

Contactless Power Transfer Through a Rotating Transformer for Spatial Application

Dissertation presented by
Thibaut DECEULENEER

for obtaining the Master's degree in
Mechatronics Engineering

Supervisors
Marc BEKEMANS , Bruno DEHEZ

Readers
Christophe CRAEYE, Francis LABRIQUE

Academic year 2015-2016

Abstract

In satellites, the energy required to supply the entire equipment is provided by the solar arrays. They are in constant rotation with respect to the satellite platform in order to face the sun at all time and maximize their efficiency. Therefore, the power is transferred through a rotating device which is called the Solar Array Drive Mechanism. The current technology ensuring the power flow is the slip rings system. However, it is limited by a low voltage operation because of electrostatic discharges.

In this master thesis, an alternative solution is explored with an inductive power transfer system. This system is composed of a rotating transformer and a Dual Active Bridge allowing to connect the solar array, the bus and the batteries in a single device. At first, models are defined in order to represent and evaluate important design features of the rotating transformer such as losses and thermal management. An optimization is then performed to reach the best efficiency for a maximal power density. Secondly, the results are validated by finite elements simulations and a prototype is manufactured for further tests and measurements.

Acknowledgements

This master thesis required an important amount of time. Still, many people were involved in this project and helped to carry out the work successfully. Therefore, I would like to acknowledge them personally.

First of all, I would like to address my acknowledgements to my supervisor, Prof. Marc Bekemans, for its continuous support, guidance and expertise. Also, I would like to thank Thales Alenia Space Belgium for the financial support of the system prototype. Moreover, I am highly grateful to the staff of the company, especially to Fabien Meinguet and Philippe Vandeplasseche, for giving me a precious and expert help along the semester.

I further would like to thank, Prof. Bruno Dehez, for his advice on the mechanical assembly. I thank the EPL FABLAB for their manufacturing quality on the mechanical assembly.

Finally, many thanks to Prof. Juliette Soulard (KTH) for her last minute reading of the report and the advice to improve it.

Introduction	1
1 Context	3
1.1 Overview of satellites	3
1.1.1 Satellite environment	4
1.1.2 Satellite power system	4
1.2 Solar Array Drive Mechanism	8
1.2.1 Slip rings system	8
1.2.2 Inductive Power Transfer	10
1.2.3 Capacitive Power Transfer	11
1.2.4 Comparison	11
1.3 Transformer geometry topologies	12
1.4 Power Electronics for Inductive Power Transfer	14
1.4.1 Electronic switches	14
1.4.2 Half-bridge and Full-bridge converters	16
1.4.3 Inductive Transformer Model	17
1.4.4 Dual Active Bridge	17
1.5 Specifications of the rotating transformer	21
2 Transformer Modelling	22
2.1 Magnetic circuit modelling	24
2.1.1 B-H curve	24
2.1.2 Reluctance model	25
2.1.3 Magnetizing inductance (L_μ)	26
2.1.4 Leakage inductances ($L_{f,1}$ and $L_{f,2}$)	27
2.1.5 External inductor (L_{ext})	28
2.2 Core loss modelling (R_{ml} and $R_{p,ext}$)	30
2.2.1 Core selection	30
2.2.2 Core loss for trapezoidal waveforms	31
2.3 Winding loss modelling (R_1 , R_2 and $R_{s,ext}$)	33
2.3.1 Skin effect	33
2.3.2 Proximity effect	34
2.3.3 Litz wire	36
2.4 Thermal modelling	38
2.4.1 Thermal resistance concept	38
2.4.2 The lumped parameter thermal model	38
2.4.3 Winding equivalent thermal conductivity	40
2.4.4 Core-Winding interface resistance	41

2.4.5	Temperature computation	41
3	Optimization	43
3.1	Optimization variables	43
3.2	Objective functions	45
3.3	Optimization method	45
3.4	Optimization results	47
3.4.1	Frequency impact	48
3.4.2	Results	48
3.5	External inductor design	51
3.5.1	Fringing effect	52
4	Finite Element Analysis	54
4.1	Transformer analysis (2D)	55
4.1.1	Magnetic field simulation (2D)	55
4.1.2	Thermal transfer simulation (2D)	56
4.2	Transformer analysis (3D)	57
4.2.1	Winding connectors	57
4.2.2	Magnetic field simulation (3D)	57
4.3	Inductor analysis (2D)	58
4.3.1	Magnetic field simulation (2D)	59
4.3.2	Thermal transfer simulation (2D)	60
5	Construction	61
5.1	Electric assembly	61
5.1.1	Components connection	61
5.1.2	Transposition in transformer windings	62
5.1.3	Tri-port mode	62
5.2	Mechanical assembly	63
5.2.1	Magnetic core	63
5.2.2	Support	64
5.3	Ferrite manufacturing	64
5.4	Visual prototype	65
	Conclusion and further work	66
	Bibliography	68
A	Theory appendices	71
A.1	Electrostatic Discharge and Paschen's law	71
A.2	Zero Voltage Switching	73
A.3	Voltage, current and flux density waveforms in the transformer	74
A.4	Trapezoidal waveform : harmonic content	75
A.5	Reluctance formulae derivation	76
A.6	Iron losses : improved Generalized Steinmetz Equation for trapezoidal waveforms	78
A.7	Proximity effect	80
A.8	Thermal model : equivalent thermal conductivity of Litz Wire winding	83
A.9	Thermal model : thermal resistances calculations	85
B	Datasheets	87
B.1	Solar Array Drive Mechanism : RUAG Septa 24®	87
B.2	Power Conditioning Unit Next Generation	89
B.3	Dual Active Bridge Transistor : SiC MOSFET	91
B.4	Rupalit Litz Wire	93
B.5	Ferroxcube 3C90 Ferrite : Material Specifications	94
B.6	Ferroxcube 3C92 Ferrite : Material Specifications	95
B.7	Ferroxcube Ferrite : E71/33/32	96

List of Figures

1.1	Satellite : Earth orbits representation	4
1.2	Satellite : schematic of the Power System	5
1.3	Solar Array : model and characteristic curve of the Photovoltaic cell	6
1.4	Battery : comparison of energy storage technologies (Ragone plot)	6
1.5	PCU : schematic	7
1.6	PCU : solar array power control schematics (S3R and MPPT)	8
1.7	SADM : slip ring technology	9
1.8	SADM : slip rings : Paschen curve	10
1.9	SADM : Inductive Power Transfer schematic	11
1.10	SADM : Capacitive Power Transfer schematic	11
1.11	Transformer geometry : comparison of technologies	13
1.12	Transformer geometry : representation and dimensions	14
1.13	Power Electronics : active switch	15
1.14	Power Electronics : passive switch	15
1.15	Power Electronics : switching losses representation	16
1.16	Power Electronics : half-bridge and full-bridge circuit representations	16
1.17	Power Electronics : transformer inductive model	17
1.18	Dual Active Bridge : circuit representation	17
1.19	Dual Active Bridge : currents and voltages	19
1.20	Dual Active Bridge : phase-shift analysis	20
2.1	Transformer modelling : circuit model of transformer and inductor	22
2.2	Transformer modelling : circuit model of combination of transformer and inductor	23
2.3	Transformer modelling : main dimensions	23
2.4	Magnetic circuit : B-H curve comparison between air and a magnetic material	25
2.5	Magnetic circuit : magnetic core circuit	26
2.6	Magnetic circuit : compete magnetic circuit with leakage flux considered	27
2.7	Magnetic circuit : airgapped core for external inductor	29
2.8	Core loss : hysteresis and eddy current effects	30
2.9	Core loss : comparison of magnetic materials	31
2.10	Winding loss : skin effect	33
2.11	Winding loss : representation of skin and proximity effects	34
2.12	Proximity effect : Dowell's equation curve	35
2.13	Proximity effect: harmonic impact	36
2.14	Proximity effect : efficient winding geometry	36
2.15	Proximity effect : Litz wire technology	37
2.16	Thermal modelling : thermal circuit over the transformer	39
2.17	Thermal modelling : complete equivalent thermal circuits	39

2.18	Thermal modelling : winding cross representation for the equivalent thermal conductivity	40
2.19	Thermal modelling : thermal interface material	41
3.1	Optimization : input variables diagram	44
3.2	Optimization : optimization diagram	46
3.3	Optimization : losses with respect to power density for all combinations	48
3.4	Theoretical results : transformer cross section representation	50
3.5	Theoretical results : transformer temperature	51
3.6	Inductor : fringing effect on windings	53
4.1	FEM : flux density in the transformer (2D)	56
4.2	FEM : thermal transfer in the transformer (2D)	56
4.3	Assembly : 3D representation of the transformer	57
4.4	FEM : flux density in the transformer (3D)	58
4.5	FEM : Flux density in the inductor (2D)	60
4.6	FEM : Thermal transfer in the inductor (2D)	60
5.1	Electric assembly : bi-port connections	62
5.2	Electric assembly : transposition	62
5.3	Electric assembly : tri-port mode	63
5.4	Mechanical assembly : magnetic cores layers constitution	64
5.5	Mechanical assembly : Solidworks representation of the assembly	64
5.6	Manufacturing : ferrite broken blocks	65
5.7	Visual prototype	65
A.1	Paschen curve	72
A.2	Paschen curve : explanation	72
A.3	Zero Voltage Switching : Principle	73
A.4	Transformer waveforms : representation of voltage, current and flux density	74
A.5	Trapezoidal waveform	76
A.6	Reluctances calculation : flux direction and distribution representation	76
A.7	Trapezoidal waveform	79
A.8	Proximity effect : two wires effect ($\delta \gg d$)	80
A.9	Proximity effect : transformer case ($\delta \gg d$)	81
A.10	Proximity effect : Dowell's equation curve	82
A.11	Thermal modelling : winding cross representation for the equivalent thermal conductivity	83
A.12	Thermal resistance calculation : thermal circuit representation	85

List of Tables

1.1	Solar Array Drive Mechanism : Septa 24 [®] specifications	9
1.2	SADM : comparison of technologies	12
1.3	Dual Active Bridge : main characteristics	21
2.1	Magnetic circuit : Hopkinson equivalences	25
2.2	Core loss : comparison of magnetic cores	31
3.1	Design material : Litz Rupalit [®] AWG41 specifications	47
3.2	Design material : Ferrite Ferroxcube 3C90 specifications	47
3.3	Theoretical results : transformer dimensions and characteristics with theoretical calculations	49
3.4	Design material : Ferrite Ferroxcube 3C92 specifications	52
3.5	Theoretical results : external inductor dimensions and characteristics	52
4.1	FEM : comparison on transformer characteristics between theory and 2D FEM simulations	55
4.2	FEM : comparison on transformer characteristics between 2D and 3D FEM simulations .	58
4.3	FEM : comparison on inductor characteristics between theory and 2D FEM simulations .	59
5.1	Final results for the Inductive Power Transfer for a Solar Array Drive Mechanism	67
A.1	Reluctance model : distance between nodes and the axis	85
A.2	Reluctance model : reluctances expression	86

Acronym List

AC	Alternating Current
AWG	American Wire Gauge
BCR	Battery Charge Regulator
BCDR	Battery Charge & Discharge Regulator
BDR	Battery Discharge Regulator
CPT	Capacitive Power Transfer
DAB	Dual Active Bridge
DC	Direct Current
EM	Electromagnetic
EMF	Electromagnetic Force
EMI	Electromagnetic Interference
ESD	Electrostatic Discharge
FEM	Finite Element Method
GEO	Geostationary Orbit
IGBT	Insulated Gate Bipolar Transistor
IPT	Inductive Power Transfer
ISGE	Improved Steinmetz Generalized Equation
MPPT	Maximal Power Point Tracker
MOSFET	Metal Oxide Semiconductor Field Effect Transistor
OC	Open Circuit
PCDU	Power Conditioning & Distribution Unit
PCU	Power Conditioning Unit
PDU	Power Distribution Unit
PV	Photovoltaic
PWM	Pulse Width Modulation
RMS	Root Mean Square
S3R	Sequential Switching Shunt Regulator
SA	Solar Array
SADM	Solar Array Drive Mechanism
SC	Short Circuit
SOC	State of Charge
SMPS	Switched Mode Power Supply
TFO	Transformer
ZVS	Zero Voltage Switching
WPT	Wireless Power Transfer

Introduction

Nowadays, satellites have a considerable impact on Earth-used technologies. The descendants of *Sputnik* are used by humans to execute several different missions such as telecommunication, navigation, observation, meteorology, etc. A satellite is furnished with an equipment which is dedicated for a mission and which requests more and more energy. Typically, the on-board power for a telecommunication mission is around 20 kW. This power is provided by two solar arrays located around the platform. In order to gather as much solar energy as possible, the solar arrays are continuously spinning and facing the sun. However, this rotation complicates the power transfer from the energy source to the equipment. Indeed, the constant rotation does not allow to make the electric connection with a simple wire. The device ensuring this power transfer is called the *Solar Array Drive Mechanism* (SADM). The current technology used for this system is the well-known slip rings. However, this technology has some limiting features.

In the frame work of this master thesis, an other system has been studied in order to replace the slip rings. The project aims to achieve a 10 kW power transfer, by means of a rotating transformer. This corresponds to the power provided by a single solar array. The system has to be able to compete with the current technology. The entire project has been undertaken in collaboration with Thales Alenia Space Belgium which is a leader company in power electronics on satellites. Moreover, this work falls under the continuity of a first master thesis executed in 2015 by Patrick Duvivier [1]. In his study, he has established the state of the art of inductive power transfer through rotating transformers. In the present work, the project becomes more concrete. The main goal is to build a prototype of the rotating transformer. In order to do that, the design of the system has to be optimized and validated by analytical and numerical tools.

This document aims to give a description of the work that has been done during the project. It also explains the reasoning that has been followed. The results from analytical and numerical simulations are presented and analysed.

Chapter 1 aims at providing a general background and overview of the issue. The satellite power system is introduced in order to identify the components interacting with the SADM. The different technologies for Wireless Power Transfer (WPT) are then presented and compared to the current slip ring technology. After, the transformer geometry is discussed with respect to conclusions drawn in the previous master thesis. Finally, the power electronics involved for the Inductive Power Transfer (IPT) is introduced.

The objective of Chapter 2 is to develop models for the important aspects of a transformer. Those models are tools that will help for the design of an efficient and optimized transformer. It also allows to have a better understanding of the losses in the transformer. The thermal management is a significant point of design for spatial application. Therefore, a thermal model is also presented and applied on the transformer.

An optimization is performed in Chapter 3 in order to make the transformer as efficient as possi-

ble. The different input variables and the objective functions for the optimization are introduced. The characteristics of the optimal transformer are then analysed. This chapter is ended by the design of a supplementary inductor.

Chapter 4 presents finite elements simulations that have been carried on in order to validate the results obtained in Chapter 3. Both 2D and 3D analysis of the flux density inside the transformer core are performed. Likewise, the thermal management is also validated with a 2D simulation.

Finally, Chapter 5 presents the electric and mechanic assemblies of the prototype. No measurement or performance validations are provided in this chapter due to manufacturing issues discussed in this section. However, a visual prototype is presented.

This chapter aims to give an overall idea of the context within which the rotating transformer is used. The transformer is part of the Solar Array Drive Mechanism (SADM) which is the interface between the solar array and the platform of a satellite.

First, an overview of a typical satellite is presented. The principal constraints of the satellite environment are expressed. The satellite power system which includes the SADM is described. Afterwards, the dedicated device is deeply presented and a comparison between different SADM technologies is made in order to highlight the issues of the current technology and the potential of a rotating transformer to ensure the power flow function. Then, the transformer geometry is discussed. This discussion is based on the state of the art of rotating transformers and the previous work done in the project [1]. After that, the *Dual Active Bridge* (DAB), the power electronics associated to the transformer is introduced and the main equations associated to this converter are derived. Finally, the specifications of the rotating transformer are defined..

1.1 Overview of satellites

Satellites are man made systems placed into orbit. This orbit can be around a planet as the Earth, around a star as the sun or even around natural satellites as the moon. Those are used by different entities (Defense, Civil and Science) and for different purposes (navigation, telecommunication, observation, meteorology, etc.). The first satellite, *Sputnik*, was sent in orbit of Earth in 1957 by the Soviet Union. After that, more than 6600 satellites have been launched in space and the space launch market has become a real competition.

Earth satellites are classified with respect to their orbit radius (Figure 1.1) : Low Earth Orbit (LEO), Medium Earth Orbit (MEO), Geosynchronous Orbit (GSO) and High Earth Orbit (HEO). A GSO satellite completes one full orbit around Earth in one day. It can be named as geostationary (GEO) when the satellite stands exactly above the equator. The satellites considered in this document are geosynchronous.

In this section, an overview on the satellites is made. First, the satellite environment is analysed. Then, the main devices composing the satellite power system are described.

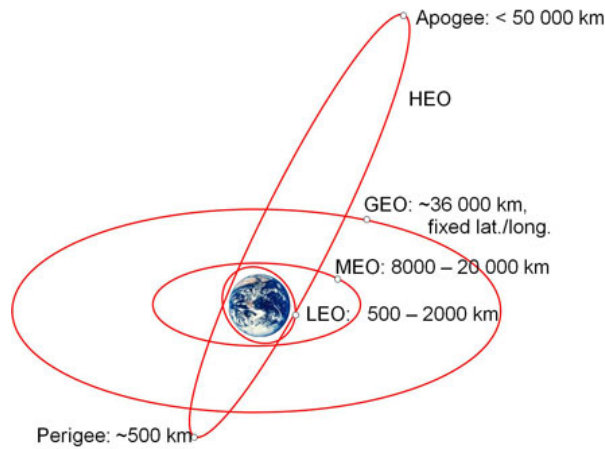


Figure 1.1 – Earth orbits representation

1.1.1 Satellite environment

A satellite operates in a constrained environment. The constraints need to be considered for the design of any device of a satellite. The major constraints in space are :

- **Vacuum** : that implies the impossibility to cool a satellite with convection as there is no gas. Cooling of the systems is done by conduction only. This also implies an important temperature gradient in the satellite : the side facing the sun is heated while the side facing the space darkness is cooled by radiation. This temperature gradient can reach 320°C [2]. The satellite temperature also decreases to -70°C when an eclipse occurs. Therefore, the device to design has to be able to face cold and warm temperatures.
- **No gravity** : This makes it impossible to use convection in a sealed system. However, zero gravity allows to decrease weight forces and frictions.
- **Radiation exposure**: The radiations that might be encountered in space are protons, electrons or cosmic rays. Those can cause important damages. Therefore, the different technologies on board have to be robust to those radiations. The technology which is the most vulnerable to radiations is the semiconductor electronics (transistors, photovoltaic panels, IGBT, etc.).
- **Plasmas**: A plasma is a ionised gas produced by ionisation of atmosphere molecules due to solar radiations. Those are located around Earth and can cause different problems in a satellite. For example, it promotes the electric discharge (ESD) phenomenon between conductors. This can lead to the destruction of a system. This specific point will be explained more in a next section.
- **Maintenance**: No maintenance is available in space. A satellite mission lasts up to 15 years. Therefore, the systems on board have to be highly robust and fault-tolerant.
- **Heavy vibrations**: During its launch, the satellite undergoes heavy vibrations. The systems composing the satellite have to be able to withstand vibrations that can reach an amplitude of 10 g.

Finally, an important point when designing a satellite device is to minimize the weight. On the orbit, this change nothing. However, weight has an impact on the satellite launch cost. Indeed, the cost per kilogram of a satellite launch is around 16 k€/kg.

1.1.2 Satellite power system

Regarding electric management, a satellite is composed of different systems which can be divided into three categories : avionic systems related to the satellite behaviour such as the on-board computer and the ion thrusters; instrumentation systems related to the satellite mission such as antennas for communication; and the power system which is in charge of the electric power management inside the satellite.

This last category is described in this section.

Figure 1.2 presents the main devices involved in the satellite power system. They are :

- **Solar Array** (SA) which is the power source in a satellite;
- **Batteries** which are used to store energy and to supply the entire satellite when an eclipse occurs;
- **Solar Array Drive Mechanism** (SADM) which is in charge of the solar array orientation and of power transmission from the solar array to the satellite platform. This is the system which is studied in this document and a more deepen overview will be given in next section;
- **Power Conditioning and Distribution Unit** (PCDU) which is in charge of controlling the power flow exchange between the solar array, the batteries and the different satellite systems. It can also regulate the bus voltage (V_{bus}).

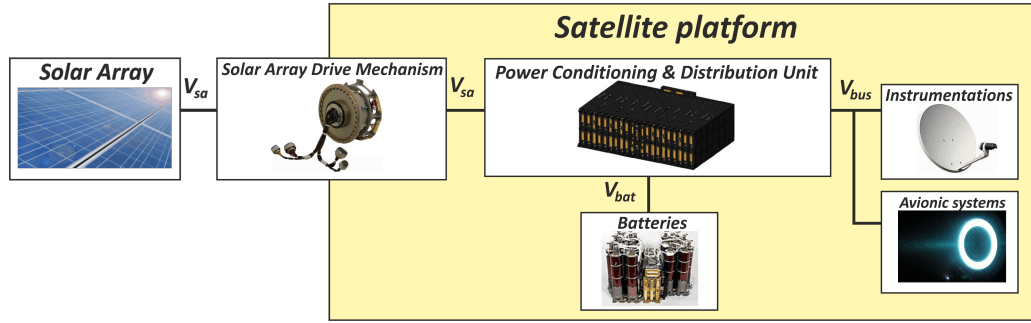


Figure 1.2 – Schematic of the satellite power system

Solar Array

The solar array (SA) is the principal energy source in a satellite. Its role is to convert light energy coming from the sun into electric current by photovoltaic effect. A first part of this current is directly used by the satellite equipment while a second part is used to charge the batteries. In a geosynchronous satellite, the total power needed can reach 20 kW. Usually, that power is provided by two SA, meaning a production of 10 kW per array.

The SA is composed of photovoltaic panels which are connected either in parallel or in series. The photovoltaic panels are themselves composed of a number of photovoltaic cells. All the cells are both connected in parallel and in series in order to reach the rated voltage and rated current. The photovoltaic effect is explained in more depth in [1].

A simplified model of a photovoltaic cell is depicted in figure 1.3a[4]. That model is represented by a parallel connection of a current source standing for the photovoltaic current generation and a reversed diode representing the semi-conductor effect. The parasitic resistances are not represented in order to simplify the explanation and the equations. The photovoltaic cell equation related to the model of figure 1.3a is found with the diode characteristic equation ¹ :

$$I_{cell} = I_{ph} - I_0 \left(e^{-V/V_T} - 1 \right) \quad (1.1)$$

I_{ph} is the photovoltaic current related to light radiation and cell composition; I_0 is the reverse bias saturation current related to the cell material; V_T is the thermal voltage related to the cell temperature. Figure 1.3b represents the I-V characteristic curve corresponding to equation (1.1). The red dot on the curve represents the maximal power available for a given photovoltaic current. This is at that point that the cell is the most efficient. Therefore, this is interesting to reach this point at any moment.

¹Diode equation : $I_D = I_0 \left(e^{-V/V_T} - 1 \right)$

This is done with a *Maximal Power Point Tracker* (MPPT). This device is explained in the PCDU section.

The PV cell efficiency depends on the technology used for the materials. Currently, PV cells dedicated to spatial applications can reach an efficiency up to 30%. For more information about the photovoltaic technology in space sector, it is advised to read the state of the art of the project [1].

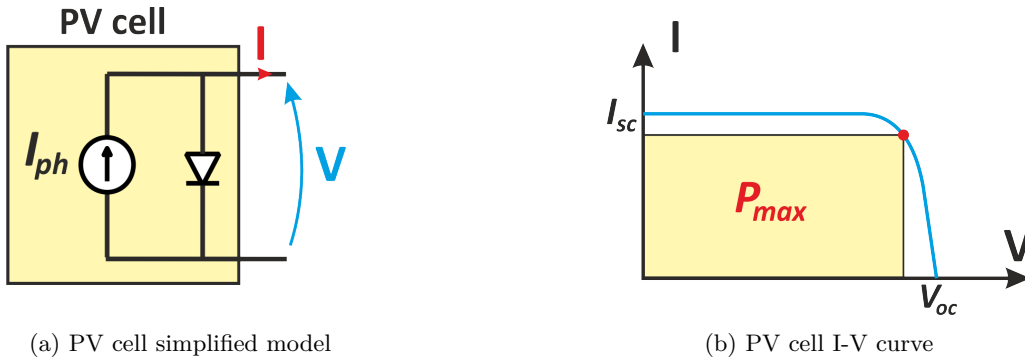


Figure 1.3 – Photovoltaic cell : model and characteristic curve

Batteries

Batteries have an important task in the satellite power system. They store energy coming from the solar array before an eclipse occurs in order to provide this energy to the satellite equipment during this dark interval of time. The battery is sized to be able to provide the necessary energy during an entire eclipse, at end of life. For a geostationary satellite, eclipses occur during half time of the year, once a day and with a maximal duration of 72 minutes. Therefore, the batteries can be charged during more than 22 hours. This means that the needed battery discharging flow rate is much larger than the charging flow rate ($P_{discharge} > 20 \cdot P_{charge}$).

In all the industrial sectors, the most used battery type is the Lithium-ion battery. Indeed, it offers a high specific energy for a relatively high specific power. Figure 1.4 is a comparison of the Ragone plots of different energy storage devices. A high specific energy allows to reduce the battery weight and volume inside the satellite. Lithium-ion batteries have also a good charge/discharge efficiency (up to 90%). A disadvantage of a battery is that its voltage, which is the voltage reference for the satellite bus, is dependant on its state of charge (SOC). Therefore, a regulation of the bus voltage is needed.

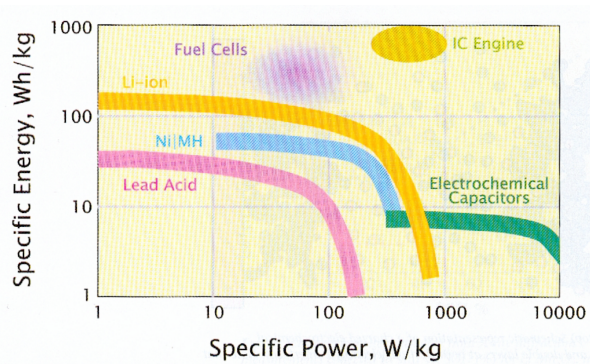


Figure 1.4 – Ragone plots for different energy storage devices [5]

Power Conditioning and Distribution Unit

The Power Conditioning and Distribution Unit (PCDU) is divided in two parts : the Power Conditioning Unit (PCU) which controls the power flowing from/to the energy sources (solar array and batteries)

and the Power Distribution Unit which is in charge of distributing power to the satellite systems. It is important to understand the tasks of the PCU because this device has a high interaction with the SADM.

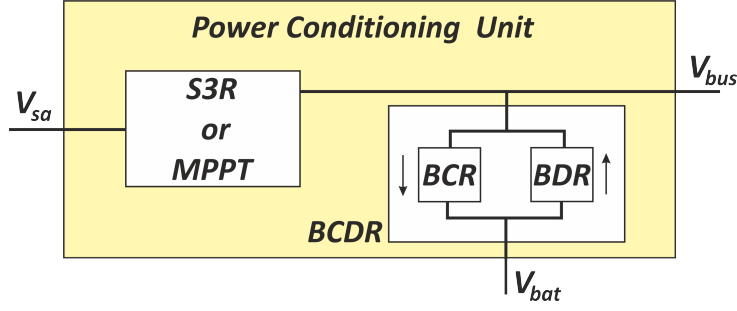


Figure 1.5 – Power Conditioning Unit schematic

Figure 1.5 is a schematic of the PCU. This device has two fundamental tasks : the solar array power regulation and the batteries power regulation. The solar array power regulation is needed to regulate the power flow that the SA delivers to the batteries and the other systems. Indeed, all the energy that the SA can provide is not always needed by the satellite equipment. This power regulation is done either with a *Sequential Switching Shunt Regulator* (S3R) or with a *Maximum Power Point Tracker* (MPPT). Both method schematics are represented in figure 1.6 and are explained in the following:

- Sequential Switch Shunt Regulator

With the S3R method, the SA is directly connected to the bus ($V_{sa} = V_{bus}$). Then, a switch is connected in parallel of each PV section (Figure 1.6a). The output current is modulated by short-circuiting some PV sections with activation of the associated switches. In order to avoid a short-circuit on the bus, diodes are connected between the bus and each PV section.

As for all electric devices, losses are generated in the S3R. The major part of losses is generated by the conducting diodes (D_i) :

$$P_D = (V_{th,D} + R_{s,D} \cdot I_o) \cdot I_o \quad (1.2)$$

where $V_{th,D}$ is the diode threshold voltage, $R_{s,D}$ is the diode series resistance and I_o is the output current.

- Maximal Power Point Tracker

The MPPT method consists of tracking a given power point with a regulation of the solar array voltage on the I-V characteristic curve. The SA voltage is regulated by a DC/DC converter (Figure 1.6b). When the batteries are not fully charged or when the satellite needs full power, a maximal power point is tracked on the I-V curve. Therefore, the SA efficiency is maximal. That is a major advantage of MPPT compared to S3R where V_{sa} is fixed.

Unfortunately, DC/DC converters generate conducting and switching losses. The total losses generated in the MPPT DC/DC converter are higher than the losses generated in the S3R diodes. That is why the MPPT is less popular even if it can extract maximal power from the SA.

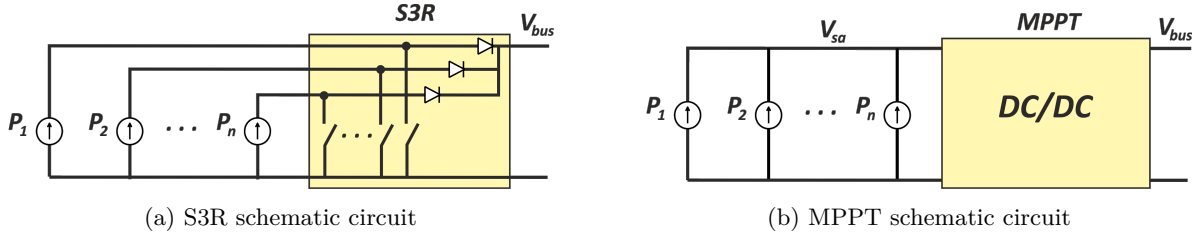


Figure 1.6 – Power Conditioning Unit : solar array power control schematic circuits

The second PCU task is to regulate the power flowing from/to the batteries. This is achieved by a *Battery Charge and Discharge Regulator* (BCDR) presented in figure 1.5. This device is composed of two unidirectional DC/DC converters between the batteries and the bus. In the *Battery Charge Regulator*, the current flows from the bus to the battery while charging. In the *Battery Discharge Regulator*, the current flows from battery to the bus while discharging. Those two regulators are Buck or Boost converters depending on the battery to bus voltage ratio. The BCDR is also used to regulate the bus voltage and cancel its dependency on the battery state of charge. Once again, the use of DC/DC converters implies conducting and switching losses. Sometimes, the PCU does not have a BCDR and the bus is directly connected to the battery. In this case, the bus is called unregulated.

In order to give an idea on a typical PCU specifications, the *Power Conditioning Unit Next Generation* developed by *Thales Alenia Space* is available in Appendix B.2. The S3R efficiency is higher than 99% and the BCDR efficiency is higher than 96%. The weight of this device is 50 kg.

1.2 Solar Array Drive Mechanism

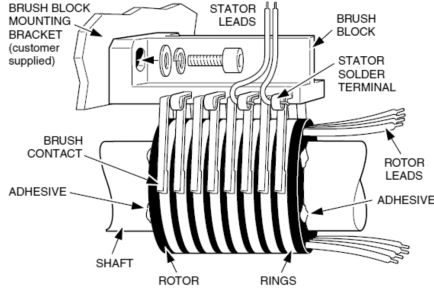
During a complete revolution, the solar array needs to face the sun with a 90° angle in order to maximize the photovoltaic efficiency. However, the satellite turns around the Earth and is always facing it with the same side. Therefore, the solar arrays need to rotate around the satellite axis. For a geosynchronous satellite, the rotation speed is 1 *turn/day*. The interface between the SA and the satellite platform is provided by a Solar Array Drive Mechanism (SADM) (Figure 1.5). The SADM has three tasks to complete :

- Perform the rotation of the SA with respect to the satellite platform. This is done by a stepper motor;
- Transfer power from the SA to the satellite platform;
- Transfer signals between the SA and the satellite platform.

From those tasks, the one that is involved in this document is the power transfer. The SADM can be seen as composed of a rotor on the SA side and a stator on the platform side. The rotor rotates with respect to the stator. In this document, it is assumed that the rotor is the inner part of the machine. The issue with the SADM is that high power has to be transferred from the rotor to the stator and wire connection is impossible because the rotation is multi-turn. Currently, the best technology for the SADM power transfer is the slip rings system.

1.2.1 Slip rings system

A slip ring is a device used to create an electric connection between a rotor and a stator. This kind of connector is used in many electrical machines such as DC-excited synchronous motors. Figure 1.7a represents the composition of a slip ring. The slip ring system is composed of conducting rings on the rotor surface and flexible conducting brushes on the stator. The electric connection is made by the gliding of brushes on the rings. In order to have a good mechanical contact, a high pressure is applied between brushes and rings. This allows to decrease the electric resistance of the contact.



(a) Slip ring composition [6]



(b) RUAG : Septa 24[®] [7]

Figure 1.7 – Solar Array Drive Mechanism : slip ring technology

RUAG space is a company providing SADM with slip rings. Figure 1.7b depicted the RUAG space product *Septa 24*[®] and table 1.1 contains its power specifications. The system is composed of several lines in parallel in order to reduce the current passing through a line. Each line is connected to one PV section and one switch of the S3R (Figure fig : review : S3R). If a problem happens for one line, the corresponding PV section is lost. As it consists of a simple electric contact, slip rings is an interesting technology because it provides low power losses and high efficiency : 99.6%.

Specification	
Number of power lines	12 or 24 lines
Current on a line	6.2 A _{RMS}
Voltage	up to 110 V
Power transfer	up to 16.3 kW
Connector-to-Connector resistance	< 40 mΩ
Power losses	< 62W

Table 1.1 – Septa 24[®] power transfer specifications [7]

The main disadvantage of this device is that the maximal voltage is limited to 110 V. This is due to electrostatic discharge (ESD) phenomenon. An ESD is caused by a dielectric breakdown occurring when the voltage difference between two conductors is above the dielectric breakdown voltage. In a gas, this breakdown voltage is function of the product of the pressure (p) and the distance between the conductors (d) ($V_{breakdown} = f(pd)$). This function is known as the *Paschen's law* and a representation of the function curve can be seen in figure 1.8. This function has a minimum which is the limit below which it is certain that no ESD can occur. More explanation about this *Paschen curve* is available in Appendix A.1.

In vacuum, there is a zero pressure. The breakdown voltage for zero pressure is infinite (Figure 1.8). However, around the atmosphere there exist plasmas which are gases composed of electrons and positive ions. Those come from atmospheric molecules ionized by solar radiation. A plasma has a low but non null pressure. Indeed, for relatively long gap length (d), the product (pd) can reach the minimal breakdown voltage point on the *Paschen curve* (Figure 1.8). Because of historical events, space designers decided to keep voltage below the minimum breakdown voltage for bare conductor systems in order to avoid bad surprises. With a security factor, the voltage cannot be above 110 V.

However, the increase of the bus voltage is highly interesting. Indeed, for a given load power, an increase of the voltage allows a reduction of the current. If the current is decreased, the conductive losses would be as well. A way to overcome the breakdown voltage issue is to drop out the slip rings system and to use Wireless Power Transfer (WPT) for the SADM. There exist two kinds of WPT : Inductive Power Transfer (IPT) and Capacitive Power Transfer (CPT).

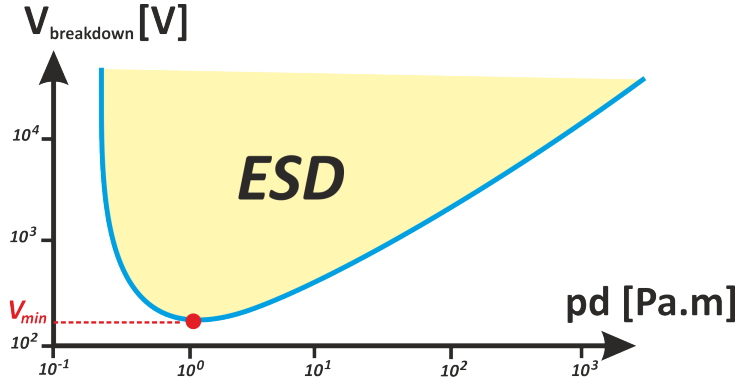


Figure 1.8 – Paschen curve : function of breakdown voltage with respect to the pressure and distance product

1.2.2 Inductive Power Transfer

The Inductive Power Transfer (IPT) is the most used WPT system. This technology is used in devices for low to high power transfer (0.1W to 30kW) in different sectors such as biomedical, automotive recharge, etc [9]. For example, this is the system used to charge electric toothbrushes. As it can be understood by its name, an IPT uses magnetic induction to transfer power from a primary side to a secondary side. The principle is that a primary winding is carrying an alternating current that generates an alternating magnetic flux in a magnetic core. This magnetic flux, in turn, induces an alternating current in the secondary winding. This device is known as the so-called *transformer*.

The first problem for a IPT is that currents and magnetic flux involved in a transformer are alternating (AC). However, the current provided by the SA is a DC current as well as the bus current. Therefore, a DC/AC power conversion before transformer and a AC/DC power conversion after it are both necessary (Figure 1.9).

Frequency is an issue when designing a transformer. Indeed, for a same power transferred, the transformer volume is inversely proportional to the frequency. Therefore, it is important to use high frequency to reduce the transformer volume and weight. The system composed of a DC/AC converter, a transformer and an AC/DC converter can be gathered in a Switched-Mode Power Supply (SMPS) which is a DC/DC converter. A SMPS works in the range of frequency from 10 kHz to 1 MHz. That converter controls the power flow with a Pulse Width Modulation (PWM). A major advantage of the SMPS for the SADM is that it can be used for the MPPT power regulation of the SA. A maximal efficiency of the SA can then be reached. However, the disadvantage is that the power electronics used in the SMPS produces switching and conducting losses which are added to the losses generated by the transformer. The switching and conducting losses of power electronics are explained in section 1.4. The transformer losses are generated by the magnetic core and the windings. The transformer losses modelling is presented in chapter 2.

An other advantage of the IPT is that a transformer can be multi-port [8]. Several different windings can be wound around the magnetic core and the same flux is seen by each winding. The use of a multi-port transformer is interesting to replace the BCDR. Indeed, a tri-port transformer would allow to connect the SA, the battery and the bus in a single device. The only constraint is that the associated power electronics needs to be bidirectional for the batteries. In that case, the transformer with its power electronics can be used for the SADM and the PCU at the same time. Therefore, cost and volume can be reduced due to less amount of devices and associated electronics. The system reliability is also increased with a decreasing number of devices.

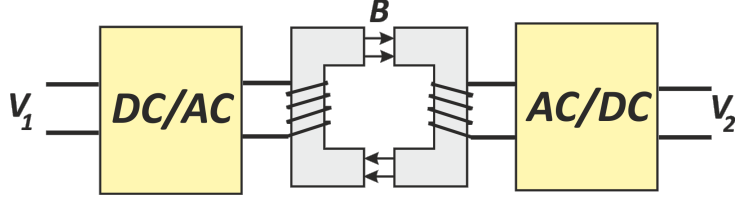


Figure 1.9 – Inductive Power Transfer

1.2.3 Capacitive Power Transfer

The Capacitive Power Transfer (CPT) is usually used in devices for very-low to medium power (0.01W to 1kW)[9]. The component used for this power transfer is a capacitor composed of two conductive plates. The power is transferred by an electric field produced with a voltage difference across the plates (Figure 1.10). Two capacitors are needed in order to close the circuit between primary and secondary sides. The capacitance of a capacitor is :

$$C = \frac{Q}{V} \quad (1.3)$$

where Q is the charge on the plates and V is the voltage difference between the plates. As the current is a charge flow ($i = \frac{dQ}{dt}$), the relation between current and voltage in a capacitor is :

$$i_c = C \frac{dV_c}{dt} \quad (1.4)$$

Likewise the IPT, the voltage of the CPT needs to be alternating. Otherwise, no current would cross the capacitor. Therefore, DC/AC and AC/DC converters are also needed around the capacitors. These capacitors can also be used in a SMPS. As for transformers, the capacitor volume is inversely proportional to the frequency for a given power transferred. However, for a same power and a same volume, a CPT requires generally a frequency five times higher than a IPT [9]. This high frequency generates high switching losses in the electronics associated.

Two kind of losses are generated in a capacitor : series resistance losses and dielectric losses. The first ones correspond to resistive losses in the conductive plates. The second ones are due to two phenomena: molecular polarization and interfacial polarization. According to [11], at high frequency, conductive losses are the dominant part of the losses.

For CPT, it is not possible to use a multi-port capacitor. However, the MPPT functionality is available like for an IPT.

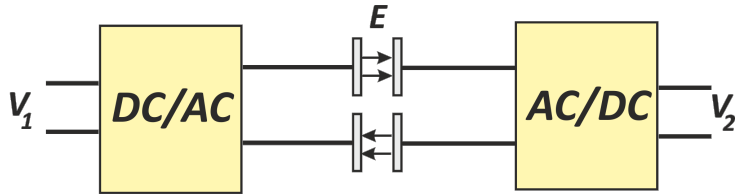


Figure 1.10 – Capacitive Power Transfer

1.2.4 Comparison

Finally, a comparison is done between the three SADM topologies introduced in this section : slip rings, IPT and CPT (Table 1.2).

On the one hand, the slip rings maximal voltage is limited to 110V where the IPT and CPT are both able to withstand a high voltage. This limitation has an impact on all the conductive losses generated in the satellite.

On the other hand, both WPT are used in a SMPS which generates high losses. For a same power density, the IPT works at high frequency (10-100 kHz) while the CPT works at very high frequency (100 kHz -1 MHz). Therefore, switching losses in the CPT electronics might reach high values. The use of SMPS allows to regulate the solar array power with a MPPT and operate at maximal efficiency. For slip rings, solar array power regulation is achieved with a S3R where losses are also generated.

For the device efficiency itself, slip rings are really better because they only generate DC conductive losses. For the IPT, losses are generated in the core and the windings which are really long. For the CPT, conductive alternating losses are generated in the capacitor plates. Finally, IPT system provides interesting functionalities such as a tri-port access which allows to use the transformer for the BCDR system.

To conclude the SADM section, the Inductive Power Transfer has the potential to compete with the currently used slip rings system. The principal advantages are high voltage and the possibility to merge the SADM with the entire PCU in a single transformer with its power electronics.

	Slip rings	IPT	CPT
High Voltage (V > 300V)	✗	✓	✓
MPPT Functionality	✗	✓	✓
Galvanic isolation	✗	✓	✗
BCDR Functionality	✗	✓	✗
Tri-port Functionality	✗	✓	✗
Device efficiency	+	–	–
Electronics efficiency	+	+	– –

Table 1.2 – SADM comparison : Slip rings, Inductive Power Transfer and Capacitive Power Transfer

1.3 Transformer geometry topologies

In this section, the transformer geometry is discussed. This discussion is based on the state of the art of IPT rotating systems made by different references [1], [12], [13], [14].

The important point for the geometry is that the transformer has to be rotating. This means that the magnetic circuit has to be divided in two parts corresponding to the rotor and the stator. Moreover, it could be interesting if the performances are not too dependent on the rotor-to-stator position. Indeed, a good symmetry of the transformer geometry can provide a limitation of the performance fluctuation with respect to the position. Therefore, an axisymmetric transformer (symmetry around the transformer axis) seems to be the perfect solution.

Figure 1.11 presents the four main topologies for a rotating transformer. For now, only one section of the transformer is represented. Readers can imagine the 3D representation as a revolution of this section around the axis. Each topology consists of the combination of a winding and a magnetic core placement. The winding placement is either adjacent (1.11a and 1.11b) or coaxial (1.11c and 1.11d). The magnetic core placement is either a pot-core (1.11a and 1.11c) or axial (1.11b and 1.11d).

First, the core placement is discussed with respect to the airgap position. Indeed, the airgap between the magnetic core parts has an important impact on the transformer efficiency. It is shown in [1] that the efficiency is higher for a shorter airgap. Therefore, it is important to minimize the airgap length. However, this airgap length is limited by the manufacturing and assembly possibilities. Indeed, machining has a certain degree of precision that is reduced for fragile materials as magnetic cores. An axial placement requires to deal with an airgap in the radial direction. It is complicated to reduce the length below 0.5 mm for radial airgaps. Actually, it is mechanically easier to deal with an airgap in the axial direction such as for pot-cores. Indeed, it is possible to use two pot-cores and to let those slip on each other. In this case, a short airgap length can be reached (< 0.1 mm). Regarding thermodynamic effects, the pot-core

is also better. Indeed, if the core heats up, it expands. For a too low radial airgap, an expansion of the rotor can cause a mechanical blocking.

The winding placement is also important. The best placement for a high frequency transformer is the coaxial one. This is due to the effect of high frequencies on the winding losses. This will be explained in the winding losses modelling section (section 2.3). For now, it is asked to the readers to trust this assumption.

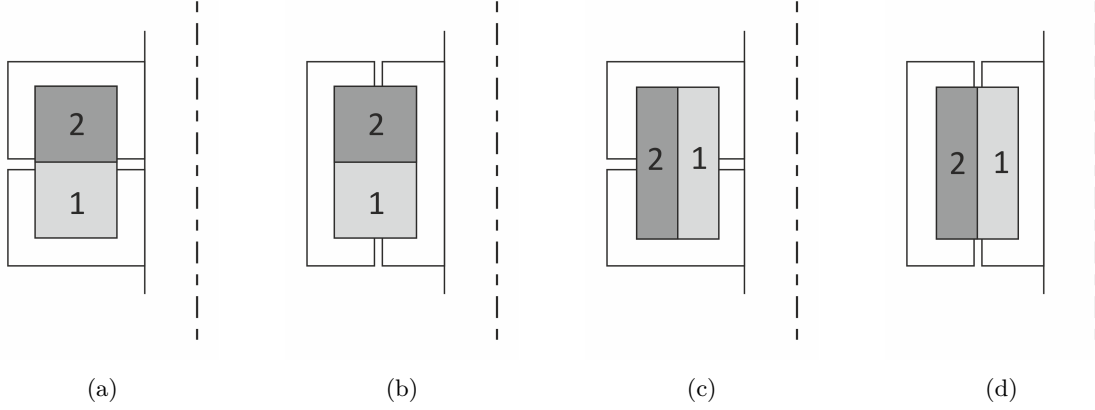


Figure 1.11 – Transformer topologies - Winding placement/Core placement : a) Adjacent/Pot Core; b) Adjacent/Axial; c) Coaxial/Pot Core; d) Coaxial/Axial

Finally, the best topology is the combination of the coaxial windings and the pot-core placement (Figure 1.11c). However, the airgap position is not fixed. Indeed, the two airgaps do not need to be in the same horizontal plan as in figure 1.11c. Considering the thermal management, it is necessary to maximize the conducting surface between each winding and its associated magnetic core. Indeed, the heat generated in the windings is drained by the magnetic core. A larger exchanging surface allows a more efficient cooling of the windings.

The best topology geometry for the rotating transformer is depicted in figure 1.12. Each airgap is located at the top of a winding. This maximizes the interface area between each winding and its associated magnetic core. The transformer core is composed of two coaxial tubes stuck between two plates. The rotor core is composed of a plate and the inner tube while the stator core is composed of the other plate and the outer tube. As it can be seen, eight dimensions characterize the transformer geometry. Amongst those, R is the shaft radius and e is the airgap length. The transformer outer radius (R_o) and height (H) expressions are :

$$\begin{aligned} R_o &= R + L_3 + 2 * L_4 + L_5 + L_6 \\ H &= 2 * L_1 + L_2 + e \end{aligned} \quad (1.5)$$

Relations can be made to link the different dimensions. First, it is supposed that both windings have the same cross section area (S_{wind}):

$$L_2 = \frac{S_{wind}}{L_4} \quad (1.6)$$

More relations can be found for the core dimensions. The idea is to keep constant the minimal core cross section along the magnetic flux path. Therefore, the cross section of the inner and outer tubes are equal to S_{core} . In the plates, the cross section is growing with the radius. In order to respect the rule, the plate

cross section is equal to S_{core} at radius ($r = R + L_6$). The following relations are then found :

$$\begin{aligned} L_6 &= \sqrt{\frac{S_{core}}{\pi} + R^2} - R \\ L_3 &= R_o - \sqrt{R_o^2 - \frac{S_{core}}{\pi}} \\ L_1 &= \frac{S_{core}}{2\pi(L_6 + R)} \end{aligned} \quad (1.7)$$

Those relations will be used for the optimization of the transformer in chapter 3.

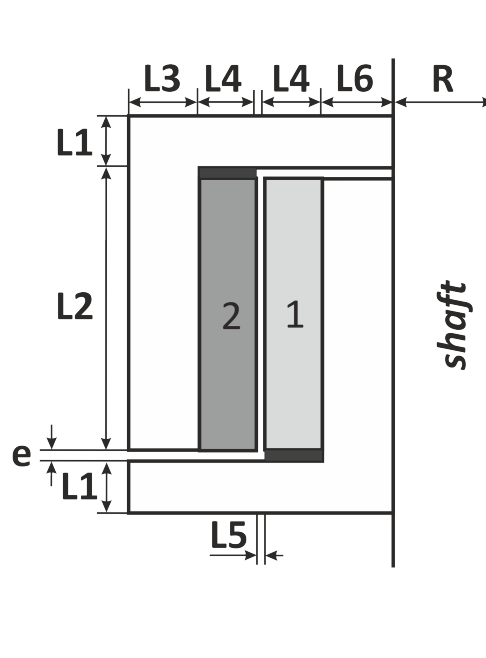


Figure 1.12 – Transformer geometry : representation and dimensions

1.4 Power Electronics for Inductive Power Transfer

An Inductive Power Transfer requires to operate with alternating currents. This implies to use a DC/AC converter and an AC/DC converter. Both of those can be gathered into a Switching-Mode Power Supply which is a DC/DC converter. The transformer of the IPT is located in this converter. Moreover, the converter has to be bidirectional for the BCDR functionality. A bidirectional DC/DC converter is introduced in this section. This topology is called the *Dual Active Bridge* (DAB) and consists in two *Full-bridges* disposed around the transformer. First, an introduction is done on the electronic switches and the full-bridge. And then, the DAB principle is briefly explained and the main equations involved are expressed.

1.4.1 Electronic switches

Electronic switches are the principal components of a SMPS. Those give the opportunity to control the power flows with a PWM control. However, those switches generate a major part of the converter losses. Electronic switches are either *active* or *passive* [15] [16].

Active switch

An active switch is controlled by an input terminal. The *on*-state or *off*-state of the switch are controlled by the signal on this terminal. There are several active switches such as MOSFET, IGBT, etc. The most

adapted active switch for spatial application is the MOSFET transistor. Indeed, it allows to use high frequency without generating high losses and those transistors can manage the power requested.

An active switch conducts current in both direction during *on*-state. In the ideal case, there is no voltage drop across the active switch during *on*-state (Figure 1.13b). However, in reality there always exists an internal resistance in the active switch. In a MOSFET, that resistance is called $R_{ds,on}$ and it generates a voltage drop during *on*-state mode (Figure 1.13c). It is possible to reduce the conduction resistance by connecting several transistors in parallel². The voltage drop during *on*-state is :

$$V_{on,MOS} = R_{ds,on} \cdot I_{on} \quad (1.8)$$

During the *off*-state, no current can flow through the switch. This switch can withstand an *off*-state voltage until the voltage rating. The internal structure of the switch breaks down beyond this maximum voltage causing irreversible damage.

In this project, the technology used for the active switches are SiC MOSFET (Datasheet available in Appendix B.3)

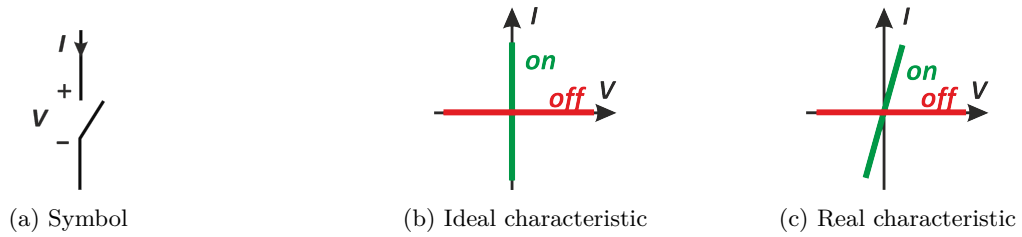


Figure 1.13 – Active switch : symbol, ideal characteristic and real characteristic

Passive switch

A passive switch has no input terminal. Therefore, that is the external circuit that controls the *conducting* or *blocking* states of the switch. The typical passive switch component is the diode.

A diode can drive the current in one direction only. In the ideal case, a current flows through the switch when the voltage across it is zero (Figure 1.14b). However, in reality, the voltage difference needs to be higher than the threshold voltage (V_{th}) of the diode which is around 1-3V. Moreover, diodes have an internal resistance as for transistors. (Figure 1.14c). The voltage drop in conducting mode of the diode is :

$$V_{on,D} = V_{th} + R_{on} \cdot I_{on} \quad (1.9)$$

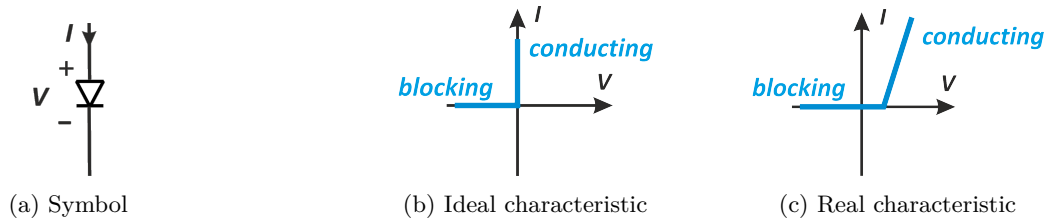


Figure 1.14 – Passive switch : symbol, ideal characteristic and real characteristic

Conduction and Switching Losses of Power Switches

In all kind of electronic switches, losses are generated. In a diode, most part of the losses are conduction losses which correspond to the losses generated at steady-state when the switch is conducting. A current is flowing while a voltage drop is carried. The power losses of a diode is :

$$P_{diode} = V_{on} I_{on} \cdot D \quad (1.10)$$

²However, parallel transistors have a higher parasitic capacitance and implies a higher cost and higher volume

where $D = \frac{t_{cond}}{T}$ is the duty-cycle, the time ratio during which the diode conducts current.

The major losses happening in a transistor are conduction losses and switching losses. Switching losses happen when the transistor is turned-on and turned-off. Because of parasitic capacitances, the switching time is not null and can last nano to microseconds. During these transitions, the switch conducts the entire load current and meanwhile carry a voltage drop corresponding to the *off*-state voltage (Figure 1.15). Therefore, a small amount of energy is lost during each transition. This can result in huge losses for high frequencies. The total losses generated in a MOSFET used in a switch mode is :

$$P_{MOSFET} = V_{on} I_{on} \cdot D + \left(V_{off} I_{on} \cdot \frac{t_{on} + t_{off}}{2} \cdot f \right) \quad (1.11)$$

where t_{on} and t_{off} are respectively the turn-on and turn-off delays and f is the operating frequency.

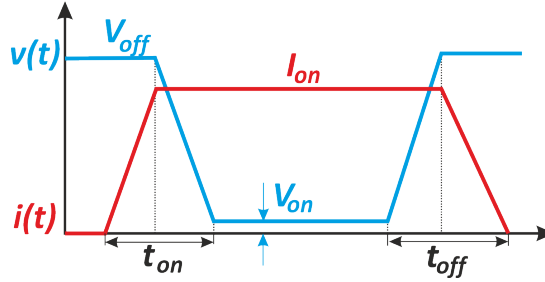


Figure 1.15 – MOSFET losses : switching and conducting losses

1.4.2 Half-bridge and Full-bridge converters

A *half-bridge* converter is a device composed of two active switches connected in series and supplied by a DC source (Figure 1.16a). The working principle of the *half-bridge* is to generate an AC voltage output by activating the switches one-by-one. The output is a square wave voltage with frequency and duty-cycle fixed by the switches control. The output signal varies from 0V (when S_2 is activated) to V_{dc} (when S_1 is activated)(Figure 1.16a). It is important to not activate both switches simultaneously in order to avoid a short circuit of the DC supply.

A *full-bridge* converter is similar to the *half-bridge* but it has a second leg in order to increase the output signal amplitude (Figure 1.16b). In this case, the switches are activated two-by-two. The output signal varies from V_{dc} (when S_1 and S_4 are activated) to $-V_{dc}$ (when S_2 and S_3 are activated). It is also possible to reach 0V by activating either S_1 and S_3 or S_2 and S_4 . However, this mode is not used in this document. The mode depicted in figure 1.16b is called *full wave mode*.

For this document, the duty-cycle of a full-bridge is considered to be 0.5. That means that the average output voltage of each full-bridge is equal to zero.

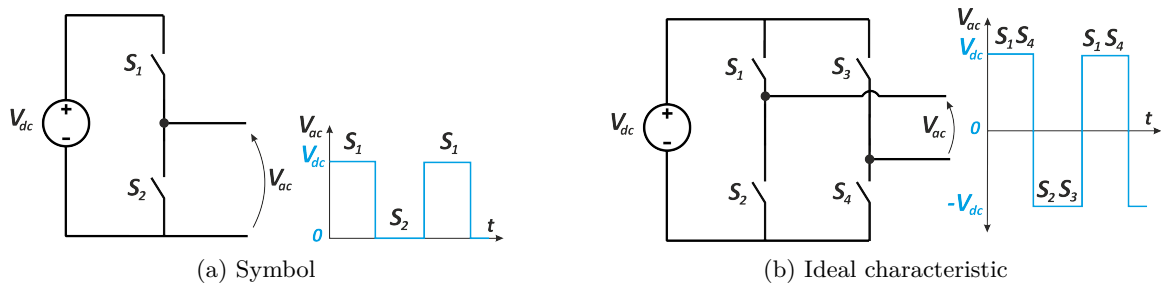


Figure 1.16 – Power Electronics : Half-bridge and Full-bridge

1.4.3 Inductive Transformer Model

A transformer can be modelled by an equivalent circuit. For the explanation of the *Dual Active Bridge* principle, a first model of the transformer is introduced. This is called the inductive transformer model because it considers the inductive components only and the transformer is supposed to be loss free (Figure 1.17a)³. Indeed, for high frequencies, resistances are not significant with respect to reactances. The transformer consists of a quadripole with the leakage inductance in series (L_f) and the magnetizing inductance in parallel (L_μ). However, the magnetizing inductance is often much larger than the leakage inductance and the load because of the magnetic core permeability. A simplification can be made by neglecting the magnetizing inductance. Therefore, the inductive model of the transformer consists of a simple series inductance corresponding to the leakage inductance (Figure 1.17b).

As a reminder, the definition of an inductance is :

$$L = \frac{\Phi}{i} \quad (1.12)$$

The relation between voltage across the inductance and the current carried is :

$$v(t) = L \frac{di(t)}{dt} \quad (1.13)$$

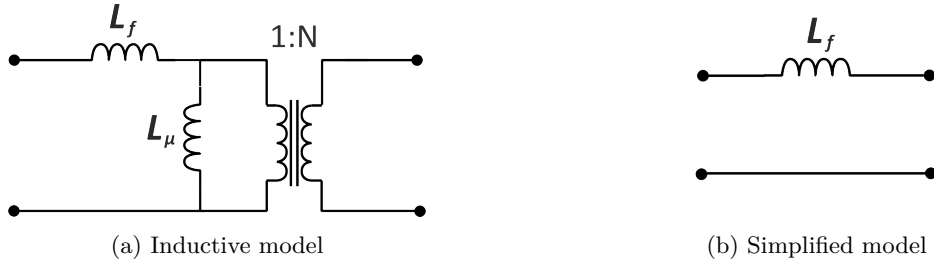


Figure 1.17 – Transformer inductive models

1.4.4 Dual Active Bridge

The Dual Active Bridge (DAB) is an isolated bidirectional DC-DC converter which consists of a transformer and two full-bridges [17][18]. A first full-bridge is connected to the primary side of the transformer and the second is connected to the secondary side. The transformer provides the isolation. The working principle is to store energy instantaneously in the transformer series inductance (L) which is, generally, the transformer leakage inductance itself. Figure 1.18 represents the DAB schematic used between the solar array which is a DC-current source and the load which is the bus ($V_{out} = V_{bus}$).

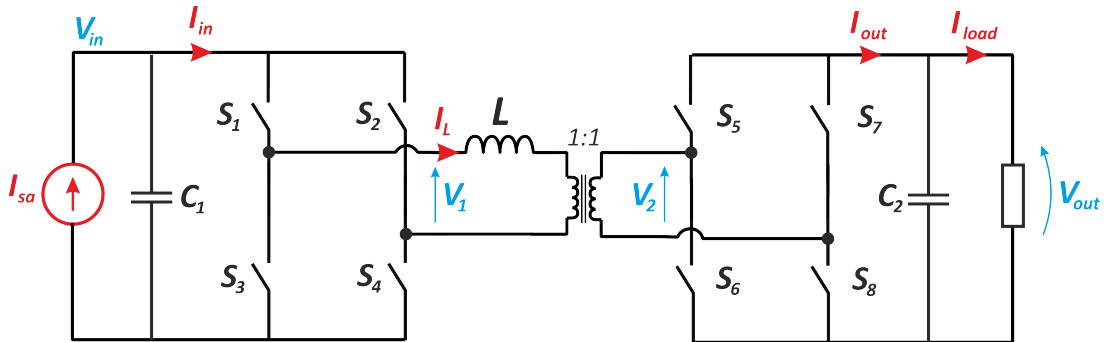


Figure 1.18 – Dual Active Bridge : circuit schematic

³A complete model considering winding and core losses will be introduced in chapter 2

Signals and Power Transferred

In this part, the signals involved in the DAB are explained and the power transferred is calculated. The currents waveform is important. Indeed, depending on the current waveform, the harmonics can have a significant impact on the different losses generated in the transformer. The purpose of this section is not to provide an advanced control of the DAB. This is why the equation derivation is made for a basic case where $V_{in} = V_{out}$ and the transformer has a unity ratio. Figure 1.19 represents the voltages and currents in the converter.

First, $V_1(t)$ and $V_2(t)$ are defined as the output voltage of the two full-bridges. As explained before, the two voltages have a 50% duty-cycle ($D_1 = D_2 = 0.5$) and those signals vary respectively between $[-V_{in}; V_{in}]$ and $[-V_{out}; V_{out}]$. As it can be seen in figure 1.19, there is a phase-shift between $V_1(t)$ and $V_2(t)$ which is expressed as $D = \frac{\phi}{\pi}$. The inductance voltage is equal to the difference between $V_1(t)$ and $V_2(t)$:

$$V_L(t) = V_1(t) - V_2(t) \quad (1.14)$$

Because of the phase-shift, the inductance voltage is not null all the time (Figure 1.19). As $V_{in} = V_{out}$, $V_L(t)$ is null during $(1 - D)T$ of a period.

With the inductance voltage $V_L(t)$ and equation (1.13), the current flowing through the inductance $I_L(t)$ can be calculated. The current corresponds to the integration of the voltage and is a trapezoidal waveform with a duty-cycle equal to D^4 . The peak-to-peak value of the inductance current can be expressed as :

$$\Delta I_L = \frac{V_{in} + V_{out}}{L_s} \cdot \frac{DT}{2} = \frac{V_{in} + V_{out}}{2fL_s} \cdot D = \frac{V_{in}}{fL_s} \cdot D \quad (1.15)$$

The maximal inductance current is equal to the half peak-to-peak current :

$$I_{max} = \frac{\Delta I_L}{2} \quad (1.16)$$

The rms value of a trapezoidal waveform such as $I_L(t)$ is :

$$I_{L,rms}(t) = I_{max} \cdot \sqrt{1 - \frac{2D}{3}} \quad (1.17)$$

Afterwards, the inductance current $I_L(t)$ is rectified by the second full-bridge from which comes out the output current $I_{out}(t)$. The alternating part of $I_{out}(t)$ flows through the capacitor C_2 while the mean value flows through the DC load. The mean current can be calculated :

$$I_{load} = \bar{I}_{out} = I_{max} \cdot \frac{(1 - D)T}{T} = \frac{V_{in}}{2fL_s} \cdot D(1 - D) \quad (1.18)$$

Finally, the power transferred is evaluated. If no losses are considered in the different components, the output power is equal to the input power :

$$P_{out} = P_{in} = \frac{V_{in}^2}{2fL_s} \cdot D(1 - D) \quad (1.19)$$

The input and output capacitors (C_1 and C_2) are used to filter the alternating tension. The larger are the capacitors, the more efficient is the filter. However, a large capacitor reduces the converter time response for the control.

⁴The duty-cycle of a trapezoidal waveform corresponds to the time of ramp over the period : $\frac{t_{ramp}}{T}$

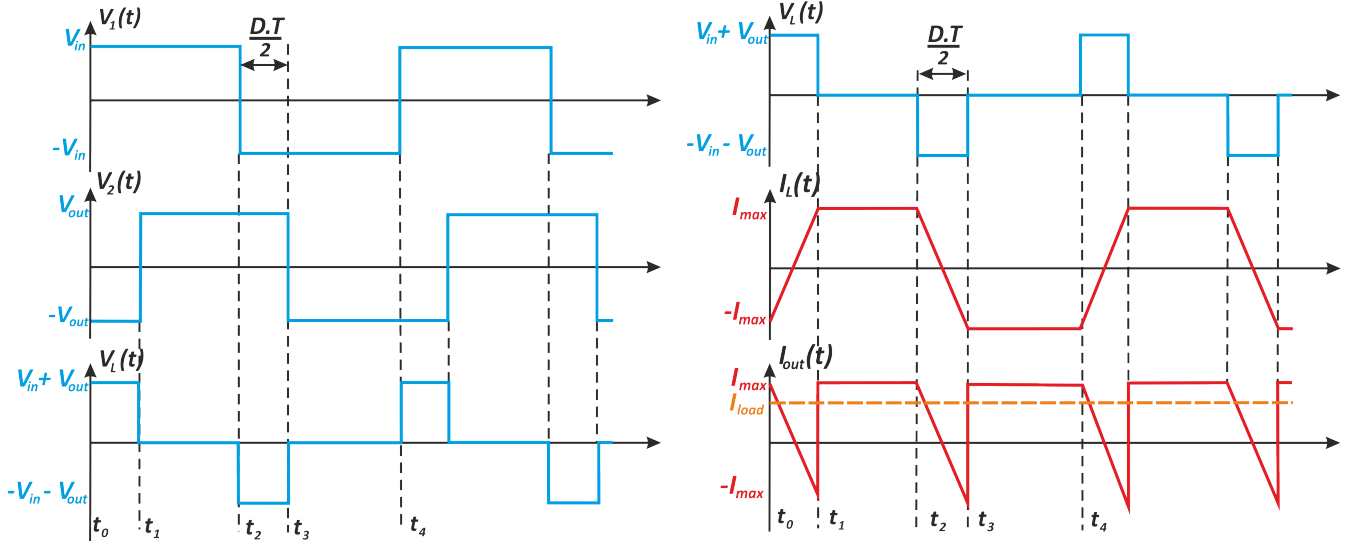


Figure 1.19 – Dual Active Bridge : voltage and current signals for $V_{in} = V_{out}$ and a unity transformer ratio

Phase-shift analysis

For the DAB full wave mode, the phase-shift between the bridges (D) is the only degree of freedom for the regulation of the transferred converter power. In this part, the impact of this variable on the DAB is analysed and an efficient rated value is searched.

By equation (1.19), it is tempting to choose a phase-shift $\phi = 90^\circ$ ($D = 0.5$) because it maximizes the active power transferred. However, reactive power (Q) is also involved in the DAB. The apparent power ($S = \sqrt{P^2 + Q^2}$) is related to the total current flowing through the inductance ($I_{L,rms}$) while the active power is related to the load current (I_{load}). The ratio between those currents is :

$$\frac{I_{L,rms}}{I_{load}} = \frac{\sqrt{1 - \frac{2D}{3}}}{1 - D} \quad (1.20)$$

This allows to calculate the apparent and reactive powers involved in the DAB :

$$S = P_{in} \cdot \frac{\sqrt{1 - \frac{2D}{3}}}{1 - D} \quad (1.21)$$

$$Q = \sqrt{S^2 - P_{in}^2} \quad (1.22)$$

The function of those different powers with respect to phase-shift (D) are depicted in figure 1.20. Each curve is normalized with respect to the maximal active power ($P_{in,max}$). As expected, the maximal active power corresponds to a phase-shift $D = 0.5$. However, the apparent power is relatively high for this point : $\frac{S}{P_{in}}(D = 0.5) = 1.63$. Too much current is flowing in the inductance (and transformer) for a given load current. The system is not efficient because more conducting losses are generated in the transformer.

Regarding figure 1.20, it could be interesting to have a minimal phase-shift ($D \rightarrow 0$) to minimize the reactive power (Q). However, when the phase-shift decreases, the trapezoidal waveform tends to a square waveform and the harmonic content increases (Appendix A.4). The impact of harmonics on the losses will be analysed in the transformer modelling (chapter 2). For now, it is just said that a larger harmonic content generates more losses.

Finally, a trade-off is made by choosing a phase-shift between the two extreme cases. The phase-shift chosen is $D = 0.25$ ($\phi = 45^\circ$). This is the value which is typically selected.

Note: the phase-shift which is calculated here is the rated phase-shift. During regulation, the phase-shift is not constant and varies around the rated value.

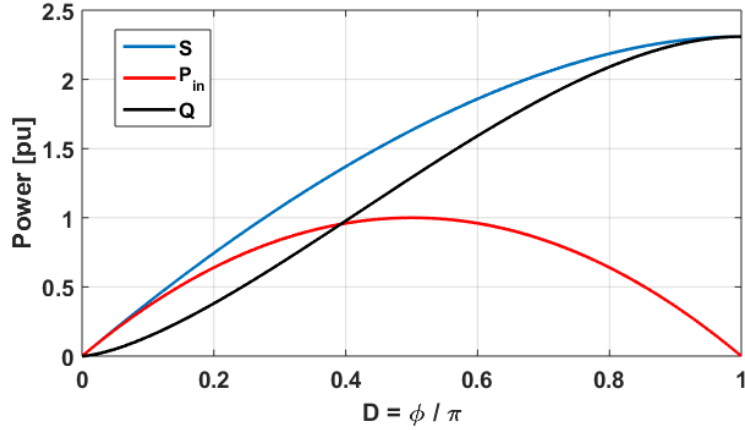


Figure 1.20 – Dual Active Bridge : apparent power, active power and reactive power with respect to phase-shift (normalized)

Losses analysis

The losses generated by the two full-bridges are mostly generated by the active switches. As explained before, there are two kind of switch losses : conducting and switching losses.

The conducting losses are generated when a current flows through the switch during *on*-state. At any moment, two switches of each full-bridge conducts the same current that is flowing through the inductance (I_L). Therefore, the total conducting losses are :

$$P_{cond} = 4 * R_{ds,on} I_{L,rms}^2 \quad (1.23)$$

where $R_{ds,on}$ is the *on*-state resistance of one switch.

The second losses are switching losses. Those losses happen while a switch turns on or turns off. In reality, a transistor has a parasitic capacitance between the drain and the source (C_{ds}). This capacitance implies that the voltage across the transistor cannot change instantaneously. Therefore, during a switching, both voltage and current are applied on the switch and a large amount of energy is lost. However, it is possible to cancel those losses by the use of a *Zero Voltage Switching* (ZVS) control. The ZVS principle consists of waiting a short time so that those parasitic capacitances get discharged. During this time called *dead-time* all the switches are off. The ZVS principle is explained more in appendix A.2. As a consequence, the total losses generated in the DAB are conducting losses only (1.23).

1.5 Specifications of the rotating transformer

In this chapter, the context within which the rotating transformer is involved was introduced. Now, the objectives of the project can be defined.

The project objective is to design an Inductive Power Transfer (IPT) for the Solar Array Drive Mechanism (SADM) of a satellite. The project aims to replace the current slip rings technology which is limited to a medium voltage use because of electric discharge (ESD). The IPT is composed of a rotating transformer and a Dual Active Bridge for the power control. The transformer topology and geometry are defined in order to allow a rotation of the rotor with respect to the stator and to optimize the transformer efficiency.

The specifications for the IPT are the following : the IPT-SADM has to be able to transfer a rated power of 10 kW under 300V with a total efficiency of 98%. 10 kW corresponds to the power delivered by one solar array. The efficiency represents an allowed total loss of 200W for the transformer and the Dual Active Bridge.

With equations from (1.15) to (1.23), and the specifications of the transformer, the main results of the Dual Active Bridge can be calculated. Those are presented in table 1.3. The switch resistance value ($R_{ds,on}$) is taken from a DAB prototype using two SiC MOSFETs in parallel (Appendix B.3). The conductive losses are evaluated to be 83 W. However, this value does not consider other losses such as gate drive losses. The total losses of the DAB can be estimated to 100 W. This corresponds to a 99% efficiency for the DAB only. The inductive reactance of the DAB is relatively high. Indeed, for a frequency of 100 kHz, the series inductance needed is $L_s = 8.5\mu H$. This value cannot be reached with the leakage inductance (L_f) of a transformer only. Therefore, an external inductor is connected in series of the transformer. To respect the total efficiency, the total power that can be lost in the transformer and the external inductor is 100W.

Characteristic			
Nominal power	P_{out}	10	kW
Input voltage	V_{in}	300	V
Output voltage	V_{out}	300	V
Load current	I_{load}	33.3	A
Phase-shift	D	0.25	
Inductive reactance	X_L	5.3	Ω
Inductive current	$I_{L,rms}$	40.5	A
Maximal current	I_{max}	44.6	A
Switch resistance	$R_{ds,on}$	12.5	$m\Omega$
Conductive losses	P_{cond}	83	W

Table 1.3 – Dual Active Bridge main characteristics

Now, the transformer is ready to be designed. This transformer has to be able to transfer 10 kW under 300V with a 99% efficiency. But, before the design, an advanced modelling of the transformer has to be done in order to get the "tools" that will help to evaluate the transformer losses and characteristics. Moreover, the transformer has to survive in the hostile environment that space is. Therefore, the thermal management needs to be investigated carefully.

Transformer Modelling

The objective of this chapter is to develop models for the important aspects of a transformer. Those models are tools that will help for the design of an efficient and optimized transformer. But, before going further, let us begin with a reminder of the transformer principle.

The main goal of a transformer is to transfer power from a primary side to a secondary side. This is achieved by the magnetic coupling of the primary and secondary windings. Figure 2.1a represents the equivalent electric circuit of a standard transformer [19]. Resistors R_1 and R_2 are respectively the primary and the secondary winding resistance. Inductors $L_{f,1}$ and $L_{f,2}$ represent the leakage inductance of the primary and secondary winding respectively. L_μ is the magnetizing inductance of the transformer and R_{ml} is the equivalent resistance representing the magnetic losses taking place in the magnetic core. The transformer turns ratio N is the characteristic that defines the induction relation in the transformer¹:

$$N = \frac{n_1}{n_2} = \frac{V_1}{V_2} = \frac{I_2}{I_1 - I_\mu} \quad (2.1)$$

In the specifications, it is expressed that the transformer has a unity turns ratio ($N = 1$). The magnetizing current I_μ corresponds to the current flowing in the magnetizing inductance. In this application, this current is not important and it generates losses in the primary winding resistance. Therefore, a first objective is to reduce this current by increasing the magnetizing inductance. However, the magnetizing current can help for the Zero Voltage Switching mode at low power (Appendix A.2).

Two kinds of losses are generated in a transformer : winding losses related to R_1 and R_2 , and core losses related to R_{ml} . In order to design a transformer as efficient as possible, both loss components have to be minimized. To do so, the best way is to understand their origin and to study what factors influence them.

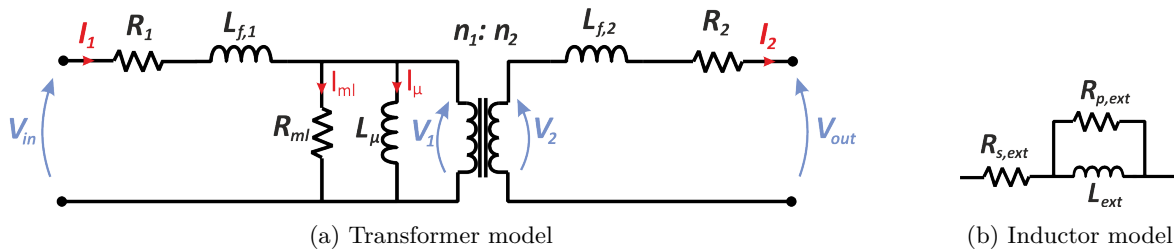


Figure 2.1 – Representation of the equivalent circuit models for a standard transformer and an external inductor

Previously, it was said that the *Dual Active Bridge* is a switch-mode power supply where energy is stored in the series inductance. However, the inductance energy needed is too important and the leakage

¹The resistance R_{ml} is really high and the current flowing in it can be neglected

inductance of the transformer is not enough. The solution is to add an external inductor in series with the transformer. In that case, a small part of the energy is stored in the transformer leakage inductance and most of the energy is stored in the additional inductor. However, the use of an external inductor has the disadvantage to bring more copper and core losses in the system. Figure 2.1b depicts the equivalent model of such an inductor. In this case, winding losses are represented by an equivalent series resistance $R_{s,ext}$ and the core losses are represented by equivalent parallel resistance $R_{p,ext}$. The equivalent model of the transformer and the external inductor can be gathered in one single equivalent model (Figure 2.2). The resistances related to the core losses are connected in parallel and the other components of the external inductor are connected in series to the transformer.

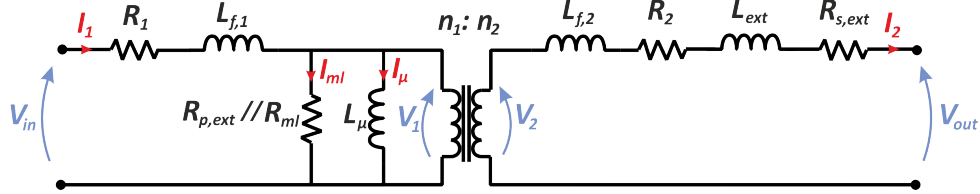


Figure 2.2 – Equivalent circuit model of combination of the transformer and the external inductor

In this chapter, each component of the transformer and inductor models are analysed and calculated. First, the magnetic components are represented with the magnetic circuit model. With this model, the equations of each inductance involved in the transformer circuit are derived. Afterwards, the core losses are explained and an adaptation of the *Steinmetz equation* is applied for the trapezoidal and triangular waveforms involved in the magnetic cores. A comparison of different core materials is also made. After, the winding losses are analysed with a special focus on the high frequency effect. The Litz wire is then introduced to face the problems generated by high frequency on the winding losses. And last but not least, a model is proposed for the thermal management of the transformer. Figure 2.3 has already been introduced in chapter 1 and is repeated here as a reminder of the transformer geometry and dimensions.

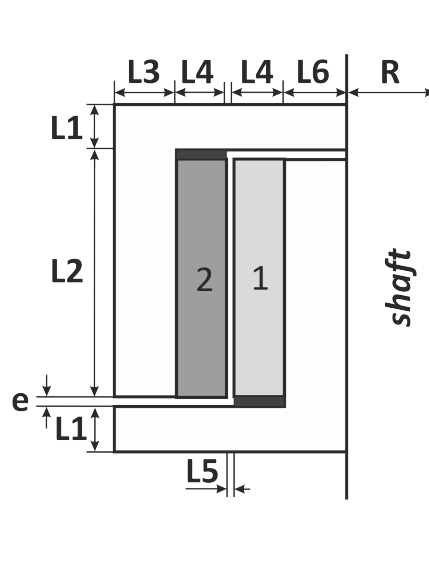


Figure 2.3 – Transformer modelling : main dimensions

2.1 Magnetic circuit modelling

The design of a transformer or an inductor often requires a core made of a magnetic material. Those are used in order to concentrate the magnetic flux, lead it through the windings and reduce the leakage flux. Each core material has its own characteristics depending on its constitution.

2.1.1 B-H curve

An important relation in a magnetic material is the B-H curve. It summarizes the main characteristics of a core and is really useful to see its behaviour. But first, a definition of each variable B and H in the case of winding induction needs to be done [20] [21].

- The magnetic flux density (**B**) is determined by the induction voltage (v) that induces a magnetic flux (ϕ) linking the n -loop windings and the surface area of the winding (A_c).

$$v(t) = n \cdot \frac{d\phi(t)}{dt} = \frac{d\Phi(t)}{dt} \quad (2.2)$$

$$\phi = \int_{A_c} \mathbf{B} \cdot d\mathbf{A} \quad (2.3)$$

Considering a uniform distribution of the magnetic flux over the cross section, the flux density can be expressed as :

$$\mathbf{B}(t) = \frac{n}{A_c} \int v \, dt \quad (2.4)$$

- The magnetizing field (**H**) is determined with the *Ampere's law* stating that the total MMF (*magnetomotive force*) around a close path is equal to the current passing through the inner area of this path :

$$\oint_{l_{path}} \mathbf{H} \cdot d\mathbf{l} = \int_{S_{path}} \mathbf{J} \cdot d\mathbf{A} \quad (2.5)$$

Assuming that the current density (**J**) is constant over the surface and separating the path by material, the previous equation can be simplified as :

$$\sum_j \mathbf{H}_j \cdot \mathbf{l}_j = ni \quad (2.6)$$

The relationship between both the flux density (**B**) and the magnetizing field (**H**) varies from one material to the other. This relation can be expressed as follows :

$$\mathbf{B} = \mu_0 \mu_r(H) \mathbf{H} \quad (2.7)$$

where μ_0 is the vacuum permeability and μ_r is the relative permeability of the material. The relative permeability is non-linear because of saturation but it can be seen as a composition of two linear curves : before and after saturation. Figure 2.4 depicts the B-H curves of air and a magnetic material. It can be seen that a magnetic material has a stiffer B-H curve than the air ². This is due to the high permeability of the magnetic material. In order to keep the relation linear and avoid signal distortion, the maximum flux density (B_{max}) allowed in the core is limited to 75% of the saturation :

$$B_{max} \leq 0.75 \cdot B_{sat} \quad (2.8)$$

The surface area located below the B-H curve corresponds to the magnetizing energy density:

$$u_{mag} = \int B dH \quad (2.9)$$

²The curves represented in figure 2.4 are not scaled in order to have a good visibility. Indeed, in reality, there is a 1000 factor between the curves slope.

In the linear part of the B-H curve, the energy density can be expressed as :

$$u_{mag} = \frac{1}{2} \frac{B^2}{\mu_0 \mu_r} \quad (2.10)$$

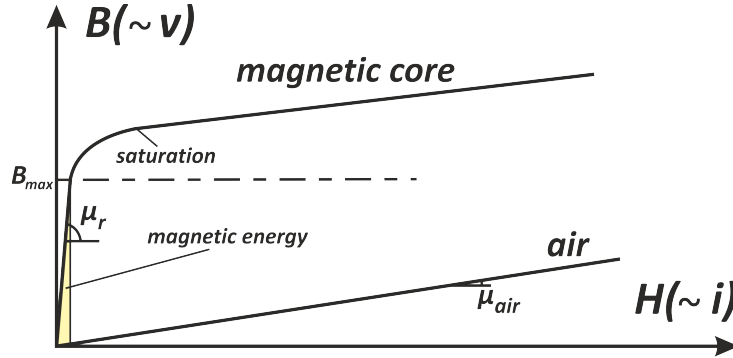


Figure 2.4 – B-H curve for a magnetic material and air

The goal of the transformer is to transfer power from the primary winding to the secondary winding. In the ideal case, no magnetizing energy would be necessary for this operation. However, this is impossible because it would imply a zero flux density (2.10) and a zero voltage (2.2). Therefore, the best is to minimize the magnetizing energy by using a material with a high relative permeability (μ_r). This corresponds to the case of the magnetic core in figure 2.4.

2.1.2 Reluctance model

Another way to deal with the different magnetic variables B and H and their relation inside a transformer or an inductor is to use the reluctance model. The idea is to make an analogy between electric circuits and the transformer/inductor equivalent magnetic circuit [21].

This analogy begins early in the electromagnetic concepts where the *magnetizing field* (H) is equivalent to the *electric field* (E) and the *magnetic flux density* (B) is equivalent to the *electric current density* (J). From those fundamental equivalences, other equivalences between electric and magnetic relations can be made. This finally leads to equivalences between electric and magnetic circuits (Table 2.1). In the magnetic circuit, the MMF is equivalent to a voltage supply in the electric circuit. Similarly, the magnetic flux is equivalent to the electric current. The *Hopkinson's law* is a relation between MMF and magnetic flux which corresponds to the *Ohm's law* in electricity.

Magnetic	Electric
Magnetizing field : H	Electric field : E
Magnetic flux density : B	Current density : J
MMF : $\mathcal{F} = \int_{P_1}^{P_2} \mathbf{H} d\mathbf{l} = n \cdot i$	Voltage : $V = \int_{P_1}^{P_2} \mathbf{E} d\mathbf{l}$
Magnetic flux : $\phi = \frac{\Phi}{N} = \int_A \mathbf{B} d\mathbf{A}$	Current : $I = \int_A \mathbf{J} d\mathbf{A}$
Hopkinson's law : $\mathcal{F} = \mathcal{R}\phi$	Ohm's law : $V = RI$

Table 2.1 – Hopkinson equivalences

The *Hopkinson's law* introduces the concept of reluctance ($\mathcal{R}$) which can be described as the ability

of a material to oppose a magnetic field :

$$\mathcal{R} = \frac{\mathcal{F}}{\phi} = \frac{\int \mathbf{H} \cdot d\mathbf{l}}{\int \mathbf{B} \cdot d\mathbf{A}} \quad (2.11)$$

Likewise the resistance, the reluctance only depends on geometry and material properties.

Reluctance can be easily calculated for a uniform magnetic flux distribution over a cross section. The reluctance expression depends on the magnetic flux direction³ :

- For uniform axial flux :

$$\mathcal{R}_{cst,axial} = \frac{l}{\mu \cdot A} \quad (2.12)$$

where l is the axial length and A is the cross section.

- For uniform radial flux :

$$\mathcal{R}_{cst,radial} = \frac{\ln(r_o/r_i)}{\mu \cdot 2\pi \cdot h} \quad (2.13)$$

where r_i and r_o are respectively the inner and outer radius and h is the axial length.

Finally, there is a link between reluctance and inductance expressions :

$$L = \frac{\Phi}{i} = \frac{n \cdot \phi}{\mathcal{F}/n} = \frac{n^2}{\mathcal{R}} \quad (2.14)$$

This relation explains why the concept of reluctance was introduced. In order to get the inductance value, the method is to calculate the total reluctance of a magnetic circuit and to compute the inductance associated with equation (2.14).

2.1.3 Magnetizing inductance (L_μ)

The magnetizing inductance represents the magnetizing energy stored in the core of the transformer. As said previously, this magnetizing energy has to be minimized. Moreover, the magnetizing current contributes to winding losses in the primary winding. Thus, it is important to limit this current by having a large magnetizing inductance. Figure 2.5a depicts the magnetic circuit of the transformer with the reluctance network and the MMF sources corresponding to the primary and secondary windings. For this magnetic circuit, it is assumed that all the magnetic flux generated is contained in the core and airgaps. The leakage flux located in the winding window and the fringing effect⁴ of the airgaps are neglected. The *Hopkinson's law* provides the expression of the magnetizing current i_μ :

$$\mathcal{F} = n \cdot (i_1 - i_2) = n \cdot i_\mu = \mathcal{R}_\mu \cdot \phi_{core} \quad (2.15)$$

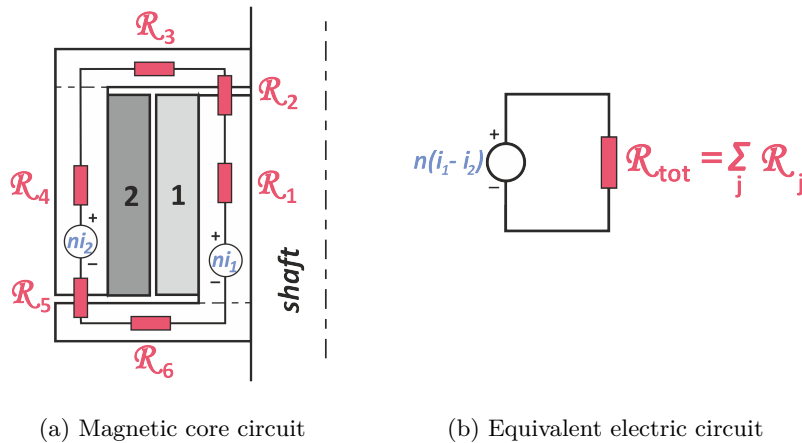


Figure 2.5 – Transformer magnetic core circuit and the equivalent circuit

³Those expression are calculated in Appendix A.5.

⁴This concept is explained in section 3.5.1

Reluctances calculation

Figure 2.5b represents the electric circuit that is equivalent to the magnetic circuit of figure 2.5a. The equivalent reluctance corresponds to the sum of the series reluctance of figure 2.5a. In the magnetic circuit, the flux flows in axial direction for $\mathcal{R}_1$, $\mathcal{R}_2$, $\mathcal{R}_4$ and $\mathcal{R}_5$ and it flows in radial direction for $\mathcal{R}_3$ and $\mathcal{R}_6$. As said previously, the cross section area of the inner and outer tubes of the transformer are equal and both plates are identical. Using (2.12) and (2.13), the reluctances expressions are ⁵ :

$$\begin{aligned}\mathcal{R}_1 = \mathcal{R}_4 &= \frac{1}{\mu_0 \mu_r} \cdot \frac{L_2}{\pi[(L_6 + R)^2 - R^2]} \\ \mathcal{R}_2 = \mathcal{R}_5 &= \frac{1}{\mu_0} \cdot \frac{e}{\pi[(L_6 + R)^2 - R^2]} \\ \mathcal{R}_3 = \mathcal{R}_6 &= \frac{1}{\mu_0 \mu_r} \cdot \frac{\ln(\frac{D_o}{2}/R)}{2\pi L_1}\end{aligned}\quad (2.16)$$

The magnetizing inductance can then be expressed with the number of loops and the reluctances with (2.14) :

$$L_\mu = \frac{\Phi_{tot}}{i_\mu} = \frac{n^2}{\mathcal{R}_\mu} = \frac{n^2}{2 * [\mathcal{R}_1 + \mathcal{R}_2 + \mathcal{R}_3]} \quad (2.17)$$

In order to minimize the magnetizing current, the magnetizing inductance needs to be high which means as many turns as possible and a total reluctance as low as possible. As the total reluctance is dominated by the airgap reluctance, it is important to decrease it by taking an airgap as thin as possible.

2.1.4 Leakage inductances ($L_{f,1}$ and $L_{f,2}$)

The leakage inductance is associated to the leakage flux generated by the windings. Indeed, the entire flux generated is not only composed of the magnetizing flux located in the core. As it can be seen in figure 2.6a, a small amount of flux called leakage flux is passing through the transformer window where the windings are located. This flux is much lower than the magnetizing flux because of the low permeability of air and copper ($\mu_{air} = \mu_{copper} = \mu_0$). This leakage inductance can be evaluated with the reluctance model as well. For this, the reluctance of the transformer window needs to be defined. However, this is not as simple as for the magnetizing reluctance where the MMF is considered as uniform over the core section. In the case of the window, the MMF is varying with radial position (Figure 2.6a). Actually, the window MMF is composed of three different parts : primary winding, secondary winding and the winding-gap.

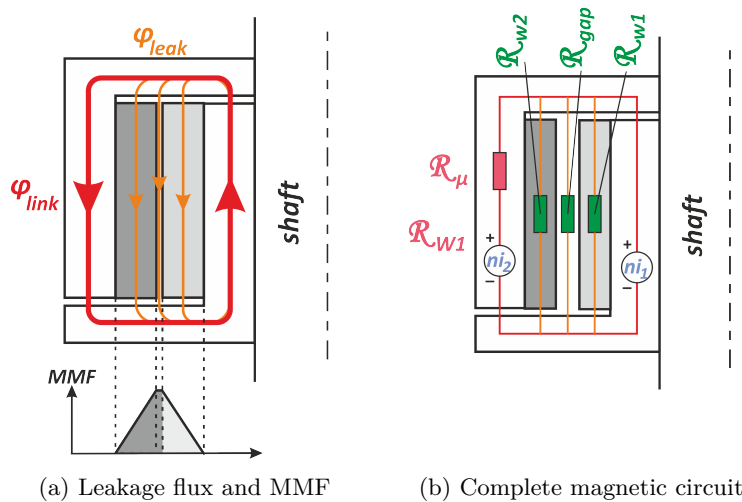


Figure 2.6 – Representation of the window leakage flux and the equivalent reluctances

⁵All the dimensions are represented in figure 2.3

Reluctance calculation

As the MMF is constant in the winding-gap, its associated reluctance can be expressed with the standard reluctance formula :

$$\mathcal{R}_{gap} = \frac{1}{\mu_0} \cdot \frac{L_2}{\pi[(R + L_6 + L_4 + L_5)^2 - (R + L_6 + L_4)^2]} \quad (2.18)$$

For the leakage reluctance of windings, the MMF is not uniform. It increases/decreases linearly with radial position (Figure 2.6a). Therefore, equation (2.12) cannot be used for the reluctance calculation. The expression of reluctance with linear MMF is derived in Appendix A.5. The result is that a reluctance with linear MMF is three times the reluctance with uniform MMF:

$$\mathcal{R}_{lin,axial} = 3 * \frac{l}{\mu \cdot A} = 3 * \mathcal{R}_{cst,axial} \quad (2.19)$$

The winding reluctances can then be expressed as:

$$\begin{aligned} \mathcal{R}_{w1} &= \frac{3}{\mu_0} \cdot \frac{L_2}{\pi[(R + L_6 + L_4)^2 - (R + L_6)^2]} \\ \mathcal{R}_{w2} &= \frac{3}{\mu_0} \cdot \frac{L_2}{\pi[(R + L_6 + 2 \cdot L_4 + L_5)^2 - (R + L_6 + L_4 + L_5)^2]} \end{aligned} \quad (2.20)$$

Finally, the leakage inductance can be computed by connecting the reluctances in parallel :

$$L_f = L_{f,1} + L_{f,2} = \frac{n^2}{\mathcal{R}_{w1} // \mathcal{R}_{w2} // \mathcal{R}_{gap}} = n^2 \cdot \left(\frac{1}{\mathcal{R}_{w1}} + \frac{1}{\mathcal{R}_{w2}} + \frac{1}{\mathcal{R}_{gap}} \right) \quad (2.21)$$

Individual winding leakage inductances are :

$$\begin{aligned} L_{f,1} &= \frac{n^2}{\mathcal{R}_{w1}} \\ L_{f,2} &= \frac{n^2}{\mathcal{R}_{w2} // \mathcal{R}_{gap}} = n^2 \cdot \left(\frac{1}{\mathcal{R}_{w2}} + \frac{1}{\mathcal{R}_{gap}} \right) \end{aligned} \quad (2.22)$$

From those equations, it can be seen that the leakage inductance increases with the window width ($2 * L_4 + L_5$) and even more with the winding-gap width (L_5). However, it decreases with an increasing window height (L_2).

2.1.5 External inductor (L_{ext})

In opposition to transformers, inductors are used to store magnetic energy. By looking at equation (2.10), the best way to store a high amount of energy is to use air for the core material ($\mu_r = 1$). However, as it can be seen in figure 2.7, an air core requires an important magnetizing current (H) for a high amount of energy. That implies a large number of turns (n) for a given current (i). An increasing number of turns will proportionally increase the winding resistance and the losses associated.

On the one hand, the use of magnetic material implies a small amount of energy. On the other hand, the use of air implies high copper losses. A trade-off can be found either by using a magnetic material with low permeability as powders or by inserting an airgap in a high permeability material (Figure 2.1b). These two methods are equivalent. Indeed, an equivalent permeability of an airgapped core can be expressed by using the *reluctance model*. For an airgapped core design, the easiest way is to use two E cores (Fig 2.9c) facing each other (Figure 2.1b). In this kind of geometry, the rectangular core section (A_{core}) is kept constant over the entire flux path. From the reluctance network of the E core inductor, an equivalent reluctance can be expressed :

$$\begin{aligned} \mathcal{R}_e &= 2 * (\mathcal{R}_{core} + \mathcal{R}_g) \\ \frac{l_e}{\mu_e \mu_0 \cdot A_{core}} &= \frac{2 * l_{core}}{\mu_r \mu_0 \cdot A_{core}} + \frac{2 * l_g}{\mu_0 \cdot A_{core}} \end{aligned} \quad (2.23)$$

where l_{core} and l_g are respectively the flux path length in the core and the airgap, l_e is the entire path ($l_e = l_{core} + l_g$), μ_r is the core magnetic permeability and μ_e is the equivalent magnetic permeability of the magnetic circuit such as :

$$\mu_e \simeq \frac{\mu_r}{1 + \mu_r l_g / l_e} \quad (2.24)$$

This solution gives the opportunity to tune the equivalent permeability by acting only on the airgap length. Figure 2.7b is a comparison of B-H curves for a magnetic core, an air-core and an airgapped core. It can be seen that the airgapped core curve is located between the two others and a higher amount of energy can be stored for a reasonable magnetizing field.

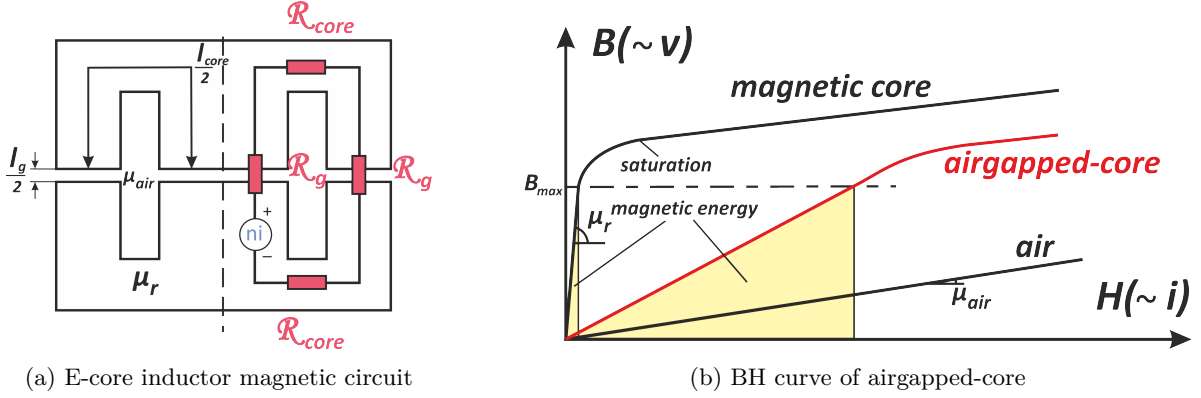


Figure 2.7 – Airgapped core equivalent permeability

Key points

In this section, the *reluctance model* was introduced. It helps to evaluate the different inductances involved in the transformer :

- For transformers, the magnetizing energy needs to be as low as possible meaning a magnetizing inductance (L_μ) as high as possible. However, a minimal value for the magnetizing current I_μ is interesting for the ZVS mode at low power transfer.
- An airgap has a high reluctance because of low permeability. Therefore, it decreases the magnetizing inductance and increases the magnetizing energy.
- For a transformer, it is important to have an airgap as thin as possible in order to minimize this magnetizing energy.
- Leakage inductance can also be computed with the *reluctance model*. It increases with the window width ($2 * L_4 + L_5$) and decrease with the window height (L_2).
- For low frequency, the leakage inductance is not enough. Then, an external inductor needs to be designed to store energy. The most efficient way is to use an airgapped-core such as *E-cores*.

2.2 Core loss modelling (R_{ml} and $R_{p,ext}$)

The disadvantage of using a magnetic core for a transformer or an inductor is that it generates losses. Those losses are generated by two phenomenons :

- **Hysteresis effect:** in reality the B-H curve of a magnetic material has hysteresis. This effect happens for an alternating flux density (sinusoidal or trapezoidal waveform) and implies that the B-H curve followed for an increasing flux density is not the same as for a decreasing flux density (Figure 2.8a). With equation (2.9), it can be concluded that surface area between the curves corresponds to the energy lost per cycle. The power dissipated by hysteresis is linearly proportional to frequency. Its dependency to the flux density is difficult to evaluate but an estimation of the power dissipated by hysteresis can be expressed as follows [22] :

$$P_{hyst} = k_{hyst} \cdot f \cdot \hat{B}^m \cdot V \quad (2.25)$$

where k_{hyst} and m are constants specific for each magnetic material, $\hat{B}$ is the peak value of flux density and V is the core volume.

- **Eddy currents:** when the flux density changes inside an electric conductor, currents appear in order to oppose to this change (Figure 2.8b). Unfortunately, magnetic materials have not an infinite resistivity and eddy currents are generated inside the material. Losses are then generated by *Joule's law*. It is still possible to limit the effect with lamination of the magnetic material. The equation of eddy currents power loss is [20] :

$$P_{eddy} = \frac{\left(\frac{d\phi}{dt}\right)^2}{R_{core}} = k_{eddy} \cdot f^2 \cdot \hat{B}^2 \cdot V \quad (2.26)$$

This means that high resistance material generates less eddy losses.

For a zero-offset sinusoidal flux density, these two sources of losses can be gathered into a single equivalent volumetric power loss equation called the *Steinmetz equation* [23]:

$$P_{v,core} = k \cdot f^\alpha \cdot \hat{B}^\beta \quad (2.27)$$

k , α and β are constants specific to each magnetic material and are determined by experimentation.

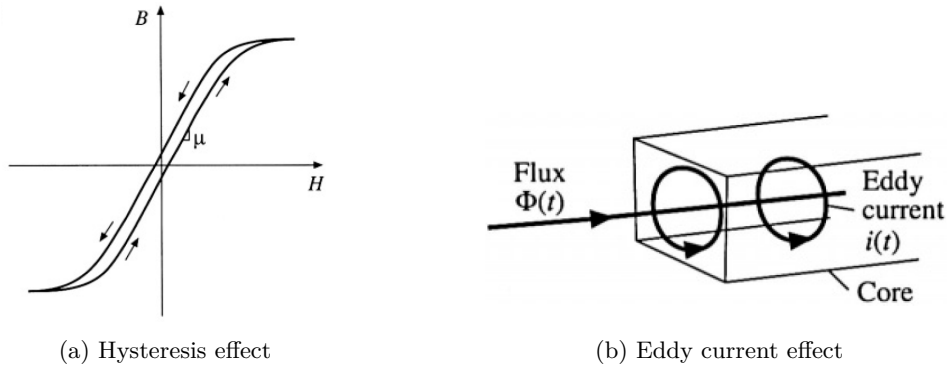


Figure 2.8 – Core loss effects [20]

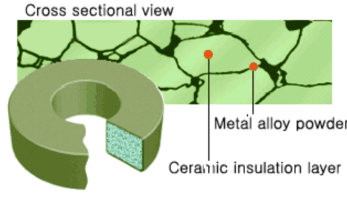
2.2.1 Core selection

In the industry, different kind of magnetic materials are used. Each material has its own properties which are more adapted for a task or an other. It is really important to choose the most suitable core material for the transformer in order to optimize the performances. The main requirements are high permeability, low losses, high Curie temperature and low cost. Of course, the perfect material for transformers does not exist. For each material, there are pros and cons. Here is a short summary of three of the most used magnetic materials : Metal Alloy Tape-Wound core, Powder Metal core and Ferrite core [22], [24], [25].

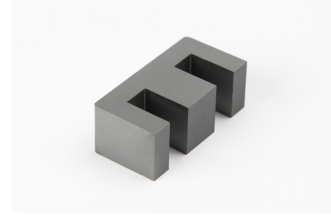
- Metal Alloy Tape-Wound core : this material consists of alloys such as Permalloy which has good magnetic properties : high permeability ($\mu_r > 60000$), high saturation flux density ($B_{sat} > 0.9T$) and low hysteresis. Unfortunately, alloys have a really low electric resistivity ($\rho = 50-150\mu\Omega \cdot cm$) that implies high losses by eddy currents. A way to decrease eddy currents is to laminate the alloy with an electric insulation. The material is then a tape composed of thin alloy layers (up to $12.5 \mu m$ thickness) separated by insulation layers. This tape is wound in order to create a composite material (Figure 2.9a). However, the tape resistivity is still too low. This material is then used for low frequency tasks (0 - 400 Hz).
- Powder Metal core: this material is a composite powdered-metal core (Figure 2.9b). The properties of the material are defined by the powder structure. The idea is to spread a uniform airgap between powder elements. The equivalent permeability is then decreased ($\mu_r = 14 - 500$) and a high amount of energy can be stored. So, this kind of material is not suitable for transformers where the magnetizing energy needs to be low. The powder metal core has a high electric resistivity because of the small airgaps in the powder.
- Ferrite core: this material is a ceramic made by a composition of iron oxide such as the hematite (Fe_2O_3) mixed with other materials such as nickel, manganese or zinc (Figure 2.9c). It has good magnetic and electric properties : high permeability ($\mu_r > 2000$), low hysteresis effect and high resistivity (200 - 2000 $\Omega \cdot cm$). This material is a trade-off between the two others and it generates a low amount of losses. However, ferrite cores have some weakness. First, the saturation flux density is low ($B_{sat} < 0.5T$). The maximal flux density is then limited and it needs to be compensated with an increase in the frequency and/or core section (2.4). Secondly, its Curie temperature is low ($T_c = 220^\circ C$). However, it will be seen later that this temperature is not critical. Finally, a ferrite is extremely fragile. The machining process needs to be chosen wisely.



(a) Metal Alloy Tape-Wound [25]



(b) Powder Metal



(c) Ferrite [26]

Figure 2.9 – Magnetic core materials

Table 2.2 contains a comparison of these three core materials. Finally, the ferrite core is chosen because of its low losses generation and it has no major weakness.

Material	Metal Alloy Tape-Wound	Powder Metal	Ferrite
Permeability (μ_r)	+	--	+
Saturation (B_{sat})	+	+	-
Hysteresis	+	--	+
Resistivity (ρ)	--	+	+
Curie temperature (T_c)	+	+	-

Table 2.2 – Magnetic cores comparison

2.2.2 Core loss for trapezoidal waveforms

In the case of a Dual Active Bridge converter, the magnetic flux density involved are either trapezoidal or triangular waveforms⁶. In literature, an adaptation of the *Steinmetz equation* for non-sinusoidal waveforms was made [23]. It is called the *improved Generalized Steinmetz Equation (iGSE)* and is expressed

⁶For more explanation about the flux density waveforms, see Appendix A.3

as follows :

$$P_{v,core} = \frac{1}{T} \int_0^T k_i \left| \frac{dB}{dt} \right|^\alpha (\Delta B)^{\beta-\alpha} dt \quad (2.28)$$

where ΔB is the peak-to-peak flux density and

$$k_i = \frac{k}{(2\pi)^{\alpha-1} \int_0^{2\pi} |\cos\theta|^\alpha 2^{\beta-\alpha} d\theta} \quad (2.29)$$

The case of a zero-offset trapezoidal waveform is developed in Appendix A.6. The resulting equation is :

$$P_{v,core} = k \cdot f^\alpha \cdot \hat{B}^\beta \cdot 0.915 \cdot D^{1-\alpha} \quad (2.30)$$

where k , α and β are the same coefficients as for the original *Steinmetz equation*. D is the trapezoid duty cycle ($D = \frac{t_{ramp}}{T}$). It can be seen that equation (2.30) is composed of the original *Steinmetz equation* multiplied by a term depending on the trapezoid duty-cycle. For a ferrite, the coefficient α is equal to 1.5. Therefore, the losses for a ferrite are expressed as :

$$P_{v,core}(\alpha = 1.5) = k \cdot f^{1.5} \cdot \hat{B}^\beta \cdot 0.915 \cdot D^{-0.5} \quad (2.31)$$

For a triangular waveform ($D = 1$), the core losses generated are lower than for a sinusoidal waveform with same frequency and same peak flux density ($P_{v,tri} = 0.915 P_{v,sin}$). This is the case of the core losses generated in the transformer magnetic core.

However, for the trapezoidal waveform involved in the inductor magnetic core ($D = 0.25$), the losses are higher than for a sinusoidal waveform with same characteristics ($P_{v,trap} = 1.83 P_{v,sin}$). Actually, the stiffer is the waveform slope, the higher is the correction factor due to the flux density waveform. This also explains why a low phase-shift is not chosen for the DAB in section 1.4. For a phase-shift that tends to 0 ($D \rightarrow 0$), equation (2.31) tends to infinity.

Finally, the expression of the core loss equivalent resistances (R_{ml} and $R_{p,ext}$) can be evaluated thanks to the primary voltage and the core losses of both the transformer and the external inductor :

$$\begin{aligned} R_{ml} &= \frac{V_1^2}{P_{core,tf}} \\ R_{p,ext} &= \frac{V_1^2}{P_{core,ext}} \end{aligned} \quad (2.32)$$

Key points

In this section, the impact of the core losses was explained. Here is a summary of the important points :

- The magnetic losses are generated by two different effects : hysteresis and eddy currents.
- There are several magnetic materials with different properties. Ferrite was selected as the most suitable magnetic core because it generates reasonable amount of eddy losses and hysteresis losses.
- The expression of core losses for a non-sinusoidal flux density was obtained by the *improved Generalized Steinmetz Equation*. It showed that the core losses are increased by 83% for a trapezoidal waveform such as in the external inductor. For a triangular waveform such as in the transformer, the losses are reduced by 8.5% with respect to a sinusoidal waveform.

2.3 Winding loss modelling (R_1 , R_2 and $R_{s,ext}$)

The winding resistances of a transformer are an important point to take care about during the design. Indeed, for a fixed current, the copper losses are directly proportional to this resistance. Moreover, it is harder to cool windings in space because of lack of convection. Therefore, it is important to reduce winding losses as much as possible.

The dc-resistances of the windings are easily calculated :

$$\begin{aligned} R_{1,dc} &= n^2 \cdot \frac{2\pi \cdot (R + L_6 + \frac{1}{2}L_4)}{\sigma_{cu} \cdot L_2 L_4 \cdot k_{f,wind}} \\ R_{2,dc} &= n^2 \cdot \frac{2\pi \cdot (R + L_6 + L_5 + \frac{3}{2}L_4)}{\sigma_{cu} \cdot L_2 L_4 \cdot k_{f,wind}} \end{aligned} \quad (2.33)$$

where $k_{f,wind}$ is the winding fill factor and σ_{cu} is the copper electric conductivity depending on the winding temperature :

$$\sigma_{cu} = \sigma_{cu}(T) = 17 \cdot 10^{-8} \cdot (1 + 0.004 \cdot (T[^\circ C] - 20)) [\Omega^{-1} m^{-1}] \quad (2.34)$$

The operating frequency of a transformer used for DC/DC conversion is usually in the range from 10 kHz to 1 MHz. The pros and cons of high frequency in transformers will be discussed for the optimization in chapter 3. However, high frequencies have a bad influence on the wire resistance. Indeed, for a high operating frequency, the leakage flux interacts with the current. The consequence is that the equivalent wire ac-resistance is increased [20]. The frequency influence on the wire resistance is analysed in this section.

2.3.1 Skin effect

The skin effect appears first at high frequencies. When a varying current flows through a conductor, it generates a varying flux density. This varying flux, in turn, generates eddy-currents that change the current distribution inside the conductor (Figure 2.10). The current density is reduced in the center of the conductor and increased on its surface. Finally, it can be said that the net current is "pushed" to the conductor surface. The *skin-depth* is the equivalent thickness wherein all current is flowing (Figure 2.10). It is expressed as [20] :

$$\delta = \sqrt{\frac{\rho_{cu}}{\pi \mu_0 f}} \quad (2.35)$$

The skin depth value depends on the copper permeability μ_0 and resistivity ρ_{cu} . Because of the skin effect, the effective cross section of the wire is decreased. Therefore, the resistance is increased and the winding losses as well. For a 50 kHz AC current at ambient temperature, the skin depth is $\delta = 0.34 \text{ mm}$. A way to avoid the loss of useful copper area is to use a wire composed of several strands thinner than twice the *skin depth* in diameter. But, the increase of the strand layers has also a negative impact on the ac-resistance because of *Proximity effect*.

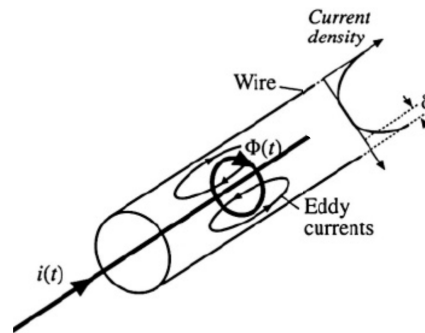


Figure 2.10 – Skin effect representation with eddy-currents and final current density distribution [20]

2.3.2 Proximity effect

The *Proximity effect* is a phenomenon happening in multi-layers and multi-strands windings and which is often neglected. However, that has an important impact on the equivalent ac-resistance R_{ac} [20]. This is why this effect is deeply analysed in this section.

A varying current $i(t)$ flowing through a strand generates a varying magnetic flux. Inside of this strand, the flux produces eddy-currents and the net current is pushed to the wire surface. This is the skin effect (Figure 2.11). In a same way, the varying flux density produces eddy-currents in the adjacent strands (located next to the first strand). As for the skin effect, the eddy-currents interfere with the initial current flowing in the adjacent strands. The current distribution is influenced and the net current in the adjacent strand is pushed on the side furthest away from the first strand. The effect is represented in figure 2.11. In the representation, the proximity effect generated by the first strand (S1) pushes the net current of S2 in the right side of the section. The effective section of S2, colored in yellow, is reduced. Similarly, the current of S2 (which is already concentrated in the right side) pushes the current of S3 in the right side of its section. This "force" applied by the current of S2 on the current of S3 is even stronger than the "force" applied by the current of S1 on the current of S2. Thus, the effective strand section for S3 is lower than for S2. Actually, there is an amplification of the proximity effect over the strands and the effective section area decreases with the strands.

Figure 2.11 represents the lost of the strands effective cross area. For a DC current, the current flows in the entire wire section. When Skin effect is added, the current is pushed on the surface in the same way for each strand and the strand effective section is decreased. Finally, the proximity effect decreases the effective strand section with an amplification over the strands. For readers who would like to know more about the proximity effect amplification, an introduction of the concept is presented in Appendix A.7.

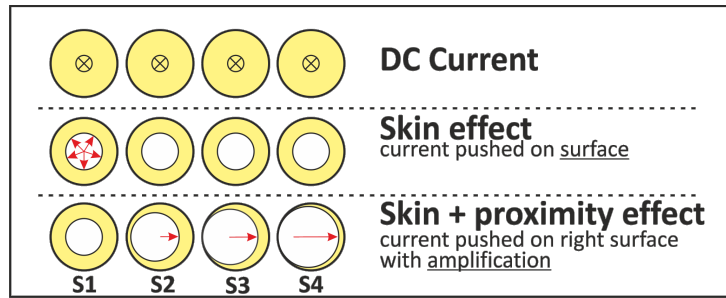


Figure 2.11 – Representation of the skin and proximity effects on four strands with same current. The yellow color represents the effective area within which the net current flows. The red arrows represents the force on the current

The proximity factor (F_R) is defined as the ac-to-dc resistance ratio. This factor represents the skin effect and proximity effect impact on the winding resistance and the power losses. *Maxwell's equations* are used to evaluate an analytical equation of this factor also known as *Dowell equation* [27] :

$$F_R = \frac{R_{ac}}{R_{dc}} = \Delta \left[\frac{\sinh(2\Delta) + \sin(2\Delta)}{\cosh(2\Delta) - \cos(2\Delta)} + \frac{2}{3}(M^2 - 1) \frac{\sinh(\Delta) - \sin(\Delta)}{\cosh(\Delta) + \cos(\Delta)} \right] \quad (2.36)$$

with the skin factor $\Delta = \sqrt{\frac{\pi}{4}} \frac{d}{\delta}$ and $M = p * s$ the number of strand's layer s^7 contained in p winding layers (Figure 2.12b).

Figure 2.12a represents the evolution of the *proximity factor* F_R with respect to the *skin factor* (Δ) and the number of layers (M). An approximation of this equation for the case $\Delta < 1$ is given by [27] :

$$F_R = 1 + \frac{5M^2 - 1}{45} \Delta^4 \quad (2.37)$$

⁷For a round wire composed of N_s strands, a good approximation is $s = \sqrt{N_s}$

Two points are important to understand in this equation :

- The *proximity factor* (F_R) decreases with a decreasing *skin factor* (Δ). Then, increasing the frequency f increases the proximity effect. Similarly, decreasing the strand diameter d reduces the proximity effect. However, for a same winding thickness L_4 , that means increasing the number of layers M . This can be seen in figure 2.12a. The black curve represents the proximity factor for a constant winding thickness.
- The *proximity factor* (F_R) increases with the number of layers squared (M^2). Then it is interesting to use a few strands with large diameter to reduce the number of layers.

These two points are actually opposed. A solution is to keep the same total winding cross section (S_{wind}) and decrease the winding thickness L_4 which implies to increase the winding height L_2 (Figure 2.12b). An important geometry factor of the winding is then introduced : $k_g = \frac{L_4}{L_2^2}$. Regarding proximity effect, the optimal winding shape is a long tube with thin thickness ($k_g \rightarrow 0$). This shape is illustrated in figure 2.14a. That geometry reduces the number of layers for a same winding cross section. However, this is not optimal for the core losses as the core volume around the winding would increase. Moreover, the core reluctance would increase which means a lower magnetizing inductance.

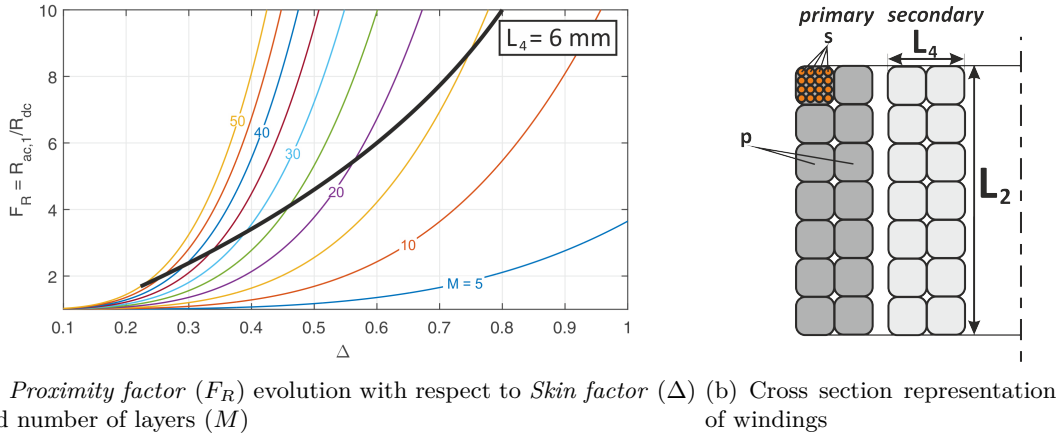


Figure 2.12

Harmonics effect

The *Proximity factor* (F_R) only considers the fundamental frequency of the current. This is enough for sinusoidal AC currents. However, in the Dual Active Bridge currents are not sinusoidal but trapezoidal. It is important to evaluate the harmonics impact on the *proximity effect*. This is expressed with the *proximity harmonics factor* (F_H) which is the ratio of the sum of harmonics's *proximity factor* over the one of the fundamental:

$$F_H = \frac{\sum_{j=0}^{\infty} P_{cu,ac}(f_j)}{P_{cu,ac}(f_0)} = \frac{\sum_{j=0}^{\infty} (F_R(f_j)|I(f_j)|^2)}{F_R(f_0)|I(f_0)|^2} \quad (2.38)$$

where f_0 is the fundamental, f_j is the j^{th} -harmonic. Appendix A.4 summarizes the frequency spectrum of those waveforms. A final ac-to-dc resistance ratio considering the entire *proximity effect* is expressed by :

$$F_P = \frac{R_{ac}}{R_{dc}} = F_R \cdot F_H \quad (2.39)$$

Figures 2.13a and 2.13b represent respectively the *harmonics proximity factor* (F_H) curves for a square waveform ($D = 0$) and a trapezoidal waveform ($D = 0.25$). It can be seen that the stiffer is the current ($D \rightarrow 0$), the larger is the harmonic content and the larger is the harmonics proximity effect. This also explains why a low phase-shift is not chosen for the DAB in section 1.4.

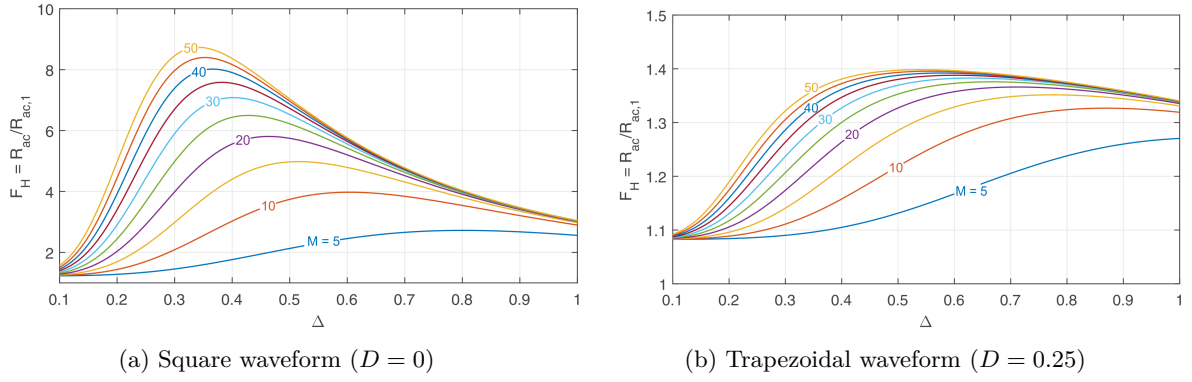


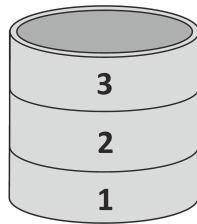
Figure 2.13 – Proximity harmonic factor (F_H) evolution with respect to Skin factor (Δ) and number of layers (M)

Adjacent winding

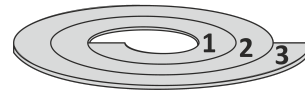
In chapter 1, the optimal transformer geometry for the rotating transformer was introduced. In this section, it was said that the coaxial winding is more suitable than the adjacent winding. Here is the explanation.

Regarding proximity effect, the optimal shape for a coaxial winding is a long and thin tube (Figure 2.14a). Indeed, the thin thickness allows to reduce the number of layers, and then to reduce the proximity effect. In contrast, the optimal shape for an adjacent winding is a thin and large plate (Figure 2.14b). Indeed, for an adjacent winding, the MMF increases in the axial direction and not in the radial direction as for the coaxial windings. Therefore, this is the axial length of the winding that needs to be reduced in order to reduce the proximity effect. However, for an adjacent winding, the wire is wound in a spiral shape where each turn has a longer length than the previous one. Each turn is even longer when the wire width is long. Therefore, regarding dc-resistance, it is better to have a thicker spiral. Thus, there is a trade-off on the adjacent winding : a thin plate implies high dc-resistance and a thick plate implies high proximity factor. In both cases, the ac-resistance is high and the adjacent winding is not adapted.

The coaxial winding is more adapted because the loop length does not depend on the winding height.



(a) Efficient winding for coaxial placement



(b) Efficient winding for adjacent placement

Figure 2.14 – Representation of the windings geometry for a winding coaxial placement and adjacent placement in order to decrease the proximity effect

2.3.3 Litz wire

A way to reduce the skin effect is to use wire composed of thin strands in parallel with a diameter lower than twice the *skin depth*. However, this kind of technology needs to be used carefully. If the strands are simply connected in parallel their length would differ regarding the radial localisation inside the winding. A longer strand has a higher dc-resistance and a higher inductance. Therefore, the strands would not have the same impedance and the current would not be uniformly distributed over all the strands. A

solution used to fix that is to transpose the strands inside a wire such that each strand passes equally through each position inside the wire cross section along the wire. The average resistance and flux density exposition are then equal for each strand. This kind of wire is called a *Litz wire* (Figure 2.15b) and a representation of the transposition method is shown in figure 2.15b.

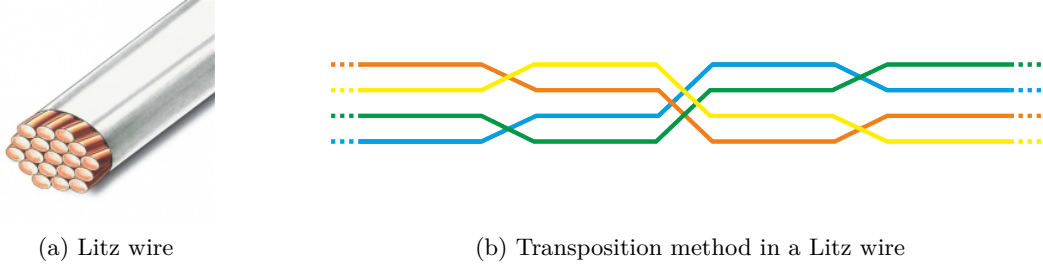


Figure 2.15 – Litz wire technology

The *Litz wires* are sorted by wire diameter, number of strands, strand diameter, isolation material, wire cross section shape, etc. The strand diameter can be really thin. For example, the wire used for the prototype has a 3 mm diameter and is composed of 840 strands of 71 μm diameter⁸ (Datasheet available in Appendix B.4)

A problem with *Litz wire* composed of round strands is the *filling factor* ($k_{f,wire}$). This factor corresponds to the copper area (A_{cu}) over the total wire cross section area (A_{wire}) :

$$k_f = \frac{A_{cu}}{A_{wire}} \quad (2.40)$$

Indeed, as all strands need to be electrically isolated, the amount of isolation in the wire cross area is higher when the number of strands increases. The fill factor of the given Litz wire is $k_{f,wire} = 0.44$.

Key points

For DC/DC converters, a relatively high frequency is requested. The high frequency has a significant impact on the transformer winding resistances and losses :

- *Skin effect* and *proximity effect* produce an increase of the windings ac resistance.
- Those effects can be limited by using stranded wires with strand diameter as thin as possible (typically $\delta \gg d$).
- *Proximity effect* can be limited by reducing the total number of layers (M). It is achieved by reducing the thickness of windings (L_4) and the geometry factor $k_g = \frac{L_4}{L_2}$.
- The harmonic content of the current has an important impact on the *proximity effect*. A stiffer current produces more losses.
- The use of *Litz wire* offers a transposition of every strand of the wire. This implies that every strand has the same inductance, the same resistance and, then, the same flowing current. However, it has a poor fill factor (k_f).

⁸AWG (*American Wire Gauge*) is a measure unit for wires : AWG41 $\rightarrow$ 71 μm

2.4 Thermal modelling

The thermal management is an important aspect when designing a transformer. Indeed, losses produced by the windings and the magnetic core are transformed into heat. This heat has to be cooled as efficiently as possible in order to avoid overheating and high temperature inside the transformer. The task is even more complicated because of the lack of convection in spatial applications. In this section, an advanced but simplified thermal model of the transformer is introduced. This is called the *lumped parameter thermal model*. This model is based on the analogy between heat transfer and electric circuits.

In a satellite, the only way to cool the different systems is to conduct heat to a *heat rejection system*. This system consists of radiators composed of white panels radiating heat to the darkness of space. The heat flow process to the heat rejection system is not mentioned in the specifications but it can be assumed that the maximal baseplate temperature is $T_{bp,max} = 65^\circ C$.

The critical point in the thermal management is the windings. Indeed, they are really hard to cool as they are located in the center of the transformer. Moreover, the windings temperature cannot exceed the insulation melt point. Otherwise, a short circuit would be created between the loops of the windings and the system would be damaged. However, the insulation melt point is quite high ($T_{melt} = 180^\circ C$) (Appendix B.4). But, a margin of $20^\circ C$ is applied on this temperature and the maximal temperature accepted in the windings is $T_{max} = 160^\circ C$.

2.4.1 Thermal resistance concept

For the reluctance model, an analogy between magnetism and electricity was made. In a same way, the thermal model is based on an analogy between heat transfer and electricity. In this analogy, temperatures (T) are equivalent to voltages (V), and heat ($\dot{Q}$) is equivalent to current (I). With these equivalences, the idea of a thermal resistance can be developed : the thermal resistance is the ability of a material to generate a temperature drop while heat crosses it :

$$\dot{Q}_{cond} = R_{cond} \cdot \Delta T \quad (2.41)$$

From the thermal resistance definition and the *Fourier's law*, it is possible to find the thermal resistance equations depending on the geometric dimensions and the material properties. The thermal resistance for heat conduction in the axial direction is :

$$R_{cond} = \frac{L}{k \cdot A} \quad (2.42)$$

with k the thermal conductivity, A the surface area crossed by the heat flux and L the length in the flux direction.

For heat conduction in the radial direction, the thermal resistance is :

$$R_{cond,rad} = \frac{\ln(r_o/r_i)}{2\pi \cdot h \cdot k} \quad (2.43)$$

with r_i the inner radius, r_o is the outer radius and h is the height. Note that the thermal resistances in equations (2.42) and (2.43) are similar to the reluctances equations of the reluctance model. Indeed, the same geometry dimensions are involved. The only difference is the material property involved. The reluctance is characterized by the material magnetic permeability (μ) and the thermal resistance is characterized by the thermal conductivity (k).

2.4.2 The lumped parameter thermal model

The *lumped parameter thermal model* aims to represent the transformer as a network composed of different nodes[28]. These nodes represent the different geometrical parts of the systems. Each node is characterized by a mean temperature and is assumed to be located in the center of its associated geometric part. So, a first strong assumption is that the temperature is assumed to be uniform over a same geometry part. In reality this is not the case but a good choice of the nodes network allows this assumption (final

temperatures will not differ too much). For each point, a heat generator is associated. This represents the losses generated in the geometric parts. Finally, the different nodes are connected through thermal resistances when a physical connection between the associated transformer's parts exist. When there is no connection between nodes, that means that there is no heat exchange. For example, as there is no physical contact between rotor and stator, there is no heat exchange between them as convection is not happening.

Figure 2.16 presents the thermal resistance network over the transformer topology. It can be seen that there is no connection at all between rotor and stator. So, it is possible to compute the temperatures of rotor and stator separately. As said before, each point stands for a geometrical part of the transformer. The boundaries of those parts are chosen such as each heat flux in radial or axial direction can pass through a thermal resistance. Both stator and rotor have five different parts : tube (1,6), winding (5,10), inner plate ring (2,7), middle plate ring (3,8) and outer plate ring (4,9). The green boundaries represent the baseplate temperature. In this representation, the cooling is supposed to happen by the shaft for the rotor and the outer cylinder for the stator. A thermal glue is also considered at the winding-core interfaces. Those are represented in black.

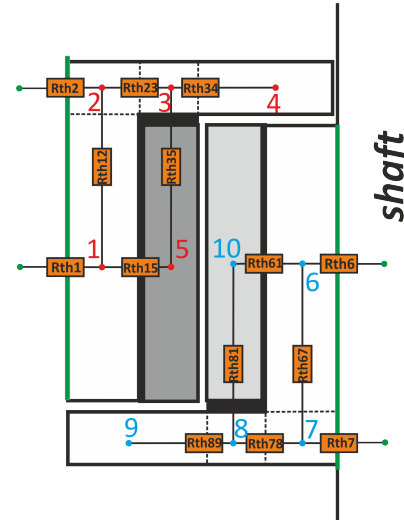


Figure 2.16 – Thermal circuit representation over the transformer topology

A complete circuit of both rotor and stator can be seen in figure 2.17. On these circuits, both the thermal resistances and the power generators are represented. The electric ground represents the baseplate temperature. So, each temperature is relative with respect to this baseplate temperature. The thermal study is at steady state only. The use of equivalent thermal capacitance is not considered in this document.

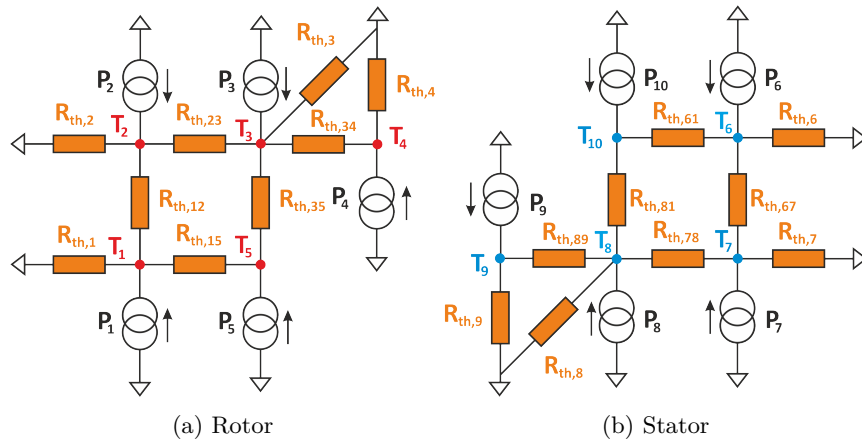


Figure 2.17 – Complete thermal circuits

2.4.3 Winding equivalent thermal conductivity

The thermal resistance of windings is an important point for the thermal management. A high amount of heat is generated in those because of the winding resistances. The heat is conducted either radially or axially. Indeed, it is supposed that no heat flows in the tangential direction because the winding length is too long. The problem is that the windings have a poor thermal conductivity in both radial and axial directions because of insulation. Indeed, the heat produced in each wire strand needs to flow through strand and wire insulation. Unfortunately, insulation which is used for their low electric conductivity have also a low thermal conductivity ($k_{ins} = 0.15 \text{ W/mK}$). Moreover, the wires in a winding for spatial application are impregnated with a resin in order to avoid the presence of vacuum space with zero conductivity inside the winding. The resins also have a low conductivity ($k_{res} = 0.15 \text{ W/mK}$). Finally, regarding thermal management, a Litz wire can be seen as a composite material composed of insulation (including resins) and copper. The aims of this subsection is to evaluate an equivalent thermal conductivity of this composite material.

Figure 2.18a represents the cross section of a winding with a zoom on a Litz wire. Each Litz strand is covered by an insulation (green) and each wire is also covered by an insulation (white). Between strands and wires the resin can be seen in yellow. The fill factors of the winding and the wire can be expressed as follows :

$$k_{f,wind} = \frac{\text{wire area}}{\text{total winding area}} = N_w \cdot \frac{\pi \left(\frac{d_w}{2}\right)^2}{L_2 L_4} \quad (2.44)$$

$$k_{f,wire} = \frac{\text{strand area}}{\text{total wire area}} = N_s \cdot \frac{d_s^2}{d_w^2}$$

where N_w is the number of wire in the window, d_w is the wire diameter, N_s is the number of strands in a wire and d_s is the strand diameter.

In practice, wires and strands have a round cross section (Figure 2.18a). However, it is too hard to evaluate an equivalent thermal conductivity with round elements. In order to address the issue, the original winding (Figure 2.18a) is modelled by an equivalent simplified winding (Figure 2.18b) composed of the same number of wires, the same number of strands, and each wire and strand has a square shape with same section area as the originals:

$$h_w = \sqrt{\frac{\pi}{4}} \cdot d_w \quad (2.45)$$

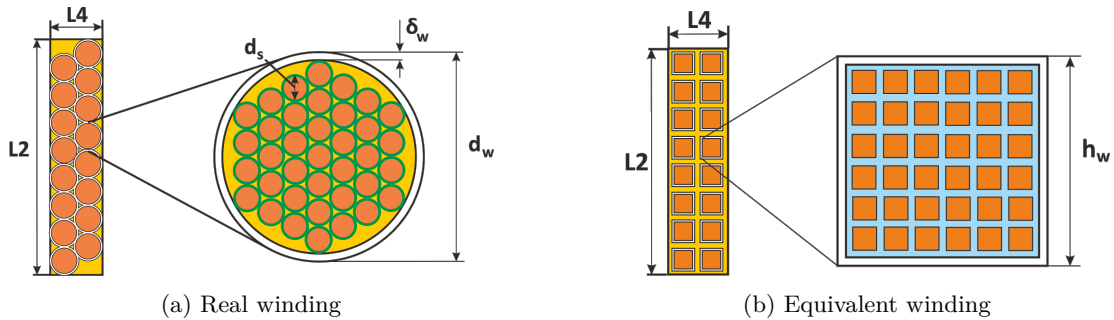


Figure 2.18 – Cross section representation of the real and equivalent winding

In that way, the heat flux is assumed to flow either in radial or axial directions. With square shapes, the equivalent thermal conductivity of the composite winding is easier to compute. The derivation was developed for this project and is available in Appendix A.8. The resulting formula for the winding equivalent thermal conductivity is :

$$k_{wind} = k_{ins} \cdot \left(1 + \frac{(k_e^{-1} - 1) \cdot \sqrt{k_{f,wind}}}{1 + (k_e^{-1} + 1) \cdot \sqrt{1 - k_{f,wind}}} \right)^{-1} \quad (2.46)$$

where k_e is the equivalent thermal conductivity factor of a Litz wire :

$$k_e = \frac{k_{wire}}{k_{ins}} = \frac{\left(\frac{(1-k_{f,wire})}{\sqrt{1-k_{f,wire}} + \sqrt{k_{f,wire}}} \right) + \sqrt{k_{f,wire}}}{\sqrt{1-k_{f,wire}}} \quad (2.47)$$

For a winding made of Litz wire with fill factor $k_{f,wire} = 0.44$ and winding fill factor $k_{f,wind} = 0.75$, the winding equivalent conductivity is $k_{wind} = 1.17 \cdot k_{ins}$. So, in a typical winding, the thermal resistance is about 15% better than the case of a winding filled of insulation only. That makes sense because the low thermal conductivity dominates the equivalent thermal conductivity that is slightly increased thanks to the copper.

This equivalent thermal conductivity can be applied for winding thermal resistance in radial direction with (2.43) :

$$R_{th,wind,r} = \frac{\ln\left(\frac{r_i + L_4}{r_i}\right)}{k_{wind} \cdot 2\pi \cdot L_2} \quad (2.48)$$

where r_i is the inner radius of winding. It can be seen that this resistance is reduced when the winding height L_2 is increased and when its thickness L_4 is decreased ($k_g \rightarrow 0$). This is another argument in favour of a thin long tube for the winding shape (Figure 2.14a).

2.4.4 Core-Winding interface resistance

When assembling the winding on the magnetic core, it is important to take care of the thermal contact resistance between those. Indeed, surfaces that look smooth have still a microscopic roughness. The roughness of two objects in contact generates enclosures filled of vacuum (Figure 2.19a). The contact area between the winding and the core is reduced because of the enclosures. Therefore, the thermal contact resistance between the materials increases and a large temperature gradient appears around the core-winding interface. Hence this is important to fill the enclosures with a *thermal interface material* (TIM) in order to reduce the contact thermal resistance (Figure 2.19b). There exist several kinds of TIM: thermal grease, pads, tape, phase changing materials (PCM), gel, glue, etc. According to [31], the best TIM for a transformer application is the glue because it needs no pressure, it has good mechanical and thermal properties and there is no pump-out phenomenon involved. The disadvantage of glue is that it is not possible to disassembly. For our application, the glue used has a thermal conductivity of $k_{glue} = 0.1 \text{ W/mK}$ and a thickness of $\delta_{glue} = 0.5 \text{ mm}$. The thermal resistance of the thin glue contact is then:

$$R_{th,glue} = \frac{5 \cdot 10^{-3}}{A_{glue}} \quad (2.49)$$

where A_{glue} is the glue surface area.



Figure 2.19 – Microscopic view of interface contact between core and winding

2.4.5 Temperature computation

Before computing the temperatures, all the components involved in the thermal circuits (Figure 2.17) need to be calculated :

- For the thermal resistances, the inputs needed are the thermal conductivity of the materials and the dimensions of each transformer part. The winding thermal conductivity have just been derived and the ferrite thermal conductivity is $k_{fer} = 3.5 \text{ W/mK}$.

- For the heat generators, the inputs are the losses generated in the core and in the windings. For core losses, the core dimensions are also needed.

All thermal resistance expressions are given in Appendix A.9.

With power losses for each node and the thermal resistances connecting those in a network, it is possible to compute the temperature of each node. To do so, the conservation of heat flux for each node is used. This is the same operation as for *Kirchhoff's current law*. For each node, incoming heat flux has to be equal to outgoing heat flux. For example, the heat flux conservation on *node 1* (Figure 2.17a) gives :

$$P_1 = \frac{(T_1 - T_0)}{R_{th,1}} + \frac{(T_1 - T_2)}{R_{th,12}} + \frac{(T_1 - T_5)}{R_{th,15}} \quad (2.50)$$

Heat flux conservation is then applied for each node of the network except for the baseplate because it is used as reference temperature. So, for n nodes, n equations are gathered into a system :

$$\delta \mathbf{T} = \mathbf{G}^{-1} \mathbf{P} \quad (2.51)$$

$$\begin{bmatrix} T_1 - T_0 \\ T_2 - T_0 \\ \vdots \\ T_n - T_0 \end{bmatrix} = \begin{bmatrix} \sum_{i=1}^n \frac{1}{R_{1,i}} & -\frac{1}{R_{1,2}} & \cdots & -\frac{1}{R_{1,n}} \\ -\frac{1}{R_{2,1}} & \sum_{i=1}^n \frac{1}{R_{2,i}} & \cdots & -\frac{1}{R_{2,n}} \\ \vdots & \vdots & \ddots & \vdots \\ -\frac{1}{R_{n,1}} & -\frac{1}{R_{n,2}} & \cdots & \sum_{i=1}^n \frac{1}{R_{n,i}} \end{bmatrix} \begin{bmatrix} P_1 \\ P_2 \\ \vdots \\ P_n \end{bmatrix} \quad (2.52)$$

where $R_{x,y}$ is the thermal resistance between nodes x and y . T_0 is the reference temperature which is equal to the baseplate temperature T_{bp} . Temperatures are computed with a Matlab function.

Key points

Thermal management is an important issue when designing a transformer for spatial application :

- For thermal modelling, the *lumped parameter thermal model* is used. It is an analogy between heat transfer and electric circuits.
- The critical point is located in the windings as there is no direct connection to the baseplate. The magnetic core is also used to drain the heat from the windings.
- Windings have a low equivalent thermal conductivity $k_{wind} = 0.18 \text{ W/mK}$ because of insulation and resin. This equivalent thermal conductivity was calculated thanks to a simplified winding model.
- The low thermal conductivity of the windings can be compensated by reducing the geometric factor ($k_g \rightarrow 0$). This reduces the winding thickness (L_4) and increases the exchange surface area ($A_e \propto L_2$).
- A *thermal interface material* (TIM) is needed to reduce the contact thermal resistance between ferrite and windings.
- The maximal temperature allowed in the windings is 160°C for a maximal baseplate temperature of 65°C .

Now that the different models of the transformer have been introduced, it is time to design the transformer. In this chapter, a numerical optimization is considered in order to find the optimal transformer dimensions. First, the different variables involved in the optimization are presented. The impact of those variables on the optimization objectives is analysed. Afterwards, the optimization method is introduced. Then, the resulting design is given and analysed. Finally, the external inductor design is described.

3.1 Optimization variables

In section 1.3, the transformer geometry was introduced. The optimal geometry is characterized by 8 dimensions (Figure 3.1b): L_1 , L_2 , L_3 , L_4 , L_5 , L_6 , R and e .

The 8 dimensions can be divided in three categories (Figure 3.1a) : dimensions associated to the winding cross section (L_2 and L_4), dimensions associated to the core (L_1 , L_3 and L_6) and the other dimensions (L_5 , R and e). The last category corresponds to dimensions which are not involved in the optimization. Indeed, in the previous work [1], it was concluded that each of those dimensions need to be minimized in order to maximize the transformer efficiency. However, because of mechanical constraints, those dimensions cannot be below a certain value :

- For the airgap length (e), a realistic minimal value needs to be considered. Indeed, machining of magnetic core has not a high precision. The value taken is $e = 0.1$ mm.
- For the shaft radius (R), it is necessary to have a mechanical shaft that can withstand mechanical efforts due to the solar array weight. The shaft radius considered is the same as the one used for the *RUAG Septa 24*[®] : $R = 15$ mm.
- The winding gap length (L_5) is the distance between the windings. For assembly reasons, a length value $L_5 = 3$ mm is imposed in order to avoid a mechanical contact between primary and secondary windings.

The two winding dimensions (L_2 and L_4) can be determined by two equations (3.1). A first one is a relation with the winding cross section S_{wind} and the other one is the winding geometry factor k_g which was introduced in section 2.3 :

$$\begin{aligned} L_2 &= \frac{S_{wind}}{L_4} \\ k_g &= \frac{L_4}{L_2} \end{aligned} \tag{3.1}$$

The three core dimensions (L_1 , L_3 and L_6) can be determined by three equations (3.2) considering a uniform core section S_{core} along the magnetic path :

$$\begin{aligned} L_6 &= \sqrt{\frac{S_{core}}{\pi} + R^2} - R \\ L_1 &= \frac{S_{core}}{2\pi(L_6 + R)} \\ L_3 &= R_o - \sqrt{R_o^2 - \frac{S_{core}}{\pi}} \end{aligned} \quad (3.2)$$

The five dimensions that are not determined yet can be summarized by three geometry factors (Figure 3.1a) : S_{core} , S_{wind} and k_g . The two sections (S_{core} and S_{wind}) can be themselves determined by physical variables :

- The winding section can be obtained by the peak current equation :

$$\hat{I}_2 = \frac{k_f \cdot S_{wind} \cdot \hat{J}}{n} \quad (3.3)$$

where $\hat{I}_2$ is the peak current of the secondary winding, $\hat{J}$ is the peak current density, n is the winding number of turns and k_f is the winding fill factor. The peak current and the fill factor are both known. The winding cross section is then determined by two variables (Figure 3.1) : n and J .

- The core section can be obtained by the peak voltage equation¹:

$$\hat{V} = n \cdot \frac{d\phi}{dt} = 4 * f \cdot n \cdot \hat{B} \cdot S_{core} \quad (3.4)$$

where $\hat{V}$ is the peak voltage of the transformer, f is the switching frequency and $\hat{B}$ is the peak flux density. The peak voltage is known. The cross section is then determined by three variables (Figure 3.1) : n , f and $\hat{B}$.

Finally, the entire geometry can be determined by 5 physical variables : k_g , J , n , f and $\hat{B}$. Those variables are the input data for the optimization.

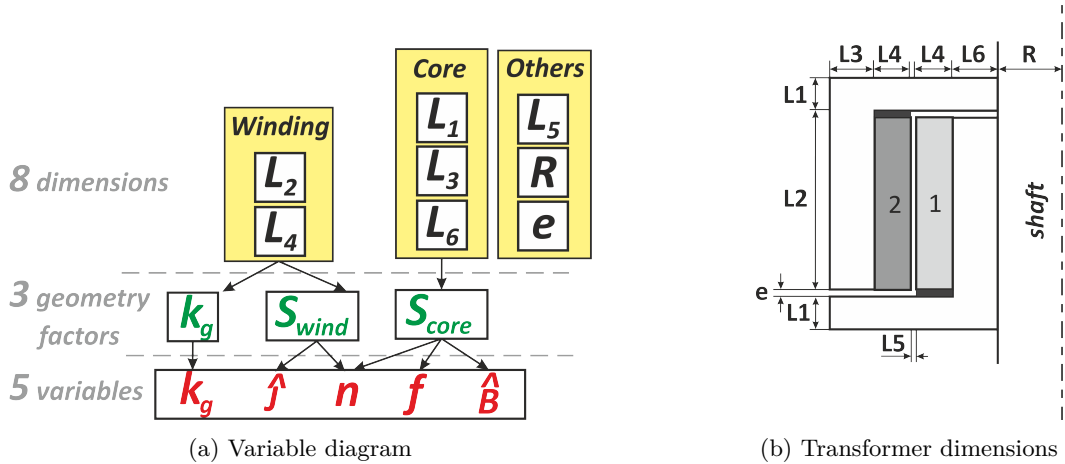


Figure 3.1 – Variables : variable diagram and dimensions representation on transformer geometry

¹The flux density waveform of the transformer is triangular (Appendix A.3)

3.2 Objective functions

A mathematical optimization is a selection of the best combination of input variables that maximizes or minimizes the objective functions. In this optimization, there are two objective functions : reduce the amount of losses generated ($P_{tot} = P_{cu} + P_{core}$) and reduce the transformer weight (M). The first objective implies to reach the highest efficiency while the second aims to maximize the power density (p) of the system as it is critical for spatial applications.

For an optimization with several input variables, it is sometimes impossible to find a single decoupled equation for each objective. Indeed, there is a high degree of interaction between all the variables. Equations (3.5), (3.6) and (3.7) show the dependency of each objective function with respect to the input variables and sections. The idea of these equations is not to be understood but it shows the high degree of interaction between all the variables :

$$P_{core} = f(\hat{B}, f, S_{core}) \quad (3.5)$$

$$P_{cu} = f(n, J, f, k_g) \quad (3.6)$$

$$M = f(S_{wind}, S_{core}, k_g^{-1}) \quad (3.7)$$

Each variable has an impact on both objective functions and the variation of one variable will have a positive impact for one objective function but a negative impact for the other one.

Example : let us consider the impact of the peak flux density ($\hat{B}$). To do so, all the other variables (n , f , $\hat{J}$ and k_g) are fixed. By (3.5), the core losses are proportional to the flux density. Therefore, it is interesting to decrease the flux density as much as possible. However, decreasing the flux density implies to increase the core section S_{core} in (3.4) and the transformer weight (M) is proportional to the core section (3.7). On the one hand, the flux density ($\hat{B}$) needs to be low for good core losses. On the other hand, it has to be high to reduce the transformer weight.

This exercise shows that a change of a variable does not favour both objective functions at a same time. The exercise can be done for the other input variables. For each of those, it comes that it is not possible to decrease the losses and the transformer's size at the same time by modifying only one variable. However, it is possible by changing all variables simultaneously.

3.3 Optimization method

A numerical optimization was operated with *Matlab*®. The chosen optimization method diagram is depicted in figure 3.2. The method consists in taking combinations of the five input variables and evaluating the transformer characteristics for each combination. For each variable, 10 values are considered in a range pre-determined. The number of combinations is then 10^5 .

The ranges used for the variables are :

$$\begin{array}{rclcl} 0.02 & \leq & k_g & \leq & 0.2 & [\text{m/m}] \\ 10^6 & \leq & \hat{J} & \leq & 5 \cdot 10^6 & [\text{A/m}^2] \\ 4 & \leq & n & \leq & 13 & [\text{turns}] \\ 10^4 & \leq & f & \leq & 10^5 & [\text{Hz}] \\ 0.05 & \leq & \hat{B} & \leq & 0.25 & [\text{T}] \end{array}$$

From each combination of inputs, the geometry factors (S_{core} , S_{wind} and k_g) can be determined. And from those geometry factors, the transformer dimensions ($L_1 \rightarrow L_6$, R and e) are known. Thanks to those dimensions and the models developed in chapter 2, the entire transformer characteristics can be determined. The steps followed by the Matlab program are as follows (Figure 3.2):

- **The core losses** (P_{core}) can be determined by equation (2.30) which is the *improved Steinmetz Generalized Equation* for triangular flux density waveform.

Inputs : core dimensions, the ferrite Steinmetz coefficients (k , α and β), the peak flux density ($\hat{B}$) and the frequency (f)

- **The magnetic equivalent components** (L_μ , $L_{f,1}$ and $L_{f,2}$) are determined with equations (2.17) and (2.22) resulting from the reluctance model.

Inputs : Transformer dimensions, number of turns (n) and ferrite relative permeability (μ_r).

- **The winding resistances** (R_1 and R_2) are determined with equation (2.33) which is the DC resistances formula and equations (A.36) to (2.39) which are the link to the AC resistances.

Inputs : winding dimensions, Litz wire characteristics, number of turns (n), frequency (f) and copper resistivity ρ_{cu} .

- **The core equivalent resistance** (R_{ml}) is obtained with equation (2.32).

Inputs : input voltage V_1 and core losses (P_{core}).

- **The currents** involved in the transformer (I_μ and I_1) can be determined with the *Kirchhoff's laws* and all the components of the transformer model.

Inputs : secondary current (I_2), transformer input voltage (V_{in}) transformer model components (L_μ , $L_{f,1}$, $L_{f,2}$, R_1 , R_2 and R_{ml}).

- **The winding losses** (P_{cu}) can then be determined

Inputs : the AC series resistances (R_1 and R_2) and the input and output currents (I_1 and I_2).

- **The temperatures** in the transformer are determined with the thermal model developed in the previous chapter.

Inputs : transformer dimensions, power losses (P_{core} and P_{cu}) and thermal conductivity (k_{th}).

The transformer efficiency (η) can then be calculated with the different losses :

$$\eta = \frac{P_{out}}{P_{out} + P_{cu} + P_{core}} \quad (3.8)$$

The power density (p) can be found with the transformer weight which depends on the transformer dimensions and the different material densities (ρ).

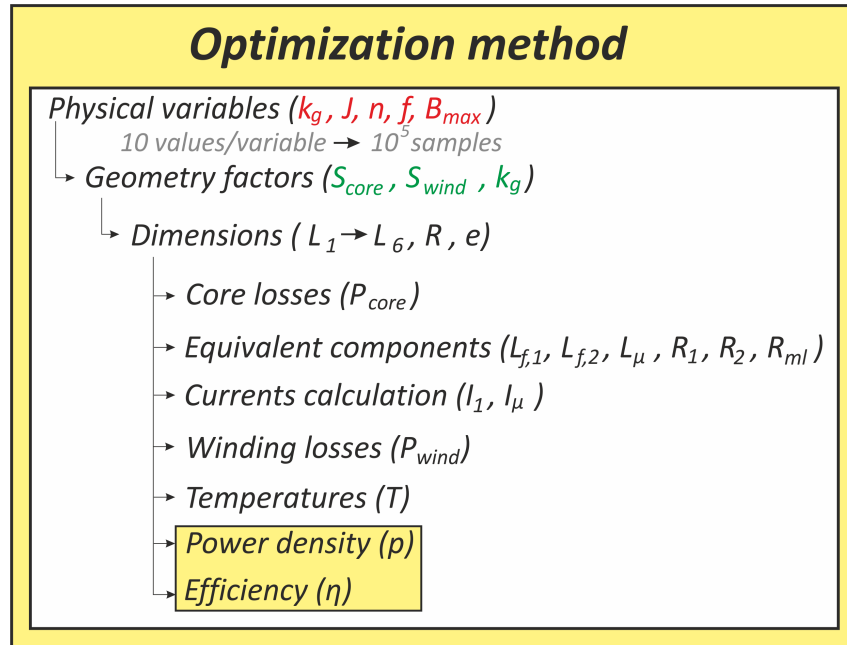


Figure 3.2 – Optimization method diagram

Two constraints are imposed for the optimization. First, the maximal temperature of the transformer cannot exceed 160 °C. Doing this avoids an over-heating in the transformer windings. If the temperature

of a combination exceeds this maximal temperature, the combination is erased from the list. Secondly, the magnetizing current (I_μ) cannot exceed 10% of the output current (I_2). Otherwise, the input current (I_1) would increase too much and the losses generated in the power electronics in the primary side would increase too.

In the calculations, different parameters characterizing the wire and the magnetic core are considered. The wire used is a Litz wire composed of 840 strands of 71 μm diameter (Table 3.1). The core used is a ferrite 3C90 made by Ferroxcube (Table 3.2).

Wire specifications		
Wire diameter (d_w)	3	mm
Number of strands	840	-
Strand diameter (d_s)	71	μm
Wire fill factor ($k_{f,wire}$)	0.44	-
Wire resistivity (ρ_{wire})	5.2	$m\Omega/m$
Thermal conductivity ($k_{th,e}$)	0.18	W/mK
Wire linear density (λ_{wire})	37	g/m
Maximal temperature (T_{max})	165	$^{\circ}C$

Table 3.1 – Litz specifications : Rupalit[®] AWG41 (Appendix B.4)

Core specifications		
Relative permeability (μ_r)	2300	-
Saturation flux density (B_{sat})	0.48	T
Steinmetz coefficient (k)	3.2e-3	-
" " (α)	1.5	-
" " (β)	2.75	-
Thermal conductivity ($k_{th,core}$)	3.5	W/mK
Core density (ρ_{core})	4800	kg/m^3
Maximal temperature (T_{max})	215	$^{\circ}C$

Table 3.2 – Ferrite specifications : Ferroxcube 3C90 (Appendix B.5)

3.4 Optimization results

The optimization method described above was implemented in a Matlab program. The results of the optimization can be seen in figure 3.3. Each combination is pointed on a graph which represents the losses generated ($P_{tot} = P_{cu} + P_{core}$) and the power density (p). For now, the different colors do not matter. The optimization is looking for a point which has minimal losses for a maximal power density. This virtual point would be located in the right-bottom corner of the graph.

From all the combinations, a front can be seen in the lower side of the graph. This front is called the *Pareto Front*. Each combination located on this line is an optimal combination. For all of those, there exists no other combination which has both less losses and a higher power density. The chosen combination for the prototype needs to be located on this curve.

An other front can be seen in the upper side of the graph. This front corresponds to the combinations that have been cancelled because too much losses are generated and the maximal temperature in the transformer is higher than the temperature allowed. Other combinations located in the middle of the graph have also been cancelled because of an excess of temperature. Actually, their geometry does not allow an efficient cooling.

3.4.1 Frequency impact

In figure 3.3, the color of the points represents the operating frequency of the associated combination. It can be seen that the frequency has an impact on the transformer weight. Indeed, low frequency points (in blue) are located on the left side of the graph which corresponds to low power density and high weight. The high frequency points (red) are more located in the right side of the graph where the transformer weight is lower.

The frequency impact on the power losses (P_{tot}) is not clear on the graph. This is due to the interaction of the frequency with the other input variables. Indeed, the frequency has an impact on both winding and core losses. In the windings, it is better to use a low frequency in order to decrease the proximity effect. For a same weight, a high frequency allows to reduce the peak flux density ($\hat{B}$) and reduce the core losses. Indeed, in the *Steinmetz equation* (2.27), the exponent on the flux density is higher than the exponent on the frequency. Then, the optimal frequency is around the middle of the frequency range (≈ 50 kHz).

A similar analysis as the one that just was operated on the frequency impact can be done for the other input variables. But, as the results depend on all the input variables together, it is really hard to draw good conclusions on the impact of a certain variable on the losses from those graphs.

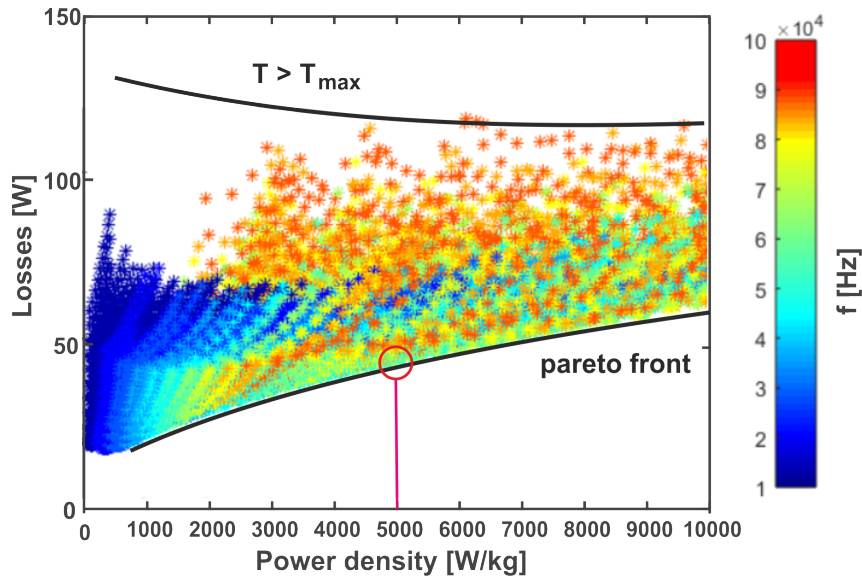


Figure 3.3 – Optimization : Power losses with respect to Power density and peak flux density ($\hat{B}$)

3.4.2 Results

There is not one optimal combination for the transformer. There exist several combinations which are located on the *Pareto Front*. In order to choose a point, it is considered that the transformer weight has to be more or less equal to the slip rings used in a SADM. The weight considered is $M = 2$ kg for a power density $p = 5000$ W/kg. This point is red-circled in figure 3.3. The combination associated is : $\{ k_g = 0.1 ; \hat{J} = 2.33[A/m^2]; n = 8[turns]; f = 40[kHz]; \hat{B} = 0.16[T] \}$.

However, some small modifications on the variables were made in order to fit the winding dimensions with the given Litz wire. Moreover, the winding height is increased in order to let some space for the winding connections. Table 3.3 contains the characteristics of the optimal transformer. Those values are based on the theoretical modelling of the transformer.

Performances			
Rated power	P	10	kW
Rated peak voltage	$\hat{V}$	300	V
Efficiency	η	99.6	%
Core losses	P_{core}	18.5	W
Winding losses	P_{wind}	19.7	W
Total losses	P_{tot}	38.2	W
Transformer ratio	N	1	-
Magnetic force	F_{mag}	9.6	N
Dimensions			
Shaft radius	R	15	mm
Airgap length	e	0.1	mm
Plate thickness	L_1	8	mm
Winding height	L_2	70	mm
Outer tube thickness	L_3	5	mm
Winding thickness	L_4	6	mm
Winding gap	L_5	3	mm
Inner tube thickness	L_6	10	mm
Transformer height	H	86	mm
Transformer diameter	D_o	90	mm
Volume	V	5.47	dm^3
Masse	M	2.2	kg
Variables at rate			
Geometry factor	k_g	0.085	-
Peak current density	$\hat{J}$	2	A/mm^2
Number of turns	n	9	$turns$
Peak flux density	$\hat{B}$	0.16	T
Operating frequency	f	40	kHz
Equivalent components			
Primary dc-resistance	$R_{1,dc}$	2.61	$m\Omega$
Primary ac-resistance	$R_{1,ac}$	5.18	$m\Omega$
Secondary dc-resistance	$R_{2,dc}$	3.45	$m\Omega$
Secondary ac-resistance	$R_{2,ac}$	6.84	$m\Omega$
Primary leakage inductance	$L_{f,1}$	0.47	μH
Secondary leakage inductance	$L_{f,2}$	0.56	μH
Magnetizing inductance	L_μ	450	μH
Core losses resistance	R_{ml}	3	$k\Omega$
Parasitic capacitance	C_p	400	pF

Table 3.3 – Optimized transformer dimensions and characteristics with theoretical calculations

Analysis

The theoretical total losses generated at rated power are $P_{tot} = 38W$. This represents an efficiency of 99.6% which is totally acceptable. As a reminder, the efficiency of the combination of the transformer and the external inductor has to be higher than 99%. The losses generated are well-balanced (48% core losses and 52% winding losses). However, the core losses are underestimated because of the assumption that flux density is constant over the entire core. The transformer weight is 0.2 kg higher than the expected value. This is due to the increase of the window height (L_2) for the wire connections.

Figure 3.4 represents the transformer cross section at scale 1:2. This gives an idea of the dimensions given in table 3.3. The winding is composed of four Litz wires connected in parallel. Two wires in the radial direction and two wires in the axial direction (Figure 3.4). The winding geometry factor k_g is low in order to reduce the proximity effect in the windings ($k_g = 0.085$). As explained in section 2.3, the proximity factor has a large impact on the ac-resistances. The winding thickness $L_4 = 6$ mm corresponds

to two Litz wires in parallel. The total number of layers in parallel is equal to $M = 2 \cdot \sqrt{N_s} = 58$ where N_s is the number of strands ($N_s = 840$). With *Dowell's equations*, the proximity factor for a 40 kHz operating frequency is : $F_R = 1.8$. The harmonics composing the trapezoidal signal increase the proximity effect by a factor $F_H = 1.1$ (10% are added because of harmonics). The total proximity factor that is considered is $F_P = \frac{R_{ac}}{R_{dc}} = 1.98$. Even though the wire strands have a $71 \mu\text{m}$ diameter, the proximity effect at $f = 40 \text{ kHz}$ has a large impact on the winding losses. Those are actually doubled because of this effect.

A magnetic force appears because of the airgaps. Indeed, this force tends to close the magnetic circuit. The stator pulls the rotor in the axial direction. This force is calculated by the *Maxwell's tensor*: $F_{mag} = 9.6 \text{ N}$. Fortunately, this value is not significant and will not have an impact on the mechanic assembly.

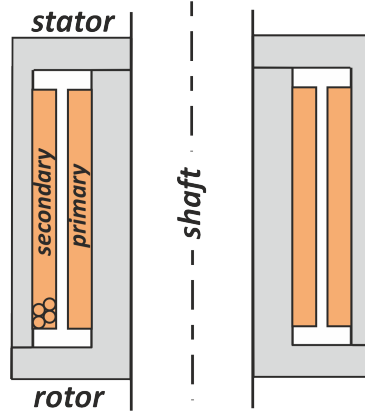


Figure 3.4 – Transformer cross section representation (scale : 1:2)

Figure 3.5 represents the temperature of each transformer part at rated power. The thermal management was modelled by the *lumped parameter thermal model* in section 2.4. The baseplate temperature is fixed to $T_{baseplat} = 65^\circ\text{C}$. As expected, the highest temperature is located in the rotor winding. This temperature reaches $T_{max} = 85^\circ\text{C}$. This is largely below the maximal temperature allowed. The losses generated in the rotor can be assumed to be equal to losses generated in the stator. Indeed, the rotor and stator cores have the same volume. For the winding losses, the secondary winding resistance is slightly higher than the primary winding resistance. However, it is compensated by a higher current in the primary side than in the secondary side because of the magnetizing current ($I_1 = I_2 + I_\mu$). The total thermal resistance of the rotor and the stator can be computed :

$$\begin{aligned} R_{th,rot} &= \frac{85 - 65}{19.1} = 1.05[\text{K/W}] \\ R_{th,rot} &= \frac{78 - 65}{19.1} = 0.68[\text{K/W}] \end{aligned} \tag{3.9}$$

Those values are really low and shows that the transformer geometry is well designed for the thermal management.

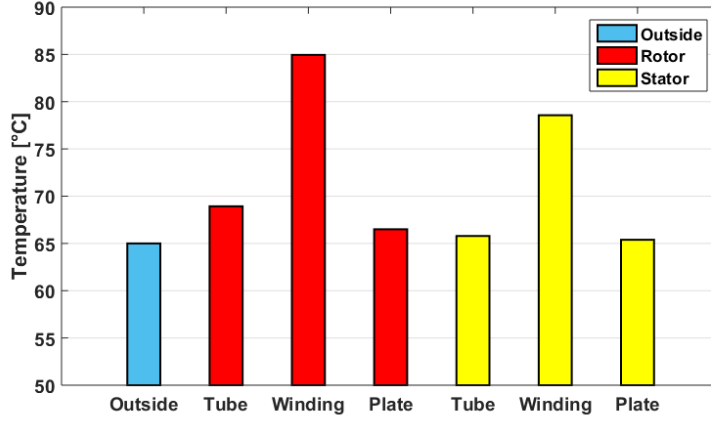


Figure 3.5 – Transformer temperatures at rated power

Finally, it is now possible to evaluate the series inductance value L_s necessary to transfer 10 kW with the Dual Active Bridge. As a reminder, the inductive reactance needed for the power transfer was calculated in section 1.5 : $X_L = 5.3 \Omega$. For an operating frequency of 40 kHz, the series inductance necessary is :

$$L_s = \frac{X_L}{2\pi f} = 21[\mu H] \quad (3.10)$$

As it was expected, this value is largely higher than the transformer leakage inductance ($L_f \simeq 1 \mu H$). Therefore, an external inductor is needed in order to reach the inductance value.

3.5 External inductor design

The design of the external inductor is not as complicated as the transformer design. As explained in section 2.1, the best inductor topology is an airgapped core inductor. This allows to store a high amount of magnetic energy without requiring a too high MMF. This has the advantage to decrease the winding losses. However, as a magnetic core is involved, core losses appear as well.

The external inductor considered has been divided in two equal inductors connected in parallel. This has no direct impact on the inductor design. This choice is made in order to anticipate the tri-port functionality of the transformer. The four parallel wires coming from the transformer secondary side are divided in two pairs. Each pair of wire goes to a single inductor. As those two inductors are connected in parallel, their own value is doubled : $L_{ext,1} = L_{ext,2} = 2 * L_s$. The current flowing through each single inductor is half of the total current. It is possible to have a look at section 5.1.1 for more details about the electric assembly that has just been explained.

For the inductor, the design is not conducted with an optimization. Indeed, it was decided to consider the use of "on the shelf" magnetic cores which are cheaper than machining cores. The core shape considered is composed of two E-cores positioned face-to-face (Figure 3.6). The *E-core* considered is the E71/33/32 made by Ferroxcube (Datasheet available in Appendix B.7). The core material used is the 3C92 ferrite (Table 3.4). The wire used is the same as for the transformer (Table 3.1). For a single inductor, two wires are connected in parallel. The number of turns that can enter the core window is $n_L = 12$.

In section 2.1, the reluctance model was used in order to find the equivalent permeability of an airgapped core inductor. With this model, it is possible to calculate the inductor airgap length l_g and the associated equivalent permeability μ_e . Afterwards, the peak flux density in the inductor can be calculated

Core specifications		
Relative permeability (μ_r)	1390	-
Saturation flux density (B_{sat})	0.5	T
Steinmetz coefficient (k)	3.2e-3	-
" " (α)	1.5	-
" " (β)	2.75	-
Thermal conductivity ($k_{th,core}$)	3.5	W/mK
Core density (ρ_{core})	4800	kg/m ³
Maximal temperature (T_{max})	280	°C

Table 3.4 – Ferrite specifications : Ferroxcube 3C92 (Appendix B.6)

with the peak magnetic energy contained in the inductor airgap :

$$W_{max} = \frac{LI_{max}^2}{2} = \frac{\hat{B}^2}{2\mu_0} \cdot V_{gap} \quad (3.11)$$

The total losses generated in the inductor are calculated in a same way as for the transformer. The core losses are computed by using the *improved Steinmetz Generalized Equation*. This time, the flux density waveform is trapezoidal and the harmonics generate more losses. For a trapezoidal waveform a duty-cycle $D = 0.25$, the losses are increased by 83% (Appendix A.7). For the winding losses, the ac-resistance has to be considered by applying the same proximity factor as for the transformer $F_P = 1.98$ as the same wire is considered and the same current waveform is flowing through.

Table 3.5 contains the main characteristics of a single inductor designed by applying the models introduced in last chapter. The total losses of the two inductors gathered are $P_{tot} = 19W$. The thermal management of the inductor is not theoretically computed. But this point will be analysed in FEM simulations.

Inductor specifications			
Inductance	L_{ext}	40	μH
Rated current	I_{rms}	20.25	A
Number of turns	n_L	12	turns
Airgap length	l_g	1.5	mm
Maximal current	I_{max}	22.3	A
Peak flux density	$\hat{B}$	0.08	T
Equivalent permeability	μ_e	49	-
DC resistance	R_{dc}	6	mΩ
AC resistance	R_{ac}	12	mΩ
Winding losses	P_{cu}	5	W
Core losses	P_{core}	4.5	W
Total losses	P_{tot}	9.5	W
Dimensions			
Height	H	66.5	mm
Length	L	70.5	mm
Width	W	32	mm
Core cross section	S_{core}	683	mm ²
Volume	V	0.23	dm ³
Weight	M	0.65	kg

Table 3.5 – Inductor dimensions and characteristics with theoretical calculations

3.5.1 Fringing effect

In the airgapped-core model, the airgap was modelled by a simple reluctance with a same area than the core cross section. Indeed, it was considered that all the magnetic flux flows through the airgap. However,

as the airgap has the permeability of air, the magnetic flux is not concentrated in the airgap space. A part of the flux spreads and leaves the airgap. This phenomenon is called the *fringing effect*. The longer is the airgap length, the stronger is the fringing effect. This is why it is neglected for the transformer.

The fringing effect has two consequences. First, it reduces the inductor reluctance which implies an increase of the inductance. Secondly, the fringing magnetic flux can generate eddy-currents in the windings. Actually, it has the same consequence as the skin effect and the proximity effect : the current is pushed to the wire strand surface and the effective wire cross section is reduced which increases the ac-resistance.

In order to avoid an impact on the ac-resistance, the winding position is considered. Figure 3.6 represents two kind of windings. The normal winding is not adapted to the fringing effect (Figure 3.6a). As it can be seen, the middle of the winding is really close to the airgaps where the flux density is high. The resistance of this local part increases. That can produce a local over-heating of the winding. The other winding is more adapted because it avoids the window part close to the airgaps (Figure 3.6a). As a rule of thumb, the minimum distance between the winding and the airgap center has to be longer than 3 times the airgap length. But, the flux density generated by the fringing effect is not null after this distance. The winding is still located in a magnetic flux. Unfortunately, it is hard to model the fringing effect on the winding resistance. Nevertheless, this point needs to be considered carefully for the FEM simulations in next chapter.

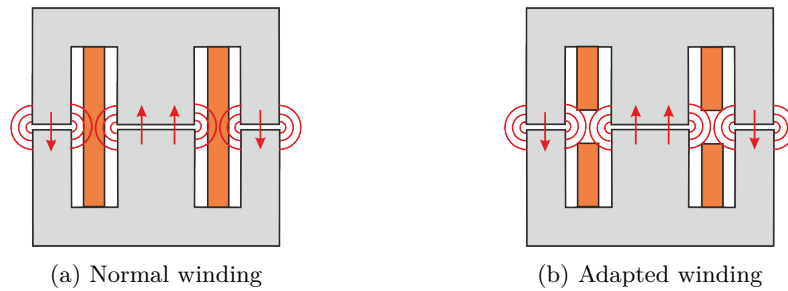


Figure 3.6 – Inductor : fringing effect on windings

Finite Element Analysis

After getting values from analytical simulations, it is time to simulate with a *Finite Element Method* (FEM) software. FEM simulations are more accurate since this is the application of the Maxwell equations on small elements. With the transformer modelling, some assumptions were made and the results obtained were less accurate. The objective of this chapter is to validate the results found by optimization.

The Finite Element software used is *Comsol Multiphysics*. This software is able to simulate different physics in 2D and 3D. Moreover, it has the capacity to simulate different physics (modules) at the same time and to link those. For the transformer and inductor simulations, three modules were used at the same time:

- *Magnetic Field* : Maxwell equations are applied on the transformer geometry. The input variable is the current which is calculated in the electric circuit module. The output is the flux density for each element.
- *Electric Circuit* : Currents and voltages are computed in function of the circuit components. The inductances and resistances in the transformer are also considered by a connection with the magnetic field module
- *Thermal Transfer* : Fourier equations are applied on the transformer geometry. The input variables are the heat generated in the core and the windings which depends on the current and the flux density. The output is the temperature for each element.

Flux densities are the real improvement of Finite Elements (FE) with respect to the models established in chapter 2. Indeed, in the model, the flux density was assumed to be constant over the core cross section. With FE, the flux density can be calculated locally.

Unfortunately, it is not possible to validate the winding losses with the FEM. Actually, it is really hard to simulate the Litz wire behaviour because it requires too many elements and too much computation time. The characteristic parameters that will be validated in this section for the transformer and the inductor are : the average peak flux density ($\hat{B}_{av}$), the temperature (T), the series and magnetizing inductances (L_{ext} , L_f and L_μ), the core losses (P_{core}) and the magnetic force (F_{mag}) for the transformer.

First, the transformer is analysed in 2D for the magnetic field and the thermal transfer analysis. Afterwards, the impact of the winding connector slots on the magnetic fields behaviour is analysed in 3D. Finally, the inductor is analysed in 2D for the magnetic field and the thermal transfer analysis.

4.1 Transformer analysis (2D)

In this section, the transformer is simulated in 2D for the magnetic field and the thermal transfer. Table 4.1 is a comparison between the analytical results given by the models introduced in chapter 2 and the results obtained with 2D FEM simulations. The relative difference is also given for information.

Magnetic parameters			Analytical	FEM 2D
Average peak flux density	$\hat{B}_{av}$	[T]	0.16	0.157 (-2%)
Magnetizing inductance	L_μ	[μH]	450	460 (+2%)
Leakage inductance	L_f	[μH]	1.04	1.13 (+8%)
Core losses	P_{core}	[W]	18.5	20 (+8%)
Magnetic force	F_{mag}	[N]	9.6	8.2 (-14%)
Thermal parameters				
Maximal temperature	T_{max}	[$^{\circ}C$]	85	85.1 (+0.5%)
Rotor resistance	$R_{th,rot}$	[$^{\circ}C/W$]	1.05	1.01 (-4%)
Stator resistance	$R_{th,sta}$	[$^{\circ}C/W$]	0.68	0.6 (-10%)

Table 4.1 – Comparison on transformer characteristic parameters for analytical results and 2D FEM simulations

4.1.1 Magnetic field simulation (2D)

Figure 4.1 presents the peak flux density in the transformer. As there is a large difference between the magnetizing flux in the core and the leakage flux in the window, two figures are used to clearly see the flux density. Figure 4.1a presents the flux density in the magnetic core and the airgaps while figure 4.1b illustrates the flux density in the window.

As expected, the flux density is not constant around the magnetic core (Figure 4.1a). In the plate, the core section perpendicular to the flux direction increases with radius and the flux density is then reduced. Likewise, the flux density in the core is concentrated locally at the corners. At these points, it can reach 0.34 T which comes close to saturation ($B_{sat} = 0.5T$). However, it is local and it has no impact on the transformer behaviour. It can be seen that the corners located on an airgap have a lower flux concentration. This is due to the low fringing effect happening at the corner because of the airgap. A part of the magnetic flux "leaves" the airgap and reduces the flux concentration.

However, the concentrations are really local and the core losses calculated with FEM are 8% larger than the losses analytically calculated (1.5 W). This is due to the assumption considering that the flux density is constant over the core section. The magnetizing inductance is similar to the value calculated previously. The 2% difference comes from the fringing effect which was neglected in the reluctance model. The fringing effect has also an impact of 14% on the magnetic force between the rotor and the stator.

In figure 4.1b, it can be seen that the fringing effect of the airgaps does not have a significant impact on the windings as those are not close to each other. Moreover, it can be seen that the flux density in the windings area has the same trapezoidal shape as the MMF. For the calculation of the leakage inductance with FEM simulations, a short-circuit test was simulated. The increase of 8% with respect to the analytical leakage inductance is caused by side effects of the winding leakage energy and also by the fringing effect.

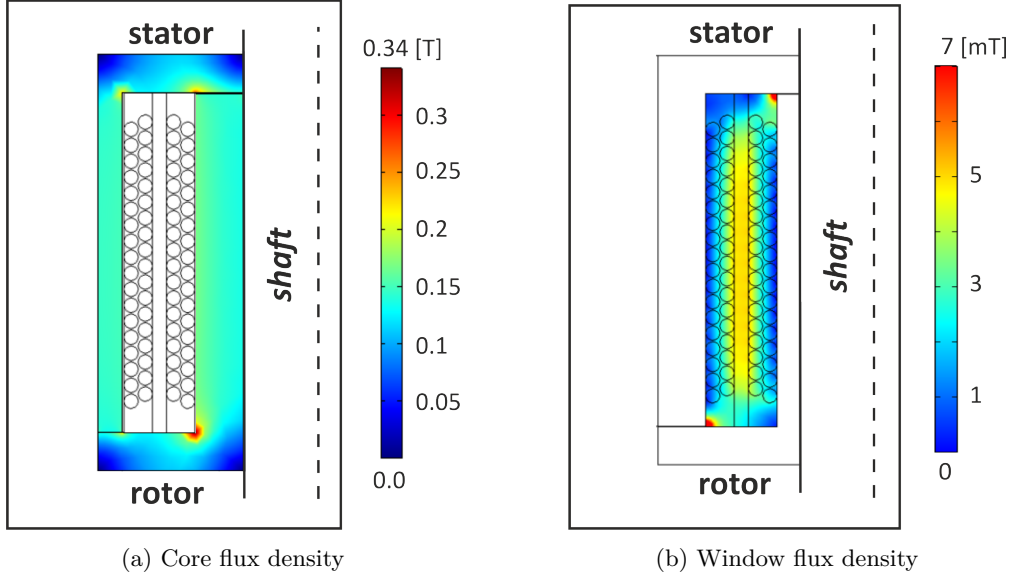


Figure 4.1 – Flux density in the transformer (2D)

4.1.2 Thermal transfer simulation (2D)

The thermal transfer of the transformer was simulated at steady-state. Indeed, with a 40 kHz frequency, the thermal capacitances are neglected. The heat generated by the windings are taken from the analytical results but the heat generated in the core is computed by applying the *improved Steinmetz Generalized Equation* locally on each element. Boundary conditions are represented in figure 4.2a. There are two kind of boundary conditions : fixed temperature boundaries and adiabatic boundaries. The first ones correspond to the baseplate temperature in green which is fixed to 65°C . The adiabatic boundaries (in orange) correspond to boundaries where no heat can flow because of the vacuum. As explained before, there is no heat exchange between the rotor and the stator as there is no contact.

Figure 4.2b represents the temperature in the transformer. As expected, the maximal temperature is located on the outer surface of the rotor winding. The maximal temperature estimated with the analytical model fits the maximal temperature found in the simulation : $T_{max} = 85,1^{\circ}\text{C}$. Actually, all the temperatures calculated with the *lumped parameter thermal model* are close to the simulation results. The simulation confirms that the transformer geometry is well designed for the thermal management.

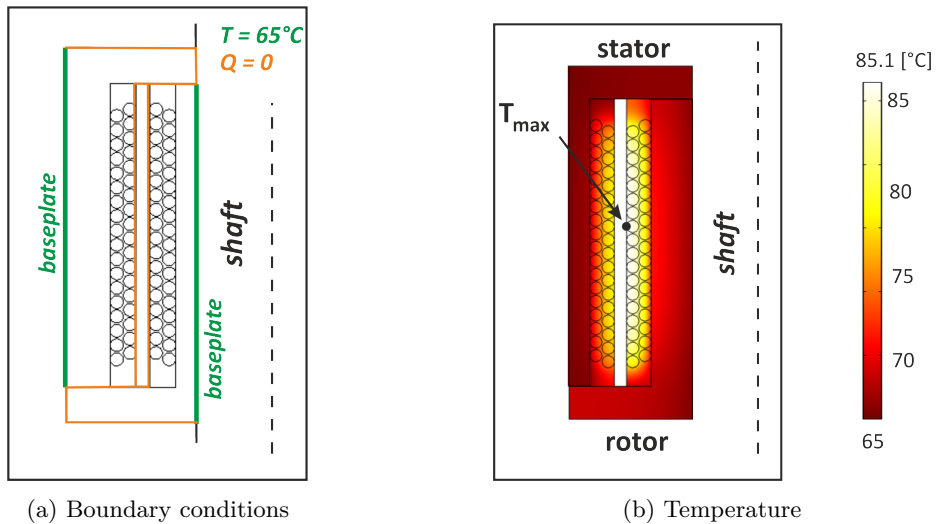


Figure 4.2 – Thermal transfer in the transformer (2D)

4.2 Transformer analysis (3D)

So far, the transformer has always been considered in a 2D plan because of its axisymmetry. However, the winding connectors cannot be considered in 2D. Indeed, it is compulsory to have a connection between the windings and the outside of the transformer. In this section, the winding connectors implemented are explained and a magnetic field simulation in 3D is operated in order to evaluate the impact on the magnetic characteristics.

4.2.1 Winding connectors

The winding connectors are designed with the objective to have a low impact on the magnetic characteristics of the transformer. Moreover, as there is no contact between the rotor and the stator, each magnetic core has its own connectors.

There are several possibilities for the connectors geometry. At first, it was considered to use a single hole through the tubes, along their length and with a thickness equal to the winding thickness ($L_4 = 6$ mm). However, this connectors had two disadvantages. First, it generates a loss of the tube cross section area. This can be compensated by increasing the tubes thickness but it implies to increase the weight and the losses. Secondly, a single connector generates a loss in the degree of symmetry in the transformer.

The chosen connector geometry is depicted in figure 4.3. For each core part, four connectors are used. As the winding is composed of four Litz wire in parallel, this corresponds to one connector per wire. The connector slots are located in the plates. The same wire comes in and out from the same slot. For the rotor (Figure 4.3a), the wire comes in by the external side of the plate slot, then it is wound around the tube, and it finally leaves the transformer by a groove located on the tube surface. For the stator (Figure 4.3b), the wire comes in by the internal side of the plate slot, then it is wound inside the tube, and it finally leaves by a groove located on the tube. Each connector slot has a thickness of 3 mm which corresponds to a single wire diameter. The resulting transformer assembly is presented in figure 4.3c.

This method has many advantages. First, the fact to separate the four wires for the connector allows a better malleability of the wires that offers a gain of space in the window. Afterwards, the degree of symmetry in the transformer is kept high which implies a good magnetic field distribution. There is at least two symmetry plans for each rotor position.

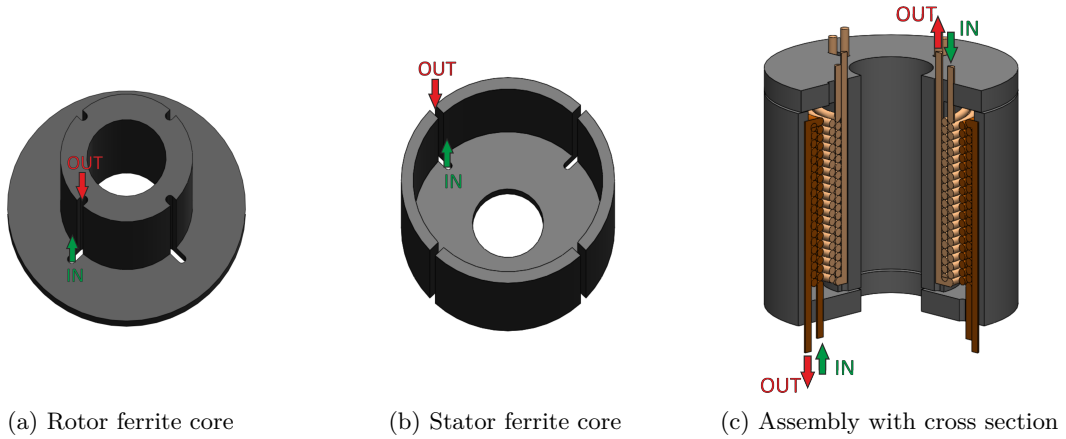


Figure 4.3 – Transformer : 3D representation of the magnetic cores (rotor and stator) and the assembly (rotor, stator and windings)

4.2.2 Magnetic field simulation (3D)

A 3D analysis was simulated with the FE software in order to evaluate the impact of the connector slots and grooves on the magnetic characteristics of the transformer. Although the grooves in the tubes reduce

the core section (-3% for the rotor tube), it was not compensated.

Table 4.2 is a comparison between results from the 2D and 3D FEM simulations. As it was expected, the average peak flux density $\hat{B}_{av}$ is increased by 3% because the core section has decreased. The impact is the same for the magnetizing inductance L_μ which is slightly decreased.

Magnetic parameters			FEM 2D	FEM 3D
Average peak flux density	$\hat{B}_{av}$	[T]	0.157	0.162 (+3%)
Magnetizing inductance	L_μ	[μH]	460	440 (+4%)
Core losses	P_{core}	[W]	20	21.5 (+7%)
Magnetic force	F_{mag}	[N]	8.2	7.5 (-8%)

Table 4.2 – Comparison on transformer characteristic parameters for 2D FEM simulation and 3D FEM simulation

Figure 4.4 represents the magnetic field in the transformer core sections close to the airgaps. Figures 4.4a and 4.4b are sections located around the external airgap while figures 4.4c and 4.4d are sections located around the internal airgap. Contrarily to what it can be expected, the magnetic flux concentrations do not happen at the corners with an airgap but it only happens for the slots located on the rotor plate. This can be easily explained. In figure 4.4a, the magnetic flux flows in the radial direction from the inside to the outside of the plate. The flux part located behind a slot needs to flow around this slot. Therefore, the flux density at the sides of the slot is increased because of this flux concentration. This can be avoided by cutting the rotor tube as it is done for the stator tube. In that case, there would not be any flux stuck behind a slot. However, this is critical for the rotor tube cross section which would be decreased by 10%. Moreover, the increase generated by the concentration around the rotor plate slots is not too high and has not a significant impact.

The impact of the incoming and outgoing wires was also evaluated. Those wires are directed in the axial direction and perpendicular to the windings. As each wire conducts a current of 10A, it can produce a magnetic flux density which is perpendicular to the magnetizing field. However, this is not the case. Indeed, as the incoming and outgoing wires are passing in the same slot, the magnetic flux generated by one wire is cancelled by the other. Therefore, no effect is generated by the incoming and outgoing wires.

The core losses P_{core} are slightly increased by +7% because of the increase of the average flux density and the local flux concentration around the slots.

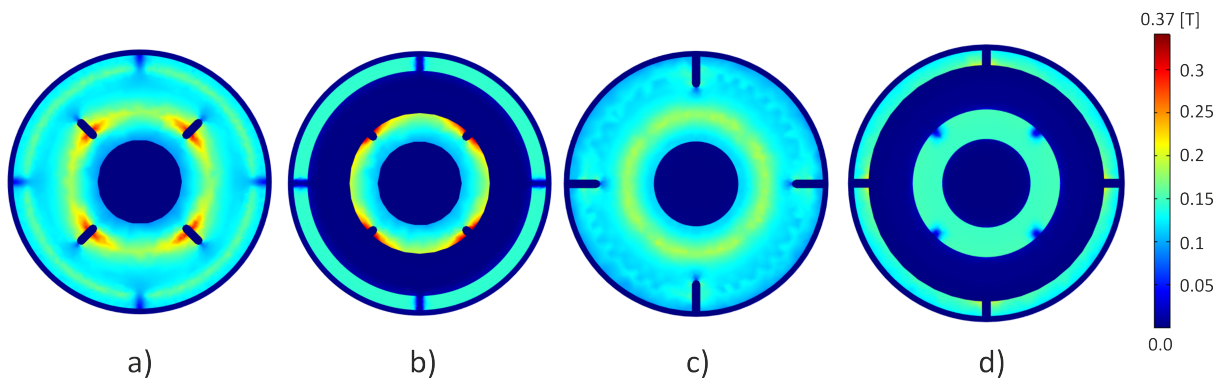


Figure 4.4 – Cross section representation of the magnetic field close to the airgaps in the transformer (3D)

4.3 Inductor analysis (2D)

In this section, the external inductor is simulated in 2D for the magnetic field and the thermal transfer. The cross section representations in figures 4.5 and 4.6 shows only one half of the *E-core* inductor. Indeed, symmetry was applied on the simulations for computation time benefit.

4.3.1 Magnetic field simulation (2D)

After a first set of simulations, it comes that the fringing effect has a significant impact on the inductance of the inductor which is 20% higher than the value calculated. This increase is caused by the window reluctance which is not neglectable. In order to reach the desired inductance ($L_{ext} = 40\mu H$), the airgap is increased to 2 mm. Table 4.1 is a comparison between the results found with a 1.5 mm airgap in section 3.5 and the results obtained with a 2 mm airgap for the 2D FEM simulations.

The average peak flux density in the core $\hat{B}_{av}$ has increased compared to the calculations because a part of the flux is flowing through the window. The consequence is that the core losses are increased, in fact doubled. Figure 4.5a represents the flux density in the magnetic circuit. Once more, there is a flux concentration in the window corners.

Magnetic parameters			Analytical	FEM 2D
Inductance	L_{ext}	$[\mu H]$	40	39.3
Airgap length	l_g	$[mm]$	1.5	2
Average peak flux density	$\hat{B}_{av}$	$[T]$	0.08	0.105
Core losses	P_{core}	$[W]$	4.5	9
Total losses	P_{tot}	$[W]$	9.5	14

Table 4.3 – Comparison on inductor characteristic parameters for analytical results and 2D FEM simulations

Figure 4.5b shows the fringing effect caused by the airgaps. The flux density is relatively high in the winding. This varying flux density increases the equivalent skin depth of the wire and this has an impact on the ac-resistance of the wire. This is critical for the wires which are close to the airgaps where the flux density is higher. Indeed, it can cause a local hot point because of a local high resistivity. It is really hard to quantitatively evaluate the fringing effect impact on the ac-resistance. Actually, it is not possible to add the fringing effect in Dowell's equations directly. Some papers have brought up the topic [32] but the method does not seems consistent. The best way to evaluate the effect is to test it with a prototype and to probe temperature in the windings. If the impact is too large, some solutions can be brought :

- The number of turns can be decreased to $n = 10$ but it would imply an increase of the flux density in the core and the core losses would increase by 5 W;
- The winding height can be reduced by making a two-layer winding. However, the proximity effect would be increased by a 2.4 factor. Then, the winding losses would be increased by 7 W;
- It is also possible to divide each airgap in two smaller airgaps. The flux density distribution in the window would be more uniform but still high;
- The use of a powder core might be more suitable in this case. Indeed, powder cores are equivalent to ferrites with a uniformly spread airgap. However, the core losses are higher in a powder core than in a ferrite.

All solutions are interesting but the first step consists of the evaluation of the fringing effect with a 2 mm airgap.

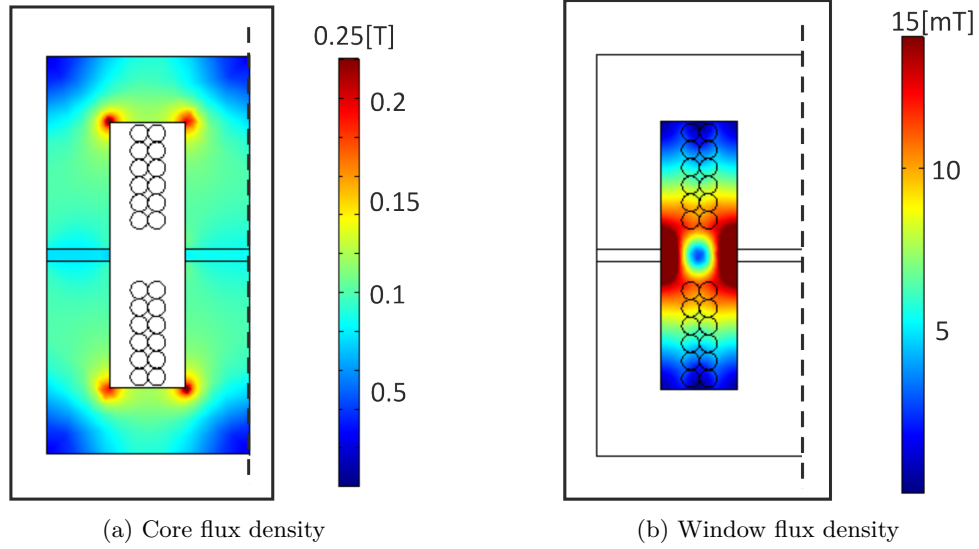


Figure 4.5 – Flux density in the inductor (2D)

4.3.2 Thermal transfer simulation (2D)

A thermal simulation in 2D has also been performed on the inductor. The methodology is the same as for the transformer thermal analysis.

The core losses are locally computed thanks to the *improved Steinmetz Generalized equation*. The winding losses calculated in the previous chapter were added manually. For the boundary conditions (Figure 4.6a), the baseplate temperature was imposed on the top and bottom of the inductor and the adiabatic boundary is set on the sides. The window is filled of a resin to conduct the heat generated in the windings.

Figure 4.6b shows that, once again, the maximal temperature is located in the winding. This maximal temperature ($T_{max} = 101^{\circ}C$) is quite high because the inductor geometry is not perfectly suitable for the thermal management. Indeed, there is a thick layer of resin between the winding and the cores. As resin has a low thermal conductivity, the thermal gradient between the core and the winding is high. However, this temperature is still really far from the maximal temperature allowed : $160^{\circ}C$.

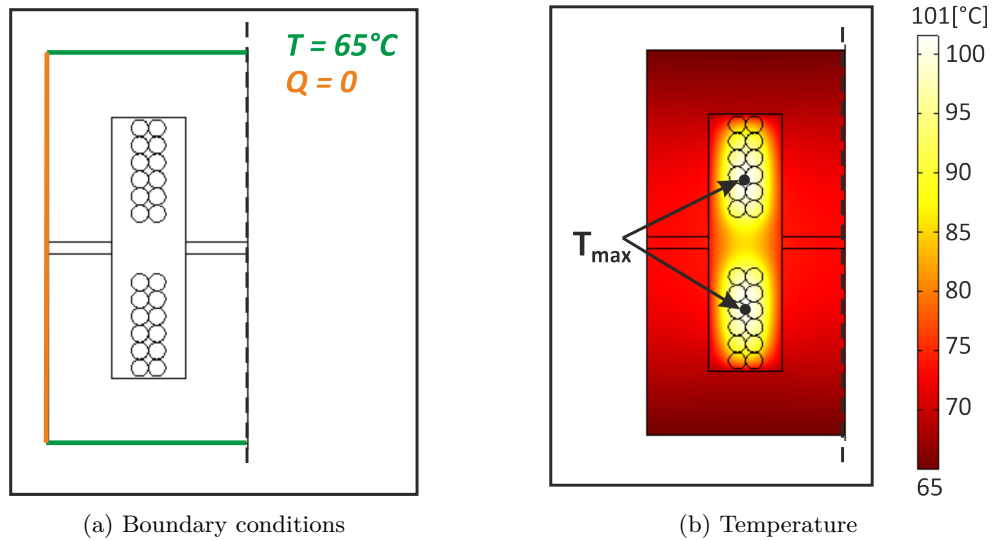


Figure 4.6 – Thermal transfer in the inductor (2D)

This last chapter is dedicated to the construction of the Inductive Power Transfer through a rotating transformer. Unfortunately, a functional transformer prototype has not resulted. Indeed, the manufacturing of the core material went wrong and breakage happened. Therefore, the system has not been validated yet because no measurement has been achieved. However, a visual prototype has been realized in order to get a concrete visual appearance of the system. That also validates the mechanical assembly of the rotating transformer.

First, the electric and mechanical assemblies are introduced. Afterwards, the ferrite manufacturing issue is explained. Finally, the visual prototype is presented.

5.1 Electric assembly

5.1.1 Components connection

Figure 5.1 represents the electric assembly for a bi-port function. The main component of the system is the rotating transformer which has been optimized in chapter 3. In the transformer, each winding is composed of four Litz wires connected in parallel. These four wires are represented in the connections of the assembly representation (Figure 5.1). For each wire, there is one entry connection and one exit connection.

As explained in section 1.4, the power electronics associated to the transformer is a Dual Active Bridge which is composed of two full-bridges : DC/AC at the primary side and AC/DC at the secondary side. The Dual Active Bridge requires a high series inductance which cannot be provided by the transformer. Therefore, an external inductor is added in series. This inductor is divided in two equal parts in order to allow a tri-port functionality which is introduced in next section. Thereby, the four wires coming from the transformer secondary winding are divided in two pairs. Each wire pair goes to a single inductor. For a bi-port functionality, the wires coming out from the two inductors are connected to the same full-bridge. In that case, those are connected in parallel and the equivalent inductor can store a large amount of energy.

Regarding fault tolerance, the system presented in figure 5.1 has a high robustness. Indeed, using four wires in parallel increases the fault tolerance degree. The condition is that each wire needs to be connected to its own full-bridge. In that case, if a problem happens for one wire, it can be disconnected and the system can continue to work. The impact is an increase of windings resistance implying winding losses increase of 33%.

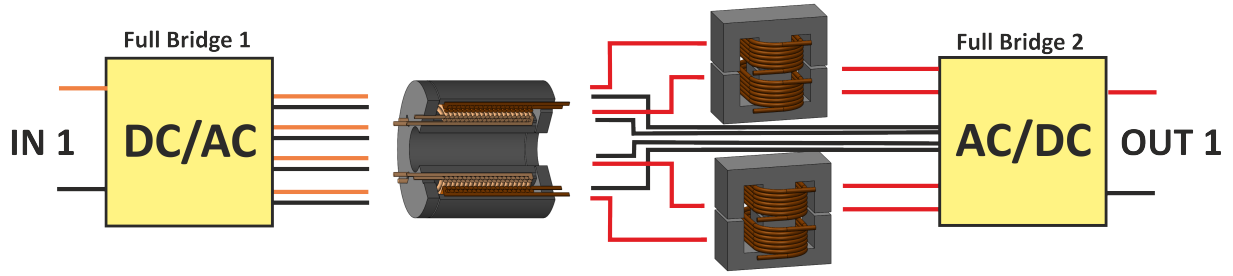


Figure 5.1 – Electric assembly : bi-port (1 in / 1 out)

5.1.2 Transposition in transformer windings

This subsection is a side note about the necessity to add a transposition in the transformer windings.

The winding is composed of four wires in parallel. Since the beginning, it is assumed that a quarter of the total current is passing through each wire. However, this is true only if each wire has the same impedance. At high frequency, this is the series inductance that dominates the wire impedance. Indeed, the ac-resistance is around $10\text{ m}\Omega$ while the series inductive reactance is around $1\text{ }\Omega$. Therefore, it can be assumed that each wire has the same impedance if these have the same leakage inductance, and thus see the same leakage flux. An example is described to demonstrate the necessity of the transposition.

Figure 5.2 represents two disposition of a coil composed of two wires in parallel : an inner wire and an outer wire. For the first case, no transposition is operated (Figure 5.2a). The inner loop has an area lower than the outer loop. Then, the magnetic flux is higher in the outer loop, its leakage inductance is higher and less current flows in it. In the second case (Figure 5.2b), a transposition is operated at halfway. Each wire is half of the way in the inside and half of the way in the outside. As it can be seen, each wire has the same loop area. Therefore, the flux and the current are the same.

This example can be applied to the transformer. If no transposition is operated, a larger part of the current would flow in the inner wire of windings. With only one transposition (after $\frac{n}{2} = 4.5$ turns), the problem is solved. The transposition is compulsory for the primary side where the series inductance is composed of the leakage inductance only. For the secondary side, the series inductance is dominated by the external inductance. The leakage inductance is not significant and the transposition is not compulsory.

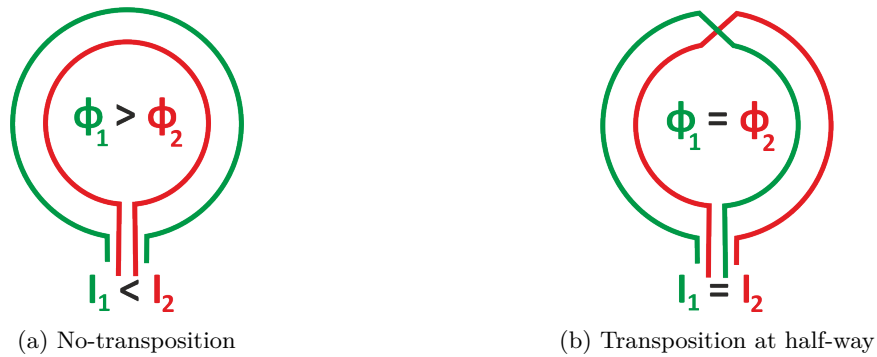


Figure 5.2 – Effect of transposition on the current distribution for a loop composed of parallel wires

5.1.3 Tri-port mode

An advantage of the Inductive Power Transfer is that a tri-port isolated connection can be achieved thanks to the transformer. The transformer winding and the external inductors are designed in consideration of the tri-port functionality. Figure 5.3 represents the tri-port mode electric assembly. The difference with respect to the bi-port mode (Figure 5.1) is that the wires coming out from the two external inductors are

not connected. Each pair of wire is connected to a different full-bridge corresponding to the two output ports.

As a reminder, for the SADM application, the input port is connected to the solar array. The two output ports are the battery and the bus. The power control of a tri-port mode with a Dual¹ Active Bridge is not analysed in this document. However, the topic has been brought up in different papers [33].

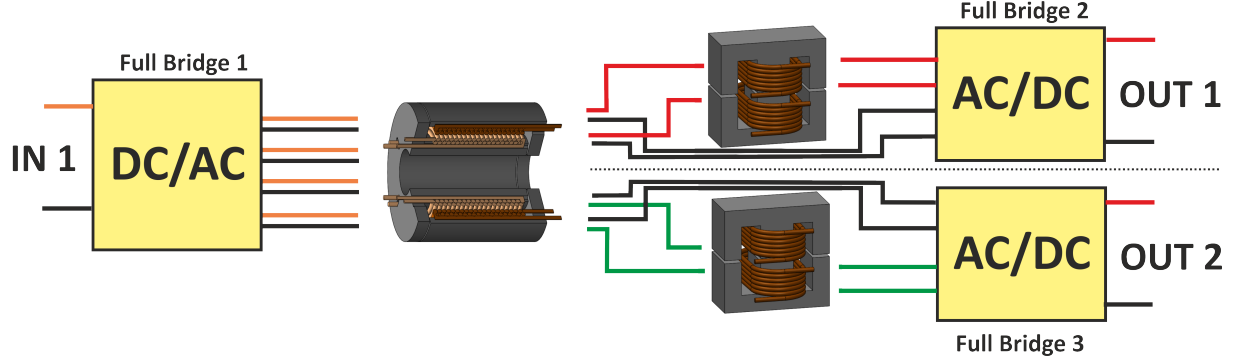


Figure 5.3 – Electric assembly : tri-port (1 in / 2 out)

5.2 Mechanical assembly

The prototype mechanical assembly is composed of two main parts : the magnetic core made of ferrite and the mechanical support made of aluminium. The mechanical support is used to simulate the mechanical interface of the SADM in the satellite.

5.2.1 Magnetic core

The two magnetic core parts have a particular shape. Indeed, the shape was optimized for the dedicated application without considering an "on the shelf" product. Without surprise, these ferrite shapes do not exist on the market. Even if pot cores can be machined to fit the same shape, the pot cores on the market are not large enough for the transformer. The use of a mold is not possible because it is too expensive. Also, ferrite manufacturing is complicated because of its fragility. The 3D machining use on ferrite is really risky and it is easier to machine in 2D (in a plan).

Innovation was needed to obtain the desired core shapes. The idea selected consists of decomposing the rotor core and the stator core in layers. Those layers are represented in figure 5.4. Each plate is a layer of 8 mm and each tube is composed of two layers of 35 mm. For each layer, a ferrite block with a height of either 8 or 35 mm is used. The blocks are cut in the plan in order to obtain the desired geometries. Finally, the different layers are glued together. The glue generates a supplementary airgap that can reach 10 μm . This can be compensated by decreasing the real airgap from 100 μm to 80 μm .

¹Note : the word "Dual" has lost sense

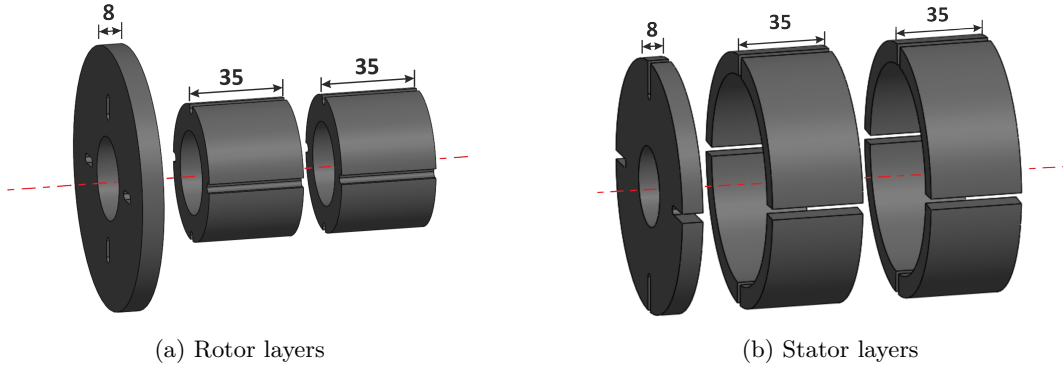


Figure 5.4 – Rotor and stator layers constitution

5.2.2 Support

The mechanical support is used to simulate the mechanical interface of the SADM in the satellite during the tests and measurements. The mechanical assembly is represented in figure 5.5. First, the rotor is fixed on a hollow shaft in order to ease the thermal convection of the heat during the tests. This shaft is cantilevered with a plain bearing. Afterwards, the stator is glued on the cantilever block. Then, the airgap of $80\ \mu\text{m}$ is controlled via a thin teflon sheet (PTFE) pressed between the rotor and the stator. The pressure on the PTFE sheet is given by the magnetic force which tends to close the magnetic circuit. This force is in the range of 7 to 10 N. After, a groove is made on the shaft surface in order to avoid a direct contact between the stator core and the shaft. Moreover, it allows to avoid a local heating of the shaft because of the small fringing effect happening at the airgap. Finally, a fixing nut is screwed on the shaft to avoid that the shaft goes away.

The rotation does not need to be provided by an external motor as this rotation is really slow (1 turn per day). Therefore, the rotor position is controlled by rotating the shaft with hands. Also, the friction forces are not considered as this is not the point of the tests and measurements. That is why a simple bearing is used instead of using a rolling-bearing.

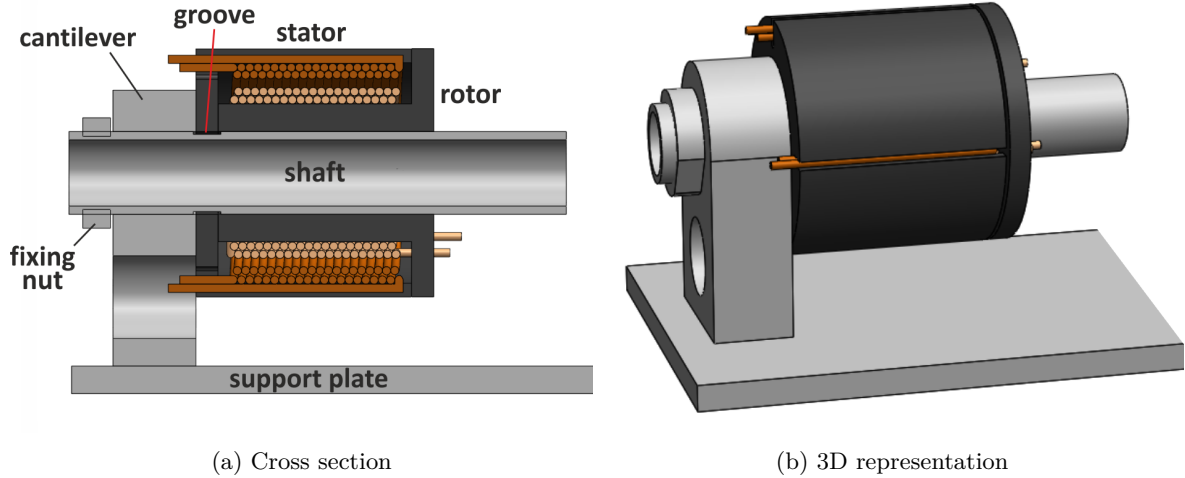


Figure 5.5 – Mechanical assembly : CAD representation of the assembly

5.3 Ferrite manufacturing

As explained earlier, the ferrite manufacturing is applied on blocks of 8 mm and 35 mm height. The manufacturing consists of cutting these plates in 2D. However, ferrite is a ceramic and those materials are highly fragile. Therefore, it is not possible to machine a ferrite block with conventional machines.

There exist different methods for the ferrite machining. The two most efficient methods are the water jet and the femto laser. The water jet method consists of propulsion of a water jet on a small area at a pressure over 200 MPa [34]. The high local force allows to break atoms connections. The femto laser consists of producing laser pulses with high energy during a short time (a few femtoseconds)[35]. A typical laser cannot be used because it generates a lot of heat. The ferrite temperature would reach its Curie temperature from which its magnetic characteristics would be modified.

The chosen machining technique was the water jet cutting. However, the process did not go well. As explained in the introduction, some ferrite blocks broke off during manufacturing. Figure 5.6 are pictures of the ferrite blocks after machining. The first picture is the stator plate which is nearly perfect. The two other pictures are broken blocks of each size. The failure does not seem to come from the water jet itself. The hypothesis is that the clamping applied a too high stress in the plan. When a part of the ferrite was cut, the stress increased locally and the ferrite blocks broke off. If the clamping is done in the perpendicular direction of the plan, the local stress concentration should be minimized. Studies on the topic are still in process.

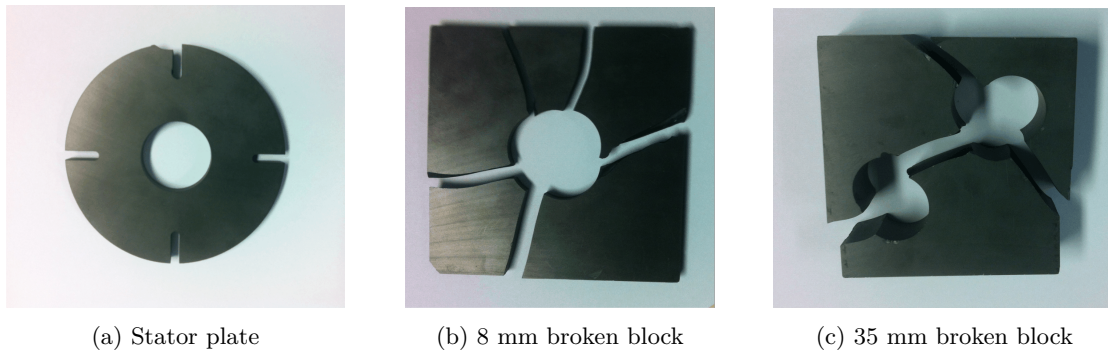


Figure 5.6 – Pictures of the ferrite manufacturing results

5.4 Visual prototype

In order to give a mechanical validation and a visual idea of the system, the core parts were printed with additive manufacturing. Of course, as the core is in a polymer, the transformer does not work. The goal of this visual prototype was to give a mechanic view of the IPT. The assembly was performed without any problem and the mechanic system is validated. Figure 5.7 is a set of pictures of the visual prototype.

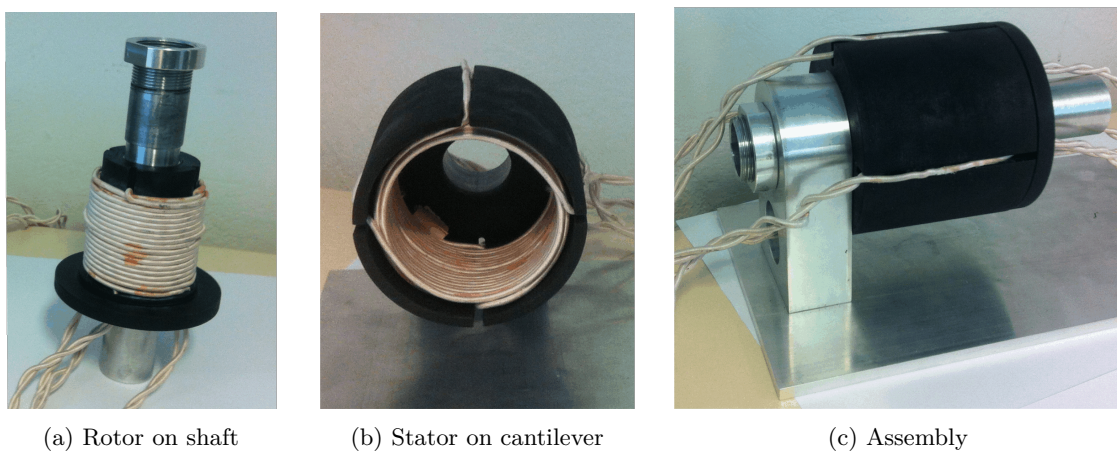


Figure 5.7 – Visual prototype : transformer prototype with core made by additive machining

Conclusion and further work

The objective of the project was to design an Inductive Power Transfer through a rotating transformer for the Solar Array Drive Mechanism of a satellite. The system had to perform the transfer of a 10 kW power under 300V with a minimum efficiency of 98% in order to compete with the current slip rings technology.

First, an introduction of the satellite power systems helped to understand the interaction between the SADM and the other systems. From that, it comes that the Inductive Power Transfer turns out to be a good solution to gather the SADM and the Power Conditioning Unit (PCU) in a same single system. Indeed, the use of an IPT requires a DC/DC converter. That same converter can be used for the Maximum Power Point Tracker (MPPT) of the solar array power regulation. Moreover, the transformer ensuring the rotation can be used to provide three ports (tri-port mode). With a Dual Active Bridge (DAB) offering a bidirectional power transfer, the Battery Charge and Discharge Regulator (BCDR) can be added in the system functionalities. That would allow to make a connection between the solar array, the batteries and the bus in a single device. The number of systems would be considerably decreased implying a volume, weight and cost reduction. The DAB generates currents with a trapezoidal waveform and a transformer flux density with a triangular waveform. The DAB also requires to store a high amount of inductive energy. Unfortunately, the series inductance necessary cannot be assumed by the transformer leakage inductance alone. Therefore, an external inductor had to be connected in series. The total losses generated in the full-bridges of the DAB are 100 W. That represents half of the maximum losses specified in the objectives.

Before the design, some models were defined in order to represent analytically what happens in the transformer. Those models are tools on which the design has been based. First, the *Steinmetz equation* was generalized for trapezoidal and triangular flux density waveforms. The harmonic content has an impact on the core losses. This can be neglected for a triangular waveform but it is significant for a trapezoidal waveform such as for the external inductor. Afterwards, the impact of a high frequency current on the winding losses was studied. The skin effect and the proximity effect imply the use of Litz wire to reduce the losses. Moreover, in order to reduce proximity effect, the most suitable winding shape is a long and thin tube. However, the resulting proximity factor considering trapezoidal harmonics still generates a 98% increase of the winding losses. Finally, the thermal management was modelled by a lumped parameter model. The Litz wire thermal conductivity was derived. It comes that the equivalent thermal conductivity of windings is barely 15% better than the thermal conductivity of an insulation.

An optimization has been performed for the transformer design. The outcomes resulting to the optimization were validated by Finite Elements simulations in 2D and 3D. Those simulations also allowed to reach more accurate results. It comes that the winding connectors do not have a significant impact on the performances. The windings are composed of four Litz wires connected in parallel. This allows to have a good symmetry and to have a high fault-tolerance degree. Finally, for 10 kW transferred, the transformer generates 41.5 W of losses. The inductor, which was quickly designed with on the shelf cores, generates 27 W of losses. All in all, the total losses are estimated

to 70 W. However, this does not consider other losses such as the wire connecting each component or the increase of winding losses in the inductor due to fringing effect. But, it can be assumed that the 98% efficiency for the entire Inductive Power Transfer is achieved.

The maximal temperature drop inside the transformer is $20^{\circ}C$. This high thermal management efficiency is due to a suitable transformer geometry that maximizes the winding contact area in the radial direction. The ferrite acts as a thermal drain. All the characteristic values of the IPT designed are contained in table 5.1.

Unfortunately, a functional prototype has not resulted because of manufacturing issues of ferrite. However, a visual prototype was built in order to validate the mechanical assembly. This also gives a nice visual appearance of what the functional prototype will look like.

The work on this project is not finished. Actually, it has barely begun and several steps have to be done. First, a functional prototype has to be realized. For that, a material study on ferrite cutting is currently in process. If it comes that ferrite is not mechanically adapted, other possibilities for magnetic materials will be explored. Afterwards, with a functional prototype, tests and measurements could be performed. Another point to develop is the Dual Active Bridge control for this transformer. It has to be done for bi-port and tri-port modes. Talking about tri-port, a second rotating transformer should be specially designed for the tri-port mode. Indeed, the current prototype is designed for a bi-port mode and has the tri-port functionality but it is not designed for it.

At the end of this project, I would like to share my personal feeling on it. Since the beginning of this master thesis, I have been highly motivated on the subject. I make no secret by saying that, when the manufacturing went bad, I was really disappointed. This is even more true when you have spent weeks to adjust again and again the design with Solidworks and Comsol. But at the end, I am glad of the results and I am convinced that the Inductive Power Transfer is able to be competitive. I am motivated to continue on the project and I hope that this work is the foundation of a future product that will go up there.

Specifications			
Rated power	P_{out}	10	kW
Rated load voltage	V_{load}	300	V
Rated load current	I_{load}	33.3	A
Losses & efficiency			
Total efficiency	η	> 98	%
Dual Active Bridge losses		100	W
Transformer losses		41.5	W
External inductor losses		27	W
Transformer characteristics			
Diameter	D_o	90	mm
Height	H	86	mm
Airgap length	e	0.1	mm
Weight	M	2.2	kg
Maximal temperature	T_{max}	85	$^{\circ}C$
Core material		Ferrite	
Winding material		Litz wire	

Table 5.1 – Final results for the Inductive Power Transfer for a Solar Array Drive Mechanism

Bibliography

- [1] P. Duvivier, "Transfert de puissance sans contact par transformateur tournant pour application spatiale", *Master's thesis, Louvain-la-Neuve (Belgium)*, 1st June, 2015.
- [2] Y. Combet, CNES, ESA, Thales Alenia Space. 2014. "Des cellules solaires à haut rendement issues du spatial". In *EDF Pulse : Le fil de l'innovation*. Online. <http://pulse.edf.com/fr/des-cellules-solaires-a-haut-rendement-issues-du-spatial>. Consulted : 20 March 2016.
- [3] ESA. 2014. "Introduction to space environment". In *European Space Agency : Space Environment*. Online. http://www.esa.int/Our_Activities/Space_Engineering_Technology/Space_Environment/Introduction_to_space_environment. Consulted : 20 March 2016.
- [4] D. Lincot, Institut de Recherche et Développement de l'Energie Photovoltaïque (IRDEP). 2007. "Nanostructures et conversion photovoltaïque". Online . 50 p. <http://sfp.in2p3.fr/expo/Nanosciences/2007/LINCOT.pdf>. Consulted : 30th November 2015.
- [5] G. Lindbergh, "Energy storage for e-mobility – batteries and supercapacitors", *Course : EJ2410 - "Hybrid Vehicle Drives", KTH, Sweden*, 2015.
- [6] Fabricast, "Slip ring mounting methods". Online. http://www.fabricast.com/wp-content/uploads/Slip_Ring_Mounting_Methods-2.gif. Consulted : 20 March 2016.
- [7] RUAG Space, "Septa 24, Solar Array Drive Assembly", *Product Datasheet, Zurich*, 2015.
- [8] C. Zhao, S. D. Round, W. Kolar, "An Isolated Three-Port Bidirectional DC-DC Converter With Decoupled Power Flow Management", *IEEE Transactions on Power Electronics*, vol. 23, no. 5, September, pp. 2443-2453, 2008.
- [9] J. Dai, D. C. Ludois, "A Survey of Wireless Power Transfer and a Critical Comparison of Inductive and Capacitive Coupling for Small Gap Applications", *IEEE Transactions on Power Electronics*, vol. 30, no. 11, November, pp 6017-6029, 2015.
- [10] J. Dai, D. C. Ludois, "Single Active Switch Power Electronics for Kilowatt Scale Capacitive Power Transfer", *IEEE Journal of Emerging and Selected Topics in Power Electronics*, vol. 3, no. 1, March, pp 315-323, 2015.
- [11] QuadTech, Inc., "Equivalent Series Resistance (ESR) of Capacitors", *Application note, September*, 2007.
- [12] K. Papastergiou, D. Macpherson, "Contact-less Transfer of Energy by means of a Rotating Transformer", *IEEE ISIE 2005, 20-23 June, Dubrovnik, Croatia*, pp.1735-1740, 2005.

- [13] J.P.C. Smeets, L. Encica, E.A. Lomonova, "Comparison of Winding Topologies in a Pot Core Rotating Transformer", *International Conference on Optimization of Electrical and Electronic Equipment (OPTIM)*, 12th May, pp.103-110, 2010.
- [14] J. P. C. Smeets, D. C. J. Krop, J. W. Jansen, M. A. M. Hendrix and E. A. Lomonova, "Optimal design of a pot core rotating transformer," *2010 IEEE Energy Conversion Congress and Exposition, Atlanta, GA*, pp. 4390-4397, 2010.
- [15] R. Ottersten, "Component Systems - Power Electronics", *Hybrid Vehicle Handbook, Göteborg, February*, 2004.
- [16] M. Bekemans, "Composants et technologies", *Course : LELEC2660 - "Power Electronics", UCL, Belgium*, 2014.
- [17] C. Mi, H. Bai, C. Wang, S. Gargies, "Operation, design and control of dual H-bridge-based isolated bidirectional DC-DC converter", *IET Power Electron.*, vol. 1, no. 4, pp. 507-517, 2008.
- [18] A. Rodríguez Alonso, J. Sebastian, D. G. Lamar, M. M. Hernando and A. Vazquez, "An overall study of a Dual Active Bridge for bidirectional DC/DC conversion," *2010 IEEE Energy Conversion Congress and Exposition, Atlanta, GA*, pp. 1129-1135, 2010.
- [19] B. Dehez, "Les transformateurs monophasés", *Course : LELEC1310 - "Convertisseurs électromécaniques", UCL, Belgium*, 2014.
- [20] R. Erickson, D. Maksimovic, "Fundamentals of Power Electronics", *second edition, Colorado*, 2004.
- [21] A. Van den Bossche, V. Cekov Valchev, "Inductors and Transformers for Power Electronics", *Ghent University, Belgium*, 2005.
- [22] Texas Instruments, "Magnetic Core Characteristics", *Texas Instrument Corporated, Texas*, 2001.
- [23] J. Mühlethaler, J. Biela, J. Kolar, A. Ecklebe, "Improved Core-Loss for Magnetic Components Employed in Power Electronic Systems", *IEEE transactions on power electronics*, vol. 27, no. 2, February, pp. 964-973, 2012.
- [24] C. Wm. T. McLyman "Transformer and Inductor Design Handbook", *Handbook, third edition, California*, 2004.
- [25] J. Mühlethaler, J. Kolar, "Optimal Design of Inductive Components Based on Accurate Loss and Thermal Models", *IEEE Applied Power Electronics Conference, Zurich*, 2012.
- [26] Ferroxcube, "Soft Ferrites and Accessories", *Data Handbook*, 2013.
- [27] K. Iyer, W. Robbins, N. Mohan, "Design and Comparison of High Frequency Transformers using Foil and Round Windings", *IEEE International Power Electronics Conference*, pp. 3037-3043, 2014.
- [28] J. Lindström, "Development of an experimental permanent-magnet motor drive", *Licentiate thesis, Chalmers University of Technology*, 1999. (link to scanned version of Paper D "Thermal model of a permanent magnet motor for a Hybrid electric vehicle")
- [29] Eric Granryd, "Heat transfer : Collection of formulas and Tables of Thermal Properties", *Course Compendium, Stockholm*, 2009
- [30] Ferroxcube, "Soft ferrites : Ferrite materials survey", *Product Datasheet*, 2008.
- [31] Loctite, "Thermal management materials", *Product Datasheet, Henkel Electronics Materials*, 2014.
- [32] W.A. Roshen, "High-Frequency Fringing Fields Loss in Thick Rectangular and Round Wire Windings", *IEEE transactions on magnetics*, vol. 44, no. 10, October, pp.2396-2401, 2008.

- [33] C. Zhao, S. D. Round, J. W. Kolar, "An Isolated Three-Port Bidirectional DC-DC Converter With Decoupled Power Flow Management", *IEEE transactions on power electronics*, vol. 23, no. 5, September, pp.2443-2452, 2008.
- [34] Wikipedia. 2016. "Water jet cutter", In *Wikipedia - The free encyclopedia : Water jet cutter*, Online : http://en.wikipedia.org/wiki/Water_jet_cutter . Consulted : 30 January 2016.
- [35] Wikipedia. 2016. "Laser femtoseconde". In *Wikipedia - The free encyclopedia : Laser femtoseconde* Online : https://fr.wikipedia.org/wiki/Laser_femtoseconde . Consulted : 30 January 2016.