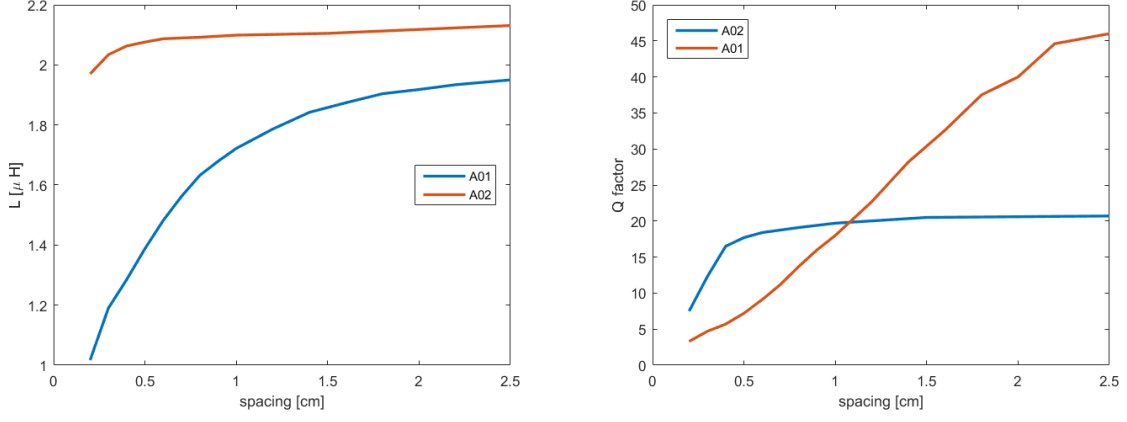


Figure 4.12: Self-inductance and quality factors of ANFCA A02 and A01 next to a 3 plies thick CFRP.

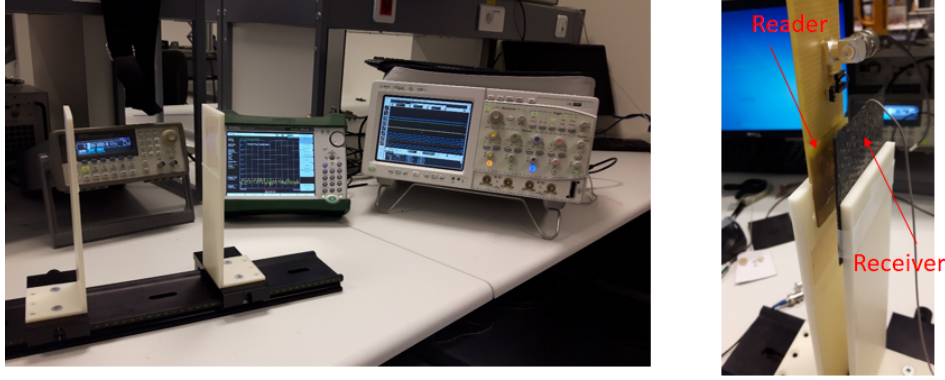


Figure 4.13: Set-up used for the coupling coefficient and power measurements

4.3.2 Coupling coefficient

Figures 4.14 and 4.15 display plots of the coupling coefficient in function of the transmitter and reader spacing. To obtain these results, the transmitter coils were stimulated with sinusoidal voltages of 1, 1.5 and $2.5V_{rms}$ and the open voltages at the receiver side measured. The coupling coefficient is then retrieved through

$$k = \frac{M}{\sqrt{(L_1 L_2)}} = \sqrt{\frac{L_1}{L_2}} \frac{V_2}{V_1}$$

Although the open voltages measured at the secondary are really small, these values can be assumed correct. The measurements were conducted for multiple input voltages and the results correlated. The variations of the readers self-inductances in proximity

of the composite sheet were taken into account as well as the small variations of the transmitter input voltages due to those self-inductance variations.

The inherent measurement offset was also subtracted (receiver voltage measured when 0V applied at the transmitter). Again, the measurement set-up was built precisely for these measurement allowed for an accurate setting of small spacing between the coils.

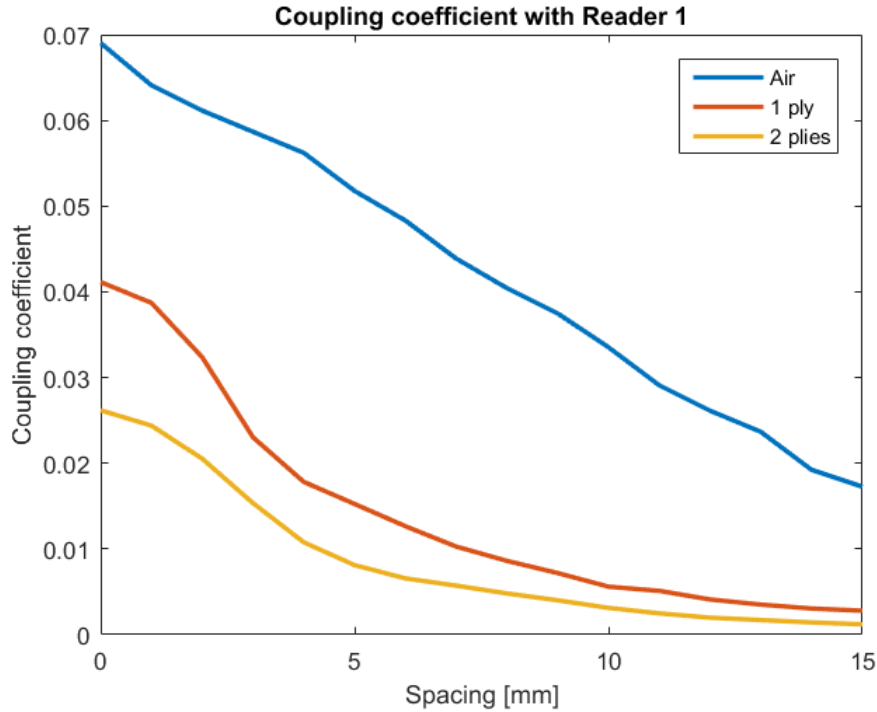


Figure 4.14: Coupling coefficient between reader 1 and the coil ANFCA A02 in air and embedded under 1 and 2 plies of CFRP.

4.3.3 AC power measurements

Figure 4.16, 4.17, 4.19 and 4.20 show the AC transmitted power measurements results for both readers and receiving coils in air and embedded under 1 and 2 plies of CFRP. The measurements were done for transmitter to receiver coil every millimeter for spacing going from 1 to 25 mm. The transmitter were excited through a $2.5V_{rms}$ input voltage. The input power was obtained measuring the potential over a 50Ω series resistor. To do so, the transmitter coil were mounted on a PCB and a resistor and connectors soldered to it. Those measurements were done with the Infiniium MSO8104A oscilloscope and lead to 30.1 mW input power for reader 1 and 26.8 mW for reader 2. All the received power values were obtained using an Anritsu MS2713E spectrum analyzer which was directly connected

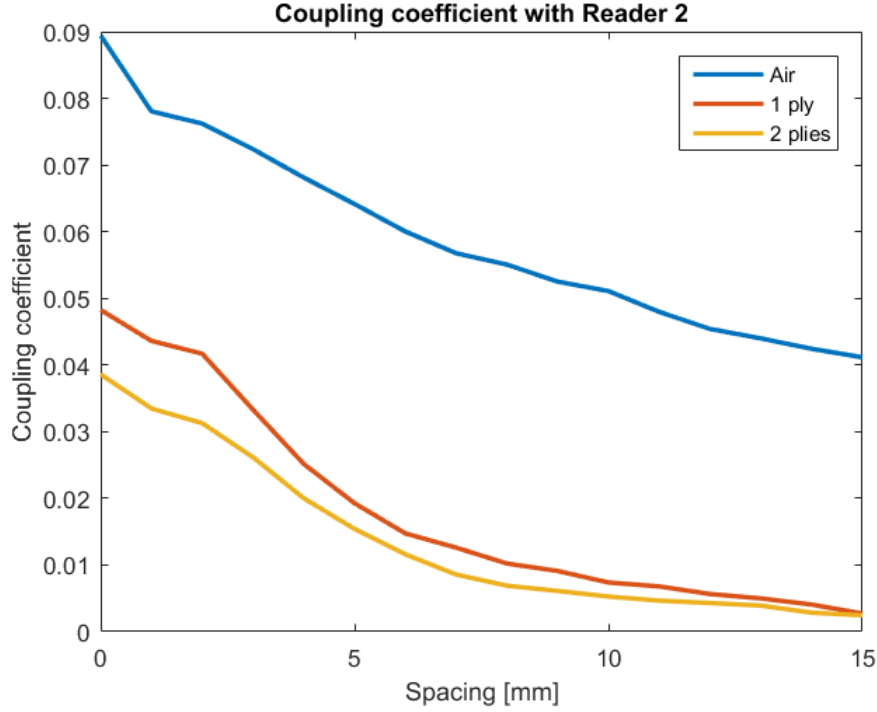


Figure 4.15: Coupling coefficient between reader 2 and the coil ANFCA A02 in air and embedded under 1 and 2 plies of CFRP.

to the receiver coils (no matching circuit). First, it can be observed that a larger power transfer than originally expected was achieved (the initial analytic case study had led to expect a power efficiency of less than 10^{-2} %). For very small distances (1-2mm), $230\mu\text{W}$ were transferred 1 ply deep by reader 1 and with a 0.6% efficiency. Even counting on a low AC/DC conversion efficiency of 25%, still over $50\mu\text{W}$ of DC power could be transfer

The important thing to notice here is the power transfer efficiency in CFRP compared to the efficiency in air. Figure 4.18 shows the power transfer in CFRP normalized by the power transferred in air. For distances of 1 mm, the power transferred is mainly divided by three when the receiver coil is embedded at 1 ply deep. Putting it differently, in this case transferring a certain amount of power for reader 1 to that receiver for a spacing of 1 mm is as efficient as transferring the same amount of power to that receiver outside of CFRP for a spacing of 5 mm (Figure 4.16).

4.3.4 DC power measurements

Power measurements were also done connecting a AC/DC converter followed by a resistive load to the receiving coil. The converter used is described in [64] and its circuit and picture

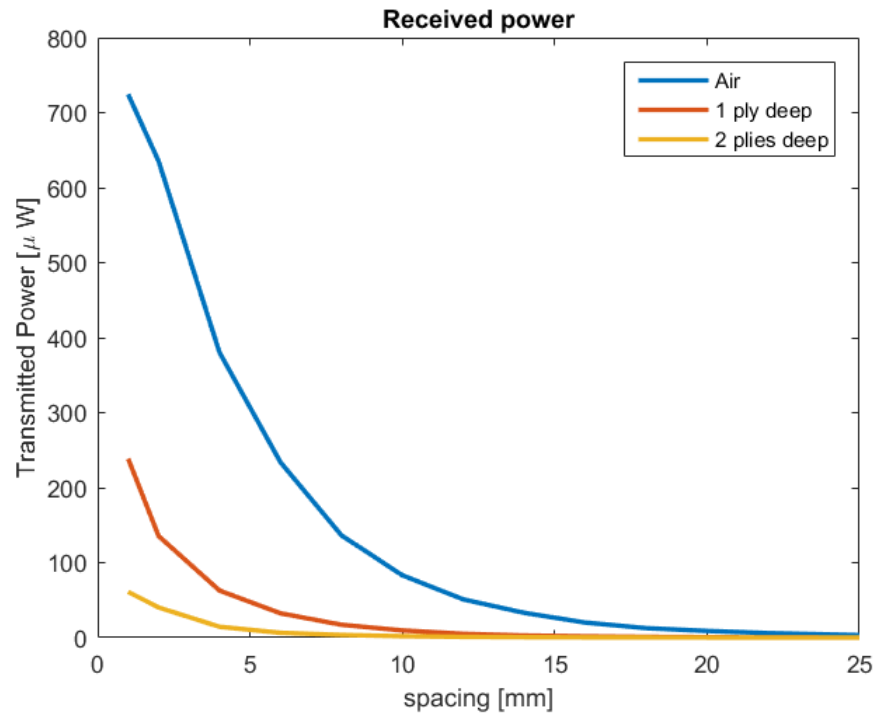


Figure 4.16: AC Output power obtained with reader 1.

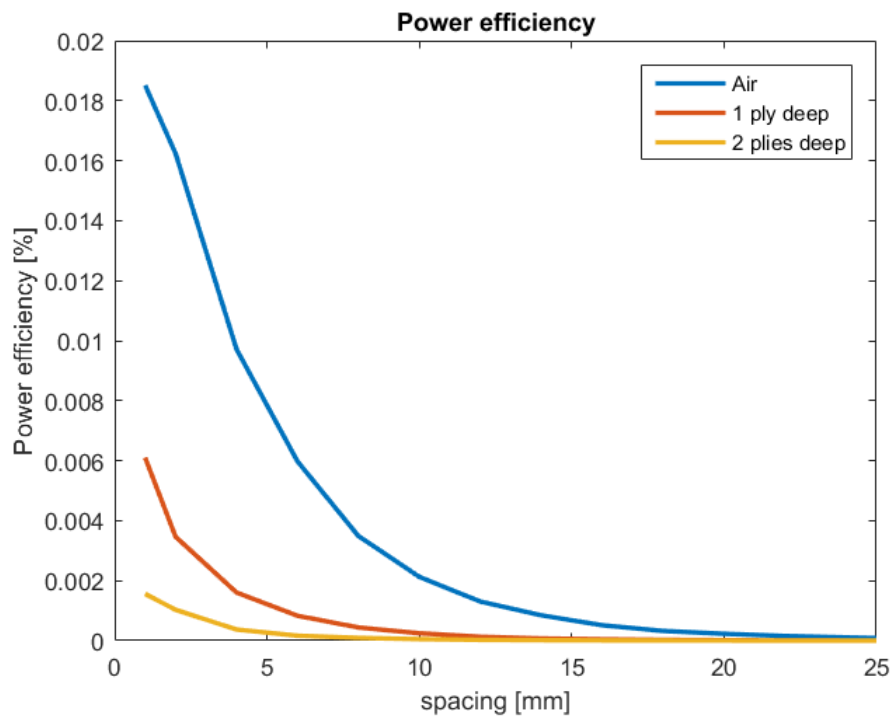


Figure 4.17: AC power transfer efficiency obtained with reader 1.

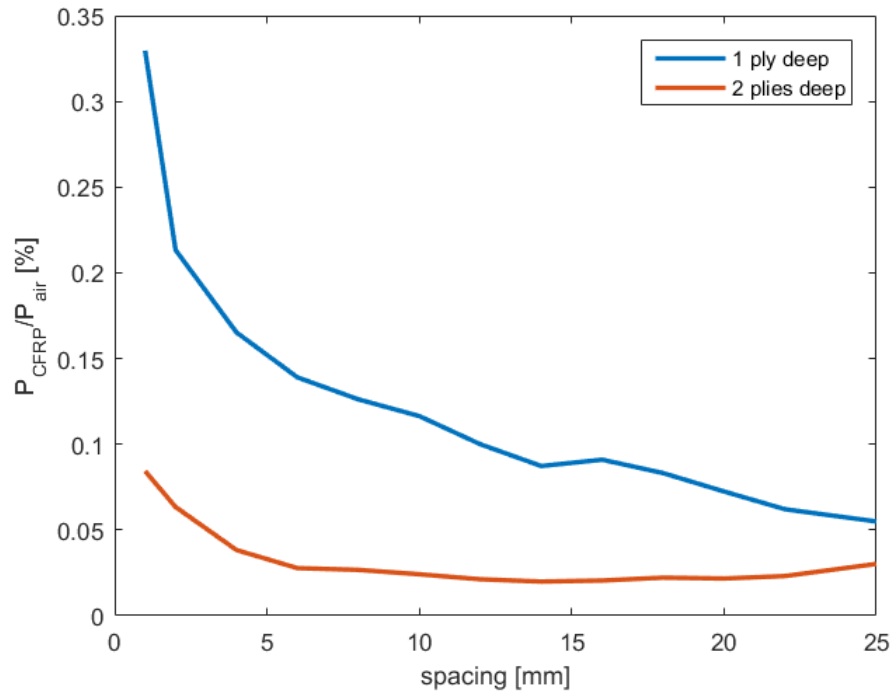


Figure 4.18: Power transferred in CFRP scaled by the power transferred in air.

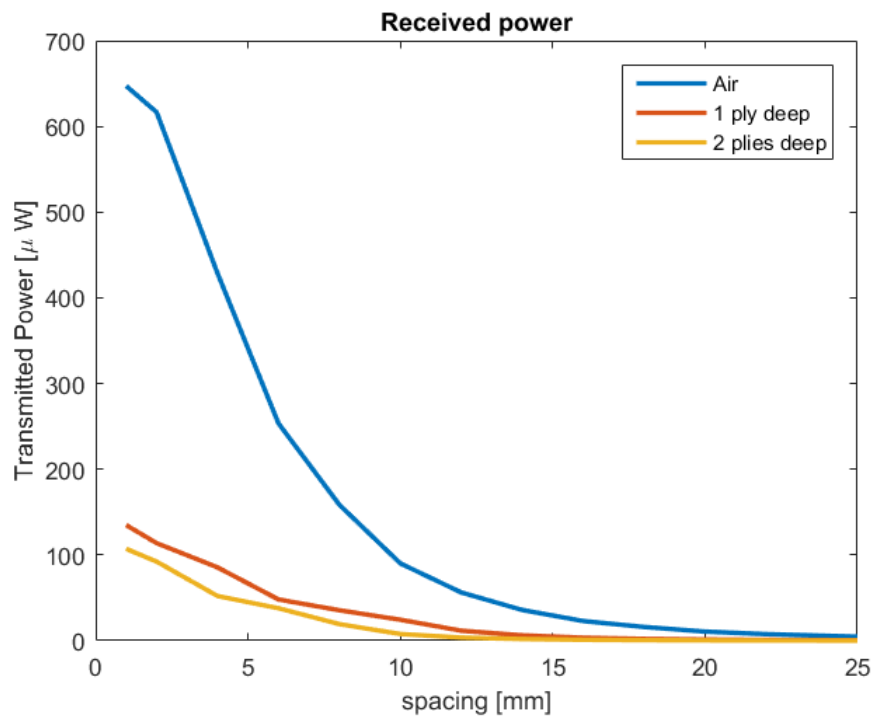


Figure 4.19: AC Output power obtained with reader 2.

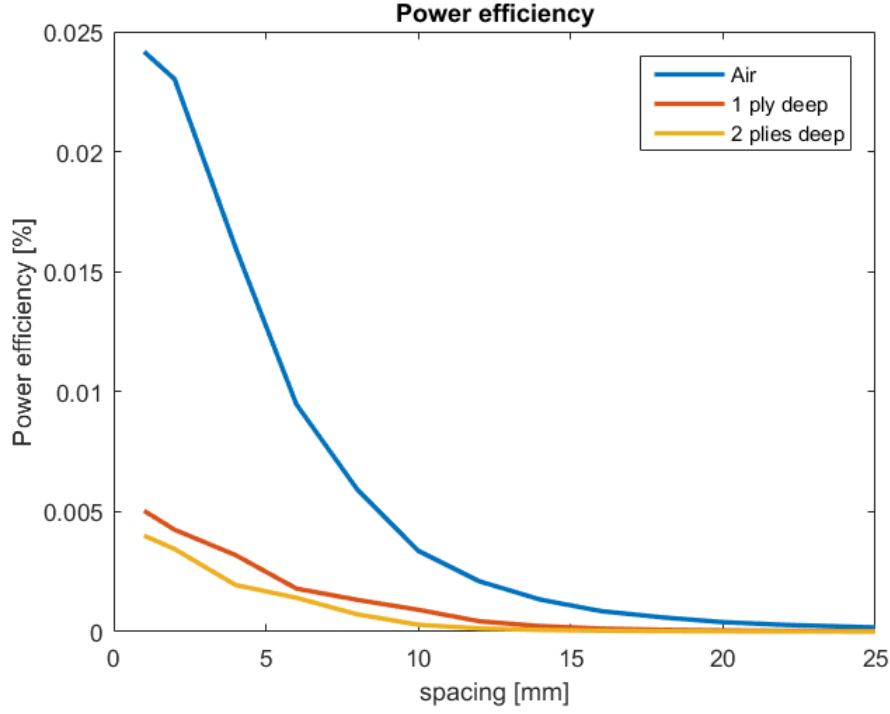


Figure 4.20: AC power transfer efficiency obtained with reader 2.

can be found in Figure 4.21. It is precisely design for this kind of low power application and seem like a good option to make DC power. measurements.

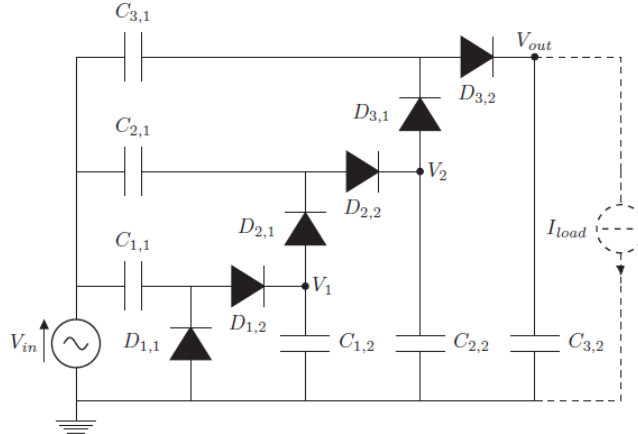


Figure 4.21: AC/DC converter circuit and picture.

Figures 4.22 and 4.23 shows the output power measured on three different resistive load. The measurement were made taking the voltage across the load with the same oscilloscope as before. The first thing to mention is that the converter features an on-chip

ESD protection circuit which activates for voltages higher than 2.5V or under -7 Volts. The measurements here were conducted using a symmetric signal. The flat zones on the plot of Figure 4.22 are due to an activation of those safety diodes due to a received voltage going under -0.7 Volts. For those zones, it is so expected that a higher output power would be obtained if using a asymmetric input signal.

Another thing to mention is that the efficiency of the converter cannot be assessed by comparing the AC power from the previous section with the power measured here. The input impedance of the spectrum analyzer used before and the input impedance of the converter are most likely different. Looking back at the circuit model of chapter 1, this means that the reflected loads are different and so the power originally transferred.

The important thing here is that if using an input power of 1W, the specification of 200μ will very largely be met. Considering the conversion efficiency obtained here give around 0.1 to 0.2% total power transfer efficiency. For an input power of 1W this corresponds to 1 to mW so basically 10 more power than originally desired.

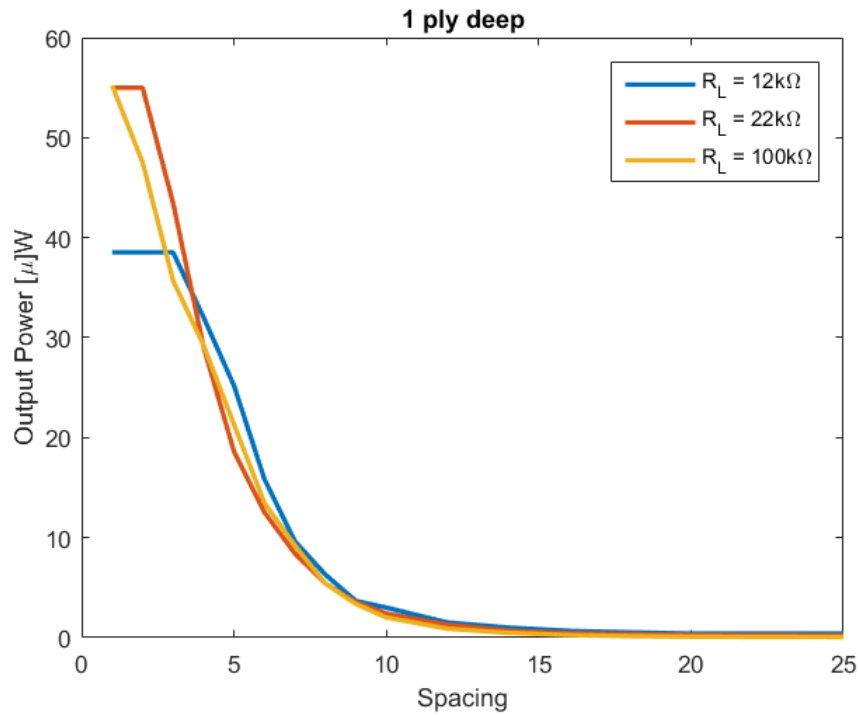


Figure 4.22: DC output power obtained for the receiver coil embedded 1 ply deep, with reader 2 and for an input power of 26.8 mW.

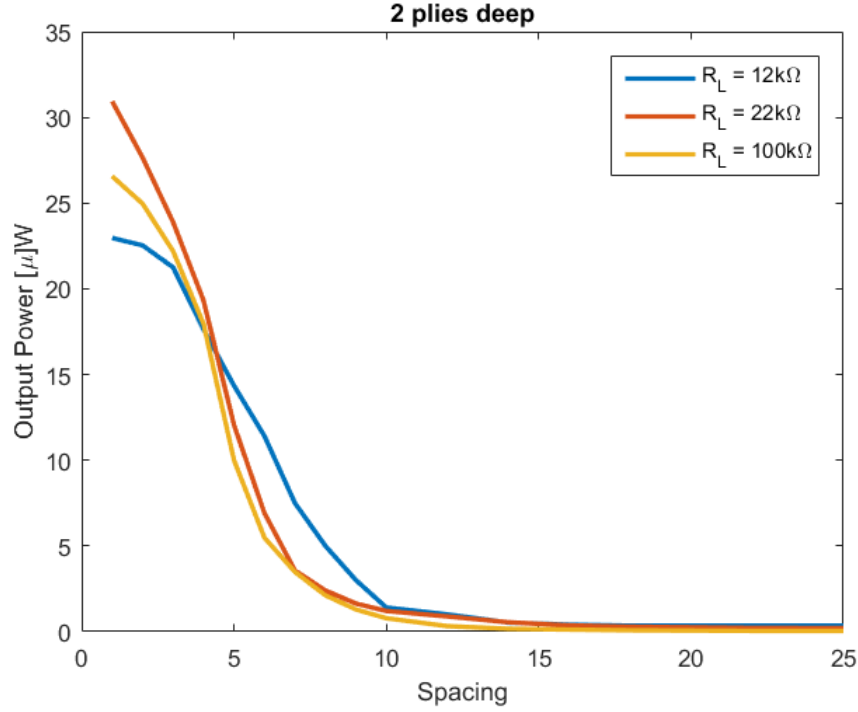


Figure 4.23: DC output power obtained for the receiver coil embedded 2 ply deep, with reader 2 and for an input power of 26.8 mW.

4.3.5 Matching PCB

In order to see the improvement obtained when matching and mismatching the receiver, a simple PBC was fabricated in order to be able to connect various components in series and in parallel with the receiver coil. Figure 4.24 shows a picture of the device. For flexibility, a series and a parallel IC Socket were used as well as a fine tuning parallel capacitor. This capacitor covers a range going from 7 to 50 pF⁴ and can so be set to have a virtually nonexistent impact on the receiver (setting it to the minimum value 7pF). The two SMA connectors are used to connect the receiver coil on one side and either the AC/DC converter or a measuring unit on the other side.

In order to accurately tune the trimmer capacitor, a fixed value for the serie capacitor is chosen as close as possible to the value obtained with equation from chapter 1. The PCB is then connected to an LCR meter and the second capacitor is adjusted in order to achieve $\omega_0 = 1/\sqrt{LC}$. This procedure was applied and the same AC and DC measurements before were conducted. It was observed that the matching circuit had very little impact on the power transfer efficiency. This is not very surprising considering the extremely low quality

⁴Reference manufacturer Vishay: BFC2 809 080

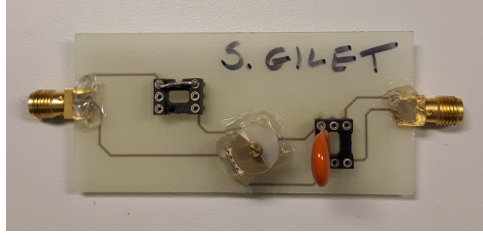


Figure 4.24: PCB used to connect extra components to the receiver coil.

factors of both the transmitter and receiver coil. Such Q factor leads to a broadening and flattening of the resonance peak of the coils. Due to the fact the necessary equipment was out of order at the time of these measurements, it was not possible to confirm this hypothesis with a vector network analyzer. Further measurements using a vector network analyzer to verify the resonance of the system coil+PCB are required in order to draw any conclusion from these results.

4.3.6 Conclusion

To conclude this last Chapter, the feasibility of powering a 10mmX10mmX0.3mm coil embedded 1- and 2-ply deep into CFRP prepeg was experimentally proved. The positive impact of adding ferrite on the back of the receiver coil was showed. COTS were selected and small ferrite backed coils were inserted into composite sheets. Using larger coils as readers, the coupling coefficient and AC and DC power transfer were measured. Over $200\mu\text{W}$ of AC power were transferred 1-ply deep for a efficiency of 0.6% with reader 1 and over $100\mu\text{W}$ of AC power were transferred 2-ply deep for an efficiency of 0.45% with reader 2.

Conclusion

The first thing to conclude with is to repeat that the composite properties vary with fabrication process, frequency and the type of plies used. This means that if elsewhere someone was to reproduce the experiments conducted here, that person might end up finding different results. Considering this, the procedure itself is really one of the main output of this work.

Here the feasibility of wireless power transfer of embedded sensor node for structural health monitoring was assessed. The overall context and the initial constraints for aerospace applications were laid down. First and after a review of powering option, it was decided to focus on magnetic resonant coupling. A transferred power of $200\mu\text{W}$ at an operating frequency of 10MHz and to a receiver featuring dimensions of $10\text{mm}\times 10\text{mm}\times 0.3\text{mm}$ was targeted. In Chapter 2, a complete analytic model of a two circular coils system separated by a conductive sheet was derived. This model, based on Dood and Deeds' vector potential calculations, allows to obtain the self-inductances, quality factor and mutual inductance of a given system. The equations were numerically solved and compared to experimental results in order to verify the model accuracy.

In Chapter 3, a circuit model of the inductive link was constructed. It was shown that the frequency had a very strong impact on the power transfer efficiency and an operating frequency of 10MHz was selected. If using the self-inductance, quality factors and mutual inductance evaluated by the analytic model, the circuit model allows to estimate the power transfer efficiency between the transmitter and the receiver coil separated by a conductive sheet. To conclude that section, the need for a more accurate model was shown through a case study and a numerical evaluation of all the equations here described. There, we considered standard values of the order the microHenry for coils of radius 0.5 cm and 2.5 cm for the receiver and transmitter respectively and $10\text{-}100\text{k}\Omega$ for the load. Using an initial CFRP conductivity of $5\cdot 10^4\text{ S/m}$ from the literature, power transfer efficiencies going from 0.45% to 0.0016% were obtained for primary to secondary spacing going from 1mm to 5mm .

Using 3D simulation, a design flow was proposed. This allows to extract the equivalent circuit parameter of a 3D coil simulation in order to take the impact of the composite into account when designing the inductive link and the matching network. To do so, CFRP was modeled as an homogeneous material of equivalent electrical conductivity, dielectric constant and permeability. This modelisation was possible using theoretical approximations for the dielectric constant and the permeability. The conductivity was measured according to various techniques and eventually evaluated by doing a comparison between experimental measurement of the impact of CFRP on test coils self-inductance and the analytic model developed in Chapter 1. This technique gave an initial conductivity value of $\sigma_{CFRP} \approx 10^4$ S/m. To show and verify the accuracy of the homogeneous material approximation, three PCB coils were fabricated and embedded into CFRP. They were simulated with CST in air and in a rectangular bloc made of an new material having those effective properties. The results of those test showed the conductivity at 10 MHz and for our custom fabrication process is actually closer to 7500 S/m. It was also proved that this model is able to predict a coil self-inductance embedded in CFRP as accurately as to predict it in air.

Finally, the feasibility of the powering of a wireless sensor embedded into a composite structure was proved. Off-the-shelf coils were embedded 1 and 2-ply deep into samples. Even though the efficiency is significantly lower than a similar system in the air, with external readers and at 10 MHz it was possible to transfer:

- Over $200\mu\text{W}$ of AC power were transferred 1-ply deep for a efficiency of 0.6% with reader 1.
- Over $100\mu\text{W}$ of AC power were transferred 2-ply deep for an efficiency of 0.45% with reader 2.

Considering the initially targeted AC power was $200\mu\text{W}$, it can be concluded that the power transfer fulfills the specifications for a receiver 1-ply deep. The embedded coils used respected the maximum element inclusion raised by aircraft structure manufacturer SABCA (10mmX10mmX0.3mm) which had never been done before. The system was also connected to a AC/DC converter and around $55\mu\text{W}$ of DC power were supplied to load of 12 to $100\text{k}\Omega$. This conversion efficiency is low and is due to the safety ESD diodes of the converter. This still corresponds to a power transfer efficiency of around 0.165% and 0.124% which allows us to say that a minimum of 1.64mW and 1.24mW of DC power will be harvested by the receiver for the maximum transmitter input power of 1W.

Future work

Some options for future work and to take this study further counts :

- It can be speculated that the power transfer efficiency could simply be improved by redesigning the readers. Both readers used were very sensitive to the composite proximity effect which led to a decrease of their self inductance (so of the mutual inductance) and to an increase of their quality factor. Using ferrite to back shield them could improve the transmission which could allow to embed the receiver deeper into the composite.
- Designing a proper matching network based on prediction made by the CST simulation using the homogeneous model.
- Building a load modulation circuit connectable to the embedded coils to validate properly a data transmission through capacitive or resistive load modulations.
- Embedded larger system containing the receiving coil, a matching network and a AC/DC converter while keeping a Ufl cable to connect to the outside. The different components could be mounted on a flexible PCB. Considering the composite fabrication process, it is expected that wire bonding the converter is not an option, the connection will not survive. An alternative could be the flip chip technique or ACA bonding described in [65] [66].
- Embedding sensor to see how they react after the curing cycle and inside the composite. The bare die could be flip chip on a flexible PCB which would itself extend to the outside of the composite test sample in a similar manner to the test coils of Chapter 3.
- Considering the outcome of all those test, an entire system could be mounted on a flexible PCB and tested inside the composite laminate.
- A similar system but on-chip could be designed in order to minimize its size and so its impact on the structural integrity of the composite. Considering the difficulties to achieve high Q-factor for inductors on chip, the coil might need to be separately fabricated and the IC mounted on it after.

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