METAMATERIALS APPLIED TO INDUCTIVE WIRELESS POWER TRANSMISSION

Jorge Virgilio de Almeida

Projeto de Graduação

**METAMATERIALS APPLIED TO INDUCTIVE
WIRELESS POWER TRANSMISSION**

Aluno(s): Jorge Virgilio de Almeida

Orientador(es): Marco Antonio Grivet Mattoso Maia

Marbey Manhães Mosso

Trabalho apresentado com requisito parcial à conclusão do curso de Engenharia Elétrica na Pontifícia Universidade Católica do Rio de Janeiro, Rio de Janeiro, Brasil.

Acknowledgments

No man is an island... And that is why I dedicate this work to those who direct or indirectly contributed to make it possible:

I'd like to thank my supervisors in Laboratoire Ampère at École Centrale de Lyon, Prof. Christian Vollaire and Prof. Laurent Krähenbühl, for introducing me to the topic.

I'd like to thank my advisors, Prof. Marco Antonio Grivet Mattoso Maia and Prof. Marbey Manhães Mosso, for believing in this project and in me.

I'd like to thank Prof. Ana Maria Beltran Pavani, Prof. Carlos Frederico Borges Palmeira and Prof. Enio Frota da Silveira for their support.

I'd like to thank the members of the Fundo Emergencial de Solidariedade da PUC-Rio for making the road here less harsh.

I'd like to thank my college friends, from this side of the Atlantic and the other, for their fellowship and indispensable help.

I'd like to thank my fellow soldiers at Base Aérea dos Afonsos for the lessons on loyalty and civism.

I'd like to thank to my old teachers at CIEP GP 175 José Lins do Rego and Escola Estadual Antônio de Oliveira Salazar for their exceptional effort.

And finally I'd like to thank to my parents and family for their love and everything else.

"Todo o futuro da nossa espécie, todo o governo das sociedades, toda a prosperidade moral e material das nações dependem da ciência, como a vida do homem depende do ar. Ora, a ciência é toda observação, toda exatidão, toda verificação experimental. Perceber os fenômenos, discernir as relações, comparar as analogias e as dessemelhanças, classificar as realidades, e induzir as leis, eis a ciência; eis, portanto, o alvo que a educação deve ter em mira. Espertar na inteligência nascente as faculdades cujo concurso se requer nesses processos de descobrir e assimilar a verdade, é o a que devem tender os programas e os métodos de ensino."

(Rui Barbosa)

Parecer sobre a Reforma do Ensino Secundário e Superior

Abstract

Over recent years, the interest in using inductive wireless power transmission for many applications has grown. One of the major limitations of this technology is the reduced operating distance. Some recent works have suggested using artificial materials known as metamaterials to improve the power transfer efficiency over distance. Due to their unique electromagnetic properties, such as negative permeability, metamaterials can be used to enhance the evanescent waves of the near field. In the present work, the usage of an electromagnetic metamaterial to increase the inductive coupling by means of enhanced evanescent waves is studied. Analytical calculations and numerical simulations of the proposed metamaterial are presented. The improvement of the power transfer efficiency is supported by empirical evidences. This work is written integrally in the English language.

Keywords: wireless power transmission; metamaterial; superlens; inductive coupling; resonant coupling;

Metamateriais aplicados à transmissão de potência sem fio por acoplamento indutivo

Resumo

Nos últimos anos, tem aumentado o interesse no uso da transmissão de energia sem fio por acoplamento indutivo em muitas aplicações. Uma das maiores limitações dessa tecnologia é a distância de operação reduzida. Alguns trabalhos recentes sugerem usar materiais artificiais conhecidos como metamateriais para aprimorar a eficiência da transferência de potência ao longo da distância. Devido às suas propriedades eletromagnéticas únicas, tais como permeabilidade magnética negativa, metamateriais podem ser usados para amplificar as ondas evanescentes do campo próximo. No presente trabalho, é estudado o uso de metamateriais eletromagnéticos para aumentar o acoplamento indutivo por meio da amplificação do campo próximo. São apresentados cálculos analíticos e simulações dos metamateriais propostos. O melhoramento da eficiência na transferência de potência é suportado por evidências experimentais. Este trabalho está redigido integralmente na língua inglesa.

Palavras-chave: transmissão de potência sem fio; metamaterial; superlente; acoplamento indutivo; acoplamento ressonante;

List of figures

Figure 1: Wardencliff Tower: Nikola Tesla's vision of a global wireless power transmission system	2
Figure 2: WPT is considered promising for recharging electrical vehicles [1]	2
Figure 3: Kiva robots transport goods at an Amazon Fulfillment Center	3
Figure 4 : lamp is lit on wirelessly over a two-meter distance [3]	4
Figure 5: EM field zones as a function of the wavelength	5
Figure 6: Schematic of the electromagnetic near-field and far-field zones	5
Figure 7: Wave impedance as a function of the normalized distance from the source	6
Figure 8: LaserMotive's logo	7
Figure 9 : Schematic of an electrical transformer	8
Figure 10 : Duracell Powermat®	9
Figure 11 : Panasonic ChargePad®	9
Figure 12 : Energizer Inductive Charger®	9
Figure 13 : Operational principle of a Powerpad	10
Figure 14 : isolated resonator modelled as a RLC circuit	10
Figure 15: Demonstration of resonant WPT at Intel Developer Forum in San Francisco	10
Figure 16: High-Q resonant coupling as a mean to compensate the decaying of the coupling coefficient ..	11
Figure 17: Strongly coupled resonances are quite robust to misalignment	11
Figure 18: equivalent circuit for the coupled resonator system	12
Figure 19: Frequency splitting phenomenon: PVS , PS and PL are respectively the output power of the AC voltage source, the dissipation power of the source and the load resistances [13]	13
Figure 20: Evolution in the number of publications related with metamaterials	14
Figure 21: Material classification, from [18], Copyright © 2006 by the IEEE, Inc.	15
Figure 22: MTM wedge with SRR and a lattice of wires deposited on opposite sides	15
Figure 23: MTM wedge used in the experiment (right) and the plot of the detected power as a function of the angle of refraction (left), from [16]	16
Figure 24: Group velocity (left) and phase velocity (right)	16
Figure 25: Perfect lensing: focalization of the far field (A) and the near field (B) [16]	17
Figure 26: Attenuating solutions are cancelled out inside the MTM slab	17
Figure 27: Experimental results, from [20], copyright © 2004 by the American Physical Society	18
Figure 28: Image currents for a PEM (left) and for a PMC (right)	18
Figure 29 : Asymmetric Split-Ring Resonators, from [21]	19
Figure 30: Ref [21] results. Solid line (experiment), filled circles (theory, method of moments), empty circles (theory for reference structure with symmetrically split rings)	20
Figure 31: Current distribution in structure A, from [21]	20
Figure 32: Q-factor as a function of the coefficient of asymmetry γ [21]	21
Figure 33: MTM inclusions based on grid-graph Hamiltonian paths	22
Figure 34: Multiple Split-Ring Resonators ($N=4$)	22
Figure 35 : Quasi-static equivalent circuit of the MSRR	23
Figure 36: Resonant frequency of the MSRR as a function of the number N	24
Figure 37: Square Spiral Resonator with $N=12$	24

Figure 38: Quasistatic equivalent circuit of a SR.....	25
Figure 39: Resonant frequency of SRs as a function of the number of turns N	25
Figure 40: Comparison between the Q-factor of MSRR and SR systems.....	26
Figure 41: MTM with lumped capacitors added to the structure	26
Figure 42: Square SR in series with a lumped capacitor via holes.....	28
Figure 43: Representation of the first prototype.....	29
Figure 44: Image of the Impedance Analyzer	30
Figure 45: Vector Network Analyzer	31
Figure 46: Experimental setup to measure the current in the transmitting coil	31
Figure 47: Experimental setup to measure the S-parameters of the coils	32
Figure 48: Experimental setup to measure the voltage of the secondary	32
Figure 49: Experimental setup	33
Figure 50: Current in the transmitter as a function of the frequency	34
Figure 51: Current in the receptor as a function of the frequency.....	35
Figure 52: Voltage in the receiving coil with and without the MTM slab.....	36
Figure 53: 3D view of the IWPT system with no MTM slab	37
Figure 54: Magnitude of the current in the primary (right) and secondary (left)	38
Figure 55: Magnitude of the voltage in the primary (right) and secondary (left)	38
Figure 56: Impedance in the primary: real part (right) and imaginary (left)	38
Figure 57: Real power (right) and reactive power (left) in the primary	39
Figure 58: Power transfer efficiency (left) and power delivered to the load (right)	39
Figure 59: S-parameters of the system in decibels: S_{11} (left) and S_{21} (right)	39
Figure 60: 3D view of the IWPT system with one MTM slab positioned at 75mm from the transmitter	40
Figure 61: Magnitude of the current in the primary (right) and secondary (left)	40
Figure 62: Magnitude of the voltage in the primary (right) and secondary (left)	41
Figure 63: Impedance in the primary: real part (right) and imaginary (left)	41
Figure 64: Real power (right) and reactive power (left) in the primary	41
Figure 65: Power transfer efficiency (left) and power delivered to the load (right)	42
Figure 66: S-parameters of the system in decibels: S_{11} (left) and S_{21} (right)	42
Figure 67: 3D view of the IWPT system with one MTM slab positioned at 20mm from the transmitter	42
Figure 68: Magnitude of the current in the primary (right) and secondary (left)	43
Figure 69: Magnitude of the voltage in the primary (right) and secondary (left)	43
Figure 70: Impedance in the primary: real part (right) and imaginary (left)	43
Figure 71: Real power (right) and reactive power (left) in the primary	44
Figure 72: Power transfer efficiency (left) and power delivered to the load (right)	44
Figure 73: S-parameters of the system in decibels: S_{11} (left) and S_{21} (right)	44
Figure 74: 3D view of the IWPT system with one MTM slab positioned at 1mm from the transmitter	45
Figure 75: Magnitude of the current in the primary (right) and secondary (left)	45
Figure 76: Magnitude of the voltage in the primary (right) and secondary (left)	45
Figure 77: Impedance in the primary: real part (right) and imaginary (left)	46
Figure 78: Real power (right) and reactive power (left) in the primary	46
Figure 79: Power transfer efficiency (left) and power delivered to the load (right)	46
Figure 80: S-parameters of the system in decibels: S_{11} (left) and S_{21} (right)	47

Figure 81: 3D view of the IWPT system with one MTM slab positioned at 1mm from the receiver	47
Figure 82: Magnitude of the current in the primary (right) and secondary (left)	47
Figure 83: Magnitude of the voltage in the primary (right) and secondary (left)	48
Figure 84: Impedance in the primary: real part (right) and imaginary (left)	48
Figure 85: Real power (right) and reactive power (left) in the primary	48
Figure 86: Power transfer efficiency (left) and power delivered to the load (right)	49
Figure 87: S-parameters of the system in decibels: S_{11} (left) and S_{21} (right)	49
Figure 88: 3D view of the IWPT system with two MTM slabs	49
Figure 89: Magnitude of the current in the primary (right) and secondary (left)	50
Figure 90: Magnitude of the voltage in the primary (right) and secondary (left)	50
Figure 91: Impedance in the primary: real part (right) and imaginary (left)	50
Figure 92: Real power (right) and reactive power (left) in the primary	51
Figure 93: Power transfer efficiency (left) and power delivered to the load (right)	51
Figure 94: S-parameters of the system in decibels: S_{11} (left) and S_{21} (right)	51
Figure 95: 3D view of the IWPT system with two MTM slabs positioned at 1mm from each coil	52
Figure 96: Magnitude of the current in the primary (right) and secondary (left)	52
Figure 97: Magnitude of the voltage in the primary (right) and secondary (left)	52
Figure 98: Impedance in the primary: real part (right) and imaginary (left)	53
Figure 99: Real power (right) and reactive power (left) in the primary	53
Figure 100: Power transfer efficiency (left) and power delivered to the load (right)	53
Figure 101: S-parameters of the system in decibels: S_{11} (left) and S_{21} (right)	54
Figure 102: 3D view of the IWPT system with two MTM slabs with 20cm of misalignment	54
Figure 103: Magnitude of the current in the primary (right) and secondary (left)	54
Figure 104: Magnitude of the voltage in the primary (right) and secondary (left)	55
Figure 105: Impedance in the primary: real part (right) and imaginary (left)	55
Figure 106: Real power (right) and reactive power (left) in the primary	55
Figure 107: Power transfer efficiency (left) and power delivered to the load (right)	56
Figure 108: S-parameters of the system in decibels: S_{11} (left) and S_{21} (right)	56
Figure 109: Corresponding magnetic field distributions along the z-axis: with no MTM (left) and with the MTM excited at the resonant frequency (right). See Ref [25]	57

List of tables

Table 1: Theoretical and experimental parameters of the first prototype	33
Table 2: Voltage in the secondary with and without the MTM slab.....	35
Table 3: Comparison between the IWPT system assisted and not assisted by MTMs (in the best configuration)	58

Acronyms

ADS	Advanced Design System
ASRR	Asymmetric Split Ring Resonator
DNG	Double Negative
DPS	Double Positive
EM	Electromagnetic
ENG	Epsilon Negative
IA	Impedance Analyzer
IWPT	Inductive Wireless Power Transmission
LHM	Left-Handed Material
MNG	Mu Negative
MSRR	Multiple Split Ring Resonator
MTM	Metamaterial
NRI	Negative Refractive Index
PCB	Printed Circuit Board
PDL	Power Delivered to the Load
PTE	Power Transfer Efficiency
SNG	Single Negative
SR	Spiral Resonator
SRR	Split Ring Resonator
VNA	Vector Network Analyzer
WPC	Wireless Power Consortium
WPT	Wireless Power Transmission

Contents

1. Introduction	2
1.1. Historic of the research.....	2
1.2. Far field and near field regions	4
1.3. Wireless power transmission: state-of-the-arts	6
1.3.1. Radiative waves	6
1.3.2. Inductive coupling.....	7
1.3.3. Resonant magnetic coupling.....	10
2. Theory of metamaterials.....	14
2.1. Artificial complex media	14
2.2. Perfect Magnetic Conductors	18
2.3. Inclusions with broken symmetry.....	19
3. Design of μ -Negative Metamaterials	22
3.1. Introduction	22
3.2. Multiple Split-Ring Resonator Model.....	22
3.3. Spiral Resonator Model	24
3.4. Comparison of MSRR and SR Models	25
3.5. Control of the resonant frequency	26
4. Design and fabrication	28
5. Experimental protocol	30
5.1. The Impedance Analyzer.....	30
5.1.1. Details of the measurement	30
5.2. Vector Network Analyzer	30
5.2.1. Details of the measurement	31
6. Experimental results and discussion.....	33
6.1. Verification of the electromagnetic characteristics of the unit cell	33
6.2. Verification of the current in the transmitting coil.....	34
6.3. Verification of the current in the receiving coil	35
6.4. Verification of the enhancement of the PTE and the PDL	35
7. Numerical simulation.....	37

7.1.	Agilent Advanced Design System (ADS).....	37
7.2.	Configuration 1: with no MTM slab	37
7.3.	Configuration 2: with a MTM slab positioned at 75mm from the transmitter	40
7.4.	Configuration 3: with a MTM slab positioned at 20mm from the transmitter	42
7.5.	Configuration 4: with a MTM slab positioned at 1mm from the transmitter	45
7.6.	Configuration 5: with a MTM slab positioned at 1mm from the receiver.....	47
7.7.	Configuration 6: with two MTM slabs one positioned at 20mm from the transmitter and one positioned at 1mm from the receiver.....	49
7.8.	Configuration 7: with two MTM slabs one positioned at 1mm from the transmitter and one positioned at 1mm from the receiver.....	52
7.9.	Configuration 8: with two MTM slabs one positioned at 1mm from the transmitter and one positioned at 1mm from the receiver with 20cm of misalignment.....	54
7.10.	Results and discussion	56
8.	Conclusions	59
8.1.	Global analysis	59
8.2.	Future work.....	59
9.	Bibliography	60

1. Introduction

1.1. Historic of the research

Figure 1: Wardenclyffe Tower: Nikola Tesla's vision of a global wireless power transmission system

The dream of wireless power transmission (WPT) dates back to Tesla. Almost a century after his first attempts to create a unified global wireless system for telecommunications and power transmission, the subject is reemerging due to the increasing demand on autonomy. Most of the daily life gadgets (like smartphones, tablets or laptops) and the robots employed in industry (like the Kiva robots at Amazon's warehouses) are cordless. Nonetheless, their autonomy is limited by their dependence on batteries. Battery technology is not only an expensive and environmentally unfriendly technology, it is also highly time-consuming. For that reason, more efficient WPT systems are quite promising in every level: from implantable medical devices to electric vehicles.

Figure 2: WPT is considered promising for recharging electrical vehicles [1]

According to the Forbes' list [2], the six more prominent companies working on the development of WPT systems are: Powermat, PowerByProxi, Wireless Power Consortium (WPC), Mojo Mobility, WiPower and Mopar.

Figure 3: Kiva robots transport goods at an Amazon Fulfillment Center

WPC is an open-membership of more than 150 companies from Asia, Europe and North America, including big mobile manufactures like Samsung, who is working since 2008 toward a global standardization of wireless charging technologies. Since 2009, companies who are members of WPC adopt Qi (the Chinese word for "energy") as their low-power inductive wireless power transmission standard.

Although the electromagnetic field is much more efficient transmitting power over large distance in its radiative form, the human exposure to electromagnetic waves carrying power on the order of tens of watts is tremendously risk. So, the energy transfer by non-radiative modes is considered much more suitable concerning health and safety. Once the magnetic field is supposed to interact very weakly with biological beings, the majority of the WPT systems exploit the magnetic near field instead of the electric one.

One of the main limitations of inductive wireless power transmission (IWPT) is its short operating distance (on the order of the coil diameter). In 2007, a group of MIT researchers demonstrated that by achieving strong magnetic coupling efficient WPT over midrange (a few times the coil diameter) was possible. This technique, named Witricity by its creators, exploits the high-Q resonant coupling of a 4-coil topology as mean to compensate the decay of the coupling coefficient with the increasing distance between source and load drivers. It can be demonstrated that power transfer efficiency (PTE) can be made arbitrarily close to 100% by increasing the Q-factor of the receiver and transmitter coils, which also makes those systems quite robust to misalignment.

Figure 4 : lamp is lit on wirelessly over a two-meter distance [3]

Besides Witricity virtues, it is too susceptible to splitting frequency, especially in a multi-receiver scenario, which imposes a great limitation to the power delivered to the load (PDL). Another problem considering this technique is the energy waste due to leakage flux of a non-directional magnetic field. If the magnetic field could be made more directional, PTE could be highly increased without the need of High-Q resonant coupling between the coils. Some solutions, like strategically designing the coils or, more elegantly, shaping the magnetic field by phase controlling the current loops, have been proposed [4].

However, recent works indicate that metamaterials (MTM) are potentially the key technology to give designers total control of the electromagnetic field [5] [6]. Being artificial media whose electromagnetic parameters can be fully engineered, magnetic field shaping is not only possible but also dynamically controllable with MTM.

In practice, concerning this particular case of IWPT, MTM are used to create a region of paramagnetic response in space where PTE can be highly increased due to the enhancement of the magnetic potential, and by consequence, the coupling coefficient between coils.

The present work is based on that approach.

1.2. Far field and near field regions

The space surrounding an antenna is usually subdivided into three regions: the reactive near-field region, the radiating near-field region (also known as Fresnel region) and the far-field region. However, for an electric small antenna, these are only separated two regions: the quasistatic field and the radiating field.

Figure 5: EM field zones as a function of the wavelength

The boundary of the two regions is defined as:

$$r = \frac{\lambda}{2\pi} \quad (1)$$

Figure 6: Schematic of the electromagnetic near-field and far-field zones

In the quasistatic field, one of the oscillating fields dominates the near-field region and the energy remains stored either as magnetic potential or as electric potential. Hence, almost no energy in the quasistatic field is lost by the source since EM waves keep coming and going until a receiver gets close enough to make power flows irreversibly from one circuit to another by means of near-field coupling. Whilst in the radiating region, the real part of the Poynting vector dominates and the energy propagates far from the source as EM radiation due to the magnetic and electric field circulations interdependency (the fields are proportional to each other by a ratio named wave impedance).

The wave impedance is defined as the ratio of the transverse components of the magnetic and the electric field (the components of the EM field that make a right angle with the direction of propagation):

$$|Z_W| \triangleq \frac{E}{H} \quad (2)$$

It measures the relative strength of electric and magnetic fields as a function of the source structure. In the near-field region, for an electric field source (e.g., a dipole antenna) wave impedance is typically very high (E-field much greater than H-field) whereas for a magnetic field source (e.g., plane coil) it is very low (E-field much smaller than H-field). In both cases, in the far-field region, the wave impedance is determined by the propagating medium. In this particular example, the sources are supposed to be in free space whose characteristic impedance is 377 Ohms.

Figure 7: Wave impedance as a function of the normalized distance from the source

1.3. Wireless power transmission: state-of-the-arts

Each application has its specific requirements for transfer distance, device size, delivered power and packaging that determine the best suitable design. However, most of the technologies available in the market today rely on the following fundamental techniques: radiative waves, inductive coupling or resonant magnetic coupling. Those technologies can be categorized as far-field based (radiative waves) or near-field based (inductive coupling, capacitive coupling, resonant coupling) WPT systems.

1.3.1. Radiative waves

Earlier researches of wireless power transmission systems have proposed using radiative waves. Radiative wireless power transmission uses directed beams in the far field of omnidirectional antennas to send power from a transmitter to a receiver, analogously to the telecommunication systems.

A Seattle-based company *LaserMotive*, for example, is developing wireless power systems using laser beams to transmit electricity without wires. The Rayleigh criterion dictates that any electromagnetic wave will spread and become weaker and diffuse over distance. So the tighter the beam is the less it will spread as a function of distance. Very small laser aperture in comparison with its wavelength gives laser beams high directivity gain, permitting them to cross a very long distance without too much attenuation. The maximum directivity gain is limited by the diffraction limit.

Figure 8: LaserMotive's logo

Although the directivity gain for lasers is much greater than for microwaves, due to the atmospheric effects and the interference generated by obstacles like buildings, the using of microwaves instead of lasers have been considered more promising [7].

Even if it is very efficient to send high power over large distances (in the order of kilometers), it is usually avoid for consumer electronics devices charging, especially indoor applications, for many reasons. First of all, radiation processes are irreversible: once the transmitter radiates power it will never return to the source whether or not there is a receiver available to make use of it (this is referred to in the literature as the time arrow of radiation). Secondly, to achieve high power transfer efficiency it requires highly directive antennas with a clear line-of-sight, which implies complex tracking mechanisms if multiple receivers are to be considered. Finally, as substantial power would be carried by the electromagnetic waves and knowing that electromagnetic radiation interacts strongly with humans and electronic devices, it would be more dangerous, cause more interference and waste much more energy than non-radiative technologies.

However, it remains a good alternative for sending energy to devices separated from the source by a large distance in environments of difficult access or inhospitable facilities like power solar satellites to feed space stations and vessels or sending highly directed power beams to the planetary surface in order to enhance the efficiency of solar power plants. Or still in replacement of transmission lines in places where it is impractical and uneconomical or to recharge the sensors installed along the lines.

That last argument is very important in countries such as Brazil with a vast territory where geography nor always favors the installation of transmission lines and its monitoring. The present work is based only on non-radiative WPT techniques. Nonetheless, in regard of the potential of this technology to facilitate and democratize the access to electric energy in the world, more researches on radiative power transmission are strongly recommended.

1.3.2. Inductive coupling

Considering the risks of transmitting power by means of radiation, the majority of the WPT technologies being commercialized are based on non-radiative inductive coupling. The most well-known IWPT system is the electrical transformer.

Figure 9 : Schematic of an electrical transformer

According to the transformer theory, the voltage and current characteristics of the transmitting and the receiving coils in an ideal IWPT are described by the following equations:

$$\begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = j\omega \cdot \begin{bmatrix} L_1 & M_{12} \\ M_{21} & L_2 \end{bmatrix} \cdot \begin{bmatrix} I_1 \\ I_2 \end{bmatrix} \quad (3)$$

$$k = \frac{M_{12}}{\sqrt{L_1 L_2}} \quad (4)$$

Where V_1 is the voltage at the transmitting coil, I_1 is the current at the transmitting coil, V_2 is the voltage at the receiving coil, I_2 is the current at the receiving coil, L_1 is self-inductance of the transmitting coil, L_2 is self-inductance of the receiving coil, M_{12} and M_{21} are the mutual inductance of the coils, and k is the coupling coefficient between the coils.

By circuit analysis, the power in the primary and the secondary are found to be equal to:

$$P_s = \frac{V_s^2 Z_2}{Z_1 Z_2 + (\omega M_{12})^2} \quad (5)$$

$$P_L = \frac{V_s^2 (\omega M_{12})^2 R_L}{(Z_1 Z_2 + (\omega M_{12})^2)^2} \quad (6)$$

Hence, the PTE of the system is:

$$\eta = \frac{P_L}{P_s} = \frac{R_L (\omega M_{12})^2}{Z_2 (Z_1 Z_2 + (\omega M_{12})^2)} \quad (7)$$

From those equations, it can be deduced that the best efficiency will be reached out when the primary and the secondary are both in resonance. Another implication is that efficiency grows with frequency. But it is in contrary with antenna theory that states that in the longer distance, the higher frequency would be weak. So, as it was stated before, there is a compromise between frequency and distance. Since the reactive near-field zone diminishes with frequency.

In the last years, numerous solutions for charging electronic devices based on conventional inductive coupling power transmission have been proposed.

Figure 10 : Duracell Powermat®

Figure 11 : Panasonic ChargePad®

Figure 12 : Energizer Inductive Charger®

The greatest advantage of this solution is that its circuitry is cheap and very simple to implement. However, the need of a contacting surface between the device and the charging pad due to the high k restriction makes the entire support infrastructure required too expensive and limiting in terms of mobility.

Figure 13 : Operational principle of a Powerpad

An additional problem is that the performance of such systems is easily disturbed by the magnetic interference of near objects, which generates usually an extra cost on shielding.

1.3.3. Resonant magnetic coupling

Resonance is a natural phenomenon that occurs when the energy of the system is oscillating between different storage modes. In this particular case, the energy of the system oscillates between the inductor (when it is stored as a magnetic field) and the capacitor (when it is stored as an electric field). When the system is at resonance, it is possible to build-up stored energy, even with the slightest excitation. This accumulation effect occurs when the rate of energy injection into the system is greater than the rate of energy loss by the system.

Figure 14 : isolated resonator modelled as a RLC circuit

Resonant coupling based WPT has been long reported [8]. Improvement in power transfer efficiency was already noted by adding a shunt or a series capacitance to the secondary coil and tuning the system to its resonance frequency [9].

Figure 15: Demonstration of resonant WPT at Intel Developer Forum in San Francisco

Nevertheless, its application was very limited until 2007 when MIT researchers demonstrated that WPT based on strongly coupled resonances, this is to say a 4-coils based topology where the two extra coils, called the added resonators, present very high Q-factors, could transfer energy efficiently even at mid-range: 60W power through a distance as large as 2 meters with 40% efficiency [3]. This experiment was later replicated by Intel during the Intel Developer Forum and, by replacing the solenoids coils for planar coils, 75% efficiency was attained [10].

Figure 16: High-Q resonant coupling as a mean to compensate the decaying of the coupling coefficient

Since then, many other researchers around the world struggle to analyze, improve and develop technologies based on resonant coupling [11] [12]. One of the most important features of these resonant systems is that it can operate efficiently even if k is too small (<0.2) thanks to a second and independent parameter named the quality factor or Q -factor. It makes possible for traditional wireless power transmission systems based on induction to work even better, but more importantly it makes possible to conceive systems with great mobility by eliminating the need of precise alignment. Another advantage is that strongly coupled resonances interact weakly with other off-resonant environmental objects minimizing the problems of magnetic induction, making this technique much safer to users.

Figure 17: Strongly coupled resonances are quite robust to misalignment

According to the experiments described in Ref [3], if one of the circuitry is detuned from resonance, even slightly, the power transmitted to the load drops sharply. Thus, both the transmitting and receiving circuitries must be tuned on resonance.

For an isolated resonator, the rate of energy injection is determined by its resonant frequency ω_0 .

$$\omega_0 = \frac{1}{\sqrt{LC}} \quad (8)$$

$$\Gamma = \frac{R}{2L} \quad (9)$$

$$Q = \frac{\omega_0}{2\Gamma} \quad (10)$$

Where Γ is the intrinsic rate of energy loss, and Q , the quality factor. Figure 18 shows an equivalent circuit of coupled resonators, where V_s is a sinusoidal voltage source with internal resistance R_i , L_1 , C_1 and R_1 are the inductance, the capacitance and the resistance of the source resonator, L_2 , C_2 and R_2 are the inductance, the capacitance and the resistance of the receiver resonator, R_L is the load and M their mutual inductance.

Figure 18: equivalent circuit for the coupled resonator system

By circuit analysis, the PTE for this system can be written as:

$$\eta = \frac{P_L}{P_s} = \frac{4k^2 Q_1 Q_2}{\left(\left(1 + \frac{R_i}{R_1} \right) \left(1 + \frac{R_L}{R_2} \right) + k^2 Q_1 Q_2 \right)^2} \cdot \frac{R_i R_L}{R_1 R_2} \quad (11)$$

Equation (11) makes explicit the contribution of both Q to compensate a low k and in so enhancing the efficiency of the system.

Consequently, the figure of merit of most designers is:

$$FOM = k\sqrt{Q_1 Q_2} \quad (12)$$

$$\eta_{opt} = \frac{FOM^2}{(1 + \sqrt{1 + FOM^2})^2} \quad (13)$$

Concerning Q -factor, it's important to mention that the Q -factor in a parallel RLC circuit is the inverse of the Q -factor in a series RLC circuit:

$$Q_{series} = \frac{\omega_0 L}{R} = \frac{1}{R} \sqrt{\frac{L}{C}} \quad (14)$$

$$Q_{parallel} = \frac{R}{\omega_0 C} = R \sqrt{\frac{C}{L}} \quad (15)$$

So, for self-resonant coils (depending on stray capacitance only) and coils embedded with lumped capacitors, the maximization criteria of the Q-factor are quite different. Once transmitter and receiver coils have usually different sizes, then coils embedded with lumped capacitors are much more common in practice. However, High-Q is easily achieved with self-resonant coils. That is why the Witricity products operate usually in the MHz range.

Nonetheless, the main problem with strong coupled resonances is their limitation in power due to splitting frequency phenomenon. As shown in the figure below, although both resonant circuits are tuned to the same frequency and impedance-matched, two different resonant frequencies will be produced when they are magnetically coupled together. This arises because the energy exchange between the two resonant circuits modulates the oscillations at the primary and secondary windings. The envelopes produced have the same shape as the beat envelope produced when two dissimilar frequencies interfere constructively and destructively. The modulation envelope is sinusoidal in shape. This can be seen in the analogous case of two masses connected by a spring. In this classic physics example, the two masses form a single system which can oscillate in two modes: one of higher frequency (even mode) and one of lower frequency (odd mode) than the fundamental frequency of an individual pendulum.

Figure 19: Frequency splitting phenomenon: P_{Vs} , P_S and P_L are respectively the output power of the AC voltage source, the dissipation power of the source and the load resistances [13]

2. Theory of metamaterials

2.1. Artificial complex media

In 1968, Veselago presented a study on the plane-wave propagation in a hypothetical material whose permittivity and permeability were simultaneously negative. He first predicted its exotic properties such as negative refractive index due to the fact that in such medium the phase vector would be antiparallel to the Poynting vector [12]. As no natural material with negative response was found, the study remained only as a theoretical curiosity until 1999, when it was shown that an artificial complex media based on periodic structures, later known as metamaterials (materials whose properties are beyond the properties of the individual constituents), may present a frequency band over which ϵ and μ are both negative [14].

That promise of physically realizable artificial materials capable of mimicking the response functions of known materials or synthesizing new, exotic response functions that do not occur or may not be easily available in nature opened a whole new frontier, not only for electrical engineers, but also for mechanical and materials ones. For example, a nearly perfect sound absorber based on acoustic metamaterials was presented last year [15].

Figure 20: Evolution in the number of publications related with metamaterials

Electromagnetic metamaterials are usually periodic structures whose inclusion dimensions and inter-inclusion distances are much smaller than a wavelength and, as a consequence, they can be mathematically described as homogeneous media, with effective electromagnetic properties such as electric permittivity ϵ , magnetic permeability μ or chirality κ [16].

A huge variety of fascinating devices and applications relying in their design on various tools have been proposed. For the purpose of this work, the MTM of interest are those capable of almost perfect lensing, this is to say, capable of focusing both the propagating and the evanescent modes of the electromagnetic wave [17]. They are useful concerning inductive WPT systems because magnetic resonant coupling is essentially evanescent wave coupling, which imposes, as discussed previously, serious limitations to the distance of operation and to the density of the transferred power. The mutual inductive coupling between the coils can be enhanced due to the fact that the focusing of the evanescent modes implies the local amplification of the near field, increasing the magnetic flux in a specific direction as a consequence of the anisotropic response of the MTM.

To build a MTM behaving as a perfect lens, this is to say with $n < 0$, both $\varepsilon < 0$ and $\mu < 0$ are required. For double-negative materials (DNG), two sets of artificial structures are then needed, which can be complicated in both design and fabrication.

$$n = -\sqrt{|\varepsilon\mu|}, \text{ if } \mu < 0 \text{ and } \varepsilon < 0 \quad (16)$$

Figure 21: Material classification, from [18], Copyright © 2006 by the IEEE, Inc.

As show by Ref [19], for lattices based on split-ring or spiral resonators, μ assumes negative values around the resonant frequency of the cells. While for a lattice of wires, ε is negative below its cutoff frequency. Then, by choosing the parameters of the lattice of wires such that the cutoff frequency was significantly above the SRR resonant frequency, a MTM can be conceived as a composite of two single-negative MTMs separated by a dielectric with negligible thickness presenting an overlapping region where $\varepsilon < 0$ and $\mu < 0$, then synthesizing a DNG material.

In Ref [16], the results of experiments (Figure 23) on refraction with a MTM (Figure 22) formed by SRRs and wires deposited on opposite sides (red curve) and a Teflon sample (blue curve) are presented. The position of the peak of the power detected as a function of the angle according to Snell's law (17) in the experiment indicates quite clearly that the refractive index of both materials have different signs.

$$\sin(\theta_1) = n \sin(\theta_2) \quad (17)$$

Figure 22: MTM wedge with SRR and a lattice of wires deposited on opposite sides

Figure 23: MTM wedge used in the experiment (right) and the plot of the detected power as a function of the angle of refraction (left), from [16]

This particular behavior of the MTM can be explained in terms of group velocity v_g , which characterizes the flow of energy, and phase velocity v_p , which characterizes the movement of the wave fronts. In conventional materials, the group and phase velocities are parallel. However, in negative refractive index media (NRI), the group and phase velocities have opposite directions implying that the sign of the refractive index n must be taken as negative [16].

Figure 24: Group velocity (left) and phase velocity (right)

As notice by Ref [16], assuming the reality of NRI, there is no fundamental reason why an image could not be created with arbitrarily high resolution, once the wavelength limitation of the diffraction limit is a result of the conventional imaging. A NRI lens exhibits a totally different focusing phenomenon of the propagating rays and amplifies the fine details included in the non-radiative evanescent electromagnetic waves of the near field.

Figure 25: Perfect lensing: focalization of the far field (A) and the near field (B) [16]

Since a NRI adds a negative phase shift to the propagating rays, the image is resolved firstly inside the material, with a 180° phase shift, and then outside the material with no phase shift.

Within the reactive near-field zone, multiple reflections result in both growing and attenuating evanescent waves. However, when there is a boundary discontinuity at the interface such as from one side $n > 0$ and from the other side $n < 0$, a resonant phenomenon occurs in which the attenuating solution is canceled out, thus resting only the growing wave solution (the waves in the same direction of the magnetic flux). According to the law of conservation of energy, energy cannot be destroyed. So, the power carried out by the attenuating solutions is transferred to the growing solutions, enhancing their amplitude. This effect contributes to diminish the total magnetic reluctance of the magnetic link between the transmitting and receiving coils by introducing a negative magnetic reluctance that increases the magnetic potential, analogously to the negative resistance in an electric circuit.

Figure 26: Attenuating solutions are cancelled out inside the MTM slab

$$\mathcal{R} = \frac{\ell}{\mu A} < 0, \text{ if } \mu < 0 \quad (18)$$

In Ref [20], an experiment with a Planar Left-Handed Transmission-Line has been used to demonstrate that the diffraction limit could be overcome. As shown in Figure 27, the image recovered at the probe level is much better than the image achieved with traditional focusing, although it is far from being a perfect representation of the input due to the losses in the system.

Figure 27: Experimental results, from [20], copyright © 2004 by the American Physical Society

However, those requirements may be simplified for IWPT systems. Once in such systems the near field is dominated by its magnetic component, only the permeability needs to be negative to achieve a similar effect of amplification of the near field. Moreover, by using a single negative MTM (SNG), the loss would be reduced thanks to a simplified structure [17], given that MTMs have an “activation cost”: the induced currents in their lattice. By reducing the quantity of conductor material, the losses by Joule’s effect are minimized.

2.2. Perfect Magnetic Conductors

A μ -negative MTM behaves as artificial magnetic conductor (AMC).

According to the Image Theory, the image current induced in a perfect magnetic conductor (PMC) has the same orientation than the original current flow. Hence, the image currents contribute to the enhancement of the magnetic near field distribution in the transversal direction of the magnetic conductor’s surface.

Figure 28: Image currents for a PEM (left) and for a PMC (right)

Besides that, PMC surfaces include no phase shift in the signal. Then, the splitting frequency phenomenon, so common in resonant coupling, is greatly reduced when using μ -negative MTMs to optimize inductive links as the one proposed in this work.

2.3. Inclusions with broken symmetry

The exotic behavior predicted for μ -negative MTMs is based on their resonant response. So high-quality resonances are essential to conceive very efficient MTMs.

The quality factor is also defined as (19), where $\Delta\omega$ is the angular half-power bandwidth. Therefore, a narrow resonant frequency is required in order to achieve a high value for the Q-factor. In other words, it means that the coupling between the resonant system and free space should be as weak as possible.

$$Q = \frac{\omega_0}{\Delta\omega} \quad (19)$$

Traditional MTMs as SRR periodic structures present usually small Q-factor values. It is due to the fact that their unit cells are strongly coupled to free space and cannot confine a large volume of electromagnetic field.

In Ref [21], it is shown that by introducing a certain degree of asymmetry in the shape of these unit cells high-Q resonances can be achieved.

The breaking of the structural symmetry of the system introduces new symmetric current modes, known as “trapped modes”, which when excited weaken the coupling to free space. As a result, the resonance bandwidth is narrowed.

Figure 29 : Asymmetric Split-Ring Resonators, from [21]

Figure 29 shows the two Asymmetric Split-Ring Resonators (ASRR) studied in Ref [21]. It is shown that spectral selectivity is sharper when the electrical field is perpendicular to the mirror line of the ASRRs, this is the x-polarization for structure A and y-polarization for structure B. Orthogonal polarizations present no remarkable spectral feature.

In the case of structure A, an antisymmetric current mode dominates the usual symmetric one (dipole mode). High-Q transmission resonance is achieved when the two halves of the unit cell are excited in antiphase with currents of approximately the same magnitude (Figure 31, configuration II). Like this, the electromagnetic fields generated by individual induced currents on the slab are reduced, minimizing the coupling to free space. As a consequence, the scattering of the electromagnetic field is minimized and therefore radiation losses.

Figure 30: Ref [21] results. Solid line (experiment), filled circles (theory, method of moments), empty circles (theory for reference structure with symmetrically split rings)

Figure 31: Current distribution in structure A, from [21]

Also, asymmetry can be seen as a mean to compensate the imperfections of the system. In a lossless substrate, with $\varepsilon'' = 0$ (ε'' corresponding to the imaginary part of the permittivity) and a perfect conductor metal pattern, Q-factor decreases monotonously with the coefficient of asymmetry γ . However, if the dielectric losses are taken in account, the curve of the Q-factor as a function of γ alters drastically. As shown in Figure 32, for a realistic MTM ($\varepsilon'' \neq 0$), there is an optimum value of γ that maximizes the Q-factor [21].

Figure 32: Q-factor as a function of the coefficient of asymmetry γ [21]

In Ref [22], a much more recent study supporting these conclusions is presented.

As losses are inevitable, asymmetry is an important design feature to be considered in order to achieve highly efficient MTM based systems.

3. Design of μ -Negative Metamaterials

3.1.Introduction

Some exotic shapes allowing the electrical dimensions of the resonant magnetic inclusions of the MTM slab to be hugely reduced have been proposed, such as inclusions based on grid-graph Hamiltonian paths [23].

Figure 33: MTM inclusions based on grid-graph Hamiltonian paths

However, the absence of simple and reliable models to study their performance makes traditional geometries much more suitable for any analytical consideration.

For this reason, analytical models for Multiple Split-Ring Resonators and Spiral Resonators only will be presented in the present work in order to conceive μ -negative metamaterials. Nonetheless, it would be desirable for future works to develop the subject. Particularly, additional investigations on homogenization of periodic structures with lumped elements embedded or more complex geometries, like fractals or asymmetric inclusions, are needed.

3.2.Multiple Split-Ring Resonator Model

A Multiple Split-Ring Resonator (MSRR) consists of inclusions of N concentric split-rings with splits placed on the opposite sides of two consecutive rings (Figure 34). So the Split-Ring Resonator (SRR) mentioned in the previous sections can be defined as a particular case of the MSRR for which $N=2$.

Figure 34: Multiple Split-Ring Resonators ($N=4$)

Considering the hypothesis that the dimensions of the split-rings are very small compared to the working wavelength, the quasi-static model presented in Ref [24] can be used to describe the electromagnetic behavior of MSRRs.

Figure 35 : Quasi-static equivalent circuit of the MSRR

In Figure 35, $C_{12}, C_{23}, \dots, C_{(N-1)N}$ represent the distributed capacitances of the rings, from the outer one to the most inner one, where the subscript reports the numbers associated to the two rings between which the distributed capacitance is calculated.

The derived expressions for total distributed inductance and capacitance are defined as:

$$L_{MSRR} = \frac{\mu_0 l_{avg}}{2} \frac{4.86}{4} \left[\ln \left(\frac{0.98}{\rho} \right) + 1.84\rho \right] \quad (20)$$

$$C_{MSRR} = \frac{N-1}{2} [2l - (2N-1)(w+s)] C_0 \quad (21)$$

Where w is the strip width, s the separation between two adjacent rings, l the side length of the outer ring, l_{avg} the average strip length calculated over all the N rings.

The fill ration ρ and the per-unit-length capacitance between two parallel strips C_0 are defined as:

$$C_0 = \epsilon_0 \frac{K(\sqrt{1-k^2})}{K(k)} \quad (22)$$

$$\rho = \frac{(N-1)(w+s)}{l - (N-1)(w+s)} \quad (23)$$

Where K is the complete elliptic integral of the first kind and k is:

$$k = \frac{\frac{s}{2}}{w + \frac{s}{2}} \quad (24)$$

Figure 36: Resonant frequency of the MSRR as a function of the number N

As shown in Figure 36, it is clear that after a certain value of N, the increasing in the number of rings will not contribute to diminish the resonant frequency. Further inclusions will only serve to finer the tune.

3.3.Spiral Resonator Model

A spiral resonator consists in a planar coil.

Figure 37: Square Spiral Resonator with N=12

The quasistatic equivalent model of the square spiral resonator shown in Figure 37 is given by [24] and presented in Figure 38.

Figure 38: Quasistatic equivalent circuit of a SR

Based on Ref [24], the approximations for total capacitance and inductance are:

$$C_{SR} = C_0 \frac{l}{4(w+s)} \frac{N^2}{N^2 + 1} \sum_{n=1}^{N-1} \left[l - \left(n + \frac{1}{2} \right) (w + s) \right] \quad (25)$$

$$L_{SR} = \frac{\mu_0}{2\pi} l_{avg}^{SR} \left[\frac{1}{2} + \ln \left(\frac{l_{avg}^{SR}}{2w} \right) \right] \quad (26)$$

Where, C_0 is also given by (22) and l_{avg}^{SR} , the single turn average length, is given by:

$$l_{avg}^{SR} = \frac{4lN - [2N(1+N) - 3](w+s)}{N} \quad (27)$$

Figure 39: Resonant frequency of SRs as a function of the number of turns N

Analogously to the MSRR case, the resonant frequency tends to a limit value after a certain number of turns.

3.4. Comparison of MSRR and SR Models

The main conclusions concerning the two models are summarized in this section.

The maximum linear dimension of the outer length of a MSRR is on the order of $\frac{l}{30}$ until $\frac{l}{40}$.

For a SR, miniaturization can be extended to the order of $\frac{l}{65}$ until $\frac{l}{250}$.

For equivalent MSRR and SR systems, with the same value of N , the Q -factor of SR is 2 or 3 times greater than the Q -factor of MSRR.

Figure 40: Comparison between the Q -factor of MSRR and SR systems

As a high Q implies a narrow bandwidth, the best compromise between miniaturization and the bandwidth desired must be found when choosing between MSRR or SR topologies.

Considering that for most WPT applications a broad bandwidth is not usually important, SR inclusions are a much more suitable choice.

3.5. Control of the resonant frequency

As the operating distance of the IWPT system is expected to be in the order of meters, the working frequency should be in the order of MHz. However, as shown in the previous section, there is a limit resonant frequency value that can be achieved increasing the number of turns of a SR inclusion. So, for any unit cell's practical size (supposing that the MTM would be expected to work inside of objects as small as a smartphone), the resonant frequency will be in the order of GHz or greater. Thus, the resonant frequency control must be made by adding more capacitance or inductance to the circuit.

Figure 41: MTM with lumped capacitors added to the structure

SR inclusions behave like a parallel LC circuit if only stray capacitance is considered. On the other hand, they behave as a series LC circuit when a lumped capacitor is embedded to their terminals.

This fact has an important consequence concerning Q-factor, as discussed in section 1.3.3, once Q-factor will increase with the square root of the capacitance and not with the square root of the inductance, when the inclusions are behaving as parallel LC circuit, and the opposite, when they are behaving as series LC circuit. In the first case, the operating frequency of the inclusions can be reduced by assembling an interdigital capacitor to the inclusion. In the second one, the inductance of the inclusions must be greatly increased in order to compensate the degradation of the Q-factor caused by the lumped capacitor.

4. Design and fabrication

In this chapter, the design of the MTM is explained.

The structure of the MTM has to provide a response function such as it synthesizes the behavior of a magnetic conductor with negative magnetic permeability. The prototype of the MTM uses a lumped capacitor approach to regulate the resonant frequency keeping it in the range of interest (tens of MHz).

The inductor is a square SR connected in series with a CMS capacitor on the back of the slab via two holes.

Figure 42: Square SR in series with a lumped capacitor via holes

Prototype characteristics:

- $d_{int} = 1\text{ cm}$
- $d_{ext} = 2\text{ cm}$
- $s = 1\text{ mm}$
- $w = 1\text{ mm}$
- $CMS\ capacitor = 100\text{ pF}$
- Period of the lattice (center to center of the unit cells) = 2.3 cm

Electric characteristics of the PCB used in the project:

- $\rho = 10\mu\Omega.mm$
- $\epsilon_r = 4.6$
- Loss tangent = 0.01
- Thickness of the slab = 1.5mm

Calculating the total length of the conductor of the unit cell and its section, its electrical resistance can be estimated:

- $l = 179\text{ mm}$
- $A = 0,017\text{ mm}^2$
- $R = 0,2\ \Omega$

Using the model for the inductance of a square planar resonator presented in section 3.3:

- $L_{SR} = 240\text{ nH}$

The resonant frequency and the Q-factor are estimated:

- $f_0 = 33\text{ MHz}$
- $Q = 180$

Figure 43: Representation of the first prototype

The total dimension of the MTM slab is 20cm².

The printed circuit was drawn using the software Eagle 6.5.0.

5. Experimental protocol

5.1. The Impedance Analyzer

Figure 44: Image of the Impedance Analyzer

The Impedance Analyzer (IA) can measure impedance (inductance, capacitance, and resistance) at spot frequencies or across a range of frequencies. Then, it is possible to determine if the theoretical impedance and capacitance calculated are coherent in the desired frequency range. Therefore, it is possible to verify the operating frequency of the MTM.

Finally, the experimental resonance frequency is determined and compared with the theoretical one.

The impedance analyzer used was the Agilent Model 4294A 40Hz-110MHz.

5.1.1. Details of the measurement

To measure the inductance of the MTM inclusions, the probes were positioned on the terminals of one of the coils. The circuit between this coil and its lumped capacitor was opened to measure pure inductance. Then, the inductance depending on frequency was plotted, and experimentally measured.

The capacitance of the lumped capacitors used in the prototype 1 was also measured to assure that they could work at the desired frequency range.

5.2. Vector Network Analyzer

A Vector Network Analyzer (VNA) is an instrument that measures the network parameters of electrical networks. In the present work, the VNA was used to measure S-parameters because reflection and transmission of electrical networks are easy to measure at high frequencies, although there are other network parameter sets, such as Y-parameters, Z-parameters and H-parameters, it can measure. VNAs are often used to characterize two-port networks, like amplifiers and filters, but they can be used on networks with an arbitrary number of ports.

The VNA used has the following characteristics: Rohde&Schwarz – 20 kHz / 300kHz – 8GHz ZVC.

Figure 45: Vector Network Analyzer

5.2.1. Details of the measurement

To calculate the PTE with and without the MTM, an appropriated probe (that does not interfere in the desired frequency range) was employed to measure the electric current.

Figure 46: Experimental setup to measure the current in the transmitting coil

For the PTE measurements, the VNA is connected to the transmitting coil in two ways: through a conventional probe that acts as a voltage source and through a special probe that measures the electric current indirectly by measuring the variations of the magnetic field surrounding the coil's wire. For the S-Parameter measurements, one probe is connected to the transmitting coil and the second one to the receiving coil. The current in the primary and the S-parameters are measured and plotted by the VNA as a function of frequency.

Figure 47: Experimental setup to measure the S-parameters of the coils

Since the impedance in the receiving coil is purely resistive (hence, constant), it was easier to measure the voltage of the secondary coil and calculate the electric current.

Figure 48: Experimental setup to measure the voltage of the secondary

6. Experimental results and discussion

In this chapter, the experimental results are presented and compared with the analytical ones. They were obtained in Ref [25] and first presented in Ref [26].

Figure 49: Experimental setup

The transmitting and receiving coils have the same characteristics of the ones used in the numerical simulations (coil's diameter equals to 10cm, copper wire of 18 AWG) and are separated by 15cm.

The printed circuit board used in the project was a standard FR4 substrate covered with copper on both sides. The spirals and the transmission lines were etched by standard photoetching procedures on FR4 substrate.

The load connected to the secondary is a resistance of 50 Ohms.

6.1.Verification of the electromagnetic characteristics of the unit cell

The measurements of EM characteristics of the first prototype are summarized and compared with the theoretical ones:

First prototype	Inductance (nH)	Capacitance (pF)	Resonant frequency (MHz)	Q-factor
Analytical	246	110	32	180
Experimental	231	115	36	-

Table 1: Theoretical and experimental parameters of the first prototype

The theoretical and the experimental values are quite similar, proving the accuracy of the model used to design the MTM unit cells.

6.2.Verification of the current in the transmitting coil

Assuming the electric potential of the transmitter as a constant, the current is expected to vary linearly and inversely with the frequency:

$$V = \text{constant} \quad (28)$$

$$Z(\omega) = jL\omega \quad (29)$$

$$I(\omega) = \frac{V}{Z(\omega)} = \frac{V}{jL\omega} \quad (30)$$

Effectively, the relationship obtained by measuring the current as a function of the frequency with the VNA is almost linear as shown by Figure 50, except around the activation frequency of the MTM, where the current response is highly non-linear. When the MTM slab gets close to the transmitting coil, the real part of the transmitter's impedance starts to grow.

Figure 50: Current in the transmitter as a function of the frequency

This effect disappears if the MTM is separated from the transmitter by a distance 3 times larger than the largest dimension of the lattice's unit cell. However, the more the distance between the transmitter and the MTM increases, the more the flux disperses and the less the MTM is capable of amplifying the magnetic potential of the near field, since the amplification power of the MTM comes from the cancellation of the attenuating solutions of EM waves and the waves of the near field are evanescent ones. Hence, there is an optimal distance between the transmitting coil and the MTM that maximizes the PTE and the PDL.

6.3.Verification of the current in the receiving coil

As predicted by the numerical simulations, the experiment shows that the current in the receptor remains the same, no matter how arbitrarily close to the receiving coil the MTM is positioned. Then, it is possible to make very compact receptors assisted by MTM, improving the range of application of this technology.

Figure 51: Current in the receptor as a function of the frequency

6.4.Verification of the enhancement of the PTE and the PDL

To performance this experiment, the MTM slab was positioned at 7.5cm from the transmitter. Around 36 MHz, the electrical potential of the receptor is greatly enhanced by the presence of the MTM slab, as shown in 36Figure 52.

	Frequency (MHz)	Voltage (mV)
Without MTM	36	8.7
With MTM	36	43.7

Table 2: Voltage in the secondary with and without the MTM slab

Figure 52: Voltage in the receiving coil with and without the MTM slab

The experimental gain in PTE is around 400%, what is even better than the one predicted by the simulation for the same configuration (300%).

7. Numerical simulation

In this chapter, some simulations of the IWPT system are presented. The main objective of the simulations is to achieve a better understanding of the mechanisms behind the enhancement of magnetic coupling caused by the introduction of the MTM slab in the inductive link.

In all simulations, the IWPT considered consists of two identical coils with 10cm of diameter (copper wire 18 AWG). The load connected to the secondary is a resistance of 50 Ohms. The numerical method used to obtain the results was the momentum method of the software Agilent ADS. Since the resonant frequency of the MTM slab is 36MHz, the simulations were run in a frequency range between 5MHz and 50MHz.

7.1. Agilent Advanced Design System (ADS)

Advanced Design System (ADS) is an electronic design automation software system produced by Agilent EEsof EDA. It provides an integrated design environment to designers of RF electronic products such as mobile phones, wireless networks, satellite communications, radar systems, and high-speed data links.

Agilent ADS supports every step of the design process—schematic capture, layout, frequency-domain and time-domain circuit simulation, and electromagnetic field simulation—allowing the engineer to fully characterize and optimize an RF design without changing tools [34].

Integrated with ADS, there is Momentum: 3D planar EM (electromagnetic) simulation software for electronics and antenna analysis, a partial differential equation solver of Maxwell's equations based on the method of moments. It accepts arbitrary design geometries (including multi-layer structures) and uses frequency-domain Method of Moments (Momentum) technology to accurately simulate complex EM effects including coupling and parasitic.

In this project, it was used ADS 2011.

7.2. Configuration 1: with no MTM slab

Figure 53: 3D view of the IWPT system with no MTM slab

Figure 54: Magnitude of the current in the primary (right) and secondary (left)

Figure 55: Magnitude of the voltage in the primary (right) and secondary (left)

Figure 56: Impedance in the primary: real part (right) and imaginary (left)

Figure 57: Real power (right) and reactive power (left) in the primary

Figure 58: Power transfer efficiency (left) and power delivered to the load (right)

Figure 59: S-parameters of the system in decibels: S_{11} (left) and S_{21} (right)

7.3.Configuration 2: with a MTM slab positioned at 75mm from the transmitter

Figure 60: 3D view of the IWPT system with one MTM slab positioned at 75mm from the transmitter

Figure 61: Magnitude of the current in the primary (right) and secondary (left)

Figure 62: Magnitude of the voltage in the primary (right) and secondary (left)

Figure 63: Impedance in the primary: real part (right) and imaginary (left)

Figure 64: Real power (right) and reactive power (left) in the primary

Figure 65: Power transfer efficiency (left) and power delivered to the load (right)

Figure 66: S-parameters of the system in decibels: S_{11} (left) and S_{21} (right)

7.4.Configuration 3: with a MTM slab positioned at 20mm from the transmitter

Figure 67: 3D view of the IWPT system with one MTM slab positioned at 20mm from the transmitter

Figure 68: Magnitude of the current in the primary (right) and secondary (left)

Figure 69: Magnitude of the voltage in the primary (right) and secondary (left)

Figure 70: Impedance in the primary: real part (right) and imaginary (left)

Figure 71: Real power (right) and reactive power (left) in the primary

Figure 72: Power transfer efficiency (left) and power delivered to the load (right)

Figure 73: S-parameters of the system in decibels: S_{11} (left) and S_{21} (right)

7.5.Configuration 4: with a MTM slab positioned at 1mm from the transmitter

Figure 74: 3D view of the IWPT system with one MTM slab positioned at 1mm from the transmitter

Figure 75: Magnitude of the current in the primary (right) and secondary (left)

Figure 76: Magnitude of the voltage in the primary (right) and secondary (left)

Figure 77: Impedance in the primary: real part (right) and imaginary (left)

Figure 78: Real power (right) and reactive power (left) in the primary

Figure 79: Power transfer efficiency (left) and power delivered to the load (right)

Figure 80: S-parameters of the system in decibels: S_{11} (left) and S_{21} (right)

7.6.Configuration 5: with a MTM slab positioned at 1mm from the receiver

Figure 81: 3D view of the IWPT system with one MTM slab positioned at 1mm from the receiver

Figure 82: Magnitude of the current in the primary (right) and secondary (left)

Figure 83: Magnitude of the voltage in the primary (right) and secondary (left)

Figure 84: Impedance in the primary: real part (right) and imaginary (left)

Figure 85: Real power (right) and reactive power (left) in the primary

Figure 86: Power transfer efficiency (left) and power delivered to the load (right)

Figure 87: S-parameters of the system in decibels: S_{11} (left) and S_{21} (right)

7.7.Configuration 6: with two MTM slabs one positioned at 20mm from the transmitter and one positioned at 1mm from the receiver

Figure 88: 3D view of the IWPT system with two MTM slabs

Figure 89: Magnitude of the current in the primary (right) and secondary (left)

Figure 90: Magnitude of the voltage in the primary (right) and secondary (left)

Figure 91: Impedance in the primary: real part (right) and imaginary (left)

Figure 92: Real power (right) and reactive power (left) in the primary

Figure 93: Power transfer efficiency (left) and power delivered to the load (right)

Figure 94: S-parameters of the system in decibels: S_{11} (left) and S_{21} (right)

7.8.Configuration 7: with two MTM slabs one positioned at 1mm from the transmitter and one positioned at 1mm from the receiver

Figure 95: 3D view of the IWPT system with two MTM slabs positioned at 1mm from each coil

Figure 96: Magnitude of the current in the primary (right) and secondary (left)

Figure 97: Magnitude of the voltage in the primary (right) and secondary (left)

Figure 98: Impedance in the primary: real part (right) and imaginary (left)

Figure 99: Real power (right) and reactive power (left) in the primary

Figure 100: Power transfer efficiency (left) and power delivered to the load (right)

Figure 101: S-parameters of the system in decibels: S_{11} (left) and S_{21} (right)

7.9.Configuration 8: with two MTM slabs one positioned at 1mm from the transmitter and one positioned at 1mm from the receiver with 20cm of misalignment

Figure 102: 3D view of the IWPT system with two MTM slabs with 20cm of misalignment

Figure 103: Magnitude of the current in the primary (right) and secondary (left)

Figure 104: Magnitude of the voltage in the primary (right) and secondary (left)

Figure 105: Impedance in the primary: real part (right) and imaginary (left)

Figure 106: Real power (right) and reactive power (left) in the primary

Figure 107: Power transfer efficiency (left) and power delivered to the load (right)

Figure 108: S-parameters of the system in decibels: S_{11} (left) and S_{21} (right)

7.10. Results and discussion

Concerning the configuration with one MTM slab only, as deduced from the experimental results, there is an optimal distance from the transmitter that maximizes both PTE and PDL. This optimal distance is of the order of the period of the array (2 cm). Configuration 3 is the optimal configuration when the IWPT is assisted by one MTM only.

When the MTM is too close from the transmitter, as in configuration 4, the reactive power of the transmitting coil, which is basically inductive, is felt by the MTM, putting the image current on the surface of the MTM out of phase with the electric current in the coil. The destructive interference between the currents diminishes the total current provided by the transmitter (and the total magnetic flux as a consequence). Effectively, the real impedance felt in the primary in configuration 4 (Figure 77) is smaller than the real impedance felt in the primary in configuration 3 and 2 (Figure 70 and Figure 60). As a consequence of the destructive interference, the resonant frequency of the MTM splits, worsening the performance of the system since PTE and PDL achieve their maximum at different frequencies.

As shown in chapter 6, the MTM can be positioned arbitrarily close to the receptor without introducing any significant interference in the magnetic link. On the other hand, when the MTM is put too far from the transmitter, the evanescent waves that reach the slab are already too attenuated in amplitude to induce enough current to activate it, creating an even greater reluctance to the magnetic flux. Therefore, most of the energy remains stationary, stored in the free space as magnetic potential energy. Also, the great augmentation of the real impedance felt in the primary increases the real power consumed by it. For such reasons, the PTE (Figure 86) obtained for configuration 5 is even worse than the one obtained in configuration 1, where there is no MTM at all inside the magnetic link (Figure 58). However, in such configuration the presence of the MTM causes a peculiar effect on the transmitter that has a significant importance. As show in Figure 85, the reactive power in the primary goes to zero at the resonant frequency since most of the magnetic potential energy gets stored close to the MTM surface when it is activated, forcing the impedance felt in the primary to be purely resistive.

Figure 109: Corresponding magnetic field distributions along the z-axis: with no MTM (left) and with the MTM excited at the resonant frequency (right). See Ref [25].

Another important fact concerning configurations 4 and 5 is that their S-parameters are exactly the same, so they can be seen as perfectly symmetric configurations. Notice that even PDL is the same for both of them.

As a matter of fact, by comparing the results of configurations 6 and 7, it is clear that when operating with two MTM, the optimal position for the MTM close to the transmitting coil is not 2cm anymore. Its new optimal position is as close to the transmitter as possible since the second MTM positioned close to the receptor cancels the reactive power in the primary. Hence, the first MTM can be put arbitrarily close to the transmitting coil in order to maximize PDL. The behavior of configuration 7 could be described as the superposition of the behaviors of configurations 4 and 5. Since those two configurations are symmetrical, the two MTMs are perfectly matched and do not disturb one another.

In fact, in configuration 7 the effective real impedance felt in the primary is approximately 50 Ohms (Figure 98). Looking at the system based on an antenna approach, the two transducers have matched impedances what maximizes the flow of energy between them. The losses in configuration 7 are mostly due to the fact that there is a significant part of the magnetic potential energy stored outside of the magnetic link. If the magnetic field could be confined in the region of interest, this is to say in the space between the antennas, PTE could be improved.

In configuration 8, the transmitter and the receiver of configuration 7 are positioned with a horizontal misalignment of 20cm. As shown in Figure 107, PTE and PDL are still high in comparison with configuration 1 and other configurations assisted by only one MTM. The robustness to misalignment is an important feature considering that in real applications the terminals would be misaligned most of the time.

	PTE (%)	PDL (Watts)
Configuration 1	1	0.28μ
Configuration 7	50	2.4m

Table 3: Comparison between the IWPT system assisted and not assisted by MTMs (in the best configuration)

As seen in Table 3, in the best configuration (configuration 7), the IWPT system assisted by MTMs have a PTE fifty times higher and a PDL almost one thousand times higher than the same system without the MTM lenses (configuration 1).

8. Conclusions

8.1. Global analysis

The main conclusions from the results found during the realization of this work are:

- A μ -negative MTM only suffices to effectively focalize the magnetic near field;
- When operating with one MTM slab only, there is an optimal distance that maximizes PTE and PDL. This distance is equivalent to the period of the MTM lattice;
- These MTM lenses are more efficient when operating in pairs;
- When operating in pairs, the MTM slab can be arbitrarily close to the transmitting and the receiving coils. Therefore, transmitters and receptors assisted by MTMs could be designed to be as compact as necessary without diminishing the efficiency of the magnetic link (which is essential concerning mobile applications);
- When operating in pairs, MTMs make the magnetic link quite robust to misalignment;
- The tremendous improvement in both PTE and PDL confirms that MTMs are a viable technology to make safe WPT a reality. Besides that, MTMs have potential to revolutionize the entire antenna and EM filter fields.

8.2. Future work

During the realization of this work, there was no available time to conduct new experiments to verify all the new results found in the numerical simulations. Then, the next step is run those experiments to confirm the simulation findings.

Other future developments intended for this work are:

- Simulation of the proposed IWPT system in a multi-receiver scenario
- Optimization of the MTM lenses via asymmetric resonators
- Simulations of the EM near field distribution along the magnetic link
- Optimization of the geometry of the coils to improve the near field confinement
- Verification of the behavior of the MTM lenses when transmitting higher power levels
- Verification of the behavior of the MTM lenses in higher temperatures
- Verification of the impact of the proposed IWPT system on communication systems

9. Bibliography

- [1] [Online]. Available: <https://harshparmar.wordpress.com/2011/08/17/wireless-battery-charger-inductively-coupled-universal-battery-charger-charging-batteries-without-wires-inductive-power-transfer-inductive-charging/>. [Accessed 30 06 2014].
- [2] "6 Companies Pushing Wireless Charging Into The Mainstream," *Forbes*, 15 10 2014. [Online]. Available: <http://www.forbes.com/sites/drewhendricks/2014/10/15/6-companies-pushing-wireless-charging-into-the-mainstream/>. [Accessed 31 6 2014].
- [3] A. Kurs, A. Karalis, R. Moffatt, J. D. Joannopoulos, P. Fisher and M. Soljic, "Wireless power transfer via strongly coupled magnetic resonances," *Science*, no. 317, p. 87, 2007.
- [4] Y. Lim and J. Park, "A novel phase control based energy beamforming techniques in non-radiative wireless power transmission," *IEEE transactions on power electronics*, 2013.
- [5] B. Wang, W. Yezauris and K. Teo, "Wireless Power Transfer: Metamaterials and Array of Coupled Resonators," *Proceedings of the IEEE*, 2013.
- [6] B. Wang, K. H. Teo, T. Nishino, W. Yezauris, J. Barnwel and J. Zhang, "Wireless Power Transfer with Metamaterials," in *European Conference on Antennas and Propagation*, Rome, Italy, 2011.
- [7] J. L.-W. Li, "Wireless Power Transmission: State-of-the-Arts in Technologies and Potential Applications," in *Proceedings of the Asia-Pacific Microwave Conference*, 2011.
- [8] N. d. N. Donaldson and T. A. Perlins, "Analysis of resonant coupled coils in the design of radio frequency transcutaneous links," *Med. & Biol. Eng. & Comput*, vol. 21, pp. 612-627, 1983.
- [9] H. Abe, H. Sakamoto and K. Harada, "A Noncontact Charger Using a Resonant Converter with Parallel Capacitor of the Secondary Coil," *IEEE TRANSACTIONS ON INDUSTRY APPLICATIONS*, vol. 36, no. 2, pp. 444-451, 2000.
- [10] T. Sawant, D. Pilankar, R. Sule and S. Mahadeshwar, "An overview of technological advancements and future possibilities in wireless power transmission," *International Journal of Research in Engineering and Technology*, 2013.
- [11] A. Karalis, J. Joannopoulos and M. Soljic, "Efficient wireless non-radiative mid-range energy transfer," *Annals of Physics*, vol. 323, pp. 34-48, 2008.
- [12] D. Ahn, K. Mehdi and M. Ghovanloo, "Enhanced Wireless Power Transmission Using Strong Paramagnetic Response," *IEEE TRANSACTIONS ON MAGNETICS*, vol. 50, march 2014.
- [13] X. Wei, Z. Wang and H. Dai, "A Critical Review of Wireless Power Transfer via Strongly Coupled," *Energies*, vol. 7, pp. 4316-4341, 2014.
- [14] D. R. Smith and N. Kroll, "Negative Refractive Index in Left-Handed Materials," *Physical Review Letters*, vol. 85, no. 14, pp. 2933-2936, 2000.
- [15] G. Ma, M. Yang, S. Xiao, Z. Yang and P. Sheng, "Acoustic metasurface with hybrid resonances,"

Nature Materials, vol. 13, pp. 873-878, 2014.

- [16] D. R. Smith, J. B. Pendry and M. C. K. Wiltshire, "Metamaterials and negative refractive index," *Science*, vol. 305.
- [17] J. B. Pendry, "Negative refraction makes a perfect lens," *Phys. Rev. Lett.*, vol. 85, 2000.
- [18] N. Engheta and R. Ziolkowski, *Metamaterials: physics and engineering explorations*, Wiley & Sons, Inc, 2006.
- [19] J. B. Pendry, A. J. Holden, D. J. Robbins and W. J. Stewart, "Magnetism from conductors and enhanced non-linear phenomena," *IEEE Transactions*, 1999.
- [20] A. Grbic and G. V. Eleftheriades, "Overcoming the Diffraction Limit with a Planar Left-Handed Transmission-Line Lens," *Physical Review Letters*, vol. 92, no. 11, 2004.
- [21] V. A. Fedotov, M. Rose, S. L. Prosvirnin, N. Papasimakis and N. I. Zheludev, "Sharp Trapped-Mode Resonances in Planar Metamaterials with a Broken Structural Symmetry," *Physical Review Letters*, vol. 99, 2007.
- [22] S. Engelbrecht, F. Kraushofer, A. M. Shuvaev, B. Jin, S. A. Kuznetsov and A. Pimenov, "Asymmetric split-ring resonators: a way toward high-quality metamaterials," *Optical Engineering*, vol. 53, 2014.
- [23] J. McVay, V. Pierro, V. Galdi, A. Hoorfar, N. Engheta and I. M. Pinto, "Metamaterial inclusions based on grid-graph Hamiltonian paths," in *Proc. 3rd Workshop on Metamaterials and Special Materials for Electromagnetic*, Rome, Italy, 2006.
- [24] F. Bilotti, A. Toscano and L. Vegni, "Design of Spiral and Multiple Split-Ring Resonators for the Realization of Miniaturized Metamaterial Samples," *IEEE TRANSACTIONS ON ANTENNAS AND PROPAGATION*, vol. 55, no. 8, pp. 2258-2265, 2007.
- [25] S. I. Nishimura, "Enhancing the inductive coupling and efficiency of wireless power transmission system by using metamaterials," *Escola Politécnica da Universidade de São Paulo and Ecole Centrale de Lyon, Ecully*, 2014.
- [26] S. I. Nishimura, J. V. De Almeida, C. Vollaie, C. A. F. Sartori, A. Breard, F. Morel and L. Krähenbühl, "Enhancing the inductive coupling and efficiency of wireless power transmission system by using metamaterials," in *16º SBMO - Simpósio Brasileiro de Micro-ondas e Optoeletrônica e 11º CBMag*, Curitiba, 2014.
- [27] F. Capolino, *Applications of Metamaterials*, CRC Press, 2009.