



Résonateurs rf homogènes à haute sensibilité pour des applications en ingénierie tissulaire utilisant un scanner irm de paillasse : exploration d'une approche 3d plastronique et cryogénique

Valernst Martial Gilmus

► To cite this version:

Valernst Martial Gilmus. Résonateurs rf homogènes à haute sensibilité pour des applications en ingénierie tissulaire utilisant un scanner irm de paillasse : exploration d'une approche 3d plastronique et cryogénique. Medical Imaging. Université Claude Bernard - Lyon I, 2025. English. ⟨NNT : 2025LYO10359⟩. ⟨tel-05530070⟩

HAL Id: tel-05530070

<https://theses.hal.science/tel-05530070v1>

Submitted on 27 Feb 2026

HAL is a multi-disciplinary open access archive for the deposit and dissemination of scientific research documents, whether they are published or not. The documents may come from teaching and research institutions in France or abroad, or from public or private research centers.

L'archive ouverte pluridisciplinaire **HAL**, est destinée au dépôt et à la diffusion de documents scientifiques de niveau recherche, publiés ou non, émanant des établissements d'enseignement et de recherche français ou étrangers, des laboratoires publics ou privés.



Distributed under a Creative Commons CC BY-NC-ND 4.0 - Attribution - Non-commercial use - No Derivative Works - International License



THESE de DOCTORAT DE L'UNIVERSITE CLAUDE BERNARD LYON 1

**Ecole Doctorale N° 160 :
Electronique, Electrotechnique et Automatique (EEA)**

Discipline : Ingénierie pour le vivant

Soutenue publiquement le 18/12/2025, par :
Valernst Martial GILMUS

High Sensitivity Homogenous RF Resonators for Large Tissue Engineering Applications Using a Benchtop MRI Scanner

Exploring a 3D Plastronics and Cryogenic Approach

Devant le jury composé de :

Pinzano,	Astrid	Directrice de Recherche	Université de Lorraine/CNRS	Présidente
Potelon,	Benjamin	Professeur des Universités	IMT Atlantique	Rapporteur
Zivkovic,	Irena	Associated Professor	Eindhoven University of Technology	Rapportrice
Jannin,	Sami	Professeur des Universités	Université Claude Bernard Lyon 1	Examineur
Lambert,	Simon	Maître de conférence, HDR	Université Claude Bernard Lyon 1	Directeur de thèse
Cabrera,	Michel	Chargé de recherche	INSA Lyon / CNRS	Co-encadrant de thèse
Petiot,	Emma	Chargé de recherche	Université Claude Bernard Lyon 1 / CNRS	Invitée
Gerges,	Tony	Docteur	INSA Lyon	Invité

Acknowledgments

I would first like to express my sincere gratitude to the members of the jury for their time, their careful evaluation of this work, and the insightful discussions that contributed to improving this manuscript. I am particularly honored by the presence of Dr. Astrid Pinzano as president of the jury, and I warmly thank Prof. Benjamin Potelon and Prof. Irena Zivkovic for their thorough reviews and valuable comments.

My deepest thanks go to my PhD supervisor, Dr. Simon Lambert, for his constant guidance, scientific rigor, and trust throughout this doctoral journey. His vision and high standards have been a major driving force behind this work. It is often said in academia that the success of a doctoral project relies greatly on the quality of the supervisor-student relationship. In this regard, I consider myself particularly fortunate. Our shared scientific curiosity, aligned expectations, and complementary working styles created an exceptionally productive and motivating research environment. From the very beginning, there was a strong intellectual and human alignment, which fostered both scientific rigor and the freedom necessary to explore ambitious ideas.

I am also grateful to my co-supervisor, Dr. Michel Cabrera. Throughout these years, I was able to fully benefit from his technical expertise, practical insight, and extensive experience. His depth of knowledge, combined with our shared enthusiasm for 3D Plastronics and its many applications, greatly contributed to the success of this project. I sincerely thank him for his guidance and support.

I would like to warmly thank the advanced packaging team at the Ampère Laboratory, Philippe Lombard, Vincent Semet, Tony Gerges, and Bruno Allard, for their support, stimulating discussions, and collaborative spirit. My sincere appreciation also goes to the ICBMS GAMBAS team, especially Emma Petiot and Christophe Marquette, for their scientific input and fruitful interdisciplinary exchanges.

More broadly, I thank all members of the Ampère Laboratory for their ever-present support and the excellent working environment they provide.

I am very grateful to my fellow PhD candidates, Jean-Lynce Ganago, Linh Nguyen, Camilla Pereira Sousa, Caroline Goument, and all colleagues who, in one way or another, contributed to this journey, supported me during challenging moments, and helped create a motivating and friendly research atmosphere.

Special thanks go to Yilbert Henrique Gimenez for his support during his post-doctoral work and for his valuable help in reaching the experimental milestones required for this thesis.

Beyond the laboratory, I wish to thank my friends Blandine, Lulu, Pierre R., Anaëlle, Lisa, Lara, and Pierre K. for their encouragement, their patience, and the balance they brought throughout these demanding years.

I would like to express my deepest gratitude to my childhood friend, Kevin, whose constant support, motivation, and unwavering belief in me have pushed me to persevere and to live up to the ideals of excellence and resilience we have always strived for.

My heartfelt thanks go to my parents, Ernst and Valiollah, for their unconditional love, sacrifices, and lifelong support. They have been, and continue to be, the rock and the foundation that have allowed their children to reach ever greater heights.

I also warmly thank my siblings, Valérie and Valer, for their patience, their understanding, and their constant presence throughout this journey. Sharing this period of my life with you both and knowing I could rely on you during the more demanding moments, meant far more than words can adequately convey.

Last but certainly not least, I wish to thank Caroline. Your unwavering support, kindness, and not-so-quiet strength have been my anchor throughout this demanding adventure. Your presence has brought balance, light, and meaning to this journey far beyond what words can fully express.

Going through life surrounded by such an amazing support system is an honor and a privilege. Thank you all.

« *Tu serres les dents et tu continues...* »

— Ernst Gilmus

Abstract

Recent advances in large 3D bio-engineered tissue constructs require new handling and imaging solutions. While complex 3D constructs demand advanced perfusion-based culture chambers for nutrient flow and waste removal, existing systems are often bulky, complex, and poorly suited for integration with imaging tools. Traditional imaging modalities are frequently destructive and fail to address the depth and scale of these tissues.

Among available modalities, magnetic resonance imaging (MRI) stands out for its ability to provide high-resolution, noninvasive imaging. However, high-field MRI systems are impractical for routine laboratory use due to cost and accessibility. Low-field benchtop MRI systems have emerged as a practical solution, but their use is associated with a decrease in signal-to-noise ratio (SNR), which limits image quality. A critical factor in achieving optimal results lies in efficiently using the available space while integrating necessary functionalities, including optimized, highly sensitive, shielded radiofrequency (RF) coils.

To take full advantage of MRI as an efficient modality for 3D living samples, it is necessary to design a setup combining a 3D perfused chamber with a form-fitted shielded coil in a single device. Moreover, the setup must fit into the limited available space inherent to benchtop MRI systems. A 3D Plastronics approach, enabling the full integration of electronic circuits on 3D surfaces, offers a promising solution. Using this method to integrate well-dimensioned coils directly onto the surface of the culture chamber allows for the creation of a cohesive system that combines imaging and tissue culture functionalities, ensuring optimal coil positioning, maximizing sensitivity, and maintaining a controlled, vivo-like environment.

This thesis aims to present a scalable and cost-effective platform for characterizing 3D tissue-engineered constructs by combining the adaptability of 3D Plastronics manufacturing with a cost-effective, accessible, and portable low-field MRI scanner for in-line integration in the tissue fabrication process.

Chapter I establishes the critical need for a practical imaging modality in tissue engineering. It delves into MRI principles, and identifies the radiofrequency (RF) coil as one of the most critical components for SNR optimization in low-field systems. It considers using 3D Plastronics to create coils with optimized geometries and enhanced sensitivity.

Chapter II details a systematic, simulation-driven process to design and optimize an RF coil for a low-field benchtop MRI scanner, beginning with a theoretical review and moving to comprehensive 3D electromagnetic co-simulations to predict real-world performance and choose the design best suited.

Chapter III covers the development, manufacturing, and validation of an integrated 3D Plastronics culture chamber. The chapter details the design process, the integration of a high-performance RF coil using a Rapid 3D Plastronics (R3DP) approach. It demonstrates the system's performance through a high-resolution, in situ MRI of a perfused, bio-printed hydrogel.

Chapter IV explores another strategy to overcome sensitivity limitations, using cryogenic cooling of the RF coil. The design, manufacturing, and validation of a novel, proof-of-concept cryogenic setup are presented, laying the groundwork for future refinements, such as leveraging 3D Ceratronics to create coils with integrated microfluidic cooling channels for superior thermal management.

Résumé

Les avancées récentes en ingénierie tissulaire 3D pour des tissus bio-conçus nécessitent de nouvelles solutions d'expérimentation et d'imagerie. Alors que les bio-constructions 3D complexes exigent des chambres de culture avancées basées sur un système de perfusion pour le flux des nutriments et l'élimination des déchets, les systèmes existants sont souvent encombrants, complexes et peu adaptés à l'intégration avec les outils d'imagerie. Les modalités d'imagerie traditionnelles sont fréquemment destructrices et ne parviennent pas à mesurer la profondeur et l'échelle de ces tissus.

Parmi les modalités disponibles, l'imagerie par résonance magnétique (IRM) sort du lot par sa capacité à fournir une imagerie non invasive à haute résolution. Cependant, les systèmes d'IRM à haut champ sont peu pratiques pour une utilisation quotidienne en laboratoire en raison de leur coût et de leur accessibilité. Des systèmes d'IRM de paillasse à bas champ ont vu le jour, mais leur utilisation est associée à une diminution du Rapport Signal sur Bruit (RSB), ce qui limite la qualité de l'image. Un facteur critique pour obtenir des résultats optimaux réside dans l'utilisation efficace de l'espace disponible tout en intégrant les fonctionnalités nécessaires, y compris des résonateurs radiofréquence (RF) blindés, optimisés et très sensibles.

Pour tirer pleinement parti de l'IRM en tant que modalité efficace pour les échantillons vivants 3D, il est nécessaire de fabriquer un dispositif combinant une chambre perfusée 3D avec un résonateur blindé épousant sa forme en un seul appareil. De plus, le dispositif doit s'intégrer dans l'espace restreint disponible propre aux systèmes d'IRM de paillasse. Une approche Plastronique 3D, permettant l'intégration complète de circuits électroniques sur des surfaces 3D, offre une solution prometteuse. L'utilisation de cette méthode pour mettre en place des résonateurs convenablement dimensionnés directement sur la surface de la chambre de culture permet la création d'un système cohésif qui allie les fonctionnalités d'imagerie et de culture tissulaire, assurant un positionnement optimal du résonateur dans la zone d'imagerie, maximisant la sensibilité et conservant un environnement contrôlé reflétant les caractéristiques *in vivo*.

Cette thèse vise à présenter une plateforme agile, évolutive et rentable pour la caractérisation de bioconstructions tissulaires 3D en combinant la flexibilité de la fabrication Plastronique 3D avec un scanner IRM à bas champ peu coûteux, accessible et portable pour une intégration en ligne lors de la production des tissus.

Le chapitre I présente le besoin critique d'une modalité d'imagerie pratique et modulable en ingénierie tissulaire. Il approfondit les principes de l'IRM et identifie le résonateur radiofréquence (RF) comme l'un des éléments les plus critiques pour l'optimisation du RSB dans les systèmes bas champ. L'utilisation de la Plastronique 3D pour créer des résonateurs aux géométries optimisées et à la sensibilité améliorée est considérée.

Le chapitre II détaille un processus guidé par la simulation et les calculs analytiques pour concevoir et optimiser un résonateur RF pour un scanner IRM de paillasse à bas champ, en commençant par une revue théorique, puis en s'orientant vers des co-simulations électromagnétiques 3D complètes pour prédire les performances en conditions réelles et sélectionner la conception la plus adaptée.

Le chapitre III couvre le développement, la fabrication et la validation d'une chambre de culture Plastronique 3D. Le chapitre détaille le processus de conception et l'intégration d'un résonateur RF haute performance en utilisant une approche de Plastronique 3D Rapide (R3DP). Il démontre les performances du système à travers une IRM *in situ* à haute résolution d'un hydrogel bio-imprimé et perfusé, ainsi qu'une caractérisation détaillée.

Le chapitre IV explore une autre stratégie pour surmonter les limites de sensibilité, en utilisant un système de refroidissement cryogénique pour le résonateur RF. La conception, la fabrication et la validation d'une preuve de concept de dispositif cryogénique sont présentées, posant les bases d'optimisations futures, tels que l'exploitation de la Cératronique 3D pour créer des résonateurs avec des canaux de refroidissement microfluidiques intégrés pour une gestion thermique améliorée.

Table of Contents

ACKNOWLEDGMENTS.....	2
ABSTRACT.....	5
RESUME	6
TABLE OF CONTENTS	8
LIST OF FIGURES	11
LIST OF TABLES	19
LIST OF ACRONYMS	20
NOMENCLATURE	24
GENERAL INTRODUCTION.....	28
CHAPTER I : TOWARDS A PRACTICAL AND VERSATILE IMAGING MODALITY FOR TISSUE ENGINEERING.....	31
Chapter Overview	31
I. Introduction: Towards a Better Characterization Tool for Large Living 3D Bio-Engineered Constructs	32
I.A. Optical Modalities for 2D and thin samples	32
I.B. 3D Characterization Modalities for Large and Thick Samples	32
II. Magnetic Resonance Imaging as Versatile Characterization Tool	34
II.A. Overview of Classical MRI Scanners	34
II.B. Fundamental Principles: From Nuclear Spins to Images	36
II.C. Tailoring MRI for Complex Tissue Construct Assessment.....	48
III. Overcoming MRI's Limitations for Laboratory Use.....	53
III.A. Benchtop MRI as a Practical Solution	53
III.B. The Signal-to-Noise Ratio (SNR) Challenge	57
III.C. The RF Coil: The Gateway to Signal.....	60
III.D. Benefits of a 3D Plastronics Approach to Coil Fabrication	75
IV. Consolidating a Practical Approach Towards Tissue Engineering Applications	83
CHAPTER II : OPTIMAL HOMOGENOUS RF RESONATORS FOR SMALL SAMPLE IMAGING USING A BENCHTOP MRI SCANNER.....	85
Chapter Overview	85
I. Introduction: The Imperative for Simulation-Driven Design	86
II. Overview of the Selected Benchtop MRI Scanner	87
III. Theoretical Foundations for Homogenous RF Resonators and Preliminary Analysis	88
III.A. Cylindrical Distribution Based Designs (Solenoid Type Coils)	89
III.B. Spherical Uniform Current Density Designs.....	101
III.C. Interpretation and Discussion.....	111

IV.	Realistic Coil Topology Modeling for Further Investigation	113
IV.A.	Simulation Methodology	113
IV.B.	Bridged Loop-Gap Resonator (BLGR)	121
IV.C.	Varying Width Ribbon Solenoid (VWRS)	126
IV.D.	H&D Extended N-Ring Spherical Coil	131
V.	Comparative Performance of The Coils Under Investigation	135
V.A.	Homogeneity	136
V.B.	Noise Dominance	136
V.C.	Sensitivity	137
VI.	Conclusion and Discussion	138
CHAPTER III : 3D PLASTRONICS CULTURE CHAMBER FOR TISSUE ENGINEERING APPLICATIONS.		140
Chapter Overview		140
I.	Introduction	141
II.	Manufacturer Provided Coil and Sample Fixation Mechanism	142
II.A.	Manufacturer Provided Shielded Coil	142
II.B.	Sample Fixation Mechanism	142
III.	3D-Bioprinted Scaffold Design	143
IV.	3D Plastronics Perfused Chamber Design and Fabrication	144
IV.A.	Custom Made Culture Chamber and Assembly	144
IV.B.	Fluidic System Validation	145
IV.C.	Fitted RF Coil Integration	153
IV.D.	RF Validation Through EM Simulations.....	156
IV.E.	R3DP Fabrication	159
V.	3D Plastronics Setup Characterization and Validation	164
V.A.	Materials and Methods.....	164
V.B.	3D Plastronics Setup Characterization Results	166
V.C.	3D Application to High Resolution Scaffold Morphology Assessment.	169
V.D.	Comparative Performance Against a More Classical Approach	171
VI.	Conclusion and Discussion	175
CHAPTER IV : A CRYOGENIC SOLUTION FOR BENCHTOP MRI SCANNERS - PROOF OF CONCEPT.....		177
Chapter Overview		177
I.	Introduction:.....	178
II.	Theoretical Background: Cryogenically Cooled Coils for Improved SNR.....	178
II.A.	Electrical Properties of Cryogenic Copper	178
II.B.	Impact on Coil Performance and SNR.....	180
III.	Design of the cryo-cooled MRI probe	182

III.A.	Proposed Basic Concept	182
IV.	EM Modeling of a Cryogenically Cooled BLGR Design	183
IV.A.	Materials and Methods.....	183
IV.B.	Simulation Results	185
V.	Practical Cryogenic Coil Setup Implementation	187
V.A.	Cryostat Design	187
V.B.	Integrated RF Coil	189
V.C.	Shielded Enclosure with Incorporated Tuning Mechanism	189
V.D.	Culture Chamber.....	194
VI.	Results.....	195
VI.A.	Fabricated Cryostat Setup	195
VI.B.	Cooling Results.....	196
VI.C.	Cryogenic Coil Performance.....	199
VII.	Discussion and Conclusion.....	200
	GENERAL CONCLUSION	202
	BIBLIOGRAPHY.....	206
	ANNEX I.....	219
	ANNEX II	222
	ANNEX III	224
	ANNEX IV.....	225
	ANNEX V.....	226
	ANNEX VI.....	227
	ANNEX VII.....	229
	ANNEX VIII: RESUME ETENDU EN FRANÇAIS.....	230

List of Figures

Figure 1 : Comparison between traditional 2D and advanced 3D cell cultures. (adapted from [7]).	28
Figure 2 : Comparative overview of imaging modalities based on sample size applicability and the depth of information provided. (Adapted from [13]).	33
Figure 3 : Classical clinical MRI System and its components	35
Figure 4 : Color-coded classification of MRI scanners by magnetic field strengths.	35
Figure 5 : Spin representation (inspired by [40]).	36
Figure 6 : Precessional motion of the magnetic moment.	38
Figure 7 : Energy levels ΔE , as a function of magnetic field strength (inspired from [40]).	39
Figure 8 : Bulk Magnetization and Spin Alignment in a Static Magnetic Field.	40
Figure 9 : Schematic example of the relaxation dynamics for a flip angle $\theta=90^\circ$	42
Figure 10: Overview of Magnetization Dynamics after RF pulse excitation	43
Figure 11 : Gradient base frequency encoding principle along the x-axis.	46
Figure 12 : A) frequency encoding along x-axis B) phase encoding along y-axis C) spatially encoded plane with phase and frequency encoding.	47
Figure 13 : Spin Echo example of a cherry tomato k-space data acquisition (On the left) and its corresponding image (On the right) after the inverse Fourier transform (IFT). The images were obtained at the AMPERE laboratory using 300mT benchtop MRI scanner from Pure Devices GmbH's Research Lab line (Pure Devices GmbH, 300 mT Research Lab, Rimpf, Germany).	48
Figure 14 : Example of axial MR Images of a human head taken with (A) proton density weighting: a long TR and short TE, (B) T_1 weighting: a reduced TR and short TE, and (C) T_2 weighting: a long TR and increased TE. (Origin [52])	49
Figure 15 : Images of commercial permanent magnet systems: Bruker ICON scanner (Left) and Aspect Imaging M [72] (Right).	54
Figure 16 : Images of benchtop systems dedicated to research and education: custom 0.18T design by Cooley et al. [27] (Left), Niumag EDUMR 0.5T scanner [96] (Center) and Pure Devices GmbH 0.55T Research Lab [95] (Right)	56
Figure 17 : Diagram Presenting each type of coil and their relation to a spherical sample. (Left) Volume Coil, (Center) Surface Coil and (Right) Array coils.(inspired from [100])	60
Figure 18 : RLC equivalent model of an RF resonator A) R-L-C series configuration and B) (R-L)//C parallel configuration.	62
Figure 19: Simplified acquisition chain schematic.	63
Figure 20 : Cylindrical conductor cross section: A) DC current distribution (uniform) B) AC current distribution (surface concentrated due to skin effect)(inspired by [106]).	67
Figure 21 : Illustration of proximity effects between adjacent conductors (inspired by [106]).	68
Figure 22 : Comparison between a conventional assembly (A) and an IME part (B): the IME version is 77% lighter (150 g vs. 650 g), 93% thinner (3 mm vs. 45 mm), has 96% fewer mechanical	

parts (2 vs. 64), and the printed circuit board is 25% smaller (10×3 cm vs. 10×4 cm), adapted from [117]).	76
Figure 23 :Example of a 3D Plastronics device: integration of a 3D-molded polymer substrate with surface-mounted electronic circuitry, combining structural and electronic functions in a single component (adapted from [118]).	76
Figure 24: A) Example of a personalized, conformal receive-coil array in the shape of a neck brace, spray-coated with a silver-based conductive layer used to maximize SNR [119]. B) Example of a complex 3D Plastronics setup incorporating multiple functions including an imaging coil a tuning system, a heated sample/mouse bed with embedded fluidic warming channels and an anesthesia gas manifold mounted on a three-axis translation stage for precise alignment [120].	77
Figure 25: Rapid 3D Plastronics process: (a) 3D-printed part with integrated grooves fabricated via vat photopolymerization; (b) surface roughening by sandblasting to enhance metal adhesion; (c) uniform application of colloidal palladium catalyst; (d) selective retention of catalyst within grooves after surface sanding; (e) copper deposition by electroless plating; (f) finalized metallized 3D prototype [121].	78
Figure 26: Pyramid of thermoplastic polymers, adapted from [125], [126].	79
Figure 27: Metal thickness in the case of ELD (a) and electrodeposition (b).	80
Figure 28: Schematic diagram of the electroplating process.	82
Figure 29:Open C-shaped permanent magnet of the 0.3T benchtop MRI scanner, featuring a 60 mm magnet gap (with a 55 mm usable width).	88
Figure 30: Representation of the different ways in which a current distribution can create a homogenous field distribution. (adapted from [104])	89
Figure 31: Representation of the ideal cylindrical surface-current distribution producing a homogeneous axial magnetic field.	90
Figure 32: Example of varying pitch winding solenoids: (A) Variable-pitch transmit coil optimized for use in a Halbach-array-based portable MRI scanner (66 mT) [142].(B) Variable-pitch transmit coil for a low-field (50 mT) portable MRI system [143].	92
Figure 33: (A) Example of a Bridged Loop Gap Resonator BLGR. (B) Expected current intensity profile on the cylinder surface[140], [144].	94
Figure 34: Magnetic field magnitude map for classical solenoids with a form factor (FF) of 0.5. The maps are shown for (A) N=10 turns and (B) N=20 turns. The dashed white line indicates the cylindrical Region of Interest (ROI) used for homogeneity analysis. (C) Normalized Population of Deviation (NPD) versus Percentage Field Deviation (PFD) for classical solenoids within the ROI. The plot compares a 10-turn coil (CoV 3.16%) and a 20-turn coil (CoV 2.92%), showing only a slight improvement in homogeneity with more turns.	96
Figure 35: Magnetic field magnitude maps for varying-pitch solenoids (FF=0.5) across a range of taper strengths (α) for N=10 (top row) and N=20 (bottom row) turns. The maps illustrate how increasing (α) extends the field distribution to improve homogeneity within the ROI (dashed white line).	97
Figure 36: Homogeneity analysis of a 20-turn varying-pitch solenoid (FF=0.5). (A) Normalized axial magnetic field profiles for different taper strengths (α), showing how the field is flattened and even raised at the ends. (B) Corresponding PFD histograms, demonstrating an optimal homogeneity at $\alpha=0.25$ (CoV 0.99%).	98

Figure 37: Magnetic field homogeneity of a loop-gap resonator ($FF=0.5$). (A) The magnetic field magnitude map, showing high uniformity within the ROI (dashed white line). (B) The corresponding NPD vs. PFD histogram, which has a sharp peak indicating excellent homogeneity (CoV 0.45%).	99
Figure 38: Normalized axial magnetic field profiles for a classical solenoid ($N=20$) at different form factor (FF) values. The plot shows that field homogeneity decreases significantly as FF increases leading to the coil becoming shorter and wider.	100
Figure 39: Comparison of normalized axial magnetic field profiles for different form factors (FF). (A) A varying-pitch solenoid with an optimized taper α and Form Factor FF sets. (B) A loop-gap resonator. The loop-gap design maintains a flatter, more homogeneous profile across a wider range of geometries compared to the VPS. This might be mitigated by systematically reoptimizing α for a given FF.	101
Figure 40: Representation of the ideal spherical surface-current distribution producing a homogeneous axial magnetic field.	101
Figure 41: Practical implementations of a spherical birdcage resonators: (A) Fabricated by Harpen in 1991 [146]. (B) Fabricated at the AMPERE laboratory in 2021 using a 3D Plastronics approach as an initial practical implementation to verify fabrication feasibility and to serve as a reference for the analytical and simulation-based study presented here.	103
Figure 42: Comparison of the extended Hoult & Deslauriers (H&D Extended) and Harpen methods for $N=4, 6, 8$, and 10 conductors. (A) 2D magnetic field maps for the Harpen method and (B) for the H&D Extended method. The central green area indicates high field homogeneity, and the dashed white circle the ROI (80% of the coil volume). (C) Corresponding histograms of the Normalized Population of Deviation (NPD) versus the Percentage Field Deviation (PFD) within the ROI. The H&D Extended method consistently shows a narrower distribution, indicating superior homogeneity.	110
Figure 43: Comparison of the extended Hoult & Deslauriers (H&D Extended) and Harpen methods for $N=20$. (A) 2D magnetic field maps for the Harpen method and (B) for the H&D Extended method. (C) Corresponding histograms of the Normalized Population of Deviation (NPD) versus the Percentage Field Deviation (PFD) inside the dashed white circle ROI (80% of the coil volume).	111
Figure 44: Preliminary convergence analysis for optimal boundary condition definition near 13 MHz. Comparison between a shielded and shield less configuration with bounding box size affecting cylindrical coil quality factor.	115
Figure 45: Sample geometries used for loaded simulations: (A) The 38 mm diameter spherical sample used to load the spherical shaped coils. (B) The 38×38 mm cylinder used to load the cylindrical coils (VPS and BLGR)	116
Figure 46: Shield enclosure model and ground rails. (A) Copper shield enclosure (25 μm thick) (B) Symmetrical floating copper rails (25 μm thick) placed 5 mm below the shield to maintain structural symmetry and allow artificial grounding. (C) Partial co-simulation schematic showing shield and rail connection common with ground reference.	118
Figure 47: Example of the single port and balanced L-network configuration used in all co-simulation setups.	119
Figure 48: Smith charts illustrations of the graphical Q-factor evaluation method proposed by Kajfez and Hwan [156] for different coupling configurations. (A) Over coupled, (B) under coupled, and (C) critically coupled.	119

Figure 49: 3D model of the selected 13 capacitor bridged loop-gap resonator. (A) Full geometry showing the placement of 13 1111-size MLCCs. (B) Detailed view of the discrete lumped element ports used to model each capacitor and its associated solder pads.	123
Figure 50: Co-simulation schematic for the 13-capacitor BLGR (BLGR_13C). Each port in the 3D EM model is connected to a corresponding lumped element model representing SMD capacitors.....	124
Figure 51: Simulated performance comparison between the 20-capacitor (BLGR_20C) and 13-capacitor (BLGR_13C) designs. (A) S-parameter (S_{11}) response, showing the resonant frequency and loaded quality factor (Q_l). (B) Normalized distribution of the Percentage Field Deviation (PFD), quantifying the magnetic field homogeneity.....	125
Figure 52: Simulated magnetic field (B_1) distribution for the $N=20$ (top row) and $N=13$ (bottom row) configurations. The left column shows the field in the axial plane, and the right column shows the transverse plane at the resonator's center. The dashed lines indicate the region of interest for homogeneity analysis.	125
Figure 53: 3D geometric model of a VWRS coil ($N=10$, $\alpha=0.7$, $d=0.5$) showing the small added protrusions for port connection and the discrete port used for excitation.	127
Figure 54: Field distribution histograms for varying taper strengths (α) in (A) $N=10$ and (B) $N=20$ turn coils.....	128
Figure 55: Loaded quality factor (Q) as a function of the taper parameter (α) for $N=10$ and $N=20$ coils under loaded conditions.....	128
Figure 56: Field distribution histograms for the $N=10$ coil ($\alpha=0.45$) showing the impact of varying the inter-turn gap size (d). Homogeneity, indicated by the CoV, improves as the gap size decreases.....	129
Figure 57: Co-simulation schematic in CST Design Studio. The circuit model includes a matching and tuning network connected to the 3D full-wave model of the VWRS coil.	130
Figure 58: Co-simulation results for the optimized VWRS. (Left) The S_{11} parameter shows a well-matched resonance at 13.03 MHz, yielding a loaded Q-factor of 180. (Right) The final magnetic field distribution histogram, resulting in a CoV of 1.58%.	130
Figure 59: Magnetic field distribution maps for the final optimized VWRS coil. The axial view (left) and radial view (right) both demonstrate a high degree of field uniformity within the central volume of interest (indicated by dashed lines).....	131
Figure 60: 3D models of the $N=4$, $N=6$, $N=8$, and $N=10$ H&D spherical coil geometries as constructed in the EM solver.....	132
Figure 61: Co-simulation schematic example for the $N=4$ H&D spherical coil.	133
Figure 62: Simulated magnetic field magnitude maps in the central transverse plane for the $N=4$, $N=6$, $N=8$, and $N=10$ H&D spherical coil designs.....	134
Figure 63: (A) Shielded coil housing positioned inside the magnet gap. (B) Close-up of the shielded coil housing designed for both vertical and horizontal orientation within the magnet bore. (C) Brass EMI contact fingers used for sliding connection to guide rail ground.....	142
Figure 64: 3D-printed dual-rail sample insertion guide. (A) CAD model showing the graduated PLA base and sliding block designed for precise sample positioning inside the $\varnothing 38$ mm coil bore. (B) Photograph of the assembled guide atop the shielded coil housing.....	143

Figure 65: (A) Isometric CAD rendering of the scaffold showing the cylindrical envelope with two flat sides to prevent rotation inside the culture chamber. (B) Top view illustrating the 4×4 array of 2.5 mm square pores. 143

Figure 66: “Sprinkler” for uniform media flow distribution. (A) Longitudinal sectional view showing the central inlet, conical flow distribution chamber, $\varnothing 1.5$ mm outlet ports aligned with the Scaffold pores. (B) Isometric rendering, highlighting the external fastening thread, flat inner surfaces to prevent scaffold rotation inside the sprinkler and sliders to prevent sprinkler rotation inside the culture chamber. 144

Figure 67 : Longitudinal sectional CAD rendering illustrating the main components of the $\varnothing 35 \times 185$ mm chamber: fastening mechanism, flow distributor (“sprinkler”), 3D bio-printed scaffold, main chamber body, and cap. Culture medium enters through an inlet at the base of the flow distributor, which ensures uniform distribution across the scaffold, and exits through an outlet connected to the return channel (flow path indicated with red arrows). 145

Figure 68: Geometry and mesh used for CFD simulations of the perfusion culture chamber. (A) CAD rendering of the vertical-orientation culture chamber showing the main vessel and the defined inlet/outlet. (B) Corresponding finite-element discretization employed in COMSOL Multiphysics, featuring an extra-fine tetrahedral mesh. Dashed box indicates the region magnified at the right, where the mesh density and element quality are easier to observe. 147

Figure 69: Schematic of the all-purpose coil provided by the MRI manufacturer and the 3D FDM-printed sliding mechanism used for positioning. (A-1) Perfused culture chamber inserted into the sliding mechanism. (A-2) 3D-printed sliding mechanism used for precise alignment within the coil bore. (B-2) Sliding mechanism fully inserted into the shielded enclosure. (B-3) Shielded enclosure housing the BLGR coil. 148

Figure 70 : Schematic of the complete fluidic perfusion setup for MRI velocimetry validation. A Masterflex peristaltic pump (left) drives medium through autoclavable tubing into a dual-reservoir buffer system (center) to smooth pulsatile flow, then onward to the MRI-compatible culture chamber mounted inside a volume coil (right). Red arrows indicate the forward perfusion path. 149

Figure 71: Comparison between experimentally measured (A-1, A-2) and CFD-simulated (B-1, B-2) steady-state velocity magnitude maps inside the culture chamber scaffold region for two different flow rates. The top row (A-1, B-1) corresponds to a flow rate of 20 mL/min, while the bottom row (A-2, B-2) corresponds to a flow rate of 7 mL/min. Velocity maps are overlaid with a red dashed rectangular ROI indicating the region used for quantitative analysis. 151

Figure 72: Comparison of measured and simulated mean velocity profiles across a single 2.5×2.5 mm scaffold channel for flow rates of 7 mL/min (left) and 20 mL/min (right). Experimental data (solid markers) are compared with CFD simulation results (open markers). 151

Figure 73: Comparison of the calculated measured and simulated absolute shear stress profiles across a single 2.5×2.5 mm scaffold channel for flow rates of 7 mL/min (left) and 20 mL/min (right). Experimental gradients of the velocity field (solid markers) and CFD-derived shear stress (open markers) both exhibit the expected peak at the channel walls and a minimum at the center, confirming Poiseuille-type behavior. 153

Figure 74: A) 3D Plastronics-based RF coil integrated design: front and side views. B) Equivalent circuit of the unshielded design. 154

Figure 75: Fully integrated 3D Plastronics-based shielded coil: (A) Shielded enclosure: 3D-printed housing lined with 25 μ m copper foil to ensure RF shielding and low-impedance grounding via SMA feedthrough and structural rails. (B) Functionalized culture chamber: High-temperature

SLA-printed chamber featuring a directly integrated BLGR coil and embedded tuning-matching circuit fabricated using the R3DP method. (C) Sliding mechanism: Custom FDM-printed support designed to precisely position the chamber inside the enclosure and ensure repeatable placement. (D) Handle: External grip allowing easy insertion and removal of the enclosure from the MRI system. (E) RF connection: SMA interface providing electrical access to the embedded coil circuitry for transmit/receive operation. (F) Fluidic inlet/outlet: Ports allowing media circulation and nutrient delivery during *in situ* imaging. (G) Internal connection cable: Coax cable leading from through-panel SMA connector grounded through the shield to the SMA connector. 155

Figure 76: BLGR coil model and co-simulation setup (A) CAD of the 3D-plastronics BLGR: HTV2 substrate (grey) with copper conductive traces with discrete ports/pads for the distributed series capacitors are highlighted in red and numbered along the resonator. (B) Loaded configuration showing the culture chamber and a cylindrical 16.3 mM Copper (II) Nitrate Trihydrate contrast agent solution phantom occupying the entire chamber. (C) CST Design Studio schematic used for co-simulation. 157

Figure 77: Simulated B_1 +magnitude maps. Central slices from the full-wave FEM model with the chamber and shield included: (A) sagittal, (B) axial, and (C) coronal planes. The cylindrical region of interest ($\varnothing 20 \times 40$ mm) used for homogeneity analysis is outlined with white dashed lines. Within this ROI, the field is highly uniform (coefficient of variation $\approx 2.07\%$). Black overlays indicate the chamber and coil geometry. 158

Figure 78: Simulated matching behavior. (A) $|S_{11}|$ versus frequency for unloaded (dashed) and 0.9% NaCl-loaded (solid) conditions, showing a sharp resonance tuned near 13.03 MHz with a deep return loss (~ -50 dB under load). (B) Corresponding Smith-chart trajectories, confirming good 50Ω tuning/matching with minimal load-induced shift. 159

Figure 79 :Structure of a palladium-tin (Pd/Sn) colloid used for surface activation in electroless metallization. The colloidal particle consists of a Pd/Sn alloy core surrounded by tin ions (Sn^{2+} , Sn^{4+}), chloride ions (Cl^-), and protons (H^+) in an acidic medium. 161

Figure 80: Overview of the 3D Plastronics fabrication process used to integrate the RF coil and matching network onto the culture chamber. (A) SLA-printed culture chamber produced using high-temperature resin. (B) Electroless copper deposition applied uniformly across the outer surface. (C) Electrolytic copper plating performed to increase the overall copper thickness and improve conductivity. (D) Selective masking applied to define coil and circuit geometries using spray paint over pre-designed grooves. (E) Ferric chloride etching is used to remove exposed copper, leaving behind the patterned conductive tracks. (F) Masking removed to reveal clean copper traces corresponding to the final circuit layout. (G) Final assembly with passive components and an SMA connector soldered onto the structure, completing the fully integrated coil with embedded tuning and matching circuitry. 162

Figure 81: Fully integrated benchtop MRI setup using 3D Plastronics: (A) Close-up of the culture chamber with an embedded RF coil and integrated tuning/matching components directly on the outer surface. (B) Final assembly showing the shielded enclosure with sliding mechanism, fluidic and RF interfaces. 163

Figure 82: Measured and simulated S-parameter results for the loaded 3D Plastronics RF coil. (A) Magnitude of the reflection coefficient $|S_{11}|$ of the loaded (A-1) and unloaded (A-2) setup as a function of frequency, showing excellent agreement between simulation (blue dashed line) and measurement (red) near the tuned frequency of 13.1 MHz. (B) Corresponding Smith chart for the loaded (B-1) and unloaded (B-2) setups, illustrating the impedance matching behavior and confirming a close match between measured and simulated coil impedance. 167

Figure 83: **B1** + field maps for the 3D Plastronics RF coil. (A) Measured **B1** + amplitude distribution in the central sagittal plane, showing uniform excitation within the defined region of interest (ROI, white dashed outline). (B) Simulated **B1** + amplitude distribution in the same plane, computed under equivalent loading conditions with culture chamber shaped zero value mask overlay to indicate the chamber geometry. The simulated field pattern closely reproduces the measured profile, enabling quantitative comparison of coil efficiency and homogeneity. 168

Figure 84: *In situ* extrusion-based 3D bioprinting of a hydrogel scaffold directly within the flow distributor (“sprinkler”) structure. (A) Bioprinter setup showing the syringe-based extrusion system. (B) Close-up of the printing process with the 800 μm nozzle aligned to the sprinkler.. 170

Figure 85: MRI of the in-situ 3D printed scaffold using a 3D Gradient Echo sequence (FA: 60° , TE: 4 ms, TR: 200 ms, voxel size: $234 \mu\text{m}^3$, FOV: $30 \times 30 \times 30 \text{ mm}^3$, AQ time = 19m 47s) with the scaffold immersed in 0.9% NaCl saline. (A) Axial and (B) sagittal slices show the internal structure of the scaffold with well-resolved macroporous channels. (C) 3D reconstruction obtained from segmented MRI data, illustrating the overall geometry and spatial fidelity of the printed construct..... 171

Figure 86: 16.3 mM Copper (II) Nitrate Trihydrate solution filled chambers GRE acquisitions (FA = 30° , TE = 4ms, TR=12ms, AQ time = 3m 17s) for SNR measurements of A) the commercial coil setup and B) the 3D Plastronics setup..... 172

Figure 87 : Field maps obtained using the dual-angle mapping (DAM) technique for (A) the commercial coil setup and (B) the 3D Plastronics setup, with both axial and coronal slices shown for each case. Cylindrical regions of interest (ROIs), representing the maximum sample dimensions accommodated by the chambers, are outlined using dashed black lines. 173

Figure 88: Copper Resistivity vs. Temperature for various RRR values. The graph illustrates how electrical resistivity decreases as temperature drops, with higher purity copper (higher RRR) showing a more significant drop in resistivity at cryogenic temperatures due to reduced impurity scattering. The performance enhancement plateaus are clearly visible at lower temperatures. ... 180

Figure 89: Schematic of the working principle of the proposed cryogenic setup, designed for low-cost coil cooling..... 182

Figure 90: Coil and Sample Geometry for Cryogenic Simulation..... 184

Figure 91:: Effect of Temperature on Coil Frequency Response 185

Figure 92: (A) Simulated Q_u and Q_ℓ . (B) Coil Noise Figure vs. Temperature..... 186

Figure 93: Simulated SNR gain vs the coil temperature (T_c). The SNR gain is normalized to the value at room temperature, with the sample temperature ($T_s=293.15 \text{ K}$). 186

Figure 94: Cross-sectional view of the proposed cryostat design. 187

Figure 95: Close up of the sample insulation mechanism..... 188

Figure 96: Detailed schematic of the radio frequency (RF) cryo-coil assembly. It shows the 70 μm contour-cut copper foil applied to the hollow alumina tube, with integrated matching and bridging capacitors. Also shown are the soldered non-magnetic SMA cable for signal transmission and the PT1000 RTD for temperature monitoring. 189

Figure 97: Detailed top view of the shielded enclosure illustrating the proposed tuning mechanism and highlighting the closed conductive loop created by the enclosure (in red)..... 190

Figure 98: A detailed side view of the shielded enclosure, illustrating the proposed tuning mechanism. Symmetrically opposed copper-clad plates are moved via external knobs to adjust the coil's inductance. 191

Figure 99: Simulated S_{11} reflection coefficient showing the tuning range of the RF coil. The resonant frequency shifts by approximately 205 kHz as the tuning plates are moved from a separation of $\Delta x = 175$ mm to $\Delta x = 125$ mm.	193
Figure 100: Impact of tuning plate position on coil performance. The plot shows the relationship between the Q-factor and plate separation, illustrating the trade-off between tuning and coil efficiency.	193
Figure 101 : Longitudinal sectional CAD rendering illustrating the main components of the new $\varnothing 25 \times 85$ mm chamber (flow path indicated with red arrows).	194
Figure 102: Detailed view of the culture chamber assembly within the borosilicate bore. The main image shows the threaded rod mechanism for precise vertical positioning of the chamber. The inset provides a magnified view of the PT1000 RTD sensor integrated at the base of the chamber for temperature monitoring.	195
Figure 103: Images of the fabricated cryostat setup. A) Whole cryostat with the shielded reservoir on top of the shielded RF coil enclosure. B) Inside view of the shielded enclosure showcasing the tuning mechanism and the insulated coil C) Close up of the fabricated coil on alumina substrate. A trimmer capacitor was added to help with mismatch due to cooling.	196
Figure 104: Temperature and loaded Q-factor ($Q\ell$) evolution over the cooling period. The primary y-axis shows the coil temperature (T_c , solid black line) and sample temperature (T_s , dashed gray line) in Kelvin. The secondary y-axis shows the corresponding loaded Q-factor ($Q\ell$, dotted black line), which increases as the coil temperature drops, indicating a reduction in coil resistance.	197
Figure 105: Measured vs. simulated resonant frequency shift during cooling. The primary y-axis shows the coil temperature (T_c , solid black line). The secondary y-axis displays the measured resonant frequency (dashed black line) and the simulation-predicted frequency (dashed gray line) in MHz.	198
Figure 106: GRE acquisitions ($FA = 30^\circ$, $TE = 4$ ms, $TR = 12$ ms, AQ time = 1m 11s) of 16.3 mM Copper (II) Nitrate Trihydrate solution filled chambers for SNR measurements of A) the coil at 300K B) the cooled coil at 90K.	199
Figure 107: Varying-Pitch Solenoid Trajectory an Spacing.	222
Figure 108: Spin Echo MRI acquisitions comparing signal quality and noise performance between a shielded commercial benchmark coil (left) and an unshielded 3D Plastronics coil (right). A cylindrical saline-filled phantom ($\varnothing 20$ mm, L 60 mm) was imaged using ultra-low resolution (A-1, B-1) and low resolution (A-2, B-2) SE sequences. The shielded coil yields markedly higher SNR and lower noise background across both protocols. Imaging with the unshielded coil at low resolution yielded excessive noise and barely noticeable phantom noticeable.	226
Figure 109 : Proposed Improved 3D Plastronics Bioreactor Design.	229

List of Tables

Table 1 : Comparison of Imaging Modalities for 3D Tissue-Engineered Constructs	34
Table 2 : Gyromagnetic ratio of some commonly used nuclei in NMR and MRI.....	37
Table 3: Tissue properties and corresponding MR techniques for tissue-engineering monitoring	52
Table 4 : Overview of volume, surface, and array RF resonators	61
Table 5: Detailed comparison of coil field uniformity metrics: min/max homogeneity, Coefficient of Variation (CoV), and Field Histogram.....	74
Table 6: Useful quantities for cosine-squared based taper winding solenoid geometry	93
Table 7: Optimal taper α and Form Factor FF sets for maximum homogeneity	98
Table 8: Coefficient of Variation (CoV) of the magnetic field, calculated within an 80% spherical ROI, for the H&D Extended and Harpen methods with N=4, 6, 8, and 10 conductors.	109
Table 9: Loaded Quality Factors for Various Inter-turn Gap Sizes	129
Table 10: Positioning angles (θ) and current ratios for the H&D spherical coil designs.....	132
Table 11: Quantitative performance summary for the four H&D spherical coil designs.	134
Table 12: Comprehensive performance metrics of investigated cylindrical coil topologies ..	135
Table 13 : Comprehensive performance metrics of investigated spherical coil topologies ...	136
Table 14: Values of the High Q non-magnetic MLCCs (Knowles Syfer) and High Q PTFE non-magnetic trimmers used in the fabricated setup. All values were obtained from manufacturer provided datasheet and models [154], [170].....	156
Table 15: Comparison of measured and simulated coil performance parameters. Loaded and unloaded quality factors (Q), coefficient of variation (CoV) within the ROI, and mean B1+ amplitude values are reported for both simulation and measurement. The measured Q and mean B1+ amplitude exceed simulated values, while the CoV is lower, indicating slightly improved homogeneity in practice.	169
Table 16: Theoretical SNR Gain Calculated from the simulated and measured loaded Q-factors the coil dimensions and the noise factors (F).....	172
Table 17 : Summary of SNR measurements and gain from cryogenic experiment.....	199
Table 18: Varying-Pitch Solenoid Trajectory Useful Quantities	223
Table 19: Quantitative Comparison of Measured Noise and SNR Performance with and without Shielded Enclosure.....	226
Table 20 :Electrical properties of the two electrolyte solutions at 25 °C and $f = 13$ MHz.	228
Tableau 21 : Techniques IRM et propriétés tissulaires correspondantes pour de l'observation en ingénierie tissulaire.....	231

List of Acronyms

ABS:	Acrylonitrile Butadiene Styrene
AC:	Alternating Current
ADC:	Apparent Diffusion Coefficient
AQ time:	Acquisition Time
ASA:	Acrylonitrile Styrene Acrylate
ASL:	Arterial Spin Labeling
BLGR:	Bridged Loop-Gap Resonator
BT-MRI:	Benchtop Magnetic Resonance Imaging
BW:	Bandwidth
CAD:	Computer-Aided Design
CFD:	Computational Fluid Dynamics
CoV:	Coefficient of Variation
CST-MWS:	CST Microwave Studio
DAM:	Dual Angle Method
DC:	Direct Current
DMEM:	Dulbecco's Modified Eagle Medium
DSC:	Dynamic Susceptibility Contrast
DTI:	Diffusion Tensor Imaging
DWI:	Diffusion-Weighted Imaging
ECM:	Extracellular Matrix
EEA:	Electronique, Electrotechnique, Automatique et Traitement du signal
ELD:	Electroless Deposition
EM:	Electromagnetic
EMI:	Electromagnetic Interference
ESR:	Equivalent Series Resistance
FA:	Flip Angle
FDM:	Fused Deposition Modeling
FEA:	Finite Element Analysis
FEM:	Finite Element Method
FF:	Form Factor
FFT:	Fast Fourier Transform
FID:	Free Induction Decay
FIT:	Finite Integration Technique

FOV: Field of View
GRE: Gradient Echo
H&D: Hoult & Deslauriers
HDR: Habilitation à diriger des recherches
HDT: Heat Deflection Temperature
HDPE: High-Density Polyethylene
HF: High-Field
hiPSCs: Human induced pluripotent stem cells
HIPS: High Impact Polystyrene
HTS: High-Temperature Superconductors
ID: Inner Diameter
IFT: Inverse Fourier Transform
IME: In-Mold Electronics
IPA: Isopropyl Alcohol
LCP: Liquid Crystal Polymer
LGR: Loop-Gap Resonator
LN2: Liquid Nitrogen
MID: Molded Interconnect Device
MLCC: Multilayer Chip Capacitors
MoM: Method of Moments
MRE: Magnetic Resonance Elastography
MRI: Magnetic Resonance Imaging
MRM: Magnetic Resonance Microscopy
MT: Magnetization Transfer
NDT: Non-Destructive Testing
NF: Noise Figure
NMR: Nuclear Magnetic Resonance
NPD: Normalized Population of Deviation
NPR: Noise Power Ratio
OCT: Optical Coherence Tomography
OD: Outer Diameter
PA: Polyamide
PAI: Polyamide-imide
PARA: Polyarylamide
PAR: Polyarylate
PBA: Perfect Boundary Approximation

PBI: Polybenzimidazole
PBT: Polybutylene Terephthalate
PC: Polycarbonate
PCB: Printed Circuit Board
PD: Proton Density
PE: Polyethylene
PEC: Perfect Electric Conductor
PEEK: Polyether Ether Ketone
PEI: Polyethyleneimine
PESU: Polyethersulfone
PET: Positron Emission Tomography
PETG: Polyethylene Terephthalate Glycol
PFD: Percentage Field Deviation
PI: Polyimide
PLA: Polylactic Acid
PMMA: Polymethyl Methacrylate
POM: Polyoxymethylene
PP: Polypropylene
PPA: Polyphthalamide
PPS: Polyphenylene Sulfide
PPSU: Polyphenylsulfone
PSU: Polysulfone
PVC: Polyvinyl Chloride
R3DP: Rapid 3D Plastronics
RF: Radiofrequency
RLC: Resistor-Inductor-Capacitor
ROI: Region of Interest
RRR: Residual Resistivity Ratio
RTD: Resistance Temperature Detector
Rx: Reception
SAR: Specific Absorption Rate
SE: Spin Echo
SLA: Stereolithography
SLS: Selective Laser Sintering
SMB: SubMiniature version B (coaxial RF connector)
SMD: Surface-Mount Device

SNR: Signal-to-Noise Ratio
SPECT: Single Photon Emission Computed Tomography
SPIO: Superparamagnetic Iron Oxide
SRP: Self-Reinforced Polymer
STL: Stereolithography (file format)
SVD: Singular Value Decomposition
T/R: Transmit/Receive
TCC: Temperature Coefficient of Capacitance
TE: Echo Time
TPI: Thermoplastic Polyimide
TR: Repetition Time
TST: Thin Sheet Technology
Tx: Transmission
UHF: Ultra-High Field
USB: Universal Serial Bus
VENC: Velocity Encoding
VNA: Vector Network Analyzer
VPS: Varying-Pitch Solenoid
VWRS: Varying Width Ribbon Solenoid
XRF: X-ray Fluorescence

Nomenclature

General Physics & Electromagnetism

Symbol	Description	Unit
A	Magnetic vector potential	T m
B	Magnetic field (induction)	T
B_0	Main static magnetic field	T
B_1	Radiofrequency magnetic field	T
C	Capacitance	F
E	Electric field	V m ⁻¹
e	Electromotive force (voltage)	V
F	Force	N
I	Electric current	A
J	Current density	A m ⁻²
L	Inductance	H
p	Pressure	Pa
q	Electric charge	C
R	Resistance	Ω
Z	Impedance	Ω
γ	Gyromagnetic ratio	rad s T ⁻¹
ε	Permittivity	F m ⁻¹
ε_r	Relative permittivity (dielectric constant)	Dimensionless
η	Filling factor	Dimensionless
Θ_D	Debye temperature	K
κ	Coupling correction factor	Dimensionless
μ	Permeability	H m ⁻¹

Symbol	Description	Unit
μ_r	Relative permeability	Dimensionless
μ_{visc}	Dynamic viscosity	Pa s
ρ_{charge}	Electric charge density	C m ⁻³
ρ_{fluid}	Fluid density	kg m ⁻³
$\rho_{\text{resistivity}}(T)$	Temperature-dependent resistivity	Ω m
$\sigma_{\text{conductivity}}$	Electrical conductivity	S m ⁻¹
τ_{torque}	Torque	N m
Φ	Magnetic flux	Wb
ω	Angular frequency	rad s ⁻¹
ω_0	Larmor frequency	rad s ⁻¹

Nuclear Magnetic Resonance (NMR) & MRI

Symbol	Description	Unit
ADC	Apparent diffusion coefficient	m ² s ⁻¹
b	b-value (diffusion weighting factor)	s m ⁻²
δV	Voxel volume	m ³
FA (θ)	Flip angle	degrees or radians
G	Magnetic field gradient strength	T m ⁻¹
J_{angular}	Total angular momentum	J s
M	Magnetization vector	A m ⁻¹
M_0	Equilibrium magnetization	A m ⁻¹
s	Spin quantum number	Dimensionless
T_1	Longitudinal (spin-lattice) relaxation time	s
T_2	Transverse (spin-spin) relaxation time	s

Symbol	Description	Unit
T_2^*	Effective transverse relaxation time	s
TE	Echo Time	s
TI	Inversion Time	s
TR	Repetition Time	s
μ_{moment}	Magnetic moment	J T ⁻¹
ϕ	Phase	radians

RF Coil & System Performance

Symbol	Description	Unit
Δf	Bandwidth	Hz
F	Noise Factor	Dimensionless
H_{uni}	Min/max field homogeneity	%
NF	Noise Figure	dB (decibel)
NPR	Noise Power Ratio	Dimensionless
Q	Quality factor	Dimensionless
Q_l	Loaded quality factor	Dimensionless
Q_u	Unloaded quality factor	Dimensionless
R_c	Coil resistance	Ω
R_s	Sample-induced resistance	Ω
R_{rad}	Radiation resistance	Ω
R_{eff}	Effective resistance	Ω
Re	Reynolds number	Dimensionless
RRR	Residual Resistivity Ratio	Dimensionless
S_{ij}	Scattering parameter (S-parameter)	Dimensionless

Symbol	Description	Unit
SNR	Signal-to-Noise Ratio	Dimensionless
T_c	Coil temperature	K
T_s	Sample temperature	K
T_{eff}	Effective noise temperature	K
Γ	Reflection coefficient	Dimensionless
σ_{noise}	RMS noise voltage	V
σ_{std}	Standard deviation	(same as data)
μ_{mean}	Mean (average)	(same as data)

General Introduction

The multidisciplinary field of tissue engineering aims to create functional biological tissues by combining living cells, biomaterials, and bioactive molecules. It is used both to restore or replace damaged tissues in clinical settings and to develop physiologically relevant *in vitro* models for research purposes.

Since the 1980s, the field has evolved significantly [1], mainly due to the transition from 2D to 3D cell cultures, allowing for more complex 3D constructs. Thanks to these developments, today, tissue engineering is not only leveraged to develop clinical therapies such as skin grafts, cartilage repair, and bone regeneration but is also becoming an essential tool in pharmaceutical research, drug development, and disease modeling [2], [3]. By offering the potential to replace animal models with human-relevant constructs [4], 3D tissue engineering also promises to enhance both ethical practices and predictive accuracy in biomedical science.

Central to this 3D approach to cell culture is the recreation of the cellular microenvironment, which plays a critical role in regulating cell behavior. In traditional 2D cultures, cells are grown on flat, rigid substrates that fail to reproduce the complex three-dimensional architecture and biochemical cues of native tissues[5], [6]. As a result, cells in 2D often exhibit altered morphology, polarity, and gene expression profiles compared to their *in vivo* counterparts, leading to limited physiological relevance.

In contrast, 3D cultures aim to more closely mimic the *in vivo* conditions by providing a spatially organized environment that allows for natural cell–cell and cell–matrix interactions. In native tissues, cells are not isolated but are embedded within a dynamic extracellular matrix (ECM) that not only offers mechanical support but also regulates key processes such as migration, proliferation, and differentiation through biochemical and biophysical cues. These multidimensional interactions in 3D environments better replicate tissue dynamics and reveal greater insights into cell behavior and communication. Figure 1 illustrates the difference between 2D and 3D cell cultures. In 2D systems, cells (in grey) adhere to a flat substrate (teal) only on their bottom surface and establish limited cell–cell interactions (in yellow). In contrast, 3D cultures allow cells to form extensive attachments with both neighboring cells and the surrounding matrix across their entire surface, closely replicating the *in vivo* microenvironment.

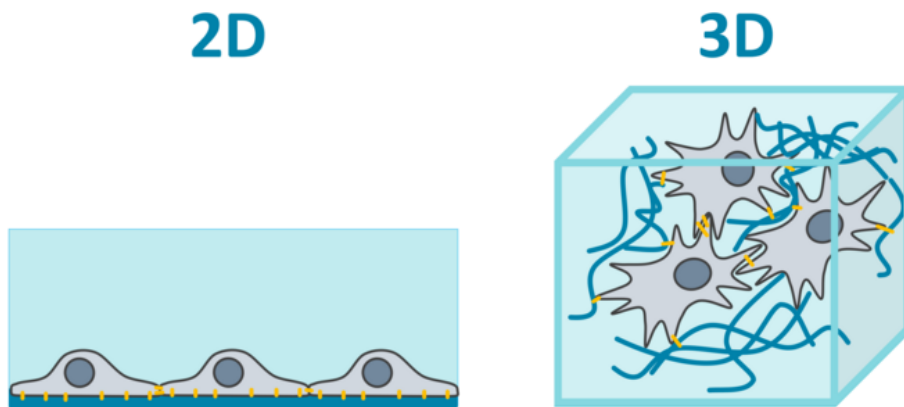


Figure 1 : Comparison between traditional 2D and advanced 3D cell cultures. (adapted from [7]).

Despite the numerous advantages offered by 3D tissue engineering, the complexity of these constructs presents new handling challenges. While 2D cultures often involve thin, monolayered

samples, 3D constructs may extend hundreds of microns to tens of millimeters in depth [8], demanding more advanced culture systems.

An optimal nutrient flow within the culture chamber is crucial for the uniform distribution of essential nutrients, which sustains cell viability, supports metabolic activity, and promotes physiological and biochemical processes necessary for cellular function. In 3D cell structures, passive flow diffusion is inadequate, especially in core regions, necessitating an active flow mechanism and flow distributor to prevent necrosis due to nutrient and oxygen deprivation [9], [10], [11]. Additionally, maintaining a sterile field during the bioprinting process and insertion is equally critical to prevent microbial contamination, which can compromise cell viability and affect tissue development.

Numerous multifunctional perfusion-based culture chambers have been developed to support such growth, incorporating active nutrient flow, waste removal, and mechanical support [12]. However, these systems are often bulky, complex, and poorly suited for integration within imaging tools.

Traditional imaging modalities optimized for 2D systems often fail to address the depth, scale, and heterogeneity of 3D tissues. Techniques like confocal and multiphoton microscopy offer high spatial resolution but are limited by penetration depths of around 3 mm, making them unsuitable for thick or opaque samples [13], [14]. Furthermore, their destructive nature limits longitudinal studies as they compromise the integrity of constructs. The need for imaging tools capable of integrating high-resolution, deep penetration, and noninvasive characterization has driven the exploration of advanced techniques.

Among available modalities, magnetic resonance imaging (MRI) stands out for its unique ability to provide high-resolution, noninvasive imaging with excellent soft-tissue contrast. MRI enables simultaneous structural and functional analysis, making it well-suited for 3D tissue engineering applications. Techniques such as diffusion-weighted imaging (DW-MRI), perfusion measurements, and magnetic resonance elastography (MRE) extend MRI's utility by evaluating tissue mechanics (i.e. mainly extracellular matrix), cellular density, and nutrient flow dynamics [15], [16], [17]. However, while high-field MRI systems (7T–11T) deliver superior sensitivity and resolution, their high cost, susceptibility to artifacts, and limited accessibility render them impractical for routine use in a bioengineering laboratory environment. Mid-field systems (1.5T–3T) provide an alternative [18], but they are still expensive and less adaptable for laboratory-scale tissue engineering workflows.

Low-field benchtop MRI systems ($\leq 0.5T$) have emerged as a practical solution, offering portability, cost-effectiveness, and sufficient resolution for many tissue engineering applications. In the literature, many studies have already showcased the use of low-field MRI machines for small animal studies [19], [20]. There are only a few promising results on the use of a 0.5T benchtop MRI (BT-MRI) for tissue engineering, highlighting its utility, particularly in characterizing different scaffold types and monitoring mouse tumor model progression [21], [22]. However, the use of low-field MRI is associated with a decrease in SNR, which limits the achievable spatiotemporal resolution and the quality of the parametric reconstructed images. Additional SNR at low field can be obtained using dedicated acquisition sequences, postprocessing tools (e.g., denoising filters, bias-field correction, AI-based enhancement etc.) [23], or MRI coils [24], [25], [26]. Moreover, these systems generally operate in unshielded environments, which is associated with an additional reduction in SNR exceeding fivefold compared to their shielded counterparts [27]. To compensate, it requires adding a local RF shield, which complicates setup design and fabrication. A critical factor in achieving optimal results lies in efficiently utilizing the available space while integrating all the necessary functionalities, including optimized shielded radiofrequency (RF) coils.

This thesis is structured to address these challenges by developing and validating optimized hardware solutions based on a 3D Plastronics and cryogenic approach for low-field benchtop MRI tailored to tissue engineering.

Chapter I delves into the fundamental principles of MRI, from the physics of nuclear spin to the intricacies of signal detection, spatial encoding, and image reconstruction. It further explores how specialized MRI sequences can be tailored to create a powerful "toolbox" for monitoring key tissue engineering parameters, such as cellularity, matrix deposition, and nutrient flow. Despite some advantages, the practical barriers of conventional high-field MRI have hindered its widespread adoption in standard laboratory settings. The discussion pivots to low-field benchtop MRI systems as a viable solution to these challenges. However, this introduces the primary trade-off of lower signal-to-noise ratio (SNR). The analysis identifies the radiofrequency (RF) coil as the most critical and effective component for SNR optimization, especially for the high-resolution imaging of small samples.

Chapter II focuses on the simulation-driven design and comparative analysis of various homogeneous RF resonator topologies. Through a combination of analytical models and comprehensive 3D electromagnetic simulations, this chapter investigates and optimizes cylindrical and spherical coil designs to maximize field homogeneity and sensitivity. The goal is to identify the most promising coil architecture for small-sample imaging within the physical constraints of a selected benchtop MRI scanner.

Chapter III presents the practical realization of an optimized RF coil by developing a fully integrated, perfused culture chamber using a 3D Plastronics approach. This chapter details the design, fabrication, and validation of the multifunctional device, which combines a form-fitted Bridged Loop-Gap Resonator (BLGR) with a life-sustaining perfused culture environment. Its performance is validated through high-resolution imaging of a bio-printed scaffold and a direct quantitative comparison against a commercial benchmark coil.

Chapter IV explores a complementary strategy for SNR enhancement by developing a proof-of-concept cryogenic cooling system. This chapter details the design, simulation, and preliminary experimental validation of a low-cost cryostat for cooling a copper RF coil to liquid nitrogen temperatures. The aim is to demonstrate how this approach can increase sensitivity by reducing the coil's thermal noise within the constrained environment of a benchtop MRI setup with the restrictions imposed by tissue engineering applications, offering an alternative path to high-performance imaging on a benchtop system. The proposed design emphasizes simplicity, avoiding overly complex cryogenic techniques to facilitate the custom fabrication of cryostats adapted to versatile coil setups for benchtop MRI applications.

Chapter I: Towards a Practical and Versatile Imaging Modality for Tissue Engineering.

Chapter Overview

This chapter establishes the critical need for a practical and versatile imaging modality within the field of tissue engineering. It begins by contextualizing the shift from two-dimensional to complex three-dimensional cell cultures, highlighting the subsequent demand for non-invasive techniques capable of characterizing thick, opaque biological constructs over time.

A review of current imaging strategies reveals the limitations of established methods, particularly optical modalities, which, despite their high resolution, are constrained by shallow penetration depths. This analysis positions Magnetic Resonance Imaging (MRI) as a uniquely suitable candidate, owing to its ability to provide deep, non-invasive, and multiparametric assessment of both tissue structure and function without the use of ionizing radiation.

The chapter then delves into the fundamental principles of MRI, from the physics of nuclear spin to the intricacies of signal detection, spatial encoding, and image reconstruction. It further explores how specialized MRI sequences can be tailored to create a powerful "toolbox" for monitoring key tissue engineering parameters, such as cellularity, matrix deposition, and nutrient flow.

Despite some advantages, the practical barriers of conventional high-field MRI have hindered its widespread adoption in standard laboratory settings. The discussion pivots to low-field benchtop MRI systems as a viable solution to these challenges. However, this introduces the primary trade-off of lower signal-to-noise ratio (SNR). The analysis identifies the radiofrequency (RF) coil as the most critical and effective component for SNR optimization, especially for the high-resolution imaging of small samples.

Finally, the chapter consolidates these points to propose a focused approach to overcome the SNR limitations of benchtop MRI by developing novel, high-performance RF coils. It introduces the concept of using advanced fabrication techniques, specifically 3D Plastronics, as a promising solution to create coils with optimized geometries and enhanced sensitivity, able to unlock the full potential of benchtop MRI as an accessible, everyday tool in a tissue engineering laboratory.

I. Introduction: Towards a Better Characterization Tool for Large Living 3D Bio-Engineered Constructs

During the tissue engineering development process, monitoring is essential to assess and validate cell viability and phenotype, support structures integrity and function, culture conditions (biochemical and mechanical), and tissue integration. To accomplish this task, many imaging strategies have been explored. Non-destructive multiparametric imaging modalities have become essential tools for monitoring the evolution of engineered tissues over time. These tools need to enable the visualization of cellular distribution, matrix remodeling, nutrient flow mechanisms, and structural integrity within the scaffold parameters that are otherwise difficult to assess without destructive sampling [13]. Techniques such as optical microscopy, X-ray/micro-CT (micro-computed tomography), ultrasound, and magnetic resonance imaging (MRI) offer varying advantages in terms of resolution, depth penetration, and compatibility with live systems.

I.A. Optical Modalities for 2D and thin samples

Optical modalities have long been identified as the gold standard for tissue engineering characterization [28]. However, most of the classical optical methods for imaging two-dimensional samples rely heavily on light transmission through transparent samples and cannot be used for thick, opaque samples. From a technical perspective, optical modalities have excellent planar resolution on the order of a micron or even below, allowing them to access information at the cellular and subcellular scale. However, this is limited by a very shallow penetration depth, which often requires preparing destructive histological slices for analysis [29].

Several strategies have been explored to correct the limitations of traditional bright-field microscopy, including confocal microscopy [30], multiphoton microscopy [31], and Optical Coherence Tomography (OCT) [32], which are able to achieve varying penetration depths of up to 3 mm using lasers and light interferometric techniques. While each of these methods offers varying degrees of spatial resolution and penetration depth, they remain quite limited when it comes to the assessment of thicker 3D structures.

I.B. 3D Characterization Modalities for Large and Thick Samples

The increasing complexity and size of large 3D-engineered tissues introduces significant characterization and monitoring challenges. The limitations of optical modalities have led to the necessity for new dedicated tools for 3D tissue engineering. These tools need to be capable of complex detailed characterization, including subcellular to macroscopic resolution, and assessment of all relevant information covering morphological, physiological, and molecular changes within the sample. They also need to enable safe, quantitative, and noninvasive monitoring with sample thickness penetration depths. So far, there have been a few promising candidates [13]. For example, X-ray and micro-CT provide excellent resolution and contrast for mineralized structures but remain limited in soft tissue imaging due to radiation exposure and poor contrast [15], [16]. Ultrasound offers real-time imaging for mechanical properties and scaffold integrity, though it is hampered by limited soft-tissue contrast [17], [18]. PET (Positron Emission Tomography) and SPECT (Single Photon Emission Computed Tomography) deliver molecular sensitivity but suffer from poor spatial resolution and the need for ionizing tracers [19], [20]. In contrast, magnetic resonance imaging (MRI) stands out for its unique ability to provide high-resolution (up to 20 μm [13]), noninvasive imaging with excellent soft-tissue contrast while enabling simultaneous structural and functional analysis [15]. Figure 2 and Table 1 show a detailed summary of a few interesting imaging modalities with their advantages and drawbacks. Optical modalities provide

high-resolution molecular and cellular information but are limited to small samples. Ultrasound and X-ray/micro-CT extend to larger samples, primarily offering morphological and mechanical insights. PET/SPECT focuses on functional and biochemical imaging across a wide range of sample sizes, without direct morphological or mechanical detail. MRI uniquely combines broad sample size coverage with the ability to simultaneously provide detailed anatomical, physiological, and microstructural information.

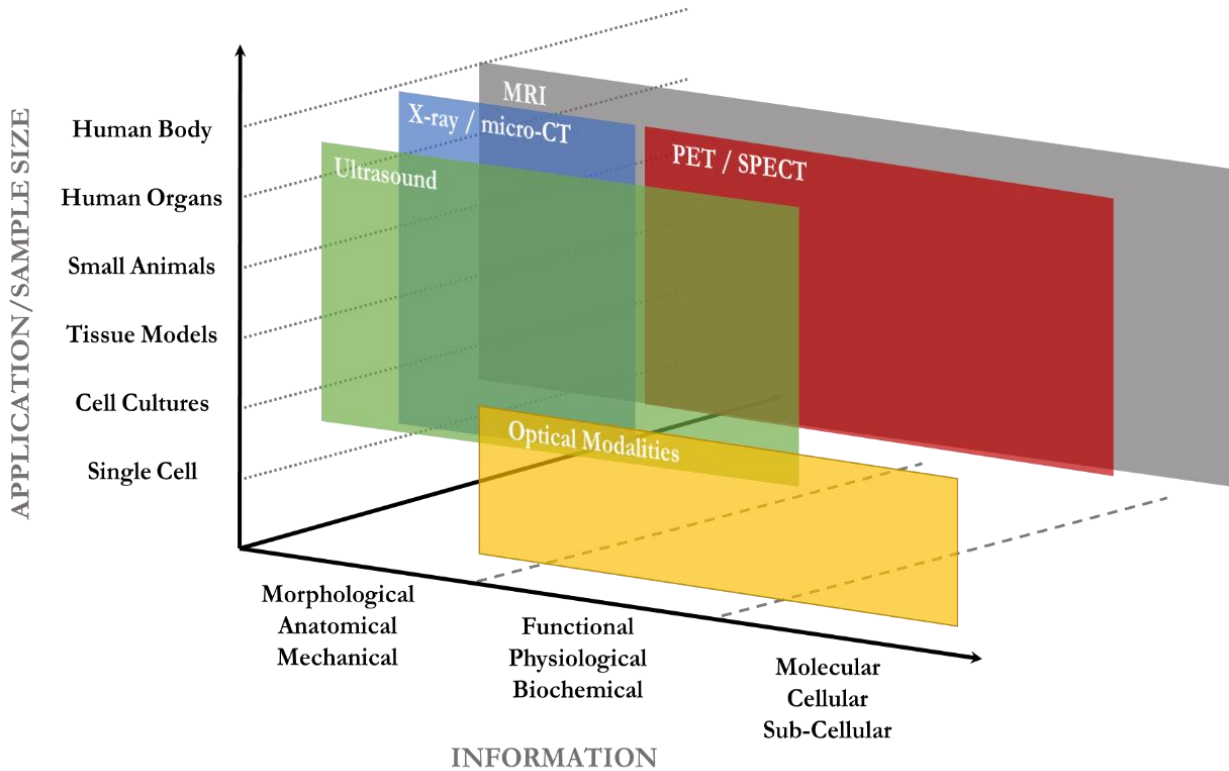


Figure 2 : Comparative overview of imaging modalities based on sample size applicability and the depth of information provided. (Adapted from [13]).

Among available modalities, Magnetic Resonance Imaging (MRI), although lacking the cellular-level resolution of optical microscopy and the high sensitivity of nuclear imaging, is the only technique which enables the non-invasive, three-dimensional assessment of both tissue structure and function deep within optically opaque constructs on multiple scales, making it uniquely suited for 3D tissue engineering applications. While optical methods provide unparalleled subcellular details and PET can track specific molecular pathways, they cannot provide a truly comprehensive view of large, dense living tissues (at the exception of light sheet microscopy which is however only suited non-living tissues [33]). MRI overcomes this critical challenge of penetration depth without using ionizing radiation, permitting longitudinal studies of the same tissue construct over its entire development. Furthermore, MRI's power lies in its multiparametric nature. From a single imaging session, it can yield detailed anatomical maps revealing cell distribution and scaffold integrity, while advanced techniques such as diffusion MRI and perfusion MRI (exposed in more details later in the section 3.2.1) can provide crucial functional and quantitative insights, respectively, on cellularity and nutrient flow/waste management. This rich, quantitative dataset enables a correlation between structural growth and functional viability, offering a comprehensive understanding of the engineered tissue's health and development. Therefore, despite its trade-offs in resolution and cost, MRI's unique combination of deep penetration, excellent soft-tissue contrast, and versatile functional analysis establishes it as an indispensable tool for characterizing and optimizing the complex biological processes at play in large-scale living tissue constructs.

Modality	Signal Type	Spatial Resolution	Temporal Resolution	Penetration Depth	Sensitivity	Strengths	Limitations
Optical (widefield, confocal, multiphoton, OCT)	Light absorption, reflection, fluorescence	~0.2–5 μm (modality-dependent)	Seconds to minutes per frame/volume	~100 μm (widefield) to ~3 mm (OCT, multiphoton)	High (fluorescence labeling)	Cellular/subcellular detail; high specificity; dynamic tracking	Limited penetration; often destructive; unsuitable for thick living constructs
X-ray / micro-CT	Ionizing radiation attenuation	~1–50 μm (micro-CT)	Minutes to hours per scan	Full Sample Size (high energy)	Low	High-resolution 3D imaging; mineral/tissue contrast	Ionizing; poor soft-tissue contrast
Ultrasound	Acoustic wave reflection (impedance mismatch)	~50–500 μm (frequency-dependent)	Seconds	Several cm	Moderate	Real-time, noninvasive; good for mechanical/elastic properties	Limited soft-tissue detail; operator dependence
PET / SPECT	Radiotracer-based gamma photon detection	~1–6 mm (PET), ~8–12 mm (SPECT)	Second to minutes	Full Sample Size	Very high	Molecular imaging; tracks metabolism, cell fate	Ionizing; poor resolution; requires tracers
MRI	RF signal from spin relaxation	~20–500 μm (Magnetic Field Strength Dependent)	Minutes to hours per scan	Full Sample Size	Low–moderate	Excellent soft-tissue contrast; structural + functional data	Expensive; slow; magnetic compatibility constraints

Table 1 : Comparison of Imaging Modalities for 3D Tissue-Engineered Constructs : Signal Types, Resolution, Penetration, Sensitivity, and Practical Trade-offs ([13], [29], [34], [35]).

II. Magnetic Resonance Imaging as Versatile Characterization Tool

II.A. Overview of Classical MRI Scanners

Magnetic Resonance Imaging (MRI) is a non-invasive imaging technique based on the principle of nuclear magnetic resonance. It exploits the behavior of nuclear spins in the presence of a strong magnetic field and radiofrequency (RF) excitation. When placed in a magnetic field, these nuclei align with the field, and after excitation with an RF pulse, they emit signals as they relax back to equilibrium. These signals are detected and processed to generate high-resolution images of anatomical structures or materials.

Standard clinical MRI systems operate at high magnetic field strengths, most commonly in the range of 1.5 to 3.0 tesla [36]. These systems employ superconducting magnets cooled with liquid helium, and they require extensive infrastructure, including shielded rooms, cryogenic support, powerful gradient and RF subsystems. Classically, as presented in Figure 4, the core components of a conventional MRI scanner include:

- **Main Magnet:** Provides the strong, static magnetic field necessary to align nuclear spins.
- **Gradient Coils:** Generate spatially varying magnetic fields to encode spatial information into the MR signal.
- **Radiofrequency Coils:** Transmit RF pulses to excite nuclei and receive the emitted MR signals.

- **Computer System:** Orchestrates data acquisition and reconstruction into interpretable images.

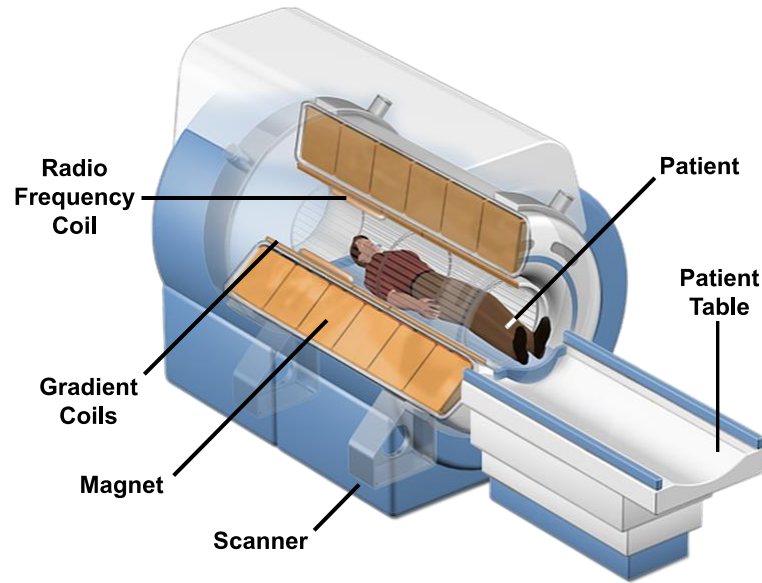


Figure 3 : Classical clinical MRI System and its components (adapted from [37])

Scanners can be classified by size (from small – Preclinical, to large – Clinical) and by static magnetic field strengths. Figure 4 shows a color-coded classification of MRI scanners depending on field strength: from ultra-low field (<0.1 T, peach), low field (0.1 – 1 T, green), high field (1 – 7 T, blue), to ultra-high field (7 – 11.7 T, purple) on a logarithmic scale.

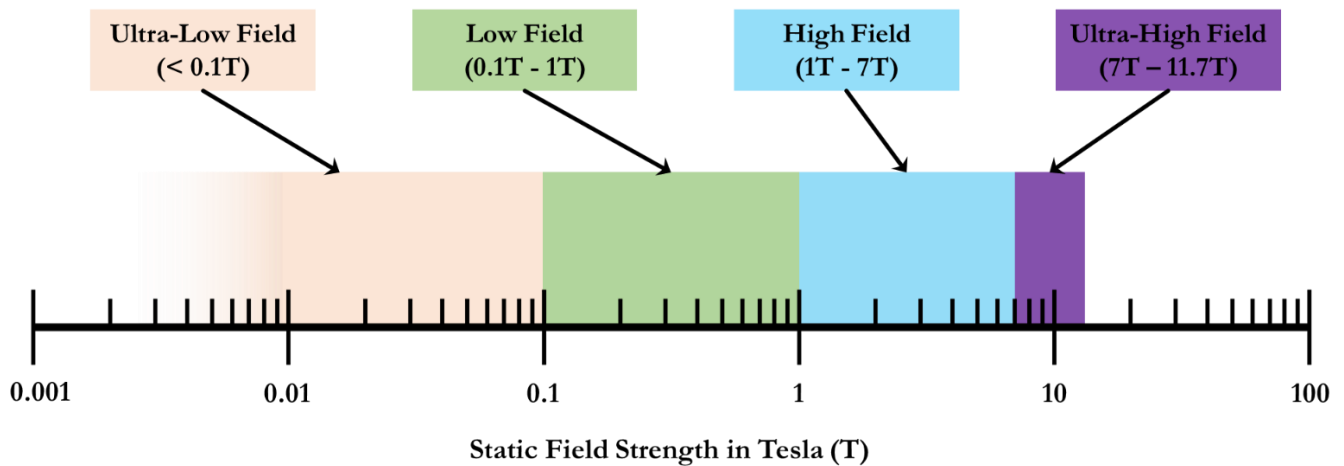


Figure 4 : Color-coded classification of MRI scanners by magnetic field strengths.

It will be explained further in this chapter how the static field directly relates to the scanner sensitivity through the available magnetization amplitude.

The performance of an MRI system is largely determined by its magnetic field strength, which affects signal-to-noise ratio, spatial resolution, and acquisition speed. Therefore, there has long been a concerted effort within the scientific and engineering communities to develop systems capable of operating at increasingly higher field strengths [38]. This is due to the exceptional image quality provided by High-Field (HF) and Ultra-High Field (UHF) systems. However, these high-field systems come with significant cost, size, and operational constraints. Such systems often

require specialized facilities and trained personnel for safe and effective use. State-of-the-art high-field units can exceed several million euros.

Despite their capabilities, high-field MRI systems are not practical in all settings. In recent years, there has been increasing interest in alternative MRI configurations that prioritize portability, affordability, and ease of installation. This has led to a resurgence in the development and deployment of low-field MRI systems, which operate at significantly reduced magnetic field strengths [24].

II.B. Fundamental Principles: From Nuclear Spins to Images

1. Origin of the MR Signal

1.1. Nuclear Spin

The principles of Nuclear Magnetic Resonance (NMR) are deeply rooted in quantum mechanics and the intrinsic properties of subatomic particles. In quantum mechanics, the “spin” refers to a fundamental property of particles akin to an intrinsic angular momentum. However, unlike classical angular momentum, which comes from an object's rotation in space, spin is an intrinsic property of particles.

The total angular momentum (**J**) of a nucleon can be expressed as the vector sum of the orbital and spin angular momenta:

$$\mathbf{J} = \mathbf{L} + \mathbf{S}, \quad (\text{I.1})$$

with the orbital angular momentum (**L**) resulting from the motion of particles in its orbit and the spin angular momentum (**S**) the intrinsic property of the particle.

However, protons and neutrons are tightly bound in the nucleus, and their orbital motion is strongly suppressed ("quenched") due to the nuclear shell structure [39]. This means the orbital contribution to the total angular momentum is negligible compared to the intrinsic spin angular momentum of the nucleus. Therefore, in the context of NMR, we can consider that the total angular momentum is almost entirely determined by the spin properties.

$$\mathbf{J} \approx \mathbf{S} \quad (\text{I.2})$$

To visualize spin (Figure 6), one may consider nucleons as small spheres with a certain mass (*m*) and a charge (*q*) rotating at high speed around their axis and generating an angular momentum. The rotating protons possessing a positive distributed charge and neutrons possessing non-uniformly distributed positive and negative charges circulating around their axis generate small magnetic fields called “magnetic moments.”

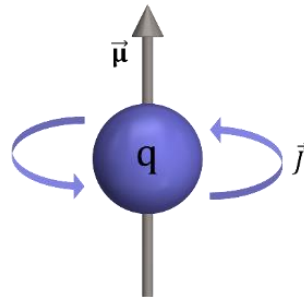


Figure 5 : Spin representation (inspired by [40])

For a given configuration of nuclei, the total magnetic moment ($\vec{\mu}$) generated is proportional to the total angular momentum by a factor γ known as the gyromagnetic ratio[41].

$$\vec{\mu} = \gamma \cdot \vec{J} \quad (I.3)$$

$$\gamma = \frac{q}{2m} \quad (I.4)$$

Spin is quantized, meaning it can only take on specific values defined by the spin quantum number (s). The spin quantum number determines the number of possible spin states a particle can occupy. For a given nucleus, the total number of spin states m_s , referred to as the spin magnetic quantum number, and the magnitude of the total angular momentum $|\vec{J}|$ are given by:

$$m_s = 2s + 1 \quad (I.5)$$

$$|\vec{J}| = \sqrt{s(s+1)}\hbar \quad (I.6)$$

where $\hbar$ is the reduced Planck's constant.

The spin quantum number is critical in defining whether a nucleus exhibits a magnetic moment. When we consider the Pauli exclusion principle and a pair of matching nucleons, to maintain the most stable arrangement, each individual subatomic particle must assume opposite spin configurations, producing a zero net angular momentum [42]. This means that nuclei possessing an even number of protons and/or an even number of neutrons will generate a zero net magnetic moment. On the other hand, for nuclei possessing an odd number of protons and/or an odd number of neutrons, it is impossible to arrange the spins to produce a zero net magnetic moment.

For example: - Particles with $I = 0$ have no spin and are magnetically inactive. - Particles with $I = 1/2$ (e.g., protons in hydrogen) have two spin states, often referred to as “spin-up” and “spin-down.” - Particles with $I = 1$ or higher possess more complex spin configurations with additional magnetic states (Table 2). Such nuclei are considered to have “nuclear spin” and include ^1H , ^{13}C , ^{17}O , ^{19}F , ^{23}Na , and ^{31}P , which can be used in NMR spectroscopy and imaging.

Nucleus	^1H	^{13}C	^{17}O	^{19}F	^{23}Na	^{31}P
I	1/2	1/2	5/2	1/2	3/2	1/2
$\frac{\gamma}{2\pi}$ (MHz/T)	42.58	10.71	-5.77	40.07	11.27	17.2

Table 2 : Gyromagnetic ratio of some commonly used nuclei in NMR and MRI.

1.2. Behavior Under an External Magnetic Field

When atoms possessing a non-zero magnetic moment are in a condensed matter state and subjected to an external static magnetic field ($\vec{B}_0$) (conventionally along the z-axis), the magnetic moments of their nuclei assume specific quantum equilibrium states determined by the spin quantum number.

Firstly, let's consider the effect of the external magnetic field on the motion of the nuclear magnetic moment at equilibrium. The forces created by the static magnetic field generate a torque acting on the magnetic moment.

$$\boldsymbol{\tau} = \boldsymbol{\mu} \times \mathbf{B}_0 \quad (\text{I.7})$$

$$\boldsymbol{\tau} = \frac{d\mathbf{J}}{dt} \quad (\text{I.8})$$

Combining these torque equations with the linear relation between the magnetic moment ($\boldsymbol{\mu}$) and the angular momentum ($\mathbf{J}$) gives a first-order differential equation governing the time evolution of the magnetic moment under a constant magnetic field $\vec{B}_0$.

$$\frac{d\boldsymbol{\mu}}{dt} = \gamma(\boldsymbol{\mu} \times \mathbf{B}_0) \quad (\text{I.9})$$

Assuming that $\mathbf{B}_0$ is constant and oriented along the z-axis ($\mathbf{B}_0 = B_0 \mathbf{e}_z$), the equation simplifies to:

$$\begin{cases} \frac{d\mu_x}{dt} = \gamma B_0 \mu_y, \\ \frac{d\mu_y}{dt} = -\gamma B_0 \mu_x, \\ \frac{d\mu_z}{dt} = 0 \end{cases} \quad (\text{I.10})$$

with the x- and y-components of the magnetic moment describing a simple harmonic motion and the z-component remaining constant. Solving these coupled equations yields a time-dependent solution for the magnetic moment dynamics under a constant magnetic field.

$$\boldsymbol{\mu}(t) = (\mu_{x,0}\cos(\omega t) + \mu_{y,0}\sin(\omega t))\mathbf{e}_x + (-\mu_{x,0}\sin(\omega t) + \mu_{y,0}\cos(\omega t))\mathbf{e}_y + \mu_{z,0}\mathbf{e}_z \quad (\text{I.11})$$

This solution (illustrated Figure 7) describes a precessional motion of the magnetic moment around the z-axis at a certain angular frequency of precession known as the Larmor frequency (ω_0), directly proportional to the strength of the static magnetic field:

$$\omega_0 = \gamma B_0 \quad (\text{I.12})$$

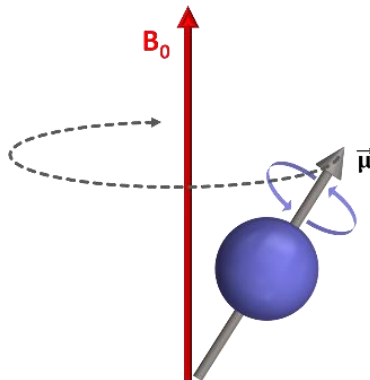


Figure 6 : Precessional motion of the magnetic moment around the $\mathbf{B}_0$ axis at a certain angular frequency $\omega_0 = \gamma B_0$ under external magnetic field (inspired from [40]).

This dynamic precession, defined by the Larmor frequency, not only describes the time evolution of the magnetic moment but also connects directly to the energy state of the system. The corresponding magnetic potential energy E_p of this equilibrium state (Zeeman Splitting) can be given by:

$$E_p = -\boldsymbol{\mu} \cdot \mathbf{B}_0 = -\gamma m_s \hbar B_0 \quad (\text{I.13})$$

Where $\hbar$ is the reduced Planck constant. The component of the magnetic moment along the field axis can be given by:

$$\mu_z = \gamma m_s \hbar \quad (\text{I.14})$$

For $I = 1/2$ nuclei, only two states exist. They correspond to either an alignment with the external field (“parallel”) or an opposition (“anti-parallel”), both associated respectively with lower (E^-) and higher (E^+) energy levels:

$$\begin{cases} E^+ = \frac{1}{2} \hbar \gamma B_0 \\ E^- = -\frac{1}{2} \hbar \gamma B_0 \end{cases} \quad (\text{I.15})$$

$$\Delta E = \hbar \gamma B_0 \quad (\text{I.16})$$

with a gap in energy levels ΔE , for a given gyromagnetic ratio γ , increasing proportionally with the static field strength, as shown Figure 8.

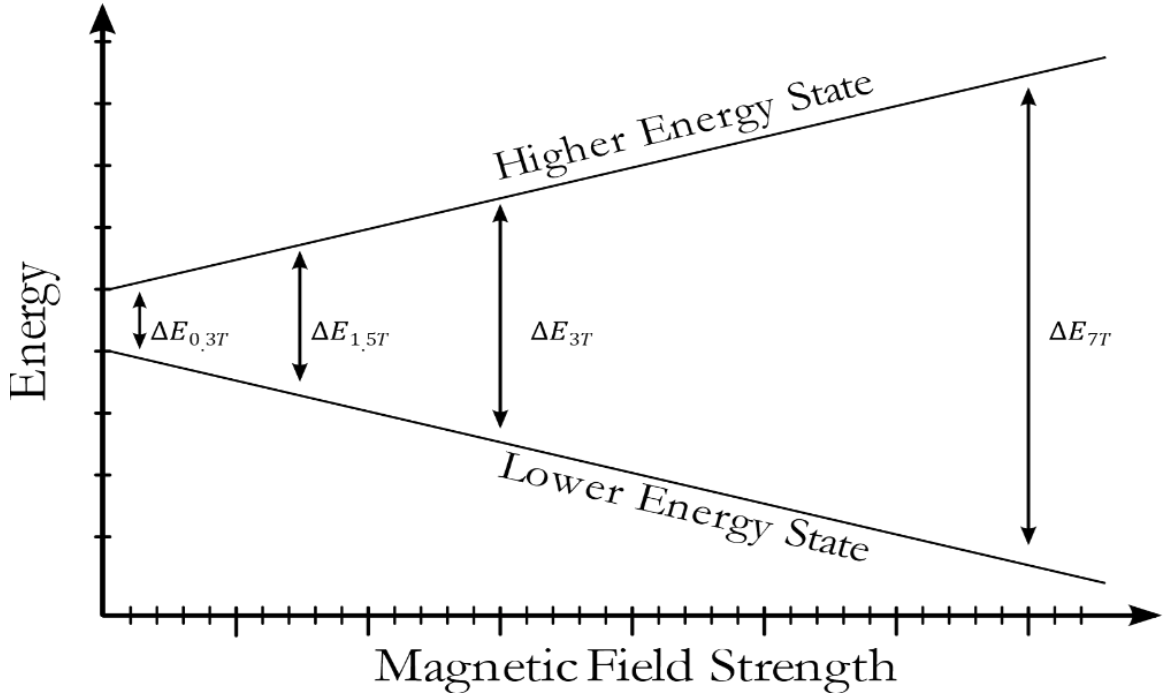


Figure 7 : Energy levels ΔE , as a function of magnetic field strength (inspired from [40])

Although both states are considered stable, the parallel orientation is statistically preferential due to the lower energy associated.

1.3. Bulk magnetization

At thermal equilibrium, the ratio of population between higher (N^+) and lower (N^-) states follows the Boltzmann distribution:

$$\frac{N^+}{N^-} = e^{-\frac{\Delta E}{k_B T}} \quad (I.17)$$

Where k_B is the Boltzmann constant, T is the absolute temperature.

Considering the bulk magnetization of a pure water sample ^1H protons at 300 K and static field strengths of 0.1T and 7T we respectively get repartitions of population (N^+/N^-)_{0.1T} $\approx$ 0.99999932 and (N^+/N^-)_{7T} $\approx$ 0.999952, with a corresponding spin polarization $P = \frac{N^- - N^+}{N^- + N^+} \approx \frac{\Delta E}{2k_B T}$, $P(0.1 \text{ T}) \approx 3.4 \times 10^{-7}$ (~ 0.34 ppm) and $P(7 \text{ T}) \approx 2.4 \times 10^{-5}$ (~ 24 ppm). This means not only that a preferential orientation exists along the direction of the magnetic field, but also that this alignment becomes increasingly pronounced as the strength of the magnetic field increases. As a result, the collective effect of the magnetic moments produces a macroscopic nuclear magnetization ($\mathbf{M}_0$) at equilibrium, aligned along the ($\mathbf{B}_0$) axis and proportional to the strength of the static field. Figure 8 illustrates this phenomenon.

At lower field strengths, the extremely small population difference between spin states leads to a markedly reduced net magnetization. This reduced polarization will have direct implications for signal strength, resulting in weaker signals and making measurements inherently more challenging.

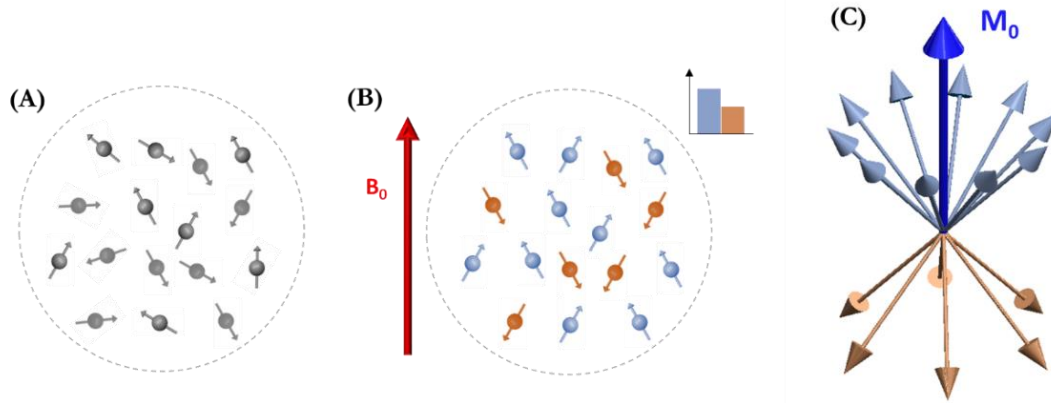


Figure 8 : Bulk Magnetization and Spin Alignment in a Static Magnetic Field : (A) In the absence of an external magnetic field, nuclear spins are randomly oriented, resulting in no net macroscopic magnetization. (B) When subjected to a static magnetic field $\mathbf{B}_0$, nuclear spins align preferentially either parallel (in blue) or antiparallel (in orange) to the field direction, with a greater population in parallel orientation. (C) The vector sum of all individual magnetic moments yields a net macroscopic magnetization $\mathbf{M}_0$, aligned with the direction of $\mathbf{B}_0$.

Among the many nuclei possessing nuclear spin, hydrogen (^1H) imaging is the most commonly used primarily due to its abundance in biological tissues and significant magnetic moment. Despite this, the overall magnetization remains very weak; for instance, the magnetization of pure water at 300 K in a magnetic field of 1 Tesla is approximately $3.2 \times 10^{-3} \text{ A} \cdot \text{m}^{-1}$. Because of this, under typical NMR conditions, the static magnetization cannot be directly measured due to its low amplitude compared to that of the main magnetic field ($B_0 \approx 0.8 \times 10^6 \text{ A} \cdot \text{m}^{-1}$). To enable detection, an oscillating radiofrequency (RF) field ($\mathbf{B}_1$) is required to perturb the magnetization vector away from equilibrium and enable measurement.

2. RF Coils for Signal Detection

RF coils (or resonators) are the vital components used for signal detection in magnetic resonance experiments. An RF resonator is a structure designed to sustain oscillating electromagnetic fields at a particular resonant frequency, determined by its physical and electrical properties. It typically consists of conductive elements arranged to provide the required inductance and capacitance, allowing the structure to resonate and generate a radiofrequency field at its designated frequency. They serve two key roles: excitation through a transmitted oscillating RF magnetic field ($\mathbf{B}_1^+$) and signal reception through a perceived field ($\mathbf{B}_1^-$). When a current $I(t)$ flows through the coil, it produces a linearly polarized magnetic field:

$$\frac{\mathbf{B}_1(\mathbf{r}, t)}{I} = \underbrace{\frac{1}{2}(\mathbf{B}_{1x} + i\mathbf{B}_{1y})}_{\mathbf{B}_1^+} e^{-i\omega_0 t} + \underbrace{\frac{1}{2}(\mathbf{B}_{1x} - i\mathbf{B}_{1y})}_{\mathbf{B}_1^-} e^{+i\omega_0 t}. \quad (\text{I.18})$$

This means that when transmitting, only the co-rotating component $\mathbf{B}_1^+$ “locks” onto the spins’ Larmor motion as it appears static in the rotating frame spinning at ω_0 while the counter-rotating part $\mathbf{B}_1^-$ oscillates at twice the Larmor rate in that frame and averages to zero. On reception, however, the precessing spins generate an oscillating transverse magnetization which induces voltage in the coil in exactly the ω_0 sense. That induced field pattern matches the coil’s $\mathbf{B}_1^-$ sensitivity profile, so only $\mathbf{B}_1^-$ effectively “picks up” any signal [43].

2.1. Excitation and Relaxation Dynamics

During the transmission (Tx) or excitation, the perturbation of the nuclear magnetization vector ($\mathbf{M}$) is possible due to the resonant interaction between a transmitted oscillating radiofrequency field ($\mathbf{B}_1^+$) and the nuclear spins. This interaction occurs when the RF frequency matches the Larmor frequency of the nuclei, enabling energy absorption that transitions the spins to higher energy states. The RF field disrupts the equilibrium magnetization along ($\mathbf{B}_0$), causing it to deviate towards the direction of ($\mathbf{B}_1^+$), with ($\mathbf{B}_1^+$) perpendicular to ($\mathbf{B}_0$). The magnitude of this deviation is quantified by the flip angle (FA) θ , the angle through which the net magnetization is “tipped” away from the static ($\mathbf{B}_0$) axis.

In the idealized case of a spatially uniform ($\mathbf{B}_1^+$) field across the sample volume and an RF pulse duration τ , the flip angle is given by the pulse area integral of the RF field:

$$\theta = \gamma \int_0^\tau \mathbf{B}_1^+(t) dt \approx \gamma \mathbf{B}_1^+ \tau \quad (\text{I.19})$$

The synchronized response of the nuclear spins to the RF field is a direct result of their quantized energy states and the coupling between their magnetic moments and the applied field. The perturbation alters the spin states of the nuclei, creating observable transitions.

The Bloch phenomenological equations provide an approximate description of the dynamic behavior of magnetization out of equilibrium. They describe the dynamics of the magnetization vector ($\mathbf{M}$) (with components M_x , M_y , M_z) in a magnetic field. They account for both the precession of magnetization around the magnetic field and the relaxation effects.

$$\begin{cases} \frac{dM_x}{dt} = \gamma(M_y B_z - M_z B_y) - \frac{M_x}{T_2} \\ \frac{dM_y}{dt} = \gamma(M_z B_x - M_x B_z) - \frac{M_y}{T_2} \\ \frac{dM_z}{dt} = \gamma(M_x B_y - M_y B_x) - \frac{(M_z - M_0)}{T_1} \end{cases} \quad (I.20)$$

Relaxation dynamics are governed by two main mechanisms: longitudinal relaxation (T_1), which describes the recovery of magnetization along the ($\mathbf{B}_0$) axis, and transverse relaxation (T_2), which characterizes the decay of magnetization perpendicular to ($\mathbf{B}_0$). These processes produce oscillatory changes in the transverse component of magnetization, which an RF coil can detect. Figure 9 summarizes the dynamic effect of the excitation using a 90° RF pulse to tilt the magnetization vector and subsequent the relaxation phenomenon: (A) At equilibrium, the net magnetization is fully aligned along the z-axis, representing the maximum longitudinal magnetization $\mathbf{M} = M_0 \mathbf{e}_z$ with no transverse components. (B) Immediately after the RF pulse, the magnetization is tipped into the transverse plane (setting the stage for observable signal generation. (C) The recovery trajectory illustrates the relaxation process: the longitudinal (z) component recovers exponentially toward $\mathbf{M} = M_0 \mathbf{e}_z$ (governed by T_1), while the transverse (x and y) components decay and precess (governed by T_2^*), as the system returns to equilibrium.

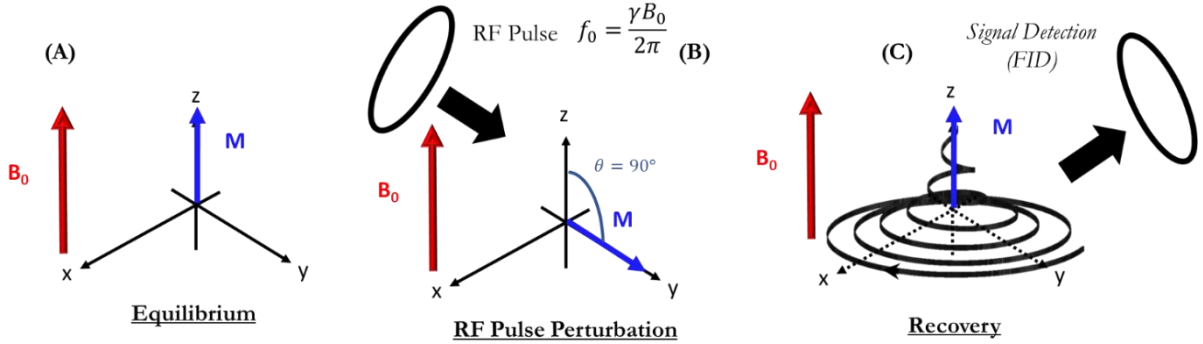


Figure 9 : Schematic example of the relaxation dynamics for a flip angle $\theta=90^\circ$.

The solutions to the Bloch equations in a constant magnetic field $\mathbf{B} = (0,0,B_0)$ are given by:

$$M_z(t) = M_0 + (M_z(0) - M_0)e^{-t/T_1}, \quad (I.21)$$

$$M_x(t) = e^{-t/T_2} [M_x(0) \cos(\gamma B_0 t) - M_y(0) \sin(\gamma B_0 t)], \quad (I.22)$$

$$M_y(t) = e^{-t/T_2} [M_x(0) \sin(\gamma B_0 t) + M_y(0) \cos(\gamma B_0 t)]. \quad (I.23)$$

Following the solutions to the Bloch equations, the behavior of the magnetization vector immediately after the RF pulse creates what is known as the free induction decay (FID).

2.2. The Free Induction Decay (FID)

As the RF pulse is turned off, the transverse magnetization components M_x and M_y , now aligned in the xy-plane, begin to precess at the Larmor frequency. However, the amplitude of this transverse magnetization component does not remain constant. Instead, it decays exponentially

over time. The observed FID is illustrated schematically (Figure 10.A), where the amplitude of the signal is seen to diminish rapidly following the RF pulse.

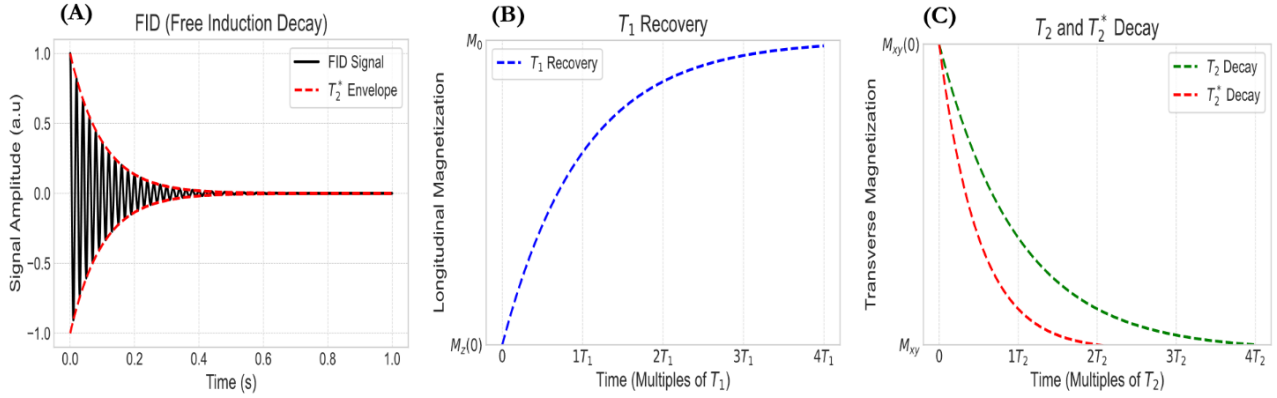


Figure 10: Overview of Magnetization Dynamics after RF pulse excitation. (A) The free induction decay (FID) curve shows the immediate post-RF pulse signal: an oscillatory waveform at the Larmor frequency with an exponentially decaying envelope governed by T_2^* . (B) The T_1 recovery plot illustrates the longitudinal magnetization M_z returning to equilibrium M_0 via spin-lattice relaxation, with approximately 63 % recovery after one T_1 period. (C) A comparison of T_2 and T_2^* decay curves are presented, where the green dashed line represents intrinsic T_2 decay due to dephasing from spin-spin interactions, and the red dashed line shows the faster decay of T_2^* which includes additional dephasing from field inhomogeneities.

This decay, observed as the FID, is governed by an effective time constant termed T_2 (or spin-spin relaxation time), as described by equations (I.22) and (I.23). While the intrinsic decay due to spin-spin interactions is characterized by T_2 , the actual observed FID decay T_2^* is accelerated by additional dephasing effects caused by magnetic field inhomogeneities, imperfections in the magnet, and susceptibility variations within the sample. Consequently, T_2^* is typically shorter than T_2 , resulting in a quicker decay (Figure 10.C). In contrast, the longitudinal component of the magnetization recovers according to T_1 relaxation (Figure 10.B), as described by equation (I.21). T_1 (or spin-lattice relaxation time), reflects the process by which the perturbed spins return to thermal equilibrium by exchanging energy with their surroundings. The recovery of the longitudinal magnetization is an exponential process, where approximately 63% of the equilibrium magnetization is restored after one T_1 period.

The free induction decay encapsulates the combined effects of both intrinsic spin-spin interactions and extrinsic field inhomogeneities. By analyzing the FID and its decay constant, it is possible to gain insights into the microstructural environment of the sample. Understanding the distinct roles of T_1 , T_2 , and T_2^* is essential for optimizing MRI protocols to achieve desired contrast and resolution through sequences. But first, the signal and subsequent FID information must be acquired.

2.2.1 Electromagnetic Detection Principle

Nuclear magnetization is a source of macroscopic electric and magnetic fields and therefore obeys Maxwell's equations. These equations encapsulate the behavior of electric and magnetic fields and their interactions with matter. They collectively describe how charges and currents generate and modify electric and magnetic fields.

Maxwell's equations can be written as [44]:

Gauss's Law for Electricity:

$$\nabla \cdot \mathbf{E} = \frac{\rho}{\varepsilon} \quad (I.24)$$

States that the divergence of the electric field equals the local charge density divided by the vacuum permittivity, expressing how charges act as sources of $\mathbf{E}$. In non-charged medium, $\nabla \cdot \mathbf{E} = 0$.

Gauss's Law for Magnetism:

$$\nabla \cdot \mathbf{B} = 0 \quad (I.25)$$

Implies there are no magnetic monopoles; magnetic field lines are continuous loops with no beginning or end. It implies implicitly that there is a conservation of magnetic flux through a closed surface.

Faraday's Law of Induction:

$$\nabla \times \mathbf{E} = -\frac{\partial \mathbf{B}}{\partial t} \quad (I.26)$$

Describes how a time-varying magnetic field induces a circulating electric field.

Ampère's Law with Maxwell's Correction:

$$\nabla \times \mathbf{B} = \mu \mathbf{J} + \mu \varepsilon \frac{\partial \mathbf{E}}{\partial t} \quad (I.27)$$

Relates the curl of the magnetic field to conduction currents and to changing electric fields (displacement current), ensuring continuity and enabling electromagnetic wave propagation, while maintaining charge conservation in dynamic fields and predicting the finite speed of electromagnetic interactions.

Here $\mathbf{E}$ is the electric field, $\mathbf{B}$ the (induction) magnetic field, ρ the electric charge density, $\mathbf{J}$ the current density, $\mu = \mu_r \mu_0$ the magnetic permeability and $\varepsilon = \varepsilon_r \varepsilon_0$ the electric permittivity of the medium.

By integrating the differential form of Faraday's law of induction over a surface, we obtain:

$$\iint_S (\nabla \times \mathbf{E}) \cdot d\mathbf{S} = - \iint_S \frac{d\mathbf{B}}{dt} \cdot d\mathbf{S} \quad (I.28)$$

And, using Stokes' theorem:

$$\oint_l \mathbf{E} \cdot d\mathbf{l} = - \iint_S \frac{d\mathbf{B}}{dt} \cdot d\mathbf{S} \quad (I.29)$$

$$e = - \frac{d\Phi}{dt} \quad (I.30)$$

where $\Phi = \iint_S \mathbf{B} \cdot d\mathbf{S}$ is the magnetic flux through the surface.

This is Faraday's law of induction: a time-varying magnetic flux through a loop induces an electromotive force e around that loop. In NMR, the precession motion obtained after RF

perturbation of the resulting magnetization vector creates a time-varying magnetic field that permeates the region surrounding the sample. Therefore, the oscillatory signal of the FID is detectable using this very law of induction.

2.3. Reciprocity Principle

In its general form, the Lorentz reciprocity theorem states that in any linear, passive, and time-invariant electromagnetic system, the interaction between two sets of sources and fields is symmetric: if a current source ($\mathbf{J}_1$) induces a voltage at the location of a second current ($\mathbf{J}_2$), then the situation is identical when the roles are reversed.

Mathematically, this is expressed as:

$$\int_V \mathbf{E}_1 \cdot \mathbf{J}_2 dV = \int_V \mathbf{E}_2 \cdot \mathbf{J}_1 dV \quad (\text{I.31})$$

In the context of NMR, this symmetry appears between the RF coil as a transmitter and the RF coil as a receiver. If a current I flows through the coil, generating a magnetic field $\mathbf{B}_1^+(\mathbf{r})$ throughout space and the same field pattern determines the sensitivity of the coil to magnetic sources during reception through $\mathbf{B}_1^-(\mathbf{r})$.

In short, the reciprocity principle tells us that the way energy is deposited into the system is exactly how it will be extracted. If a coil is efficient at exciting spins in a certain region, it will be equally efficient at detecting signals from that region. This has far-reaching implications:

- Coil Design: Optimization of the coil for homogeneous RF delivery also enhances signal reception. Coil geometry and performance affect both transmit and receive performance equally.
- Calibration: Since $\mathbf{B}_1^+$ (transmit) and $\mathbf{B}_1^-$ (receive) patterns are conjugate pairs in many practical systems, mapping one informs the other.

Reciprocity bridges transmission and reception in NMR. The magnetic field produced by a coil during excitation precisely mirrors the sensitivity it exhibits during detection. This symmetry, grounded in Maxwell's equations and formalized by the Lorentz reciprocity theorem, underlies the entire architecture of NMR hardware and enables efficient, reliable signal detection. In essence, the reciprocity principle ensures that the very same field pattern that perturbs the spin system also determines how well the resulting signal is detected.

Furthermore, thanks to the reciprocity principle, a more intuitive expression of the detectable signal can be directly defined as a relation between the $\mathbf{B}_1^-(\mathbf{r})$ field and transverse component of magnetization $\mathbf{M}_\perp(\mathbf{r}, t)$. According to the reciprocity principle, the voltage $e(t)$ induced in the coil can be written as (proof in [45]):

$$e(t) = -\frac{d}{dt} \int_V \mathbf{B}_1^-(\mathbf{r}) \cdot \mathbf{M}(\mathbf{r}, t) d^3\mathbf{r} \quad (\text{I.32})$$

Assuming harmonic precession at frequency ω_0 , this becomes:

$$e(t) = - \int_V \mathbf{B}_1^-(\mathbf{r}) \cdot \frac{\partial \mathbf{M}(\mathbf{r}, t)}{\partial t} d^3\mathbf{r} \approx i \omega_0 \int_V \mathbf{B}_1^-(\mathbf{r}) \cdot \mathbf{M}_\perp(\mathbf{r}) e^{-i\omega_0 t} d^3\mathbf{r} \quad (\text{I.33})$$

This enables a direct expression of the signal voltage detectable by the coil. Over a small voxel of volume δV if both $\mathbf{B}_1^-(\mathbf{r})$ and $\mathbf{M}_\perp$ are treated as constants with $\mathbf{B}_1^-(\mathbf{r}) = B_1/I$, the peak induced signal voltage per unit current becomes:

$$V_{\text{signal}} = |e(t)| = \omega_0 \frac{B_1}{I} M_{\perp} \delta V. \quad (\text{I.34})$$

3. Creating an Image: Spatial Encoding and Reconstruction

Following the NMR signal detection, it is imperative to address the methodologies used to localize the signals origin in space allowing for the reconstruction of images. In contrast to classical NMR which focuses on providing detailed information on a sample's chemical and molecular structure, magnetic resonance imaging (MRI) aims to produce detailed images of either morphology or function. The techniques used to produce detailed images hinge critically upon “spatial encoding” which is a process where unique spatial coordinates are assigned to nuclear spins, followed by the conversion of encoded signals into interpretable imagery reconstruction algorithms. In this section, spatial encoding strategies, specifically frequency and phase encoding, as well as the subsequent computational methods employed for image reconstruction, will be discussed.

3.1. Spatial Encoding Techniques

In MRI, spatial encoding relies on the controlled application of linear magnetic field gradients superimposed upon the homogeneous main magnetic field ($\mathbf{B}_0$). These gradients, generally generated using specialized gradient coils, cause the Larmor frequency of nuclear spins to vary depending on their location. This means that each position in space is linked to a specific signal frequency and phase.

3.1.1 Frequency Encoding

Frequency encoding, also called readout encoding, is achieved by applying a linear gradient field along one direction during the signal acquisition [46]. The introduction of a gradient field G_x , alters the magnetic field linearly in the selected direction:

$$B(x) = B_0 + G_x x \quad (\text{I.35})$$

This means that spins at different spatial locations resonate at different frequencies, depending on the new spatially dependent Larmor frequency:

$$\omega(x) = \gamma(B_0 + G_x x) \quad (\text{I.36})$$

The measured signal becomes an ensemble of frequencies, each directly mapping specific spatial coordinates along the gradient axis.

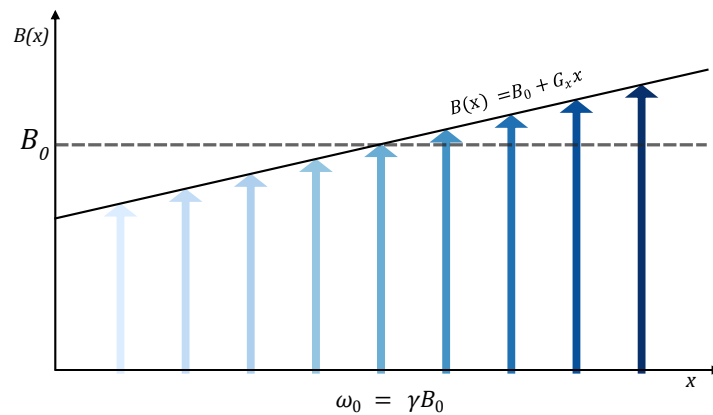


Figure 11 : Gradient base frequency encoding principle along the x-axis.

3.1.2 Phase Encoding

Phase encoding is an additional step used to create spatial encoding in another direction [47]. In this step, a gradient is briefly applied along a direction perpendicular to the frequency encoding axis, typically along the y-axis (G_y), before the acquisition. This short burst of gradient causes the spins to accumulate spatially-dependent phase shifts, expressed by:

$$\phi(y) = \gamma G_y y \tau \quad (I.37)$$

where τ is the duration of gradient application.

After gradient extinction, the spins at different positions along the y-axis keep their unique phase shifts. These phase differences help identify the location of the spins along that direction. By repeating the scan multiple times and changing the strength of the gradient in each cycle, a complete encoding along the desired spatial direction can be collected.

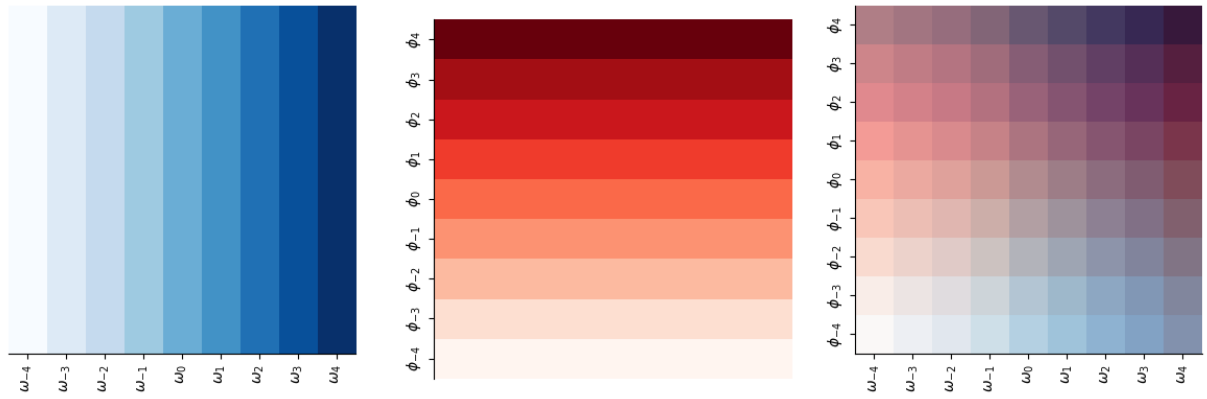


Figure 12 : A) frequency encoding along x-axis B) phase encoding along y-axis C) spatially encoded plane with phase and frequency encoding.

3.1.3 Slice Selection

In volumetric MRI, it's important to pinpoint where signals come from in all three directions. To achieve this, a technique called slice selection is used. It works by applying a radiofrequency pulse targeting a specific frequency range coupled with a gradient field (G_z) applied along the z-axis perpendicular to the imaging plane [47]. By carefully calibrating the bandwidth of the RF pulse and the gradient strength, a selective excitation occurs at a precise slice position according to:

$$z = \frac{\omega - \omega_0}{\gamma G_z} \quad (I.38)$$

Coupled with the frequency and phase encoding mentioned previously, it is possible to obtain a comprehensive spatial encoding for 3D MRI data acquisition.

3.2. Image Reconstruction

3.2.1 Raw Data and k-Space

Following the spatial encoding process, a reconstruction algorithm mathematically transforms the acquired raw signals, represented as complex-valued data points in the frequency-space domain (commonly referred to as k-space) into images. The relationship between image space $I(x, y)$ and k-space data $S(k_x, k_y)$ is described by the 2D inverse Fourier transform [47]:

$$I(x, y) = \iint_{-\infty}^{+\infty} S(k_x, k_y) e^{i2\pi(k_x x + k_y y)} dk_x dk_y \quad (\text{I.39})$$

In the k-space, each point corresponds to a specific frequency and phase combination acquired during the MRI acquisition. The coordinates in k-space are determined by the integrated gradient waveforms, mathematically described by:

$$k_x(t) = \frac{\gamma}{2\pi} \int_0^t G_x(t') dt', \quad k_y(t) = \frac{\gamma}{2\pi} \int_0^t G_y(t') dt' \quad (\text{I.40})$$

Therefore, the trajectory traversed within k-space is directly controlled by the application of gradient waveforms.

3.2.1 Fourier Reconstruction and Advanced Algorithms

The fundamental reconstruction method in MRI involves applying the inverse Fourier transform (IFT) to the acquired k-space data. While the classical Fourier approach is computationally efficient and analytically robust, certain advanced imaging protocols and non-Cartesian k-space trajectories such as radial or spiral sampling have been developed to improve acquisitions [48], [49]. These methods mostly aim to reduce scan time, improve spatial resolution, and accommodate irregular sampling patterns. Furthermore, modern reconstruction techniques like compressed sensing [50] and iterative algorithms [51] have enabled high-quality image reconstruction from under-sampled data, pushing the boundaries of acquisition speed and image fidelity beyond what traditional Fourier-based methods can achieve.

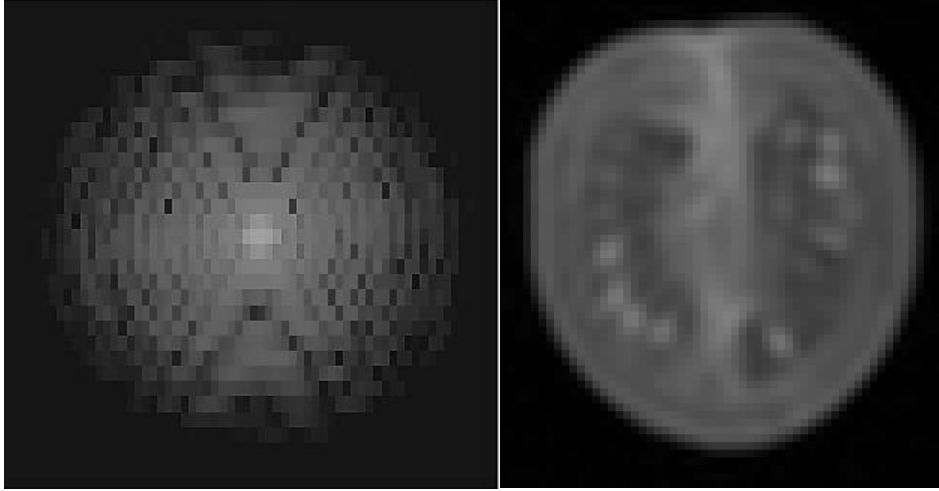


Figure 13 : Spin Echo example of a cherry tomato k-space data acquisition (On the left) and its corresponding image (On the right) after the inverse Fourier transform (IFT). The images were obtained at the AMPERE laboratory using 300mT benchtop MRI scanner from Pure Devices GmbH's Research Lab line (Pure Devices GmbH, 300 mT Research Lab, Rimpar, Germany).

II.C. Tailoring MRI for Complex Tissue Construct Assessment

Aside from the signal detection principles and image reconstruction techniques, a special focus should be put on the sequence architectures that govern contrast formation and functional data acquisition for both structural and functional imaging, as these can be tailored specifically for tissue engineering applications.

An MRI sequence consists of a temporal succession of events to generate a usable signal and always includes at least one RF pulse, spatial-encoding gradients, and a readout phase for signal reception. A distinctive advantage of MRI is the ability to tune the desired contrast by adjusting sequence parameters related to signal excitation and reception, which enables the ability to emphasize specific sample characteristics linked to longitudinal relaxation time (T_1), transverse relaxation time (T_2), or proton density (PD). The pulse sequences play a pivotal role in determining the “weighting” of the image and the sensitivity to specific sample properties.

For example: - A sample with a long T_1 and T_2 (like water) appears dark (low signal) on a T_1 weighted scan but bright (high signal) on a T_2 weighted scan – A sample with short T_1 and short T_2 (such as fat or oils) shows up bright on T_1 -weighted images and intermediate-gray to dark on T_2 -weighted scans.

Two key timing parameters in every sequence are the repetition time (TR) and the echo time (TE):

- TR is the time between two successive RF excitation pulses. It affects how much the longitudinal magnetization has time to recover before the next pulse and plays a key role in determining T_1 contrast in the image.
- TE is the time between the RF pulse and the moment the MRI signal (the echo) is measured. It reflects how much transverse relaxation (or T_2 decay) has happened, which means that it controls the T_2 contrast in the image.

Together, TR and TE help define the type of contrast seen in MRI images by balancing how media with different relaxation properties appear. A long TR and short TE sequence is usually called Proton density-weighted, a short TR and short TE sequence is usually called T_1 -weighted, while a long TR and long TE sequence is usually called T_2 -weighted.

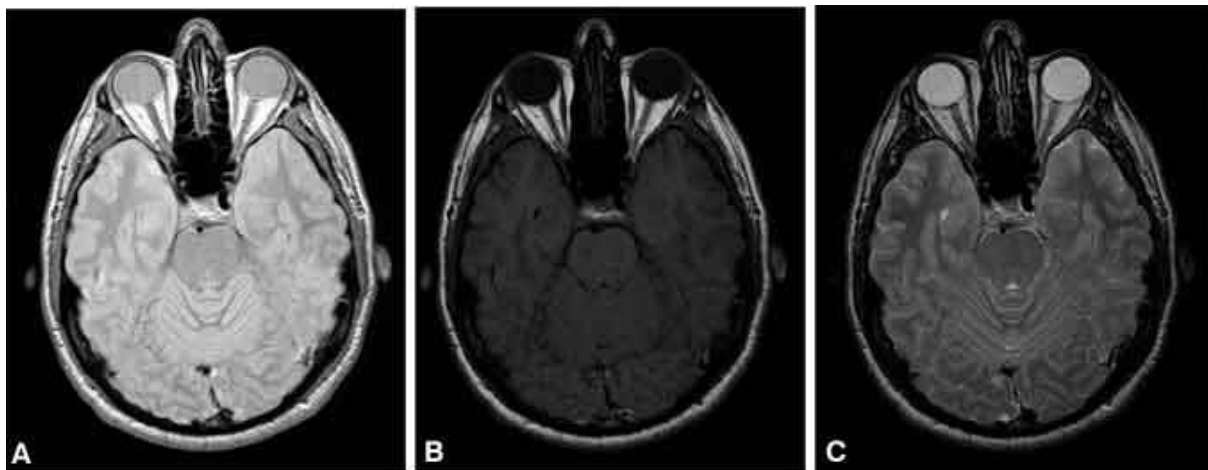


Figure 14 : Example of axial MR Images of a human head taken with (A) proton density weighting: a long TR and short TE, (B) T_1 weighting: a reduced TR and short TE, and (C) T_2 weighting: a long TR and increased TE. (Origin [52])

MRI sequences can be used to obtain both structural information through contrast manipulation, but also functional information through more specialized sequences designed to sensitize the signal to additional sample characteristics. As MRI sequence development is a broad

and complex topic, here, we will only briefly present a few useful sequences for tissue engineering applications.

1. Standard Pulse Sequences for Morphological and Structural Assessment

Structural MRI sequences form the backbone of anatomical/morphological imaging by leveraging different relaxation properties to generate contrast. In this subsection, spin-echo (SE) methods and gradient-echo (GRE) techniques are briefly covered, as together, these two sequence families establish the fundamental tools for visualizing tissue structure and set the stage for the more advanced functional and diffusion-based approaches that follow.

- **Spin Echo (SE) sequences** are foundational in MRI, primarily due to their effectiveness in compensating for magnetic field inhomogeneities and reducing artifacts and distortions [53]. These sequences typically involve two consecutive RF pulses: an initial 90° excitation pulse that tips the net magnetization into the transverse plane, followed by a 180° refocusing pulse applied at half the echo time ($TE/2$). The first pulse initiates dephasing among the spins due to spatial variations in the magnetic field, which may arise from imperfections in the instrumentation or internal inhomogeneities within the sample. The second pulse reverses the direction of phase dispersion, causing the spins to rephase and form a measurable echo at time TE . This rephasing leads to a recovery of the MR signal that would otherwise decay prematurely due to static field irregularities. The echo, observed at $TE = 2\tau$ (where τ is the time between the two pulses), reflects the transverse relaxation properties (T_2) of the sample, especially when TE is long enough to allow T_2 -weighted contrast to dominate.
- **Gradient Echo (GRE) sequences** are particularly valued for their rapid acquisition speed and sensitivity to magnetic susceptibility variations [53]. Unlike Spin Echo sequences, GRE sequences use a single RF excitation pulse (typically less than 90°) to tip the net magnetization into the transverse plane. Signal rephasing is then achieved through the application of gradient reversal rather than a 180° RF refocusing pulse. After the initial dephasing caused by spatially varying magnetic fields and applied gradients, a reversal of the readout gradient induces rephasing, producing an echo at the echo time (TE). Because GRE sequences do not employ a refocusing pulse, they are more susceptible to field inhomogeneities and T_2^* effects, making them particularly useful for applications where sensitivity to these variations is desired. The resulting echo reflects the effective transverse relaxation time (T_2^*), allowing GRE sequences to provide strong T_2^* -weighted contrast when TE is appropriately chosen.

Choosing the proper sequence tailored to the desired application is a crucial step in any MR experiment. As a rule of thumb, SE sequences should be used for high-quality T_2 -weighted morphological acquisitions with minimal artifacts and accurate T_2 quantification. In contrast, GRE sequences should be prioritized for speed, susceptibility effects-based acquisitions, and when distortions are tolerable.

2. Advanced MRI Sequences for Functional Assessment

Beyond structural imaging, MRI sequences can be tailored to quantitative functional assessments through mechanisms such as molecular diffusion and fluid flow by embedding motion-sensitive gradients into their pulse sequences. In this subsection, we introduce diffusion-weighted imaging (DWI), where paired gradients attenuate signal according to the apparent diffusion coefficient, and perfusion/velocimetry techniques that encode spin motion as a phase shift, enabling noninvasive flow quantification.

2.1. T1 and T2 Maps

While T_1 - and T_2 -weighted images provide excellent qualitative contrast, quantitative mapping of relaxation times offers a more powerful approach by generating images where each pixel's intensity directly represents the measured T_1 or T_2 value. These maps provide objective, reproducible biomarkers of tissue composition, independent of specific scanner parameters like repetition time (TR) or echo time (TE).

T_1 mapping is commonly performed using an inversion recovery sequence [54], where a series of images is acquired with different inversion times (TI). The signal intensity for each voxel is sampled at various points along the longitudinal relaxation curve. By fitting these data points to the T_1 recovery equation, a precise T_1 value can be calculated for each voxel, resulting in a quantitative T_1 map.

Similarly, T_2 mapping is typically achieved with a multi-echo spin-echo sequence [54]. This method acquires multiple echoes at different TEs within a single TR. The signal decay across these successive echoes is measured for each voxel and fitted to a mono-exponential decay function ($S(TE) = S_0 e^{-TE/T_2}$), from which the T_2 value is determined. The resulting T_2 map provides a quantitative assessment of transverse relaxation, which is sensitive to tissue water content and macromolecular environment. These quantitative maps are invaluable for characterizing subtle pathologies and for longitudinal studies tracking disease progression or treatment response.

2.2. Diffusion-Weighted Imaging

Diffusion-weighted imaging (DWI) exploits the sensitivity of the MR signal to the microscopic Brownian motion of water molecules within water-based samples [55]. This is achieved through the application of a pair of symmetric diffusion-sensitizing gradients flanking a 180° refocusing pulse. The initial gradient induces a spatially dependent phase shift in the spins, which is reversed by the second gradient in the absence of diffusion. However, in samples where water molecules move between the gradient applications, the phase reversal is incomplete, resulting in signal attenuation:

$$S = S_0 \cdot e^{-b \cdot ADC} \quad (I.41)$$

where S_0 is the signal without diffusion weighting, b is the diffusion weighting factor determined by the strength and duration of the gradient pulses, and ADC is the apparent diffusion coefficient. The degree of signal loss is governed by the b-value, a scalar representing the diffusion weighting. This means that the b-value influences the sensitivity of the measurement to diffusion, and by using different b-values, it is possible to quantify the diffusivity of water molecules in a medium. DWI provides contrast based on the apparent diffusion coefficient (ADC), enabling the detection of differences in the physical properties such as variations in structural composition, molecular organization, or the presence of barriers to diffusion. In practice, it can be used in tissue engineering applications to assess cellularity among other things [15], [56], [57].

2.3. Perfusion Imaging and Velocimetry

Perfusion MRI encompasses a range of techniques designed to assess flow dynamics and the distribution of fluid through different media. Among these, phase-contrast velocimetry uses velocity-encoded gradient pulses to quantify flow velocities. When spins in a sample are moving under the influence of a flow, they acquire an additional phase shift proportional to their velocity [58]. Under a gradient (G), a spin at position ($\mathbf{r}(t)$) processes a phase:

$$\phi(\tau) = \gamma \int_0^\tau G \cdot \mathbf{r}(t) dt, \quad (I.42)$$

when moving through the gradient field such that $\mathbf{r}(t) = \mathbf{r}_0 + \mathbf{v}_r t$, we get:

$$\phi(\tau) = \gamma G \int_0^\tau (\mathbf{r}_0 + \mathbf{v}_r t) dt = \gamma G \mathbf{r}_0 \tau + \frac{1}{2} \gamma G \mathbf{v}_r \tau^2. \quad (1.43)$$

where τ is the gradient application duration, $(\mathbf{r}_0)$ initial position vector, and $\mathbf{v}_r$ the velocity along the gradient direction.

In this context, bipolar gradient pulses are applied, and the resulting phase difference is used to infer flow velocities along the gradient direction. By applying a pair of bipolar gradients, the static phase term is canceled, isolating the second term for direct velocity measurement.

Alternative perfusion techniques include dynamic susceptibility contrast (DSC) and arterial spin labeling (ASL), each exploiting different mechanisms to estimate flow-related parameters such as volume distribution, flow rate, and transit characteristics [59]. All these methods retain the advantage of enabling direct, noninvasive flow quantification.

2.4. Magnetic Resonance Elastography (MRE)

Magnetic Resonance Elastography (MRE) is another advanced functional technique that non-invasively measures the mechanical properties of tissue, specifically its stiffness [60]. The method involves applying low-frequency mechanical vibrations to the tissue using an external driver to generate shear waves that propagate through the sample. The gradient pulses are synchronized with the vibrations so that the wave's movement is recorded as phase shifts in the MRI signal. By analyzing the “wave” images, an inversion algorithm can calculate the tissue's viscoelastic properties, producing a quantitative map of stiffness.

3. Assessing Complex Tissue Constructs Properties using a Multiparametric MRI Toolbox

All the previously presented MR Techniques can be assembled into an “MRI Toolbox” that can be tailored to different tissue engineering monitoring needs. Table 3 summarizes these MR Techniques and a few of the tissue properties that can be used to identify in this context.

Assessed Tissue Property	MR Technique	References
Tissue water content, tissue maturity, and structure. The T2/T1 ratio can be an indicator of tissue stiffness.	T1 and T2 Maps	[15], [61], [62]
Restriction of water motion by tissue microstructure, tissue growth, porosity, nutrient flow, cellularity, and oxygen tension.	Diffusion-Weighted Imaging	[15], [56], [57]
Vascularization, nutrient flow distribution and management, shear stress	Perfusion	[63], [64]
Mechanical properties, including tissue stiffness and viscoelasticity.	MRE (Magnetic Resonance Elastography)	[60], [65]

Table 3: Tissue properties and corresponding MR techniques for tissue-engineering monitoring

Although not presented here, there are quite a few more techniques, such as $T_{1\rho}$ contrasts and Magnetization Transfer (MT) measurements to assess macromolecular matrix contents[15], [66], Diffusion Tensor Imaging (DTI) to assess tissue anisotropy & fiber organization [67], [68], [69],

or even Superparamagnetic Iron Oxide Labeling (SPIO-Labeling) for cell tracking localization [70], and many more.

III. Overcoming MRI's Limitations for Laboratory Use

One of the greatest limitations preventing broader use of MRI in laboratory-based tissue engineering stems from the nature of MRI systems themselves. Conventional MRI scanners are expensive, with high acquisition and maintenance costs, often reaching millions of euros [71]. This financial burden restricts MRI access to specialized imaging centers and large hospitals, making it impractical for routine laboratory environments. Furthermore, MRI systems typically require dedicated infrastructure: shielded rooms, cryogenics for superconducting magnets and strict safety protocols to manage strong magnetic fields. The large physical footprint and technical complexity of such installations pose significant barriers to integration within standard tissue engineering facilities, where space is often limited and workflows demand flexibility. Safety concerns further complicate matters. High static magnetic fields present risks to both operators and equipment, necessitating rigorous screening and limiting compatibility with many laboratory devices. As a result, MRI, despite its powerful capabilities, remains underutilized.

III.A. Benchtop MRI as a Practical Solution

Low-field benchtop (or Tabletop/Desktop) MRI scanners have emerged as a versatile class of instruments. Designed to fit within laboratory or classroom environments, these compact systems extend the utility of MRI beyond traditional clinical and research facilities. By relaxing the requirements for size, shielding, and cryogenics, benchtop scanners offer a practical solution for a variety of specialized and emerging applications.

Benchtop scanners generally operate in the 0.01 to 1 Tesla range, much lower than standard clinical MRI (1.5–3 T). These compact systems bring MRI capabilities into laboratories, classrooms, and point-of-care settings by avoiding the bulky infrastructure of high-field machines. Reduced field strength comes with a lower signal-to-noise ratio, but modern advances [in post processing methods](#), magnet design and electronics have enabled useful imaging even at tens or hundreds of milli-tesla [26]. Benchtop MRI units are typically self-contained and easier to site, often not requiring special shielded rooms or extensive cryogenic systems. Many of these systems also feature open configurations, which enhance their versatility by allowing easier sample access, accommodating diverse specimen types, and enabling integration with complementary analytical tools.

1. Benchtop Scanners Magnets

Low-field benchtop MRI systems employ either resistive electromagnets or permanent magnets, each with distinct advantages in cost, size, and performance for specific use cases.

Resistive electromagnet systems use current-carrying coils (often with iron cores) to generate a static magnetic field. Their key advantages are field tunability and the ability to switch the field on or off, making them well suited for research, teaching, and portable applications. Coil geometries can be optimized for uniformity, and Ultra-Low Field (< 0.1T) designs remain lightweight and require minimal cooling. However, resistive magnets become impractical above ~0.3T due to escalating power and cooling demands [26].

Permanent magnet systems (e.g., rare-earth types) provide a constant (B_0) with no power input or cryogenics, offering simple, low-maintenance operation [26]. Designs such as Halbach arrays enable fields up to 1 T, exemplified by commercial systems like the Bruker ICON scanner and the Aspect Imaging M-Series [72] (Figure 15) used to produce high-resolution 3D

images of small animals and rodents. These systems demonstrate that permanent-magnet MRI can deliver useful imaging performance at low field with a self-contained, cryogen-free design.

Despite their operational simplicity, permanent magnets have notable limitations. They tend to be heavy and bulky relative to their field strength, requiring large volumes of magnetic material to achieve a strong ($\mathbf{B}_0$). Their always-on field poses safety and siting challenges without proper shielding. Field inhomogeneity and drift are also more pronounced: shaping the field is harder, and ambient temperature changes can alter field strength over time [26]. Typical designs have homogeneity of tens of thousands of ppm over the imaging volume, far worse than superconducting systems. A one-degree thermal fluctuation can shift resonance by tens of kHz, necessitating temperature stabilization or software drift correction. Material properties limit the maximum field to ~ 1 T for classical NdFeB alloys. Still, permanent-magnet benchtop MRI scanners offer low-cost, low-maintenance imaging, ideal for settings prioritizing simplicity over high spatial resolution.



Figure 15 : Images of commercial permanent magnet systems: Bruker ICON scanner (Left) and Aspect Imaging M [72] (Right)

2. Shielding Techniques

In Benchtop Systems, conventional shielded rooms are replaced with local shielding strategies. Unlike large clinical MRI scanners which require large room-scale Faraday cages, these small-scale systems rely heavily on coil and device level shielding. Effective RF shielding must block environmental RF noise, contain transmitted RF fields, and minimize artifacts from gradient switching. Multiple advanced strategies such as active RF shielding, involving the generation secondary electromagnetic fields to cancel the primary coil's external radiation, and self-shielding coil designs [73], [74], [75], [76], have been developed. However, the most basic and commonly used shielding approaches involve passive conductive enclosures near or around the RF coil. Among the many possible shielding strategies, passive shielding remains the most widely used and reliable approach for coil-level RF protection. A simple yet effective method involves enclosing the coil within a conductive layer, typically copper or brass foil ($\sim 9\text{--}25\text{ }\mu\text{m}$ thick), grounded at a single point to avoid ground loops [27], [77]. For example, wrapping a 3D-printed coil housing in copper foil, connected to the system ground via the RF coil connector, improved SNR by over fivefold in a portable brain MRI prototype [27].

Material selection is also critical. Copper is preferred due to its high conductivity and non-magnetic nature; however, alternative non-magnetic materials like aluminum may also be used. The shield's thickness must balance RF isolation with eddy current suppression. Thicker shields provide better RF blocking but can exacerbate gradient-induced eddy currents, leading to imaging artifacts [78], [79] and noticeable gradient delays [27]. Ultra-thin copper foils ($\sim 9\text{ }\mu\text{m}$) have been shown to sufficiently attenuate RF noise while minimizing eddy current artifacts at high magnetic fields [80] while, similarly, for a lower-field setup [27], a copper foil thickness of $\sim 25\text{ }\mu\text{m}$ was

sufficient to attenuate RF noise without eddy current induced gradient delays. To further address eddy currents, slotted and segmented shield designs have been developed. Longitudinal slots in cylindrical shields interrupt current loops while maintaining effective RF reflection [81].

Alternatively, phosphor-bronze meshes have demonstrated superior performance, providing excellent RF shielding (>15 dB better than segmented copper sheets) while substantially reducing eddy current-induced artifacts [81]. Also, conductive shielding fabrics draped over the entire scanner have also been shown to provide adequate shielding while minimizing eddy current artefacts [82].

3. Benchtop Applications

Low-field benchtop MRI scanners, with their compact footprints and relaxed citing requirements, have already opened the door to a diverse array of uses far beyond traditional clinical settings.

3.1. Preclinical Research and Small Animal Imaging

In Preclinical Research and Small Animal Imaging, Benchtop MRI aims to democratize small-animal MRI by using lower field magnets that dramatically reduce cost and infrastructure needs. For example, a permanent-magnet 0.5 T MRI scanner was developed as a “proof-of-concept” for mouse imaging: it has no cryogenics and much lower installation cost, yet was able to produce useful in vivo images of mouse anatomy [22], [83]. The authors emphasize that the lower initial and running costs of the benchtop MRI make it very attractive for routine preclinical studies, so long as the image quality is sufficient for the scientific questions. Moreover, low-field imaging can even have some advantages in preclinical settings: lower field means reduced magnetic susceptibility artifacts and fewer dielectric resonance issues at small scale [84]. For example, this can be useful when imaging around implanted devices or tracking cell injections, where high-field MRI might suffer distortions.

3.2. Materials Science and Industrial Applications

Beyond biomedical subjects, low-field MRI has important applications in materials science, engineering, and industry. The non-destructive imaging capability of MRI, even at low field, is valuable for examining the internal structure of objects that are opaque to visible light or X-rays (especially soft materials or those without strong density contrast). One class of devices, often called Nuclear Magnetic Resonance (NMR) analyzers or profile scanners, uses permanent magnets to analyze materials in situ.

For instance, the NMR-MOUSE (MOBILE Universal Surface Explorer) is a one-sided handheld magnet system that creates a shallow region of sensitive field outside the magnet. It can perform 1D MR depth profiling of surfaces without needing to enclose the sample [85], [86]. Using this device and similar ones, it has been possible to measure the layer structure of historical paintings [87], the hydration depth in skin [88], or burn wounds [89]. These applications take advantage of low-field MRI's ability to probe hydrogen density and relaxation in materials. Although the output is typically a depth vs. signal graph (or a series of 1D measurements), it provides unique information nondestructively. The low magnetic field is also a benefit here, because it reduces susceptibility-induced distortions when scanning heterogeneous materials (metals, pigments, etc.), and the compact permanent magnet can be brought right next to large objects (one could never do that with a 7 T superconducting magnet, for example).

Benchtop MRI scanners have also been adapted to perform full 2D or 3D imaging of material samples. In the petrochemical industry, for example, MRI is used to study porous rock cores to analyze oil and water content distribution. Low-field benchtop systems dedicated to core analysis can image rock samples of several centimeters in diameter, mapping their porosity and fluid

saturation. Pure Devices GmbH offers an MR-Core Analyzing Tomograph (MR-CAT)[90], which is essentially a low-field MRI scanner optimized for geological cores.

Other industrial uses of benchtop MRI include the imaging of extruded rubber profiles for non-destructive, in-line quality control [91], agricultural and food applications[92], and many more[93], [94] as a means of non-destructive testing (NDT). In many of these cases, low-field machines are favored because they are simpler, safer, and easier to install on a factory floor or lab bench. Also, many materials of interest contain hydrogen atoms in environments (like bound water in concrete, or oil in seeds) that give sufficient signal even at low field, so ultra-high field is not necessary to get a useful image or measurement. The reduced specific absorption rate (SAR) at low field is another bonus, meaning one can scan certain dielectric or conductive samples without significant heating [84].

3.3. Education and training

Lastly, another important application area for benchtop MRI is in education and training. MRI is a complex modality that historically has been taught with lectures and maybe demonstrations on full-size clinical scanners. However, full MRI systems are impractical for hands-on student use due to cost and safety. Low-field benchtop scanners have changed this by providing compact, user-friendly MRI devices for instructional use [27].

Several companies and academic groups have developed teaching MRI systems that allow students to experiment with all aspects of MRI, from pulse sequence parameters to image reconstruction, in a classroom or lab setting. For example, Pure Devices' Research Lab and Portable Lab benchtop MRI systems are explicitly marketed for scientific and educational use, with a permanent magnet, easy-to-use software, and a small footprint [95]. These units let students directly acquire 2D and 3D images of sample objects (typically small phantoms or test tubes of water) on a desktop. The educational benefit is significant: learners can visualize how slice selection works, see the effect of changing gradient strengths or RF pulse timings, and get a feel for MRI safety and operation, all on a system that is safe and low-cost enough for a college laboratory.

Similarly, Niumag (a Chinese manufacturer) sells an EDUMR system for experimental teaching of MRI principles, which is a table-top 0.5 T permanent magnet device designed for universities (weighing ~138 kg, easily accommodated in a classroom) [96].

Some universities have also built their own teaching MRI scanners as open-source projects. The previously mentioned design by Cooley et al. [27] is one example: they created a low-cost tabletop MRI scanner kit and even built 20 of them for use in parallel by students, demonstrating concepts like slice encoding and Fourier reconstruction in a tangible way. This scanner used a small Halbach permanent magnet (on the order of 80 mT) and simple gradients, with an Arduino-based console, showing that one can construct an MRI essentially from scratch for educational purposes, a concept that would have been science fiction in the early days of MRI education.



Figure 16 : Images of benchtop systems dedicated to research and education: custom 0.18T design by Cooley et al. [27] (Left), Niumag EDUMR 0.5T scanner [96] (Center) and Pure Devices GmbH 0.55T Research Lab [95] (Right)

4. Perspective Towards Tissue Engineering

In the context of Tissue Engineering such machines might be crucial to the insertion and the widespread adoption of MRI as an everyday tool in a laboratory setting. Their small footprint, relatively low purchase and maintenance costs and lax safety requirements make them ideal for in line integration.

However, despite their growing utility, low-field benchtop MRI systems still face significant limitations. Chief among them is reduced signal-to-noise ratio (SNR), which directly impacts image resolution and scan time. Lower field strength means inherently weaker MR signals, requiring longer acquisition times or averaging, which can hinder throughput or dynamic imaging. This also limits the sensitivity for detecting subtle contrasts or small changes.

Ultimately, while low-field benchtop MRI expands access and lowers cost, it does so with trade-offs that limit the available SNR. Although the landscape of low-field benchtop MRI is rich and continually evolving, the current promising results have only been made possible through technological advances, aiming to narrow the performance gap between low-field and mid-field MRI. In this work, the aim will be to contribute to this endeavor by exploring new approaches to tackling SNR for small samples and enhancing benchtop scanners' measurement sensitivity to achieve microscopic capabilities.

III.B. The Signal-to-Noise Ratio (SNR) Challenge

In MRI, the quality of images obtained is fundamentally dictated by the magnetic resonance signal and the presence of noise. While numerous factors including spatial resolution, contrast weighting, and imaging speed shape the MRI acquisition, no single metric more directly captures the quality of the acquired image than the signal-to-noise ratio (SNR). SNR reflects the balance between the amplitude of the induced MR signal and the background of random noise arising from both the sample and the detection system.

1. Noise in MRI

Charges in a conductive medium are never completely motionless. At any nonzero temperature, random collisions with the surrounding lattice set the charges into a Brownian-type motion. Applying the stochastic Langevin equation to a single charge reveals that this microscopic motion gives rise to two inseparable effects: random “kicks” from lattice atoms produce fluctuations (the Langevin force), while viscous drag against the environment leads to energy dissipation. The fluctuation–dissipation theorem states that at thermodynamic equilibrium, spontaneous fluctuations must be accompanied by an equivalent dissipation mechanism. Charles Kittel’s *Introduction to Solid State Physics* presents a full overview of this result [97].

Experimentally, Nyquist and Johnson confirmed that thermal agitation in a resistor generates measurable voltage and current noise under steady-state conditions. They demonstrated that the root-mean-square voltage noise (σ_{noise}) across a resistance (R) at temperature (T), measured over a bandwidth Δf , obeys:

$$\sigma_{\text{noise}} = \sqrt{4 k_B R T \Delta f} \quad (\text{I.44})$$

where k_B is Boltzmann’s constant and σ_{noise}^2 represents the variance of the noise voltage.

The noise picked up by an RF coil in an MRI experiment is governed by the same principle. Taking into account the different loss mechanisms at play, it can be expressed as:

$$\sigma_{noise} = \sqrt{4 k_B R_{eff} T_{eff} \Delta f} \quad (I.45)$$

where R_{eff} is the effective resistance seen at the coil terminals, which includes all the coil losses such as the intrinsic ohmic losses of the conductor and frequency-dependent losses arising from dielectric and radiation effects in the sample, T_{eff} is the effective noise temperature, a weighted average of the physical temperatures of the coil and sample, Δf is the effective noise bandwidth of the acquisition chain.

However, as the acquisition chain is non-ideal, it adds its own noise characterized by a noise factor F , which multiplies noise power by F (and rms voltage by $\sqrt{F}$). Hence the total rms noise is:

$$\sigma_{noise} = \sqrt{F} \cdot \sqrt{4 k_B R_{eff} T_{eff} \Delta f}, \quad (I.46)$$

2. Defining SNR

By definition, the SNR for a single voxel is:

$$SNR = \frac{V_{signal}}{\sigma_{noise}} \quad (I.47)$$

Where (V_{signal}) is the peak induced signal voltage per unit current (I.34), and (σ_{noise}) is the noise mainly arising from the thermal contributions from the coil and the sample.

Therefore, the single-scan signal-to-noise ratio gives

$$SNR_{single} = \frac{V_{signal}}{\sigma_{noise}} = F^{-1/2} \frac{\omega_0 (B_1/I) M_{\perp} \delta V}{\sqrt{4 k_B R_{eff} T_{eff} \Delta f}}. \quad (I.48)$$

Furthermore, averaging N identical scans improves SNR by ($\sqrt{N}$). For a repetition time T_R and total acquisition time (t_{acq}), ($N = t_{acq}/T_R$) which gives:

$$SNR = F^{-1/2} \frac{\omega_0 (B_1/I) M_{\perp} \delta V}{\sqrt{4 k_B R_{eff} T_{eff} \Delta f}} \sqrt{\frac{t_{acq}}{T_R}}. \quad (I.49)$$

Finally, here, Δf represents the total reception bandwidth divided by the sampling interval in frequency space. In the time domain, this concept translates to the observation time: the duration over which the signal is acquired. According to time–frequency duality of the Fourier transform, the frequency resolution is inversely proportional to the acquisition window, and, therefore, the bandwidth can be equivalently expressed as $\Delta f = 1/t_o$ where (t_o) is the observation time per pixel :

$$SNR = F^{-1/2} \frac{\omega_0 (B_1/I) M_{\perp} \delta V}{\sqrt{4 k_B R_{eff} T_{eff}}} \sqrt{\frac{t_o \cdot t_{acq}}{T_R}}. \quad (I.50)$$

3. Optimizing SNR for MR Microscopy

Compared to other imaging modalities such as optical microscopy, magnetic resonance imaging (MRI) is fundamentally limited by intrinsically low signal levels. Only a small fraction of nuclear spins contribute to the net magnetization detectable by the RF coil, as the population difference between spin states at thermal equilibrium is minimal. As a result, the MR signal is inherently weak, while thermal noise from the coil and sample remains significant.

This inherent limitation becomes even more pronounced in magnetic resonance microscopy (MRM), where both small sample volumes and extremely high spatial resolution are required. In this context, the detectable signal scales with voxel volume, which must be drastically reduced to achieve micron-scale resolution, while noise remains unchanged. As resolution increases, SNR deteriorates rapidly, posing a fundamental constraint on image quality.

This trade-off defines the core SNR challenge of MR microscopy: achieving sufficient signal to generate informative, high-resolution images while preserving the fine structural detail necessary for quantitative characterization of engineered tissues and small biological samples.

The SNR relationship presented above can be written as:

$$\text{SNR} \propto \delta V \times S_{\text{eq}} \times \sqrt{t_{\text{acq}}} \quad (\text{I.51})$$

This highlights that either, increasing the total acquisition time (t_{acq}), increasing the voxel size (δV) or the equipment sensitivity (S_{eq}) will improve overall SNR. The optimization of SNR is a multi-faceted task involving trade-offs between hardware capabilities, acquisition parameters and experimental needs.

- Increasing acquisition time: Averaging multiple scans reduces random noise through ($\sqrt{N}$) the improvement in SNR. This approach is straightforward and does not affect spatial resolution or contrast weighting. However, increased scan duration is not always practical as both throughput considerations and application specific time constraints often impose strict time limits in both clinical and experimental settings.
- Increasing voxel size: Enlarging the voxel volume directly boosts SNR, as the MR signal scales linearly with (δV). Yet, this comes at the cost of spatial resolution. In contexts such as microscopic imaging of engineered tissues, or when structural detail is critical, increasing voxel size will compromise the ability to resolve fine features.
- Increasing equipment sensitivity:
 - Static Field Strength: The MR signal scales approximately linearly with static magnetic field strength ($\mathbf{B}_0$), while noise grows more slowly, resulting in improved SNR at higher ($\mathbf{B}_0$). However, increasing field strength is not without drawbacks. The cost and complexity of higher-field magnets rise sharply with ($\mathbf{B}_0$), making such upgrades economically unfeasible. Additionally, stronger fields require more sophisticated shielding, larger infrastructure, and stricter safety protocols. Beyond financial considerations, higher ($\mathbf{B}_0$) also alters relaxation times and can exacerbate susceptibility artifacts. In many low-field MRI setups, including those designed for tissue engineering or small-animal imaging, maximizing coil sensitivity and acquisition efficiency proves to be a more practical and cost-effective path to SNR improvement than escalating ($\mathbf{B}_0$).
 - Coil Sensitivity: As one of the most effective avenues for SNR optimization, coil sensitivity plays a central role. An efficient coil provides enhanced signal induction through (B_1/I), while simultaneously minimizing perceived losses via reduced noise contributions (σ_{noise}) effectively improving SNR. Careful

coil design through tailored geometry, optimized current distribution, and performance optimization can substantially boost (S_{eq}) by locally enhancing sensitivity. In low-field or miniature systems, coil optimization often yields the most significant gains relative to other strategies.

- Acquisition Parameters: SNR is further modulated by acquisition parameters such as bandwidth, repetition time (TR), echo time (TE), and flip angle. For instance, reducing receiver bandwidth improves SNR by narrowing noise contributions but risks introducing artifacts from chemical shift and susceptibility. Flip angles can be optimized (e.g., using Ernst angle conditions [98]) to maximize SNR given specific (T_1) and TR constraints. Each parameter must be carefully selected to balance SNR against contrast, resolution, and acquisition time.

Addressing the central challenge of MR microscopy by producing sufficient signal while maintaining the microscopic resolution necessary for quantitative imaging of engineered tissues requires strategic improvements. While many parameters can be adjusted to improve SNR, not all strategies are equally practical or effective. Simply increasing acquisition time is universally applicable but constrained by time limitations in many experimental setups. Enlarging voxel size provides a straightforward gain in SNR but greatly undermines spatial resolution, which is unacceptable for high-detail applications. Raising the static field strength offers a powerful but costly route, with significant practical and infrastructural barriers limiting overall accessibility and integrability in a standard biological laboratory. By contrast, enhancing RF coil sensitivity through careful design and optimization seems to offer the most efficient, scalable, and cost-effective means of improving SNR making it the preferred strategy.

III.C. The RF Coil: The Gateway to Signal

1. Types of RF Coils

RF coils can be broadly classified into three main categories based on their geometry and application: volume coils, surface coils, and array coils. The choice of coil is critical as it dictates the signal sensitivity, field homogeneity, and spatial coverage of the experiment [99].

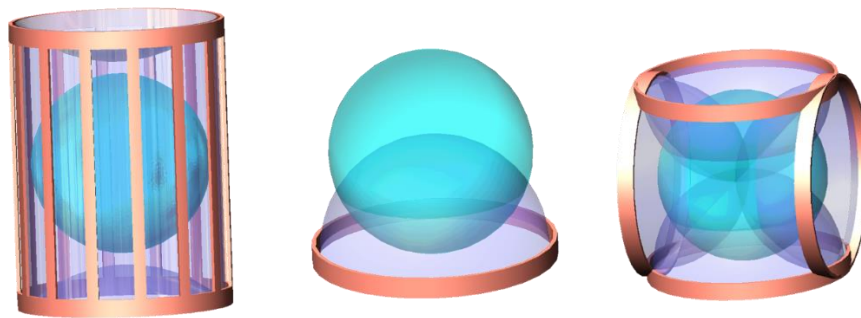


Figure 17 : Diagram Presenting each type of coil and their relation to a spherical sample. (Left) Volume Coil, (Center) Surface Coil and (Right) Array coils.(inspired from [100])

Volume Coils: These coils are designed to completely enclose the sample or the region of interest, providing a highly uniform RF magnetic field over a large volume. This homogeneity is crucial for consistent excitation and reception, making them ideal for imaging entire objects or anatomical regions where uniform sensitivity is paramount.

Surface Coils: In contrast to volume coils, surface coils are placed directly on the surface of the object being imaged. Their simple, often planar, design provides very high signal sensitivity in the region immediately beneath the coil. However, the sensitivity of a surface coil decreases rapidly with distance from the coil, and the ($\mathbf{B}_1$) field is inherently inhomogeneous. This makes them particularly well-suited for high-resolution imaging of superficial structures.

Array Coils: To overcome the limited coverage of single surface coils while retaining their high signal sensitivity, array coils were developed. These consist of multiple, smaller coil elements arranged over a larger area. Each element acts as an independent receiver, and their signals can be combined to create a composite image with both high signal sensitivity and a large field of view. A key advantage of array coils is their ability to facilitate parallel imaging techniques, which significantly reduce acquisition time. However, designing array coils requires careful consideration of the electromagnetic coupling between adjacent elements to avoid signal interference.

Table 4, summarizes the main characteristics of volume, surface, and array resonators, including their typical field profiles, transmission and reception behavior, and principal advantages and limitations

Type	Geometry and Field Profile	Transmission (Tx)	Reception (Rx)	Advantages	Limitations
Volume Coils	Surround the sample; uniform field	Highly homogeneous excitation across the sample	Uniform sensitivity across the sample volume	High field homogeneity, efficient for samples that match the coil size; Ideal for large samples	Reduced efficiency with small samples; Bulky; Less suited for localized or surface studies
Surface Coils	Planar; strong field near surface, decays with depth	Localized excitation favoring near-surface regions	High near-surface sensitivity; Rapid sensitivity decay with depth	High superficial sensitivity; Simple, adaptable design; Ideal for localized studies	Poor depth penetration; Field inhomogeneity; Lower efficiency for large or deep samples
Array Coils	Multiple small elements covering larger area or volume	Combined excitation, shaping possible but uniformity may be complex	Local high sensitivity per element; Supports parallel acquisition	Large coverage with high sensitivity; Compatible with fast parallel imaging; Modular flexibility	Complex element decoupling; Potential non-uniform excitation; Higher overall system complexity and cost

Table 4 : Overview of volume, surface, and array RF resonators summarizing their geometry, transmission and reception characteristics, and principal advantages and limitations.

Volume coils offer excellent homogeneity and are suited for larger samples but are less effective when the sample volume is small. Surface coils provide high sensitivity near the surface and are flexible in design but suffer from poor depth penetration and inhomogeneous fields. Array coils combine multiple elements to optimize sensitivity over large areas and enable parallel acquisition, but they require careful engineering to manage inter-element coupling and maintain field uniformity. They can also require larger, more complex setups that may not be compatible with the limited workable space and connectivity available with some benchtop scanners.

The choice of coil, as described above, directly influences the spatial characteristics of the transmitted RF field ($\mathbf{B}_1^+$), which is a critical factor in the subsequent excitation of nuclear spins. For tissue engineering application, which requires precise quantitative imaging, a homogenous field is necessary across the entire imaging volume.

A critical challenge in quantitative MRI arises from the spatial inhomogeneity of the transmit RF field ($\mathbf{B}_1^+$). In practice, the ($\mathbf{B}_1^+$) field is rarely uniform, meaning the flip angle becomes position dependent. If the generated field is inhomogeneous, each spin at position ($\mathbf{r}$) experiences a local flip angle. The flip angle equation, assuming spatially non-uniform ($\mathbf{B}_1^+$) field across the sample volume, becomes:

$$\theta(\mathbf{r}) = \gamma \int_0^\tau \mathbf{B}_1^+(\mathbf{r}, t) dt \quad (\text{I.52})$$

This spatial variation can become a significant source of error in quantitative imaging [101]. Specifically, for tissue-engineered constructs, which are often heterogeneous in composition and structure, if the goal is to measure parameters like cell density, extracellular matrix deposition, or scaffold degradation, these measurements depend directly on signal intensity, which is a function of the flip angle.

If the flip angle is not uniform across the construct, nominally identical regions will experience different levels of excitation. For example, a region with high cell density located in an area of low ($\mathbf{B}_1^+$) strength (and thus a lower flip angle) could produce a signal intensity equal to that of a region with lower cell density situated in an area of high ($\mathbf{B}_1^+$) strength. In that case, this would confound the interpretation of the image, as the signal variation would be due to an instrumental artifact rather than true biological differences which can lead to erroneous quantifications. Although ($\mathbf{B}_1^+$) mapping techniques coupled with correction algorithms exist [102], [103], RF coils capable of generating highly homogenous fields will be preferred for sample excitation. For larger samples, although generally less sensitive, volume coils will be selected over other topologies.

2. RF Coil Characteristics

2.1. RLC Equivalent Model for a Simplified Coil Overview

Electrically, any resonator can be modeled using a lumped-element resonant RLC circuit (Figure 18). The inductance (L) reflects the ability of the conductor geometry to store magnetic energy through the generation of magnetic flux when current flows. The capacitance (C) results from both the inherent distributed capacitance of the structure and any additional capacitive components incorporated to tune the resonant frequency. The effective resistance (R_{eff}) represents the combined losses in the system.

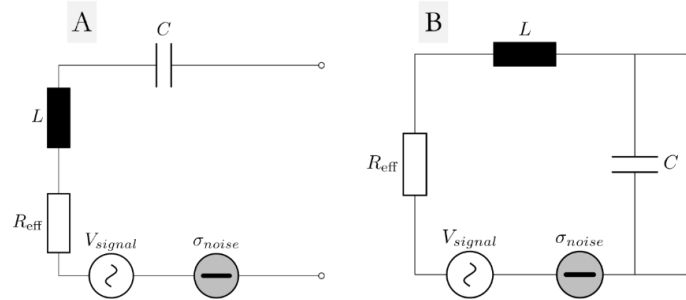


Figure 18 : RLC equivalent model of an RF resonator A) R-L-C series configuration and B) (R-L)//C parallel configuration

Applying Kirchhoff's laws to the two RLC topologies in Figure 18 gives the following resonance conditions:

- Series (R-L-C):

$$LC \omega_0^2 = 1 \quad (\text{I.53})$$

- **Parallel ((R-L)//C):**

$$LC \omega_0^2 = 1 - \frac{R_{eff}^2 C}{L} \quad (I.54)$$

In practice, for coils with ($R_{eff} \ll 1$), the shift in the parallel-mode resonance is negligible and both topologies resonate very nearly at ($\omega_0 \approx 1/\sqrt{LC}$).

In the series tuning configuration (Figure 18-A), the tuning capacitor (C_T) is placed in series with the coil inductance (L). The equivalent impedance can be written:

$$Z_{coil}(\omega) = R_{eff} + j\omega L + \frac{1}{j\omega C_T} \quad (I.55)$$

At resonance, the net reactance cancels' out which leaves:

$$Z_{coil}(\omega) = R_{eff} \quad (I.56)$$

While this form of tuning is simple, $Z_{coil} = R_{eff}$ is expected to be very small compared to the standard system impedance (50 Ω).

In the parallel tuning configuration (Figure 18-B), the tuning capacitor C_T is placed in parallel with the coil inductance L . The equivalent impedance can be written:

$$Z_{coil}(\omega) = \left[\frac{1}{R + j\omega L} + j\omega C_T \right]^{-1}. \quad (I.57)$$

At resonance this becomes [104] :

$$Z_{coil}(\omega) \approx \frac{(\omega L)^2}{R_{eff}} = R_{eff} Q^2, \quad (I.58)$$

In this configuration, a higher impedance at resonance is expected.

2.2. Coil Interfacing with The Acquisition Chain

In an MRI experiment, the acquisition chain carries the signal from the coil, with an output impedance Z_{coil} (source impedance), to the preamplifier, with an input impedance Z_{pre} (load impedance), before being transmitted to the spectrometer.

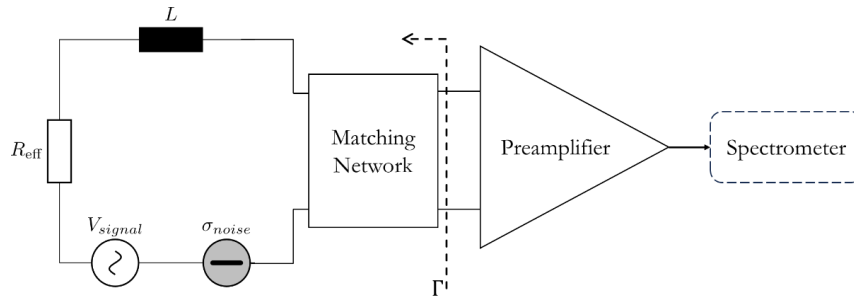


Figure 19: Simplified acquisition chain schematic

2.2.1 Power Matching

Maximum power transfer from coil to preamplifier occurs when the load impedance equals the complex conjugate of the source impedance:

$$Z_{coil} = Z_{pre}^* \quad (III-59)$$

This “impedance matching” works by canceling out reactive components (inductance or capacitance) and equalizing real resistances, so all available signal power is transferred without loss from reflection.

These mismatch losses are most conveniently expressed in terms of the complex reflection coefficient at the coil-preamplifier interface:

$$\Gamma = \frac{Z_{coil} - Z_{pre}^*}{Z_{coil} + Z_{pre}^*}. \quad (III-60)$$

The fraction of power actually delivered to the load or Power-Transfer Ratio (P_T) is:

$$P_T = 1 - |\Gamma|^2. \quad (III-61)$$

and the corresponding mismatch loss (in dB) is:

$$ML = -10 \log_{10}(P_T) \quad [\text{dB}] \quad (III-62)$$

while the return loss (a measure of reflected power) is:

$$RL = -20 \log_{10}|\Gamma| \quad [\text{dB}] \quad (III-63)$$

To mitigate these losses, a matching network is used with the aim of getting as close as possible to an ideal power matching ($\Gamma = 1$). Figure 19: Simplified acquisition chain schematic shows a simplified acquisition chain schematic with a two-port network between the coil and the preamplifier representing the matching network. In general, the matching network is implemented using low-loss reactive components so as not to introduce attenuation or transmission noise. Various matching techniques are reviewed in the work of Mispelter et al. [104]. The most conventional circuits use capacitors to realize a capacitive transformer. However, as this is quite the extensive topic, matching techniques will only be discussed as they become relevant in the following chapters.

2.2.2 Noise Matching

The source impedance that minimizes SNR degradation in the preamplifier is generally not the same as the one that maximizes average power transfer. To account for this, a noise reflection coefficient (Γ_n) is introduced. This coefficient characterizes the mismatch between the coil impedance and the preamplifier’s optimal source impedance (Z_{opt}), which minimizes the noise contribution of the preamplifier:

$$\Gamma_n = \frac{Z_{coil} - Z_{opt}^*}{Z_{coil} + Z_{opt}^*}. \quad (III-64)$$

In essence, Γ_n quantifies how far the system is from ideal noise matching conditions. In many modern preamps, designers deliberately choose a slight mismatch in favor of noise performance rather than absolute power transfer [105]. However, it is not possible to optimize both Noise

Matching and Power Matching. In cases where Power Matching is prioritized, the added noise contribution can be quantified and taken into account in calculations. It can be shown that the noise factor associated with this mismatch can be written [100]:

$$F_p = 1 + \frac{T_N}{T_{eff}} \left(\frac{1 + |\Gamma_n|^2}{1 - |\Gamma_n|^2} \right) \quad (\text{III-65})$$

Where T_N the standard preamplifier noise temperature.

Most modern MRI systems already incorporate dedicated low-noise preamplifier leading to a port of a standard impedance (typically 50Ω). In such cases a power matching will be selected such that $Z_{coil} = 50 \Omega$ and the total noise factor (F) contribution will have to be obtained through direct measurements.

2.2.3 Scattering Parameters (S-Parameters)

Traditional circuit parameters such as impedance and admittance become difficult to measure and interpret at the frequencies typically used for imaging. To address this challenge, scattering parameters, commonly referred to as S-parameters, are employed. S-parameters define relationships between incident and reflected waves at the ports of a network. At each port, two complex quantities are considered: the incident wave, denoted a_i , and the reflected wave, denoted b_i . These wave amplitudes are related to measurable voltages and currents through the expressions:

$$a_i = \frac{V_i + Z_0 I_i}{2\sqrt{Z_0}}, \quad b_i = \frac{V_i - Z_0 I_i}{2\sqrt{Z_0}}, \quad (\text{I.66})$$

where V_i and I_i represent the voltage and current at port i , and Z_0 is the characteristic impedance of the system (typically standardized to 50 ohms).

The general relationship between incident and reflected waves in an N-port network can be written in matrix form as:

$$\mathbf{b} = \mathbf{S}\mathbf{a}, \quad (\text{I.67})$$

where $(\mathbf{a})$ and $(\mathbf{b})$ are column vectors containing the incident and reflected wave amplitudes, respectively, and $(\mathbf{S})$ is the scattering matrix containing the S-parameters. Each element S_{ij} of the scattering matrix represents the ratio of the wave reflected at port i to the wave incident at port j , with all other ports terminated in matched loads:

$$S_{ij} = \left. \frac{b_i}{a_j} \right|_{a_k=0, k \neq j}. \quad (\text{I.68})$$

The physical interpretation of S_{ij} depends on whether $(i = j)$ or $(i \neq j)$:

- When $(i = j)$, S_{ii} represents the reflection coefficient at port i .
- When $(i \neq j)$, S_{ij} describes the transmission coefficient from port j to port i .

Two-port Network Configuration

For a two-port coil configuration, the relationships between incident and reflected waves simplify to:

$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix} \quad (I.69)$$

Each S-parameter in a two-port network can be formally defined based on the relationships between the incident and reflected waves. Specifically, they are given by:

$$S_{11} = \left. \frac{b_1}{a_1} \right|_{a_2=0}, \quad S_{22} = \left. \frac{b_2}{a_2} \right|_{a_1=0}, \quad (I.70)$$

$$S_{21} = \left. \frac{b_2}{a_1} \right|_{a_2=0}, \quad S_{12} = \left. \frac{b_1}{a_2} \right|_{a_1=0}. \quad (I.71)$$

Here, (S_{11}) corresponds to the reflection coefficient observed at port 1 while (S_{22}) describes the reflection at port 2. The transmission from port 1 to port 2 is captured by (S_{21}), and the reverse transmission from port 2 to port 1 is represented by (S_{12}).

It is often convenient to express the magnitude of S-parameters in decibels (dB) rather than on a linear scale. The conversion is given by:

$$|S_{ij}|_{\text{dB}} = 20 \log_{10}(|S_{ij}|). \quad (I.72)$$

A low (S_{11}) or (S_{22}) indicates efficient impedance matching of the coil to the system transmitter or receiver and minimal reflection while high (S_{21}/S_{12}) values between two coils may indicate coupling between the coils. This coupling effect can be used in certain characterization methods which will be exposed in a later section.

Measurement of S-parameters in MRI coil development is typically performed using a Vector Network Analyzer (VNA). The VNA injects known RF signals into the coil ports and measures the resulting incident and reflected waves to derive the S-parameters across the frequency band of interest, usually centered at the Larmor frequency specific to the target nucleus (for example, 13 MHz for 0.3T and 64 MHz for 1.5T systems).

Once the S-parameters are determined, critical quantities such as input impedance and inter-coil coupling can be assessed. The input impedance (Z_{in}) at port 1 and the output impedance (Z_{out}) at port 2 are related to the reflection coefficients by:

$$Z_{\text{in}} = Z_0 \frac{1 + S_{11}}{1 - S_{11}}, \quad Z_{\text{out}} = Z_0 \frac{1 + S_{22}}{1 - S_{22}}. \quad (I.73)$$

VNAs also provide several visualization modes to assist in interpreting S-parameter data. Reflection coefficients (S_{11}, S_{22}) are commonly displayed on a Smith chart, allowing immediate assessment of impedance matching and resonance behavior. In addition, Cartesian plots of magnitude ($|S|$) and phase versus frequency can be used to evaluate bandwidth, return loss, and transmission characteristics.

2.3. Impact of RF Coil Losses

Losses in an MRI RF coil arise from three principal mechanisms, each contributing to the total effective resistance (R_{eff}) seen at the coil terminals.

2.3.1 Conductor (Ohmic) Losses.

Conductor losses (R_c) are due to the finite electrical resistance of the coil's metallic conductors. At typical MRI frequencies, two effects dominate: **skin-effect losses** and **proximity-effect losses**.

a. Skin-Effect Losses

Skin-effect losses, modeled as an effective series resistance (R_{SE}), originate from the tendency of alternating RF current to flow predominantly near the surface of a conductor, rather than uniformly throughout its cross-section. This phenomenon occurs because the alternating magnetic fields associated with the current induce opposing eddy currents that cancel the interior current, confining the net current to a thin surface region whose thickness δ is given by [104]:

$$\delta = \sqrt{\frac{2\rho}{\omega\mu}} \quad (I.74)$$

where ρ is the resistivity of the conductor material, ω is the angular frequency of the RF current, and μ is the magnetic permeability of the conductor.

As frequency increases, δ decreases, causing the effective cross-sectional area available for current flow to shrink. This reduction in area increases the effective resistance of the conductor, leading to higher ohmic losses.

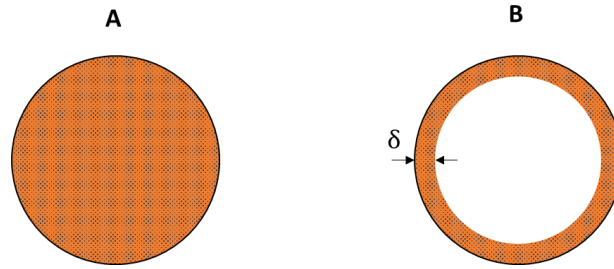


Figure 20 : Cylindrical conductor cross section: A) DC current distribution (uniform) B) AC current distribution (surface concentrated due to skin effect)(inspired by [106]).

This means that for a given conductor with a cross-section contour C , the resistance per unit length $R_{SE/unit}$ of a single winding, can be expressed as:

$$R_{SE/unit} = \frac{\rho}{C\delta}, \quad (III-75)$$

where ρ is the conductor resistivity. Here, $C\delta$ represents the effective conducting area through which current flows. To find the total resistance of the coil winding, the resistance per unit length is multiplied by the total length of the conductor l_c leading to:

$$R_c = R_{SE/unit} \cdot l_c. \quad (III-76)$$

b. Proximity Effects Losses

Proximity-effect losses arise when conductors are placed close together [107]. In this situation, the magnetic field produced by current in one conductor induces eddy currents in neighboring conductors. These induced eddy currents distort the primary current distribution, concentrating the current into localized regions within the conductors. This distortion reduces the effective cross-sectional area available for current flow, further increasing the conductor's resistance and

thereby enhancing energy dissipation. Proximity effects become more pronounced at higher frequencies and with closer conductor spacing.

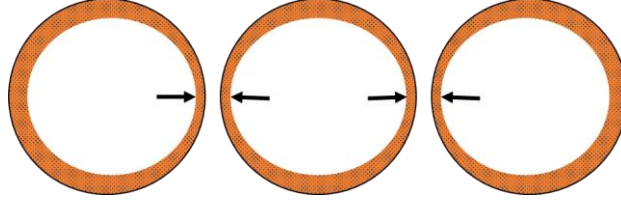


Figure 21 : Illustration of proximity effects between adjacent conductors (inspired by [106]).

In the case of a multi-turn coil, proximity effects also need to be taken into account in R_c calculations. To capture the proximity effect, a proximity factor k_p depending on geometry, winding distance and number of turns is introduced [108]:

$$R_c = R_{SE/unit} \cdot l_c \cdot k_p. \quad (\text{III-77})$$

2.3.2 Sample-Induced Losses

When an RF coil is loaded by a biological or dielectric sample, electromagnetic energy is dissipated within the sample through various mechanisms. The losses incurred by this phenomenon are referred to as sample losses ($R_s = R_{CD} + R_{CC}$).

a. Conduction and Dielectric Losses

Sample losses primarily arise from two physical mechanisms: conduction currents and dielectric polarization. Conduction losses are associated with the motion of free charges under the influence of the RF electric field, characterized by the sample conductivity σ_s . Dielectric losses, on the other hand, result from the phase lag in the response of bound charges to the applied field, quantified by the imaginary component of the complex permittivity $\epsilon = \epsilon' + j\epsilon''$.

Applying Poynting's theorem [44], the total power dissipated in the sample can be expressed as:

$$P_{CD} = \frac{1}{2} \iiint_V (\sigma_s + \omega \epsilon'') |E|^2 dV \quad (\text{I.78})$$

Where ω is the angular frequency of the RF field, E is the local electric field and V is the sample volume under consideration.

These losses can be equivalently modeled as an effective series resistance (R_{CD}) related to the current i flowing through the coil by:

$$P_D = \frac{1}{2} R_D i^2 \quad \Rightarrow \quad R_{CD} = \frac{1}{i^2} \iiint_V (\sigma_s + \omega \epsilon'') |E|^2 dV \quad (\text{I.79})$$

These losses are manifested macroscopically as heat deposition within the sample.

b. Capacitive Coupling Losses

Beyond conduction and dielectric mechanisms within the sample bulk, an additional significant loss pathway arises from capacitive coupling between the RF coil conductors and the sample surface. When the coil elements are positioned in close proximity to the sample, an effective distributed capacitance C_{cs} forms between the coil and the sample boundary.

This capacitance can be modeled as a complex quantity to account for dielectric losses:

$$C_{cs}^* = C_{cs} (1 - j \tan \delta), \quad (I.80)$$

where $\tan \delta$ is the loss tangent defined as

$$\tan \delta = \frac{\epsilon''}{\epsilon'}. \quad (I.81)$$

The RF electric potential V_{RF} across this capacitance drives a displacement current

$$I_d = j \omega C_{cs} V_{RF}, \quad (I.82)$$

which results in additional real power dissipation given by:

$$P_{CC} = \frac{1}{2} \omega C_{cs} V_{RF}^2 \tan(\delta). \quad (I.83)$$

Analogously to conduction losses, these capacitive coupling losses can be represented by an equivalent series resistance (R_{CC}):

$$R_{CC} = \frac{V_{RF}^2}{P_{CC}} = \frac{1}{\omega C_{cs} \tan \delta}. \quad (I.84)$$

Capacitive coupling becomes particularly significant when large conductive surfaces of the coil lie parallel and in close proximity to the sample, effectively increasing (C_{cs}) and enhancing the associated losses.

2.3.3 Radiation Losses (or Antenna Effect)

At sufficiently high frequencies or with electrically large coil geometries (coil dimensions comparable to a significant fraction of the RF wavelength), the coil will radiate electromagnetic energy into the surrounding space rather than confining it to the sample volume. These losses behave like an antenna's radiation resistance. The radiation resistance (R_{rad}) grows with $(ka)^2$ (where $(k = 2\pi/\lambda)$ and (a) is a characteristic coil dimension), becoming appreciable when $(a \gtrsim \lambda/10)$.

2.3.4 Effective Resistance and Temperature

The thermal noise picked up by an MRI RF coil is ultimately governed by the fundamental physical losses present within the resonator system, prior to the inclusion of any amplifier or matching network effects. These intrinsic coil-sample losses can be fully described by an effective resistance (R_{eff}) as previously outlined:

$$R_{eff} = R_c + R_s + R_{rad} \quad (I.85)$$

where R_c represents the conductor (ohmic) losses, R_s the sample induced losses, and R_{rad} represents radiation losses from the coil.

At lower magnetic field strengths ($<1T$) the contribution of (R_{rad}) is generally negligible compared to (R_c) and (R_s). This can be justified based on the scaling behavior of radiation resistance with frequency as (R_{rad}) scales approximately as $(ka)^2$. Given that $((ka)^2 \ll 1)$ under low-frequency (low-field) conditions, the radiation resistance becomes orders of magnitude smaller than the conductor and sample resistances. Moreover, the physical dimensions of MRI

coils are typically much smaller than the corresponding free-space wavelength, further suppressing radiation losses.

As a result, for small coils at low field, (R_{eff}) becomes:

$$R_{eff} \approx R_c + R_s \quad (I.86)$$

This means that at low magnetic fields, the dominant contributions to R_{eff} arise from R_c and R_s and the effective noise temperature becomes:

$$T_{eff} \approx \frac{R_c T_{coil} + R_s T_{sample}}{R_c + R_s} \quad (I.87)$$

In low-field RF resonators, the overall noise performance will be a mixture of the coil and sample thermal environments, weighted by the relative magnitudes of their associated resistances. If the sample dominates the losses ($R_s \gg R_c$), as often occurs when imaging large conductive samples, T_{eff} approaches T_{sample} . However, if the coil losses dominate ($R_c \gg R_s$), T_{eff} tends toward T_{coil} . The impact of the effective temperature will be particularly noticeable in cases where the coil and sample temperatures are very distinct (e.g., cryogenically cooled coil setups).

3. Performance Metrics

3.1. Quality Factor (Q)

The quality factor (Q) of an RF resonator is a dimensionless parameter that quantifies the ratio of maximal energy stored in the coil to the energy lost through resistive dissipation in a single cycle:

$$Q = \omega_0 \cdot \frac{\text{maximum energy stored}}{\text{energy dissipated per cycle}} \quad (I.88)$$

In both the equivalent series mode (R-L-C) and equivalent parallel mode ((R-L)//C) circuit at resonance ($\omega_0 = 1/\sqrt{LC}$), this can be expressed as:

$$Q = \frac{X_L(\omega_0)}{R_{eff}} = \frac{\omega_0 L}{R_{eff}} \quad (I.89)$$

Because (Q) is proportional to the energy stored relative to the energy lost for each oscillation, a high- Q coil accumulates more reactive energy resulting in larger oscillating currents and voltages for a given drive power, whereas a low- Q coil dissipates energy more rapidly. When an RF coil has a low-quality factor, the tiny voltage it induces by transverse magnetization decays, so quickly that important phase, frequency and amplitude information are lost before it ever reaches the receiver.

3.2. Filling Factor (η)

While Q describes the relationship between energy storage and dissipation within the coil, another important metric is the filling factor which serves to characterize how effectively the coil's magnetic energy is coupled into the sample itself.

To quantify how effectively a coil deposits (or picks up) magnetic energy in the sample, it is possible to define the filling factor (η) as the fraction of the total RF magnetic energy in the sample over total RF magnetic energy stored by the coil itself. The filling factor is given by:

$$\eta = \frac{\int_{\text{sample}} |\mathbf{B}_1(\mathbf{r})_{\perp}|^2 dV}{\int_{\text{all space}} |\mathbf{B}_1(\mathbf{r})|^2 dV}, \quad (\text{I.90})$$

where $(|\mathbf{B}_1(\mathbf{r})_{\perp}|^2)$ is the energy density of the transverse magnetic field in the sample, $(|\mathbf{B}_1(\mathbf{r})|^2)$ the total magnetic field energy density of the coil dV is the differential volume element.

3.3. Resonator Sensitivity

In practical terms, both (Q) and (η) play complementary roles in determining the overall sensitivity and efficiency of an RF coil. Whereas (Q) governs how well the coil preserves oscillating energy, (η) dictates how much of the coil's magnetic field interacts meaningfully with the sample. Consequently, for a given input power, a higher quality factor (Q) leads to a stronger magnetic field $(\mathbf{B}_1)$ because less energy is dissipated as heat. Similarly, a higher filling factor means that a greater proportion of this generated $(\mathbf{B}_1)$ field is effectively concentrated within the sample, strengthening the magnetic interaction between the coil and the sample.

This can further be reflected in the proposed definition of the SNR. Starting from the effective resistance expressed via the resonator's quality factor, with Q_{ℓ} the sample loaded quality factor,

$$R_{\text{eff}} \approx R_c + R_s = \frac{\omega_0 L}{Q_{\ell}}, \quad (\text{III-91})$$

it follows that the signal-to-noise ratio can be written as:

$$\text{SNR} = F^{-1/2} (B_1/I) M_{\perp} \delta V \sqrt{\frac{\omega_0 Q_{\ell}}{4 k_B T_{\text{eff}} L}} \sqrt{\frac{t_o t_{\text{acq}}}{T_R}}. \quad (\text{III-92})$$

with the sensitivity (S_{RF}) :

$$S_{RF} = (B_1/I) \sqrt{\frac{\omega_0 Q_{\ell}}{4 k_B T_{\text{eff}} L}} \propto \sqrt{\frac{\eta Q_{\ell}}{T_{\text{eff}}}}, \quad (\text{III-93})$$

The relationship between coil current (I) and the produced RF field $(\mathbf{B}_1)$ is obtained through the expression of the magnetic energy stored in the coil (assuming a uniform magnetic field across the sample volume (V_s)),

$$\frac{1}{2} L I^2 = \int_{\text{coil}} \frac{B_1^2}{2\mu_0} dV = \frac{1}{\eta} \int_{\text{sample}} \frac{B_1^2}{2\mu_0} dV = \frac{V_s}{2\mu_0 \eta} B_1^2. \quad (\text{III-94})$$

Giving,

$$\frac{B_1}{I} = \sqrt{\frac{\eta \mu_0 L}{V_s}}. \quad (\text{I.95})$$

Substitution into the SNR expression yields the compact form expressed as a function of (Q_{ℓ}) and (η) ,

$$\text{SNR} = F^{-1/2} M_{\perp} \delta V \sqrt{\frac{\eta \mu_0 \omega_0 Q_{\ell}}{4 k_B T_{\text{eff}} V_s}} \sqrt{\frac{t_o t_{\text{acq}}}{T_R}}, \quad (\text{I.96})$$

The coil sensitivity (S_{RF}) is then defined by:

$$S_{RF} \propto \sqrt{\frac{\eta Q_{\ell}}{T_{\text{eff}}}}, \quad (\text{I.97})$$

For a given sample volume, optimal RF coil designs will seek to maximize both (η) and (Q_{ℓ}) and minimize (T_{eff}), ensuring minimal energy loss while maximizing magnetic field overlapping with the region of interest to increase sensitivity.

3.4. Noise Dominance Domain Considerations

The relative contributions of sample and coil loss to the system noise can be quantified using the Noise Power Ratio (NPR) (or noise factor), defined as the ratio of SNR power with and without the coil noise contributions (lossless) [109]:

$$\text{NPR} = \frac{\text{Lossless Coil SNR}_{pwr}}{\text{Lossy Coil SNR}_{pwr}} = \left(\frac{R_s T_s + R_c T_c}{R_s T_s} \right) = 1 + \frac{R_c T_c}{R_s T_s} \quad (\text{III-98})$$

The coil noise figure, which characterizes the noise degradation introduced by the coil and sample loading, is then given by:

$$NF_{\text{coil}} = 10 \log_{10}(\text{NPR}). \quad [\text{dB}] \quad (\text{III-99})$$

In the coil-noise-dominated regime ($R_s T_s / R_c T_c \ll 1$), the sample introduces minimal loss, so the loaded quality factor Q_{ℓ} remains close to the unloaded value Q_u . Because the sample contributes little noise, the noise figure of the coil becomes large ($NF_{\text{coil}} \gg 0$ dB), indicating that most of the system noise originates from the coil itself. In this regime, SNR scales approximately with $\sqrt{Q_{\ell}} \approx \sqrt{Q_u}$, and improvements will have to rely on reducing coil resistance by using better conductors, adapting geometry, or applying cryogenic cooling.

On the other hand, in the sample-noise-dominated regime ($R_s T_s / R_c T_c \gg 1$), the sample causes significant energy loss, which lowers Q_{ℓ} and dominates the noise contribution. The coil noise figure in this case is close to 0 dB, since the coil adds little noise beyond that of the sample. SNR scales with $\sqrt{Q_{\ell}}$, and further optimizing the coil has little effect. Instead, reducing sample loading through dielectric matching, geometry changes, or minimizing conductive material becomes key to enhancing sensitivity. The coil is no longer the limiting factor.

In clinical MRI, it is good practice to always aim for a sample-noise-dominated regime. However, in the case of mid-range coils and small samples, coil induced losses more often dominate [100], [110]. This further emphasizes that the first approach to tackling SNR for small sample experiments should be RF coil optimization as coil performance will be the limiting factor.

3.5. B_1 Field Homogeneity

To assess the uniformity of the RF coil (B_1) field, three primary metrics are commonly used: min/max-based homogeneity, the coefficient of variation (CoV) and the field histogram method.

3.5.1 Min/Max-Based Homogeneity

This metric measures the evenness of the coil's field by comparing its weakest and strongest points within a specified ROI:

$$H_{\text{uni}} = 1 - \frac{B_{\text{max}} - B_{\text{min}}}{B_{\text{max}} + B_{\text{min}}} = \frac{2 B_{\text{min}}}{B_{\text{max}} + B_{\text{min}}} \quad (\times 100 \%), \quad (\text{III-100})$$

where B_{max} and B_{min} are the maximum and minimum field values respectively. This homogeneity assessment is useful for quick checks of “hot” and “cold” spots. MRI coil specifications often require $\geq 90\%$ homogeneity over the specified imaging volume [104].

3.5.2 Coefficient of Variation (CoV)

The Coefficient of Variation (CoV) quantifies the overall field consistency by considering the standard deviation relative to the mean across all sampled points, not just the extremes. Given N data points (B_i) within the ROI, the mean (μ) and standard deviation (σ) are calculated:

$$\mu = \frac{1}{N} \sum_{i=1}^N B_i, \quad \sigma = \sqrt{\frac{1}{N} \sum_{i=1}^N (B_i - \mu)^2}. \quad (\text{III-101})$$

The CoV is then:

$$\text{CoV} = \frac{\sigma}{\mu} \quad (\times 100\%) \quad (\text{III-102})$$

Unlike min/max homogeneity assessments, the CoV accounts for the full spatial distribution and is less sensitive to single outliers than min/max methods. A CoV under 5% is often considered very homogeneous [111].

3.5.3 Field Histogram Method

Proposed by Yang et al. in 1994 [112], this method provides a comprehensive visualization of the entire field distribution within the ROI. Instead of summarizing the field with one or two numbers, it generates a plot that reveals the global field characteristics. The process involves:

1. **Calculating Percentage Field Deviation (PFD):** Each point's field value (B_1) is normalized relative to the mean or center field value (B_{1c}) to show its deviation in percent:

$$\text{PFD} = \frac{B_1 - B_{1c}}{B_{1c}} \times 100\% \quad (\text{III-103})$$

2. **Generating a B_1 -Weighted Histogram:** A histogram of the PFD values is created. Crucially, the contribution of each data point to the histogram is weighted by its own B_1 field magnitude which gives more significance to regions with stronger fields.
3. **Normalization (NPD):** The resulting histogram is normalized so that the total area under the curve is 1. The vertical axis is the Normalized Population of Deviation (NPD).

The shape of the final plot is a direct measure of homogeneity: a tall, narrow peak centered at 0% PFD indicates a highly uniform field, while a broad, flat, or multi-peaked distribution signifies poor homogeneity.

3.5.4 Choosing the Right B_1 Field Homogeneity Metric

Table 5 summarizes a comparison between the min/max homogeneity, Coefficient of Variation (CoV) and Field Histogram methods. Although requiring less computation, min/max homogeneity metrics can only give a quick snapshot of worst-case uniformity, while CoV provides a richer, statistically robust measure of overall field consistency. However, the field histogram method is in class of its own as it offers the most detailed diagnostic insight. It allows for the visualization of the entire field distribution, making it possible to identify not just the presence of inhomogeneity but also its character (e.g., a gradual drop-off vs. distinct regions of different field strengths etc...). The choice of metric will have to depend on the application's tolerance for localized "hot" or "cold" spots (Min/Max) against its need for overall field stability (CoV) or if a deeper understanding of the field's spatial behavior is needed (Field Histogram).

Criterion	Homogeneity (Min/Max)	CoV	Field Histogram
Output Type	Single percentage value.	Single percentage value.	2D plot (NPD vs. PFD).
Primary Use Case	Quick pass/fail assessment; identifying worst-case performance.	Standardized statistical reporting; comparing overall coil quality.	Detailed coil design; troubleshooting inhomogeneity sources.
Diagnostic Value	Low. Identifies the existence of "hot" and "cold" spots but not their extent or distribution.	Medium. Provides a single, robust number for overall variance but hides spatial information.	High. Visualizes the full distribution, revealing skew, multiple peaks, and the nature of the inhomogeneity.
Robustness	Low. Highly sensitive to single-pixel outliers or noise spikes which can drastically skew the result.	High. Statistically robust as it's based on the mean and standard deviation of all points.	High. Individual outliers have minimal impact on the overall shape of the distribution.
Computational Cost	Very low. Requires only finding the minimum and maximum values.	Low. Requires calculating the mean and standard deviation of the ROI.	Low to Medium. Requires binning data and calculating weighted sums, but is still very fast on modern hardware.
Data Requirement	Only the two extreme values in the ROI.	All data points within the ROI.	All data points within the ROI.

Table 5: Detailed comparison of coil field uniformity metrics: min/max homogeneity, Coefficient of Variation (CoV), and Field Histogram.

III.D. Benefits of a 3D Plastronics Approach to Coil Fabrication

Ultimately, to achieve greater sensitivity through the RF coil, every aspect of the performance metrics presented previously has to be taken into account. An optimal coil needs to maximize both the filling factor (η) and the quality factor (Q) while minimizing the effective temperature (T_{eff}). Traditional coil fabrication methods, which often rely on manual winding of copper wire or etching of flat copper sheets, face significant limitations in achieving these goals, especially for the complex, small-scale applications found in tissue engineering. These methods offer limited geometric freedom, making it difficult to create coils that conform perfectly to irregularly shaped samples. Furthermore, integrating additional functionalities, such as supports, perfusion circuits, or sensors, is often a cumbersome process involving mechanical assembly of multiple separate parts.

1. Generalities on 3D Plastronics

3D Plastronics, offers an interesting alternative by merging the structural and electronic aspects of the device into a single, integrated component. This approach provides a powerful toolkit for designing and fabricating RF coils that are not only optimized for performance but are also tailored to the specific demands of laboratory-based MRI for tissue engineering.

Fundamentally, 3D Plastronics refers to the integration of electronic circuits and components directly into shaped polymer parts, using the polymer substrate itself as the carrier for electrical pathways. In this approach, a single component combines mechanical structure and electronic functionality. Related concepts include Molded Interconnect Devices (3D-MID) and In-Mold Electronics (IME); however, 3D Plastronics has emerged as the broader term favored by industry standards.

Key technical steps in 3D Plastronics involve [113]:

1. Shaping the polymer substrate into a 3D form.
2. Metallizing selected areas of the surface to create conductive traces.
3. Attaching and connecting surface-mount devices (SMDs) directly onto these conductive paths.

Unlike traditional flat printed circuit boards (PCBs), 3D Plastronics devices feature circuitry patterned along complex, curved surfaces. These conductive patterns not only interconnect components such as LEDs, sensors, or capacitors but may also serve as functional elements themselves, acting as antennas[114], coils[115], or heating elements[116], for example.

This integration blurs the classic separation between a device's housing and its electronics: in 3D Plastronics, the polymer structure "is" the electronic carrier. For instance, a control panel no longer requires a separate PCB mounted behind it; instead, the circuit is embedded directly onto or into the molded panel itself, streamlining the design and eliminating bulky connectors and wiring.

While flexible circuits offer some degree of bending and shaping, they are fundamentally planar by nature. In contrast, 3D Plastronics enables direct functionalization of fully 3D and intricately curved geometries [115].

By merging the mechanical properties of polymers with embedded electronic functionality, 3D Plastronics enables the creation of lightweight, compact, and highly aesthetic components. Designs can become thinner, lighter, and more organic, eliminating the need for traditional mechanical assemblies like buttons, wires, and connectors. An example is presented in Figure 22. This not only simplifies product architecture by reducing the number of parts and assembly steps, but also

opens up new possibilities such as wrap-around antennas and coils, curved touch-sensitive surfaces, and integrated sensor systems.

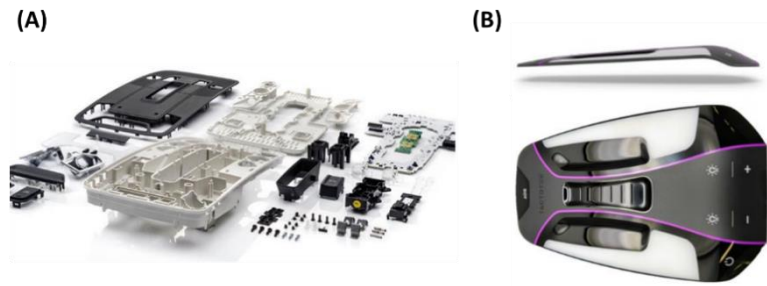


Figure 22 : Comparison between a conventional assembly (A) and an IME part (B): the IME version is 77% lighter (150 g vs. 650 g), 93% thinner (3 mm vs. 45 mm), has 96% fewer mechanical parts (2 vs. 64), and the printed circuit board is 25% smaller (10×3 cm vs. 10×4 cm), adapted from [117]).

Because 3D Plastronics devices bring together electronic circuits and polymer structural parts (Figure 23), a 3D Plastronics approach allows for a wide range of material selection. Both the polymer substrates and the electronic conductive materials can be selected to meet application-specific requirements (biocompatibility, electrical performance, thermal stability, processability, etc.). These materials are often shaped using techniques conventionally used in the polymer industry before being metallized through various means. Moreover, 3D Plastronics is not limited only to polymer substrates. Ceramic substrates can also be functionalized through the same process greatly expanding the available substrate selection while offering enhanced properties such as mechanical robustness, and high thermal stability and conductivity for more demanding applications. When ceramic-based substrates are used, the process is referred to as “3D Ceratronics”.

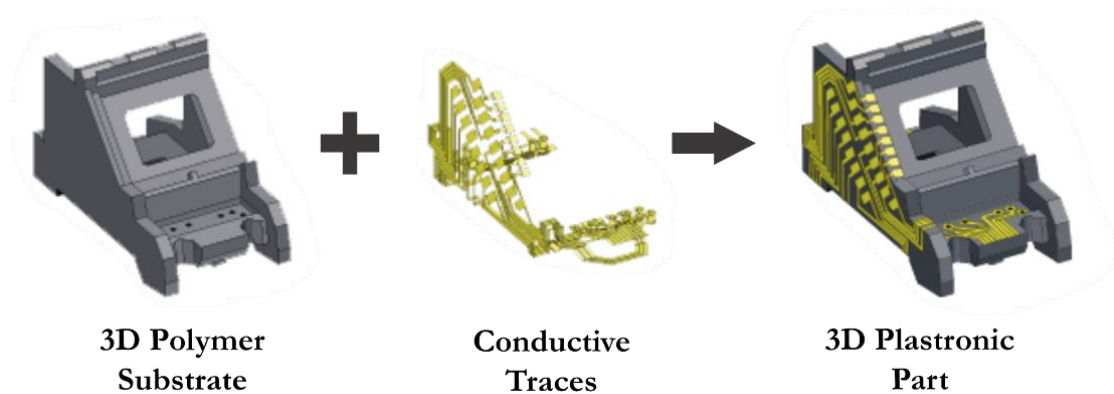


Figure 23 :Example of a 3D Plastronics device: integration of a 3D-molded polymer substrate with surface-mounted electronic circuitry, combining structural and electronic functions in a single component (adapted from [118]).

In the context of MRI applications, it has already been shown that 3D Plastronics, through its integration capabilities, has enabled coil performance optimization and complex multifunctional assemblies [119],[120].

As 3D Plastronics enables unparalleled design freedom coupled with the ability to place RF coils on any 3D surface, it is a manufacturing method perfectly suited to filling factor maximization. For example, Zamarayeva et al. [119] used 3D Plastronics to fabricate a personalized, conformal receive-coil array in the shape of a neck brace, spray-coated with a

silver-based conductive layer (see Figure 24-A). Compared to a commercial four-channel Siemens neck coil, this fabricated coil enabled up to a 500% increase in peripheral SNR and a 40% increase in central SNR. These results clearly illustrate how 3D Plastronics enables dramatically improved RF coil performance through customizable, form-conforming geometries designed to maximize filling factor.

Another non negligible benefit of using 3D Plastronics is the ability to integrate a large number of functions into a single compact device. Gerges et al.[120] illustrates such capabilities thoroughly. In a single device dedicated to small animal imaging, they have managed using 3D Plastronics to integrate a dual-loop Helmholtz coil with onboard capacitors for imaging, a heated sample/mouse bed with embedded fluidic warming channels, an adjustable external coupling drive, and an anesthesia gas manifold, all mounted on a three-axis translation stage for precise isocenter alignment (see Figure 24-B).

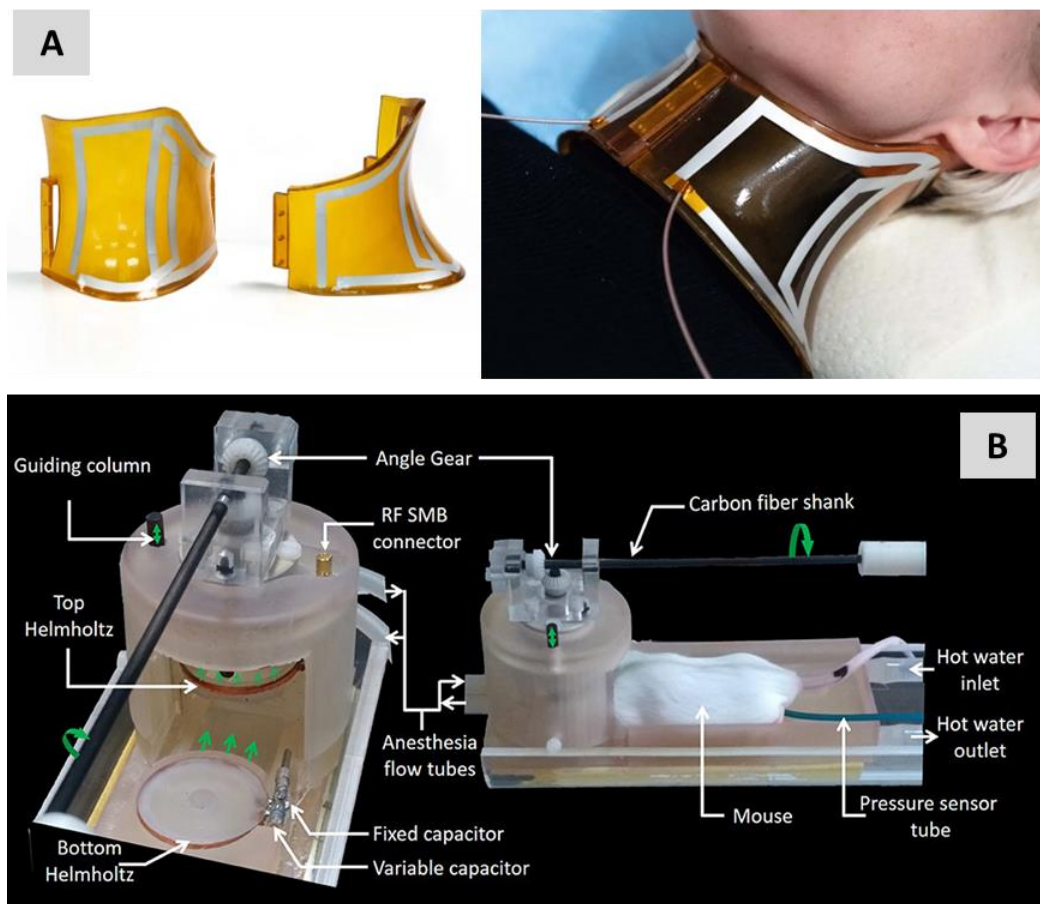


Figure 24: A) Example of a personalized, conformal receive-coil array in the shape of a neck brace, spray-coated with a silver-based conductive layer used to maximize SNR [119]. B) Example of a complex 3D Plastronics setup incorporating multiple functions including an imaging coil a tuning system, a heated sample/mouse bed with embedded fluidic warming channels and an anesthesia gas manifold mounted on a three-axis translation stage for precise alignment [120].

In the context of a small-scale benchtop laboratory MRI for Tissue Engineering applications, both capabilities will be crucial. 3D Plastronics can be used as a means to tackle the inherent SNR limitations through filling factor optimization and integrate all the necessary functionalities to recreate *in-vivo* like conditions (fluidic system etc...) as well as potentially adding control sensors (pH, temperature, etc...).

2. Rapid 3D Plastronics (R3DP)

Among the many 3D Plastronics manufacturing techniques [121], “Rapid 3D Plastronics” [122], stands out as a particularly promising process presenting a robust and accessible solution for prototyping in a laboratory setting. Developed at the AMPERE Laboratory, this process combines 3D printing with a selective copper metallization technique tailored for rapid prototyping of electronic components on 3D surfaces. Figure 25 shows an example of a device manufactured using this process.

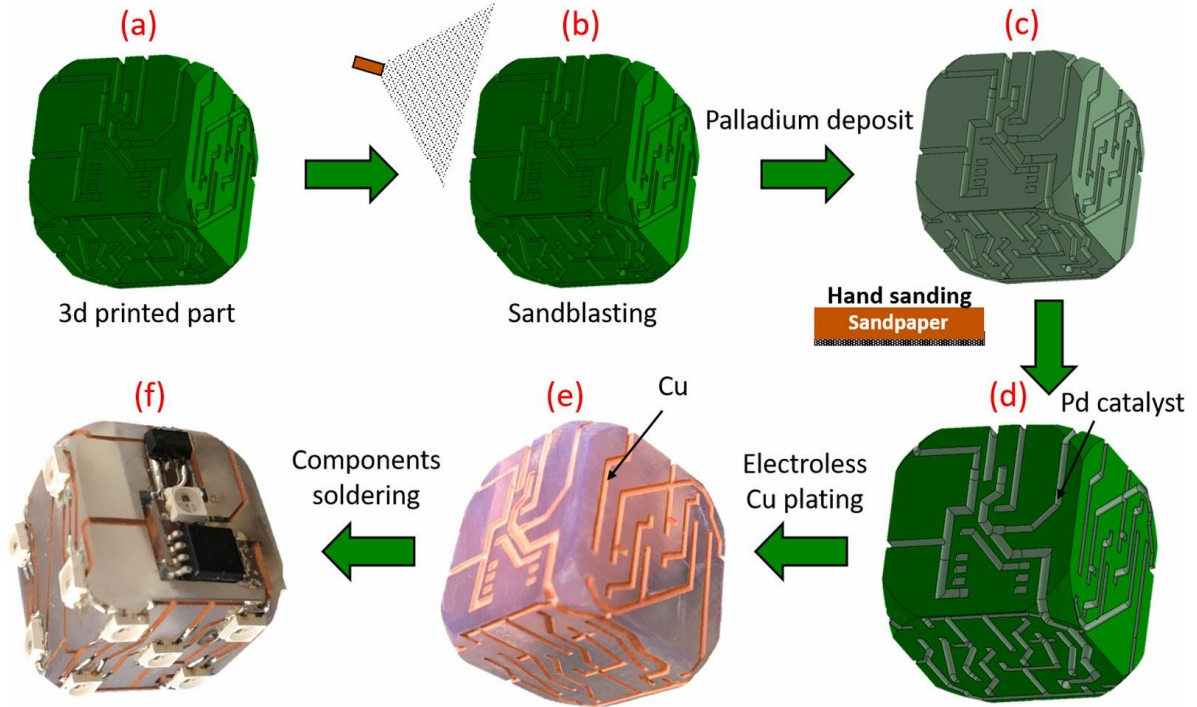


Figure 25: Rapid 3D Plastronics process: (a) 3D-printed part with integrated grooves fabricated via vat photopolymerization; (b) surface roughening by sandblasting to enhance metal adhesion; (c) uniform application of colloidal palladium catalyst; (d) selective retention of catalyst within grooves after surface sanding; (e) copper deposition by electroless plating; (f) finalized metallized 3D prototype [121].

With its low implementation cost, compatibility with standard 3D printing materials and the freedom to design application tailored devices, this method is well-suited for the iterative design and optimization cycles common in research environments. By relying on additive manufacturing (3D printing), it allows for the rapid creation of complex, custom-shaped substrates that can be perfectly tailored to the geometry of a specific tissue sample, directly addressing the challenge of maximizing the filling factor.

2.1. Material Selection and Forming

The first step in the R3DP process is material selection. Only materials compatible with the employed 3D printing process can be used. These processes include fused deposition modeling (FDM), selective laser sintering (SLS), and stereolithography (SLA).

There is a clear supremacy of thermoplastic polymers, since they can be reshaped multiple times by applying heat and pressure using suitable tooling. Their ability to soften without chemical change allows easy formation of complex 3D geometries through both the FDM and SLS techniques. In the FDM process, thermoplastic filaments are extruded through a heated nozzle and deposited layer by layer to create the 3D shape [123], while in the SLS process, a laser

selectively sinters powdered thermoplastics into solid layers without the need for support structures. This latter process allows complex shapes but often results in porous surfaces with higher roughness, which can hinder subsequent metallization [124].

Thermoplastic polymers are classified based on their thermal stability and mechanical performance, often represented by a pyramid (Figure 26): standard polymers like polylactic acid (PLA), polyethylene terephthalate (PET) and acrylonitrile butadiene styrene (ABS) lie at the base, while high-performance materials like polyether ether ketone (PEEK) and polyethyleneimine (PEI) occupy the top due to their superior thermal resistance and mechanical stability.

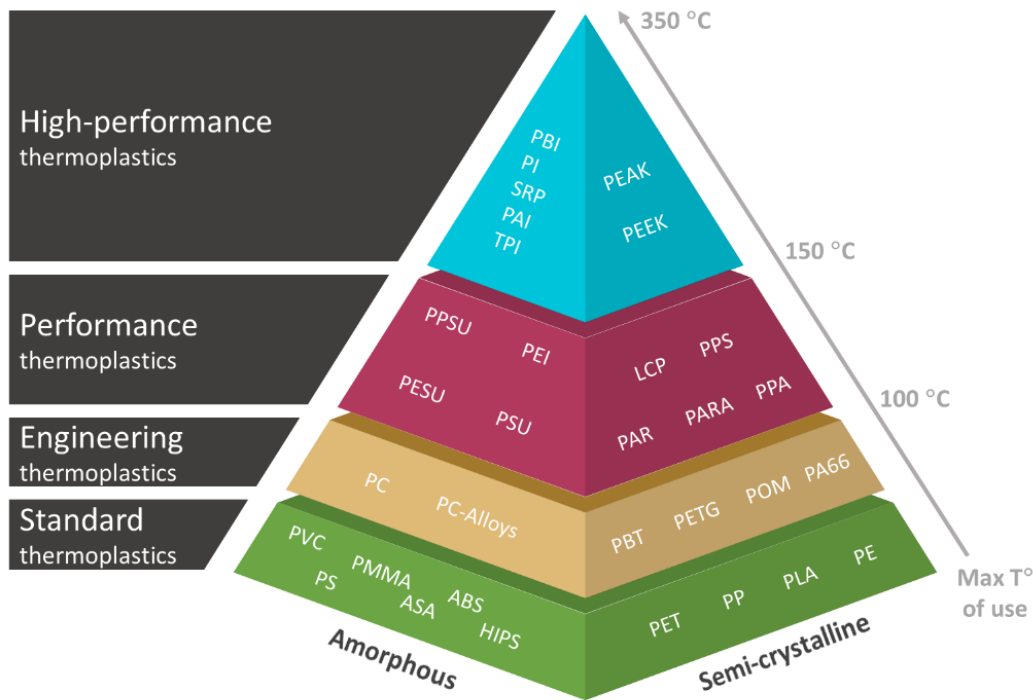


Figure 26: Pyramid of thermoplastic polymers, adapted from [125], [126].

In cases demanding exceptional performance, polymers like PEEK and PEI are selected. These high-end materials offer minimal thermal expansion, excellent chemical resistance, and the ability to endure high reflow temperatures, reducing stress between metal traces and the polymer substrate when soldering. However, their high cost often limits their use to specialized or high-reliability applications. Many thermoplastics used in 3D Plastronics can also be enhanced with fillers [127]. These additives are incorporated to improve mechanical strength, thermal conductivity, fire resistance, or dielectric properties.

As an exception to the thermoplastics supremacy, SLA employs photopolymer resins, which are thermosets, cured layer by layer using a laser or projected light [128]. These resins can achieve high resolution and smooth surfaces. When using high-temperature formulations, such as Formlabs' High Temp Resin v2, SLA parts can even withstand soldering processes [129], [130]. However, they can be brittle and susceptible to long-term dimensional changes if post-curing is incomplete. Although composites and thermosets are less commonly used overall in additive manufacturing for 3D Plastronics, SLA stands out as a valuable method when thermal performance and surface quality are prioritized.

Moreover, FDM, SLS and SLA are all additive manufacturing methods compatible with ceramic printing [131]. This means, among other things that, that the R3DP process can also be used to create 3D Ceratronics variants if the need arises. While often requiring specialized 3D printers some companies like Formlabs offer compatible solutions to their generic flagship

printers [132]. However, these products still require complex, multi-day workflows and extra equipment, including ovens for sintering, making them still more costly to deploy. Nonetheless, if the application requires ceramic properties, it is worth considering.

2.2. Metallization

Following substrate forming, the next essential phase in the R3DP process is the metallization of the surface to create conductive pathways and circuitry. A defining feature of 3D Plastronics is that the substrates are inherently non-conductive. This property prevents the direct use of electrodeposition techniques, which require a conductive base material. To overcome this limitation, a preliminary conductive "seed" layer must first be applied to the surface through electroless metallization. This method is the most widely adopted in industrial applications due to its ability to produce uniform, conformal coatings on complex 3D geometries.

2.2.1 Electroless deposition (ELD)

Electroless deposition (ELD) enables the chemical reduction and deposition of a metal layer onto a non-conductive surface without the need for electrical current. A major advantage of ELD is its ability to metallize electrically insulating materials such as polymers or even ceramics, which are common in 3D Plastronics and other advanced applications.

Unlike classical electrodeposition, which requires an electrically conductive base and is sensitive to variations in surface geometry due to its reliance on current density, ELD can be uniformly applied across both flat and contoured surfaces. This geometric independence makes it especially valuable in applications where metal must be deposited uniformly on complex 3D forms, as shown in Figure 27. Because of this uniformity, ELD is often used not only to metallize insulating polymers but also to recoat metallic components with new layers of materials such as nickel-phosphorus or chromium. These applications are widespread in industries where consistent layer thickness across curved or irregular surfaces is critical.

In contrast, electrodeposition produces coatings that vary in thickness according to the electrical field distribution and surface topology, requiring intricate process controls to achieve uniform results. This makes ELD a more homogeneous technique, and one better suited for the precision demands of 3D Plastronics systems.

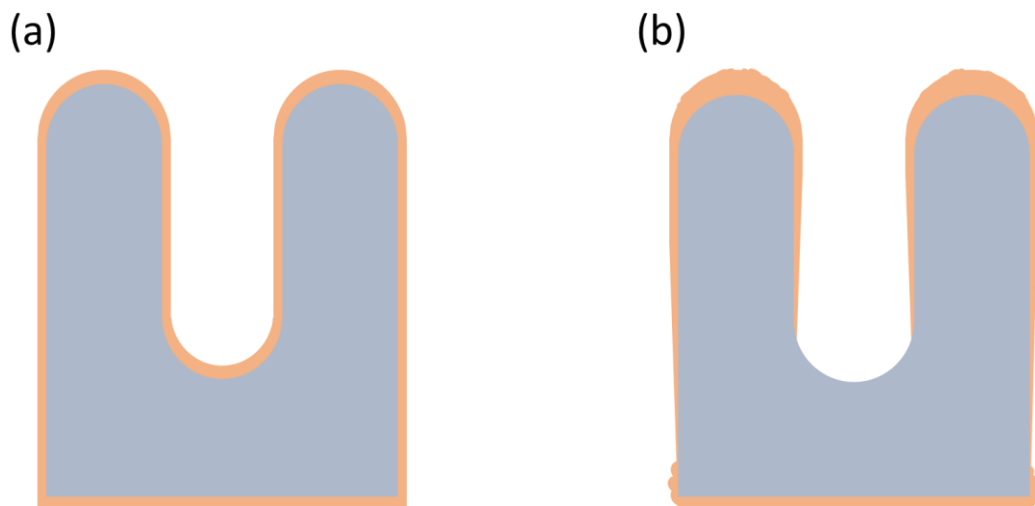


Figure 27: Metal thickness in the case of ELD (a) and electrodeposition (b).

The electroless deposition (ELD) process is based on a redox reaction that occurs only in the presence of a catalytic surface (typically palladium) and an appropriate reducing agent. To ensure

proper adhesion and controlled metal growth, the process is divided into three distinct stages, each carried out in a separate chemical bath:

1. **Surface preparation:** A chemical treatment is applied to modify the polymer surface, increasing its roughness and reactivity to promote adhesion in subsequent steps.
2. **Surface activation:** The treated surface is exposed to a catalytic solution, often containing palladium-based colloids, which deposits active catalytic sites onto the polymer.
3. **Metal deposition:** The activated surface is immersed in a plating solution containing metal ions and a reducing agent. A redox reaction occurs at the catalytic sites, leading to the uniform deposition of a metal layer on the substrate.

Once the initial copper layer forms, it becomes autocatalytic: the deposited copper surface can continue catalyzing the redox reaction, allowing continuous metal growth as long as the chemical balance is maintained. This makes ELD highly effective for covering irregular surfaces and internal cavities without the need for electrical contacts. This behavior, however, only occurs when so-called “auto-catalytic metals” are involved [133], [134]. ELD is therefore limited to such metals which include: Fe, Ru, Os, Co, Rh, Ir, Ni, Pd, Pt, Cu, Ag, and Au.

In practice, electroless metallization offers several significant advantages. Chief among them is its ability to deposit metal uniformly over intricate 3D surfaces, regardless of curvature or orientation. The process also ensures strong adhesion, particularly when preceded by proper etching, and does not require an external power source, simplifying setup. Furthermore, the thickness of the deposited layer (typically between 6 and 8 μm) can be precisely controlled by adjusting the duration of immersion in the plating bath. The autocatalytic growth process only lasts until a certain thickness is achieved ($\sim 5 \mu\text{m}$). This can be a limiting factor in applications such as RF coil fabrication, where trace thickness is directly linked to ohmic losses and performance. Furthermore, it has been shown that the deposited thin electroless copper boasts a limited conductivity of around 74% the conductivity of bulk copper [122].

2.2.2 Electrodeposition as a Follow-Up to ELD

While electroless copper provides adequate conductivity for many applications, certain 3D Plastronics systems, particularly those handling high current or operating at radio frequencies, require lower resistance and increased conductor thickness. In such cases, electrodeposition (or electroplating) is employed as a secondary step to increase the thickness of the copper layer beyond what ELD alone can provide [120], [135].

Electrodeposition involves immersing the metallized part in an electrolyte containing copper ions and applying an external electric current as presented in Figure 28. The electroless copper layer, deposited during the ELD step, serves as the conductive base needed for this current-driven process. As copper ions migrate toward the negatively charged surface, they are reduced and deposit as metallic copper, gradually building up the thickness of the trace.

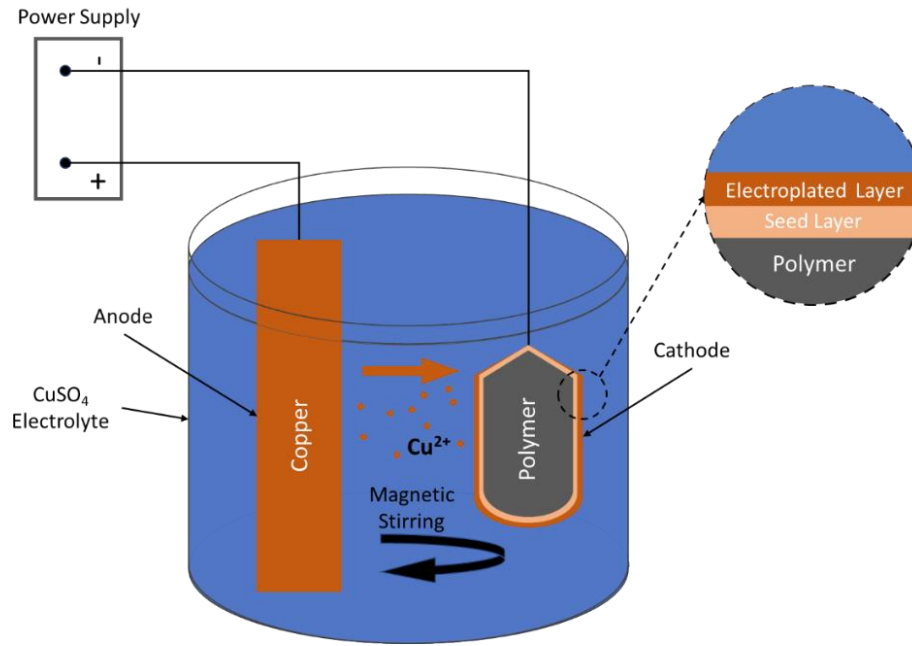


Figure 28: Schematic diagram of the electroplating process

However, unlike ELD, electrodeposition is sensitive to geometry. The distribution of current density across a complex 3D surface can vary significantly, leading to non-uniform metal thickness (particularly on edges, corners, or recessed areas). Nonetheless, when properly controlled, electrodeposition allows for thick, low-resistance copper traces that meet the demands of high-performance 3D Plastronics applications. Electroless copper generally has lower purity (around 99.2%) compared to electroplated copper (which can reach 99.9%), leading to a slightly higher resistivity of 0.020–0.025 $\mu\Omega\cdot\text{m}$, versus 0.017 $\mu\Omega\cdot\text{m}$ for bulk copper [136].

2.3. Trace Patterning

Once a metal layer has been deposited by one of the previously described methods, it must be patterned to form functional circuit traces. Two main strategies are used to define these patterns: “subtractive” and “additive” methods.

In the subtractive approach, the entire surface is uniformly metallized, and unwanted areas are subsequently removed. Techniques for this include laser ablation, micro-milling, or wet chemical etching using a patterned resist or removable adhesive mask [120]. By contrast, the additive approach deposits conductive material only where needed, avoiding the need for removal.

The choice between additive and subtractive techniques depends on the required resolution, mechanical complexity, production volume, and cost. In the R3DP process, subtractive techniques are generally employed. The favored methods include either surface sanding after metallization of predefined grooves [122] or masking coupled with chemical etching [120].

2.4. Surface Mounted Devices (SMD) Placement and Connection

A critical aspect of 3D Plastronics device fabrication is establishing reliable electrical connections between surface-mounted devices (SMDs) and the metallized traces. The choice of connection method is largely determined by the thermal stability of the underlying polymer substrate (see Figure 26). For polymers that can tolerate elevated temperatures, conventional soldering techniques, often using standard or low-temperature solder alloys, are generally preferred due to their superior electrical conductivity and long-term reliability.

However, when the substrate cannot withstand the thermal stress of soldering, alternative approaches such as conductive adhesives must be used [137]. While these materials allow for low-

temperature bonding, they typically offer lower conductivity and mechanical robustness compared to soldered joints. This limitation becomes especially significant in RF applications, as this introduces significant resistive losses through poor connection quality and directly affects signal performance and system efficiency.

IV. Consolidating a Practical Approach Towards Tissue Engineering Applications

The successful advancement of tissue engineering hinges on the ability to non-invasively monitor and characterize the development of three-dimensional constructs over time. As established in the preceding sections, this creates a fundamental requirement for the seamless integration of a life-sustaining environment with a robust imaging tool. To maintain sample viability for the longitudinal studies essential to understanding tissue maturation, the imaging apparatus must be compatible with, or ideally incorporated into, a culture chamber or bioreactor system. This integration ensures that the engineered tissue remains in a controlled, nourishing environment throughout the characterization process.

Having identified low-field benchtop MRI as the most promising imaging modality due to its practicality and versatility, the primary challenge becomes overcoming its inherent SNR limitations. While several optimization strategies exist (such as increasing acquisition time, voxel size, or static field strength), they introduce unacceptable compromises in throughput, resolution, or cost-effectiveness. Therefore, enhancing RF coil sensitivity emerges as the most effective and practical path forward.

The RF coil is the direct interface for signal transmission and reception. In the context of small-sample, high-resolution imaging, where system noise is often dominated by the coil itself, optimizing the coil's design offers the most significant potential for SNR improvement. Possible strategies to optimize the imaging tool through RF coil sensitivity include:

1. Optimizing Coil Geometry and Materials:

- **Maximizing the Filling Factor (η):** The filling factor quantifies how effectively the coil's magnetic field is coupled to the sample. A higher filling factor is achieved by designing coils that conform closely to the sample's geometry, ensuring that more magnetic energy is concentrated in the region of interest. This involves creating custom-fit coils tailored to the specific dimensions of a scaffold.
- **Minimizing Electrical Losses:** The coil's efficiency, or Q-Factor, is improved by systematically minimizing all sources of electrical loss. This includes using highly conductive materials (e.g., copper) and optimizing the conductor's geometry to counteract frequency-dependent "skin" and "proximity" effects.
- **Field Homogeneity:** A homogeneous $\mathbf{B}_1$ field is critical for enabling accurate quantitative measurements. If the field is not uniform, different parts of the sample will experience different excitation levels leading to signal variations that do not reflect the underlying tissue properties. Therefore, coil design must balance a high filling factor with a geometry that ensures a sufficiently homogeneous $\mathbf{B}_1$ field over the entire sample.

2. Employing Advanced Fabrication Methods:

Traditional, manual fabrication methods can be a bottleneck for creating coils with complex, optimized form fitting geometries. Modern techniques like 3D Plastronics, which combine the precision of electronics with the flexibility of additive manufacturing, allow for the creation of integrated devices. This enables the fabrication of a culture chamber with a built-in, perfectly-fitted RF coil, representing the ultimate optimization of the filling factor and streamlining the entire experimental workflow.

3. **Employing Cryogenic Cooling:** This is one of the most powerful strategies for boosting SNR. Reducing the physical temperature of the coil conductors to cryogenic levels (e.g., using liquid nitrogen) improves sensitivity through a dual mechanism:

- It drastically lowers the conductor's electrical resistance, which significantly increases (Q).
- It directly reduces the effective noise temperature (T_{eff}) of the system.

Since SNR is proportional to (Q/T_{eff}) , the combined effect provides a substantial improvement. In the "coil-noise-dominated" regime, which is common for small samples, this strategy can yield the most drastic gains.

By focusing on these targeted hardware innovations, we hope to leverage the accessibility of a benchtop MRI system and overcome its primary limitations. Designing and fabricating high-performance RF coils specifically tailored for small biological samples will dramatically enhance sensitivity, enable high-resolution, quantitative imaging of viable engineered tissues and establish benchtop MRI as an indispensable tool for the field.

Chapter II: Optimal Homogenous RF Resonators for Small Sample Imaging Using a Benchtop MRI scanner

Chapter Overview

Chapter II builds upon the need for high-performance RF coils by undertaking a systematic, simulation-driven design and optimization process to identify the ideal resonator topology for small-sample imaging on a low-field benchtop MRI scanner. The chapter begins by establishing the unique design challenges inherent to this environment, where the interplay between the RF coil, the sample, and the mandatory local RF shielding creates a complex electromagnetic system in which all sources of signal loss are significant.

The investigation begins with a review of the theoretical foundations for generating homogeneous magnetic fields, exploring both cylindrical and spherical current distributions. These principles are then applied in a preliminary analytical comparison of various coil geometries, including the classical solenoid, varying-pitch solenoid, loop-gap resonator, and different spherical coil approximations. This initial screening, based on idealized models, serves to evaluate the intrinsic B_1 field homogeneity of each design and identify the most promising candidates for further analysis.

From this first-order comparison, the classical solenoid and simpler spherical approximations are set aside, while the Varying-Width Ribbon Solenoid (VWRS), the Bridged Loop-Gap Resonator (BLGR), and the extended Hoult & Deslauriers (H&D) spherical coil emerge as the most viable topologies. The chapter then transitions from idealized theory to a more rigorous, realistic evaluation of these select designs.

This advanced analysis employs comprehensive 3D electromagnetic co-simulations that incorporate the full complexity of the operational environment, including the RF shield, a dielectric sample load, and the non-ideal characteristics of lumped electronic components. This crucial step moves beyond theoretical homogeneity to predict real-world performance metrics such as loaded quality factor (Q_l), sensitivity, and noise dominance.

Finally, the chapter culminates in a direct comparative assessment of the simulated performance of the final candidates. The results are synthesized to determine the optimal coil topology that delivers the best balance of high B_1 field homogeneity and maximum sensitivity.

I. Introduction: The Imperative for Simulation-Driven Design

As established in the preceding chapter, the primary obstacle to the widespread use of low-field benchtop MRI systems in laboratory settings is the inherent trade-off between their accessibility and performance. While these systems offer a practical and cost-effective solution for laboratory settings, their lower magnetic field strength inherently results in diminished SNR, which compromises image quality and extends acquisition times, particularly for high-resolution applications. The RF coil has been identified as the most critical and accessible component for substantial SNR enhancement, yet its design for low-field benchtop systems presents unique challenges.

Unlike high-field clinical systems, where the dominant source of loss is typically the lossy biological sample, or microcoil systems, where conductor and dielectric losses are minimal [110], benchtop configurations for small samples often operate in a mid-range electromagnetic regime (characterized by frequency–dimension products $f \cdot d \approx 2\text{-}30 \text{ MHz} \cdot \text{m}$). In this intermediate regime, all sources of loss, including coil conductor resistance, capacitive dielectric dissipation, sample-induced conduction losses, and eddy current damping from nearby shielding, contribute significantly to overall signal degradation.

A critical aspect of RF coil design for benchtop MRI systems is the incorporation of RF shielding. Shielding is essential to protect the sensitive MR receiver chain from external electromagnetic interference (EMI), and to minimize unwanted interactions between the RF coil and other metallic components of the scanner, such as the magnet or gradient coils [40], [99]. In compact benchtop systems, this shielding is typically "local," meaning it is placed near the RF coil, often forming part of the coil housing itself. While local shields are effective in containing RF fields, their proximity to the resonator can introduce significant challenges. The conductive shield material can lead to eddy current induction, which in turn reduces the coil's quality factor (Q), alters its resonant frequency and impedance, and can perturb the $\mathbf{B}_1$ field distribution. Therefore, the RF coil and its local shield cannot be designed in isolation. They constitute an electromagnetically coupled system that requires careful co-optimization to balance effective shielding with minimal performance degradation of the resonator. Consequently, achieving optimal RF coil performance, primarily aimed at maximizing SNR, requires a comprehensive optimization strategy that meticulously accounts for this complex interplay of loss mechanisms.

Given this design complexity, computational simulation offers the most efficient and systematic path to optimizing coil geometry. Analytical models and mathematical formulations provide valuable initial estimates but are built on idealized assumptions, such as infinitely long or thin conductors, perfect geometric symmetries, and the absence of environmental interactions[104]. They struggle to accurately model the intricate details of real-world coils with complex 3D geometries. These models typically capture only some dominant effects such as current and field patterns while neglecting other interactions and higher-order phenomena such as proximity effects between closely spaced conductors, and electromagnetic coupling with samples or nearby shields. Although these can also be modeled analytically, they generally require stringent and time-consuming case by case studies. As such, their predictive power is often limited to first-order approximations where geometries are simplified, and they are insufficient for fine-tuned optimization of complex realistic coils.

Quantitative assessment of biological samples of inhomogeneous composition and irregular shapes has been acknowledged by Hoult and Deslauriers [101] to be very difficult. As previously exposed in Chapter I, the signal strength from each small element of the sample depends on the local $\mathbf{B}_1$ field and if the field varies spatially, different regions experience different flip angles, creating uncertainty in the interpretation of the perceived signal. Unless both the spatial $\mathbf{B}_1$ distribution and the intrinsic sample properties are known precisely at any given location, accurate

and physiologically relevant interpretation become difficult. These uncertainties are compounded when the sample moves or changes size, or when the coil's Q factor and tuning vary with sample size and composition, which is inevitable in tissue engineering applications where samples will grow from cell seeds to complex 3D structures. Using coils with excellent $\mathbf{B}_1$ field homogeneity, high sensitivity, and stable resonant behavior can help drastically alleviate these difficulties. Therefore, for large sample quantitative imaging, $\mathbf{B}_1$ field homogeneity will be essential to accurate physiologically relevant quantitative acquisitions.

Using a mixture of analytical calculations and numerical simulations, this chapter focuses the optimization and comparison of different homogenous volume coil topologies. Although highly sensitive surface coil designs might offer far greater sensibility locally [104], their use in the context of tissue engineering application will be limited to thin samples and constructs (~ 5 mm) requiring very little penetration (skin, 2D organoids etc...). As surface coil's $\mathbf{B}_1$ falls off rapidly with depth (approximately half of its diameter or faster) [104], they are inherently ill-suited for full volume large sample (~ 40 mm) characterization. Similarly, array coils and dissociated Rx/Tx setups will also fall outside the scope of this study as the targeted benchtop scanner is a single-channel system, and implementing an array would require extensive hardware modifications, such as the addition of multi-channel receiver hardware or a multiplexing switch bank.

Each coil will be optimized to maximize SNR and $\mathbf{B}_1$ field homogeneity across either a 38×38 mm cylindrical or a 19 mm radius spherical region of interest (depending on the topology of the coil). They will be tailored specifically for a low-field 0.3T benchtop MRI scanner from Pure Devices GmbH (Research Lab line), operating at a ^1H Larmor frequency of 13.03 MHz with a rated usable ROI of $40 \times 40 \times 60$ mm. The design is constrained by the scanner's physical dimensions, which limit the shielded coil housing to a maximum size of $180 \times 55 \times 220$ mm. The primary objective of RF coil optimization in this context will be the maximization of coil performance metrics while maintaining high $\mathbf{B}_1$ field homogeneity across the region of interest.

As exposed in the previous chapter, from a resonator perspective SNR is enhanced primarily by maximizing the quality factor (Q) and the filling factor (η), which together quantify the efficiency of magnetic coupling between the coil and sample, while maintaining noise from external interference to a minimum. However, these metrics will only have a significant impact on SNR in a coil-noise dominance domain where most of the losses are coil dependent. Field homogeneity, however, is governed by coil geometry and winding topology, which need to be selected and fine-tuned to deliver uniform RF fields. The $\mathbf{B}_1$ field homogeneity generally required in a RF coil design is defined as either having under $\pm 5\%$ [104] field variations or a coefficient of variation (CoV) of less than (5%)[111].

The aim of this chapter will be to identify the optimal locally shielded RF coil topology for small sample imaging applications on the aforementioned 0.3T benchtop MRI scanner.

II. Overview of the Selected Benchtop MRI Scanner

The benchtop scanner used throughout this thesis was a custom-ordered low-field 0.3T benchtop MRI scanner from Pure Devices GmbH's Research Lab line (Pure Devices GmbH, 300 mT Research Lab, Rimpf, Germany). This $310 \times 400 \times 270$ mm scanner, weighing approximately 130 kg, was selected for this study as its size and weight made it suitable for placement on standard laboratory surfaces. Furthermore, as it can be operated through a Universal Serial Bus (USB) interface and operated through MATLAB, it is ideal for advanced customization, including integration of user-defined measurement sequences and custom hardware attachments.

The scanner was equipped with a 300 mT permanent neodymium magnet configured in a conventional C-shaped open design (Figure 29.A). This configuration featured a 60 mm magnet gap with a field homogeneity of less than 50 ppm within a maximum field of view (FOV)

measuring $40 \times 60 \times 60$ mm. Grounded support guard rails were integrated into the magnet structure to facilitate the insertion of a shielded housing (Figure 29.C) in either a vertical or horizontal orientation, thus accommodating either vertical or horizontal bore alignment. The magnet is maintained at a stable temperature of 29°C using a temperature stabilization unit provided by the manufacturer and connected to the yoke to ensure a ^1H Larmor frequency of 13.03 MHz.

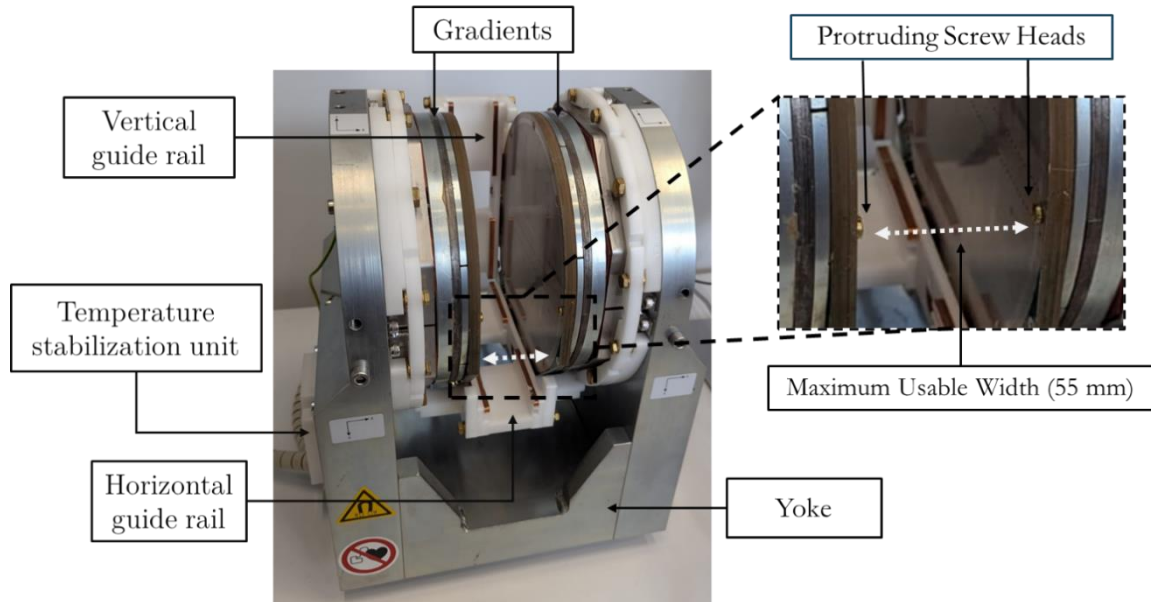


Figure 29: Open C-shaped permanent magnet of the 0.3T benchtop MRI scanner, featuring a 60 mm magnet gap (with a 55 mm usable width).

The system incorporates a 3D gradient system coupled with a High-Power Gradient Amplifier (DC 600, Pure Devices GmbH, Rimpf, Germany), enabling imaging in x-, y-, and z-directions, with a maximum gradient strength of 200 mT/m. Interestingly, the smaller bore, suited for small-sample imaging, allows the use of more powerful gradient coils compared to those found in clinical machines (~ 20 mT/m). The radiofrequency coils are connected to the transmit/receive switch inside the scanner's RF amplifier (RF-200, Pure Devices GmbH, Rimpf, Germany) output box using a SMB-female to SMB-female 2.5-meter cable (RG316) wrapped around a ferrite filter with 4 turns to suppress unwanted high-frequency noise and common-mode currents.

III. Theoretical Foundations for Homogenous RF Resonators and Preliminary Analysis

Generating a homogenous $\mathbf{B}_1$ field can be achieved one of three ways [104] :

- By running homogeneous current tangentially on the surface of an infinitely long cylinder to create a perfectly homogeneous magnetic field parallel to its axis. (Figure 30-A1)
- By using a uniform distribution of loop currents on a sphere creating surface current density following a cosine distribution. (Figure 30-A2)
- By creating a cosine dependent distribution of line currents, flowing parallel to a cylinder axis. (Figure 30-B)

As the generated magnetic field is always perpendicular to the current direction, resonators can be classified as either "Axial" or "Transverse" according to the field orientation. Figure 30 shows a graphical representation of the different ways in which a current distribution can create a

homogenous field distribution with Figure 30-A1 and Figure 30-A2 representing the axial field orientations and Figure 30-B the transverse.

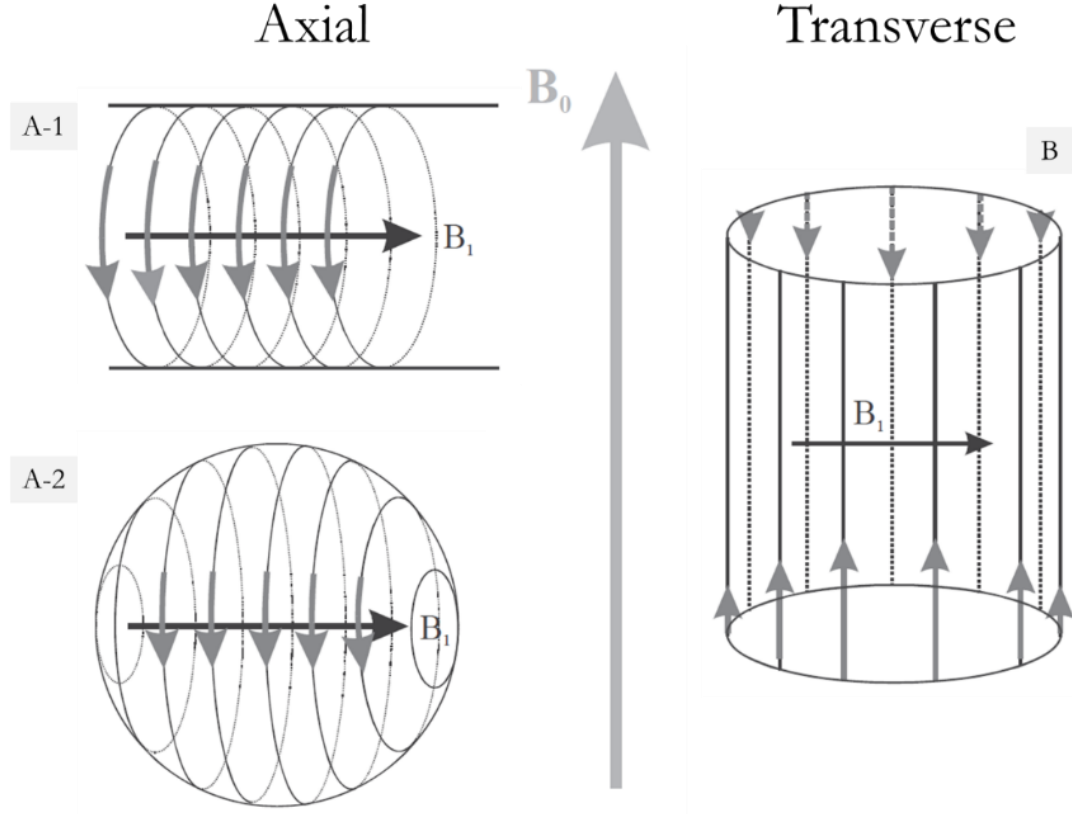


Figure 30: Representation of the different ways in which a current distribution can create a homogenous field distribution. (adapted from [104])

Both Axial and Transverse resonator topologies should enable highly homogenous $\mathbf{B}_1$ fields below the $\pm 5\%$ field variations threshold [104], [138]. However, sensitivity of transverse designs is generally worse than that obtained with axial designs [104]. For example, it has been demonstrated by Hoult and Richards [139] that the saddle coil, a transverse resonator, would boast only one third of the sensitivity of a classical solenoid axial resonator. Of course, this may vary slightly depending on the coil design and loading conditions, but transverse resonators are still considered less sensitive.

As the objective for benchtop scanners, where sensitivity is a primary concern due to their intrinsically low signal levels, axial resonators will be the topology of choice. Besides offering high sensitivity for the inherently weak signals at low field, their geometry can be mounted around the sample without obstructing access. This practicality is not available in clinical tunnel-type systems, where an axial coil would block patient loading. Therefore, only homogenous axial resonators will be investigated.

III.A. Cylindrical Distribution Based Designs (Solenoid Type Coils)

1. Ideal Distribution

The simplest model for a cylindrical resonator that produces a homogeneous axial field is the ideal solenoid. This theoretical construct consists of an infinitely long cylinder with a current

flowing purely in the tangential (azimuthal) direction. Figure 31 shows a representation of this ideal cylindrical current distribution.

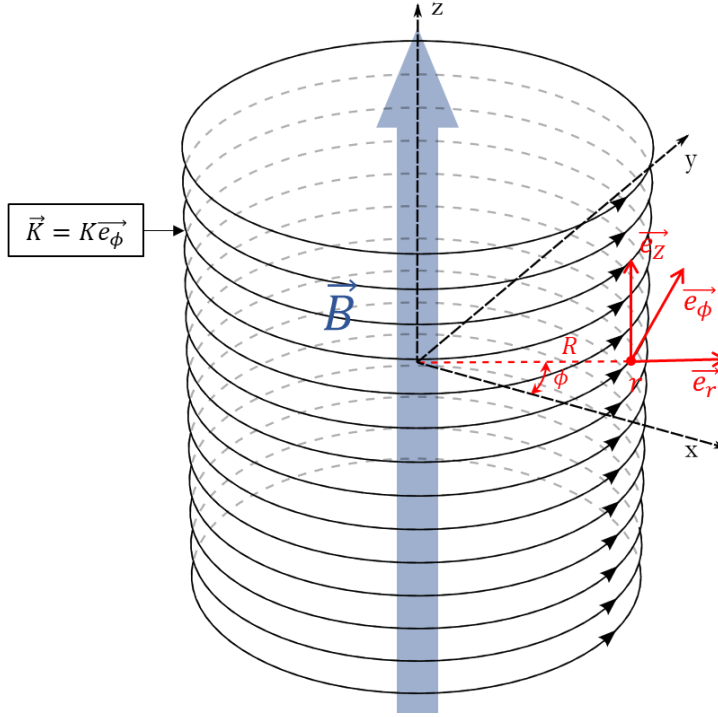


Figure 31: Representation of the ideal cylindrical surface-current distribution producing a homogeneous axial magnetic field.

In the ideal case, the cylinder is uniformly covered by a surface current density

In the ideal case, an infinitely long cylinder of radius R is uniformly covered by a surface current density

$$\vec{K} = K \vec{e}_\phi \quad (\text{III-1})$$

where K is the current per unit length (for a coil with n turns per unit length each carrying current I , $K = nI$). Taking into account the infinite symmetry, Ampere's Law in its integral form can be applied to find the magnetic field:

$$\oint_C \vec{B} \cdot d\vec{l} = \mu_0 I_{enc} \quad (\text{III-2})$$

where I_{enc} is the total current enclosed inside the cylindrical shape. By symmetry, the magnetic field outside the solenoid is zero ($\vec{B}_{out} = 0$) and inside it is non-zero, uniform, and directed purely along the axis of the cylinder ($\vec{B}_{in} = B_z \vec{e}_z$). To find B_z , an Amperian loop of length L , with one side inside the solenoid parallel to the axis, and the return path outside the solenoid.

The line integral of $\vec{B}$ around this loop is non-zero only for the segment of length L inside the solenoid, where it equals $B_z L$. The total current enclosed, I_{enc} , by this loop is the surface current density multiplied by the length L , so $I_{enc} = KL$. Applying Ampere's Law:

$$B_z L = \mu_0 (KL) \Rightarrow B_z = \mu_0 K = \mu_0 nI \quad (\text{III-3})$$

This demonstrates that an ideal, infinitely long solenoid type coil creates a perfectly homogeneous magnetic field within its volume.

2. Practical Approximation

Practically, however, any resonator must be of finite length. In this case, the magnetic field is no longer perfectly uniform, and its strength diminishes near the ends of the coil. To analyze this, the Biot–Savart law can be applied, which gives the infinitesimal field contribution from a current element $I d\vec{l}$:

$$d\vec{B} = \frac{\mu_0}{4\pi} \frac{I d\vec{l} \times \vec{e}_r}{r^2} \quad (\text{III-4})$$

where r is the distance from the element to the observation point and $\vec{e}_r$ is the corresponding unit vector.

The finite cylindrical resonator can be treated as a continuous series of circular current loops. Integrating the contribution from each loop along the solenoid's length (L) gives the on-axis magnetic field at a distance z from its center:

$$B_z(z) = \frac{\mu_0 n I}{2} \left[\frac{L/2 - z}{\sqrt{R^2 + (L/2 - z)^2}} + \frac{L/2 + z}{\sqrt{R^2 + (L/2 + z)^2}} \right] \quad (\text{III-5})$$

This equation reveals that the field is dependent on the axial position z . It is maximum at the center ($z = 0$) and falls off symmetrically towards the ends ($z = \pm L/2$) (“edge effects”).

Therefore, a simple, finite solenoid does not produce a truly homogeneous field. The degree of homogeneity is a function of the coil's aspect ratio ($L/2R$).

2.1. Traditional Solenoid Designs

The most straightforward practical implementation of a cylindrical current distribution is the simple or “traditional” solenoid, consisting of a continuous winding along a cylindrical shape [139]. Traditional solenoids are known to be approximately three times more sensitive than other classical geometries like the saddle and birdcage coils [139], which should make them highly efficient at lower fields where sensitivity is a limiting factor.

The trajectory of a solenoid can be defined parametrically by:

$$\begin{cases} x(t) = R \cos(t) \\ y(t) = R \sin(t) \\ z(t) = ct - (L/2) \end{cases} \quad (\text{III-6})$$

where R is the solenoid radius, t spans the angular domain over N loops ($t \in [0, 2\pi N]$), and c is the axial pitch per radian (the vertical rise of the helix per unit angular displacement) with $c = L/(2\pi N)$ ensuring that $z(t)$ spans a total height L .

The main difference between a practical solenoid and the ideal model is the current path. The ideal distribution assumes a perfect sheet of current flowing purely in the azimuthal ($\vec{e}_\phi$) direction. In contrast, a traditional solenoid is built from a single finite conductive trace wound in a helix. This helical path means the current has both an azimuthal component and a small but non-zero axial component ($\vec{e}_z$) which can be a source of deviation from the ideal magnetic field. However, for a tightly wound solenoid where the pitch is much smaller than the radius, the axial component

of the current can be expected to become negligible compared to the azimuthal component. In that case, the model of discrete current loops derived from Equation (III-5) becomes a very accurate approximation, and the solenoid's behavior closely approaches that of the ideal surface-current model. While this phenomenon can be mitigated with a tight winding, the primary cause of the large-scale field inhomogeneity are still "edge effects".

Furthermore, in literature, it has been found that solenoids do not only present significant field inhomogeneities at their edges, but also high-impedances, self-resonant behaviors, and higher electric and sample losses in the mid-sized coil range ($f \cdot d \approx 2\text{--}30$ MHz-m) range [110], [140]. They are also notably ill-suited when samples and environments are prone to significant changes or when proximity to conductive surfaces reduces their effective inductance [141]. Therefore, when imaging samples with varying properties and environments near a shielded enclosure, solenoids can experience significant drops in quality factor and large frequency shifts in the presence of conductive or dielectric samples, thus limiting their overall performance and versatility.

To remedy some of those limitations two variations to the classical solenoid topology stand out: The varying pitch solenoid and the single turn solenoid or loop gap.

2.2. Varying-Pitch Solenoid

A Varying-Pitch Winding Solenoid or Varying-Pitch Solenoid (VPS) is a solenoid design where the spacing between turns changes along its length, with a tighter at the ends and a loose winding at the center. The varying pitch winding approach aims to artificially increase the current density at the ends of the solenoid and extend the available homogenous field. Varying-pitch windings are used in low-field benchtop MRI primarily to improve B_1 -field homogeneity while keeping the solenoid short, which is critical in compact systems[142], [143]. Figure 32 shows examples of Varying-Pitch Solenoid used in benchtop systems.

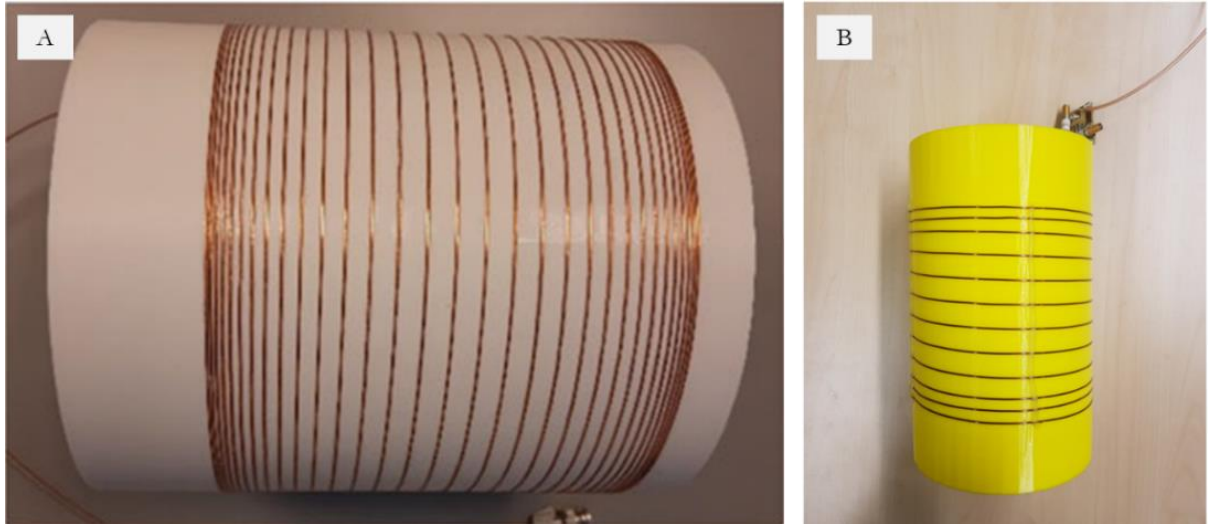


Figure 32: Example of varying pitch winding solenoids: (A) Variable-pitch transmit coil optimized for use in a Halbach-array-based portable MRI scanner (66 mT) [142].(B) Variable-pitch transmit coil for a low-field (50 mT) portable MRI system [143].

The proposed trajectory of a classical solenoid can be modified to parametrize such a winding configuration by maintaining the same trajectory in (x, y) while the axial coordinate is generated by a pitch-per-radian function:

$$\begin{cases} x(t) = R \cos(t) \\ y(t) = R \sin(t) \\ z(t) = Z(t) - \frac{L}{2}, \quad Z'(t) = c(s) \end{cases} \quad (\text{III-7})$$

with the normalized winding coordinate $s = \frac{t}{2\pi N} \in [0,1]$.

The taper function $c(s)$ must satisfy several design requirements:

- **Symmetry:** winding density should be identical near both ends ($z = \pm L/2$), hence $c(s)$ must be even about $s = 1/2$.
- **Smoothness:** pitch variation must be continuous to avoid local field ripples
- **Tunability:** a single parameter α should control the strength of the end winding.
- **Normalization:** the average pitch must remain $\langle c(s) \rangle = L/(2\pi N)$ so that the total length of the solenoid is preserved.

The cosine-squared modulation

$$c(s) = c(1 - \alpha \cos^2(\pi s)), \quad 0 \leq s \leq 1,$$

satisfies all these criteria:

- $\cos^2(\pi s)$ attains maxima at $s = 0,1$ and a minimum at $s = 0.5$, so the turn density $n(z) \propto 1/c(s)$ is largest near the coil ends and smallest in the middle, exactly what is needed to compensate for solenoid edge effects.
- $c(s)$ is continuous and differentiable, with $c'(0) = c'(1) = 0$, preventing abrupt pitch jumps.
- The average is easily evaluated as $\frac{1}{1} \int_0^1 c(s) ds = c(1 - \alpha/2)$, which yields the normalization $c = \frac{L}{2\pi N} \frac{1}{1 - \alpha/2}$.
- The end-to-center density ratio follows directly as $\frac{n_{\text{end}}}{n_{\text{center}}} = \frac{1}{1 - \alpha}$, providing a simple “knob” to tune homogeneity.

Alternative choices (linear tapers, higher-order polynomials, or $\cos(2\pi s)$ modulations) either lack smoothness, break symmetry, or complicate normalization. The cosine-squared taper is therefore the simplest smooth, symmetric, and analytically tractable profile for implementing end loading in a finite solenoid. Table 6 lists some useful quantities for this cosine-squared based taper winding.

Local pitch (per radian)	$c(t) = \frac{L}{2\pi N} \frac{1 - \alpha \cos^2(\pi s)}{1 - \alpha/2}, \quad s = \frac{t}{2\pi N}$
Local turns per unit length (approx.)	$n(z) \approx \frac{1}{2\pi c(t)}$
Minimum inter-turn spacing (near the ends)	$\Delta z_{\min} \approx 2\pi c_{\min} = \frac{L}{N} \frac{1 - \alpha}{1 - \alpha/2}$

Table 6: Useful quantities for cosine-squared based taper winding solenoid geometry

2.3. Loop Gap and Bridged Loop Gap Resonators

The loop gap resonator (or single-turn solenoid) can also be considered a variation on the traditional solenoid. Instead of a helical winding, the loop gap coil consists of a conductive loop interrupted by a capacitive gap. This gap breaks the continuity of the electric field, allowing the coil to resonate at a specific frequency.

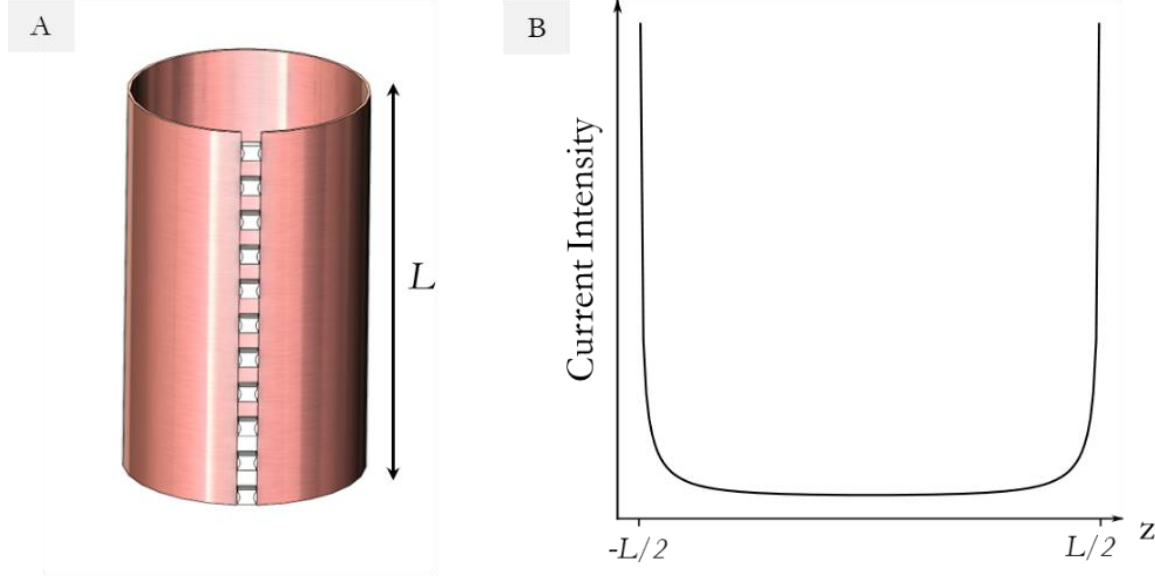


Figure 33: (A) Example of a Bridged Loop Gap Resonator BLGR. (B) Expected current intensity profile on the cylinder surface[140], [144].

Unlike solenoids, the current distribution over the surface of the finite hollow cylinder of the loop gap coil isn't homogenous throughout its entire length. Labiche et al [144] have shown it varies along the axis, concentrating near the edges, and must be evaluated numerically to predict the resulting RF field distribution which significantly extends the usable homogenous region. Similarly to a VPS configuration, this natural edge concentration of currents explains why the axial field inside a loop gap resonator is expected to be more homogeneous than in a simple solenoid. In a finite hollow cylinder, the RF current cannot distribute uniformly along its length because the conductor wall blocks any radial magnetic flux from penetrating. To satisfy this boundary condition, the current is forced to adjust itself so that the radial field at the wall vanishes. The only self-consistent solution is that more current flows near the ends of the cylinder, where it can bend the field lines back and preserve flux conservation. To this, a loop-gap adds a longitudinal capacitive interruption that is only expected to introduce azimuthal nonuniformity through charge accumulation at the gap edges, the axial edge enhancement remains the same as the underlying mechanism for the finite cylinder and can therefore be modeled as such for preliminary field behavior assessment.

Loop gap resonators are used primarily at ultra-high frequencies (UHF) due to the inherent physical limitations of the capacitance created by the gap dimensions. They can however be modified to be used at lower frequencies by bridging the gap with nonmagnetic high Q multilayer chip capacitors (MLCCs), expanding the range of frequencies available for any given geometry. These modified loop-gap resonators are commonly called Bridged Loop-Gap Resonators (BLGR) and offer better performances and stability for small samples in constrained spaces, with sufficient RF field (B_1) field homogeneity to perform experiments in transmit/receive (T/R) mode [140]. BLGRs maintain higher Q values and exhibit minimal frequency shifts compared to traditional solenoids [104] despite sample and environment variability. The BLGRs function as a so-called "high current probe" or "Low-E coil" [145] characterized by their low inductance design making

them less susceptible to inductive losses from local shielding and dielectric losses, primarily associated with the near-field component of the electric field, arising from capacitive coupling between the probe and the sample, and proportional the square of the electric field amplitude E . As these coils offer more favorable quality factors and less prominent frequency shifts from conductive and dielectric samples in a shielded enclosure, they are particularly adapted to applications with widely varying load conditions making them ideal for imaging heterogeneous, dynamically evolving samples.

It is important to emphasize, however, that a preliminary Biot-Savart based analysis is limited to the B-field distribution in vacuum. While this allows a clear comparison of intrinsic field homogeneity between different geometries, it does not capture the complete performance of an RF coil under realistic loading conditions. Therefore, any perceived gain from actual electrical performance would require a more complex analysis. Thus, although BLGRs is expected to demonstrate superior electrical properties [141], a true test of its performance requires accounting for additional mechanisms which will be assessed through full-wave electromagnetic simulations in a further section.

2.4. Preliminary Homogeneity Analytical Comparison Between the Different Cylindrical Coil Designs

2.4.1 Materials and Methods

As only cylindrical coils having a relatively simple geometry were considered ((free space, no shield, no sample loading, etc...), all three cylindrical coil designs (classical solenoid, varying-pitch solenoid, and loop-gap) were compared using Python-based (NumPy, SciPy, Matplotlib) analytical models. Each geometry was represented by its current path. For the classical solenoid and varying pitch solenoid, a helix trajectory following the proposed trajectories (Equation (III-6) and (III-7)) while the loop-gap coil was modeled as a series of discretized circular loops with optimized current distribution following the method described by Labiche et al. [144].

For each coil, the form factor FF of the cylindrical coils was defined as $FF = 2R/L$ and, when applicable, the default FF value was selected to be 0.5.

Magnetic fields were computed using the Biot–Savart law and normalized to the value of the field at the center of the coils $B_1(0,0,0)$.

For each design, the script generated:

- Families of axial $B_1(z)$ profiles along the symmetry axis for different FF values.
- 2-D $|B_1|$ maps on an (x,z) slice at $y = 0$ with a resolution of 241×241 pixels.
- Homogeneity metrics in an arbitrary cylindrical ROI spanning 80% of the coil radius and only 60% of the coil length to account for the inevitable edge effects (radius = $0.8R$, length = $0.6L$). The metrics used were the CoV and PFD histogram normalized to the B_1 mean value within the defined ROI volume ($\langle B_1 \rangle_{ROI}$).

This choice of ROI although arbitrary was made for two reasons. Firstly, to exclude fringe-field regions close to the coil ends where field distortions are unavoidable, and secondly to approximate the effective sample volume typically used in practical NMR/MRI setups, where the sensitive region is smaller than the full coil volume due to the presence of a former or substrate for the conductive trace or wire, especially in small volume coils.

For the solenoid designs $N=10$ and $N=20$ turns were investigated. The loop-gap resonator, by contrast, was discretized into $N=160$ slices. Here N does not correspond to a physical number of turns but is a numerical integration parameter ensuring accurate representation of the continuous current distribution around the cylinder. This distinction is critical to avoid confusion.

2.4.2 Results and Discussion

a. Classical Solenoid

The classical solenoid served as a baseline throughout the comparison process. As predicted by theory and shown in the B-field map in Figure 34, due to its finite nature, the classical solenoid design exhibits significant "edge effects," where the magnetic field strength drops off near the ends of the coil. The Percentage Field Deviation (PFD) histogram in Figure 34-C quantifies this phenomenon, as a relatively broad distribution of field values within the region of interest (ROI) can be observed.

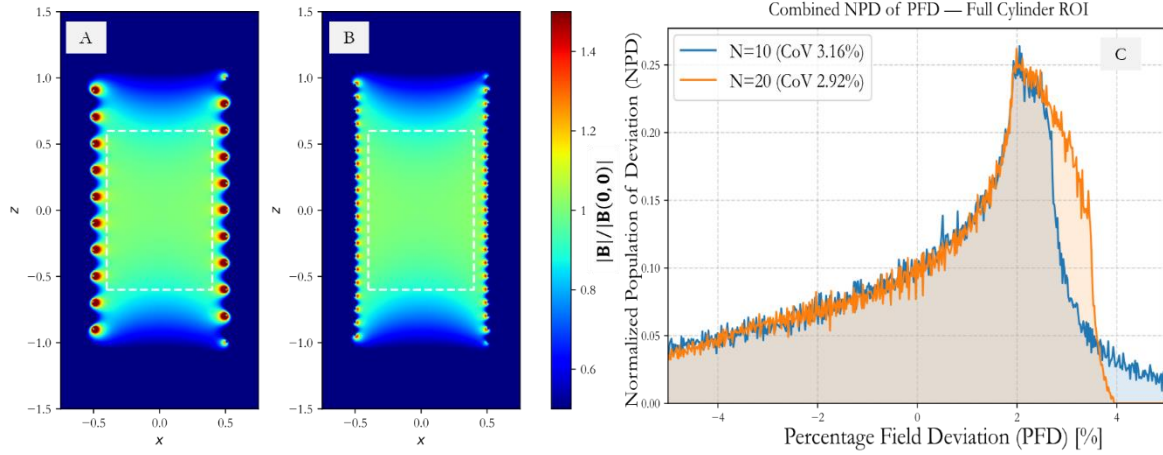


Figure 34: Magnetic field magnitude map for classical solenoids with a form factor (FF) of 0.5. The maps are shown for (A) $N=10$ turns and (B) $N=20$ turns. The dashed white line indicates the cylindrical Region of Interest (ROI) used for homogeneity analysis. (C) Normalized Population of Deviation (NPD) versus Percentage Field Deviation (PFD) for classical solenoids within the ROI. The plot compares a 10-turn coil (CoV 3.16%) and a 20-turn coil (CoV 2.92%), showing only a slight improvement in homogeneity with more turns.

Furthermore, the calculated coefficient of variation was 3.16% for a 10-turn ($N=10$) coil and 2.92% for a 20-turn ($N=20$) coil. This slight improvement confirms as expected, that a denser winding more closely approximates the ideal, continuous current sheet, by reducing localized field ripples and the small axial current components. However, this is a marginal difference (7.6%), which means that, for a tightly wound solenoid, the homogeneity is still dominated by the global edge effects, not the discrete nature of the windings.

Although the CoV values fall well within the 5% margin that was defined previously as an acceptable homogenous range, the field histogram highlights clearly that the overall homogeneity remains limited by the inherent field drop-off at the ends. This not only showcases the necessity of using multiple tools to assess coil field homogeneity but also justifies the need for better optimized coil geometries to remedy "edge effects".

b. Varying Pitch Solenoid

The varying-pitch solenoid was designed specifically to counteract the edge effects of the classical design. By adjusting the taper strength parameter (α), the current density is increased at the ends of the coil. Figure 35 shows the B-field maps for different α values.

The results clearly demonstrate the effectiveness of using a varying-pitch approach. As shown in Figure 36-A, increasing α creates "humps" in the axial field profile that compensate for the natural drop-off. The PFD analysis in Figure 36-B reveals an optimal taper of $\alpha = 0.25$, which achieved a minimum CoV of 0.99% (for $N=20$) and the tallest and narrowest histogram peak

indicating the strongest uniformity. This means that by using a varying pitch approach the homogeneity within the region of interest increased by more than three times compared to the classical solenoid. The B-field maps (Figure 35) also visually confirm that an optimized VPS produces a more uniform field within the ROI.

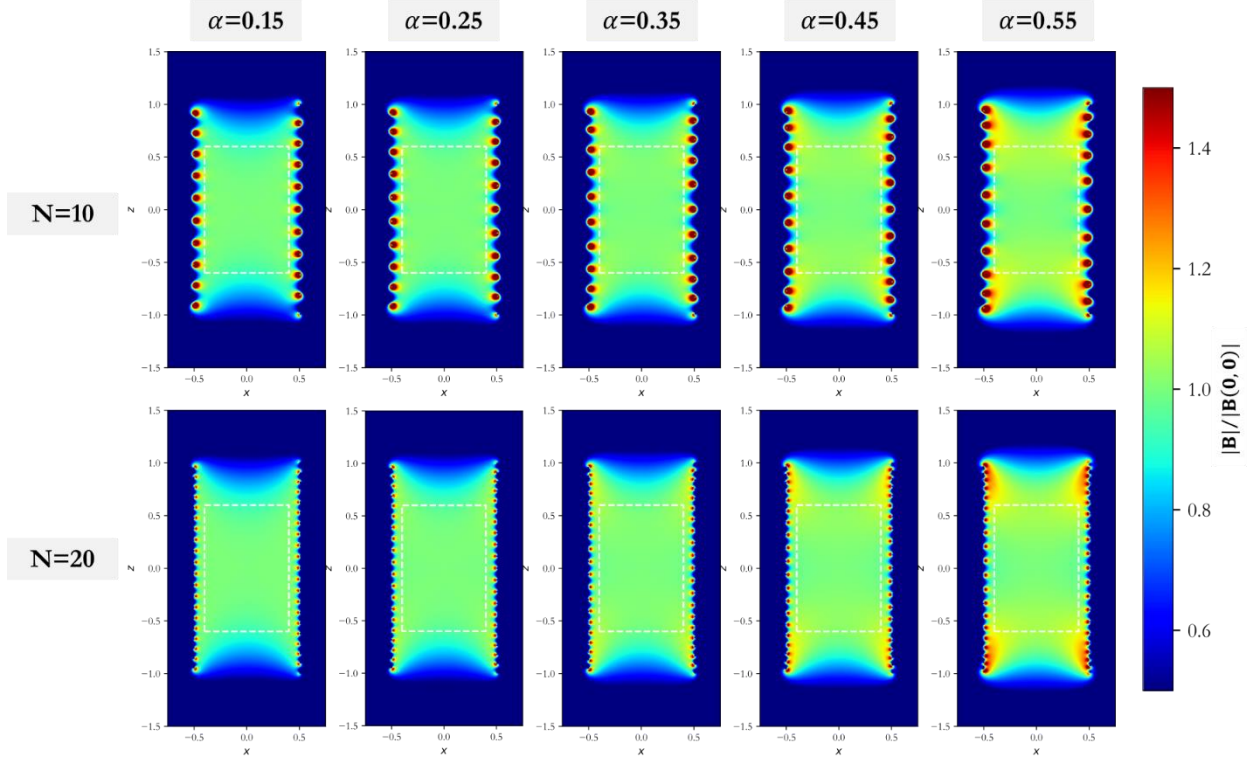


Figure 35: Magnetic field magnitude maps for varying-pitch solenoids (FF=0.5) across a range of taper strengths (α) for $N=10$ (top row) and $N=20$ (bottom row) turns. The maps illustrate how increasing (α) extends the field distribution to improve homogeneity within the ROI (dashed white line).

However, both the B-field maps and the homogeneity analysis show that tapering beyond an optimal value (here $\alpha = 0.25$) overcompensates for the edge effects by introducing localized overshoots near the ends, leading to a quick decline in homogeneity (e.g., CoV increases to 3.40% at $\alpha = 0.55$).

Furthermore, the proposed implementation of the varying-pitch solenoid is designed to maintain a fixed coil length while tightening edge windings and loosening the central windings which means that the pitch angle in the center must increase. As the central pitch widens, the axial component of the current at the center of the coil can no longer be expected to become negligible compared to the azimuthal component, creating a drop in homogeneity near the central windings and reducing the width of the available homogenous region.

This degradation is evident when comparing the field maps in Figure 35 for taper strengths $\alpha=0.15$ and $\alpha=0.55$. The strong taper at the ends (for $\alpha=0.55$) seems to be “stretching out” the magnetic field and creating a drop in field intensity on each side. While the VPS concept effectively addresses axial field drop-off, this simple implementation also compromises radial homogeneity.

Optimal taper α and Form Factor FF sets for maximum homogeneity were calculated for FF=0.2 to FF=0.6 and the values for $N=20$ are available in Table 7. The corresponding optimal axial profiles were plotted in Figure 39-A.

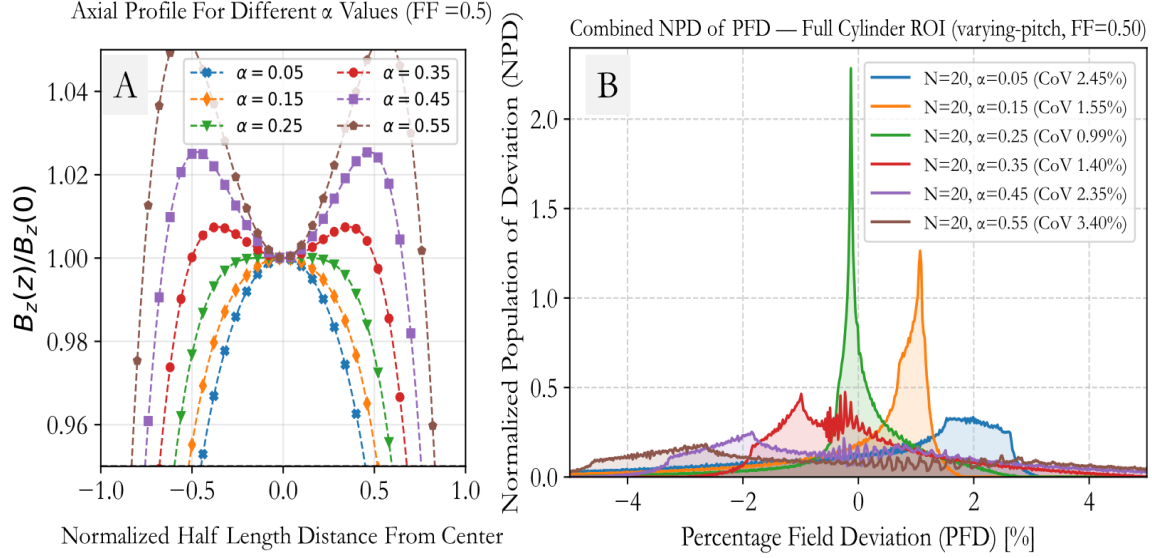


Figure 36: Homogeneity analysis of a 20-turn varying-pitch solenoid (FF=0.5). (A) Normalized axial magnetic field profiles for different taper strengths (α), showing how the field is flattened and even raised at the ends. (B) Corresponding PFD histograms, demonstrating an optimal homogeneity at $\alpha=0.25$ (CoV 0.99%).

Coil Form Factor (FF)	0.2	0.3	0.4	0.5	0.6
Optimal Taper Strength (α)	0.04	0.10	0.17	0.26	0.35
Coefficient of Variation (CoV)	2.36%	0.77%	0.71%	0.99%	1.32%

Table 7: Optimal taper α and Form Factor FF sets for maximum homogeneity

c. Loop Gap

The loop-gap resonator, with its fundamentally different current distribution that concentrates near the cylinder's edges (Figure 33-B), demonstrated the highest field homogeneity among the three designs. The B-field map presented in Figure 37 visually shows a highly uniform field throughout the ROI.

This visual observation is confirmed by the PFD histogram, which displays an exceptionally sharp and narrow peak around 0% deviation with a resulting CoV of 0.45%, approximately seven times more homogeneous than the classical solenoid and more than twice as homogeneous as the optimized VPS.

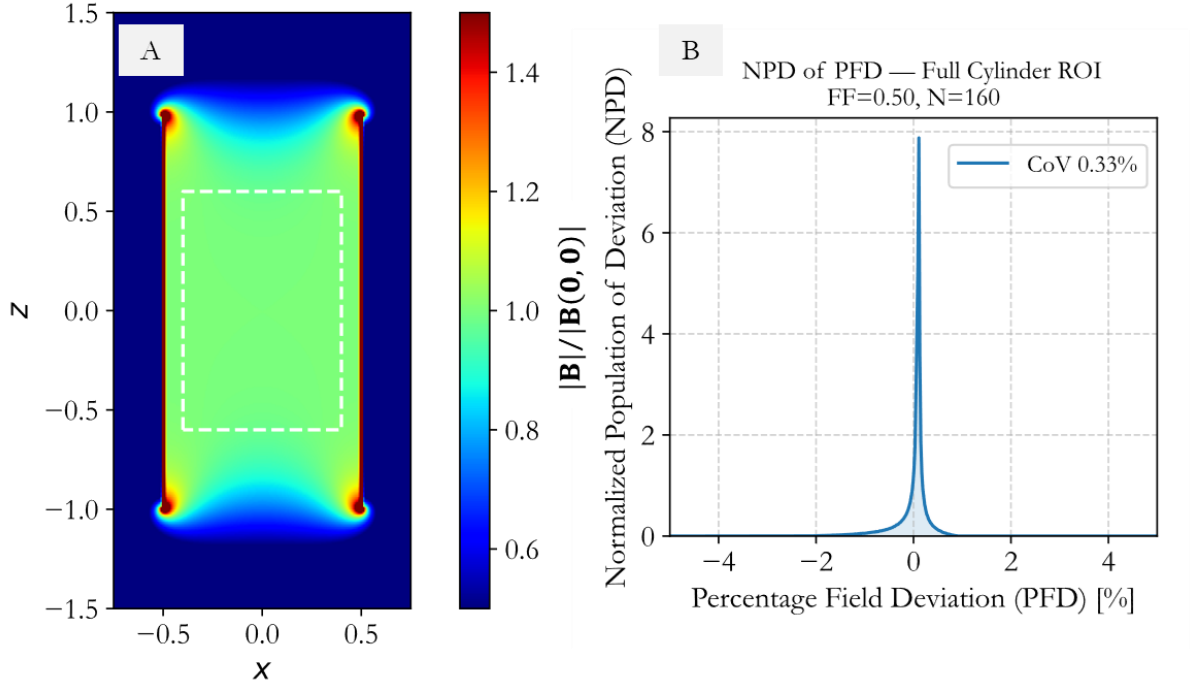


Figure 37: Magnetic field homogeneity of a loop-gap resonator (FF=0.5). (A) The magnetic field magnitude map, showing high uniformity within the ROI (dashed white line). (B) The corresponding NPD vs. PFD histogram, which has a sharp peak indicating excellent homogeneity (CoV 0.45%).

d. Comparative Analysis and Influence of Form Factor

When comparing the axial field profiles across different form factors ($FF = 2R/L$), as seen in Figure 38 and Figure 39, the loop-gap resonator consistently provided the flattest and most extended uniform field region. While the homogeneity of all the designs degrades as they become shorter and wider (i.e., as FF increases), the loop-gap design is the most resilient to these geometric changes.

The form factor is a particularly important parameter because it directly dictates the trade-off between magnetic field homogeneity and sample coverage. For a given coil topology, a low FF (long and narrow coil) generally increases the extent of the uniform field region but reduces the filling factor, lowering sensitivity for a given sample size. On the other hand, a high FF (short and wide coil) improves filling and can therefore boost the intrinsic signal-to-noise ratio, but this comes at the expense of strong edge effects and poor homogeneity across the region of interest. As such, FF serves as the critical geometric control variable when evaluating coil performance: it simultaneously governs the spatial coverage of the homogeneous region, the achievable filling factor, and therefore the effective sensitivity of the resonator for a given sample geometry. Moreover, a design that tolerates a greater range of FF values offers better adaptability to different sample sizes and shapes making them practical for MRI setups where the coil must accommodate multiple experimental configurations.

Classical solenoids seem to be the least suited to achieving the homogeneity levels required for compact MRI systems, where edge effects dominate the effective imaging region. The Axial profiles in Figure 38 confirm this behavior as even very low FF values barely provide the coverage necessary to cover 60% of the coil length. To increase the size of the homogenous region, a very long solenoid compared to the ROI would provide sufficient coverage but would in turn drastically reduce the filling and at the same time the SNR. For example, according to Figure 38, for a 40

mm long by 40 mm diameter sample would require an FF of 0.4 or below to maintain the sample within the 95% homogenous region. As a high form factor correlates directly to a higher filling factor, SNR optimization through filling factor optimization will be quite limited when using this topology.

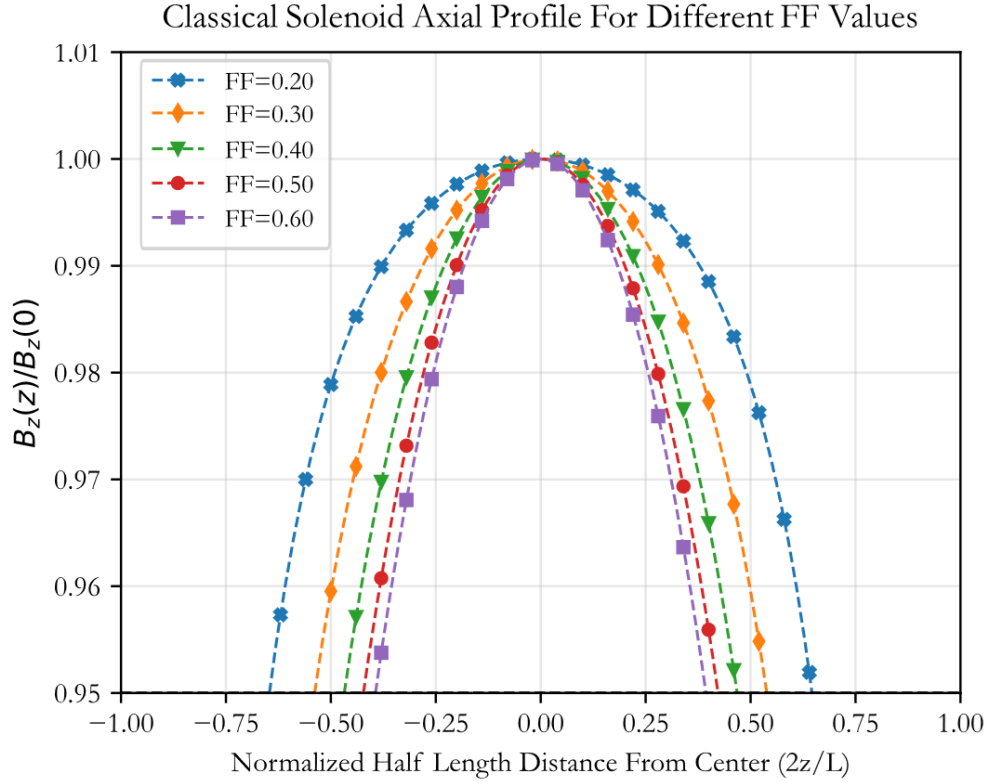


Figure 38: Normalized axial magnetic field profiles for a classical solenoid ($N=20$) at different form factor (FF) values. The plot shows that field homogeneity decreases significantly as FF increases leading to the coil becoming shorter and wider.

On the other hand, the VPS offers a significant and customizable improvement over the classical solenoid for FF values above 0.2. Taking the same 40×40 mm cylindrical ROI example, and according to Figure 39-A, with an optimized $\alpha=0.25$ an FF=0.5 or below would cover the desired ROI with a more than 95% homogenous field.

However, the analysis confirms that for applications demanding the highest possible B1 homogeneity within a compact cylindrical volume, the loop-gap resonator's inherent edge-current distribution provides a clear advantage. Taking the same 40×40 mm cylindrical ROI example, and according to Figure 39-B, an FF=0.6 or below would generate a 95% homogenous region that would largely cover the full sample.

While it has been demonstrated here that the loop gap resonator is by far the best suited to maximizing filling factor, by enabling a greater range of form factors, while maintaining high homogeneity, this remains a preliminary analysis. The choice between these designs in a practical application would therefore depend on the specific balance required between field homogeneity, coil size, and other operational factors such as sample loading, quality factors and resonant frequency tuning and stability.

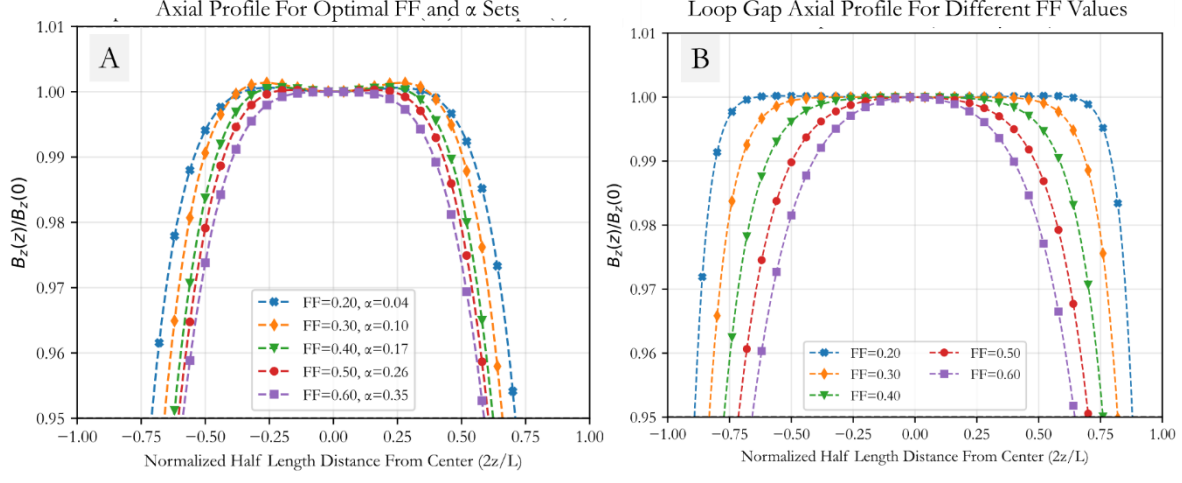


Figure 39: Comparison of normalized axial magnetic field profiles for different form factors (FF). (A) A varying-pitch solenoid with an optimized taper α and Form Factor FF sets. (B) A loop-gap resonator. The loop-gap design maintains a flatter, more homogeneous profile across a wider range of geometries compared to the VPS. This might be mitigated by systematically reoptimizing α for a given FF.

III.B. Spherical Uniform Current Density Designs

1. Ideal Distribution

The theoretical ideal for generating a perfectly homogeneous magnetic field within a spherical volume is based on a specific distribution of surface current. Starting from fundamental electromagnetic principles, it is possible to demonstrate that this ideal field is produced by a current density that flows azimuthally on the surface of a sphere, with a magnitude that varies with the sine of the polar angle [104], [138], [146]. Figure 40 shows a representation of this ideal spherical current distribution.

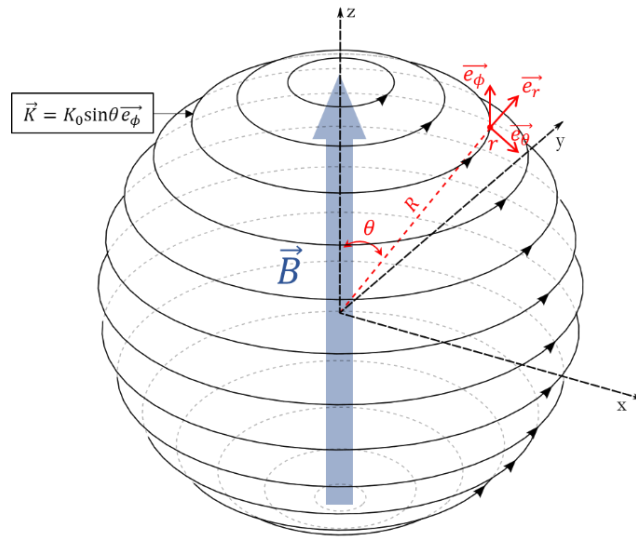


Figure 40: Representation of the ideal spherical surface-current distribution producing a homogeneous axial magnetic field.

Let's consider a sphere of radius R with a surface current density $\vec{K}$ given in spherical coordinates by:

$$\vec{K} = K \sin\theta \vec{e}_\phi \quad (\text{III-8})$$

Here, K is a constant representing the maximum current density (at the equator, where $\theta = \pi/2$), θ is the polar angle, and $\vec{e}_\phi$ is the azimuthal unit vector. This describes current flowing in loops parallel to the xy-plane.

To determine the magnetic field $\vec{B}$ generated by this current distribution, it is most convenient to first calculate the magnetic vector potential, $\vec{A}$, from which the magnetic field can be found using the relation $\vec{B} = \nabla \times \vec{A}$ [40]. The vector potential due to a surface current is given by the integral:

$$\vec{A}(\vec{r}) = \frac{\mu_0}{4\pi} \int_S \frac{\vec{K}(\vec{r}')}{|\vec{r} - \vec{r}'|} da' \quad (\text{III-9})$$

where μ_0 is the permeability of free space, $\vec{r}'$ is the position vector to a point on the surface S where the current element is, $\vec{r}$ is the position vector to the point where the potential is calculated and da' is the differential surface area element on the sphere.

While the direct integration is complex, this is a canonical problem in electromagnetism, often solved using expansions in Legendre polynomials [40]. The established solution for the vector potential inside the sphere ($r \leq R$) is remarkably simple (Annex I):

$$\vec{A}_{in}(r, \theta, \phi) = \frac{\mu_0 K_0}{3} r \sin\theta \vec{e}_\phi \quad (\text{III-10})$$

Now, the magnetic field can be given by taking the curl of $\vec{A}_{in}$ in spherical coordinates:

$$\vec{B} = \nabla \times \vec{A} = \frac{1}{r \sin\theta} \frac{\partial}{\partial \theta} (\sin\theta A_\phi) \vec{e}_r - \frac{1}{r} \frac{\partial}{\partial r} (r A_\phi) \vec{e}_\theta \quad (\text{III-11})$$

Substituting the expression for $A_\phi = \frac{\mu_0 K}{3} r \sin\theta$, the radial and polar components, respectively B_r and B_θ , become:

$$B_r = \frac{2}{3} \mu_0 K \cos\theta, \quad B_\theta = -\frac{2}{3} \mu_0 K \sin\theta \quad (\text{III-12})$$

The azimuthal component, B_ϕ , is zero as A does not depend on ϕ .

Combining the components gives:

$$\vec{B} = B_r \vec{e}_r + B_\theta \vec{e}_\theta = \frac{2}{3} \mu_0 K (\cos\theta \vec{e}_r - \sin\theta \vec{e}_\theta) \quad (\text{III-13})$$

Recognizing that the unit vector in the Cartesian z-direction can be expressed in spherical coordinates as $\vec{e}_z = \cos\theta \vec{e}_r - \sin\theta \vec{e}_\theta$, the result becomes:

$$\vec{B}_{in} = \frac{2}{3}\mu_0 K \vec{e}_z \quad (\text{III-14})$$

This result is profound: the magnetic field inside the sphere is perfectly uniform in magnitude and direction, pointing along the z-axis. It does not depend on any spatial coordinates (r, θ, ϕ) .

2. Practical Approximation

Practically, it is very difficult to achieve the ideal spherical current distribution. This theoretical current distribution is however the basis for advanced resonator designs like the spherical birdcage coil. The foundational concept was first illustrated by Clark in 1938 [147], who described an approximation of an ideal, continuous current sheet on a sphere as a new way to generate a homogenous field. The first practical approximations were developed by Everett in 1966 [148] and later improved and adapted to MRI by Bolinger et al. 1988 [149]. Their designs used a finite number of discrete, equidistant conductor loops arranged over a spherical surface to mimic the ideal current sheet. The primary drawback of this approach was that the limited number of conductors flown by the same current I and their suboptimal equidistant placement resulting suboptimal magnetic field homogeneities compared to other topologies[112].

2.1. Harpen Approximation of The Ideal Spherical Distribution

Subsequent work by Harpen in 1991 [146] introduced a more elegant solution to the spherical birdcage resonator. Instead of focusing solely on conductor placement, Harpen's design used equally spaced conductive loops and controlled the current flow by precisely varying the capacitance between them. To generate a perfectly uniform field, the design required a specific capacitance distribution given by the equation:

$$C(\theta) = C_0/\sin\theta \quad (\text{III-15})$$

where $C(\theta)$ is the capacitance per unit angle at a given polar angle θ . This electrical configuration forces the current into the desired pure sinusoidal distribution (an $l = 1$ Legendre mode), significantly improving field homogeneity. While innovative, the design remained an approximation sensitive to the precision of both the physical construction and the discrete capacitor values.

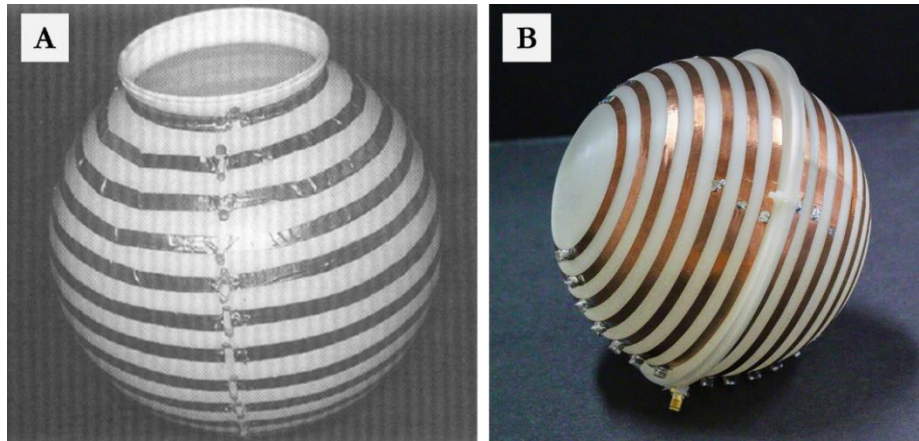


Figure 41: Practical implementations of a spherical birdcage resonators: (A) Fabricated by Harpen in 1991 [146]. (B) Fabricated at the AMPERE laboratory in 2021 using a 3D Plastronics approach as an initial practical implementation to verify fabrication feasibility and to serve as a reference for the analytical and simulation-based study presented here.

Nowadays thanks to modern fabrication methods such as 3D Plastronics capable of producing spherical objects with precisely placed imbedded conductors coupled with the ability to cleanly stack very small non-magnetic capacitors to achieve any desired value such a design becomes more achievable than before. Figure 41 shows pictures comparing the spherical birdcage resonator fabricated by Harpen in 1991 and a similar coil fabricated at the AMPERE laboratory in 2021 using a 3D Plastronics approach.

Moreover, it can be shown that this Harpen approximation of a spherical distribution of current is limited by the number of rings and their placements. This design becomes relevant only above a certain number of rings as the equal spacing between each ring doesn't allow for the cancelation of the specific magnetic field distortions (harmonics) created by using discrete current distributions (Figure 42).

2.2. Hoult and Deslauriers Four Ring Configuration

A solution to the suboptimal ring distribution of the Harpen approach was hinted at by Hoult and Deslauriers [101] a year prior in their 1990 paper. In this paper, they propose an optimized four ring coil to approximate the ideal spherical current distribution by tailoring the ring positions and the current ratio between rings to precisely cancel out the low-order even spherical-harmonic terms of the axial field.

For a circular ring lying on a sphere of radius R , at semiangle θ (measured from the $+Z$ symmetry axis), the axial magnetic field can be expanded in spherical harmonics. Along the symmetry axis, this reduces to an even-power series:

$$B_{1z}(z) = \frac{\mu_0 i}{2R} \sin\theta \sum_{n=0}^{\infty} P_{n+1}^1(\cos\theta) \left(\frac{z}{R}\right)^n, \quad (\text{III-16})$$

where P_ℓ^1 are associated Legendre functions. The coefficient of the n -th term is therefore

$$a_n(\theta) = \sin\theta P_{n+1}^1(\cos\theta). \quad (\text{III-17})$$

For a four-ring configuration, two identical pairs of rings at θ_1 and θ_2 above the equator, mirrored by identical rings at $\pi - \theta_1$ and $\pi - \theta_2$ below the equator, with equal currents within each pair are considered. This symmetric construction eliminates all odd powers automatically, so only even orders $n = 0, 2, 4, 6, \dots$ remain. The total even-order coefficients are

$$a_n^{(\text{total})} = i_{14} \sin\theta_1 P_{n+1}^1(\cos\theta_1) + i_{23} \sin\theta_2 P_{n+1}^1(\cos\theta_2), \quad (\text{III-18})$$

where i_{14} and i_{23} are the currents in the outer and inner pairs, respectively.

Hoult and Deslauriers enforced $a_4 = 0$ for each pair individually by choosing θ at the roots of $P_5^1(\cos\theta)$:

$$P_5^1(\cos\theta) = \frac{15}{8} \sin\theta (21\cos^4\theta - 14\cos^2\theta + 1) = 0. \quad (\text{III-19})$$

Excluding the trivial $\sin\theta = 0$ solutions, this condition yields

$$\theta_1 \approx 0.699 \text{ rad } (40.05^\circ), \quad \theta_2 \approx 1.282 \text{ rad } (73.45^\circ), \quad (\text{III-20})$$

so that each pair contributes no a_4 term regardless of current. So far, only the 4th order components are canceled. With θ_1 and θ_2 fixed, the quadratic term a_2 can also be canceled by enforcing

$$i_{14} \sin\theta_1 P_3^1(\cos\theta_1) + i_{23} \sin\theta_2 P_3^1(\cos\theta_2) = 0, \quad (\text{III-21})$$

which gives the current ratio

$$\frac{i_{14}}{i_{23}} = -\frac{\sin\theta_2 P_3^1(\cos\theta_2)}{\sin\theta_1 P_3^1(\cos\theta_1)} \approx +0.682. \quad (\text{III-22})$$

Similarly to the Harpen method, these current ratios can be enforced through well selected capacitances on each ring. With these angles and this current ratio, Hoult and Deslauriers further observed that the 6th order coefficient generating an even more homogenous field.

Although tailored to four ring coils, this strategy can be expanded to topologies with a greater number of rings.

2.3. Generalization to N-Ring Coils

Let us consider a set of N circular rings lying on a spherical surface of radius R , centered at the origin, with the symmetry axis along $+Z$. The rings are arranged as $M = N/2$ pairs placed symmetrically about the equator, so odd orders vanish by construction. Let

$$\theta_{Z,j} \in (0, \pi/2), \quad j = 1, \dots, M, \quad (\text{III-23})$$

denote the polar angle measured from $+Z$ of the j th pair. Each pair carries current I_j ; the two rings within a pair carry the same current and share the same $\theta_{Z,j}$.

Let a_n denote the on-axis coefficient multiplying z^n in the interior series expansion of the RF field. For a ring pair at θ_Z with current I , the even-order coefficients contribute linearly as

$$a_n(\theta_Z; I) \propto I \sin\theta_Z P_{n+1}^1(\cos\theta_Z), \quad n = 0, 2, 4, \dots, \quad (\text{III-24})$$

where P_ℓ^1 is the associated Legendre function. Summing over the M pairs gives

$$a_n = \sum_{j=1}^M I_j \sin\theta_{Z,j} P_{n+1}^1(\cos\theta_{Z,j}), \quad n = 0, 2, 4, \dots \quad (\text{III-25})$$

The design objective is to maintain $a_0 > 0$ while cancelling

$$a_2 = a_4 = \dots = a_{2K} = 0, \quad (\text{III-26})$$

for as large a K as possible given M degrees of freedom.

Using the standard definition of the associated first order Legendre function [150]:

$$P_\ell^1(x) = -\sqrt{1-x^2} P'_\ell(x), \quad (\text{III-27})$$

from which it follows that

$$P'_\ell(x_0) = 0 \quad \Rightarrow \quad P_\ell^1(x_0) = 0. \quad (\text{III-28})$$

Setting $x = \cos\theta$ and noting $\sin\theta = \sqrt{1 - \cos^2\theta}$, the order- n contribution of an individual pair in Equation (III-25) vanishes whenever

$$P_{n+1}^1(\cos\theta_z) = 0 \quad \Leftrightarrow \quad P'_{n+1}(\cos\theta_z) = 0. \quad (\text{III-29})$$

Similarly to Hoult and Deslauriers [101] the first step of the generalized strategy will rely on a geometric pre-cancellation of some higher orders.

For M ring pairs, choose the M semiangles $\{\theta_{z,j}\}$ as the angles corresponding to the M largest positive roots of

$$P'_{2M+1}(x) = 0, \quad x \in (0,1), \quad \theta_{z,j} = \arccos x_j. \quad (\text{III-30})$$

This ensures each pair individually contributes zero to the $n = 2M$ coefficient:

$$\sin\theta_{z,j} P_{2M+1}^1(\cos\theta_{z,j}) = 0 \quad \Rightarrow \quad a_{2M} \equiv 0 \quad \text{for any currents } \{I_j\}. \quad (\text{III-31})$$

This is advantageous since a_{2M} is the highest order accessible with M degrees of freedom, once this order is guaranteed zero geometrically, the remaining $M - 1$ degrees of freedom can be dedicated to cancelling all lower orders by linear combination. Eliminating a_{2M} first is not only mathematically convenient but also optimal, since it maximizes the number of even terms cancelled for a given number of pairs.

With the geometry fixed by equation (III-30), the second step aims to cancel the remaining lower even orders by solving the homogeneous linear system

$$\sum_{j=1}^M I_j \sin\theta_{z,j} P_{n+1}^1(\cos\theta_{z,j}) = 0, \quad n = 2, 4, \dots, 2M - 2. \quad (\text{III-32})$$

This is an $(M - 1) \times M$ system for the current ratios $\{I_j\}$, with a one-dimensional null space. All that is left is to pick the solution in the null space, fix the global scale by setting the innermost pair current to $+1$, and choose the overall sign so that

$$a_0 = \sum_{j=1}^M I_j \sin\theta_{z,j} P_1^1(\cos\theta_{z,j}) > 0. \quad (\text{III-33})$$

After the geometric pre-cancellation and the current cancellation of lower orders, at least the orders $n = 2, 4, \dots, 2M$ are cancelled. The first non-zero even order is generically $n \geq 2M + 2$ (additional “bonus” cancellations may occur for special geometries, just like the $M = 2$ case where the $n = 6$ term is observed to be small or vanish numerically).

Algorithmically, the strategy can be summarized by:

4. Given $N = 2M$ rings, compute the M largest positive roots $\{x_j\}_{j=1}^M$ of $P'_{2M+1}(x)$ in $(0,1)$.
5. Set $\theta_{z,j} = \arccos x_j$ and place one pair at each $\theta_{z,j}$, mirrored about the equator.
6. Assemble the $(M - 1) \times M$ matrix

$$A_{kj} = \sin\theta_{z,j} P_{2k+1}^1(\cos\theta_{z,j}), \quad k = 1, \dots, M-1, \quad j = 1, \dots, M. \quad (\text{III-34})$$

7. Compute a basis vector of $\text{null}(A)$ (the smallest right singular vector).
8. Set the innermost pair current to $+1$ and flip the overall sign if needed so that $a_0 > 0$ via (Equation (III-33)).
9. Verify cancellations by evaluating $a_2, a_4, \dots, a_{2M}$ from (Equation (III-25)); the first residual should be at $n \geq 2M + 2$.

2.4. Calculating the Capacitor Values for the H&D Extended Method

As stated previously, similarly to the Harpen method, the current ratios necessary to cancel the lower order even spherical harmonics can be enforced through well selected capacitances on each ring.

Considering N conductive loops (“rings”), each broken by a series tuning capacitor C_k (and possibly a small series resistance R_k) with the loops purely magnetic magnetically coupled, the complex current phasors can be written as:

$$\mathbf{I} = [I_1 \quad I_2 \quad \dots \quad I_N]^T, \quad (\text{III-35})$$

and the self and mutual inductances (L_k, M_{kj}) can be defined by the symmetric matrix

$$\mathcal{L} \in \mathbb{R}^{N \times N}, \quad \mathcal{L}_{kk} = L_k, \quad \mathcal{L}_{kj} = M_{kj} = \mathcal{L}_{jk}. \quad (\text{III-36})$$

Applying Kirchhoff's voltage law in the phasor domain, gives for loop k :

$$\underbrace{\sum_{j=1}^N j\omega \mathcal{L}_{kj} I_j}_{\text{Induced Voltage}} + \underbrace{\left(R_k - \frac{j}{\omega C_k}\right) I_k}_{\text{Local Voltage Drop}} = 0. \quad (\text{III-37})$$

Assuming that the small resistive terms R_k is negligible at the tuning stage, the condition for a specific operating frequency ω_0 and a desired current pattern $\mathbf{I}$ is

$$(\omega_0 \mathcal{L} - 1/\omega_0 \mathbf{C}^{-1})\mathbf{I} = \mathbf{0}, \quad (\text{III-38})$$

where $\mathbf{C} = \text{diag}(C_1, \dots, C_N)$.

Therefore, the k -th loop condition reads

$$(\mathcal{L}_{kk} - 1/\omega_0^2 C_k) I_k + \sum_{j \neq k} \mathcal{L}_{kj} I_j = 0. \quad (\text{III-39})$$

$$\frac{1}{\omega_0^2 C_k} I_k = \sum_{j=1}^N \mathcal{L}_{kj} I_j = [\mathcal{L} \mathbf{I}]_k. \quad (\text{III-40})$$

Thus, the required capacitance for ring k is

$$C_k = \frac{I_k}{\omega_0^2 [\mathcal{L} \mathbf{I}]_k}, \quad k = 1, \dots, N. \quad (\text{III-41})$$

This equation states that each capacitor enforces the cancellation of the self and mutual inductive reactance seen by its ring, such that the current pattern following vector $\mathbf{I}$ distribution becomes a natural mode of oscillation at ω_0 .

By, defining the effective inductance seen by ring k in the desired current distribution as

$$L_k^{(\text{eff})} = \frac{[\mathcal{L} \mathbf{I}]_k}{I_k}, \quad (\text{III-42})$$

it is possible to write simply $C_k = 1/(\omega_0^2 L_k^{(\text{eff})})$. This highlights that each loop must resonate with its own mode-dependent effective inductance.

Importantly, this behavior scales with frequency. For fixed geometry, loading, and current vector $\mathbf{I}$, the capacitances scale as

$$C_k(f_2) = C_k(f_1) \left(\frac{f_1}{f_2} \right)^2, \quad (\text{III-43})$$

so that capacitance ratios C_i/C_j remain invariant when returning to a new frequency.

In practice, the inductance matrix $\mathcal{L}$ can be obtained from a well-defined coil model and EM solver by exporting the multiport impedance matrix $\mathbf{Z}$ (extractable for S-parameters) at a sufficiently low “quasi-static” frequency f_s , and computing

$$\mathcal{L} \approx \frac{\text{Im}\{\mathbf{Z}(f_s)\}}{\omega_s}, \quad \omega_s = 2\pi f_s. \quad (\text{III-44})$$

2.5. Preliminary Homogeneity Analytical Comparison Between the Proposed Generalized N-Ring Coil Strategy and the Harpen Approximation

2.5.1 Materials and Methods

To validate the proposed generalized method, the two representative strategies were compared: the Harpen method, where N conductors are placed at equally spaced polar angles ($\theta = \frac{\pi/2}{N}, \dots, \pi - \frac{\pi/2}{N}$) with current weighted by $\sin(\theta)$, and the extended Hoult and Deslauriers method (H&D Extended), in which conductor positions are set by the positive roots of P'_{2M+1} ($M = N/2$) and the lowest even spherical-harmonic terms are suppressed through optimized current weighting. The corresponding current weights were optimized by solving a small linear system via singular value decomposition (SVD).

As only the spherical coils having a relatively simple geometry were considered (no shield, no sample loading, etc...) a python script with simple mathematical models for generating magnetic field maps was used for the comparison.

For each configuration, the axial field (B_z) was calculated using the analytical Biot–Savart expression with elliptic integrals, and superposition was applied to obtain the total field. Homogeneity was assessed within a spherical Region of Interest (ROI, 80% of coil radius) using the Coefficient of Variation (CoV) of the field across 100000 sampled points and histograms of the Percentage Field Deviation (PFD) were plotted, yielding the Normalized Population of

Deviation (NPD). For visualization, normalized 2D field maps were also generated. This process was repeated for $N=4,6,8$ and 10 conductors.

2.5.2 Results and Discussion

Representative results (field maps and histograms) are shown on Figure 42 and the respective coefficient of variations are detailed in Table 8.

Visually, the 2D field maps for the H&D Extended (Figure 42-B) method display a noticeably larger and more uniform central green area compared to the Harpen method (Figure 42-A) across all numbers of conductors ($N=4, 6, 8$, and 10), indicating a more homogeneous magnetic field within the region of interest (ROI). This qualitative observation is quantitatively confirmed by the field histograms and the CoV values. The histograms for the H&D Extended method consistently exhibit a taller, narrower peak at a Percentage Field Deviation (PFD) of zero. This demonstrates that a much larger fraction of the sampled points within the ROI experience a field strength very close to the target value, signifying superior homogeneity. In contrast, the Harpen method's histograms are shorter and broader, indicating a wider distribution of field deviations.

As detailed in Table 8., the CoV for the H&D Extended method is consistently and significantly lower than that of the Harpen method for the same number of conductors. For instance, at $N=4$, the H&D Extended method achieves a CoV of 4.55%, which is less than half the 10.30% CoV of the Harpen method. This performance gap, while narrowing as N increases, remains substantial. At $N=10$, the H&D Extended method's CoV is 0.20%, still more than twice as good as the Harpen method's 0.42%.

Strategy	Coefficients of Variation (CoV)			
	N=4	N=6	N=8	N=10
H&D Extended	4.55%	1.54%	0.55%	0.20%
Harpen	10.30%	3.29%	1.15%	0.42%

Table 8: Coefficient of Variation (CoV) of the magnetic field, calculated within an 80% spherical ROI, for the H&D Extended and Harpen methods with $N=4, 6, 8$, and 10 conductors.

Both the qualitative field maps and the quantitative histogram and CoV data demonstrate the superior performance of the generalized H&D strategy. By strategically placing conductors at the roots of the Legendre polynomial and optimizing current ratios to cancel lower-order field inhomogeneities, the H&D Extended method produces a significantly more homogeneous magnetic field than the simpler, equally spaced Harpen approximation.

Although it is true that the H&D Extended method is superior, theoretically, it is expected that the Harpen method will eventually catch up as the number of conductors increases and will cancel all the spherical harmonics. However, in both cases, practical implementation of such a coil for small sample imaging would prove difficult due to the sheer number of individually tuned conductors. Figure 43 shows such case with $N=20$ with exceptional field homogeneity but a disputable realistic feasibility.

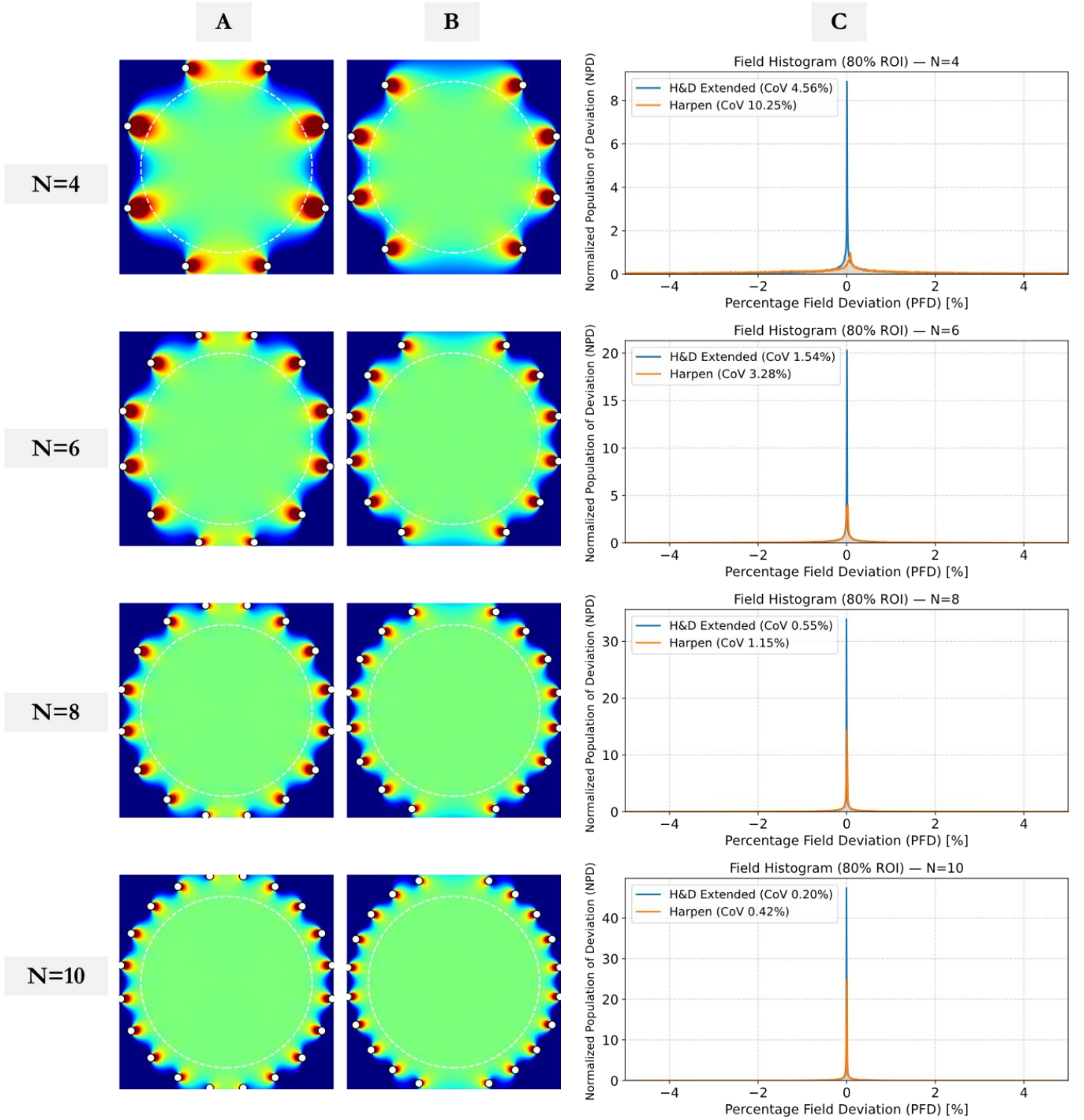


Figure 42: Comparison of the extended Hoult & Deslauriers (H&D Extended) and Harpen methods for N=4, 6, 8, and 10 conductors. (A) 2D magnetic field maps for the Harpen method and (B) for the H&D Extended method. The central green area indicates high field homogeneity, and the dashed white circle the ROI (80% of the coil volume). (C) Corresponding histograms of the Normalized Population of Deviation (NPD) versus the Percentage Field Deviation (PFD) within the ROI. The H&D Extended method consistently shows a narrower distribution, indicating superior homogeneity.

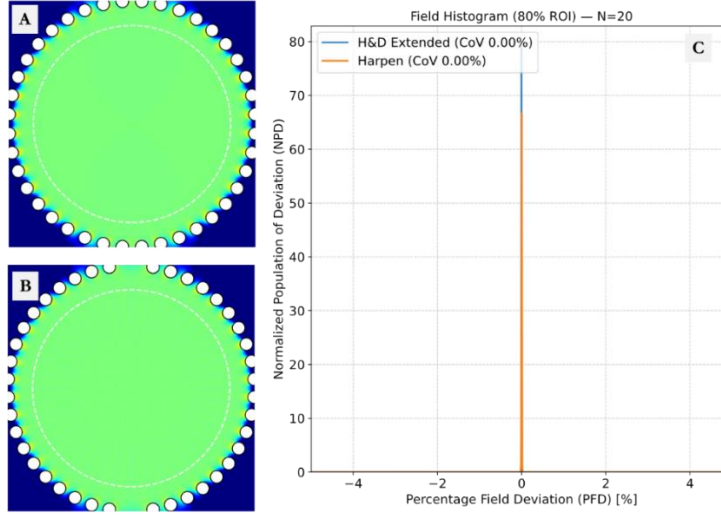


Figure 43: Comparison of the extended Hout & Deslauriers (H&D Extended) and Harpen methods for N=20. (A) 2D magnetic field maps for the Harpen method and (B) for the H&D Extended method. (C) Corresponding histograms of the Normalized Population of Deviation (NPD) versus the Percentage Field Deviation (PFD) inside the dashed white circle ROI (80% of the coil volume)

III.C. Interpretation and Discussion

The preliminary analytical comparison presented in this section provides a first-order evaluation of different RF resonator topologies. These models, based on the Biot-Savart law in a simplified free-space environment, intentionally neglect complex real-world factors such as sample loading, shield interactions, and parasitic effects to focus solely on homogeneity analysis. While they do not capture the full performance of a practical resonator, they are invaluable for establishing a theoretical baseline, understanding the fundamental origins of field homogeneity, and identifying the most promising designs for further, more rigorous investigation.

The results draw a clear distinction between the performance of the various cylindrical and spherical coil designs. Among the cylindrical topologies, the Loop-Gap Resonator (LGR) demonstrated far superior performance than the other topologies. It achieved a Coefficient of Variation (CoV) of 0.45%, which was more than twice as good as the optimized Varying-Pitch Solenoid (VPS) (0.99%) and roughly seven times better than the classical solenoid ($\sim 3\%$). This outstanding homogeneity stems from its inherent current distribution, which naturally concentrates near the cylinder edges. This physical behavior directly counteracts the "edge effects" that fundamentally limit the performance of finite solenoids. While the VPS cleverly mitigates these effects by increasing conductor density at the ends, it is essentially a remedy for an inherently ill-suited geometry for high homogeneity in constrained spaces. The LGR, by contrast, possesses a current profile that is intrinsically suited for generating a uniform field within a finite volume. Furthermore, among all the investigated models, the LGR's homogeneity proved most resilient to changes in the form factor ($FF=2R/L$), a critical parameter that links homogeneity to the sample filling factor, and by extension, to sensitivity. This means that the loop gap coils would enable homogenous form fitting coverage for a larger number of sample dimensions making it incredibly versatile.

Among the spherical topologies, the generalized Hout and Deslauriers (H&D Extended) method proved to be a significantly more efficient and powerful design strategy than the Harpen

approximation. For any given number of conductors, the H&D Extended method produced a much more homogeneous field, achieving a CoV of just 0.20% with 10 conductors, compared to 0.42% for the Harpen method. The reason for this superiority lies in its design philosophy. The Harpen method approximates an ideal continuous current distribution by simply increasing the number of equally spaced conductors while the H&D Extended method takes a more analytical approach by strategically placing conductors and optimizing their relative currents to systematically cancel specific low-order spherical harmonic terms, the primary contributors to field inhomogeneity. It can be said that the Hoult and Deslauriers approach is a targeted, "smart" design that eliminates error, while the Harpen method is a "brute-force" approximation.

In terms of design trade-offs and practical considerations, this analytical comparison highlights a clear hierarchy of performance, but the final choice of a resonator depends on a balance of competing practical factors.

- **Classical Solenoid:** This design is simple to construct but offers the worst homogeneity and is highly sensitive to its aspect ratio. Its documented instability with varying loads and susceptibility to losses make it a poor candidate for versatile, high-performance applications. It serves as a useful theoretical baseline but is not a competitive option.
- **Varying-Pitch Solenoid (VPS):** The VPS offers a substantial improvement over the classical solenoid and the advantage of homogeneity tunability via the taper parameter α . However, its performance is an optimization of a fundamentally limited solenoidal current path, and it may still inherit the stability and loading issues of its predecessor.
- **Loop-Gap Resonator (LGR/BLGR):** This design emerged as the clear winner among cylindrical topologies, not only for its superior homogeneity and form-factor resilience but also for its documented practical advantages. As a "low-E" coil, the Bridged Loop-Gap Resonator (BLGR) is known to maintain higher quality factors and exhibit minimal frequency shifts when loaded or placed near conductive surfaces, making it exceptionally stable and versatile. Its primary challenge lies in the potential complexity of fabrication, requiring the integration of capacitive gaps. The potential impact of this capacitive gap has not been evidenced through these preliminary findings. Therefore, a more thorough investigation in more realistic conditions is necessary to conclude on its supremacy.
- **Spherical Coils (H&D Extended):** This topology offers the highest theoretical homogeneity of all designs considered. The systematic method for error cancellation makes it an extremely powerful and scalable approach. However, its practical implementation is the most complex, requiring precise non-uniform conductor placement and carefully chosen capacitor values to ensure the necessary current ratios. Furthermore, a spherical geometry may not be optimal for all sample shapes, potentially leading to a lower filling factor compared to a cylindrical coil for elongated samples.

Based on this first-order analysis, the classical solenoid and the Harpen approximation are eliminated as viable candidates for a high-homogeneity resonator. Their performance is clearly surpassed by more advanced designs.

Two topologies that stand out as the most promising. The Loop Gap (LG)/Bridged Loop-Gap Resonator (BLGR) as they represent the optimal choice for a cylindrical geometry, offering an excellent combination of high homogeneity and geometric flexibility (resilience to form factor) and the Extended Hoult & Deslauriers (H&D) N-Ring Coil as it is the superior choice for a spherical geometry, providing the highest achievable theoretical homogeneity through its systematic cancellation of error fields. The Varying Pitch Solenoid (VPS) although achieving shorter homogenous regions, might outperform the loop gap in other aspects (e.g loaded quality factor) that were not explicitly investigated here and may still be a good contender despite its worst sample coverage

These three distinct yet powerful designs represent the current state-of-the-art for their respective geometries. Therefore, they will be the focus of the next section, where they will be modeled and analyzed under more realistic conditions to account for the complexities of a practical MRI environment.

IV. Realistic Coil Topology Modeling for Further Investigation

The preliminary analysis in the previous chapter served to identify the most theoretically sound RF coil topologies based on idealized free-space models. While essential for understanding fundamental design principles, these first-order approximations do not account for the complex electromagnetic interactions present in a real-world MRI environment. To obtain a more accurate prediction of performance, it's necessary to advance from analytical equations to comprehensive 3D electromagnetic (EM) simulations that incorporate conductor geometry, RF shielding, and dielectric sample loading.

Based on the superior homogeneity and design principles identified previously, three topologies have been selected for this advanced modeling phase: the Bridged Loop-Gap Resonator (BLGR) and the Varying-Pitch Solenoid (VPS) as the optimal cylindrical designs, and the Extended Hoult & Deslauriers (H&D) N-Ring Coil as the optimal spherical design. This section outlines the methodology for building and evaluating realistic models of these candidates before ultimately comparing them against each other.

IV.A. Simulation Methodology

1. Electromagnetic Simulation Environment: CST Microwave Studio

CST Microwave Studio (CST-MWS) (Computer Simulation Technology GmbH, Darmstadt, Germany) is the primary simulation tool employed for the electromagnetic simulations, RF coil characterization, and optimization of locally shielded RF resonators described in this thesis. This versatile commercial software package excels at 3D electromagnetic field simulations by solving Maxwell's equations to compute electric (E) and magnetic (H) field distributions with high accuracy.

CST-MWS offers a suite of solvers, including time-domain and frequency-domain options, enabling flexible analysis in both the time and frequency domains. Key numerical methods forming the basis of its solvers include the Finite Integration Technique (FIT), the Finite Element Method (FEM), and the Method of Moments (MoM). FIT is foundational to CST's time-domain (Transient) solver and involves the spatial discretization of the integral form of Maxwell's equations, typically on a hexahedral Cartesian grid.

The Transient Solver (often referred to as the Time Domain Solver) is highly efficient for initial broad-spectrum analysis. It calculates the electromagnetic field response over a wide frequency range from a single simulation run by exciting the structure with a time-domain pulse. Frequency-domain results, including S-parameters and resonance frequencies, are subsequently obtained via a Fast Fourier Transform (FFT).

Alternatively, the Frequency Domain Solver directly calculates electromagnetic fields at specific frequencies. This approach is advantageous for analyzing high-Q structures or when detailed field information at a known resonant frequency is required. This solver, often employing FEM, supports tetrahedral meshing, which can be more adaptive to complex or curved geometries compared to the hexahedral mesh typically used by the Transient Solver.

Another important tool is the Eigenmode Solver, which calculates the resonant frequencies and corresponding field patterns of a structure without requiring explicit excitation sources. This is particularly useful for identifying the natural resonant modes of resonators.

A key advantage of CST-MWS is its co-simulation capability, which significantly streamlines the design optimization process. This feature allows for the parametric tuning of lumped elements, such as capacitors and inductors, without needing to re-run an entire simulation each time a component value is changed. Instead, only a localized and rapid field recalculation is performed, offering substantial savings in simulation time and computational cost. This efficiency is facilitated by defining 50 Ω input ports in the coil model, which allows for evaluating impedance matching and tuning behavior through post-processing without directly inserting lumped elements into the primary simulation model.

CST-MWS also offers advanced meshing capabilities, including options such as hexahedral and tetrahedral meshes, alongside adaptive mesh refinement to improve accuracy in regions of high field concentration or complex geometry. Furthermore, specialized techniques such as "Perfect Boundary Approximation" (PBA) and "Thin Sheet Technology" (TST) enhance the accuracy of modeling curved surfaces and thin conductive layers, especially when using a Cartesian mesh, which is particularly useful for the fine features often encountered in RF resonators.

Comprehensive material modeling is another critical feature, enabling the definition of various material properties, including dielectrics (with specified permittivity and loss tangent) and conductors (with specified conductivity). For this study at 0.3T, copper traces with adequate thickness are used as the primary conductor, and appropriate phantom materials are modeled on standard biological sample properties.

2. Simulation Setup

2.1. Solver Selection

Given the focus on accurately determining resonant frequency, quality factor (Q) and B1 field homogeneity when loaded with a phantom, parameters often requiring high precision and detailed field information at specific frequencies, the Frequency Domain Solver within CST-MWS was primarily selected for this investigation. Its suitability for analyzing high-Q structures and providing detailed field characteristics at known resonant frequencies makes it the preferred choice.

As the investigated coils will only operate at the Larmor frequency of 13.03 MHz at 0.3T, a frequency-domain solver with a narrow sweep around resonance is sufficient to determine the circuit's S-matrix. Furthermore, given the coil's small dimensions compared to the wavelength at 13.03 MHz (~ 20 m in free space), wave propagation effects are minimal, and the system can be treated under the quasi-static approximation. Therefore, coils will be characterized in a steady-state regime and propagation effects will be considered negligible.

2.2. Convergence Criteria

The residual r is selected to balance simulation time and accuracy was set to the default 10^{-4} value of CST MWS as, besides lengthening simulation time, less than a 1% difference in S_{11} and quality factor with values 10^{-5} and 10^{-6} was observed.

2.3. Meshing

A tetrahedral mesh was chosen for its ability to cover complex geometries. To ensure the accuracy and independence of the results from the mesh discretization, the solver's adaptive mesh refinement feature was used. CST MWS uses a specific convergence criterion based on the difference between the S-matrices of two successive iterations of the refinement process. CST can also analyze locally how much a mesh area contributes to this difference. If the difference drops below the set criterion, the mesh is considered optimal in that specific area. Otherwise, the mesh is refined in regions that have the most significant effect on this deviation. The simulation was

configured to automatically refine the mesh until the change in the S-parameters between subsequent passes was less than 1%.

2.4. Boundary Conditions

Open boundary conditions were applied to all external faces of the simulation domain to approximate an unbounded environment and prevent artificial reflections of electromagnetic energy. The computational domain was extended by 20 mm beyond the outer surface of the RF shield to ensure sufficient clearance for the attenuation of radiated fields. This distance was selected based on a preliminary convergence test. Since the shield itself serves as a physical boundary in the real system, no additional perfect electric conductor (PEC) walls were introduced, avoiding cavity resonances or waveguide effects that could alter the coil's impedance and quality factor.

2.4.1 Preliminary Boundary Convergence Test

To figure out optimal boundary condition at 13.03 MHz a small preliminary test was performed. A makeshift ideal (a single discrete port on the gap) unloaded BLGR with a diameter of 42 mm and a length of 65 mm was used. Both shielded and unshielded configurations were explored. For the shielded configuration a 92×92×115 mm hollow copper box was placed around the centered coil. A parametric sweep was performed, and the bounding box dimensions were incrementally increased from 10 mm from the coil to 1000 mm. The coil quality factor was evaluated at each increment using CST MWS's built in coil parameter RLCQ template based post processing which calculates intrinsic coil resistance, inductance and quality factor from the S_{11} frequency response. Figure 44 shows the results of the preliminary boundary convergence test.

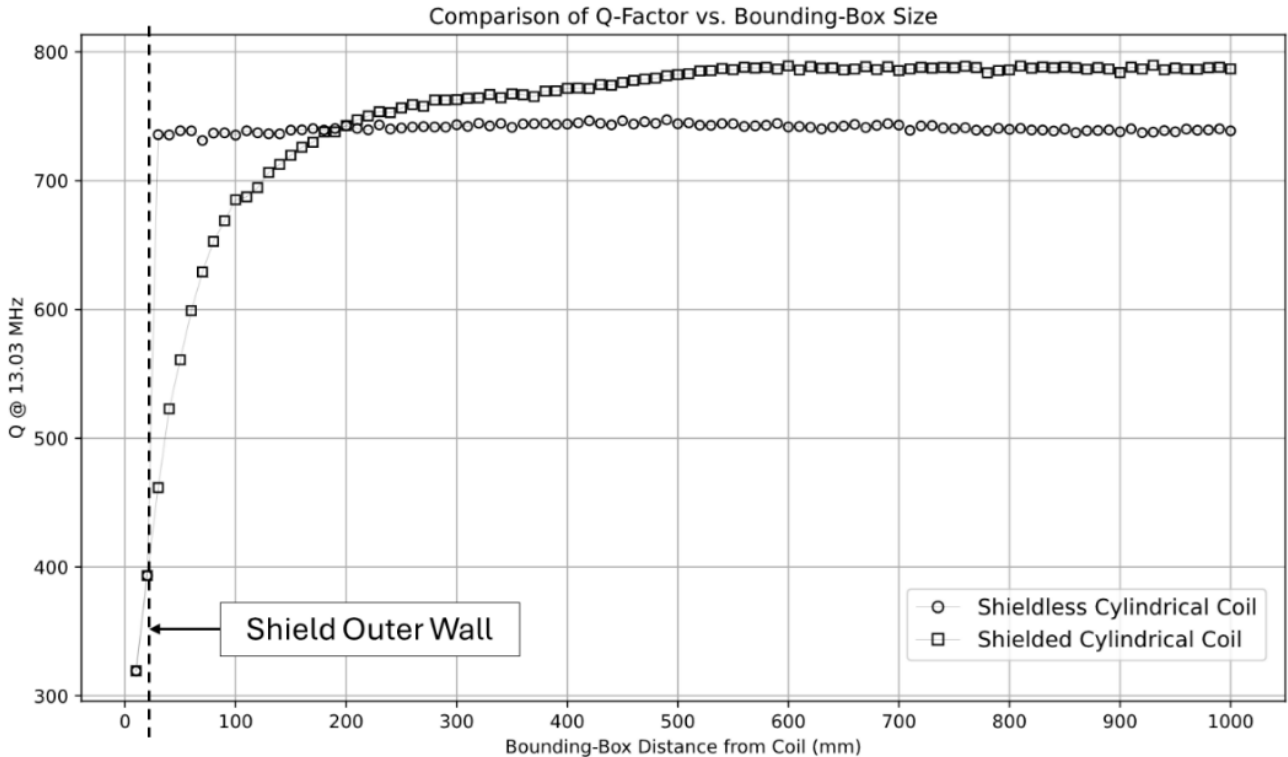


Figure 44: Preliminary convergence analysis for optimal boundary condition definition near 13 MHz. Comparison between a shielded and shield less configuration with bounding box size affecting cylindrical coil quality factor.

From the preliminary convergence test, it is possible to conclude that the bounding box size has no impact on the shielded coil and for this reason the proposed 20 mm bounding box beyond the RF shield was maintained for all shielded simulations. Conversely, the unshielded coil quality factor showed a less than 1% variation for bounding box distances from the coil above 500 mm. Therefore, if unshielded simulations are considered, the bounding box distance should be superior or equal to 500 mm.

It should be noted that this convergence test was carried out on an ideal unloaded cylindrical coil, chosen for its simplicity and well-defined resonance. While the exact numerical values may differ for other topologies or under load, the underlying conclusion that shielded configurations are largely insensitive to boundary distance while unshielded ones require larger computational domains remains. This simplification was adopted for all subsequent coil simulations, with the caveat that it represents a practical but idealized assumption.

2.5. Port definition and Excitation

To analyze the resonator's performance, the structure needs to be excited, and its response evaluated. In CST MWS, several port types can be defined (e.g., waveguide, discrete, lumped). Discrete ports are commonly used to represent lumped sources and to create intermediate connection points between the 3D electromagnetic model and a co-simulated circuit network, enabling excitation and measurement at specific locations. Discrete ports both deliver excitation and absorb reflected power, enabling direct calculation of S-parameters within the simulation at specific positions.

Different excitation signals are used depending on the selected solver. For the Frequency Domain Solver, the simulation is performed at discrete frequency points or over a defined sweep range. In this case, the default excitation is a continuous wave sinusoidal signal at the specified frequency or frequencies of interest.

In the current simulation setup, excitation power will be set to 1W, and discrete ports will be used to define each model lumped elements to later be solved through the co-simulation scheme.

3. Definition of the Standardized Models Used for Simulation

3.1. Free Space

The space within the computational domain not occupied by other materials, air or vacuum, is defined with $\epsilon_r = 1$ and $\mu_r = 1$.

3.2. Sample

Throughout the entire investigation process, the sample will be modeled as either a 38 mm diameter sphere or a 38 × 38 mm cylinder for spherical and cylindrical topologies, respectively, centered inside the coils. Figure 45 shows the proposed sample geometries.

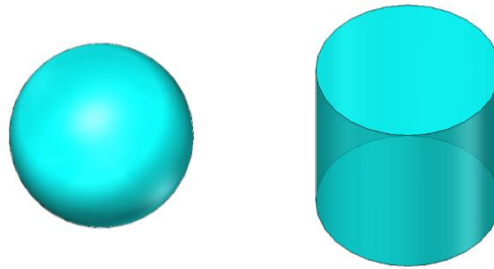


Figure 45: Sample geometries used for loaded simulations: (A) The 38 mm diameter spherical sample used to load the spherical shaped coils. (B) The 38 × 38 mm cylinder used to load the cylindrical coils (VPS and BLGR)

In the first simulation stage, the samples were defined as a 0.9% NaCl saline which was modeled with a permittivity $\epsilon_r = 79.2$ for a loss tangent $\tan \delta = 25.25$ and a conductivity $\sigma = 1.45$ S/m (values obtained using the Stogryn [151] and Klein & Swift[152] models for saline water).

3.3. The Shield

The design was constrained by the maximum available volume of the selected MRI system (180 mm×55 mm×220 mm). In benchtop systems, the shield is often integrated closely with the RF coil, forming a "local shield." The close proximity of an RF shield can induce eddy currents on its inner surfaces, which in turn modify the coil's inductance and contribute to additional resistive losses due to energy dissipation in the copper walls, leading to power dissipation. The local conductive shield generally decreases the inductance due to the formation of image currents that oppose the original magnetic field, effectively reducing the coil's self-inductance and adding another resistive loss component, further reducing the coil's Q-factor.

Therefore, the effective quality factor of a loaded shielded coil can be given by the sum of these contributions:

$$Q_l = \frac{\omega_0 L_{\text{eff}}}{R_{\text{eff}}}, \quad (\text{IV-1})$$

with:

$$L_{\text{eff}} = L - L_{\text{shield}} \pm L_{\text{sample}} \quad (\text{IV-2})$$

$$R_{\text{eff}} = R_c + R_{\text{shield}} + R_{\text{sample}}, \quad (\text{IV-3})$$

($L_{\text{shield}}, R_{\text{shield}}$) being the contributions from the shielded enclosure and ($L_{\text{sample}}, R_{\text{sample}}$) the contributions from the sample.

To mitigate these effects, the shield was modeled as a hollow copper box with 25 μm thick walls spanning the entire available volume making it as large as possible. The shield thickness was selected be greater than the skin depths at the operating frequency (18.06 μm at 13.03 MHz) while considering constraints related to gradient-induced eddy currents [27].

The shield was subsequently artificially grounded by adding a discrete port between the cooper box and floating copper rails of the same thickness placed 5 mm below on either side of the box. The placement of the rails on either side was purposefully selected to maintain symmetry and mimic the physical rails of the magnet design Figure 29. The rails were grounded in the co-simulation setup forming a common ground between the investigated coil and the shield. Figure 46 shows the shield model in detail.

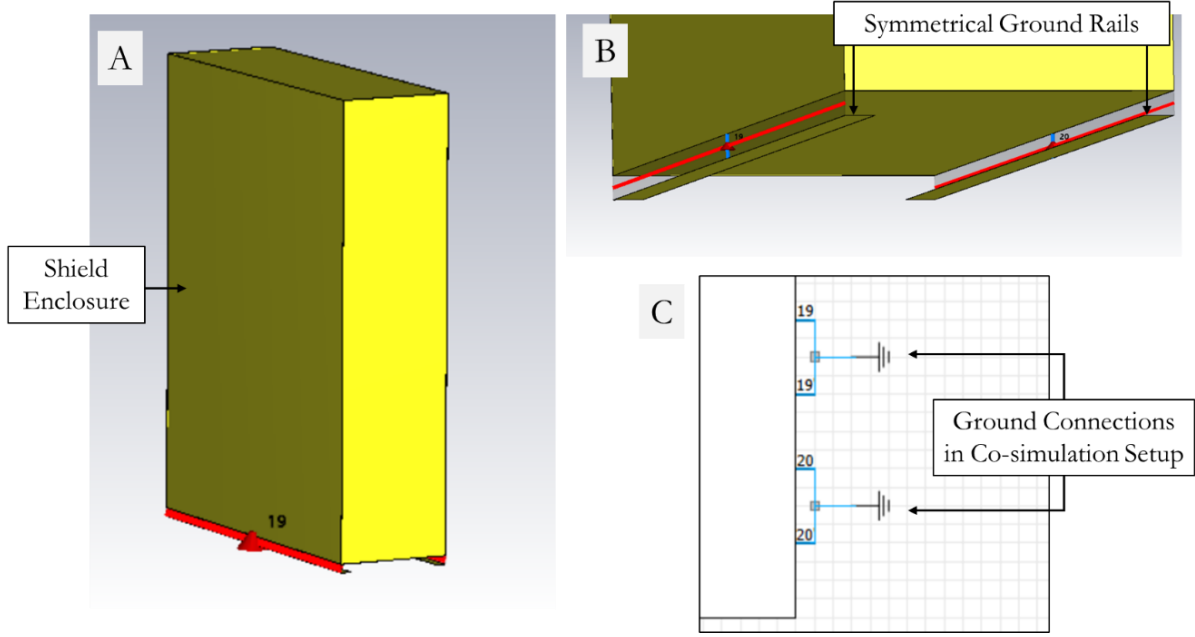


Figure 46: Shield enclosure model and ground rails. (A) Copper shield enclosure (25 μm thick) (B) Symmetrical floating copper rails (25 μm thick) placed 5 mm below the shield to maintain structural symmetry and allow artificial grounding. (C) Partial co-simulation schematic showing shield and rail connection common with ground reference.

3.4. Coil Conductor

Coil conductors were modeled as 40 μm thick pure copper strips with a conductivity of 5.8×10^7 S/m. This specific thickness was selected to be almost twice the copper skin depth at 13.03 MHz (18.06 μm).

3.5. Coil Substrate

No physical substrate was modeled in these simulations because the objective was to evaluate the intrinsic electromagnetic performance of each coil topology in the presence of the sample and RF shield, which were explicitly included, while avoiding material-specific uncertainties associated with substrate selection. Introducing a substrate at this stage would add material-dependent dielectric effects such as frequency detuning, parasitic capacitance, and additional losses that depend on the specific choice of polymer and fabrication process.

At 13 MHz, the electromagnetic wavelength in typical polymer substrates is on the order of several meters, making their influence on the magnetic field distribution negligible compared to the dominant loading induced by the conductive sample and their effect is primarily limited to minor resonance shifts and dielectric losses. For this reason, the substrate was omitted to isolate the geometric and topological contributions to coil performance during the initial design stage. Substrate-related effects can be addressed in subsequent implementation steps, once the material and fabrication parameters are defined, however, their impact on the overall performance of the topologies under investigation is considered minimal due to the low dielectric constant and low loss tangent of common polymers, which contribute far less to energy dissipation than the conductive sample and metallic shield. For example, the High Temp V2 (HTV2) resin (FLHTAM02) from Formlabs' engineering resin line, a material commonly used in the R3DP method, boasts only a permittivity $\epsilon_r = 3.33$ for a loss tangent $\tan \delta = 0.019$ (@ 13.56 MHz [153]) compared to a 0.9% saline solution with a permittivity $\epsilon_r = 79.2$ for a staggering loss tangent $\tan \delta = 25.25$.

3.6. Lumped Elements

Lumped components play a critical role in the designs under investigation. Their characteristics may very well be a strong limiting factor due to their significant number in some designs (at least one for each ring in the H&D extended approach and the entire capacitive gap in the BLGR). Therefore, High Q non-magnetic MLCCs from Knowles Syfer were selected and each element was modeled using manufacturer-provided S-parameter data, obtained from the CAPCAD library [154], to ensure accurate representation of their frequency-dependent characteristics.

3.7. Co-Simulation Setup

A single port was defined across the coil feed to emulate the preamplifier input. All EM models were co-simulated with a balanced L-shaped matching network as this balanced matching network topology has been shown to drastically improve coil stability and sensitivity[104], [155]

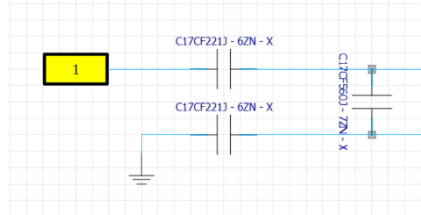


Figure 47: Example of the single port and balanced L-network configuration used in all co-simulation setups.

4. Extracting Core Performance Metrics from EM Simulations

4.1. Quality Factor

4.1.1 Smith Chart–Based Q Evaluation (Kajfez–Hwan method)

Individual coil S-parameters were exported as SnP files from the co-simulation setup. The Q-factors were evaluated directly from the frequency response S_{11} using the graphical Smith chart-based method described by KAJFEZ and HWAN [156]. This graphical method, although less widely used, was selected in order to accurately measure the Q-factor no matter the coupling/matching conditions, which can vary greatly depending on loading condition. This allows for accurate Q-factor comparisons without having to systematically retune and rematch the coil.

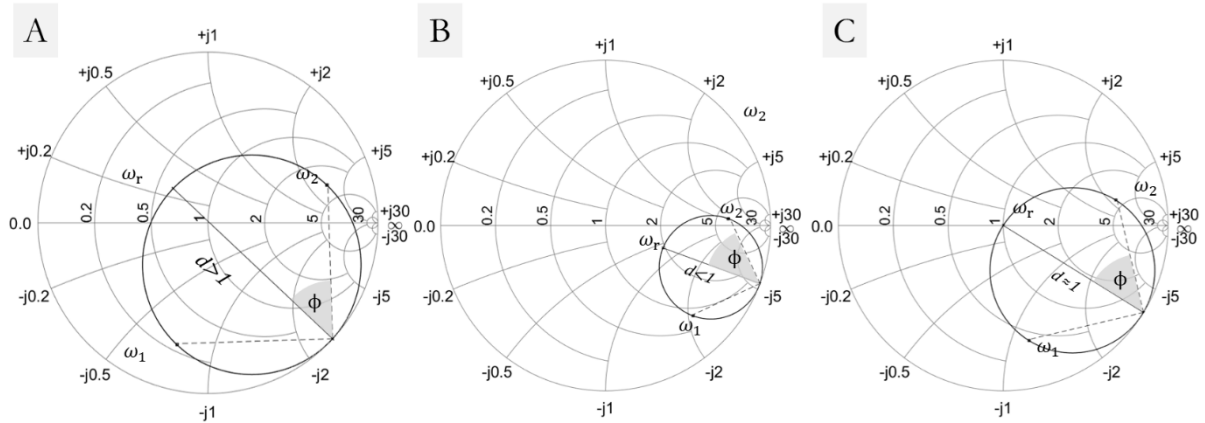


Figure 48: Smith charts illustrations of the graphical Q-factor evaluation method proposed by Kajfez and Hwan [156] for different coupling configurations. (A) Over coupled, (B) under coupled, and (C) critically coupled.

The coupling correction factor was determined from the measured diameter d of the impedance loop as

$$\kappa = \frac{d}{2 - d} \quad (\text{IV-4})$$

The measured Q-factor was determined using

$$Q_{meas} = \frac{\omega_r}{|\omega_1 - \omega_2|} \tan(\phi) \quad (\text{IV-5})$$

where ω_r is the resonant angular frequency and $|\omega_1 - \omega_2|$ is the frequency difference between two points corresponding to a chosen phase angle ϕ . The intrinsic Q is then obtained by correcting for the coupling effect:

$$Q = Q_{meas} \cdot (1 + \kappa) \quad (\text{IV-6})$$

It is interesting to note that if the coils are perfectly matched to the characteristic impedance $Z_0 = 50\Omega$, and, are therefore in a critically coupled ($\kappa = 1$) configuration, for $\phi = \pi/4$, $\Delta\omega = |\omega_1 - \omega_2|$ is the bandwidth at the -3 dB level of the frequency response which corresponds to the well-established Q-factor estimation formula for well-matched coils:

$$Q = \frac{2\omega_r}{\Delta\omega} \quad (\text{IV-7})$$

4.1.2 Q-factor from CST's "Coil_Capacitor parameters (1-port)"

Q-factors were also obtained using CST Microwave Studio's built-in "*Coil_Capacitor parameters (1-port)*" template". This routine converts the simulated one-port response into admittance

$$Y_{11} = G + jB, \quad (\text{IV-8})$$

and evaluates the parallel quality factor as

$$Q_p = \frac{|B|}{G} = -\frac{\Im\{Y_{11}\}}{\Re\{Y_{11}\}}. \quad (\text{IV-9})$$

Series equivalents are accessible from the impedance $Z = 1/Y$, with

$$R_s = \Re\{Z\}, \quad L_s = \frac{\Im\{Z\}}{2\pi f}, \quad Q_s = \frac{2\pi f L_s}{R_s}. \quad (\text{IV-10})$$

The method is restricted to single-port setups but provides a rapid, automated intrinsic Q estimate. It assumes a single-port lumped RLC representation of the coil and remains valid when shields or samples are included in the simulation, as these simply modify the effective R , L , and Q . It is however not applicable to multi-port configurations where the one-port assumption is no longer valid. This method is particularly suited to quick parametric sweep measurements.

4.2. B₁ Field and H Field Maps and Homogeneity

To extract B₁ field maps, H-field monitors, coupled with the integrated post-processing MRI Toolbox, were used [157].

On a rectilinear grid with uniform voxel volume ΔV , the extracted magnetic field magnitude is reconstructed from real/imaginary Cartesian components as

$$|\mathbf{B}|_i = \sqrt{|B_x|_i^2 + |B_y|_i^2 + |B_z|_i^2}, \quad |B_\alpha|_i^2 = (\Re\{B_\alpha\}_i)^2 + (\Im\{B_\alpha\}_i)^2, \quad (\text{IV-11})$$

so that $|\mathbf{B}|_i^2$ serves as the energy–density weight (for $\mathbf{H}$ data, $|\mathbf{B}|^2$ is replaced by $|\mathbf{H}|^2$).

The magnetic field distribution was calculated over a specified volume. 3D Field Maps were exported as HDF5 files and treated using a custom Python script to evaluate the coefficient of variation (CoV) and field distribution histograms over the sample region.

For quantitative assessment of B_1 field homogeneity, the coefficient of variation was selected as the primary metric, since it directly reflects the relative spread of field values within the region of interest. A lower CoV corresponds to a narrower distribution and improved spatial uniformity. Although histogram analysis provides complementary insight into the distribution of field deviations, the location of the histogram peak mainly indicates whether the field is centered around the mean rather than how uniformly it is distributed. Consequently, coils will be compared and ranked according to their CoV values, while the histogram peak position will be used as a secondary descriptor of field bias.

Similarly, for preliminary field analysis, H-field maps were employed as they do not require co-simulation setups and are therefore better suited for broad parametric sweeps. While these maps provide a rapid estimate of the spatial field distribution, they do not account for realistic circuit loading and losses. Such effects were only captured in the more time consuming and stringent co-simulation steps, which were used for the final quantitative assessment of each design.

4.3. Filling Factor

The magnetic energy in a region Ω can be defined as

$$U_\Omega \propto \sum_{i \in \Omega} |\mathbf{B}|_i^2. \quad (\text{IV-12})$$

With constant factors ΔV and $1/(2\mu_0)$ canceling out, the filling factor was defined as the fraction of energy inside the sample region relative to the total field energy in all space (approximated in practice by the total magnetic energy contained within the RF shield volume, assuming the shield fully encloses the field distribution):

$$\eta = \frac{U_s}{U_{\text{tot}}} = \frac{\sum_{i \in \text{sample}} |\mathbf{B}|_i^2}{\sum_{i=1}^M |\mathbf{B}|_i^2}. \quad (\text{IV-13})$$

The filling factor was therefore numerically estimated using a custom Python script from the extracted magnetic field maps over the entire shield volume.

IV.B. Bridged Loop-Gap Resonator (BLGR)

1. Modeling Approach

For the comparative study, the Bridged Loop-Gap Resonator was modeled by a 40 μm thick hollow cylinder of radius $R=21$ mm, spanning an axial length $L=65$ mm. A narrow longitudinal gap of 3.7 mm was cut on one side along the cylinder axis to serve as the resonant gap and accommodate 1111 (2.79mm W x 2.79mm L x 2.59mm H) High Q SMD Ceramic Capacitors.

Along the gap, small 0.4 mm long and 2.8 mm wide rectangular protrusions were added to mark the capacitor placements and accommodate the discrete ports which will be used for the subsequent co-simulation steps.

In order to be operated at lower frequencies, the gap of the classical loop-gap resonator must be bridged using non-magnetic, high- Q multilayer chip capacitors (MLCCs). In this work, 1111-size SMD ceramic capacitors (2.79 mm $\times$ 2.79 mm $\times$ 2.59 mm) were considered and placed equidistantly along the longitudinal gap. The number of capacitors along the axial gap strongly influences both electrical and electromagnetic performance. Ideally, the entire span of the gap is covered so that the total capacitance is distributed evenly, and the RF voltage is shared among many elements, which stabilizes tuning, reduces stress on any single lumped element, and helps preserve the current symmetry which is essential for field homogeneity. For N capacitors of value C_i in parallel across the gap,

$$C_{\text{tot}} = \sum_{i=1}^N C_i \approx N C, \quad f_0 = \frac{1}{2\pi\sqrt{L C_{\text{tot}}}}. \quad (\text{IV-14})$$

Because the goal is to maximize the unloaded quality factor Q , the choice of N is not trivial. On the one hand, increasing the number of capacitors reduces the effective series resistance and inductance per branch approximately as

$$R_{s,\text{eq}} \approx \frac{R_{s,i}}{N}, \quad L_{s,\text{eq}} \approx \frac{L_{s,i}}{N} \quad (\text{IV-15})$$

which lowers capacitive loss, spreads the voltage across the gap ($V_i \approx V_{\text{gap}}/N$), and homogenizes the electric field distribution. These effects are beneficial for improving the unloaded quality factor and preventing local dielectric breakdown. Therefore, for the present BLGR geometry (65 mm axial gap length, 40 μm copper thickness, 42 mm diameter, operated near 13 MHz) an $N=20$ capacitor covering almost the entirety of the gap was considered (BLGR_20C).

On the other hand, beyond a certain count, additional capacitors may cease to provide proportional benefit. Each component is expected to introduce pad area and solder joints, which couple capacitively to the shield and increase distributed stray capacitance. This effect would scale approximately linearly with N , particularly when the shield is close. Moreover, the intrinsic dielectric loss of the MLCCs, while low, does not vanish, and adding more dielectric volume does not necessarily reduce the total loss. From a practical standpoint, increasing the number of components also raises manufacturing complexity, assembly cost, and the likelihood of parasitic effects due to near-continuous metallization along the gap. For these reasons, a second configuration presenting a compromise was also considered with 13 equally spaced MLCCs (BLGR_13C). At this pitch (~ 5 mm center-to-center), the capacitors are sufficiently numerous to reduce per-component voltage, smooth the electric field distribution, and lower the effective ESR/ESL, while leaving adequate spacing (~ 2 mm) between packages to avoid pad-to-pad fringing and excess parasitic capacitance to the shield. Figure 49 shows an example of the BLGR model for 13 capacitors. Images of the 20-capacitor model associated co-simulation schematic will be presented in Annex III.

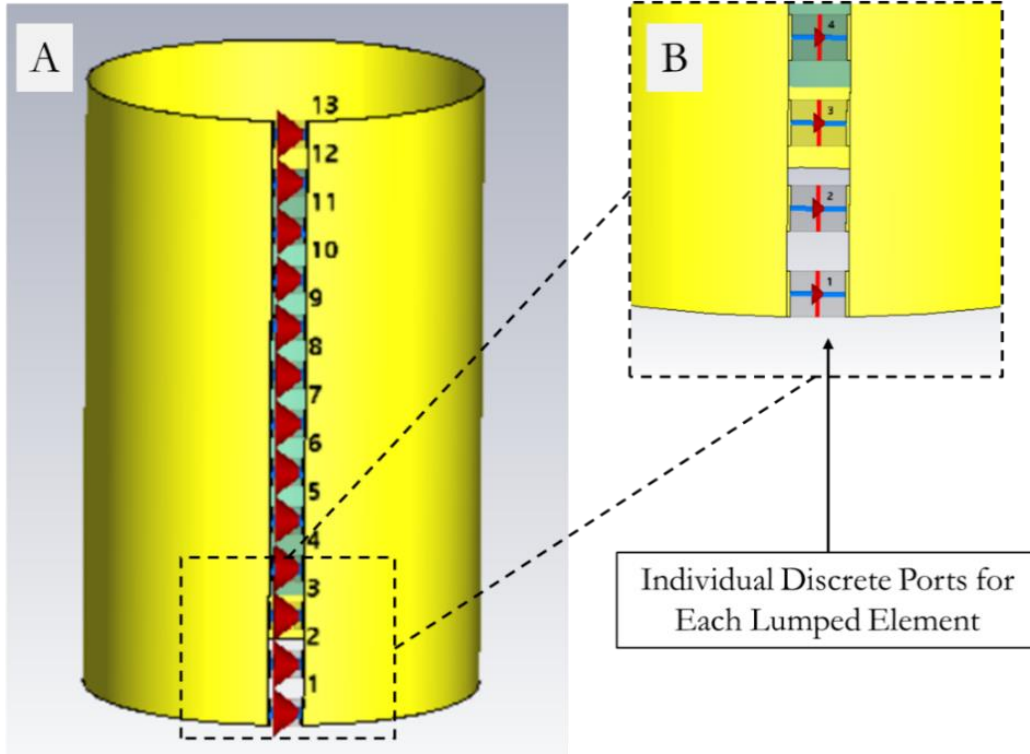


Figure 49: 3D model of the selected 13 capacitor bridged loop-gap resonator. (A) Full geometry showing the placement of 13 1111-size MLCCs. (B) Detailed view of the discrete lumped element ports used to model each capacitor and its associated solder pads.

2. Full-Wave Simulation and Co-Simulation Results

The performances of the 20-capacitor (BLGR_20C) and 13-capacitor (BLGR_13C) Bridged Loop-Gap Resonator designs were evaluated using a co-simulation approach. The schematic for the 13-capacitor co-simulation setup is shown in Figure 50, where each discrete port in the 3D model is connected to a corresponding capacitor model in the circuit solver.

The simulated S-parameter (S_{11}) results for both configurations are presented in Figure 23A. Both resonators were tuned to a resonant frequency of approximately 13.03 MHz. A significant difference in electrical performance was observed, primarily in the loaded quality factor (Q_ℓ). The BLGR_20C design achieved a Q_ℓ of 374, which is over 70% higher than the Q_ℓ of 215 obtained for the BLGR_13C. This result aligns with the theoretical model, which predicts that increasing the number of parallel capacitors reduces the total equivalent series resistance (ESR) of the capacitive bridge, leading to lower overall resonator losses and a correspondingly higher Q-factor.

Figure 50-B shows the normalized distribution of the magnetic field deviation from the mean, with the Coefficient of Variation (CoV) used as a quantitative measure of field uniformity. Both designs demonstrate excellent homogeneity, with the BLGR_13C having a slightly lower CoV of 1.58% compared to 1.63% for the BLGR_20C. This indicates that both configurations produce a highly uniform field in the region of interest.

Visualizations of the magnetic field distribution in the axial and transverse planes are provided in Figure 52. These plots confirm the quantitative findings. For the N=20 case, the field is exceptionally smooth along the entire length of the resonator. In contrast, the N=13 case exhibits a visible periodic "ripple" or modulation in the field intensity near the resonator wall, corresponding to the larger spacing between the discrete capacitors. However, consistent with the

CoV data, the field in the ROI (indicated by the dashed lines) remains highly uniform for both designs.

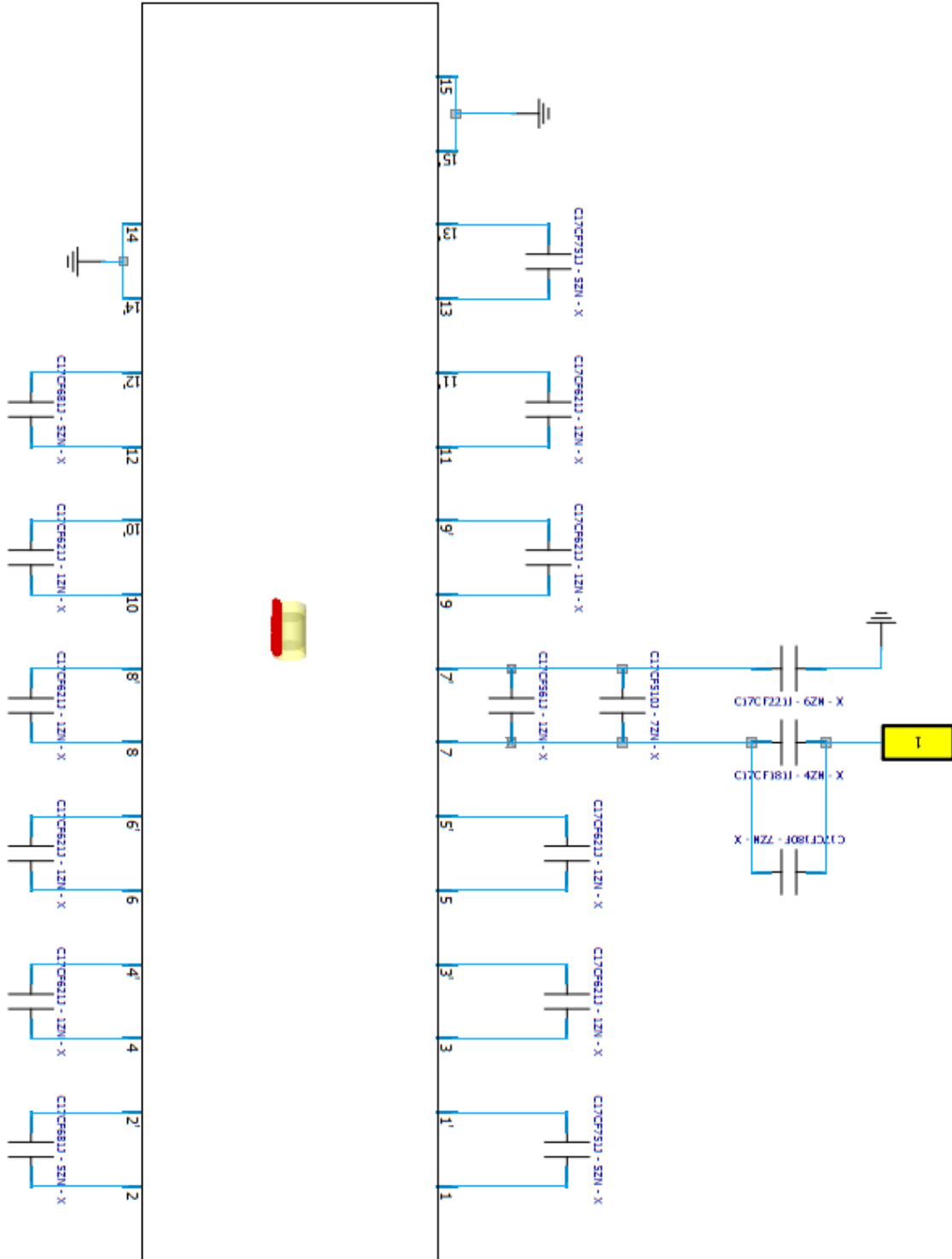


Figure 50: Co-simulation schematic for the 13-capacitor BLGR (BLGR_13C). Each port in the 3D EM model is connected to a corresponding lumped element model representing SMD capacitors.

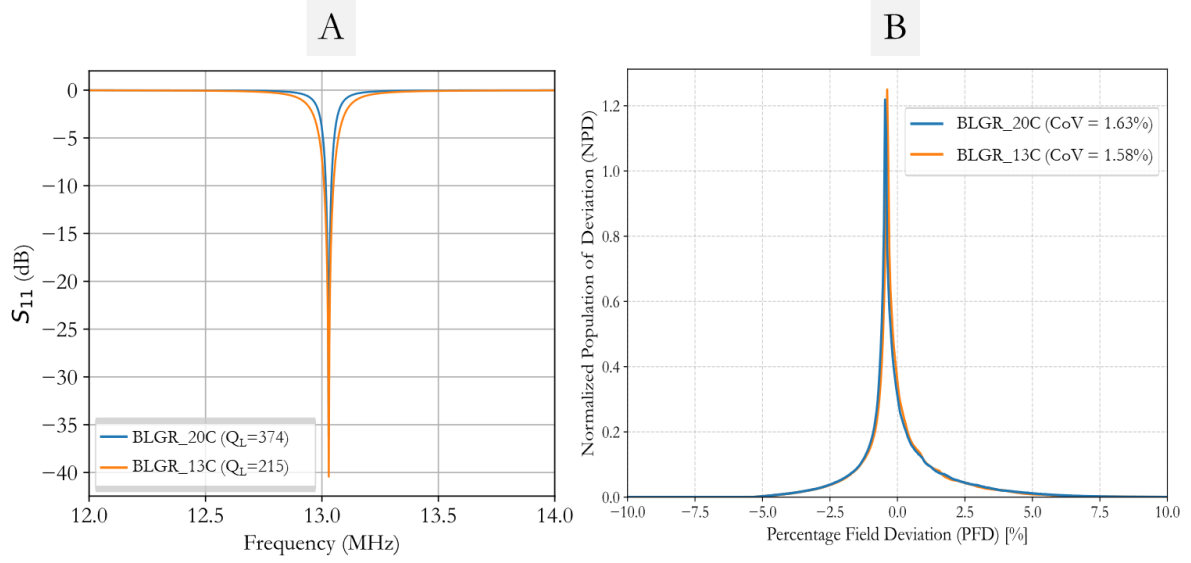


Figure 51: Simulated performance comparison between the 20-capacitor (BLGR_20C) and 13-capacitor (BLGR_13C) designs. (A) S-parameter (S_{11}) response, showing the resonant frequency and loaded quality factor (Q_L). (B) Normalized distribution of the Percentage Field Deviation (PFD), quantifying the magnetic field homogeneity.

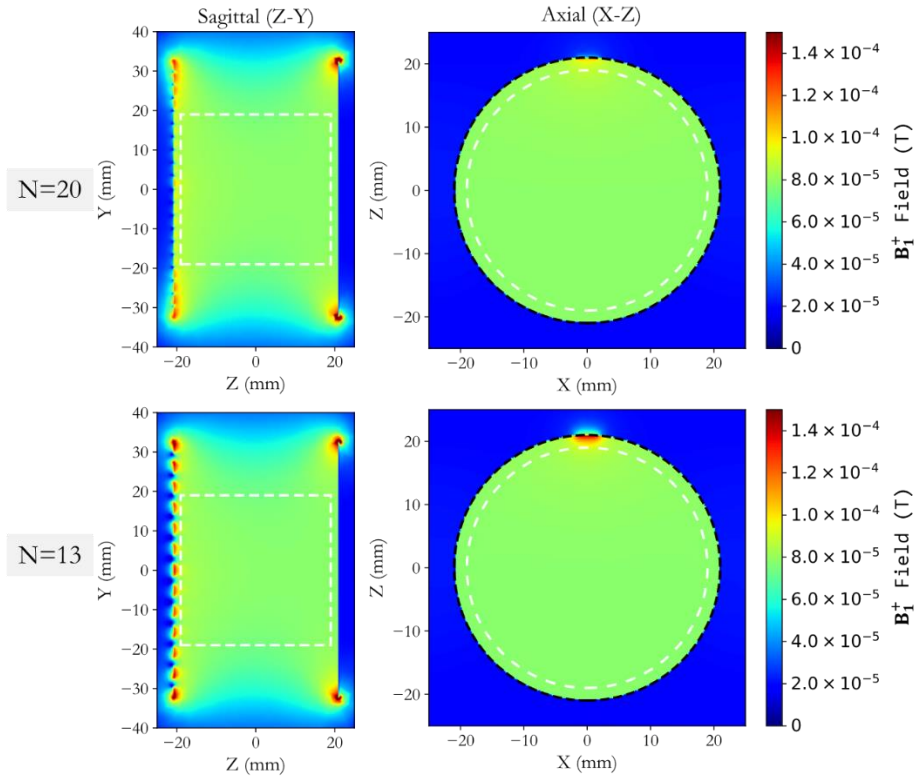


Figure 52: Simulated magnetic field (B_1) distribution for the N=20 (top row) and N=13 (bottom row) configurations. The left column shows the field in the axial plane, and the right column shows the transverse plane at the resonator's center. The dashed lines indicate the region of interest for homogeneity analysis.

The simulation results demonstrate a clear trade-off. The BLGR_20C configuration offers substantially superior electrical performance with a much higher quality factor. While both designs produce a highly homogeneous magnetic field, the BLGR_13C presents a viable alternative that simplifies manufacturing complexity at the cost of a lower Q-factor.

IV.C. Varying Width Ribbon Solenoid (VWRS)

As presented in section III.A, the VPS improves upon the classical solenoid by modulating the spacing between its windings making them tighter at the ends and looser in the center to increase the current density at the coil's edges to counteract the natural drop-off in field strength, significantly improving homogeneity within a compact form factor. Previously, the coil was described as a single ideal current path following the trajectory set by Equation (III-7).

However, to maintain a fixed coil length while loosening the central windings, the pitch angle in the center must increase. As the central pitch widens, the axial component of the current at the center of the coil can no longer be expected to become negligible compared to the azimuthal component, creating drop in homogeneity near the central windings and reducing the width of the available homogenous region (see Figure 35).

As a remedy, Privalov et al.[158] have proposed a design that utilizes a thin ribbon of continuously variable width. This method allows for an increase in the density of turns at the coil ends, which corrects for inhomogeneity in the axial direction, similar to variable pitch designs. However, by varying the ribbon's width instead of the gap between the turns, the gap can be kept constant and narrow. This minimizes field variations near the conductor, leading to improved homogeneity in the radial direction as well.

1. Modeling Approach

For the comparative study, a Varying Width Ribbon Solenoid (VWRS) following the same concept as Privalov et al.[158] was used. The VWRS was modeled by a 40 μm thick ribbon conductor wound in the shape of circular cylinder of radius $R=21$ mm, spanning an axial length $L=65$ mm over $N=10$ turns. The ribbon is described as a smooth surface embedded in $\mathbb{R}^3$ via a two-parameter map given by:

$$\begin{cases} x(u, v) = R \cos u \\ y(u, v) = R \sin u \\ z(u, v) = z_{\text{bot}}(u) + v w(u) \end{cases} \quad \begin{matrix} u \in [0, 2\pi N], \\ v \in [0, 1], \end{matrix}$$

with

$$z_{\text{bot}}(u) = -\frac{L}{2} + \frac{L}{2\pi N} u - \frac{L}{2} A [\sin(u/N) (1 + C) - \cos(u/N) S] - \frac{L}{2N} + \frac{d}{2} \quad (\text{IV-16})$$

$$w(u) = \frac{L}{N} - L A S \cos(u/N) - d \quad (\text{IV-17})$$

and

$$A = \frac{\alpha}{4\pi (1 - \alpha/2)}, \quad S = \sin\left(\frac{2\pi}{N}\right), \quad C = \cos\left(\frac{2\pi}{N}\right), \quad (\text{IV-18})$$

Here $d > 0$ is the constant inter-turn axial gap and $\alpha \in \mathbb{R}$, similarly to the previous idealized path case of the VPS, is a dimensionless taper parameter. A detailed explanation of the parametric formula can be found in Annex III. Figure 53 shows examples of the applied geometry example of the ribbon solenoid model for over $N=10$, $\alpha=0.7$, $d=0.5$.

To be able to connect a discrete port for excitation small ($1.5 \times 2 \times 4$ mm) protrusions mimicking soldering pads were added at each end of the solenoid on the last turn. These extensions provide flat metallic surfaces where the lumped element port could be reliably attached in CST, ensuring stable excitation and accurate calculation of the input impedance and S-parameters without artificially altering the coil geometry.

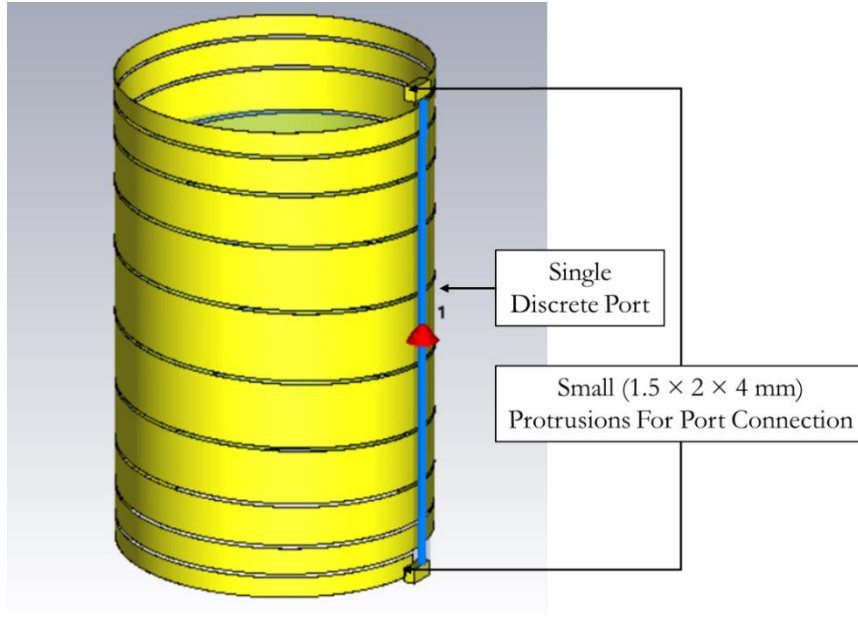


Figure 53: 3D geometric model of a VWRS coil ($N=10$, $\alpha=0.7$, $d=0.5$) showing the small added protrusions for port connection and the discrete port used for excitation.

2. Preliminary Optimization Steps

Both α and d were optimized in loaded (0.9% NaCl) and shielded conditions, through preliminary parametric simulation steps, to find the best baseline homogeneity without compromising on Q-factor.

2.1. Identifying The Optimal Taper Strength for Homogeneity

To find the taper strength (α) that produced the most uniform magnetic field, a parametric sweep was conducted for coil configurations with both $N=10$ and $N=20$ turns with a fixed gap $d=0.5$. The investigation focused on two key performance metrics: field homogeneity (Figure 54), and quality factor (Figure 55).

The analysis of the field histograms and CoV in Figure 54 reveals a clear optimal point for homogeneity. For the $N=10$ coil, homogeneity improves as α increases from 0.3, reaching a minimum CoV of 1.62% at $\alpha=0.45$. Similarly, for the $N=20$ coil, the best homogeneity is achieved at $\alpha=0.55$, with a superior CoV of 1.55%. Beyond these optimal values, field uniformity begins to degrade. Similarly to preliminary analytical study, this behavior occurs because insufficient tapering fails to fully correct the drop-off, while excessive tapering overcompensates, creating high-field regions at the ends and degrading the overall homogeneity.

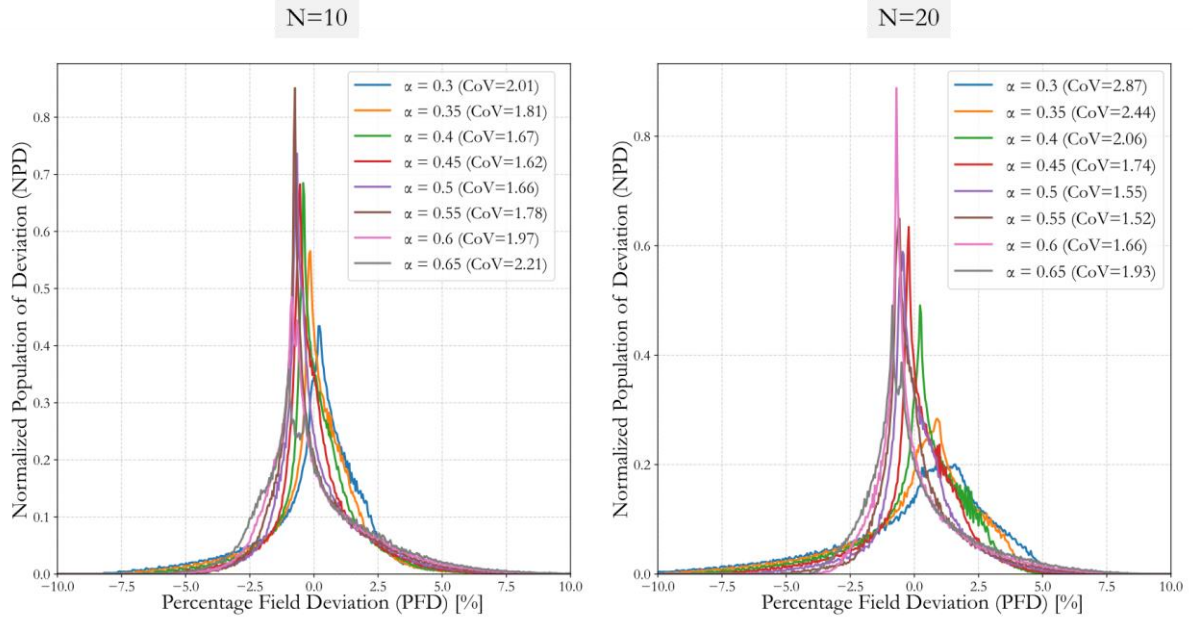


Figure 54: Field distribution histograms for varying taper strengths (α) in (A) N=10 and (B) N=20 turn coils.

Interestingly, the highest histogram peaks do not match the minimum CoV value. This is because the CoV accounts for the entire distribution's spread relative to the mean, while the peak only represents the mode. A distribution can have a high peak but also wider tails, which increases the standard deviation and thus the CoV. The optimal α is therefore the one that minimizes the overall field variation across the entire volume of interest, leading to the tightest distribution.

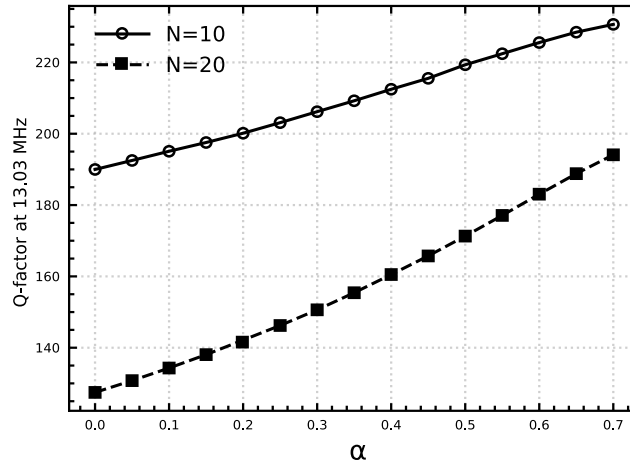


Figure 55: Loaded quality factor (Q) as a function of the taper parameter (α) for N=10 and N=20 coils under loaded conditions

In parallel, the evaluation of the Q-factor in Figure 55 shows that it increases monotonically with α . This is because a larger α corresponds to a wider conductor ribbon, which reduces ohmic losses and improves the coil's efficiency. This presents a critical design trade-off as maximizing the Q-factor by increasing α is in direct conflict with achieving the best possible field homogeneity, which requires a specific, optimal α . Given that the primary objective is to achieve the highest degree of field uniformity while maintaining the highest possible quality factor, $\alpha=0.45$ for N=10 was selected as the optimal VWRS coil in this configuration for an expected ideal loaded Q-factor of around 215 compared to 178 for $\alpha=0.55$ for N=20.

2.2. Studying The Impact of Gap Size

Following the optimization of the taper strength, the influence of the inter-turn gap size, d , on field homogeneity was investigated. For this analysis, the previously determined optimal taper parameter of $\alpha=0.45$ for the $N=10$ coil was kept constant, while the gap size was varied. Figure 56 presents the field distribution histograms for different gap sizes, ranging from $d=0.2$ mm to $d=0.8$ mm. The results clearly demonstrate that field homogeneity is highly sensitive to the spacing between the ribbon turns. As the gap size decreases, the peak of the field distribution becomes progressively narrower and taller, indicating a reduction in the overall field deviation.

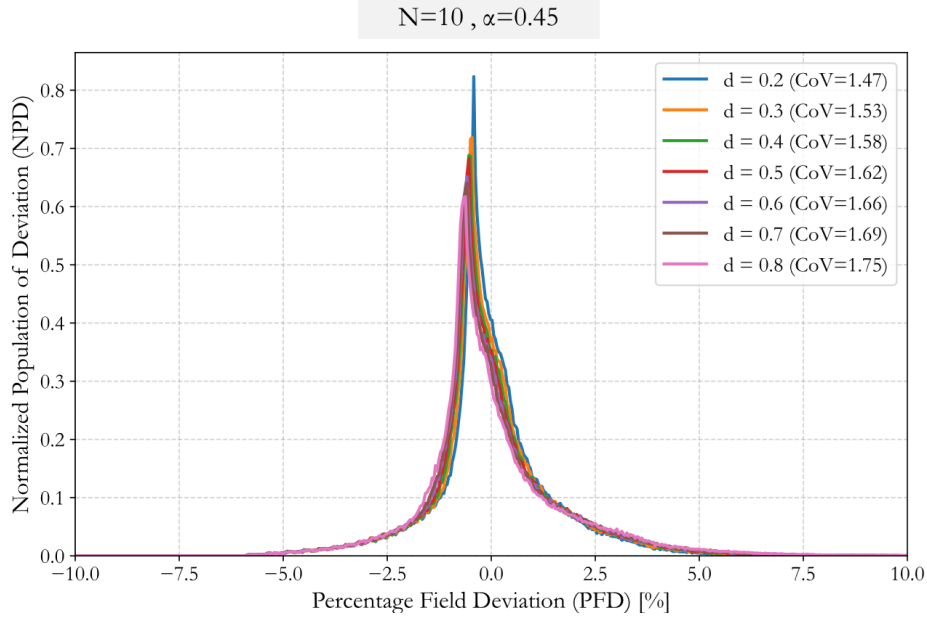


Figure 56: Field distribution histograms for the $N=10$ coil ($\alpha=0.45$) showing the impact of varying the inter-turn gap size (d). Homogeneity, indicated by the CoV, improves as the gap size decreases.

This observation is quantitatively confirmed by the Coefficient of Variation (CoV), which improves from 1.75% at $d=0.8$ mm to a minimum of 1.47% at $d=0.2$ mm. This trend suggests that a smaller gap size minimizes field variations near the conductor, leading to a more uniform magnetic field. However, this pursuit of homogeneity presents a complex practical trade-off. As shown in Table 9, the loaded quality factor is also dependent on the gap size. The Q-factor is lowest (209) at the smallest gap of $d=0.2$ mm and increases with the gap size, plateauing at 215 for $d \geq 0.4$ mm. With a difference of roughly less than 3%, it can be argued that this parameter has very little impact on the coil quality factor.

Inter-turn gap (d)	0.2	0.3	0.4	0.5	0.6	0.7	0.8
Ideal Loaded quality factors	209	213	215	215	215	215	215

Table 9: Loaded Quality Factors for Various Inter-turn Gap Sizes

While a smaller gap is electromagnetically advantageous for homogeneity, it compromises the Q-factor and imposes stricter manufacturing tolerances, increasing the risk of parasitic capacitance and potential arcing between adjacent turns. Therefore, the choice of an optimal gap size must balance the competing needs for high field uniformity, a high quality factor, and the practical constraints of fabrication and electrical safety. Keeping that in mind, an inter turn gap of 0.4 mm was selected for the final design.

3. Full-Wave Simulation and Co-Simulation Results

Following the parametric optimization, the final design for the Varying Width Ribbon Solenoid (VWRS) was established with $N=10$ turns, a taper parameter of $\alpha=0.45$, and an inter-turn gap of $d=0.4$ mm. To evaluate the performance of this optimized coil in a realistic operational context, a co-simulation was performed. The VWRS with its circuit model for impedance matching and tuning, is illustrated in Figure 57.

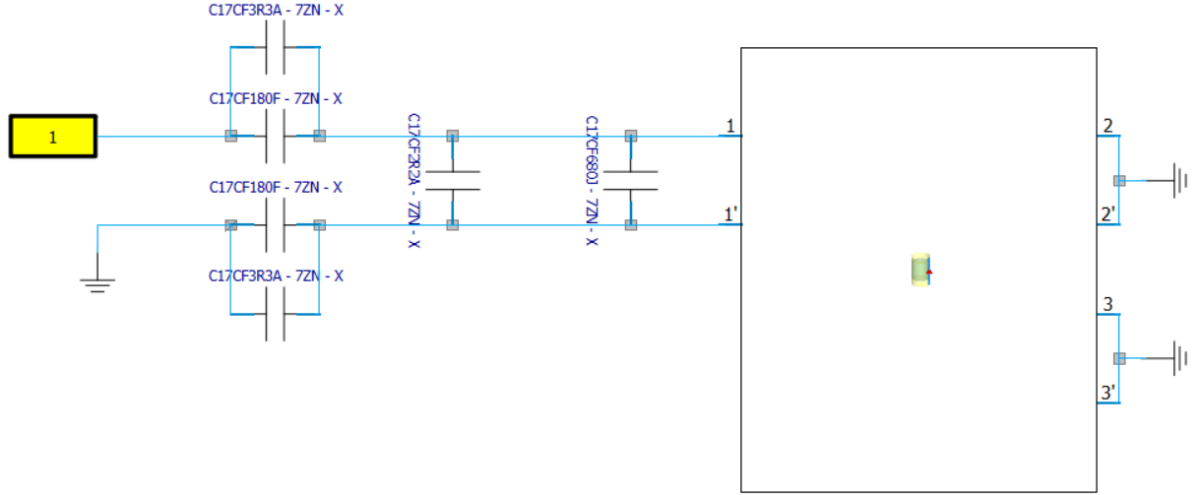


Figure 57: Co-simulation schematic in CST Design Studio. The circuit model includes a matching and tuning network connected to the 3D full-wave model of the VWRS coil.

The results of the co-simulation are presented in Figure 30. The scattering parameter (S_{11}) plot on the left shows that the coil is well-matched, with a deep resonance notch reaching below -50 dB at 13.03 MHz, indicating efficient power transfer. From this resonance curve, the loaded quality factor (Q_ℓ) was calculated to be 180. This value is slightly lower than the ideal loaded Q -factor of 215 predicted in the optimization study (Table 9). The reduction is expected and can be attributed to losses introduced by the matching network components and the port connections, which were not accounted for in the initial, purely electromagnetic simulations.

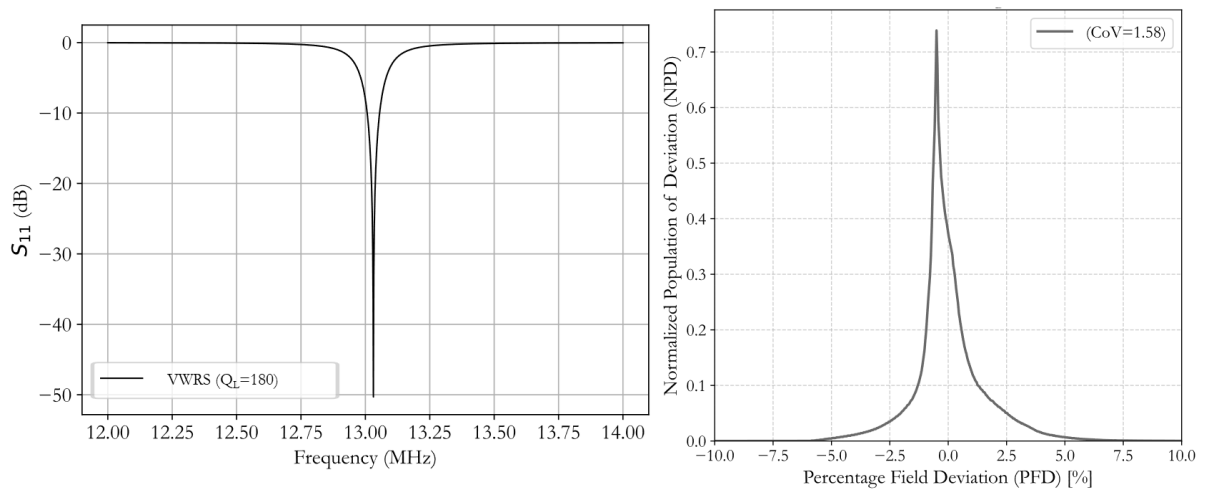


Figure 58: Co-simulation results for the optimized VWRS. (Left) The S_{11} parameter shows a well-matched resonance at 13.03 MHz, yielding a loaded Q -factor of 180. (Right) The final magnetic field distribution histogram, resulting in a CoV of 1.58%.

The plot on the right in Figure 58 displays the final magnetic field distribution histogram for the optimized and matched coil. The distribution is sharp and narrow, with a final Coefficient of Variation (CoV) of 1.58%. This demonstrates a high degree of field homogeneity, consistent with the design goals and the trade-offs identified during the optimization of the inter-turn gap. While the CoV is slightly higher than the absolute minimum achieved with a 0.2 mm gap (1.47%), it represents a significant improvement over less-optimized configurations and confirms the effectiveness of the chosen design parameters in balancing homogeneity with practical constraints like the Q-factor and manufacturing tolerances.

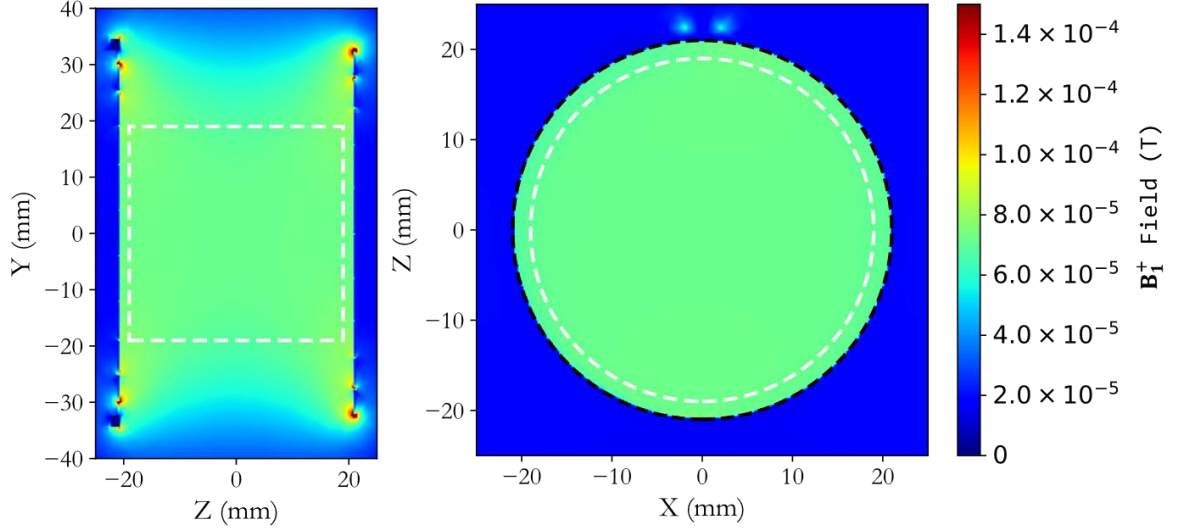


Figure 59: Magnetic field distribution maps for the final optimized VWRS coil. The axial view (left) and radial view (right) both demonstrate a high degree of field uniformity within the central volume of interest (indicated by dashed lines).

Figure 59 provides a visual representation of the magnetic field distribution within the coil. The axial cross-section (left) clearly shows the uniform field in the ROI, delineated by the dashed box representing the volume of interest. The radial cross-section (right) further confirms the excellent homogeneity across the sample area. The field is consistent and uniform throughout this region, validating the success of the VWRS design in achieving a highly homogeneous magnetic field for the target application.

IV.D. H&D Extended N-Ring Spherical Coil

This topology approximates the ideal spherical current distribution by using a small number of strategically placed conductor rings and optimized current ratios to systematically cancel low-order spherical harmonic field distortions. This method promises the highest degree of theoretical homogeneity.

1. Modeling Approach

1.1. Analytical Design

The generalized H&D algorithm presented in described in section III.B will be used to calculate the precise polar angles ($\theta_{Z,j}$) for each ring pair and the required current ratios $\{I_j\}$ needed to cancel the higher and lower order spherical harmonic terms. Positioning angles and current ratios are presented in Table 10.

N=4		N=6		N=8		N=10	
θ (in $^{\circ}$)	Current ratios	θ (in $^{\circ}$)	Current ratios	θ (in $^{\circ}$)	Current ratios	θ (in $^{\circ}$)	Current ratios
40.09	0.68	29.34	0.51	23.14	0.41	9.00	0.16
73.43	1.00	53.72	0.83	42.37	0.69	27.00	0.45
		77.92	1.00	61.45	0.89	45.00	0.71
				80.49	1.00	63.00	0.89
						81.00	1.00

Table 10: Positioning angles (θ) and current ratios for the H&D spherical coil designs.

1.2. 3D Geometry Construction

3D models of the four spherical coils were built in the EM solver based on the calculated ring positions. Each ring was modeled with a finite 4 mm width and 40 μm copper thickness including a small gap where a lumped element port was defined. Figure 60 shows the four selected coil geometries.

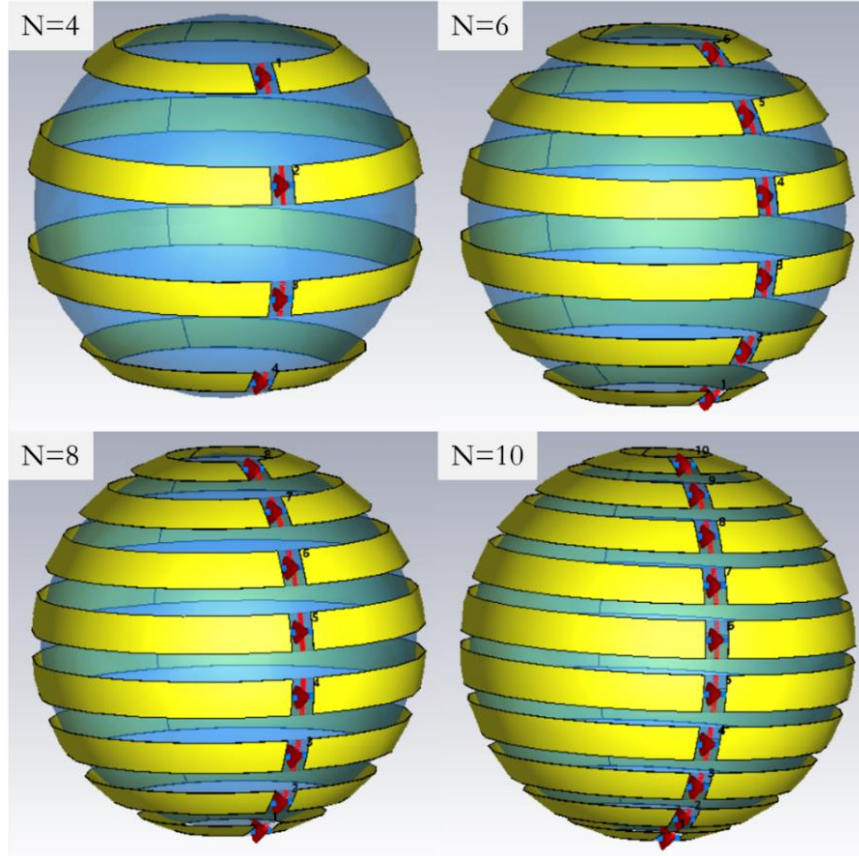


Figure 60: 3D models of the N=4, N=6, N=8, and N=10 H&D spherical coil geometries as constructed in the EM solver.

1.3. Inductance Matrix Extraction and Capacitance Calculation

For each coil, an initial simulation was performed to extract the multiport inductance matrix, $\mathcal{L}$, which quantifies the self-inductance of each ring and the mutual inductance between all ring pairs.

Using the extracted inductance matrix $\mathcal{L}$, the target current vector $\mathbf{I} = [I_1 \ I_2 \ \cdots \ I_N]^T$, and the desired resonant frequency ω_0 , the precise capacitance value C_k required for each ring is calculated using the relationship $C_k = \frac{I_k}{\omega_0^2[\mathcal{L}\mathbf{I}]_k}$, ensuring that the desired current distribution becomes a natural resonant mode of the coil.

2. Full-Wave Simulation and Co-Simulation Results

The calculated capacitor values were inserted into the 3D model's co-simulation setups as lumped elements modeled using the S-parameter files provided by the capacitor manufacturer. Figure 61 shows the Co-simulation schematic of the N=4 rings coil. (The detailed co-simulation schematics and lumped component values can be found in Annex IV)

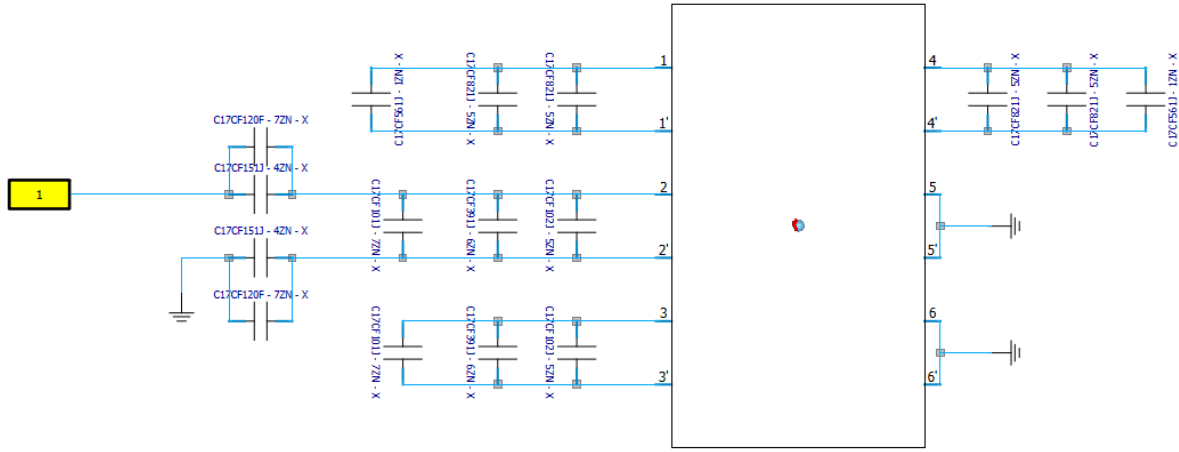


Figure 61: Co-simulation schematic example for the N=4 H&D spherical coil.

The simulated magnetic field distributions for the four H&D spherical coil designs are presented in Figure 62, which shows the magnetic field magnitude in the central transverse plane. A qualitative assessment of the plots reveals a significant improvement in field homogeneity as the number of rings (N) increases. The central region of uniform field (shown in green) becomes progressively larger and more consistent moving from the N=4 to the N=10 design which matches preliminary analytical estimates.

For a quantitative analysis, the coefficient of variation (CoV) of the magnetic field was calculated within the spherical ROI, defined by the white dashed line in Figure 62. These values, along with the loaded quality factor (Q_ℓ) for each resonant coil, are summarized in Table 11.

The quantitative results confirm the visual trend. The CoV decreases substantially with more rings, improving from 13.08% for the N=4 design to 7.08% for N=6, 4.72% for N=8, and reaching a low of 3.90% for the N=10 coil. This demonstrates a systematic enhancement of field homogeneity.

Notably, this significant improvement in homogeneity is achieved without a major impact on the coil's efficiency. The loaded quality factor (Q_ℓ) remains high and relatively stable across all four configurations, hovering between 132 and 139. This indicates that the N=10 coil provides the best

overall performance, offering the highest degree of field uniformity while maintaining an excellent quality factor.

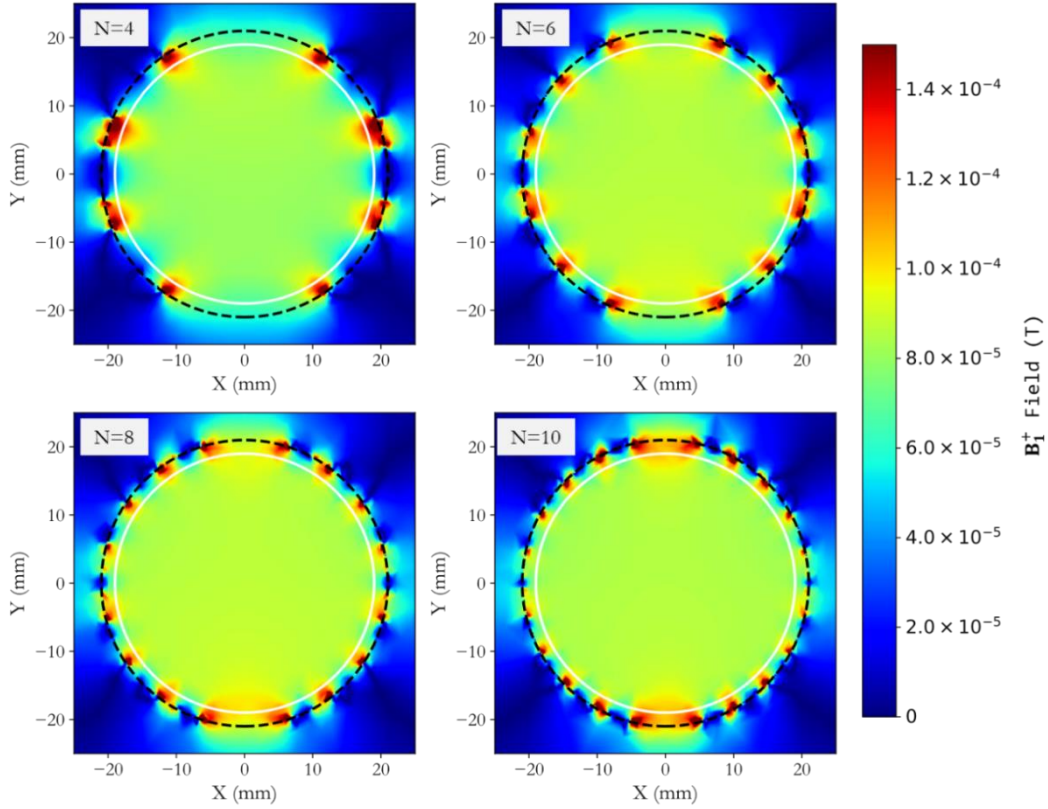


Figure 62: Simulated magnetic field magnitude maps in the central transverse plane for the N=4, N=6, N=8, and N=10 H&D spherical coil designs.

	H&D4	H&D6	H&D8	H&D10
CoV	13.08%	7.08%	4.72%	3.90%
Loaded Quality Factors (Q_L)	136	139	138	132

Table 11: Quantitative performance summary for the four H&D spherical coil designs.

3. Comparison with Analytical Model and Discussion

A critical comparison reveals a notable discrepancy between the homogeneity predicted by the preliminary analytical model and the results obtained from the full-wave co-simulation. The CoV values from the simulation (e.g., 13.08% for N=4, 3.90% for N=10) are significantly higher than those calculated analytically (4.55% for N=4, 0.20% for N=10).

This difference can be attributed to the idealizations inherent in the analytical Biot-Savart model versus the realistic conditions of the full-wave simulation. The analytical model assumes infinitesimally thin conductors and does not account for the physical geometry of the coil. In contrast, the full-wave simulation incorporates several key real-world factors:

- **Finite Conductor Dimensions:** The model includes a finite ring width of 4 mm, which alters the current distribution from a perfect line to a surface, impacting the resulting magnetic field profile.

- **Physical Gaps:** The inclusion of gaps for the lumped element ports introduces discontinuities that create field distortions not present in the idealized analytical model.
- **Realistic Components:** The use of manufacturer S-parameter files for the capacitors in the co-simulation accounts for their parasitic inductance and resistance, which can slightly alter the resonant current ratios and affect field homogeneity. This is further exacerbated by the need to stack up to 3 standard value capacitors to get close to ideal values. However, a non-negligible difference still remains between the capacitor stack value and the ideal value for maintaining proper current ratios. (e.g a stack of $680 + 680 + 1000 = 2360$ pF to get close to an ideal 2352 pF with an 8 pF difference)

Therefore, while the analytical model is invaluable for establishing the theoretical superiority of the H&D Extended strategy, the full-wave simulation provides a more accurate and realistic prediction of the performance achievable in a practical, physically implemented coil. The simulation results represent a trade-off between theoretical perfection and real-world feasibility.

Furthermore, this discrepancy underscores the manufacturing limitations that keep the ideal theoretical model out of practical grasp. The various factors that degrade performance in the simulation, finite conductor widths, necessary gaps for component placement, and the non-ideal behavior of capacitors, are all unavoidable consequences of physical fabrication. Achieving the near-perfect homogeneity of the analytical model would require manufacturing capabilities that do not exist, such as zero-width conductors and flawless components.

V. Comparative Performance of The Coils Under Investigation

This section provides a comprehensive comparison of the final coil topologies based on the key performance metrics derived from the full-wave co-simulations. The evaluation focuses on field homogeneity, sensitivity, and operational noise domains to identify the optimal design for the target benchtop MRI based tissue engineering application. Because each coil family is intended for a different sample geometry in practice, H&D spherical coils for spherical and VWRS/BLGR for cylindrical phantoms/tissue constructs the simulations were run with matching phantoms (H&D: $\varnothing$ 38 mm sphere with $V_s = 28\,731\text{ mm}^3$; VWRS/BLGR: $\varnothing$ 38 mm $\times$ 38mm cylinder with $V_s = 43\,096.37\text{ mm}^3$). Since sample geometry affects loading (Q-factor, B_1^+ efficiency, and CoV), direct numerical comparison across these two shapes would be misleading. Therefore, all results are reported in two tables, each evaluated in the geometry it was designed for, and all ranking and discussion are done within each shape. Table 12 summarizes the metric results for the cylindrical configurations and Table 13 the results for spherical configurations.

Coil Topology	VWRS	BLGR_13C	BLGR_20C
CoV	1.58%	1.58%	1.63%
Filling Factor	0.49	0.48	0.48
Unloaded Quality Factors (Q_u)	468	429	454
Loaded Quality Factors (Q_l)	180	215	374
Noise Dominance (Q_u/Q_l)	2.60	2.00	1.21
$E_{TR} = \langle B_1^+ \rangle_{ROI} (\Gamma/\sqrt{W})$	7.22×10^{-5}	7.78×10^{-5}	7.85×10^{-5}
$S_{RF} \propto E_{TR} \sqrt{V_s} (1 + \text{CoV}^2)^{1/2}$	4.74×10^{-7}	5.11×10^{-7}	5.16×10^{-7}

Table 12: Comprehensive performance metrics of investigated cylindrical coil topologies

Coil Topology	H&D4	H&D6	H&D8	H&D10
CoV	13.08%	7.08%	4.72%	3.90%
Filling Factor	0.66	0.69	0.7	0.71
Unloaded Quality Factors (Q_u)	224	246	246	228
Loaded Quality Factors (Q_l)	136	139	138	132
Noise Dominance (Q_u/Q_l)	1.65	1.77	1.78	1.73
$E_{TR} = \langle B_1^+ \rangle_{ROI} (\Gamma/\sqrt{W})$	8.20×10^{-5}	8.62×10^{-5}	8.74×10^{-5}	8.56×10^{-5}
$S_{RF} \propto E_{TR} \sqrt{V_s} (1 + CoV^2)^{1/2}$	4.44×10^{-7}	4.64×10^{-7}	4.69×10^{-7}	4.60×10^{-7}

Table 13 : Comprehensive performance metrics of investigated spherical coil topologies

V.A. Homogeneity

Field homogeneity is critical for quantitative MRI, and here, the Coefficient of Variation (CoV) serves as the primary metric for this comparison. Within the Cylindrical set The VWRS and the BLGR_13C coils emerged as the top performers, achieving an identical CoV of 1.58%. The BLGR_20C was negligibly behind at 1.63%. These cylindrical designs significantly outperformed the spherical H&D coils in the realistic simulation environment.

While the H&D topology promised the highest theoretical homogeneity, its performance was degraded by practical implementation factors such as finite conductor widths and the inclusion of physical gaps for capacitors. The H&D10 coil, the best in its class, only achieved a CoV of 3.90%, which is more than double that of the leading cylindrical designs and close to the proposed $\pm 5\%$ threshold. This discrepancy highlights the challenge of translating the theoretical elegance of the extended H&D approach into a physically realizable and robust coil without performance loss. As both the H&D8 and H&D10 coils achieved a CoV within the sample region below the 5% benchmark proposed at the beginning of the chapter, they might yet be an acceptable design for spherically shaped samples. To properly rule on their overall suitability for small sample quantitative imaging, a deeper study of the impact of homogeneity specific to our use case must be carried out.

V.B. Noise Dominance

As presented in Chapter I, the noise dominance, quantified by the ratio of the unloaded to loaded quality factors Q_u/Q_l , is a crucial metric for evaluating the efficiency of an RF coil. This ratio indicates whether the dominant source of noise in a measurement is the sample itself or the intrinsic losses of the coil hardware. In the ideal "sample-noise-dominated" regime, noise from the sample significantly outweighs the thermal noise from the coil, and the coil's noise figure is close to 0 dB. This is desirable because the MRI signal also originates from the sample; maximizing the signal-to-noise ratio (SNR) is most effective when the coil's own noise contribution is negligible. A Q_u/Q_l ratio significantly greater than 1, and conventionally targeted to be above 2, confirms that the coil is operating efficiently and that the SNR is limited by the sample properties rather than hardware imperfections.

Based on this metric, the VWRS coil exhibits the best performance with a noise dominance ratio of 2.60, clearly establishing it as a sample-noise dominated resonator. The BLGR_13C also performs well, meeting the threshold with a ratio of 2.00. In contrast, all the H&D spherical designs fall below this mark (1.65–1.78), indicating that their performance is partially limited by coil noise. Most notably, the BLGR_20C, despite having the highest loaded quality factor $Q_\ell = 374$, shows the poorest noise performance with a ratio of just 1.21. This implies that it is strongly coil-noise dominated.

V.C. Sensitivity

The classical resonator-sensitivity theory proposed in Chapter I Section III.C.3.3 describes the single-coil transmit/receive (TR) S_{RF} as

$$S_{RF}^{(\text{ideal})} \propto \sqrt{\frac{\eta \mu_0 \omega_0 Q_\ell}{T_{\text{eff}} V_s}}, \quad (\text{V-1})$$

where the filling factor η quantifies the magnetic-energy overlap between the coil field and the sample volume V_s , and Q_ℓ represents the ratio of stored to dissipated energy in the loaded resonator. This compact expression assumes a uniform B_1 field throughout the sample and corresponds to an energy-based analytical model, not to a specific excitation normalization.

In the present work, each resonator was simulated under a power-normalized excitation condition: the feed port was driven with a fixed incident power of 1 W, and the solver reported the corresponding accepted power P_{acc} after reflection losses. Under this normalization, the directly comparable quantity is the transmit efficiency

$$E_{TR} = \frac{\langle |B_1^+| \rangle_{\text{ROI}}}{\sqrt{P_{\text{acc}}}}, \quad (\text{V-2})$$

where $\langle |B_1^+| \rangle_{\text{ROI}}$ denotes the ROI-averaged transmit-field magnitude. This metric expresses the coil's RF-field generation efficiency (B_1^+ per $\sqrt{W}$) and allows direct comparison of geometries under identical loading and frequency conditions.

As proposed in Chapter I, the true NMR sensitivity S_{RF} arises from electromagnetic reciprocity. In receive mode, the voltage induced at the coil terminals by the precessing magnetization is proportional to the spatial integral of the local transverse magnetization $M_\perp(\mathbf{r})$ multiplied by the transmit field per unit current, $B_1^+(\mathbf{r})/I_{\text{rms}}$. Since the instantaneous signal voltage is proportional to this integral, the detected signal power (and therefore the measurable SNR) scales with its square. In a power-normalized formulation, where B_1^+ is expressed per $\sqrt{P_{\text{acc}}}$ instead of per ampere, this relationship introduces a square root over the integrated quadratic term, restoring the correct voltage-proportional scaling. Accordingly, for a homogeneous sample of identical composition and constant noise temperature T_{eff} , the relative sensitivity can be expressed as

$$S_{RF} \propto \sqrt{\int_{V_{\text{sample}}} \frac{|B_1^+(\mathbf{r})|^2}{P_{\text{acc}}} dV}, \quad (\text{V-3})$$

which measures the total coupling under the same power normalization. No additional $\propto \sqrt{R_{\text{eff}}}$ factor is introduced, because coil and sample losses are already embodied in the B_1^+ amplitude at constant accepted power. Including it again would double-penalize high-inductance or heavily loaded geometries.

Eq. (V-3) can be simplified to

$$S_{\text{RF}} \propto E_{\text{TR}} \sqrt{V_s} (1 + \text{CoV}^2)^{1/2}. \quad (\text{V-4})$$

The correction factor $(1 + \text{CoV}^2)^{1/2}$ originates from the difference between the mean field amplitude and the root-mean-square (RMS) field that determines the quadratic signal coupling. Let $B(\mathbf{r}) = |B_1^+(\mathbf{r})|$ and $\mu = \langle B \rangle_{\text{ROI}}$, then

$$B_{\text{rms}} = \sqrt{\langle B^2 \rangle_{\text{ROI}}} = \mu \sqrt{1 + \text{CoV}^2}, \quad (\text{V-5})$$

where $\text{CoV} = \sigma/\mu$ is the coefficient of variation of the field amplitude. Replacing the spatial integral of $|B_1^+|^2$ by $B_{\text{rms}}^2 V_s$ introduces exactly this multiplicative correction. For $\text{CoV} = 5\%$, $(1 + \text{CoV}^2)^{1/2} = 1.0025$, corresponding to only a 0.25 % deviation; the field may therefore be considered uniform to an excellent approximation.

This formulation preserves the expected $\sqrt{V_s}$ scaling of NMR sensitivity, is consistent with reciprocity, and remains fully compatible with the power-normalized simulation framework adopted in this work.

Following this more general S_{RF} definition the coil rankings by sensitivity, from the coil boasting the highest sensitivity is:

- For the cylindrical configurations: H&D8, H&D6, H&D10, H&D4.
- For the spherical configurations: BLGR-20C, BLGR-13C, VWRS.

Sensitivity is arguably the most important metric for low-field MRI. In this category, the BLGR_20C is the most sensitive among the cylindrical coils, with sensitivity metric $S_{\text{RF}} \propto 5.16 \times 10^{-7}$ which is however only approximately 1% higher than its closest competitor, the BLGR_13C with a $S_{\text{RF}} \propto 5.11 \times 10^{-7}$. Driven by its exceptionally high loaded quality factor ($Q_\ell = 374$ approximately 74% higher than the BLGR_13C with a $Q_\ell = 215$) the BLGR_20C would be expected to have a better overall sensitivity. This high Q_ℓ is a direct result of using 20 parallel capacitors, which effectively minimizes the total equivalent series resistance (ESR) and thus reduces resonator losses. However, using high number of capacitors seems to shift the noise dominance towards a coil noise dominated regime ($Q_u/Q_\ell = 1.21$) which would explain its lackluster performance compared to what one would expect solely looking at the loaded quality factor.

Among spherical coils on the other hand the highest sensitivity is obtained using the H&D8 topology with $S_{\text{RF}} \propto 4.69 \times 10^{-7}$, followed closely by the H&D6 topology ($S_{\text{RF}} \propto 4.64 \times 10^{-7}$) with a negligible sensitivity drop of around 1%.

VI. Conclusion and Discussion

This investigation conducted a systematic evaluation of three advanced RF coil topologies, the Varying-Width Ribbon Solenoid (VWRS), the Bridged Loop-Gap Resonator (BLGR), and the extended Hoult & Deslauriers (H&D) spherical coil, to identify the optimal design for small-sample quantitative imaging on a 0.3T benchtop MRI scanner. The results reveal that the choice

of the optimal coil involves a nuanced trade-off between maximizing raw sensitivity and achieving a highly homogenous field profile.

For cylindrical configurations, the analysis identifies two exceptional candidates from the BLGR family, the BLGR_20C and the BLGR_13C. Both configurations offer similar levels of field homogeneity and sensitivity. BLGR_13C emerges as a highly viable and robust alternative that offers a more balanced performance profile. It achieves the highest B1 field homogeneity of all tested designs, with a CoV of 1.58%. Crucially, it operates in the desirable sample-noise dominated regime ($Q_u/Q_\ell=2$) when highly loaded, ensuring that measurement quality is limited by the sample rather than hardware imperfections. While its sensitivity is slightly lower than the BLGR_20C, it is still the second highest in its set, with the VWRS boasting the weakest sensitivity. Furthermore, its reduced complexity (13 vs. 20 capacitors) simplifies construction, tuning, and replication. For applications requiring a large number of fabricated coils the 13-capacitor variation is the more economical option with non-magnetic MLCC's costing upwards of 1.5 to 2 euros per unit compared to a few cents per unit for standard capacitors.

The H&D spherical coils, despite their theoretical elegance, proved less effective in a realistic simulated environment. The practical necessities of finite conductor size and component integration significantly degraded their homogeneity. They cannot however be completely excluded as a practical solution as the higher ring count topologies still exhibited field homogeneity below the accepted $\pm 5\%$ threshold. Following the established field homogeneity limits in this chapter, defining coil suitability for quantitative imaging, H&D8 and H&D10 can be retained as strong viable candidates. However, since the proposed threshold was only determined from broadly accepted conventions in the literature [104] [111] rather than from a rigorous error-propagation analysis linking field non-uniformity to quantitative bias in tissue engineering applications, it remains somewhat arbitrary. Future work should therefore aim to refine these criteria by correlating spatial variations in B_1^+ with the accuracy and precision of derived quantitative parameters, which would enable a more application-driven selection of acceptable homogeneity levels.

Ultimately, the optimal design choice is application dependent. For experiments requiring spherical coil designs, matching spherical shaped samples, although barely reaching the acceptable homogeneity threshold for quantitative analysis, H&D8 can be proposed as a highly sensitive option. For cylindrical sample, the 13-capacitor variation of the BLGR coil will be preferred as the perceived advantage of adding more capacitors is mitigated by a penalizing shift in the noise dominance domain.

Chapter III: 3D Plastronics Culture Chamber for Tissue Engineering Applications.

Chapter Overview

This chapter presents the development, fabrication, and validation of a fully integrated 3D Plastronics-based culture chamber tailored for tissue engineering applications within a low-field benchtop MRI system. The work addresses the limitations of conventional culture chambers and commercial RF coils by proposing a multifunctional platform that combines active perfusion, integrated mechanical assemblies, and high MR sensitivity using a form-fitted RF coil in a single compact device.

The chapter begins by outlining the design constraints imposed by the scale and operating conditions of benchtop MRI, along with the biological requirements for sustaining viable and functional 3D cell constructs. It then details the conception of a custom perfused chamber, including its mechanical architecture, flow distribution system, and fluidic performance evaluation through both computational fluid dynamics (CFD) and MRI-based velocimetry.

Then, the integration of a bridged loop-gap resonator (BLGR) RF coil directly onto the chamber's surface using a Rapid 3D Plastronics (R3DP) approach is described, alongside a local shielding strategy coupled with positioning mechanisms. Electromagnetic simulations guide the coil geometry and matching network design, with results compared to experimental data.

Finally, the system's imaging performance is demonstrated through high-resolution, in situ MRI of a bio-printed hydrogel scaffold under perfusion, enabling non-destructive morphological assessment and a quantitative comparison between the 3D Plastronics setup and a manufacturer-provided shielded coil setup.

I. Introduction

Despite the numerous advantages offered by 3D tissue engineering, the complexity of these constructs presents new handling challenges linked to their size and controlled perfusion requirements. Although multifunctional perfusion-based culture chambers incorporating active nutrient flow, waste removal, and mechanical support have been developed to support bioengineered tissue growth [12], these systems are bulky, and poorly suited for integration within a benchtop MRI setup.

To take full advantage of Benchtop MRI as an efficient modality for 3D living samples characterization, it is necessary to design a setup combining a 3D perfused chamber with a high sensitivity form-fitted shielded coil in a single device to maximize the inherently limited SNR. Moreover, the setup must fit into the constrained available imaging volume of a benchtop MRI system. The current fabrication processes relying on conventional electronics are lacking and are unable to seamlessly integrate mechanical and electrical components. In fact, the use of conventional electronic fabrication technologies, predominantly based on rigid 2D printed circuit boards, are not suited to design complex 3D geometries required for optimizing sensitivity and ensuring homogeneous radiofrequency magnetic fields. Currently, MRI setups largely rely on the use of commercially available RF coils, which leads to suboptimal assemblies. The fixation of samples and the integration of necessary peripherals within these RF coils frequently involve unreliable methods, such as adhesive tapes and makeshift supports, which lack reproducibility and precision.

The 3D Plastronics approach, a technique that enables the full integration of electronic circuits and components on 3D surfaces, offers a promising solution. 3D Plastronics manufacturing methods rely on the integration of circuitry onto 3D polymer surfaces, with conductive tracks serving as interconnects or functional elements like antennas [114] or inductors [159]. They also enable active features such as pH sensors and microfluidics systems [160], [161], with material choices tailored to the application (biosourced/biodegradable [162], temperature-resistant [129], biocompatible, etc.).

3D Plastronics methods have already been shown to enable the seamless embedding of RF coils in complex MRI setups [120], [130], [163]. Using similar methods, integrating well dimensioned coils directly onto the surface of the culture chamber, would allow for the creation of a cohesive system that combines imaging and tissue culture functionalities. This approach ensures optimal coil positioning relative to the sample, maximizing sensitivity and field homogeneity while maintaining the culture chamber's ability to provide a controlled, *vivo*-like environment. Furthermore, the flexibility of 3D Plastronics fabrication enables the design of customized culture chamber geometries and coil configurations, accommodating a wide range of sample sizes.

This chapter will aim to present a scalable and cost-effective platform for characterizing 3D tissue-engineered constructs through an approach which not only enhances imaging capabilities but also provides new opportunities to explore the mechanisms, dynamics, and interactions underlying functional tissue development in bio-engineered systems by combining the adaptability of 3D Plastronics fabrication with a cost effective, accessible and portable low-field MRI scanner for an in-line integration in the tissue fabrication process.

To provide a performance-based comparison to another more “classical” approach, a manufacturer provided optimized volume coil will be used as a benchmark, facilitating a quantitative performance evaluation (signal-to-noise ratio (SNR), field homogeneity, assembly practicality etc...). Specifically, the manufacturer-provided optimized volume coil will serve as the reference standard against which the custom 3D Plastronics fabricated coil setup will be assessed under identical scanning protocols. Both setups will employ the same coil topology to ensure a

fair, performance-focused comparison. As shown in Chapter II, the BLGR topology demonstrated the greatest stability under varying load conditions while maintaining high homogeneity and sensitivity, and will therefore be implemented in both the manufacturer's benchmark coil and our custom 3D Plastronics fabricated setup.

II. Manufacturer Provided Coil and Sample Fixation Mechanism

II.A. Manufacturer Provided Shielded Coil

A shielded bridged loop-gap resonator (BLGR) was built and provided as an all-purpose coil by the MRI manufacturer. According to the manufacturer, the coil is centered within an inner chamber ($90 \times 51 \times 170$ mm) housed inside a larger enclosure ($165 \times 51 \times 170$ mm). The coil has a diameter of 42 mm and a length of 65 mm, wrapped around an alumina tube with an outer diameter of 42 mm, an inner diameter of 38 mm, and a length of 170 mm, forming the usable bore. The BLGR followed a classical center-fed design with a deposed tuning and matching circuit integrated onto a Printed Circuit Board (PCB) positioned inside the outer chamber of the enclosure. The shield was grounded through both the SMA feedthrough, and the support guard rails, ensuring a low impedance path to ground.

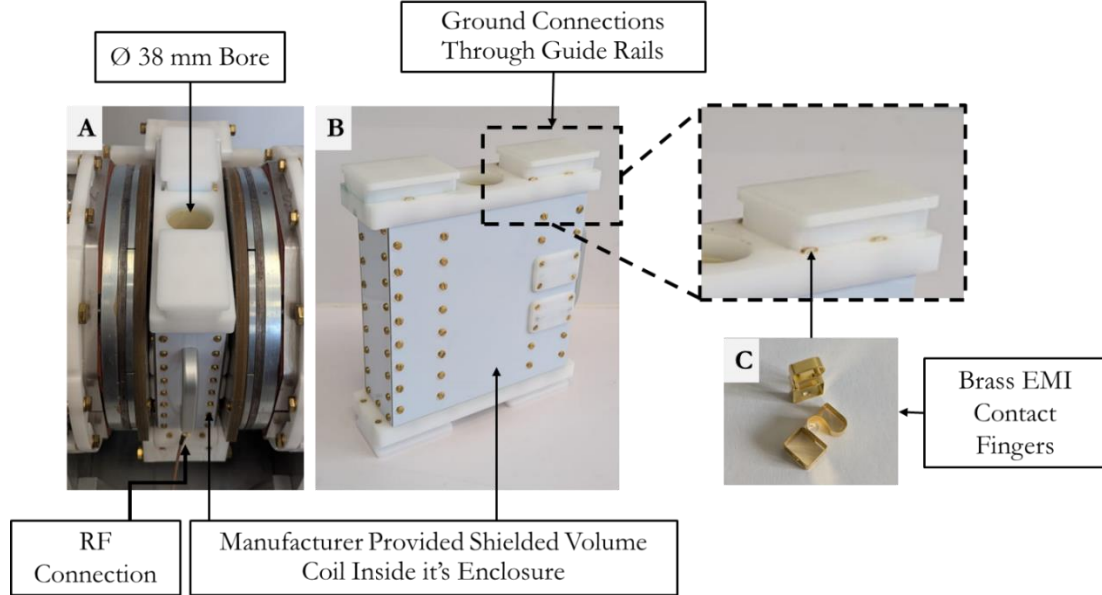


Figure 63: (A) Shielded coil housing positioned inside the magnet gap. (B) Close-up of the shielded coil housing designed for both vertical and horizontal orientation within the magnet bore. (C) Brass EMI contact fingers used for sliding connection to guide rail ground.

II.B. Sample Fixation Mechanism

To achieve reproducible sample positioning within the manufacturer-provided coil, a custom two-rail guide assembly was developed (Figure 64). The PLA 3D-printed support structure consists of two parallel rails on a fixed base plate, which features graduations to indicate insertion depth into the shielded coil bore. A matching sliding block, also 3D-printed in PLA and bored to snugly accept the cylindrical sample tube, glides along these rails. Vertical locking is accomplished via two nylon set screws on the sides of the base, which, when tightened, clamp the sliding block in place and prevent any unwanted translation. This simple sliding mechanism allows smooth lowering of

the sample into the coil bore and can be mounted in either a vertical or horizontal orientation to accommodate different magnet geometries.

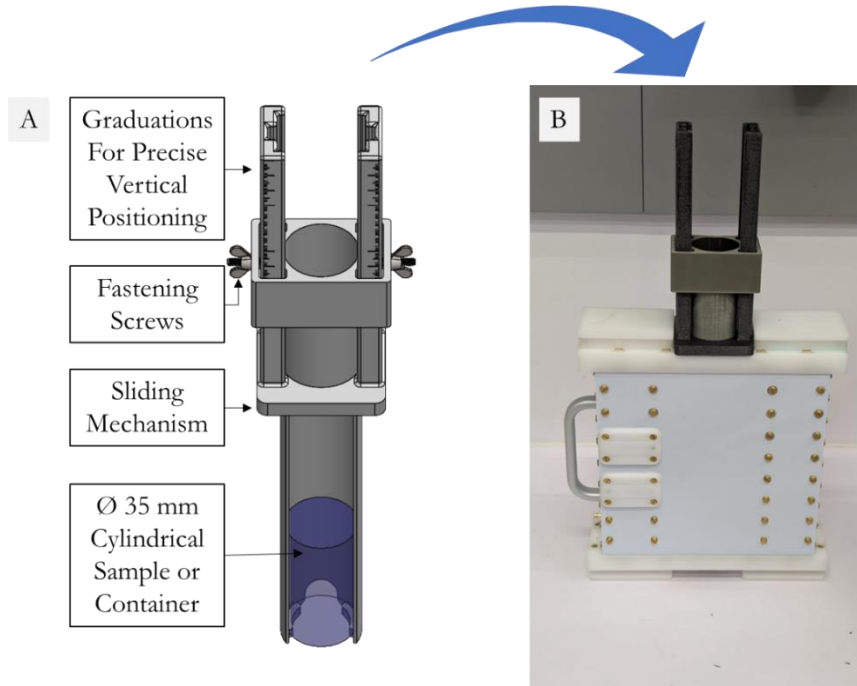


Figure 64: 3D-printed dual-rail sample insertion guide. (A) CAD model showing the graduated PLA base and sliding block designed for precise sample positioning inside the Ø38 mm coil bore. (B) Photograph of the assembled guide atop the shielded coil housing.

III. 3D-Bioprinted Scaffold Design

To produce complex 3D bio-printed constructs, scaffold structures are used as support during the tissue growth process. Defining the scaffold geometry up front guides any subsequent culture-chamber design. For this study, we selected a cylindrical shaped scaffold (Ø 18.2 × 25 mm height) with two opposing flat faces to prevent rotation inside the chamber. The scaffold contains a 4 × 4 grid of identical square pores (2.5 × 2.5 mm) to allow uniform nutrient flow and cell seeding throughout the entire structure (see Figure 4).

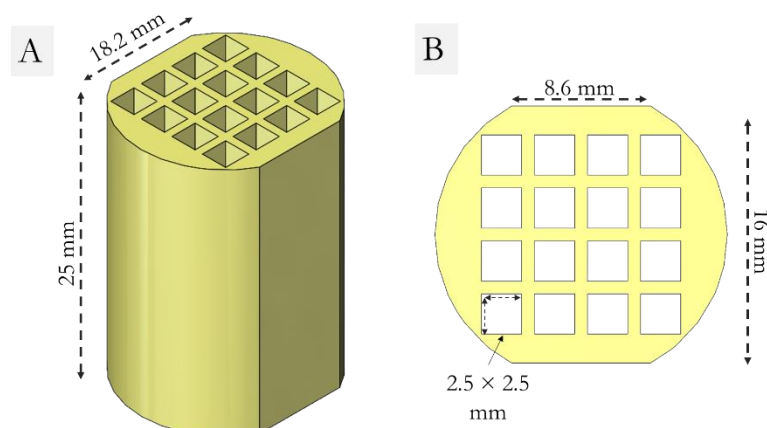


Figure 65: (A) Isometric CAD rendering of the scaffold showing the cylindrical envelope with two flat sides to prevent rotation inside the culture chamber. (B) Top view illustrating the 4 × 4 array of 2.5 mm square pores.

IV. 3D Plastronics Perfused Chamber Design and Fabrication

IV.A. Custom Made Culture Chamber and Assembly

A custom-designed $\varnothing 35 \text{ mm} \times 185 \text{ mm}$ culture chamber (Figure 67) was engineered to accommodate the 3D bio-printed scaffold presented in Figure 65. The design featured a primary cylindrical chamber measuring $\varnothing 20 \text{ mm} \times 80 \text{ mm}$, serving as the central housing for the bio-printed scaffold. The chamber was designed to be supplied with liquid through two separate 3 mm cylindrical channels: a return path running from the top of the chamber and parallel to it, leading to an outlet, and an inlet positioned at the center of a piston mechanism. This piston, also referred to as “sprinkler” due to its design, was configured to ensure uniform flow distribution across all 16 channels by aligning them with 16 precisely positioned 1.5 mm outlets. This removable part also doubles as a printing surface during the bio-printing process as it encapsulates and holds the gel-based scaffold (Figure 66).

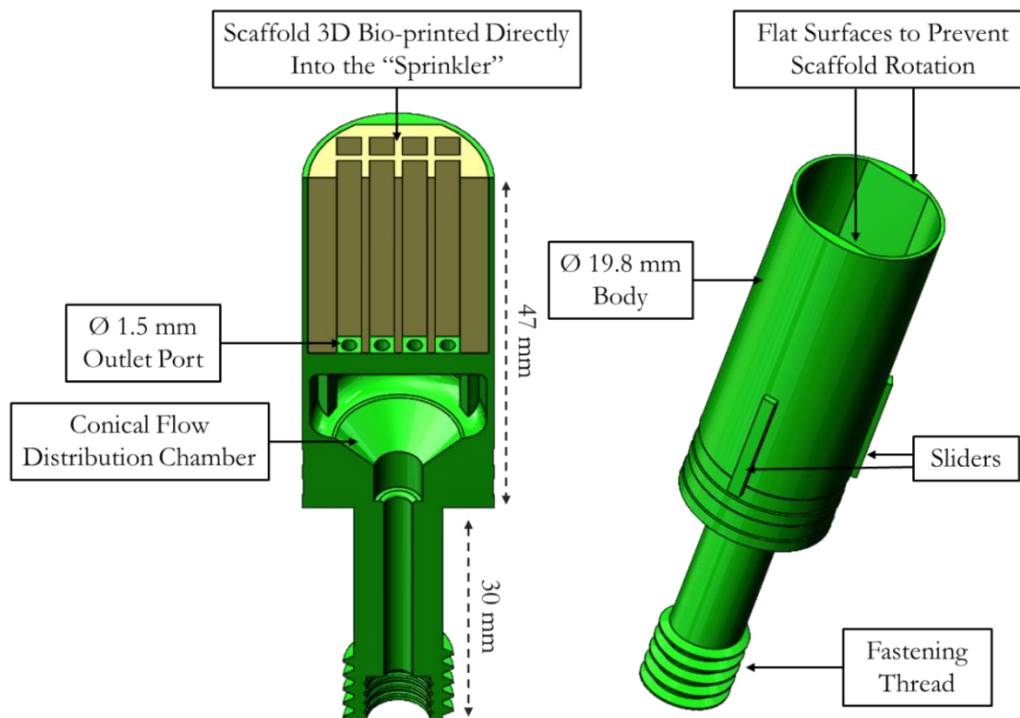


Figure 66: “Sprinkler” for uniform media flow distribution. (A) Longitudinal sectional view showing the central inlet, conical flow distribution chamber, $\varnothing 1.5 \text{ mm}$ outlet ports aligned with the Scaffold pores. (B) Isometric rendering, highlighting the external fastening thread, flat inner surfaces to prevent scaffold rotation inside the sprinkler and sliders to prevent sprinkler rotation inside the culture chamber.

Three 1 mm grooves were etched on the inside of the culture chamber to secure and prevent the rotation of a piston mechanism doubling as a 3D-printing build surface to allow for the sterile insertion of bio-printed samples once the printing process was complete. The system filled and was sealed at both ends using, at the top, a threaded 3D-printed cap and a Nitrile O-ring (RS PRO ID 26, OD 31, E 2.50, part number 255-6600), and at the bottom, a threaded 3D-printed slotted fastener applying vertical pressure to squeeze a second, smaller Nitrile O-ring (RS PRO ID 15.15, OD 20.39, E 2.62, part number BS114) between the bottom of the chamber and the piston mechanism, making it watertight.

All parts aside from the O-rings were designed using SOLIDWORKS 2023 (Dassault Systèmes, Vélizy-Villacoublay, France) to be printed using either Fused Deposition Modelling (FDM) or Stereolithography (SLA) 3D printers.

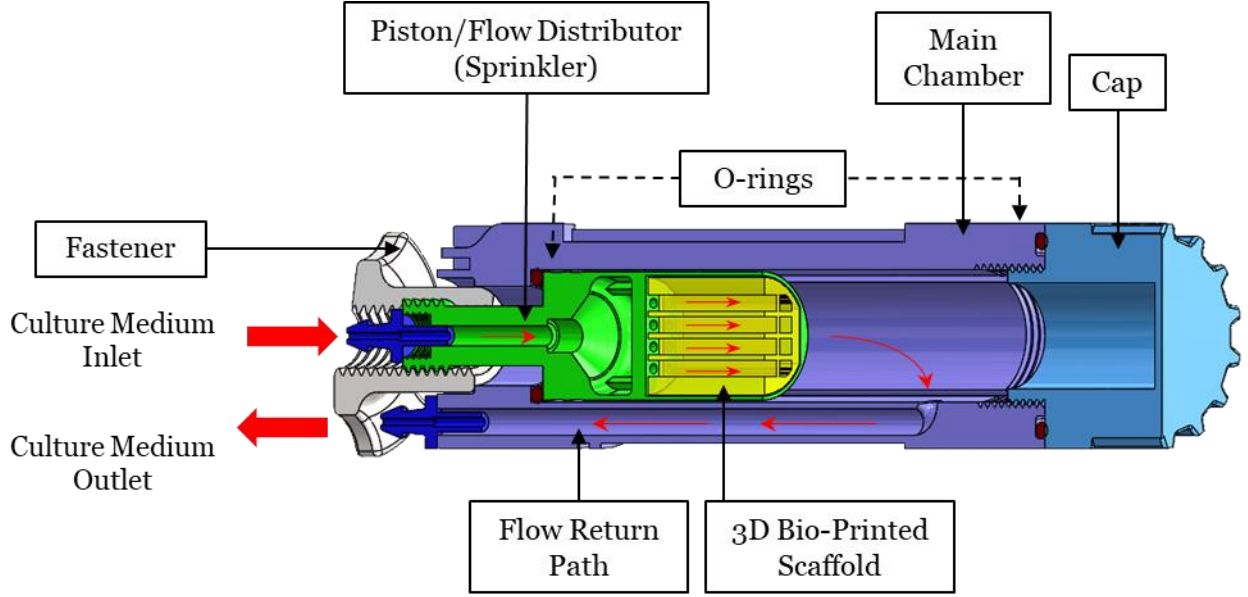


Figure 67 : Longitudinal sectional CAD rendering illustrating the main components of the Ø 35 × 185 mm chamber: fastening mechanism, flow distributor (“sprinkler”), 3D bio-printed scaffold, main chamber body, and cap. Culture medium enters through an inlet at the base of the flow distributor, which ensures uniform distribution across the scaffold, and exits through an outlet connected to the return channel (flow path indicated with red arrows).

IV.B. Fluidic System Validation

The design of a fluidic system must address several critical constraints to create a viable microenvironment for tissue development. First, It's important to ensure that the flow provides consistent nutrients and oxygen everywhere [164], [165], [166] . Additionally, a uniform flow through the scaffold helps avoiding dead zones so cells can be evenly seeded, and make sure that nutrients and waste are exchanged efficiently [166], [167]. Second, the system must carefully manage shear stress to support healthy cell growth. The quantitative link between the velocity field $u(r)$ and the resulting shear stress $\tau(r)$ is given by [168]:

$$\tau(r) = \mu \nabla u(r) \quad (IV-1)$$

where μ is the fluid’s dynamic viscosity. By controlling the velocity field through adjustments to the flow geometry and pump settings it’s possible to impose specific and beneficial shear profiles throughout the culture chamber. Addressing these factors enables the creation of an optimal microenvironment for successful tissue development.

Due to the geometric complexity of the proposed design, detailed understanding of the local velocity profiles requires computer-assisted fluid dynamics (CFD) simulations, validated with measured velocimetry maps [167].

1. Materials and Methods

1.1. Computer-Assisted Fluidic Dynamics Simulations (CFD)

1.1.1 Computational Framework

The simulation of liquid flow throughout the culture chamber, including the scaffold and associated flow pathways, was performed using COMSOL Multiphysics (version 6.2, COMSOL AB, Sweden). This computational platform employs finite element analysis (FEA) to solve the governing equations of fluid dynamics, providing a comprehensive framework for modeling flow behavior within the culture chamber.

1.1.2 Geometry and Model Setup

The culture chamber geometry, including the scaffold and inlet/outlet manifolds, was imported into COMSOL as a CAD file. The scaffold consisted of interconnected channels with a 2.5 mm square cross-section. The culture chamber was modeled in a vertical orientation to replicate experimental conditions, ensuring accurate simulation of the flow environment.

1.1.3 Governing Equations

The liquid flow within the culture chamber was modeled using the Navier-Stokes equations for incompressible fluids:

$$\rho \left(\frac{\partial \mathbf{u}}{\partial t} + \mathbf{u} \cdot \nabla \mathbf{u} \right) = -\nabla p + \mu \nabla^2 \mathbf{u} + \mathbf{F}_{\text{ext}}, \quad (\text{IV-2})$$

$$\nabla \cdot \mathbf{u} = 0, \quad (\text{IV-3})$$

where ρ is the fluid density, $\mathbf{u}$ is the velocity vector, p is pressure, μ is dynamic viscosity, and $\mathbf{F}_{\text{ext}}$ represents external forces. Gravity was included in the model to simulate its effects on the vertical culture chamber setup.

1.1.4 Flow Conditions

Two flow rates, 7 mL/min and 20 mL/min, were simulated to evaluate their effects on fluid dynamics within the culture chamber system. These flow rates (Q_{in}) were applied as boundary conditions at the inlet, with zero-gauge pressure set at the outlet ($P_{\text{out}} = P_{\text{atm}}$) (as shown in Figure 68-A). A no-slip condition was enforced on all culture chamber and scaffold walls to replicate realistic flow interactions.

1.1.5 Material Properties

Water was used to model the culture medium, with its density ($\rho = 997 \text{ kg/m}^3$) and dynamic viscosity ($\mu = 1 \text{ mPa}\cdot\text{s}$) defined based on standard physical properties.

1.1.6 Meshing and Solver Configuration

The entire culture chamber geometry, including the scaffold with 2.5 mm square channels, was discretized using a tetrahedral finite element mesh. “Extra fine” mesh resolution was applied over the entire simulation region to accurately capture velocity gradients (as shown in Figure 68-B). Adaptive mesh refinement was employed to balance computational efficiency and numerical accuracy.

Steady-state simulations were performed using segregated solvers to generate velocity fields across the culture chamber system for both simulated flow rates. The laminar flow assumption was validated during the simulations by confirming that the Reynolds number remained below the critical threshold for turbulence (see **Error! Reference source not found.**).

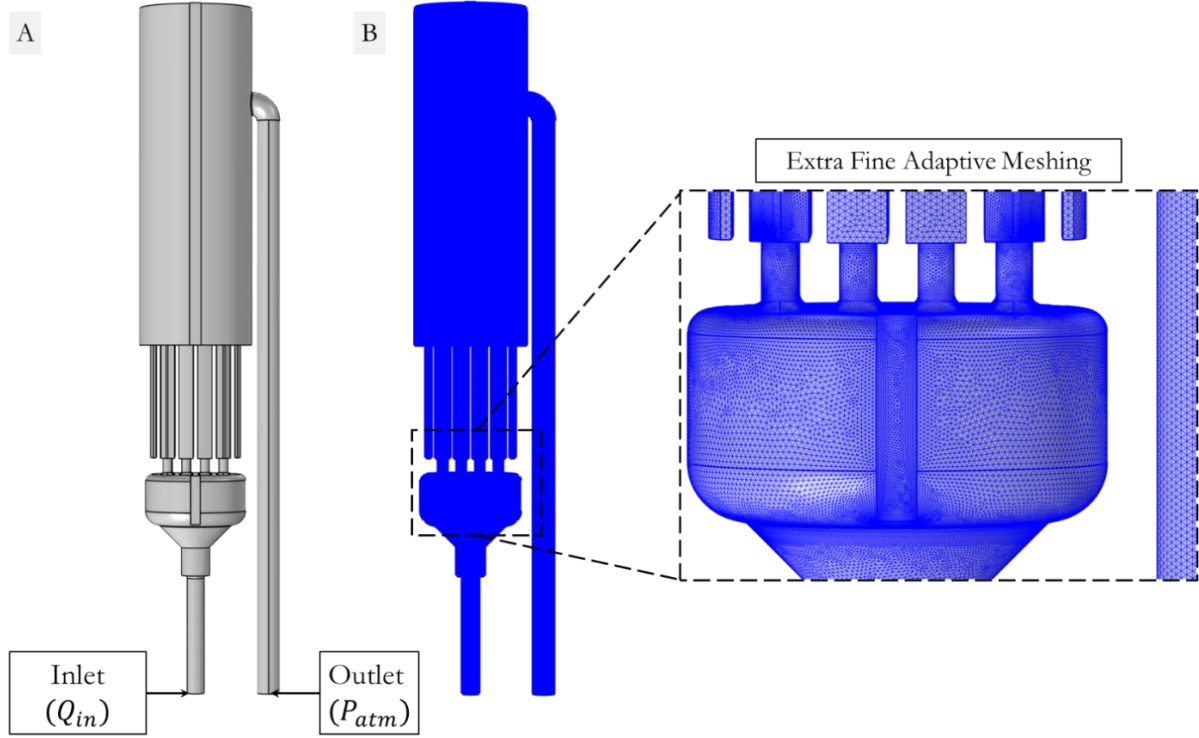


Figure 68: Geometry and mesh used for CFD simulations of the perfusion culture chamber. (A) CAD rendering of the vertical-orientation culture chamber showing the main vessel and the defined inlet/outlet. (B) Corresponding finite-element discretization employed in COMSOL Multiphysics, featuring an extra-fine tetrahedral mesh. Dashed box indicates the region magnified at the right, where the mesh density and element quality are easier to observe.

1.2. Experimental Validation using MRI Velocimetry

1.2.1 Scaffold Preparation

A makeshift 20 mm long scaffold (same design as described in Section III) was FDM 3D printed using Transparent PLA (Ultimaker) to fit inside the main chamber to simulate the scaffold. A PLA Scaffold was selected instead of a gel-based to ensure repeatability and to match theoretical values as closely as possible, avoiding variability from potential hydrogel swelling, malformations, or distortions during handling.

The scaffold was placed inside a sealed culture chamber filled with a 16.3 mM Copper (II) Nitrate Trihydrate solution in water serving as a contrast agent.

1.2.2 Flow Circuit Setup

In this setup, the culture chamber was inserted into the coil bore using a simple 3D FDM-printed Transparent PLA (Ultimaker) sliding mechanism (Figure 69) to ensure sample centering and positioning

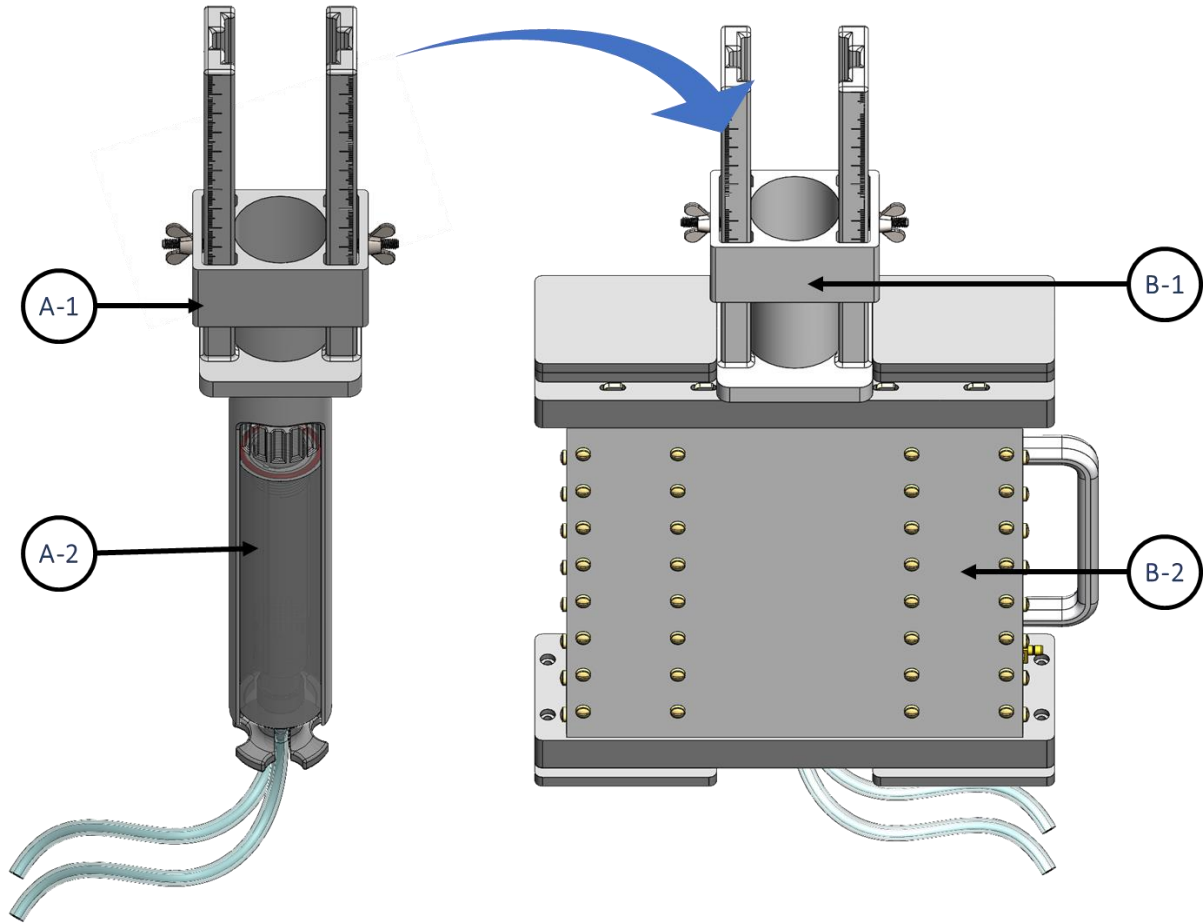


Figure 69: Schematic of the all-purpose coil provided by the MRI manufacturer and the 3D FDM-printed sliding mechanism used for positioning. (A-1) Perfused culture chamber inserted into the sliding mechanism. (A-2) 3D-printed sliding mechanism used for precise alignment within the coil bore. (B-1) Sliding mechanism fully inserted into the shielded enclosure. (B-2) Shielded enclosure housing the BLGR coil.

Perfusion of the makeshift scaffold within the culture chamber was performed using a Masterflex peristaltic pump (Ismatec ISM834C), having a speed range from 0.003 to 31 mL/min, connecting all the elements with autoclavable translucent tubing (Saint Gobain Versilic flexible tube), Ø 3mm x Ø 6mm. To mitigate the pulsatile nature of the peristaltic pump and simulate a continuous flow, a double reservoir system was implemented as a buffer between the pump and the culture chamber (see Figure 70). This setup acts as a low-pass filter, where the large surface area of the reservoirs dampens the pressure fluctuations from the peristaltic pump, delivering a more continuous flow to the culture chamber.

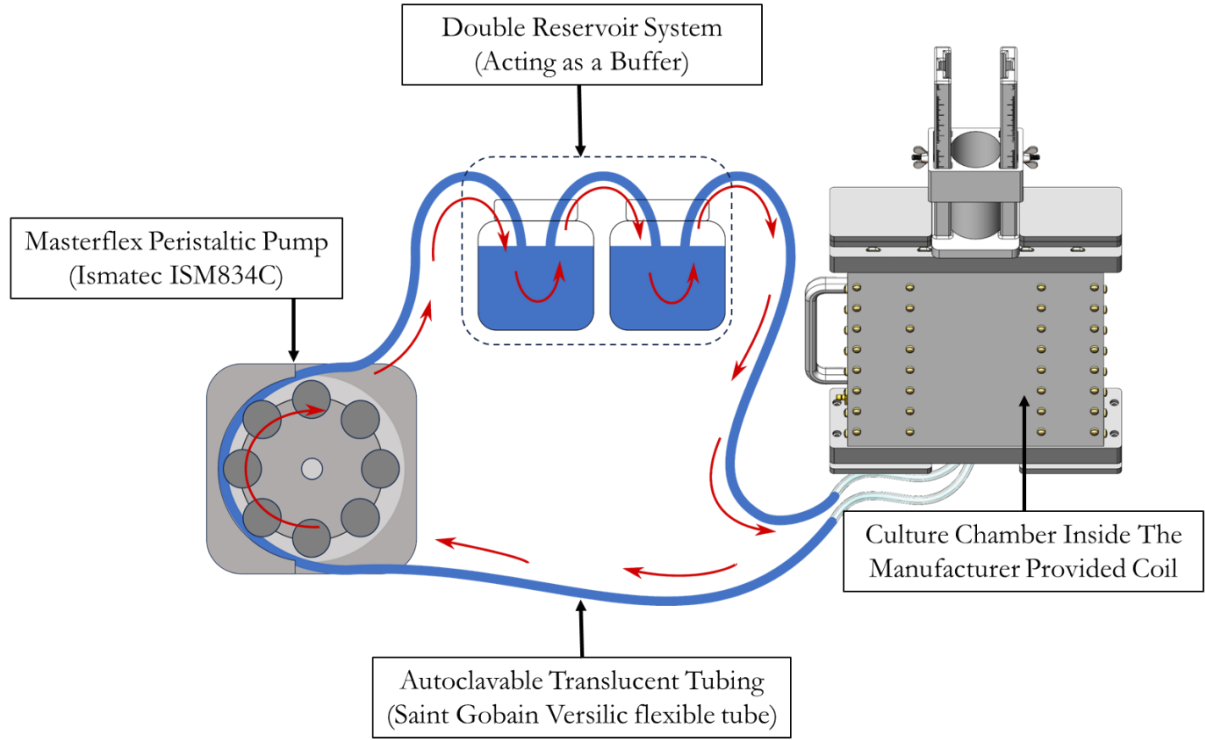


Figure 70 : Schematic of the complete fluidic perfusion setup for MRI velocimetry validation. A Masterflex peristaltic pump (left) drives medium through autoclavable tubing into a dual-reservoir buffer system (center) to smooth pulsatile flow, then onward to the MRI-compatible culture chamber mounted inside a volume coil (right). Red arrows indicate the forward perfusion path.

1.2.3 MRI Acquisition and Processing

A scan was carried out by placing the filled and sealed culture chamber inside the provided volume coil so that the scaffold was positioned in the center of the bore.

A Velocity map sequence acquisition was carried out by using a Spin Echo based phase contrast MRI acquisition. Flow encoding was performed using pairs of bipolar gradients placed before and after the 180° refocusing pulse. These gradients introduced a phase shift

$$\phi = \gamma M_1 v \quad (\text{IV-4})$$

that is proportional to the velocity v of moving spins, while stationary spins remain unaffected. Velocity was encoded in all three spatial directions (x, y, and z), using a velocity encoding value (VENC) of 20 mm/s, selected to be well above the maximum expected velocities, preventing phase aliasing, while still providing good sensitivity for the flow rates under investigation. The scan parameters included an echo time (TE) of 20 ms, a zero-filling factor of 4, for an isotropic resolution of $250 \mu\text{m}^3$ (matrix size $256 \times 256 \times 256$ across a $64 \times 64 \times 64 \text{ mm}^3$ field of view).

To correct for background phase artifacts (e.g., from magnetic field imperfections), a reference scan was acquired with no flow. The velocity-induced phase shift ($\Delta\Phi_v$) was then isolated by subtracting the phase of the reference scan (Φ_{ref}) from the phase of the flow scan (Φ_{flow}):

$$\Delta\Phi_v = \Phi_{\text{flow}} - \Phi_{\text{ref}} \quad (\text{IV-5})$$

This phase difference was converted into a velocity vector (v) using the following relationship:

$$v = \frac{\Delta\Phi_v \cdot \text{VENC}}{\pi} \quad (\text{IV-6})$$

A 3D phase unwrapping algorithm was subsequently applied to correct for any phase wrapping artifacts. The final velocity maps for each direction (v_x, v_y, v_z) were processed in MATLAB. The velocity magnitude ($\|v\|$), or speed, at each voxel was calculated as the Euclidean norm of the three vector components:

$$\|v\| = \sqrt{v_x^2 + v_y^2 + v_z^2} \quad (\text{IV-7})$$

1.3. Data and Profile Analysis

For both CFD simulations and measurements, a rectangular ROI spanning the full channel width on the central slice was defined to extract the mean axial velocity $\bar{v}(r_i)$ and its standard deviation $\sigma(r_i)$ at discrete lateral positions r_i . These measurements were fitted to a three-parameter Poiseuille model:

$$v_{\text{model}}(r) = v_0 + A(R^2 - r^2), \quad (\text{IV-8})$$

where R is the effective hydraulic radius, $A = \Delta P / (4\mu L)$ encodes the pressure gradient, and v_0 accounts for any uniform offset. Fitting was performed in Python using SciPy's "curve_fit", minimizing the weighted sum of squared residuals:

$$\min_{R, A, v_0} \sum_i \frac{\bar{v}(r_i)}{\sigma(r_i)} [\bar{v}(r_i) - v_{\text{model}}(r_i)]^2, \quad (\text{IV-9})$$

emphasizing points with lower relative variance. Initial parameter guesses and bounds were set as $R_0 = 1.25 \text{ mm}$, $A_0 = (\max \bar{v} - \min \bar{v}) / R_0^2$, $v_{0,0} = \min \bar{v}$. All fits employed a trust-region-reflective algorithm with up to 20 000 function evaluations. The best-fit values of R and A were used to confirm the laminar parabolic profile correlating to the pressure drop $\Delta P = 4\mu L A$ across the channel.

From the resulting mean velocity profile (Figure 72), absolute shear stress profiles were also calculated (using equation (IV-1))

2. Results and Discussion

The steady-state velocity fields obtained from both CFD simulations and MRI-based measurements are presented in Figure 71. Across both flow rates investigated (7 mL/min and 20 mL/min), the velocity magnitude maps show a high degree of qualitative similarity between simulated and experimental datasets. For the higher flow rate (20 mL/min), the velocity distribution exhibits stronger peak magnitudes concentrated in the central scaffold channels, as expected from the increased inlet momentum. At the lower flow rate (7 mL/min), the velocity field is reduced in magnitude but retains the same laminar, parabolic distribution across the channels. The regions of high velocity are consistently localized at the channel inlets, gradually diminishing toward the outlet due to viscous dissipation.

The defined rectangular ROI (highlighted in red in Figure 71) allowed extraction of axial velocity profiles for direct quantitative comparison. Figure 72 presents the mean speed distribution across the channel width for both flow rates.

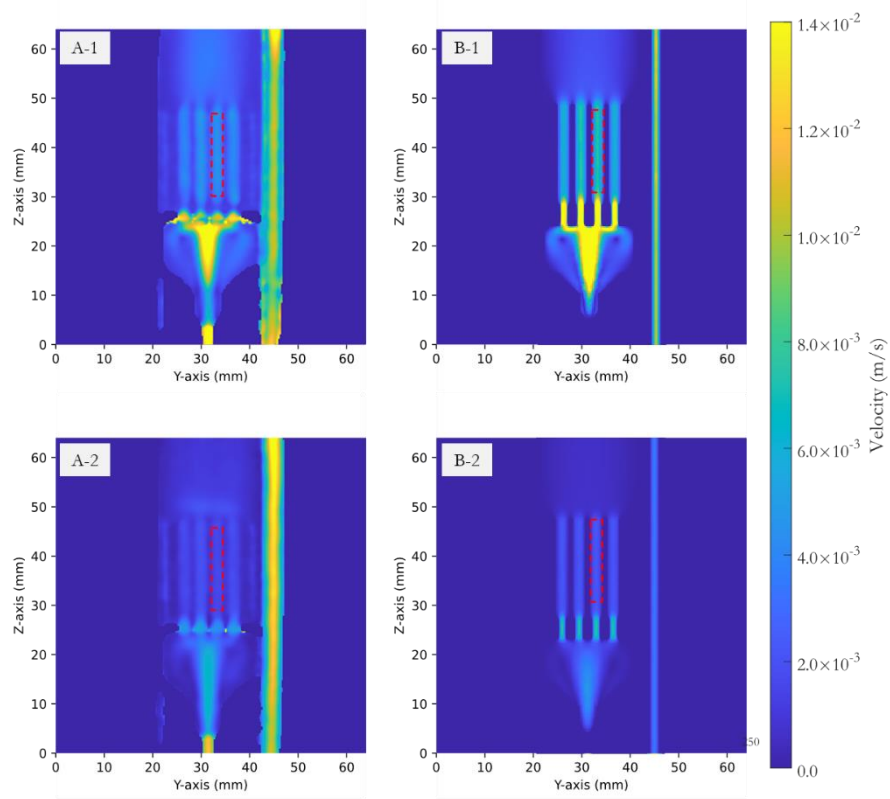


Figure 71: Comparison between experimentally measured (A-1, A-2) and CFD-simulated (B-1, B-2) steady-state velocity magnitude maps inside the culture chamber scaffold region for two different flow rates. The top row (A-1, B-1) corresponds to a flow rate of 20 mL/min, while the bottom row (A-2, B-2) corresponds to a flow rate of 7 mL/min. Velocity maps are overlaid with a red dashed rectangular ROI indicating the region used for quantitative analysis.

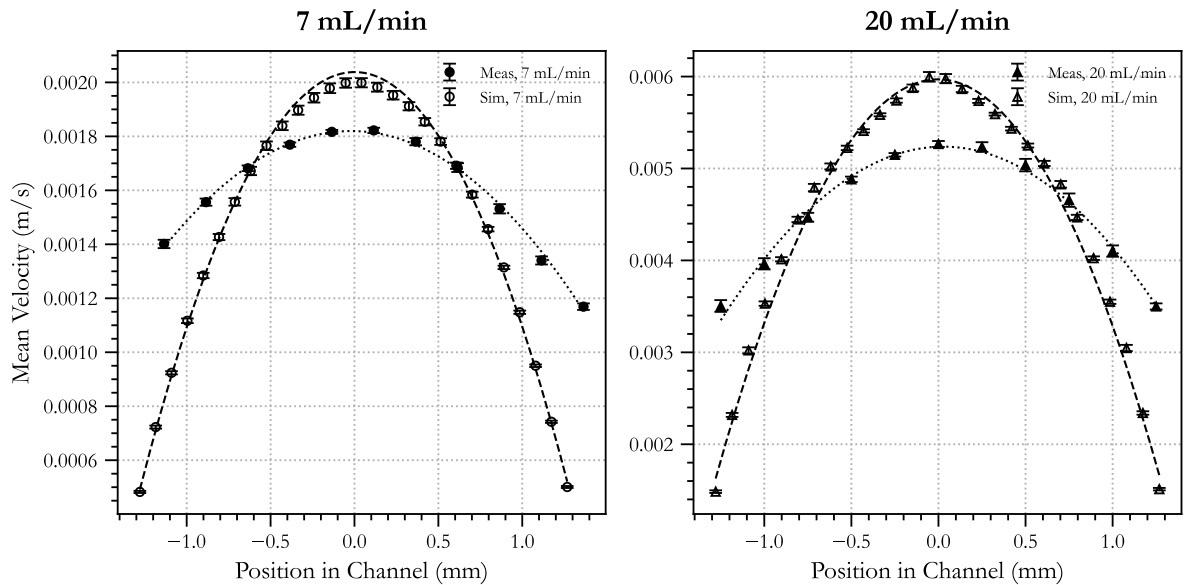


Figure 72: Comparison of measured and simulated mean velocity profiles across a single 2.5×2.5 mm scaffold channel for flow rates of 7 mL/min (left) and 20 mL/min (right). Experimental data (solid markers) are compared with CFD simulation results (open markers).

The resulting profiles conform well to the generalized Poiseuille equation, displaying a clear parabolic shape characteristic of laminar flow in rectangular channels. At 7 mL/min, the measured peak velocity reached approximately 1.82×10^{-3} m/s, whereas the 20 mL/min condition yielded a peak velocity of roughly 5.25×10^{-3} m/s.

Although at first glance, the velocity maps inside the scaffold channels presented in Figure 71 seem to loosely follow the CFD predictions, flow distribution profiles presented in Figure 72 show glaring discrepancies between CFD simulations and measurements. For example, the measured peak velocity was only 90 % of the simulated peak velocity while the parabolic shapes of the measured data profiles appear flattened mostly due to partial volume effects in the MRI acquisition. The true flow inside the channel should follow a parabolic Poiseuille profile, where the velocity decreases quadratically from the center to zero at the walls. However, due to the finite voxel size, what MRI measures is not the pointwise velocity but the average velocity within each voxel. This volume averaging has several effects. Near the tube walls, the voxels partially overlap regions where the fluid velocity is either close to zero (biased downward) or influenced by the fluid velocity of surrounding channels (biased upward). As a result, the measured velocity in these peripheral voxels is either lower or higher than the true local velocity just inside the wall, leading to an apparent flattening of the velocity profile near the boundaries. At the center, where the true velocity is maximal, the voxel still averages over a finite width, so the measured central velocity is also underestimated.

The parabolic shape of the profile is smoothed out in the MRI data, and the highest and lowest velocities are both biased. This means that while the general shape of the flow profile is preserved, partial volume effects lead to a systematic underestimation of both peak velocity and wall shear rate. This effect is expected to become more significant when the number of voxels across the channel is limited, and it highlights the importance of considering spatial resolution when interpreting flow measurements in small structures. Furthermore, artificial resolution enhancement techniques such as zero filling will only accentuates partial-volume biases, making flattened peaks look artificially crisp and wall-region smearing more pronounced. This means that, although useful to get a quick insight into the velocity profile inside the scaffold channels, MRI velocimetry methods will only be able to capture the true velocity profiles when the resolution is sufficiently high. the MRI based method used for experimental scaffold and chamber design validation can only be recycled for flow monitoring in longitudinal studies if the spatial resolution is sufficiently high. This method will show its practicality mostly during the growth process where scaffold misalignment, bubbles or changes due to tissue growth will be difficultly assessed through CFD based analysis.

Overall, CFD validates the proposed culture chamber fluidic design, confirming its ability to sustain uniform, predictable flow profiles across the scaffold channels under physiologically relevant flow rates.

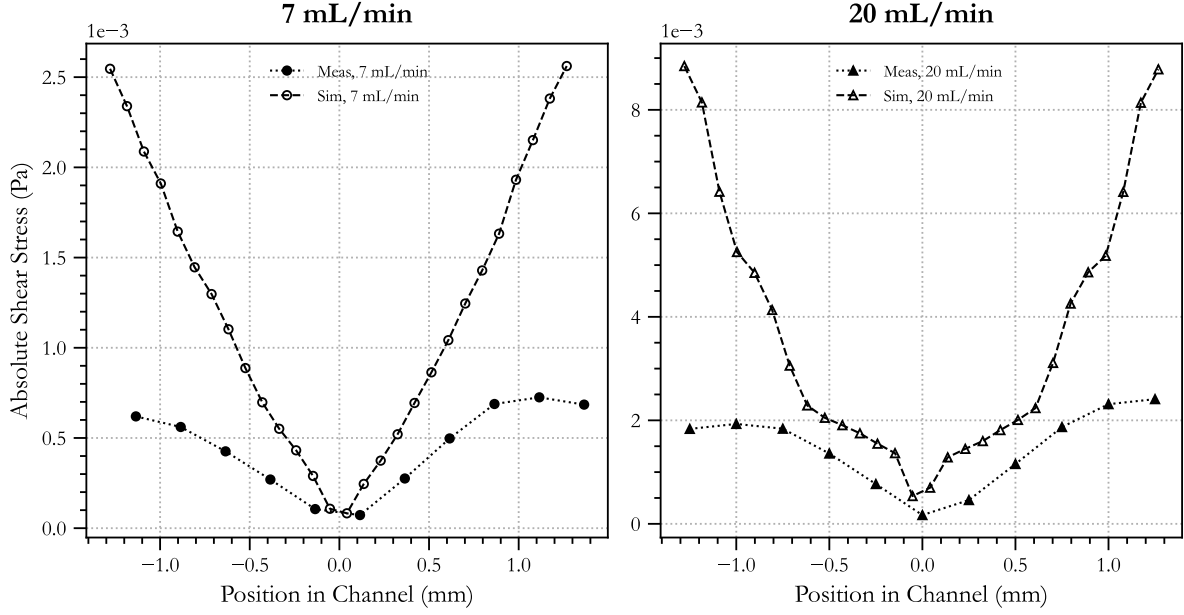


Figure 73: Comparison of the calculated measured and simulated absolute shear stress profiles across a single 2.5×2.5 mm scaffold channel for flow rates of 7 mL/min (left) and 20 mL/min (right). Experimental gradients of the velocity field (solid markers) and CFD-derived shear stress (open markers) both exhibit the expected peak at the channel walls and a minimum at the center, confirming Poiseuille-type behavior.

The validated CFD model now allows for a confident estimation of wall shear stress, confirming that the proposed flow rates (7–20 mL/min) result in wall shear stress values between 2.6×10^{-3} and 8.9×10^{-3} Pa, which is orders of magnitude below the critical shear stress for cell detachment (between 1 and 3 Pa [165]). For certain applications there optimal shear stress range typically reported for promoting cell differentiation (0.1 to 2 Pa depending on exposure time and cell type [168]) is still above the obtained values. While higher shear stresses (≥ 0.1 Pa) have been associated with enhanced cellular response, lower values, such as those obtained here, may be suitable for minimizing mechanical damage in delicate early constructs or maintaining undifferentiated states, depending on the application. Furthermore, if cellular response and differentiation are paramount, as everything from the scaffold to the chambers and other components are 3D printed, adjusting the flow rates and geometries to enable custom profiles is a non-negligible advantage of the proposed approach. In those cases, the applicability of the CFD model as a predictive tool for further optimization of the system without the need for extensive experimental trials will be instrumental. By accurately predicting local flow behavior and shear stress distributions, the model enables rational tuning of the culture environment to meet specific biological requirements. This capacity for *in silico* optimization not only streamlines the design process but also reduces time and cost, making the system particularly advantageous for applications in tissue engineering where mechanical cues must be precisely controlled to guide cell growth.

IV.C. Fitted RF Coil Integration

1. Integrated RF Coil Design

The 3D-plastronics setup included a BLGR coil with a diameter of 34 mm and a length of 65 mm directly integrated onto the surface of the culture chamber. As shown in Figure 6-B, the coil and the tune-match circuit were designed to be embedded in the outer surface of the chamber,

with the feed point positioned at the base of the BLGR. A 13-capacitor variation of the BLGR was selected here as a highly sensitive and cost effective option suited for cylindrical samples.

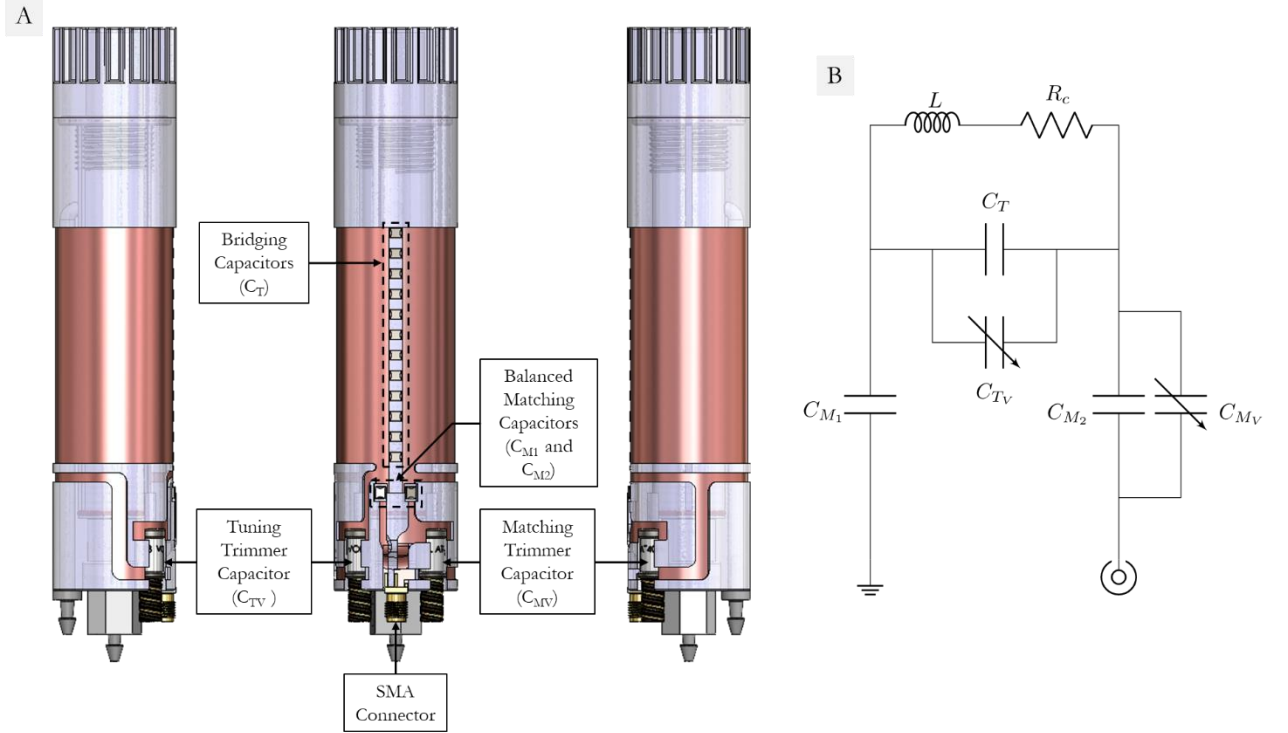


Figure 74: A) 3D Plastronics-based RF coil integrated design: front and side views. B) Equivalent circuit of the unshielded design.

Figure 74-B, shows a general equivalent circuit of the BLGR with L being the intrinsic inductance from the cylindrical copper loop, R_c the characteristic resistance of the loop, C_T the equivalent capacitance of the parallel bridging capacitors necessary to attain the lower frequency ranges doubling as tuning capacitors, C_{TV} and C_{MV} are variable trimmer capacitors for precise coil tuning and matching to compensate sample coupling related mismatch, C_{M1} and C_{M2} matching capacitors creating balanced matching circuit with $C_{M1} \approx C_{M2} + C_{MV}$. Although less critical than for other classical coil geometry, as the BLGRs are less susceptible to losses from the electric field, by balancing the probe, the electric field is still reduced by a factor of two which decreases the dielectric losses by a factor of four[104]. This is imperative in the low-field regimes where achieving the highest possible coil sensitivity is required to compensate for the inherent signal limitation.

2. Shielded Housing and Positioning Mechanism

Unlike large clinical scanners requiring room-scale Faraday cages, benchtop systems designed to operate outside such environments rely on integrated local shielding strategies. At low field, although some teams using larger scanners forgo shielding altogether [169], it has been shown that for smaller samples, a good local RF shield may improve SNR by over five times [27].

The 3D-plastronics setup included a BLGR coil with a diameter of 34 mm and a length of 65 mm directly integrated onto the surface of the culture chamber. The chamber was housed in a (178 × 52 × 218 mm) shielded 3D-printed Transparent PLA (Ultimaker) enclosure. The RF shielding was implemented by wrapping a 25 μ m thick copper foil inside the enclosure, which was then tightly closed using brass screws. The copper foil thickness was determined to prevent significant induced gradient delays [27] while maintaining adequate shielding. The shield was also grounded through both the SMA feedthrough, and the support guard rails, ensuring a low impedance path to ground. Inside the enclosure, a 3D FDM-printed White PLA (Ultimaker)

sliding mechanism was designed to hold the chamber and ensure precise sample centering and positioning. Figure 75 shows the enclosure in detail.

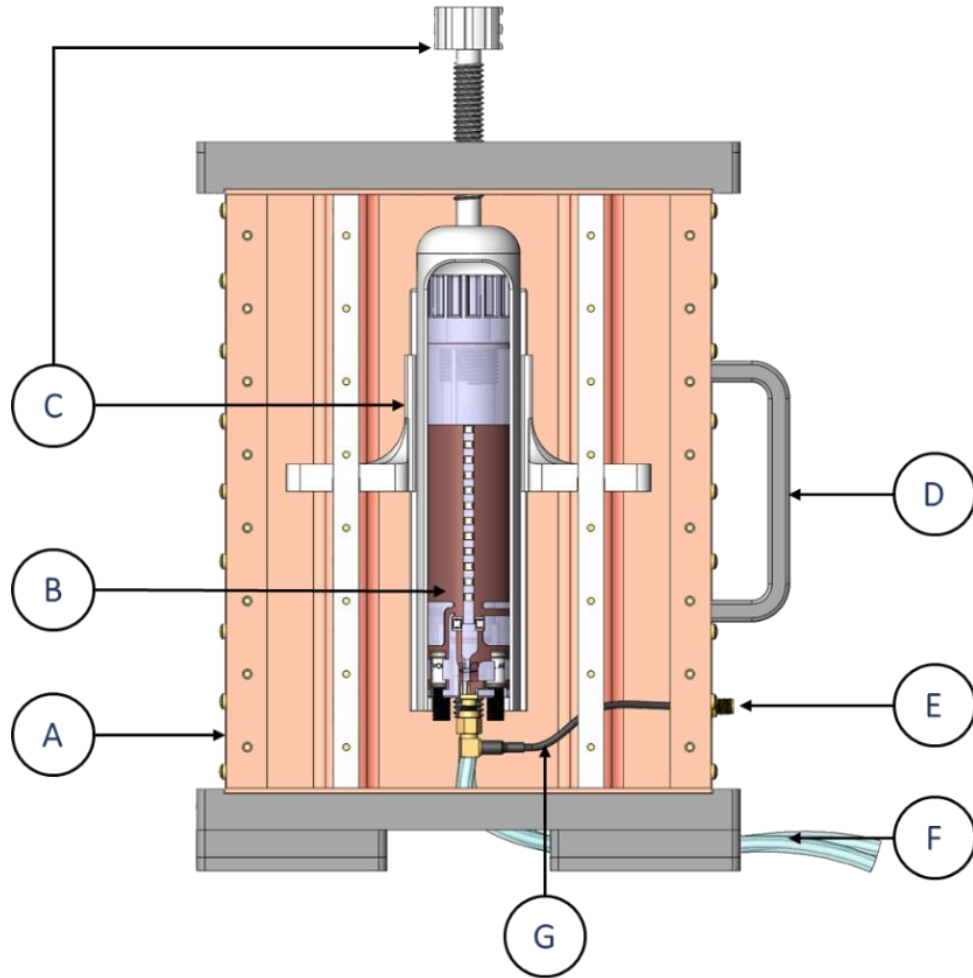


Figure 75: Fully integrated 3D Plastronics-based shielded coil: (A) Shielded enclosure: 3D-printed housing lined with 25 μm copper foil to ensure RF shielding and low-impedance grounding via SMA feedthrough and structural rails. (B) Functionalized culture chamber: High-temperature SLA-printed chamber featuring a directly integrated BLGR coil and embedded tuning-matching circuit fabricated using the R3DP method. (C) Sliding mechanism: Custom FDM-printed support designed to precisely position the chamber inside the enclosure and ensure repeatable placement. (D) Handle: External grip allowing easy insertion and removal of the enclosure from the MRI system. (E) RF connection: SMA interface providing electrical access to the embedded coil circuitry for transmit/receive operation. (F) Fluidic inlet/outlet: Ports allowing media circulation and nutrient delivery during *in situ* imaging. (G) Internal connection cable: Coax cable leading from through-panel SMA connector grounded through the shield to the SMA connector.

Furthermore, an effective RF shielding must block environmental RF noise. However, sub-optimal shielding will compound this noise in an already non-ideal transmit/receiver chain, which adds another noise contribution to the system through the noise factor F of the chain characterized by a sharp drop in SNR (Annex V shows a short experiment evidencing this phenomenon).

IV.D. RF Validation Through EM Simulations

Similarly to the fluidic system validation, a purely analytical approach to RF coil design is insufficient for complex geometries like the integrated BLGR. Electromagnetic (EM) simulations are necessary to accurately predict the coil's performance *in silico* before fabrication. This computational approach allows for the characterization and optimization of key performance metrics, such as the magnetic field (B_1) homogeneity, tuning and matching behavior, and overall efficiency (Q-factor). By simulating the coil within its complete environment, including the culture chamber, sample load, and RF shield, we can anticipate its real-world performance, significantly reducing the need for costly and time-consuming iterative prototyping.

1. Materials and Methods

1.1. Computational Framework

Electromagnetic simulations were performed using CST Microwave Studio (Dassault Systèmes, France). The Frequency Domain Solver, based on the Finite Element Method (FEM), was selected for its high accuracy and computational efficiency in analyzing electrically small, high-Q resonant structures like the one presented within a small frequency range. A co-simulation approach was employed, integrating the 3D full-wave simulation of the coil and its environment with a circuit simulation (CST Design Studio) to accurately model the behavior of the lumped-element and tuning/matching network.

1.2. Geometry, Model Setup and Material Properties

The complete CAD model of the 3D-plastronics assembly, including the BLGR coil, and the culture chamber was exported as .step files from SOLIDWORKS and imported into the simulation environment. For simplicity, the copper-lined shielded enclosure was modeled as a copper box and added directly in CST as its geometry is far less complex which should significantly reduce computational complexity without impacting the accuracy of the internal field distribution. A cylindrical phantom spanning the entire chamber was placed inside the culture chamber to represent the sample load (see Figure 76).

For the co-simulation setup, all discrete capacitors (port 4 to 18) were modeled using touchstone files provided by the manufacturer while the trimmer capacitors C_{TV} and C_{MV} were considered ideal (port 2 and 3) (see Figure 76-C). This idealization is a minor limitation, as with the trimmers high-Q values (>2000 [170]) should limit their contribution to the overall circuit loss. The passive component values used for the BLGR can be found in Table 14. The modeled capacitors were High Q non-magnetic MLCCs (Knowles Syfer) and High Q PTFE 1.5 to 40.0pF non-magnetic trimmers (AT40HV - Knowles Voltronics).

	C_T	C_{M1}	C_{M2}	C_{TV}	C_{MV}
Number of Chip Capacitors	13	1	1	1	1
Values (pF)	12820	180	150	1.5 to 40.0	1.5 to 40.0
Estimated Q -factor @13 MHz	~800	~1000	~1000	>2000	>2000

Table 14: Values of the High Q non-magnetic MLCCs (Knowles Syfer) and High Q PTFE non-magnetic trimmers used in the fabricated setup. All values were obtained from manufacturer provided datasheet and models [154], [170]

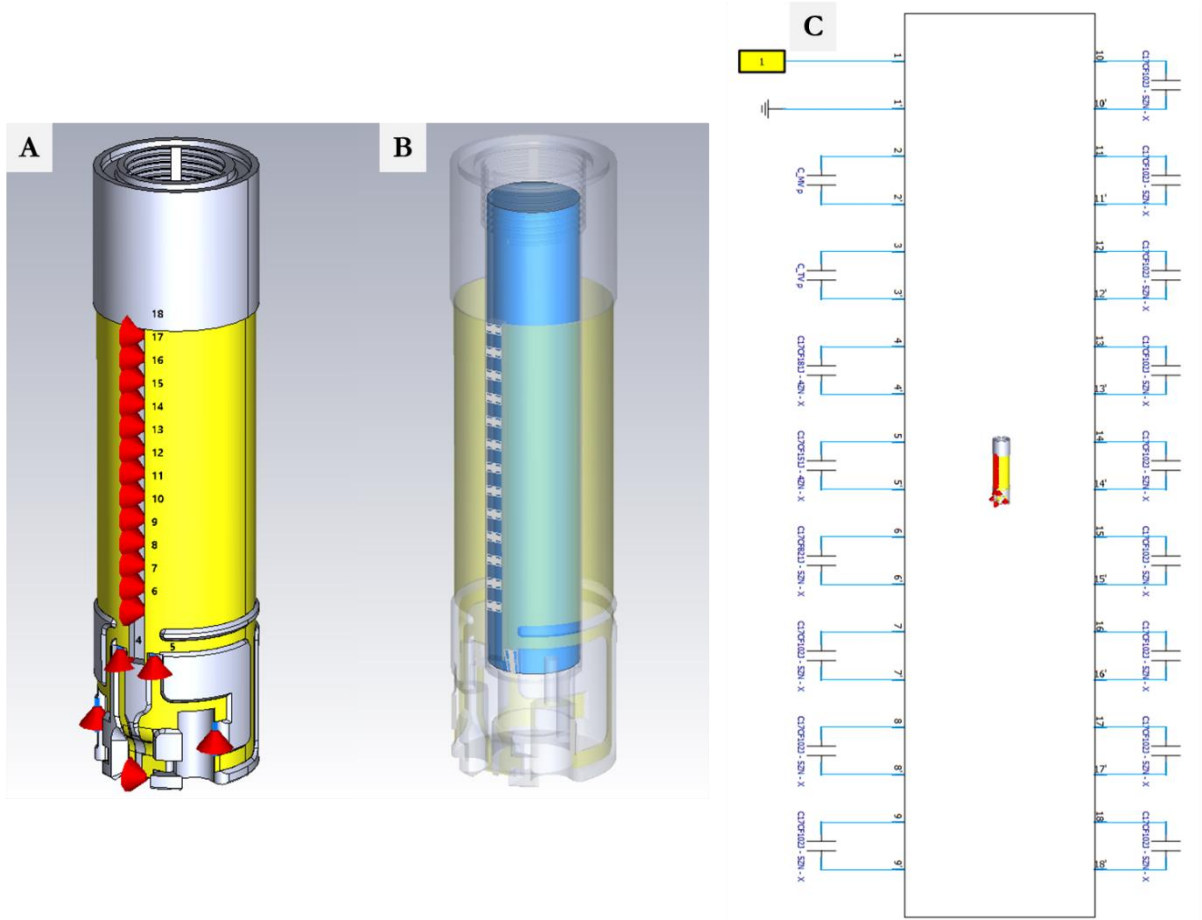


Figure 76: BLGR coil model and co-simulation setup(A) CAD of the 3D-plastronics BLGR: HTV2 substrate (grey) with copper conductive traces with discrete ports/pads for the distributed series capacitors are highlighted in red and numbered along the resonator. (B) Loaded configuration showing the culture chamber and a cylindrical 16.3 mM Copper (II) Nitrate Trihydrate contrast agent solution phantom occupying the entire chamber. (C) CST Design Studio schematic used for co-simulation.

The coil conductors were modeled as copper with a defined thickness of $40 \mu\text{m}$ and a conductivity of $5.8 \times 10^7 \text{ S/m}$. The culture chamber and sliding mechanism were assigned the properties of the SLA resin (dielectric constant $\epsilon_r \approx 3.33$, loss tangent $\tan\delta \approx 0.019$ [153]). The phantom was modeled with properties matching a 16.3 mM Copper (II) Nitrate Trihydrate contrast agent solution ($\epsilon_r = 78$, conductivity $\sigma = 0.35 \text{ S/m}$) which is used for experimental coil characterization (Annex VI shows the method used to estimate 16.3 mM Copper (II) Nitrate Trihydrate contrast agent solution properties). The passive components (capacitors) were modeled as lumped elements with their specified values and Q-factors.

1.3. Simulation Setup and Analysis

The simulation was run over a frequency range of 2 MHz centered at 13 MHz. A discrete port (Port 1) was defined at the SMA feed point to excite the coil and calculate the scattering parameters (S-parameters). To match the expected 12.5V input voltage from the scanner a port signal amplitude $1.77 \text{ W}^{1/2}$ was used for all field calculations. H-field monitors, coupled with the integrated post-processing MRI Toolbox, were used to assess the field distribution [157].

To ensure the accuracy and independence of the results from the mesh discretization, the solver's adaptive mesh refinement feature was utilized. The simulation was set to automatically

refine the mesh in areas with high field gradients until the change in the S-parameters between subsequent passes was less than 1%. This convergence criterion was met after several passes, resulting in a final, optimized mesh of approximately 500,000 tetrahedrons, which was used for the final simulations presented here.

The magnetic field (B_1) distribution was calculated on the entire chamber volume. 3D Field Maps were exported as HDF5 files and treated using a custom Python script to evaluate the coefficient of variation within a centered $\text{Ø}20 \times 40$ mm cylindrical ROI.

The simulation results were used to extract the return loss (S_{11}) to assess the coil's matching to the 50Ω system impedance and the influence of the load. The unloaded (Q_u , simulated without the phantom) and loaded (Q_l , simulated with the phantom) quality factors were calculated from the simulated S-parameters to estimate coil efficiency using the graphical Smith chart-based method described in Chapter II.

2. Results and Discussion

The results from the EM simulations provide a comprehensive prediction of the RF coil's performance. Figure 77 displays the simulated B_1 field distribution for the fully loaded and shielded coil.

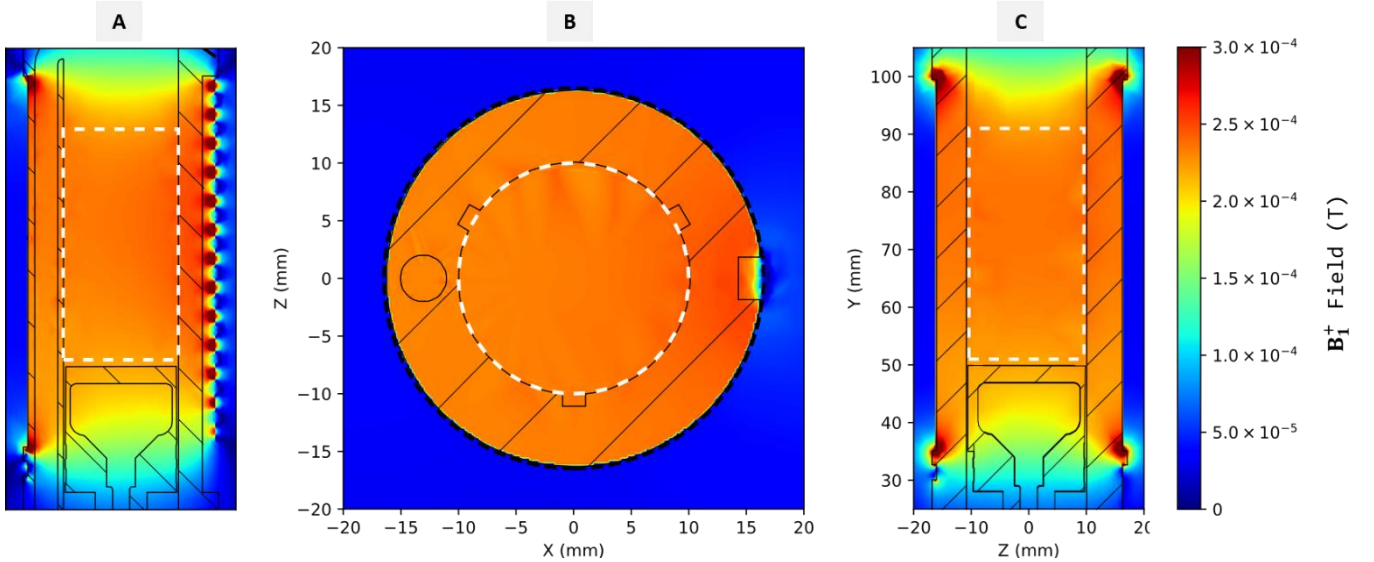


Figure 77: Simulated B_1^+ magnitude maps. Central slices from the full-wave FEM model with the chamber and shield included: (A) sagittal, (B) axial, and (C) coronal planes. The cylindrical region of interest ($\text{Ø}20 \times 40$ mm) used for homogeneity analysis is outlined with white dashed lines. Within this ROI, the field is highly uniform (coefficient of variation $\approx 2.07\%$). Black overlays indicate the chamber and coil geometry.

The simulated B_1 field map demonstrates excellent field homogeneity (CoV of 2.07 %) homogeneity across the central region of interest where the scaffold is located. The characteristic field pattern of the BLGR, with its high homogeneity along the coil axis, is well-reproduced despite the more complex geometry.

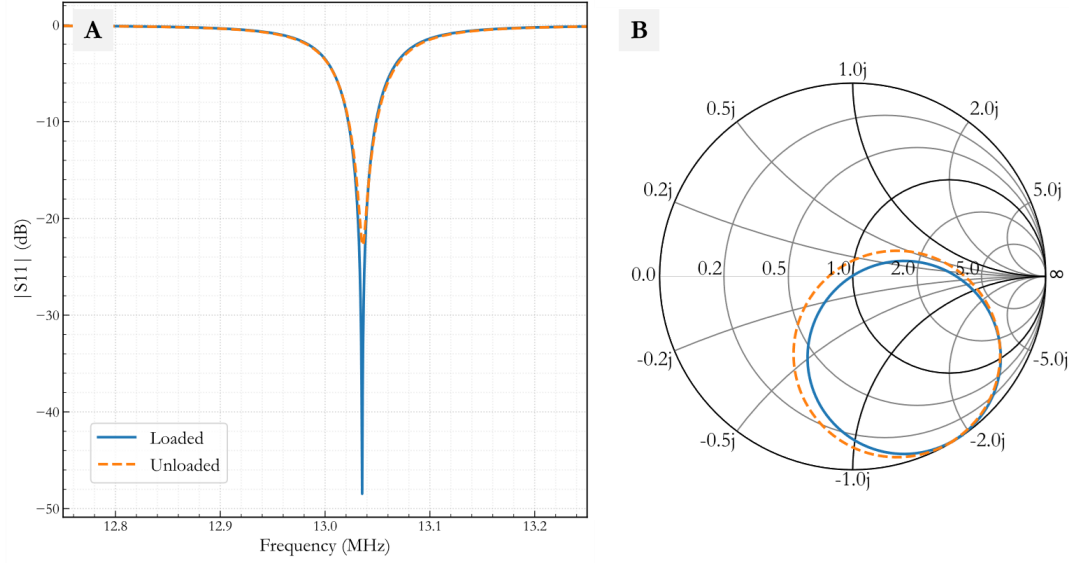


Figure 78: Simulated matching behavior. (A) $|S_{11}|$ versus frequency for unloaded (dashed) and 0.9% NaCl-loaded (solid) conditions, showing a sharp resonance tuned near 13.03 MHz with a deep return loss (~ -50 dB under load). (B) Corresponding Smith-chart trajectories, confirming good 50 Ω tuning/matching with minimal load-induced shift.

The return loss plot (Figure 78) shows a sharp resonance at the target frequency of 13.035 MHz, with a simulated S_{11} value of almost -50 dB. This indicates a highly efficient transfer of power between the MRI system and the coil, with minimal reflection ($< 0.1\%$). The simulation also allowed for the fine-tuning of the variable capacitors *in silico* to achieve this optimal match, confirming the chosen component values were appropriate.

The simulated quality factors were calculated to be $Q_u \approx 342$ and $Q_\ell \approx 327$. The ratio of these values ($Q_u/Q_\ell \approx 1.046$) confirms that under this loading condition the coil is dominated by coil noise rather than sample noise which as we have seen is suboptimal. While maximizing the SNR was a primary objective, this was ultimately constrained by the physics of the small coil operating at a low frequency, where the intrinsic resistance of the coil conductors constitutes the primary source of loss. With the SNR performance constrained by the coil-dominant noise regime, the design was therefore optimized for the other key metrics of high field homogeneity and a maximized filling factor.

Nonetheless, these simulation results collectively validate the RF design, confirming its suitability within the integrated culture system. These findings provide a strong benchmark for the subsequent fabrication and experimental validation, where the simulated S-parameters and Q-factors will be directly compared against measurements from a vector network analyzer to confirm the accuracy of the model.

IV.E. R3DP Fabrication

1. Material Selection

All the culture chamber parts were fabricated using a Form 3 SLA printer (Formlabs, Somerville, USA) with High Temp V2 (HTV2) resin (FLHTAM02) from Formlabs' engineering resin line. HTV2 was selected as the structural substrate for a 3D Plastronics culture chamber because it offers exceptional thermal stability (HDT ≈ 238 $^{\circ}\text{C}$ @ 0.45 MPa [171]) making it highly compatible to stringent autoclave cycles and dimensional fidelity during post-processing, plating,

and sterilization steps that typical thermoplastic polymers struggle with. This resin has been tested as the primary substrate for R3DP fabrication and was used for the development the techniques and therefore is highly compatible with the process [129], [122], [114], [130], [120].

Nevertheless, HTV2 is not inherently commercialized as a biocompatible material. In a study comparing the biocompatibility of 3D printed plastics for use in bioreactors and cell cultures, Licata et al. [172] and Hart et al. [173] have shown that Formlabs High Temp resin is structurally robust and autoclavable for bioreactor parts, and is cytocompatibility with common, less-delicate cell types such as fibroblasts and electrogenic cells. However, during the autoclave process HTV2 leaches cytotoxic components that kill or de-differentiate highly sensitive cultures such as Human induced pluripotent stem cells (hiPSCs). Therefore, in a context where highly sensitive cell structures are used, it will be imperative to shift towards a more cytocompatible material. Such material include nylon as presented by Licata et al. [172] or many other high performance polymers. Such polymers and composites have already been proven at the AMPERE laboratory o be highly compatible with the R3DP process [127].

2. SLA 3D Printing

Once the material selected, 3D printing was carried out to obtain the base substrate (Figure 80-A). The STL models were sliced using PreForm software with preloaded material profiles to ensure optimal print settings, and the parts were printed at a 50 µm layer resolution over approximately 6 hours, resulting in high-precision components with smooth surfaces. After printing, support structures were removed manually using cutting tools and sanding techniques, and the parts were washed in a 99% isopropyl alcohol (IPA) bath for 20 minutes to eliminate excess uncured resin before being allowed to dry completely. A final post-curing step was then performed using the Form Cure station, where the parts were placed in a heated UV curing chamber at 80 °C for 120 minutes, following manufacturer-recommended times and temperatures to fully polymerize the resin and enhance its mechanical properties.

3. Metallization

3.1. Electroless Plating

After 3D printing, a first ~5 µm thick layer of copper was deposited with electroless plating (Figure 80-B) using the following steps:

- **Surface preparation: Surface Etching / Roughening:**

Surface activation is performed using sandblasting with 50 µm corundum at 4 bar through a 1.2 mm nozzle positioned 5 cm from the surface. The abraded components undergo ultrasonic cleaning in deionized water (10 min), drying, isopropanol wiping, and a final rinse with deionized water.

- **Surface activation:**

After a thorough wash, the etched polymer is immersed in a colloidal palladium solution. These colloids typically consist of (Pd²⁺) ions stabilized by tin chloride (SnCl₂) in an acidic medium. Upon rinsing, local pH changes induce the reduction of Pd²⁺ to Pd⁰:



Metallic Pd particles deposit onto the polymer surface and act as nucleation sites for the subsequent electroless plating reaction (Figure 79).

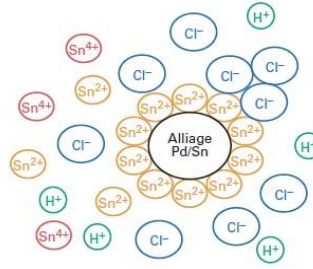


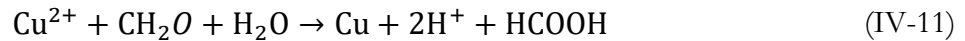
Figure 79 :Structure of a palladium-tin (Pd/Sn) colloid used for surface activation in electroless metallization. The colloidal particle consists of a Pd/Sn alloy core surrounded by tin ions (Sn^{2+} , Sn^{4+}), chloride ions (Cl^-), and protons (H^+) in an acidic medium.

Catalyst deposition is achieved by immersing the samples in a 1 L bath containing 5 mL of a palladium-based activator (Macuplex Activator-D34C) and 220 mL of 37% hydrochloric acid at 30 °C for 3 min. Excess catalyst is mechanically removed from non-recessed areas by manual sanding, leaving the activator confined to the predefined channels.

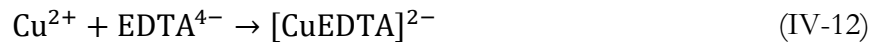
- **Metal deposition:**

Electroless copper deposition is carried out by immersing the activated parts in a commercial plating solution (Mid Copper 100XB) at 60 °C for 2 h, followed by thorough rinsing and air drying for 12 h.

In the final chemical bath, metal ions, commonly (Cu^{2+}) from copper sulfate are reduced by a chemical reducing agent such as formaldehyde (CH_2O). The overall redox reaction for copper deposition is [174]:



The presence of complexing agents such as EDTA (ethylenediaminetetraacetic acid) keeps Cu^{2+} in a metastable complex to prevent premature precipitation and maintain bath stability:



Copper deposition proceeds only at the Pd^0 catalytic sites, making the process selective and uniform over complex 3D geometries.

3.2. Electroplating

To enhance the electrical conductivity and thickness of the tracks, the part then underwent electroplating (Figure 80-C). using an electrolyte composed of 220 g/L CuSO_4 , 3.1% H_2SO_4 , 0.2 mL/L HCl , and organic additives (0.2 mL/L Rubin T200A, 8 mL/L Rubin T200G, 2 mL/L Rubin T200E). A copper anode is placed 10 mm from the substrate, and plating is performed at a current density of 10 A/dm² for 25 min at 30 °C resulting in a total copper thickness > 40 μm (as verified by X-ray fluorescence (Bowman XRF)).

3.3. Trace Patterning

To pattern the conductive tracks, protective masking is applied using RS Matte Black Spray Paint to shield the designated areas by filling predefined grooves to create sharp and well-defined track edges (Figure 80-D). The masked parts were then immersed in a ferric chloride etching solution (Bernier P-2035L) maintained at 30 °C for 30 minutes to remove unwanted copper, with the etching process carefully monitored to prevent over-etching (Figure 80-E), after which the

parts were immediately rinsed with deionized water to halt the reaction and the Spray Paint was removed using an acetone, thereby revealing the conductive tracks (Figure 80-F). Finally, the parts were dried using compressed air and a clean cloth and inspected to ensure accuracy and continuity.

4. Component Soldering

Since high-temperature SLA resin was used, capacitors and the SMA connector were soldered by hand onto the designated contact points using lead-free low-temperature SnAgCu solder and proper flux to prevent substrate damage (Figure 80-G).

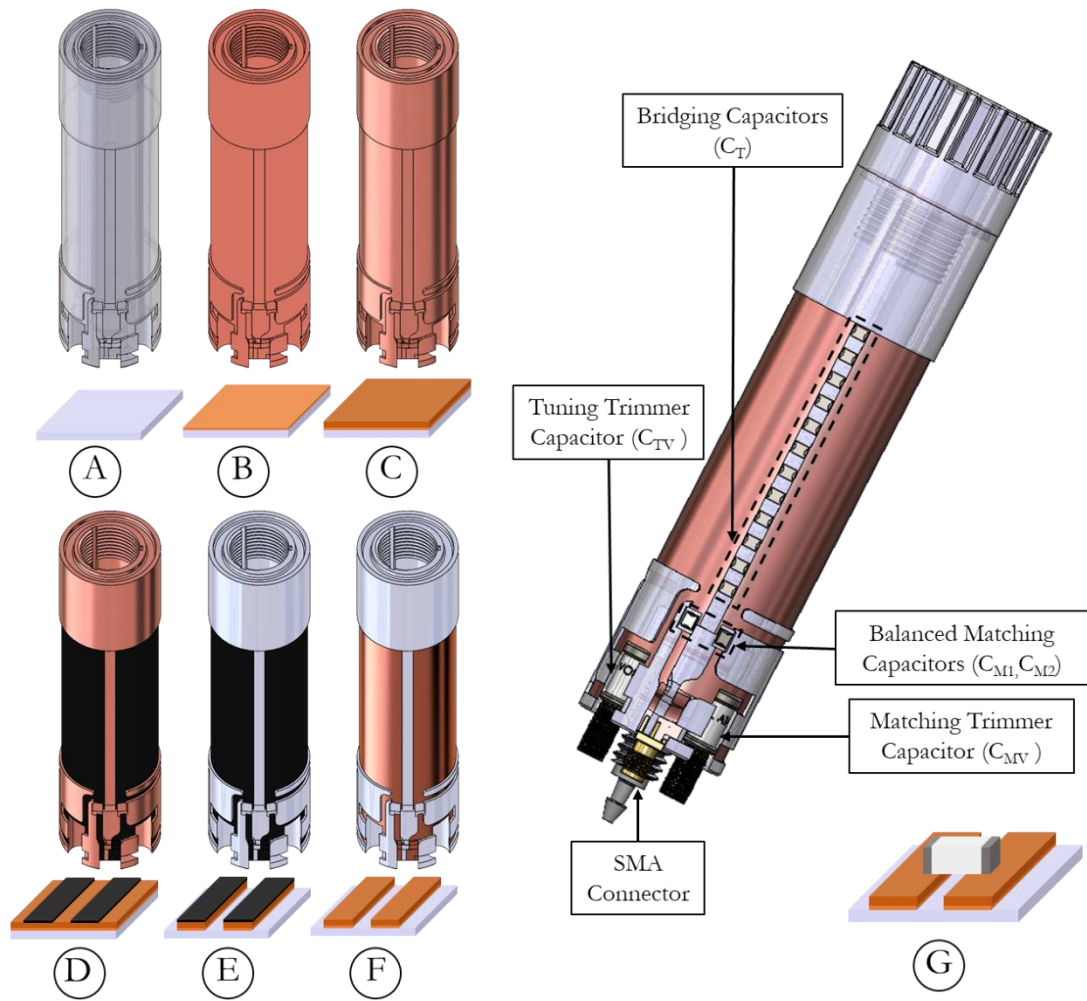


Figure 80: Overview of the 3D Plastronics fabrication process used to integrate the RF coil and matching network onto the culture chamber. (A) SLA-printed culture chamber produced using high-temperature resin. (B) Electroless copper deposition applied uniformly across the outer surface. (C) Electrolytic copper plating performed to increase the overall copper thickness and improve conductivity. (D) Selective masking applied to define coil and circuit geometries using spray paint over pre-designed grooves. (E) Ferric chloride etching is used to remove exposed copper, leaving behind the patterned conductive tracks. (F) Masking removed to reveal clean copper traces corresponding to the final circuit layout. (G) Final assembly with passive components and an SMA connector soldered onto the structure, completing the fully integrated coil with embedded tuning and matching circuitry.

5. Resulting Fabricated MRI Setup and Characterization

Photographs of the complete 3D Plastronics-based MRI setup are presented in Figure 81. The image shows, on the left (A), a close-up of the functionalized culture chamber featuring the integrated RF coil and embedded tuning-matching circuit, and on the right (B) the fully assembled system including the shielded enclosure.

Assembly of the printed and metallized components proceeded without mechanical or electrical issues. The copper traces showed good adhesion and durability throughout handling, and soldering of the surface components was performed reliably without damaging the copper traces or substrate. However, slight mechanical tension was observed at the SMA connector on the chamber side, indicating a potential point of fragility that may require mechanical reinforcement in future iterations.

The 3D-printed sliding mechanism enabled accurate and reproducible placement of the culture chamber within the shielded enclosure, while allowing for rapid repositioning and reconfiguration. During operation, the system remained mechanically stable and required no re-tuning across multiple imaging runs, demonstrating good load tolerance and mechanical robustness of the integrated circuit.

The integrated shielding effectively suppressed environmental noise without compromising access to the chamber or connectors. However, the current design limited the speed of setup and sample exchange (~ 10 minutes), primarily due to the need to fasten and unfasten multiple screws to open and close the enclosure.

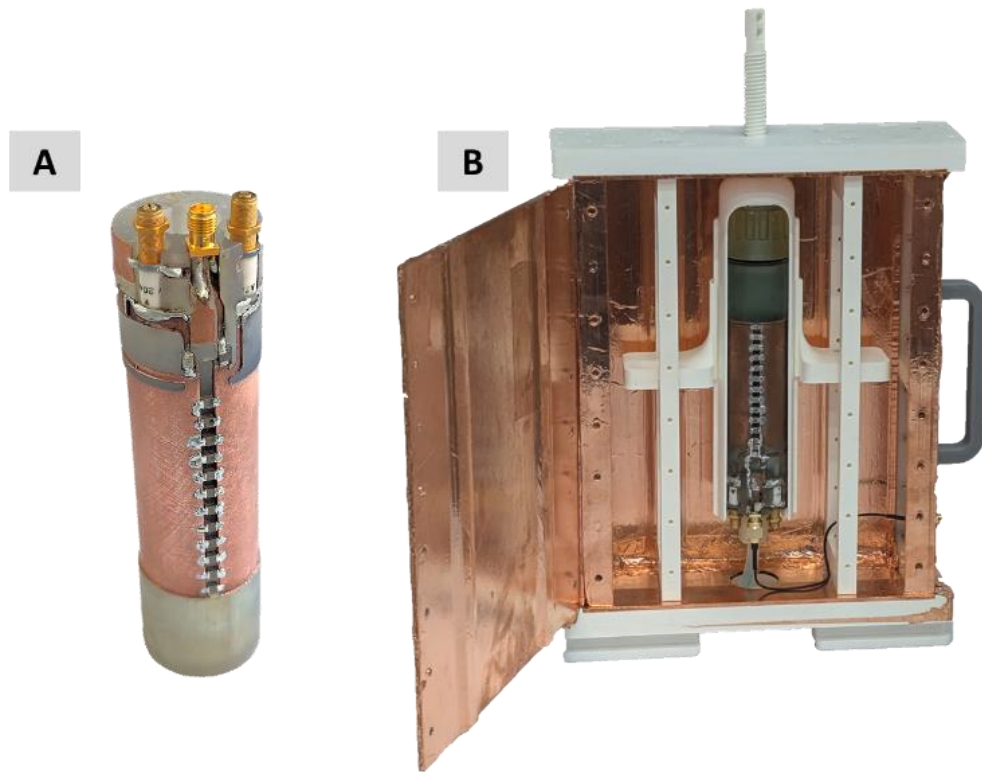


Figure 81: Fully integrated benchtop MRI setup using 3D Plastronics: (A) Close-up of the culture chamber with an embedded RF coil and integrated tuning/matching components directly on the outer surface. (B) Final assembly showing the shielded enclosure with sliding mechanism, fluidic and RF interfaces.

V. 3D Plastronics Setup Characterization and Validation

V.A. Materials and Methods

1. S-Parameters and Q-factor Measurements

The S-parameters and Q-factors were obtained for both the shielded reference coil and shielded fabricated coil under loaded and unloaded conditions. For loaded measurements, the entire culture chamber was filled with 16.3 mM Copper (II) Nitrate Trihydrate solution.

The S_{11} parameter was measured using a Rohde & Schwarz ZNL3 vector network analyzer, calibrated using the ZN-135 calibration kit. The shielded setups were connected to the network analyzer through 1 m RG316 coaxial cables, and the S_{11} parameter was obtained by measuring the reflection coefficient across the relevant frequency range (12 MHz – 14 MHz).

The Q -factors were evaluated directly from the frequency response S_{11} measurements using the graphical Smith chart-based method described in Chapter II.

2. Theoretical SNR Gain Estimation

The SNR taking into account those effects can be expressed for a solenoidal type coil geometry as a function of the loaded (Q_ℓ) quality factor, the noise factor (F), the coil volume (V_c) and the filling factor (η) [139]:

$$SNR \propto \eta \cdot \sqrt{\frac{Q_\ell \cdot V_c}{F}} \quad (V-1)$$

The filling factor (η) is given by:

$$\eta = \frac{\int_{\text{sample}} |\mathbf{B}_\perp|^2 dV}{\int_{\text{coil}} |\mathbf{B}|^2 dV}, \quad (V-2)$$

where $|\mathbf{B}_\perp|^2$ is the energy density of the transverse magnetic field in the sample, $|\mathbf{B}|^2$ the total magnetic field energy density of the coil dV is the differential volume element. For a cylindrical coil with a radius R and length l coupled to a centered homogenous cylindrical sample with a radius r_s and length l_s , assuming a uniform magnetic field, the filling factor can be simplified to [139]:

$$\eta \approx \frac{r_s^2 \cdot l_s}{2 \cdot R^2 \cdot l}. \quad (V-3)$$

Therefore, the SNR gain between the two setups, using the same sample and different coils of the same length, becomes:

$$SNR_{\text{Gain}} = \frac{SNR_2}{SNR_1} \approx \frac{R_1}{R_2} \sqrt{\frac{Q_{\ell 2} F_1}{Q_{\ell 1} F_2}} \quad (V-4)$$

where R_1 is the radius of the reference coil, R_2 the radius of the comparison coil, $Q_{\ell 1}$ and $Q_{\ell 2}$ their loaded quality factors, and F_1 and F_2 the noise factors of their respective acquisition chains.

3. Acquisition Chain Noise Factor Estimation

The noise performances of both setups were characterized one at a time by calculating the noise factor F and the corresponding noise figure NF under loaded conditions. The noise factor is defined as the ratio of the SNR at the input and output of the acquisition chain:

$$F = \frac{\text{SNR}_{\text{input}}}{\text{SNR}_{\text{output}}} = \frac{(S_i/N_i)}{(S_o/N_o)} \quad (\text{V-5})$$

where S_i and S_o are the signal power density at the input and output, while N_i and N_o are the noise power density at the input and output, respectively.

The setups were connected to the receive chain one at a time. The noise power density at the output, N_o , was measured directly using an acquisition in the absence of any external signal to isolate the setup's intrinsic noise characteristics. Acquisition chain calibration data and numerical filters provided by the manufacturer were used to ensure the measurement accuracy. On the other hand, N_i was estimated as the theoretical minimum thermal noise power density for a resistive load at thermal equilibrium over the acquisition bandwidth BW . This simplifies F to:

$$F = \frac{N_o}{N_i} = \frac{N_o}{k_B \cdot T_{\text{sys}} \cdot BW} \quad (\text{V-6})$$

with k_B the Boltzmann constant and T_{sys} the absolute temperature of the system in Kelvin. It is important to note that this estimation method works only under the assumption of a perfect impedance matching and minimal signal loss at the input. While this approach provides simple and straightforward way to estimate the noise figure, it doesn't reflect real-world conditions as factors such as mismatched components, non-linear behaviors, and additional noise from active parts like amplifiers or mixers are not considered. However, in the case of this study, since the aim is ultimately to get an estimation of the SNR gain between both setups under the same conditions a more rigorous approach was deemed unnecessary as we assume that both measurements will be affected equally.

4. SNR Assesment Through Image Acquisition

To assess and compare the efficiency of both setups in imaging conditions, SNR measurements were performed on image acquisitions of the filled culture chambers.

A circular ROI incorporating most of the phantom while avoiding the edges was drawn over a uniform region of a coronal slice of the phantom to capture the signal, and four additional ROIs were drawn in the air regions, away from the phantom, to capture noise in areas where no object or structured noise at the corners of the image, to avoid phase-encoded ghosting or other artifacts that might propagate along the phase and frequency encoding directions. This placement ensured that the noise measurement captured random background noise.

When acquiring MRI data, in regions where the signal is weak, the noise can dominate, leading to a Rician distribution characterized by its non-negative nature and a peak at zero. This distribution reflects the statistical behavior of the signal where the noise influences the magnitude of the complex signal. In particularly low signal areas where SNR values are small, the Rician distribution of MRI data can be approximated by a standard Rayleigh distribution, while in areas where the SNR is relatively higher (>2) the Rician distribution of the data adopts a gaussian behavior. Due to the non-gaussian behavior when the signal is low, to be able to calculate accurately, a correction factor must be applied to adjust for the Rayleigh distribution[175].

To get a more accurate measurement of the overall SNR across the phantom, the acquisition process was repeated on N slices along the phantom and the mean SNR value was obtained.

$$\text{SNR} = \frac{\sum_{i=1}^N \left(\frac{\text{mean}(S_i)}{\text{std}(N_i)} \cdot \sqrt{\frac{4 - \pi}{2}} \right)}{N} \quad (\text{V-7})$$

5. Experimental Field Mapping

The field generated by the coils (B_1^+) in both setups were mapped using the gradient echo based Dual Angle Method (DAM) proposed by Insko and Bolinger in 1993 [176]. This method relied on the acquisition of two separate images with different excitation flip angles, α and 2α , while maintaining all other imaging parameters constant, under conditions where the repetition time (TR) is significantly longer than the longitudinal relaxation time (T1), the acquired signal intensity in each voxel is proportional to the sine of the local flip angle. The actual flip angle distribution, which is indicative of the spatial variation of the B_1^+ field, was computed using the standard DAM formulation:

$$\alpha(r) = \arccos\left(\frac{|I_2(r)|}{2|I_1(r)|}\right), \quad (\text{V-8})$$

where $I_1(r)$ and $I_2(r)$ represent the voxel-wise signal intensities corresponding to the prescribed flip angles α and 2α , respectively. This approach provides a direct estimation of the local flip angle, which is proportional to the effective B_1^+ field strength at each spatial location:

$$B_1^+ = \frac{\alpha}{\gamma\tau} \quad (\text{V-9})$$

where α is the flip angle, γ is the gyromagnetic ratio ($\gamma \approx 2.675 \times 10^8 \text{ rad} \cdot \text{s}^{-1} \cdot \text{T}^{-1}$), and τ the duration of the applied RF pulse, assuming constant B_1^+ amplitude over the pulse duration.

For the acquisition, the chamber was filled with a 16.3 mM Copper (II) Nitrate Trihydrate solution in water with an expected T1 of around 50ms at 0.3T. To ensure accurate B_1^+ mapping, the imaging sequence was implemented using a three-dimensional spoiled gradient echo acquisition with a long repetition time (TR = 300ms > 5T1) to minimize T1-related signal dependencies. The echo time (TE = 4ms) was chosen to minimize off-resonance effects while maintaining sufficient SNR. Classical flip angles $\alpha = 60^\circ$, $2\alpha = 120^\circ$ were selected and four averages were performed. The acquired data was processed using a voxel-wise intensity ratio calculation, followed by nonlinear fitting to extract the flip angle map. Regions with extreme signal dropout or insufficient SNR were excluded from further analysis to prevent artifacts in the final B_1^+ maps.

V.B. 3D Plastronics Setup Characterization Results

1. S-Parameters and Quality factor

The reflection coefficient, $|S_{11}|$, was measured to assess the impedance matching of the loaded coil and unloaded at the target frequency. Figure 82 shows the measured and simulated S_{11} magnitude and the corresponding Smith chart representation for the loaded coil.

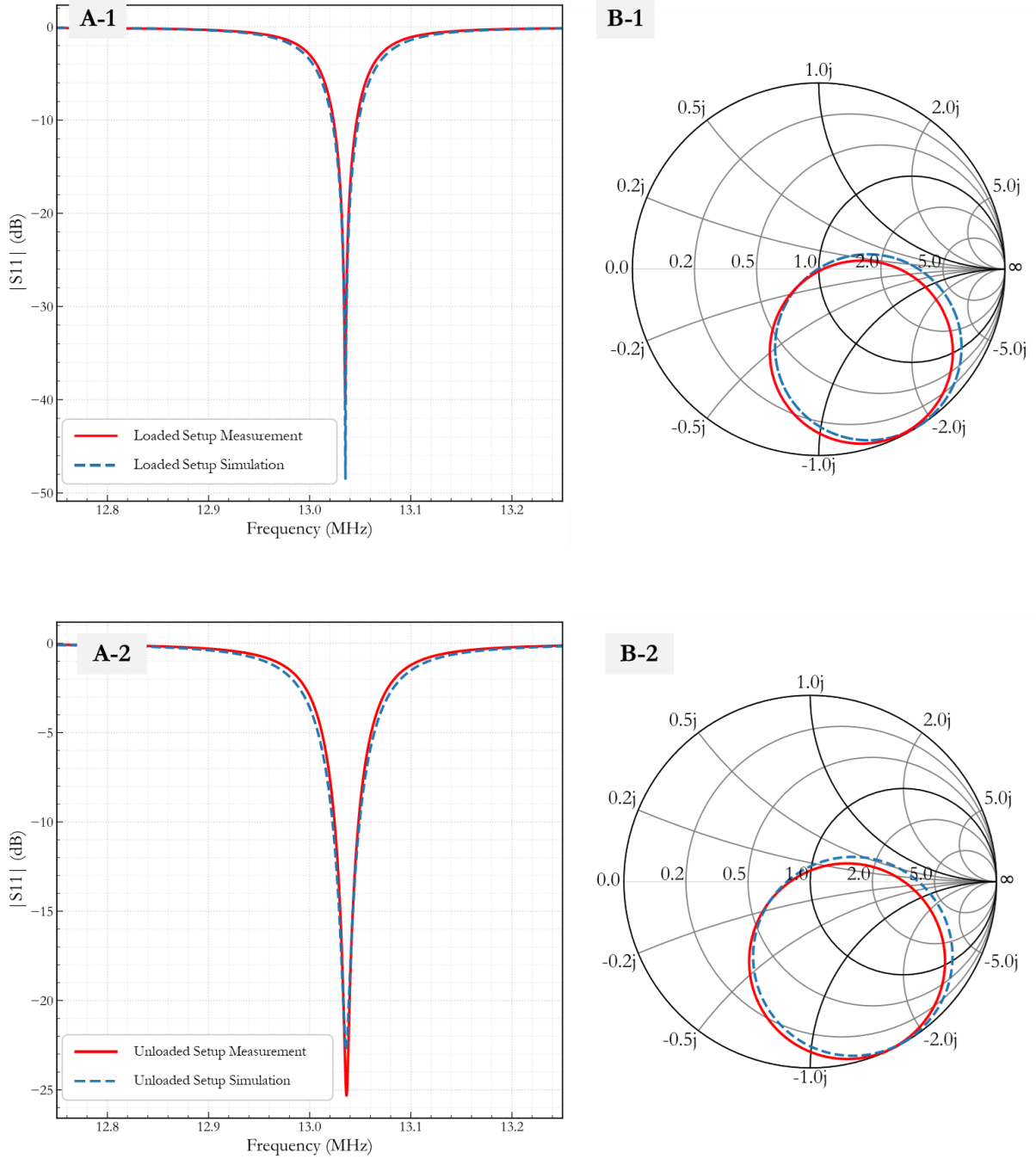


Figure 82: Measured and simulated S-parameter results for the loaded 3D Plastronics RF coil. (A) Magnitude of the reflection coefficient $|S_{11}|$ of the loaded (A-1) and unloaded (A-2) setup as a function of frequency, showing excellent agreement between simulation (blue dashed line) and measurement (red) near the tuned frequency of 13.1 MHz. (B) Corresponding Smith chart for the loaded (B-1) and unloaded (B-2) setups, illustrating the impedance matching behavior and confirming a close match between measured and simulated coil impedance.

As shown in Figure 82-A, there is excellent agreement between the measured (red) and simulated (blue dashed line) data, with both showing a sharp resonance dip at 13.035 MHz when loaded. The measured reflection coefficient at this frequency for the loaded setup was below -40 dB, indicating highly efficient power transfer and successful impedance matching to the 50 Ω system. The Smith charts in Figure 82-B further confirm the close match between the measured

and simulated impedances at resonance with a small phase shift probably due to the coax cable leading from through-panel SMA connector mounted on the shield to the SMA connector of the functionalized chamber (Figure 75-G) which was not modeled in the simulation.

From these S-parameter results, the quality factor was evaluated under both unloaded and loaded conditions to quantify the coil's energy storage efficiency and the effect of sample loading. The measured coil exhibited a higher loaded quality factor ($Q_\ell = 370$) than the simulated configuration ($Q_\ell = 327$), corresponding to a $\sim 13\%$ increase. The unloaded quality factors followed the same trend, with $Q_U = 382$ in measurement and $Q_U = 342$ in simulation ($\sim 12\%$ increase), indicating consistently lower losses in the fabricated coil than in the model. The ratio of unloaded to loaded Q was very similar in both cases (1.032 for measurement vs. 1.046 for simulation), suggesting that the relative contribution of sample loading to overall loss was small, and that the noticed improvements to the simulated model and measured setup most likely arose from overestimated coil and circuit losses.

2. B_1^+ Field Distribution and Homogeneity

The B_1^+ field generated by the coil was mapped experimentally using the dual-angle method and compared to the simulated field distribution. Figure 83 displays the measured and simulated B_1^+ amplitude maps in the central coronal plane.

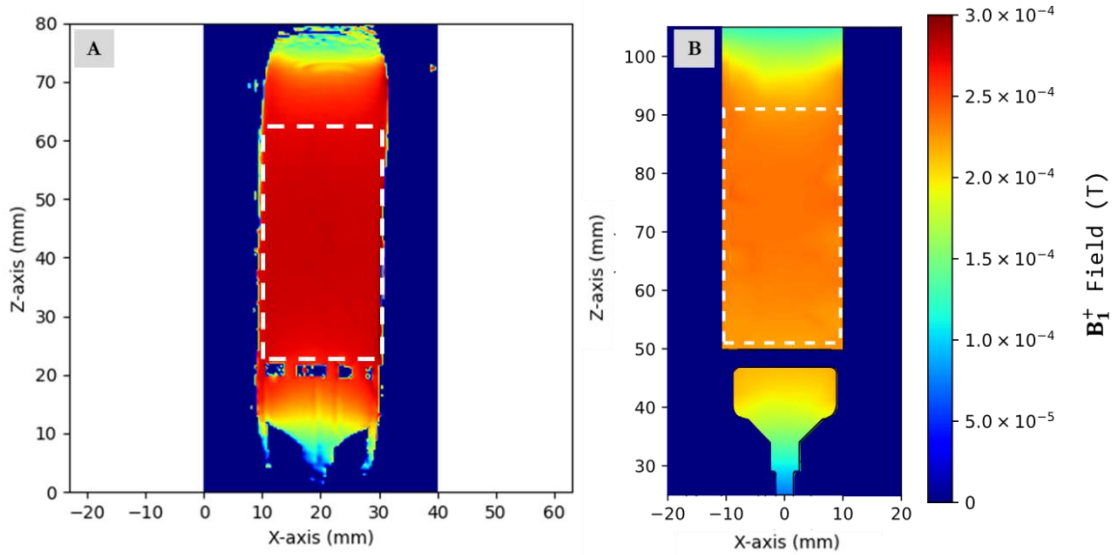


Figure 83: B_1^+ field maps for the 3D Plastronics RF coil. (A) Measured B_1^+ amplitude distribution in the central sagittal plane, showing uniform excitation within the defined region of interest (ROI, white dashed outline). (B) Simulated B_1^+ amplitude distribution in the same plane, computed under equivalent loading conditions with culture chamber shaped zero value mask overlay to indicate the chamber geometry. The simulated field pattern closely reproduces the measured profile, enabling quantitative comparison of coil efficiency and homogeneity.

Visually, the measured and simulated field patterns show strong agreement, with both demonstrating a relatively uniform field distribution throughout the designated region of interest. For a quantitative analysis of homogeneity, the coefficient of variation (CoV) was calculated within the ROI. The measured field was found to be slightly more homogeneous ($\text{CoV} = 1.53\%$) than the simulated field ($\text{CoV} = 2.07\%$), as detailed in Table 15.

3. Synthesis and Discussion

A comprehensive comparison of all measured and simulated performance parameters is provided in Table 15

	Loaded Q	Unloaded	CoV	ROI Mean B_1^+ Amplitude (T)
Simulation	327	342	2.07%	2.37×10^{-4}
Measurement	370	382	1.53%	2.79×10^{-4}

Table 15: Comparison of measured and simulated coil performance parameters. Loaded and unloaded quality factors (Q), coefficient of variation (CoV) within the ROI, and mean B_1^+ amplitude values are reported for both simulation and measurement. The measured Q and mean B_1^+ amplitude exceed simulated values, while the CoV is lower, indicating slightly improved homogeneity in practice.

The data confirms that the fabricated coil outperformed the simulation model across all key metrics. The higher measured Q-factor directly translates to a more efficient conversion of input power into the B_1^+ field. For a fixed input power, the B_1^+ amplitude scales approximately with the square root of the loaded Q:

$$\frac{B_{1,\text{meas}}^+}{B_{1,\text{sim}}^+} \approx \sqrt{\frac{Q_{\ell,\text{meas}}}{Q_{\ell,\text{sim}}}} = \sqrt{\frac{370}{327}} \approx 1.064 \quad (+6.4\% \text{ predicted}) \quad (\text{V-10})$$

Experimentally, however, the ROI mean B_1^+ amplitude increased by $\sim 18\%$ (from 2.37×10^{-4} T to 2.79×10^{-4} T) which is significantly higher than the predicted enhancement. This may be attributed to potential inaccuracies in the manufacturer-provided models of the discrete capacitors used in the simulations, which may not fully capture their real-world parasitic effects and loss characteristics as well as subtle differences in how the sample interacted with the RF field distribution (despite the overall loading magnitude being similar, as indicated by the loaded-to-unloaded Q ratios). Minor deviations between simulated and actual loss mechanisms, as well as the slightly improved field homogeneity observed in the measurements (lower CoV), likely also contributed.

While these deviations highlight the limits of purely numerical predictions, the agreement in scaling trends confirms that electromagnetic simulations remain a reliable tool for first-order performance estimation and design optimization, provided they are complemented with experimental feedback to account for real-world operating conditions.

V.C. 3D Application to High Resolution Scaffold Morphology Assessment.

To validate the functionality of the proposed setup, the full 3D *In Situ* bio-printing process followed by a high-resolution acquisition tailored to 3D scaffold morphological assessment was carried out.

1. Scaffold Geometry and *In Situ* 3D Printing Process

The imaging target was the $19.5 \times 19.5 \times 25$ mm cylindrical scaffold with 16 (2.5×2.5 mm) channels fitted to the chamber and sprinkler dimensions.

A Hydrogel-based construct was fabricated via extrusion-based 3D bioprinting using a composite bioink comprising 2% (w/v) fibrinogen (F8630-1G, Merck), 2.5% (w/v) porcine skin-derived gelatin (G1890, Sigma-Aldrich), and 2% (w/v) low-viscosity sodium alginate (A18565.36, Alfa Aesar) dissolved in calcium-free Dulbecco's Modified Eagle Medium (DMEM; 21068028, ThermoFisher). Printing was carried out at 21 °C using a COSMED333 bioprinter (TOBECA, France) equipped with a Vipro-head 3 extrusion system (VicoTec, Germany) and an 800 μm nozzle directly inside the sprinkler structure. Post-printing, structural integrity was achieved by crosslinking in a 270 mM calcium chloride bath.

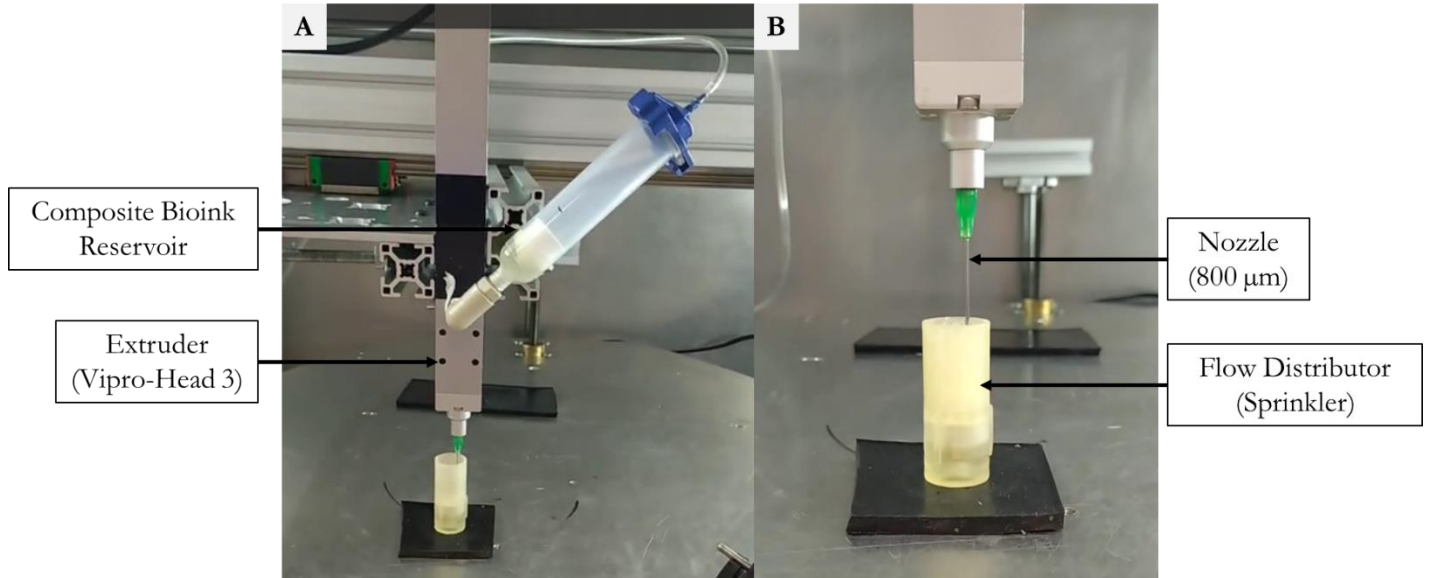


Figure 84: *In situ* extrusion-based 3D bioprinting of a hydrogel scaffold directly within the flow distributor (“sprinkler”) structure. (A) Bioprinter setup showing the syringe-based extrusion system. (B) Close-up of the printing process with the 800 μm nozzle aligned to the sprinkler.

2. High Resolution Imaging Results

Following scaffold fabrication, high-resolution imaging was performed using the custom-built 3D Plastronics setup. The sprinkler containing the 3D bio-printed scaffold was placed in the sealed culture chamber filled with 0.9% NaCl saline solution. A 3D GRE sequence (FA = 60°, TE = 4 ms, TR = 200 ms, AQ time = 19m 47s) was chosen to provide T1-weighted imaging with minimal T2* attenuation, to provide a good compromise between SNR and hydrogel structure contrast. The field of view was adapted to the culture chamber dimensions (30 $\times$ 30 $\times$ 30 mm³), with an isotropic voxel resolution of 234 μm .

Figure 85 presents representative images obtained from the scan: axial (A) and sagittal (B) slices highlight clearly defined scaffold boundaries and well-resolved macroporous channels, including occasional obstructions or trapped air bubbles. The 3D reconstruction (C), generated through segmentation of the acquired data, provides a full volumetric representation of the printed structure.

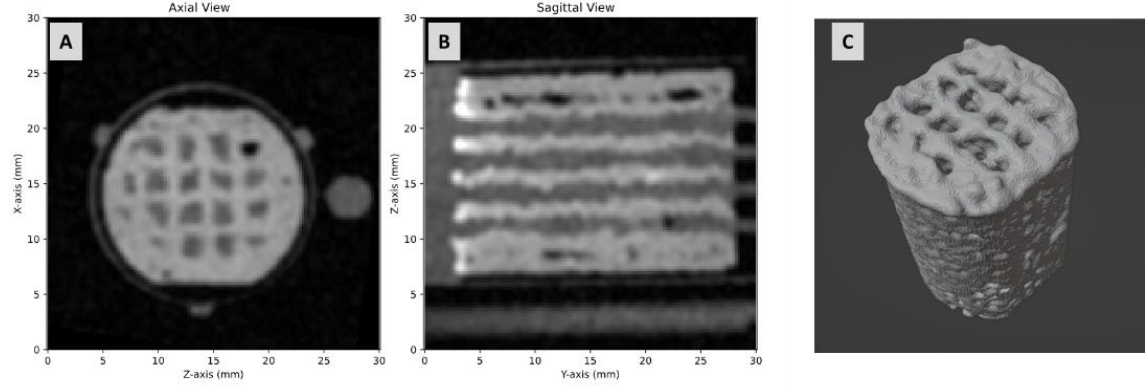


Figure 85: MRI of the in-situ 3D printed scaffold using a 3D Gradient Echo sequence (FA: 60° , TE: 4 ms, TR: 200 ms, voxel size: $234 \mu\text{m}^3$, FOV: $30 \times 30 \times 30 \text{ mm}^3$, AQ time = 19m 47s) with the scaffold immersed in 0.9% NaCl saline. (A) Axial and (B) sagittal slices show the internal structure of the scaffold with well-resolved macroporous channels. (C) 3D reconstruction obtained from segmented MRI data, illustrating the overall geometry and spatial fidelity of the printed construct.

3. Synthesis and Discussion

The fabricated 3D Plastronics setup enabled a high-resolution MRI acquisition lasting under 20 minutes on a 3D bio-printed scaffold placed within the culture chamber under perfusion. With a resolution of $234 \mu\text{m}^3$, imaging enabled clear visualization of the scaffold's internal channel structure and geometry. High image contrast and uniform signal intensity were maintained across all orientations. The scaffold remained mechanically and geometrically stable throughout the acquisition, with no visible motion or susceptibility-related artifacts. Both free-flowing and obstructed (air bubbles) channels were clearly visible with high enough contrast to be distinguishable from the 0.9% NaCl saline solution.

This kind of acquisition will be essential to enable non-destructive, quantitative assessment of scaffold morphology directly within the culture chamber, under realistic environmental and perfusion conditions. The combination of in situ fabrication and high-resolution imaging facilitates the detection of structural defects, occlusions, or deviations from the intended geometry without interrupting the culture process. Moreover, volumetric MRI datasets can be exploited for automated segmentation and morphometric analysis, providing metrics such as channel diameter, wall thickness, and porosity distribution with sub-millimeter precision. Such information is critical for correlating fabrication parameters with structural outcomes and, ultimately, for ensuring reproducibility and quality control in advanced tissue engineering applications.

V.D. Comparative Performance Against a More Classical Approach

In this Section, the two experimental setups were investigated. The first setup, consisted of the benchtop scanner combined with an all-purpose shielded reference coil provided by the manufacturer (Section-II.A) while the second setup, replaced the benchmark coil with the fabricated shielded device developed using the 3D Plastronics approach.

1. Q-factor and SNR Comparison

A comprehensive bench characterization was performed on both setups. Q-factors were measured under loaded conditions using a vector network analyzer and a graphical Smith chart

method. As summarized in Table 16, the commercial coil setup exhibited a (Q_ℓ) of 322 while the 3D Plastronics setup showed a slightly higher loaded Q-factor of 370.

To assess imaging performance, noise factor (F) measurements were obtained, and theoretical SNR gains were calculated according to equation (V-4) using the measured Q-factors and coil radii. As presented in Table 16, the commercial coil setup had a measured F of 2.01 for a radius of 42 mm, while the 3D Plastronics setup exhibited a lower F of 1.48 with a reduced coil radius of 34 mm, yielding a theoretical SNR gain of around 55%. These calculations are consistent with the measured SNR gain of 60 %, demonstrating that the integrated 3D Plastronics design not only improves electromagnetic coupling thanks to a higher filling factor but also reduces intrinsic noise thanks to an apparent higher shielding efficiency evident through the noise factors, resulting in superior imaging sensitivity. Figure 86 displays representative gradient echo (GRE) acquisitions for each setup with the different noise and signal ROIs, clearly showing the enhanced SNR in the 3D Plastronics configuration. The improved performance in the 3D Plastronics setup is attributable to both the compact integration of the RF coil directly onto the culture chamber surface, optimized coil-to-sample matching and the RF shielding efficiency.

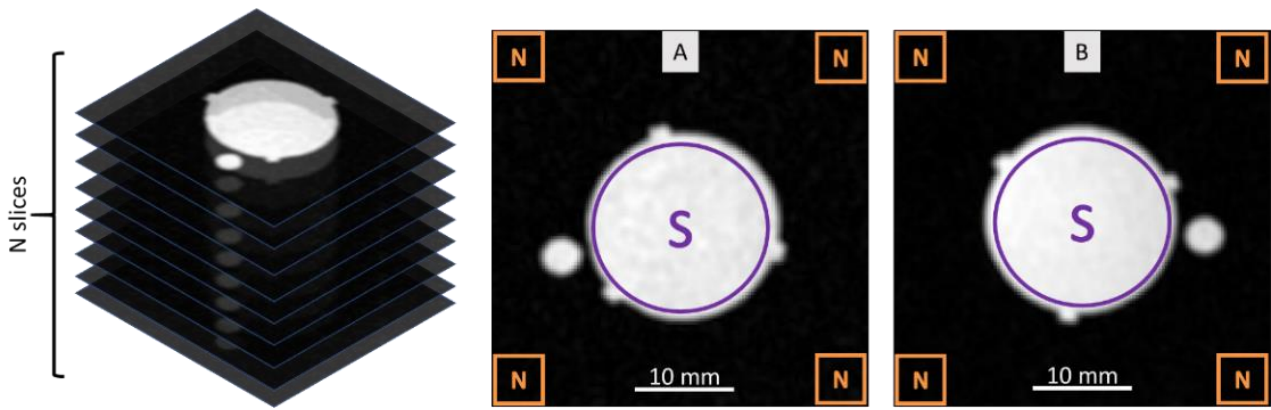


Figure 86: 16.3 mM Copper (II) Nitrate Trihydrate solution filled chambers GRE acquisitions (FA = 30°, TE = 4ms, TR=12ms, AQ time = 3m 17s) for SNR measurements of A) the commercial coil setup and B) the 3D Plastronics setup.

	Measured Noise Factor (F)	Measured loaded Q-factors (Q_ℓ)	Theoretical SNR Gain from Measured Loaded Q-factors	Mean Measured SNR from Images	Measured SNR Gain from Images
Commercial Coil Setup	2.01	322	~55 %	~ 68	~ 60%
3D Plastronics Setup	1.48	370		~ 109	

Table 16: Theoretical SNR Gain Calculated from the simulated and measured loaded Q-factors the coil dimensions and the noise factors (F).

2. B_1^+ Field Mapping and Homogeneity

The B_1^+ field distribution within the culture chambers was mapped using a dual-angle method based on the DAM approach (Figure 87) and compared. For both setups, the spatial variation of the effective flip angle was calculated on a voxel-by-voxel basis under identical excitation conditions. In The commercial coil setup, the mean B_1^+ field within the region of interest (ROI) was determined to be 1.78×10^{-4} (in Tesla) with a standard deviation of 3.82×10^{-6} , resulting in a coefficient of variation (CoV) of 2.15%. By contrast, the 3D Plastronics setup achieved a higher mean B_1^+ (2.79×10^{-4}) with a standard deviation of 4.26×10^{-6} , yielding a lower CoV of 1.53%.

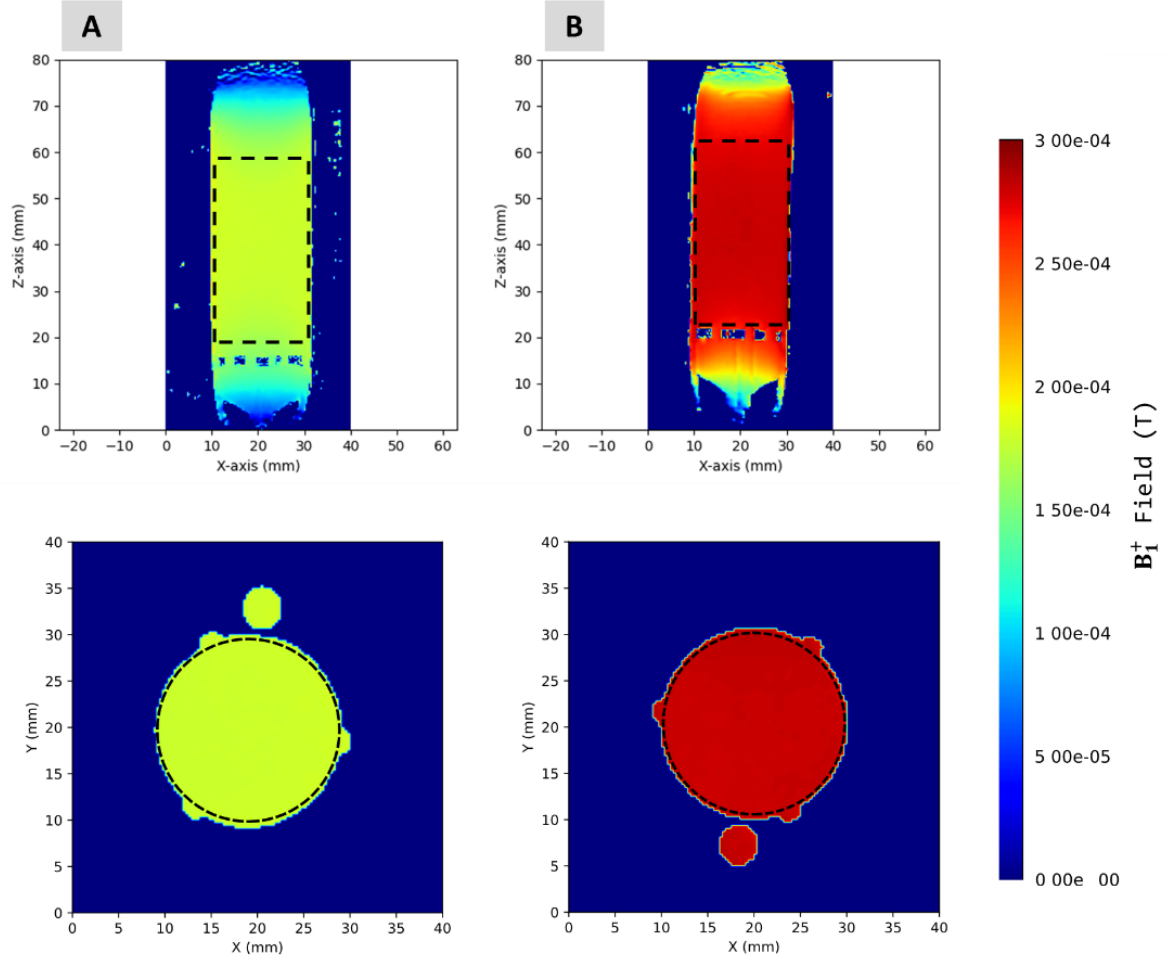


Figure 87 : Field maps obtained using the dual-angle mapping (DAM) technique for (A) the commercial coil setup and (B) the 3D Plastronics setup, with both axial and coronal slices shown for each case. Cylindrical regions of interest (ROIs), representing the maximum sample dimensions accommodated by the chambers, are outlined using dashed black lines.

These measurements indicate that the 3D Plastronics design not only enhances the absolute field strength, thereby directly contributing to improved SNR, but also improves field homogeneity across the sample volume. The more uniform B_1^+ field in the 3D Plastronics setup is essential for achieving consistent excitation and signal reception, particularly when imaging complex 3D tissue constructs and performing quantitative measurements such as ADC and Velocimetry. The direct integration of the RF coil onto the culture chamber ensures sample positioning and reduces edge effects, further contributing to the observed homogeneity.

3. Synthesis and Discussion

It has been demonstrated here that integrating a custom-designed RF coil fabricated using 3D Plastronics techniques directly onto the culture chamber yields some benefits over a conventional shielded coil system. In our comparative evaluation of two experimental setups, the 3D Plastronics approach proved advantageous in several critical areas.

First, the ability of 3D Plastronics to integrate the RF coil closer to the region of interest, resulted in an overall measured SNR gain of 60% and slightly improved $\mathbf{B}_1^+$ field homogeneity. This was particularly evident in the $\mathbf{B}_1^+$ field maps derived from the dual-angle method, which showed both increased mean $\mathbf{B}_1^+$ amplitude (from 1.78×10^{-4} T to 2.79×10^{-4} T) and reduced coefficient of variation (from 2.15% to 1.53%) in the 3D Plastronics setup compared to the commercial alternative.

The measured SNR gain is consistent with the theoretical predictions derived from our SNR model (55%), where improvements in quality factors, effective volume coverage, and reduced noise factor contribute to the observed sensitivity enhancement. The major source of the SNR improvement is the enhanced filling factor, achieved by reducing the coil's inner radius from 21 mm to 17 mm, which brings the sensor into closer proximity with the sample. The decision was made in this chamber design to have the return path of the fluidic system go through the integrated coil and to use the 13-capacitor configuration presented in Chapter II. Incorporating a return path flowing outside the integrated coil in future designs should allow for greater coil sample proximity (~ 12 mm inner radius and 58 mm length) and could yield even better coil sensitivity up to roughly 120% with the 13-capacitor design (based on estimated Q-factor from numerical simulations) or even more with a 20-capacitor design. The improved homogeneity stems from the tighter, more conformal fit and positioning, while minimizing edge effects by ensuring that the coil length is sufficiently larger than its diameter [104] with a form factor $FF \approx 0.53$ compared to $FF \approx 0.65$ for the manufacturer provided coil.

Furthermore, the 3D Plastronics setup exhibited a slightly higher loaded Q-factor (370 vs. 322) and a significantly lower noise factor (1.48 vs. 2.01), indicating more efficient energy storage and superior shielding against external RF interference. Both of these aspects can be attributed to more efficient shielding. The Q-factor difference stems from the larger shielded area compared to the coil dimensions in the 3D Plastronics setup ($178 \times 52 \times 218$ mm for a 34 mm diameter coil) as opposed to the commercial setup ($90 \times 51 \times 170$ mm for a 42 mm diameter coil), significantly reducing shield proximity effects. The 3D-printed housing in the 3D Plastronics setup enabled better coil coverage. The chamber was almost completely closed off, leaving only a small gap at the bottom to accommodate fluidic tubing.

Ultimately, these findings confirm that the 3D Plastronics approach represents a viable and advantageous alternative to conventional benchtop MRI coil configurations, particularly for applications requiring close integration with fluidic culture chambers. The combination of enhanced filling factor, improved shielding efficiency, and optimized coil-sample matching translated into measurable gains in sensitivity, field homogeneity, and noise suppression. As stated previously, this improvement is especially relevant in preclinical and tissue-engineering contexts, where high SNR and uniform excitation are essential for resolving fine structural details and obtaining quantitative MR metrics. Looking ahead, further optimization of coil geometry and chamber design offers the potential for even greater sensitivity gains, paving the way for next-generation, application-specific MRI hardware solutions enabled by 3D Plastronics.

When compared to other efforts in low-field MRI for tissue engineering, our setup demonstrates competitive performance. For instance, while Nitzsche et al.[21] successfully used a 0.5T benchtop system to characterize scaffolds, their approach relied on standard coils and 128 single scan averages, resulting in a complete measuring time of about 90 min. An SNR gain of 60%

using a 3D Plastronics setup similar to the one proposed in this work would help address the long acquisition times: an image requiring 90 minutes with the original coil could be acquired in just ~35 minutes with equivalent quality. In the original study, scaffolds were transferred from culture plates to NMR tubes for imaging, risking mechanical disruption, contamination, or misalignment. Their setup required interrupting cultivation to perform imaging with no perfusion compatibility during the imaging process, making it impossible to assess nutrient transport and cell proliferation under growth conditions. A 3D Plastronics integrated culture-imaging chamber could enable non-invasive, in-situ monitoring without interrupting sterile conditions while improving reproducibility, maintaining precise spatial alignment, and supporting perfusion-based culture, allowing imaging under physiologically relevant conditions. Our work builds on previous 3D Plastronics coils by Gerges et al.[120] and Gnanago et al.[130] by fully integrating the coil with a perfusion chamber, creating a single, optimized device specifically for *in situ* tissue monitoring.

VI. Conclusion and Discussion

This work demonstrates the feasibility and benefits of a fully integrated 3D Plastronics-based MRI setup tailored for imaging perfused, bio-printed constructs. By embedding conformal RF coils directly onto the culture chamber and incorporating local shielding within a modular mechanical design, the system achieved improved B_1^+ field homogeneity, higher SNR, and reproducible imaging performance under low-field conditions. The setup's ability to capture detailed internal scaffold structures under perfusion offers significant value for non-destructive assessment in tissue engineering.

Beyond improved imaging performance, a 3D Plastronics-based MRI setup offers practical advantages in terms of design flexibility and rapid prototyping. The fabrication process enables the direct integration of both RF coils and customized fluidic access points, such as perfusion inlets and outlets, into the body of the culture chamber. This would allow each system to be tailored to specific scaffold geometries, flow requirements, and experimental conditions, improving reproducibility and adaptability. By embedding the coil directly within the printed structure, the design avoids common limitations of externally mounted coils, including suboptimal alignment, increased distance from the sample, and degraded electromagnetic coupling.

The versatility of the 3D Plastronics process extends beyond RF coils. As demonstrated in other studies using similar fabrication techniques, it is possible to embed a variety of sensors (temperature, pH [160], [161], etc...), or even piezo-based actuators [177] for Magnetic Resonance Elastography (MRE) and studies. This would transform the setup into a multi-modal monitoring platform, providing real-time physiological data alongside anatomical MRI scans, offering a more holistic view of tissue development.

Moreover, this fabrication approach is inherently scalable. The digital design workflow (CAD, CFD and EM Analysis to 3D printing and functionalization) allows for rapid iteration and customization of the culture chamber and coil geometry to accommodate different scaffold sizes, cell types, or even integration with different benchtop MRI systems. Unlike conventional manufacturing that requires expensive and time-consuming retooling, new designs can be printed and metallized in a matter of days, making this technology highly adaptable for diverse research needs and paving the way for standardized, application-specific MR-compatible culture chambers.

Furthermore, while the shielding effectively reduces noise, it also limits physical access to the chamber and requires manual assembly steps that can slow down repeated experiments. However, further improvements in shield geometry and design could help overcome this challenge. The current need in the proposed setup to screw and unscrew the enclosure can be addressed in future iterations by adopting a mechanical latching system. To address the cumbersome nature of the current passive shield further, alternative strategies can also be explored (i.e., active shielding using secondary coils to cancel out external interference or self-shielded coil topologies with patterns

designed to inherently confine the magnetic field [73], [74], [75], [76]). This new approach would significantly improve user access and experimental throughput.

Overall, these results highlight the potential of 3D Plastronics as a scalable and customizable solution for embedding imaging capabilities into bioengineered systems, making it a very interesting solution to address the sensitivity limitations of a low-field benchtop setup.

Chapter IV: A Cryogenic Solution for Benchtop MRI Scanners - Proof of Concept

Chapter Overview

This chapter explores a different strategy for overcoming the inherent Signal-to-Noise Ratio (SNR) limitations of low-field benchtop MRI based on cryogenic cooling of the radiofrequency (RF) coil. Building upon the geometric optimizations of previous chapters, this section investigates a physics-based approach to directly reduce the primary source of system noise. The motivation is to achieve a substantial sensitivity enhancement that could unlock even higher-resolution imaging for demanding tissue engineering applications.

A review of cryogenic technologies positions cooled copper coils as a pragmatic and cost-effective alternative to more complex High-Temperature Superconductor (HTS) systems, aligning with the thesis's goal of developing accessible laboratory tools. While HTS coils offer theoretically higher performance, they present significant challenges in cost, fabrication, and operational bandwidth that make them less suitable for versatile benchtop setups.

The chapter then delves into the fundamental principles of cryogenically cooled copper conductors. It explains how reducing the coil's temperature to that of liquid nitrogen drastically lowers its electrical resistivity, which in turn increases the quality factor (Q) and directly reduces the thermal noise power. The analysis highlights that this approach is particularly effective in the "coil-noise-dominated" regime characteristic of small-sample imaging at low fields, where the coil itself is the primary performance bottleneck.

The discussion then pivots to the practical engineering challenges of implementing such a system. The core problem is designing a low-cost, easy-to-fabricate cryostat that can effectively cool the RF coil while ensuring the biological sample remains thermally insulated and viable at room temperature. This introduces trade-offs between cooling efficiency, thermal insulation, material selection, and the mechanical integration of the system within the constrained space of the benchtop scanner.

Finally, the chapter consolidates these points by presenting the design, simulation, fabrication, and experimental validation of a novel, proof-of-concept cryogenic setup. It details a low-cost cryostat built from accessible materials and demonstrates a significant, measured SNR gain upon cooling. This work serves to validate cryogenic cooling as a powerful and achievable strategy for boosting the performance of benchtop MRI. Furthermore, it lays the groundwork for future refinements, such as leveraging 3D Ceratronics, an extension of 3D Plastronics to ceramic substrates—to create coils with integrated microfluidic cooling channels for superior thermal management and even greater performance gains.

I. Introduction:

While the preceding chapter focused on enhancing SNR through the geometric optimization of the radiofrequency (RF) coil with 3D Plastronics, this chapter explores a complementary, physics-based strategy involving cryogenic cooling. This approach shifts the focus from maximizing the filling factor to directly reducing the primary source of signal degradation in the coil-noise-dominated regime, the intrinsic thermal noise of the coil conductor itself. As established for small-sample, low-field MRI, the coil's own electrical resistance is often the limiting factor for sensitivity. The principle that cryogenic cooling can yield substantial SNR improvements has been successfully studied and implemented in various NMR and MRI applications [100], [178], [179], [180]. By lowering the coil's physical temperature, it is possible to significantly decrease its electrical resistance, thereby boosting the coil's quality factor (Q) and fundamentally improving the achievable SNR. The motivation behind this strategy is to achieve a sensitivity enhancement that could unlock even higher-resolution imaging for demanding tissue engineering applications.

The objective of this chapter is to design, fabricate, and validate a novel, proof-of-concept cryogenic system to demonstrate the feasibility of this approach and to quantify the achievable enhancement in SNR when cooling. While other cryogenic technologies exist [100], [180], cryogenically cooled copper coils offer a robust, inexpensive, and versatile solution that aligns with the overarching goal of this thesis which aims to develop accessible, high-performance laboratory tools. This approach provides a reliable balance of performance, cost, and ease of fabrication, making it an appealing solution for a proof-of-concept investigation.

The core of this work lies in addressing the central engineering challenge by designing a low-cost cryostat able to cool the RF coil to near-liquid nitrogen temperatures (≈ 77 K) while simultaneously ensuring the biological sample remains thermally insulated and viable at room temperature (≈ 293 K). This dual requirement of effective cooling and robust thermal isolation is particularly demanding, as it must be met within the significant physical constraints of the benchtop scanner, where every millimeter of space is critical. The design must balance the need for high-performance thermal insulation with the goal of maintaining a good filling factor, which is essential for maximizing the final SNR.

To achieve this, the chapter will first delve into the theoretical principles governing the performance of cryogenically cooled copper conductors, explaining the relationship between temperature, resistance, and coil noise. Following this, it will detail the design, simulation, and practical implementation of the novel cryostat, emphasizing accessible materials and fabrication methods. The investigation will then proceed with an experimental characterization of the system's cooling performance and its impact on the RF coil's electrical properties, including its quality factor and resonant frequency. The chapter will culminate in a direct, image-based measurement of the SNR gain achieved, providing a quantitative assessment of this strategy's potential for high-performance benchtop imaging.

II. Theoretical Background: Cryogenically Cooled Coils for Improved SNR

II.A. Electrical Properties of Cryogenic Copper

The electrical resistivity of copper depends on temperature due to phonon scattering and impurities [181]. The total resistivity can be expressed as:

$$\rho(T) = \rho_0 + \rho_{ph}(T), \quad (II-1)$$

where ρ_0 is the temperature-independent residual resistivity due to impurities, and $\rho_{\text{ph}}(T)$ is the temperature-dependent phonon contribution. The Residual Resistivity Ratio (RRR) is given by [181]:

$$\text{RRR} = \frac{\rho(300 \text{ K})}{\rho(4 \text{ K})} = 1 + \frac{\rho_{\text{ph}}(300 \text{ K})}{\rho_0}. \quad (\text{II-2})$$

Solving for ρ_0 gives:

$$\rho_0 = \frac{\rho_{\text{ph}}(300 \text{ K})}{\text{RRR} - 1}. \quad (\text{II-3})$$

The phonon contribution is modeled using the Bloch–Grüneisen formula [181]:

$$\rho_{\text{ph}}(T) = A \left(\frac{T}{\Theta_D} \right)^5 \int_0^{\Theta_D/T} \frac{x^5 e^x}{(e^x - 1)^2} dx, \quad (\text{II-4})$$

where $\Theta_D = 343 \text{ K}$ is the Debye temperature for copper. The constant A is determined using the known resistivity of high-purity copper at $T = 300 \text{ K}$ [181]:

$$\rho_{\text{ph}}(300 \text{ K}) = 1.68 \times 10^{-8} \Omega \cdot \text{m}. \quad (\text{II-5})$$

Finally, electrical conductivity is given by:

$$\sigma(T) = \frac{1}{\rho(T)}. \quad (\text{II-6})$$

Figure 88 shows how electrical resistivity varies with temperature for different RRR values. Higher RRR values correspond to higher purity copper, which exhibits significantly higher conductivity at low temperatures due to reduced impurity scattering. For each RRR values, plateaus appear at a certain temperature showcasing the limitations of improvements using cryogenically cooled cooper coils. From to copper resistivity curves in Figure 88, it can inferred that even with cooling, the noise reduction in copper is limited by the intrinsic properties of the material, reaching a temperature-independent plateau defined by the residual resistance ratio (RRR) and anomalous skin effect limiting the decrease of the electric resistance at low temperature. This is expected to restrict the ultimate performance of cryogenic copper coils at lower temperatures [180]. However, since the target temperature for this current proof of concept implementation is 77K or above, cryogenically cooled copper conductors should still yield significant improvements even for low purity (RRR=10) materials.

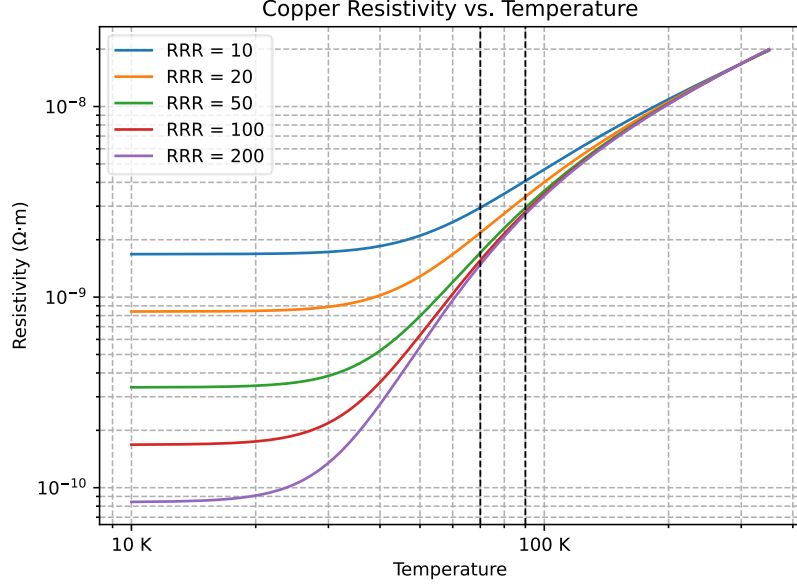


Figure 88: Copper Resistivity vs. Temperature for various RRR values. The graph illustrates how electrical resistivity decreases as temperature drops, with higher purity copper (higher RRR) showing a more significant drop in resistivity at cryogenic temperatures due to reduced impurity scattering. The performance enhancement plateaus are clearly visible at lower temperatures.

II.B. Impact on Coil Performance and SNR

The reduction in copper resistivity at cryogenic temperatures has a direct impact on coil losses, expressed through the effective coil resistance R_c under RF excitation:

$$R_c \propto \sqrt{\frac{\pi f \mu_0}{\sigma(T)}}, \quad (\text{II-7})$$

where f is the operating frequency, μ_0 is the permeability of free space, and $\sigma(T)$ is the conductivity. As $\sigma(T)$ increases with cooling, R_c decreases, leading to a reduction in ohmic losses.

This improves both the unloaded quality factor (Q_u) and the loaded quality factor (Q_ℓ) of the resonator:

$$Q_u = \frac{\omega_0 L_{eff}}{R_c}, \quad Q_\ell = \frac{\omega_0 L_{eff}}{R_c + R_s} \quad (\text{II-8})$$

where ω_0 is the angular resonance frequency, L_{eff} is the effective inductance, R_c is the intrinsic coil resistance and R_s the sample induced resistance.

With reduced R_c at cryogenic temperatures, Q_u is expected to increase substantially while the potential increase in Q_ℓ will depend entirely on the relative value of R_s . The magnitude of Q_ℓ improvement when cooling depends on the loading regime. Specifically, in the case where ($R_s \gg R_c$), Q_ℓ is effectively clamped by R_s , and cooling produces only a marginal increase. In contrast, when coil losses are comparable to or exceed sample losses, lowering R_c substantially raises both Q_u and Q_ℓ ,

Also, SNR can be expressed as [179]:

$$\text{SNR}_{\text{coil}} \propto \sqrt{\frac{Q_\ell}{T_s(1 - Q_\ell/Q_u) + T_c Q_\ell/Q_u}} \quad (\text{II-9})$$

where T_s and T_c are the sample and coil temperatures respectively. This highlights that the balance between coil and sample losses, expressed by the ratio $Q_\ell/Q_u = R_c/(R_c + R_s)$, determines the efficacy of cooling and that the impact of temperature goes beyond simple resistivity reduction but also noise repartitioning.

When Q_ℓ/Q_u is small (sample noise dominant), the effective noise temperature is governed by T_s and little benefit is obtained by lowering T_c . Conversely, when Q_ℓ/Q_u approaches unity, both the increase in Q_ℓ and the reduction in T_c translate directly into SNR gains.

1. Sample Noise Dominance

In most high-field MRI systems, the dominant noise source arises from the sample. Sample-induced losses are proportional to their conductivity σ_s , permittivity ϵ_s , and the coil geometry. The effective resistance contribution of the sample, R_s , can be approximated as:

$$R_s \propto \int_V \sigma_s |E|^2 dV, \quad (\text{II-10})$$

where E is the electric field within the sample volume V . In this regime, cooling the coil only marginally improves SNR, since sample noise dominates. This situation applies especially to human imaging at clinical field strengths. However, using low-field benchtop MRI for small sample imaging using small coils, sample noise dominance is almost never achieved due to the relatively low sample conductivity and reduced dielectric losses at lower frequencies, as well as limited sample volume, which together render the coil resistance R_c the prevailing contributor to system noise.

2. Coil Noise Dominance

At low fields and with small samples, the sample losses are comparatively small, and the coil conductor noise becomes limiting. In this regime, cryogenic cooling is highly beneficial. The SNR can also be written as [109]:

$$\text{SNR}_{\text{coil}} = \frac{\text{SNR}_{\text{lossless coil}}}{\sqrt{\text{NPR}}}. \quad (\text{II-11})$$

The relative contributions of coil and sample noise can be quantified using the Noise Power Ratio (NPR):

$$\text{NPR} = 1 + \frac{R_c T_c}{R_s T_s}, \quad (\text{II-12})$$

where R_c, T_c are the coil resistance and effective temperature, and R_s, T_s are those of the sample. Cooling reduces R_c and T_c , significantly lowering coil noise. As a result, SNR improves in proportion to the square root of the resistance reduction, provided sample noise remains negligible.

In benchtop MRI for tissue engineering, where samples are small and weakly conductive, the coil-noise-dominated regime applies. This has been showcased in chapters II and III where most coils in this regime presented loading factors (Q_u/Q_ℓ), in the best cases, for highly loaded scenarios, around or inferior to 2, and in the worst cases, where loading was suboptimal, closer to 1. Cryogenically cooled copper coils therefore represent an effective strategy to maximize SNR and enable high-resolution imaging.

III. Design of the cryo-cooled MRI probe

III.A. Proposed Basic Concept

Early in the project, a simple cryogenic solution was proposed. The proposed cryogenic approach relied on a system engineered for low-cost and effective cooling, centered around a nested dewar-like structure. Figure 89 shows a simplified schematic of the working principle of the proposed approach.

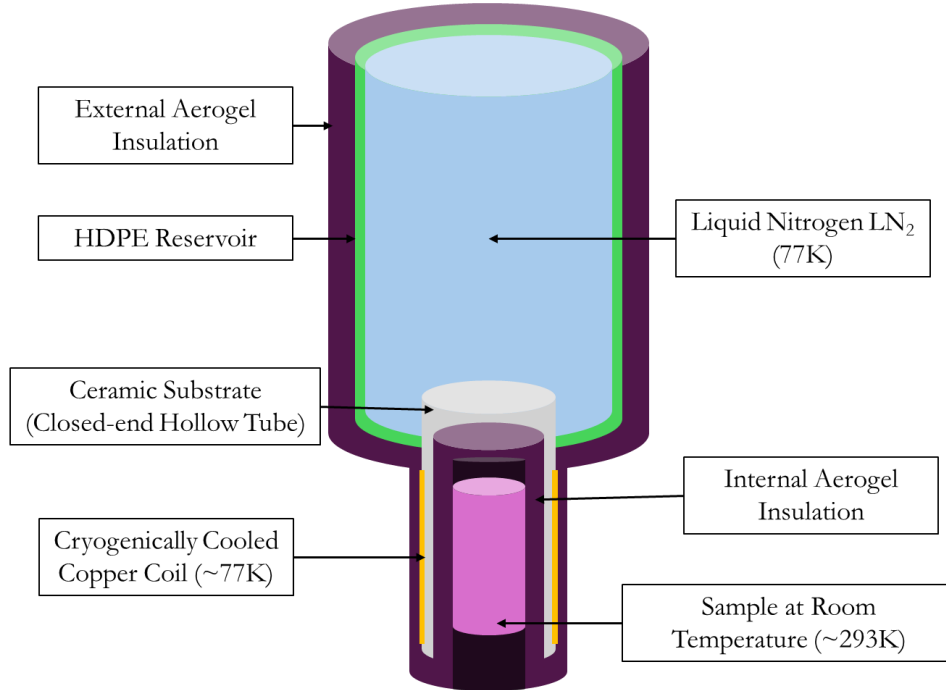


Figure 89: Schematic of the working principle of the proposed cryogenic setup, designed for low-cost coil cooling.

In the original design, most of the insulation was to be performed using high-performance Aerogel (Spaceloft Aerogel, a flexible, fibre-reinforced silica aerogel blanket insulation by Aspen Aerogels)[182]. This material would encase the entire apparatus (reservoir, substrate/temperature buffer and cryogenically cooled RF coil), providing thermal insulation barrier to maintain the internal cryogenic temperatures.

Within this insulation, the main cryogen container, a High-Density Polyethylene (HDPE) reservoir, designed to hold a substantial volume of Liquid Nitrogen (LN₂), would act as the primary cold source. HDPE was selected as a cost-effective and readily available liquid nitrogen compatible material.

A ceramic substrate plunged into the reservoir to create as much contact with the cold source as possible would serve a dual purpose. First, it would act as a precise temperature transfer

mechanism, from the cold source to the copper coil. Second, it would function as a buffer to prevent bubbling related artifacts.

The coil would be deposited on the outer surface of the ceramic substrate and be cooled by thermal conduction while samples would be insulated from the cold source using an internal layer of aerogel material.

IV. EM Modeling of a Cryogenically Cooled BLGR Design

As a preliminary validation of the proposed cryogenic concept, and to predict the potential performance enhancements before committing to fabrication, an electromagnetic (EM) simulation study was conducted. These simulations were designed to model the behavior of a Bridged Loop-Gap Resonator (BLGR) coil under cryogenic conditions, to quantify the expected improvements in quality factor (Q) and signal-to-noise ratio (SNR) and analyze temperature-dependent effects such as frequency shifts.

IV.A. Materials and Methods

1. Computational Framework

Electromagnetic simulations were performed using CST Microwave Studio (Dassault Systèmes, France). The Frequency Domain Solver, based on the Finite Element Method (FEM), was selected for its high accuracy and computational efficiency in analyzing electrically small, high-Q resonant structures like the one presented within a small frequency range. A co-simulation approach was employed, integrating the 3D full-wave simulation of the coil and its environment with a circuit simulation (CST Design Studio) to accurately model the behavior of the lumped-element and tuning/matching network.

2. Coil Selection and Sample Geometry

As presented in Chapter II, the bridged loop gap resonator (BLGR) has been proven to be the coil topology of choice for homogenous large (~ 40 mm) tissue engineering construct applications. Under optimal loading conditions in the proposed simulation setup, this topology has shown to be able to reach the limits of the coil noise dominance with a loading factor $Q_u/Q_\ell = 2$ for the 13-capacitor variant. However, in Chapter III, a practical implementation of a similar coil, with a more realistic loading scheme, has produced loading factors closer to 1 ($Q_u/Q_\ell \approx 1.04$). As it will be impossible to implement a cryogenically cooled coil setup without some kind of insulation between the coil and the sample, the regime at room temperature is expected to be squarely in the coil noise dominance domain as the distance between the coil and the sample will have to increase to accommodate insulation (negatively impacting the filling factor).

Due to these considerations, a 13-capacitor BLGR was selected for the cryogenically cooled setup.

- **The Coil:** The coil was modeled as a $\text{Ø}42 \times 65$ mm BLGR with an added “T” protrusion for realistic matching circuitry.
- **The Sample:** The sample was modeled as a $\text{Ø}20 \times 65$ mm cylinder spanning the entire coil length
- **The Substrate:** A 42 mm OD, 35 mm ID and 150 mm long cylindrical alumina substrate was selected as a former which double as the bore.

Figure 90 shows the selected geometries and the coil model setup.

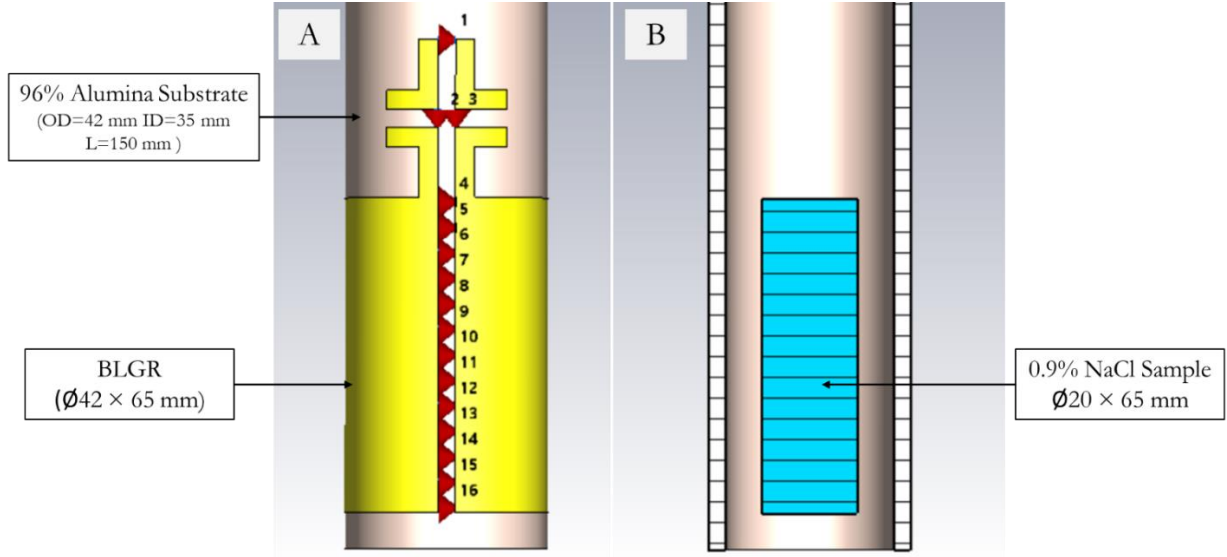


Figure 90: Coil and Sample Geometry for Cryogenic Simulation

3. Material Properties

3.1. Alumina Substrate

The alumina substrate was modeled using the built in CST MWS material library. The “96% Alumina (lossy)” model was selected with a $\epsilon_r = 9.4$ for a loss tangent $\tan \delta = 0.0004$.

3.2. Cryogenically Cooled Copper

Discrete copper resistivity values were calculated using the theoretical approach presented in section II.A and used as input in CST MWS for RRR=20 which is a typical value for standard engineering copper used in MRI [100], [180].

3.3. Sample

The sample was defined as a 0.9% NaCl saline was modeled with a permittivity $\epsilon_r = 79.2$ for a loss tangent $\tan \delta = 25.25$ and a conductivity $\sigma = 1.45$ S/m (values obtained using the Stogryn [151] and Klein & Swift [152] models for saline water).

3.4. Lumped Elements

High Q non-magnetic MLCCs from Knowles Syfer were selected and each element was modeled using manufacturer-provided S-parameter data, obtained from the CAPCAD library [154], to ensure accurate representation of their frequency-dependent characteristics. The use of Class I ceramic dielectrics, such as C0G/NP0, suggests these capacitors will remain marginally stable under cryogenic conditions due to their low temperature coefficient.

However, a degree of skepticism is warranted. The manufacturer's specified temperature rating for these components is typically from -55°C to $+125^\circ\text{C}$, and operation at cryogenic temperatures is outside this guaranteed range. The temperature coefficient for C0G/NP0 is rated at 0 ± 30 ppm/ $^\circ\text{C}$ over the specified range, which would theoretically lead to a capacitance drift of approximately 1% over the full cryogenic temperature sweep from room temperature to 70K. The S-parameter data used for modeling was obtained at room temperature and does not account for potential physical changes, such as thermal contraction, that could unpredictably alter the component's performance. Therefore, the true cryogenic performance of these components is unverified and should be viewed with skepticism until empirically tested.

IV.B. Simulation Results

The simulation results quantify the performance enhancement of the bridged loop gap resonator (BLGR) when subjected to cryogenic cooling. The analysis focuses on the frequency response, quality factors, and the operational noise domain.

1. Frequency Response and Quality Factor

The simulated S-parameter (S_{11}) frequency response of the BLGR coil is presented in Figure 91-A. A significant improvement in the coil's performance is observed when cooled from room temperature (305 K) to liquid nitrogen temperature (77 K). As shown in Figure 91-A, the loaded quality factor (Q_ℓ) increased from 342 at 305 K to 484 at 77 K, representing a 41.5% improvement. This enhancement is a direct consequence of the reduced electrical resistance of the copper conductor at cryogenic temperatures.

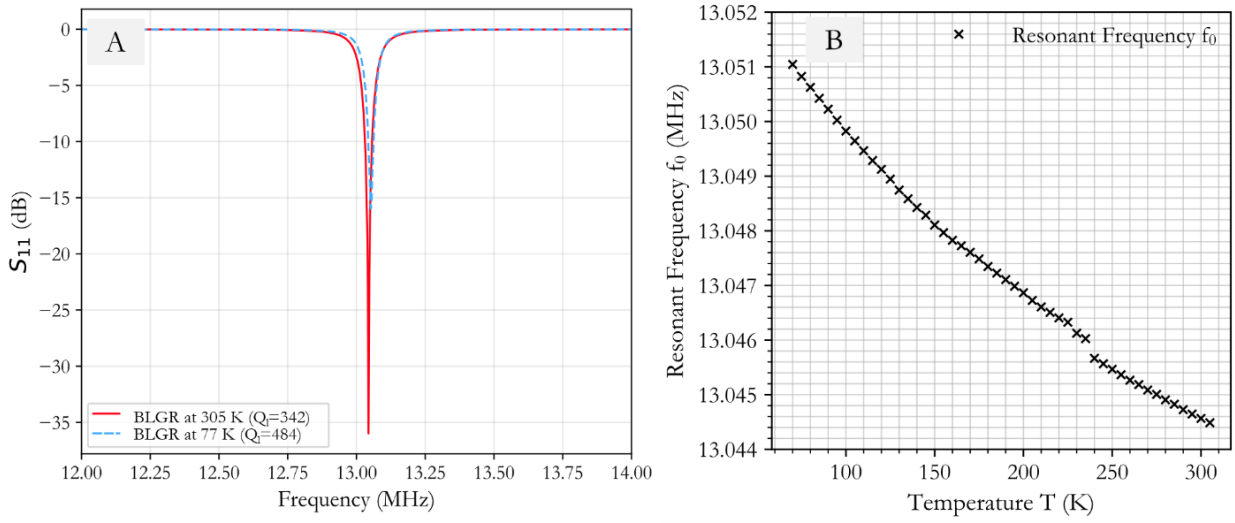


Figure 91:: Effect of Temperature on Coil Frequency Response

Figure 91-B illustrates the relationship between the resonant frequency and temperature. The resonant frequency exhibits a nearly linear upward drift as the temperature decreases, shifting from 13.044 MHz at 305 K to 13.051 MHz at 77 K. This shift of approximately 6 kHz is likely attributable to a decrease in the coil's inductance caused by the changing skin effect at cryogenic temperatures. This upward drift corresponds to $\approx 0.05\%$ on 13 MHz and originates from the reduction of the internal component of inductance as the copper resistivity decreases. At higher conductivity, the skin depth $\delta = \sqrt{2\rho/(\mu_0\omega)}$ becomes smaller, confining currents nearer to the surface and slightly lowering the magnetic energy stored inside the conductor. With geometry held constant, this loss of internal field energy reduces the total inductance by $\approx 10^{-3}$, yielding the observed frequency shift.

2. Noise Dominance Domains

A more detailed analysis of the quality factors across the entire temperature range is provided in Figure 92. Figure 92-A plots both the unloaded (Q_u) and loaded (Q_ℓ) quality factors as a function of temperature. An increase in both Q values is observed as the temperature is lowered. The unloaded Q factor improves from 362 to 573, while the loaded Q factor increases from 342 to 484 over the temperature sweep from 305 K to 77 K.

The noise dominance regime was evaluated using the coil noise figure defined as:

$$NF_{coil} = 10 \log_{10}(NPR) = 10 \log_{10} \left(1 + \frac{T_c/T_s}{(Q_u/Q_l - 1)} \right). \quad (IV-1)$$

Figure 92-B shows the coil noise figure for different temperatures. NF_{coil} decreases from around 12.3 at room temperature to 3.6 at 77 K. Since the system remains in the coil-noise dominant regime ($NF_{coil} > 3$) throughout the entire temperature range, the reduction in coil resistance had a more substantial impact than any changes in sample-induced losses. This result confirms that cryogenic cooling is an effective strategy for enhancing the signal-to-noise ratio (SNR) for this specific coil and sample configuration.

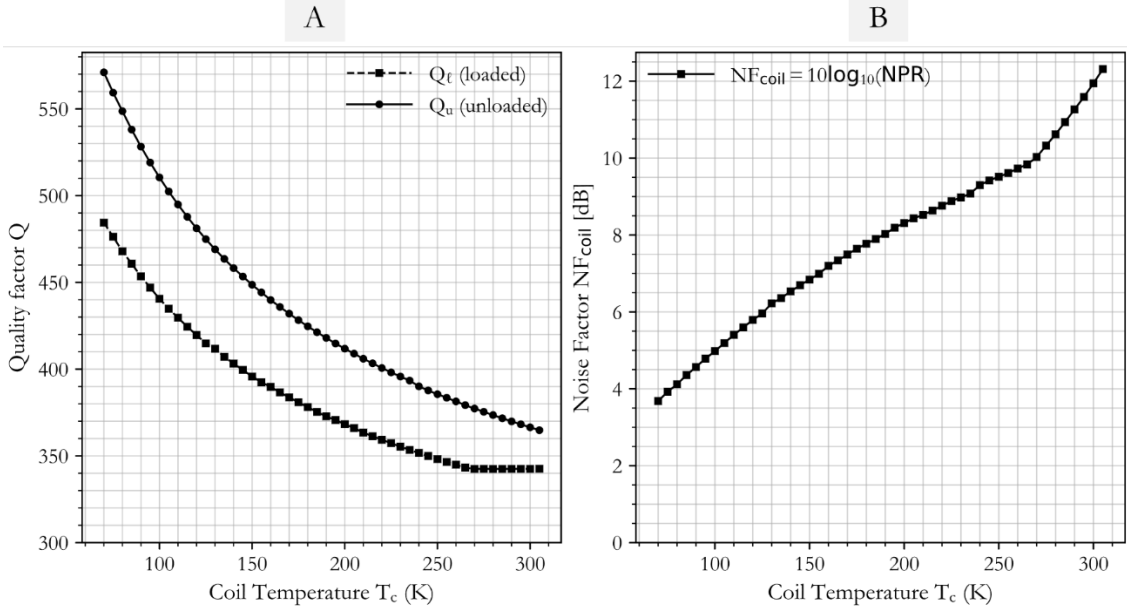


Figure 92: (A) Simulated Q_u and Q_l . (B) Coil Noise Figure vs. Temperature.

3. Expected SNR Gain from Simulation

The primary motivation for employing cryogenic cooling is the potential for significant enhancement in SNR. Using Equation (II-9), the simulated SNR gain as a function of the coil temperature (T_c), while keeping the sample at a constant room temperature (T_s), is calculated and presented in Figure 93.

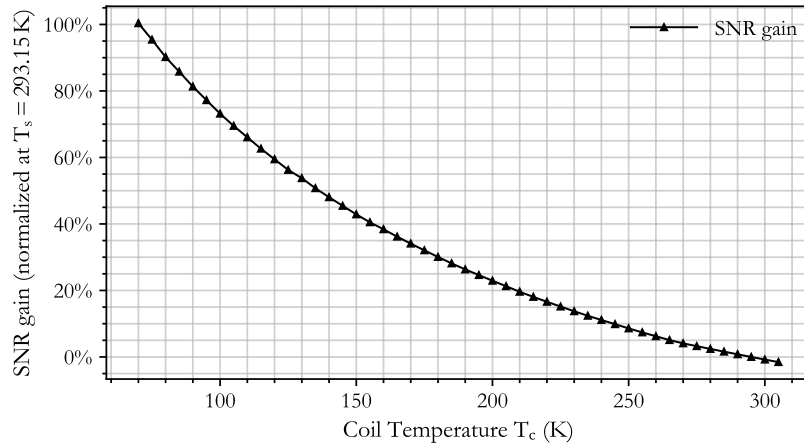


Figure 93: Simulated SNR gain vs the coil temperature (T_c). The SNR gain is normalized to the value at room temperature, with the sample temperature ($T_s=293.15$ K).

The simulation predicts a substantial improvement as the coil is cooled. The results, normalized to the SNR at room temperature, show that reducing the coil temperature from 305 K to liquid nitrogen temperature (77 K) is expected to yield an SNR gain of approximately 100%. In other words, the SNR is predicted to double. This enhancement is a direct consequence of reducing the thermal noise generated by the coil's electrical resistance. This result is consistent with the system operating firmly within the coil-noise dominant regime ($Q_u/Q_\ell < 2$), where improvements in the coil's quality factor directly translate to significant gains in overall system sensitivity.

V. Practical Cryogenic Coil Setup Implementation

The practical implementation of the cryogenic system evolved significantly from the original concept. The final design, shown in Figure 94, is a more sophisticated apparatus that separates the cryogen reservoir from the radio frequency (RF) coil assembly, employing a more “indirect” cooling method. All pieces aside from the inner reservoir, the insulation, the alumina substrate and transfer line were designed to be 3D printed using either a FDM printer or a SLA resin printer.

V.A. Cryostat Design

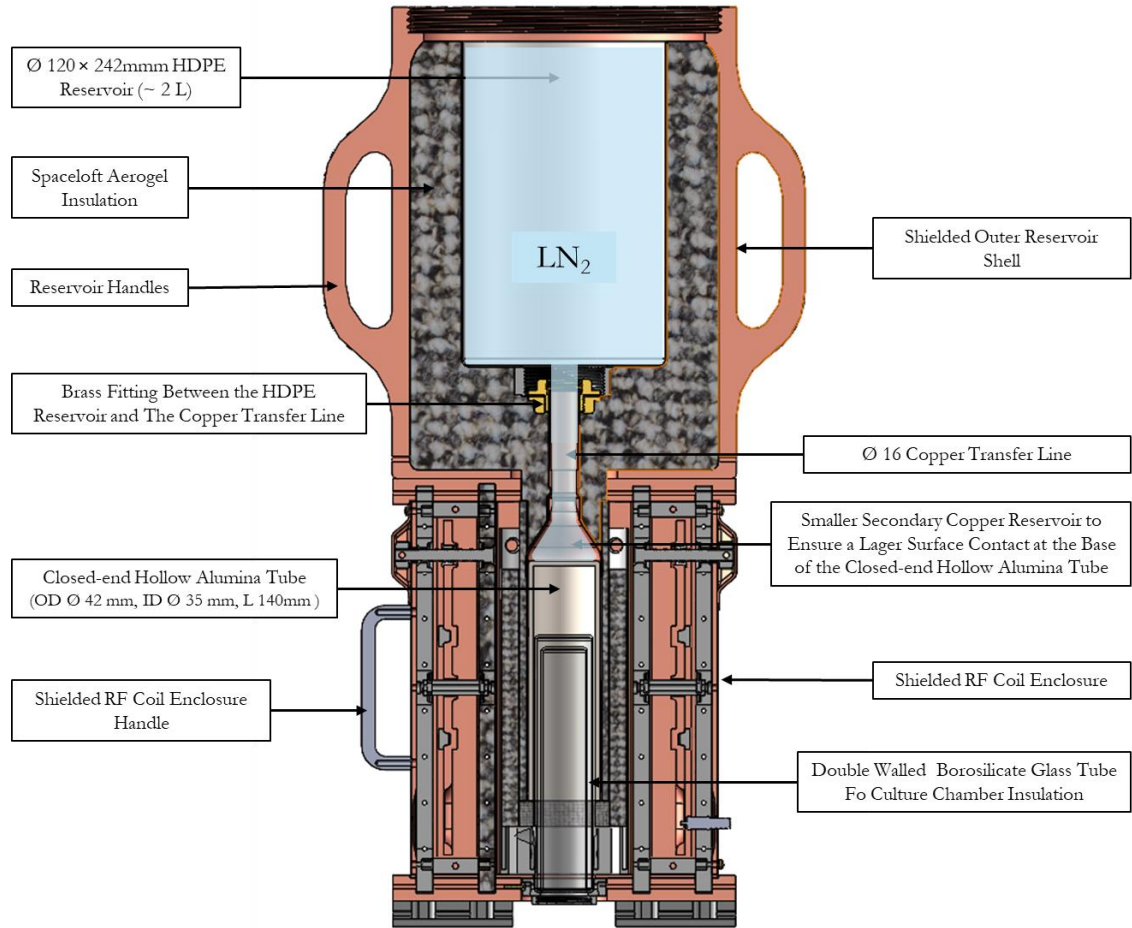


Figure 94: Cross-sectional view of the proposed cryostat design.

The core of the system is the integrated cryostat and shielded enclosure. Unlike the original proposal where the coil substrate was directly immersed in liquid nitrogen, this design utilizes a dedicated heat transfer path. A 2-liter High-Density Polyethylene (HDPE) reservoir, insulated with Spaceloft Aerogel (nominal thermal conductivity $\lambda \approx 0.015$ W/m·K, density ≈ 150 kg/m³, roll

thickness 5 mm)[182], holds the liquid nitrogen (LN_2). This entire reservoir assembly is housed within a shielded outer reservoir shell to prevent external electromagnetic interference (EMI) from coupling to the copper transfer line, which could otherwise feed noise into the RF coil enclosure.

From the reservoir, a brass fitting connects to a $\varnothing$ 16 mm copper transfer line. This line channels the LN_2 to a smaller secondary copper reservoir located at the base of the closed-end hollow alumina tube that serves as the coil substrate. This secondary reservoir is engineered to ensure a large and stable surface contact area with the alumina tube, providing consistent and efficient cooling via thermal conduction. This indirect approach allows the coil to be placed slightly further from the reservoir and ensure central positioning in the magnet, with the added bonus of preventing potential instabilities and electrical noise that could arise from the coil being in proximity to a large pot of boiling nitrogen.

The sample chamber is thermally isolated from the cryogenic coil by a double-walled borosilicate glass tube, which functions as a vacuum-insulated dewar. The double wall glass measures 35 mm in external diameter, and 28 mm of internal diameter for a length of 150 mm with an inter wall gap of 2.5 mm and a distance to the alumina substrate of 1 mm. This is a fundamental departure from the internal aerogel insulation initially proposed and is expected to provide superior thermal separation within the limited available space inside the alumina bore, ensuring the sample remains at room temperature. The choice was made to deviate from the originally proposed approach because as early tests were performed, the Spaceloft Aerogel could not provide adequate insulation while maintaining a decent bore opening due to the required thickness (>5 mm) for an effective thermal barrier. Figure 95 shows a closeup of the sample insulation mechanism

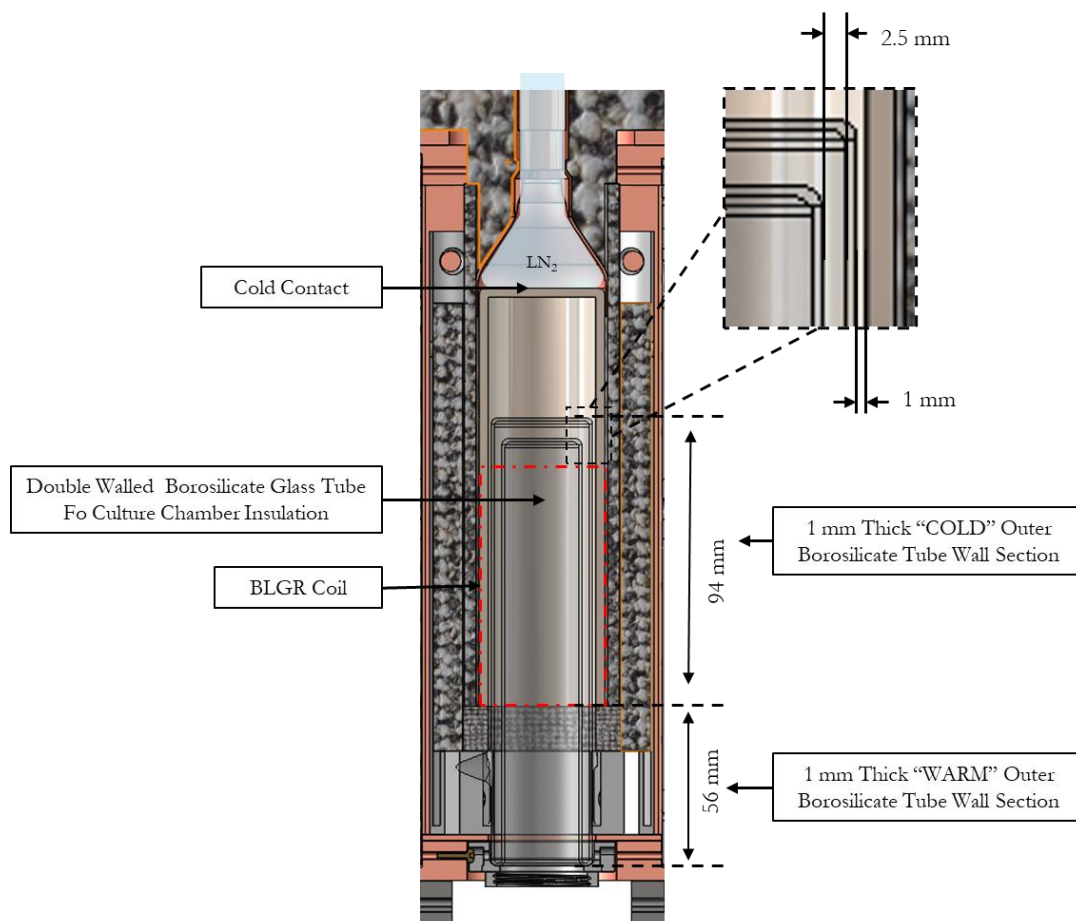


Figure 95: Close up of the sample insulation mechanism

V.B. Integrated RF Coil

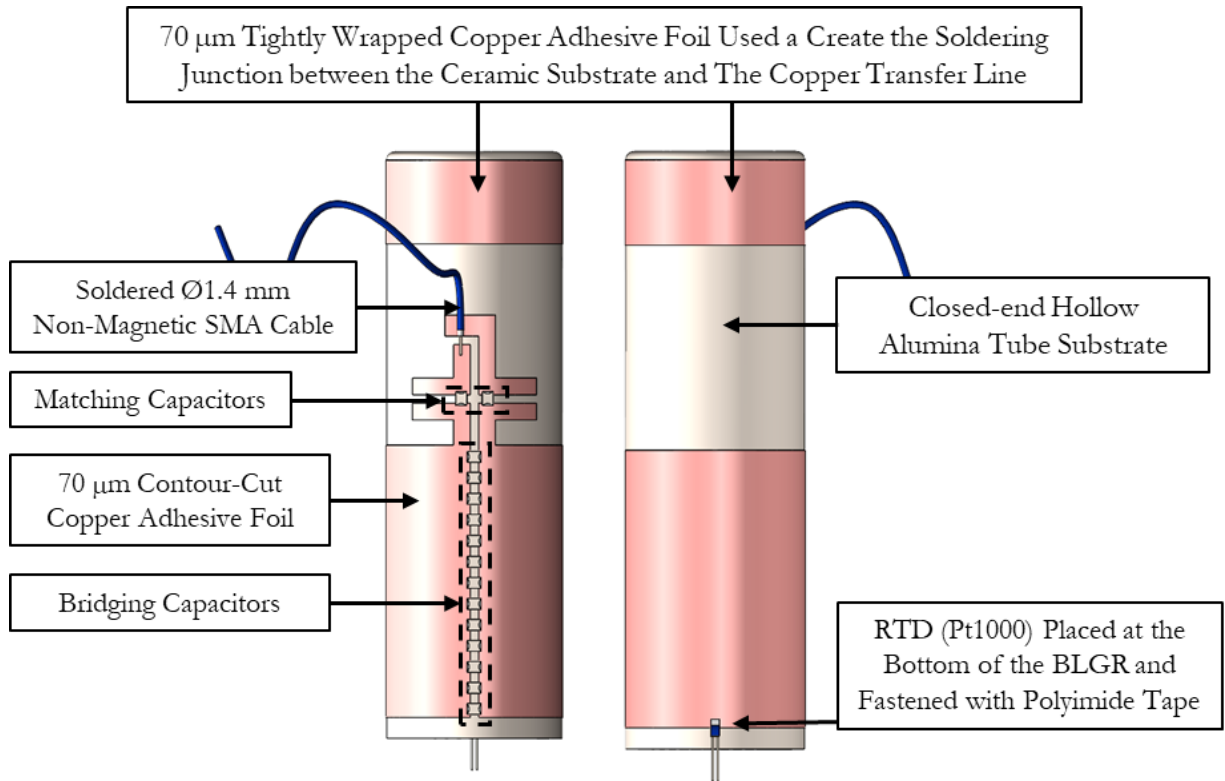


Figure 96: Detailed schematic of the radio frequency (RF) cryo-coil assembly. It shows the 70 μm contour-cut copper foil applied to the hollow alumina tube, with integrated matching and bridging capacitors. Also shown are the soldered non-magnetic SMA cable for signal transmission and the PT1000 RTD for temperature monitoring.

The details of the RF coil assembly are illustrated in Figure 96. A 70 μm thick, contour-cut copper adhesive foil is applied to the outer surface of the $\varnothing 42$ mm closed-end hollow alumina tube. This construction forms the RF coil. The necessary matching and bridging capacitors are integrated directly onto the substrate. A non-magnetic SMA cable is soldered to the feed point to transmit the RF signal.

To facilitate assembly with the transfer line, a 70 μm tightly wrapped copper adhesive foil was used to create the soldering junction pad to enable soldering.

To monitor the coil's temperature accurately, a PT1000 Resistance Temperature Detector (RTD), rated for cryogenic temperatures of up to -200°C , is fastened at the base of the alumina substrate using Polyimide Tape. The entire coil and cooling assembly is housed within a robust shielded enclosure, which provides electrical shielding and structural support.

V.C. Shielded Enclosure with Incorporated Tuning Mechanism

Description of the tuning mechanism: the shielded box dimensions are 180 mm in x, 55 mm in z and 220 mm in y. The tuning mechanism was built to reduce the size of the shielded enclosure by varying the dimensions on the y axis and changing the coil inductance. To do so, copper clad tuning plates were added on each side with flex polyimide linking them to the nearest side panel to form a closed conductive loop around the coil creating an internal box. To allow the internal

box to “shrink” on the x axis, a sliding mechanism was added to each plate and actioned by tuning knobs on both sides. When not in use, these tuning knobs are covered with a 3D printed copper clad cover to keep optimal EMI shielding.

For the new shield configuration tailored to this cryogenic setup, the shielded enclosure serves a dual role:

1. The suppression of electromagnetic interference (EMI) to preserve the intrinsic signal-to-noise ratio (SNR) of the coil,
2. An integrated tuning mechanism.

Similarly to chapter III, the external enclosure was constructed from copper-clad panels, forming a continuous Faraday cage around the coil assembly which ensures that environmental electromagnetic noise is effectively attenuated.

To reconcile the often-conflicting requirements of shielding integrity and resonance tuning flexibility and avoid adding extra trimmer capacitors inside the hard to access environment, an internal reconfigurable geometry was engineered. Following the same initial design, the enclosure dimensions were set to 180 mm (x), 55 mm (z), and 220 mm (y). However, here, tuning was achieved by incorporating sliding copper-clad plates on the x-axis, coupled via flexible polyimide copper laminates to the adjacent side walls. This configuration created a secondary, variable-sized conductive box around the coil, effectively modifying the inductive and capacitive environment. By reducing or enlarging the effective volume of this internal shield, the resonance frequency of the RF coil could be shifted without having to modify the coil components or using a trimmer.

Mechanically, the plates were mounted on sliding rails, actuated externally via SLA 3D printed precision knobs located on both sides of the enclosure. These knobs transferred motion through non-magnetic shafts, enabling fine, reproducible adjustments under operating conditions. When not in use, the knobs were sealed by 3D-printed copper-clad covers, preserving the EMI shielding. Figure 98 shows a detailed presentation of the proposed tuning mechanism and its components.

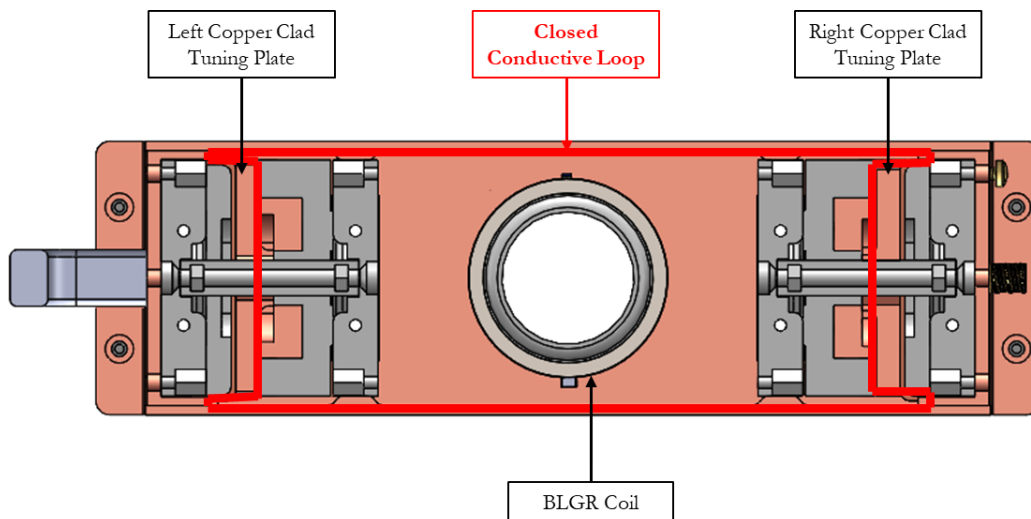


Figure 97: Detailed top view of the shielded enclosure illustrating the proposed tuning mechanism and highlighting the closed conductive loop created by the enclosure (in red)

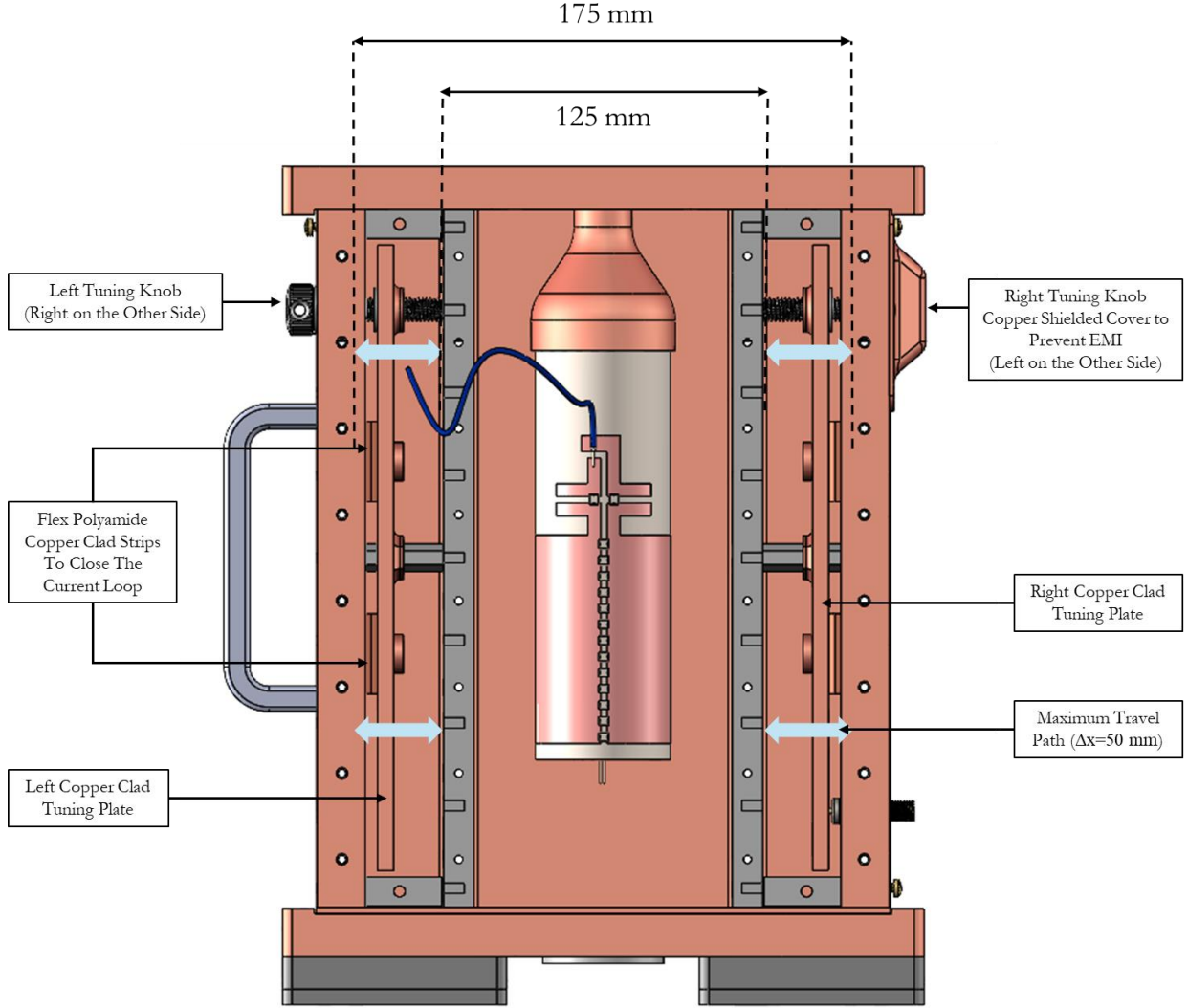


Figure 98: A detailed side view of the shielded enclosure, illustrating the proposed tuning mechanism. Symmetrically opposed copper-clad plates are moved via external knobs to adjust the coil's inductance.

1. Scientific Principle of the Tuning Mechanism

The operating principle of the tuning mechanism is based on the perturbation of the electromagnetic boundary conditions surrounding the coil. The resonance frequency of the RF resonator is governed by the general relation:

$$\omega_0 = \frac{1}{\sqrt{L_{\text{eff}}C_{\text{eff}}}}, \quad (\text{V-1})$$

where L_{eff} is the effective inductance of the coil and C_{eff} the effective capacitance of the surrounding environment and integrated matching network.

The presence of nearby conductive boundaries modifies both L_{eff} and C_{eff} . When copper-clad tuning plates are moved closer to the coil, image currents are induced, leading to a redistribution of magnetic flux lines. This reduces the net magnetic energy stored in free space, effectively decreasing L_{eff} :

$$L_{\text{eff}} = \frac{2W_m}{I^2}, \quad (\text{V-2})$$

where W_m is the total stored magnetic energy and I the coil current. A closer conductive boundary lowers W_m , reducing L_{eff} .

In parallel, the tuning plates increase the capacitive coupling to ground. The effective capacitance can be expressed as:

$$C_{\text{eff}} = C_{\text{coil}} + C_{\text{parasitic}}, \quad (\text{V-3})$$

where $C_{\text{parasitic}}$ arises from the proximity of the coil to the metallic surfaces. As the plates move inward, $C_{\text{parasitic}}$ increases due to the reduced separation, following approximately:

$$C_{\text{parasitic}} \propto \frac{\epsilon_0 A}{d}, \quad (\text{V-4})$$

with A the effective plate vis à vis area and d the distance to the conductive boundary.

The combined effect of reduced L_{eff} and increased C_{eff} results in a measurable shift of the resonance frequency ω_0 . Depending on the initial coil configuration, the dominant effect can either raise or lower ω_0 . In our particular case, the selected BLGR resonator requires large capacitor values to reach the desired resonant frequency as the coil inductance is very small (in the ~ 15 nH range). This means, with the distance from the plates ($\sim 1.5 \times$ the coil diameter), this resonator topology will be more strongly affected by the variation of inductance rather than the variation in capacitance, leading to an inductive dominant effect and an increase in resonant frequency.

The incorporation of such a boundary-based tuning mechanism is particularly advantageous in cryogenic MRI coils. After cooldown, mechanical reconfiguration of the tuning capacitors near or on the resonator or replacement of capacitors is impractical. The sliding plate system provides an external, non-invasive means of adjusting resonance frequency while maintaining shielding integrity. Although this method can be great for tuning, it might disturb the matching therefore, other methods will have to be investigated to complete the enclosure by complementing its design in future iterations.

2. Tuning Range Predictions

To have an idea of the potential tuning range of the proposed tuning mechanism, the copper clad plates connected to the enclosure were added to the simulation setup presented in Section IV. Both the tuning range and the impact on the coil quality factor were investigated.

Figure 99 presents the simulated reflection coefficient (S_{11}) of the coil for two distinct separations of the tuning plates: $\Delta x = 175$ mm (maximum separation) and $\Delta x = 125$ mm. The results demonstrate a significant shift in the resonant frequency as the plates are moved. Adjusting the plates over this 50 mm travel path shifts the resonance from approximately 12.98 MHz to 13.18 MHz. This corresponds to a continuous tuning range of about 200 kHz, which is more than sufficient to compensate for ~ 10 kHz predicted frequency drifts caused by cooling or compensate variations in sample loading.

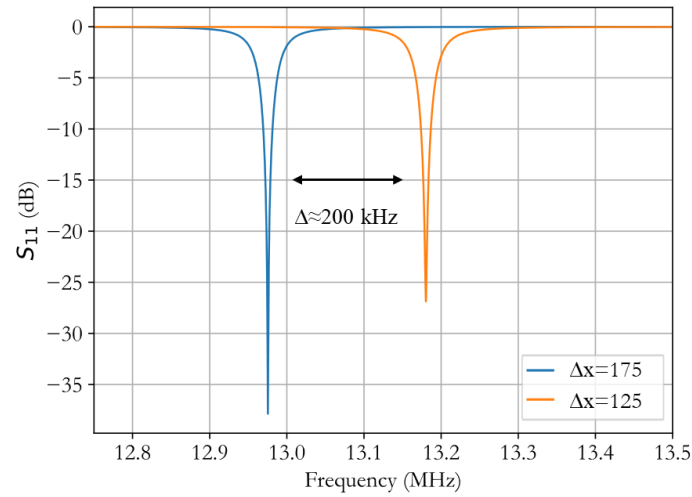


Figure 99: Simulated S_{11} reflection coefficient showing the tuning range of the RF coil. The resonant frequency shifts by approximately 205 kHz as the tuning plates are moved from a separation of $\Delta x = 175$ mm to $\Delta x = 125$ mm.

While the mechanism provides an effective means of frequency adjustment, it is crucial to assess its effect on coil performance. Figure 100 illustrates the relationship between the tuning plate separation (Δx) and the coil's loaded Q-factor. The plot reveals that the Q-factor is highly dependent on the proximity of the conductive plates. As the plates move closer to the coil (decreasing Δx), the Q-factor progressively degrades. This degradation is attributed to increased losses from eddy currents induced in the copper-clad plates increasing the effective resistance. The Q-factor reaches its maximum value of approximately 340 at the largest separation ($\Delta x \approx 175$ mm) and decreases significantly at closer distances. However, notably, the Q-factor remains relatively stable in the 125 mm to 175 mm range, with a deviation of less than 3% from its maximum value, indicating that fine-tuning within this region has a minimal impact on coil performance before the Q-factor begins to decrease more sharply at closer distances.

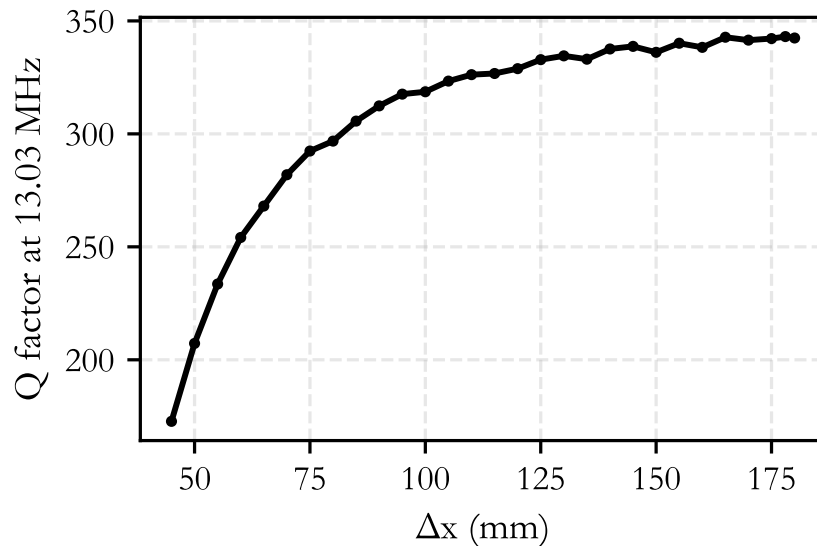


Figure 100: Impact of tuning plate position on coil performance. The plot shows the relationship between the Q-factor and plate separation, illustrating the trade-off between tuning and coil efficiency.

V.D. Culture Chamber

To fit inside the hollow double walled borosilicate bore, a smaller modified custom-designed $\varnothing 25 \text{ mm} \times 85 \text{ mm}$ culture chamber (Figure 67) was engineered to accommodate the same 3D bio-printed scaffold presented in chapter III. Similarly to the design presented in chapter III, the design featured a primary cylindrical chamber measuring around $\varnothing 20 \text{ mm} \times 50 \text{ mm}$, serving as the central housing for the bio-printed scaffold. The chamber was also designed to be supplied with liquid through two separate 3 mm cylindrical channels, and the “Sprinkler” function of the first design was directly integrated into the main body. Additionally, all the internal dimensions were maintained from the first design (Inlet and outlet ports, sprinkler cone dimensions and outlet ports, sprinkler channels etc...). The main difference was the return path. The large 3 mm return path from the first design was replaced by two 1 mm by 7 mm rectangular holes running through the flat surfaces designed to prevent sample rotation see Figure 67. This version was designed to bio-print directly inside the main chamber.

Furthermore, as there is no longer any separate sprinkler mechanism the seal is only performed by a single O-ring (RS PRO ID 17, OD 22, E 2.50, part number 196-4975) placed at the top between the main and a threaded 3D-printed cap.

Similarly to the first version, all parts aside from the O-rings were designed using SOLIDWORKS 2023 (Dassault Systèmes, Vélizy-Villacoublay, France) to be printed using either Fused Deposition Modelling (FDM) or Stereolithography (SLA) 3D printers.

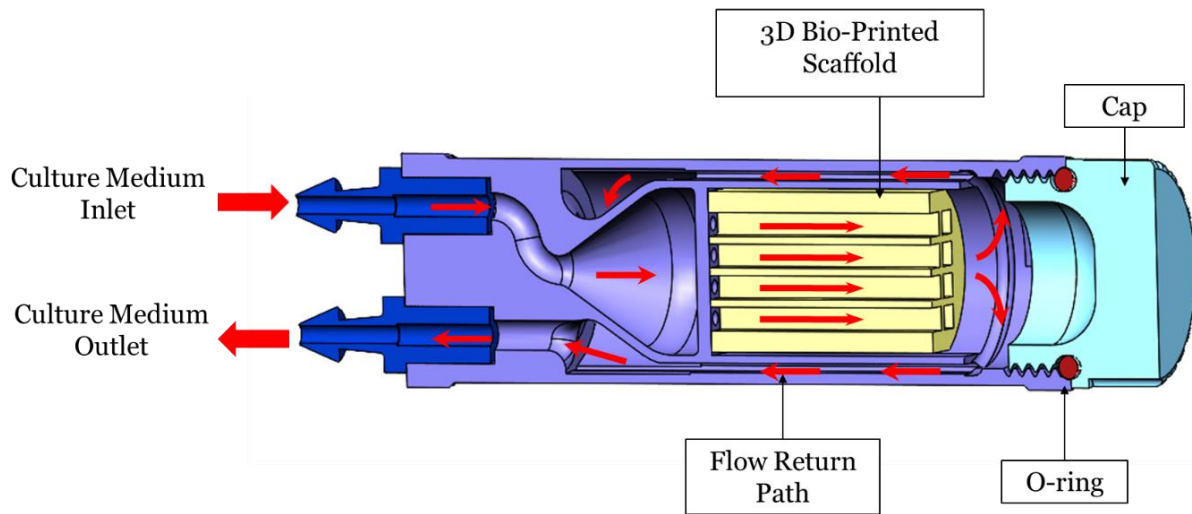


Figure 101 : Longitudinal section CAD rendering illustrating the main components of the new $\varnothing 25 \times 85 \text{ mm}$ chamber (flow path indicated with red arrows).

To enable the precise placement of this new culture chamber, another sliding mechanism was devised (Figure 102). It consisted of 3D printed threaded rod and fillet assembly fixed at the base of the hollow double walled borosilicate bore. This simple mechanism worked by tightening the rod to move the culture chamber upwards and loosening it to bring it down. Furthermore, as the temperature within the chamber would need to be recorded, another non-magnetic PT1000 RTD was added to a hollow cavity at the bottom of the chamber.

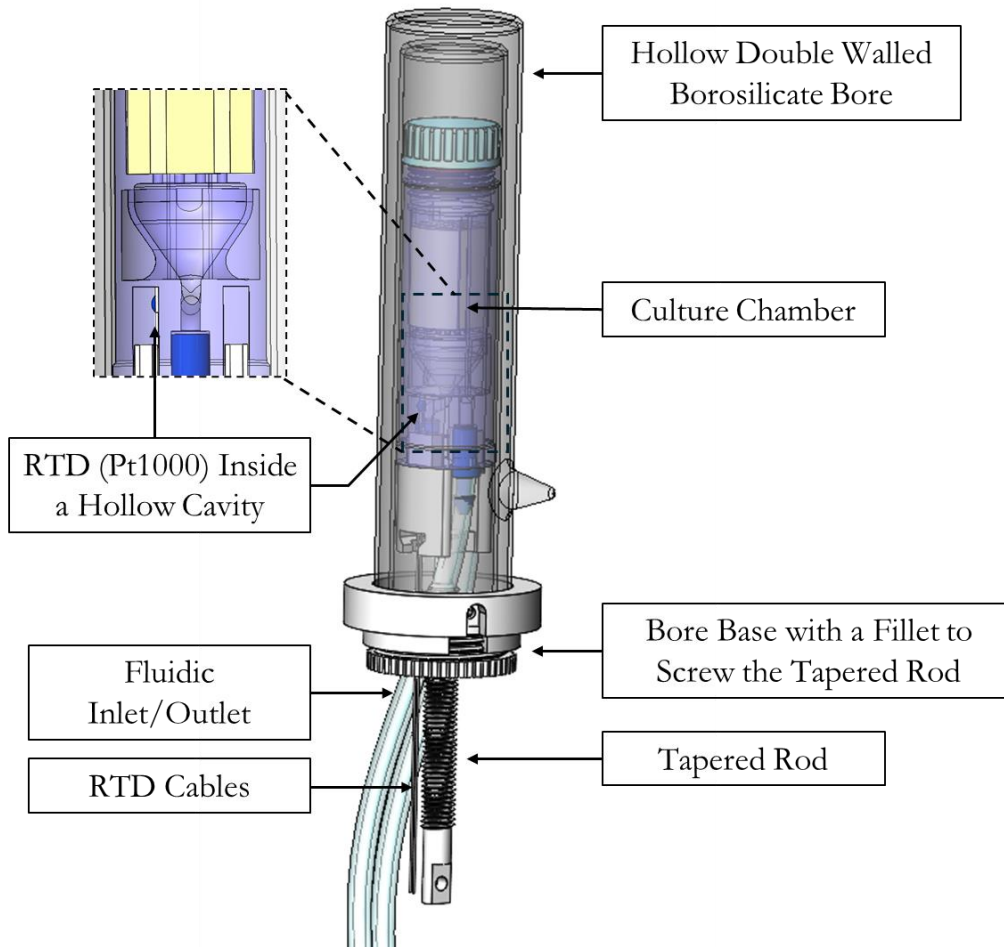


Figure 102: Detailed view of the culture chamber assembly within the borosilicate bore. The main image shows the threaded rod mechanism for precise vertical positioning of the chamber. The inset provides a magnified view of the PT1000 RTD sensor integrated at the base of the chamber for temperature monitoring.

VI. Results

VI.A. Fabricated Cryostat Setup

The structural components of the cryostat, shown in Figure 103, were 3D printed using either standard PLA on an Ultimaker printer or for the finer mechanisms HTV2 resin on a From Labs SLA printer. The cryostat consists of two main modules: a liquid nitrogen reservoir and an RF coil enclosure. Effective RF shielding was achieved by lining the external components with 25 μm copper EMI shielding foil.

The junction between the alumina substate and the transfer line was performed by soldering the end in the line directly onto the preprepared “pad” copper adhesive structure. The seal at the junction was further insured using a commercially available silicone joint (RUBSON).

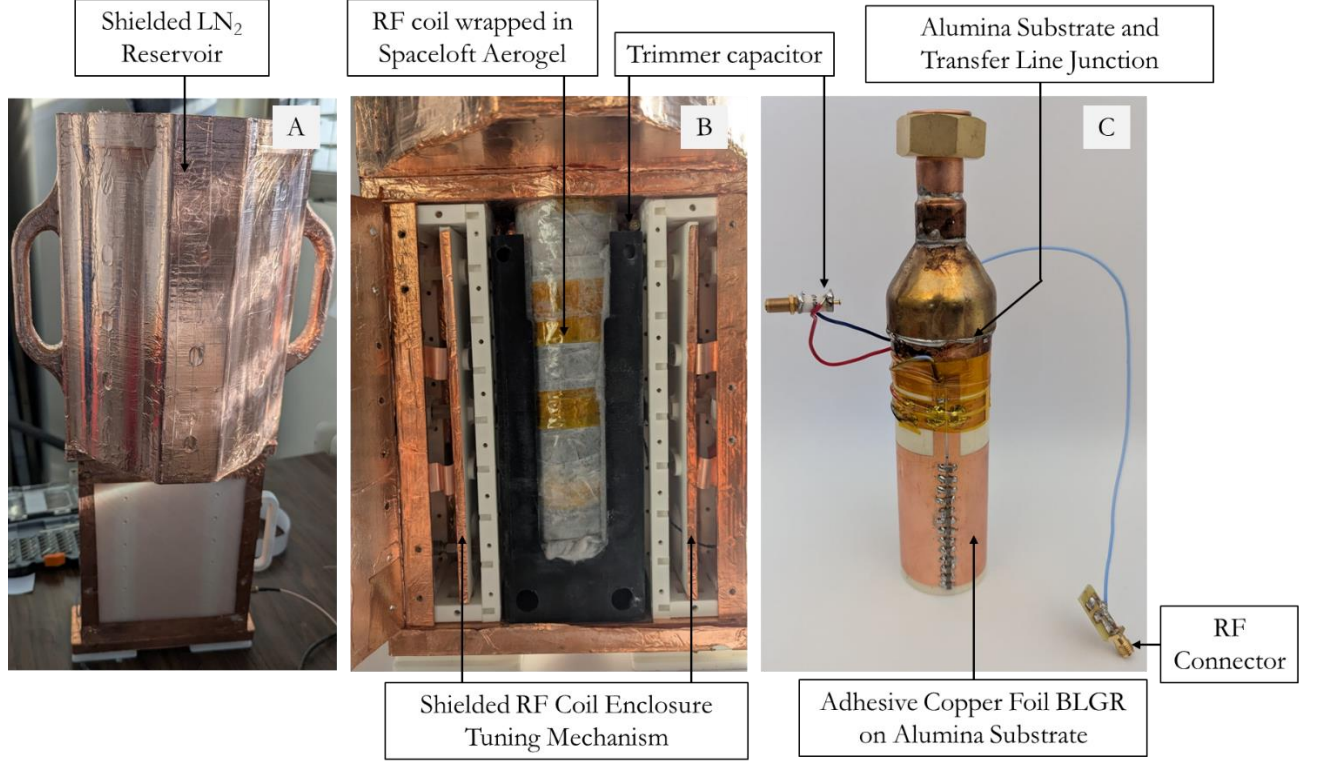


Figure 103: Images of the fabricated cryostat setup. A) Whole cryostat with the shielded reservoir on top of the shielded RF coil enclosure. B) Inside view of the shielded enclosure showcasing the tuning mechanism and the insulated coil C) Close up of the fabricated coil on alumina substrate. A trimmer capacitor was added to help with mismatch due to cooling.

Inside the enclosure, the bridged loop-gap resonator (BLGR) coil is mounted on an alumina substrate, as shown in Figure 103-C. An adjustable trimmer capacitor was added to correct for the impedance mismatch that occurs during cooling. All pieces were assembled into a robust and easy to carry cryostat able to fit precisely into the benchtop MRI scanner.

VI.B. Cooling Results

To evaluate the cryostat's effectiveness and its impact on the RF coil, key parameters were recorded throughout a complete cooling cycle, initiated by filling the reservoir with liquid nitrogen. The RTD sensors were probed every second using an Arduino-based interface, while the S-parameters were acquired periodically (every 10 s) using a VNA. The S_{11} parameters were measured using a Rohde & Schwarz ZNL3 vector network analyzer (calibrated using the ZN-135 calibration kit) and subsequently used to extract the loaded Q-factor using the KAJFEZ and HWAN method [156] previously employed in Chapters II and III. Similarly to Chapter III, all measurements were performed under loaded conditions with the culture chamber filled with a 16.3 mM Copper (II) Nitrate Trihydrate solution used in this work as a standard phantom solution.

1. Cooling Profiles

Figure 104 illustrates the thermal dynamics of the system, which can be understood in three main stages.

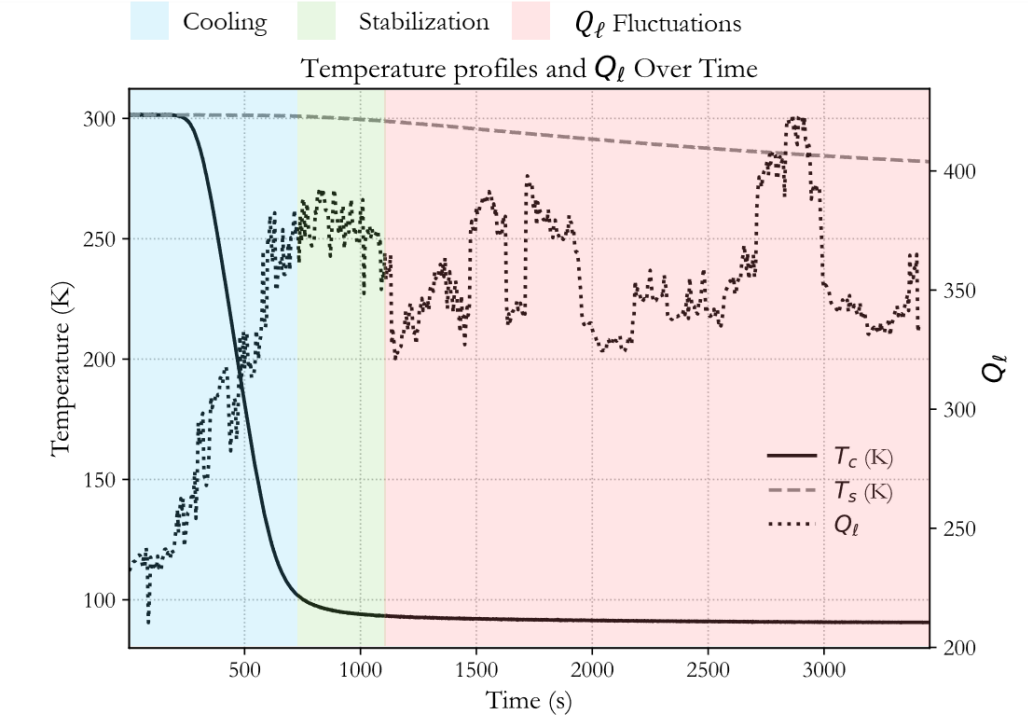


Figure 104: Temperature and loaded Q-factor (Q_l) evolution over the cooling period. The primary y-axis shows the coil temperature (T_c , solid black line) and sample temperature (T_s , dashed gray line) in Kelvin. The secondary y-axis shows the corresponding loaded Q-factor (Q_l , dotted black line), which increases as the coil temperature drops, indicating a reduction in coil resistance.

In the first stage, the coil temperature (T_c) shows a rapid decrease from ambient (~ 300 K) to a stable cryogenic temperature of approximately 90 K within less than 1000 seconds (17 minutes). This indicates efficient thermal transfer from the liquid nitrogen reservoir to the coil assembly. However, it also suggests a strong possibility of having temperature gradient between the cold source at 77K and the bottom of the alumina substrate at 90K.

During this stage, as expected, the loaded Q-factor (Q_l) of the coil shows a significant increase as the temperature drops. The Q_l rises from an initial value of ~ 240 to a peak close to 400, directly correlating with the reduction in the coil's intrinsic conductor losses. This improvement is the fundamental basis for the expected SNR gain in cryogenic coils.

The second stage consists of the temperature stabilization of the coil around 90K and a Q-factor just under 400.

Near the end of this second stage, the sample temperature (T_s) exhibits a slow, gradual decline, indicating that the culture chamber is not fully thermally insulated from the cold source. By replacing the internal Aerogel of the original conceptual design with the borosilicate double walled tube, although conduction losses were significantly decreased, radiation losses seem to have been severely underestimated. As such, at around 1250 s a thin layer of ice was observed to be forming at the top of the culture chamber (near the cap). This was later mitigated by adding a continuous flow of room temperature solution through the chamber, but it still highlights a major avenue of improvement for the proposed design.

Finally, the third stage is characterized by fluctuations observed in the Q_l measurement after the initial, rapid cooling phase and short stabilization phase have concluded. As seen in Figure 104, while the coil temperature (T_c) remains stable, the sample temperature (T_s) continues its slow decline. The Q_l fluctuations are therefore possibly attributed to variations in the sample load.

caused by the phantom media's gradual phase transition from liquid to solid. This hypothesis is further supported by the resonant frequency data in Figure 105, which also exhibits similar fluctuations, suggesting changes in the capacitive load as the phantom transitions.

2. Frequency Response

The measured frequency response of the coil during cooling revealed a significant downward shift that contrasted with initial simulation predictions. While simulations suggested a slight increase in frequency of approximately 6 kHz due to changes in the coil conductor's skin effect, the experiment showed a consistent decrease in resonant frequency. As shown in Figure 105, the measured frequency closely mirrors the temperature curve, dropping from its initial value as the coil cools and stabilizing once the target temperature is reached with a drop in frequency of a staggering -53 kHz.

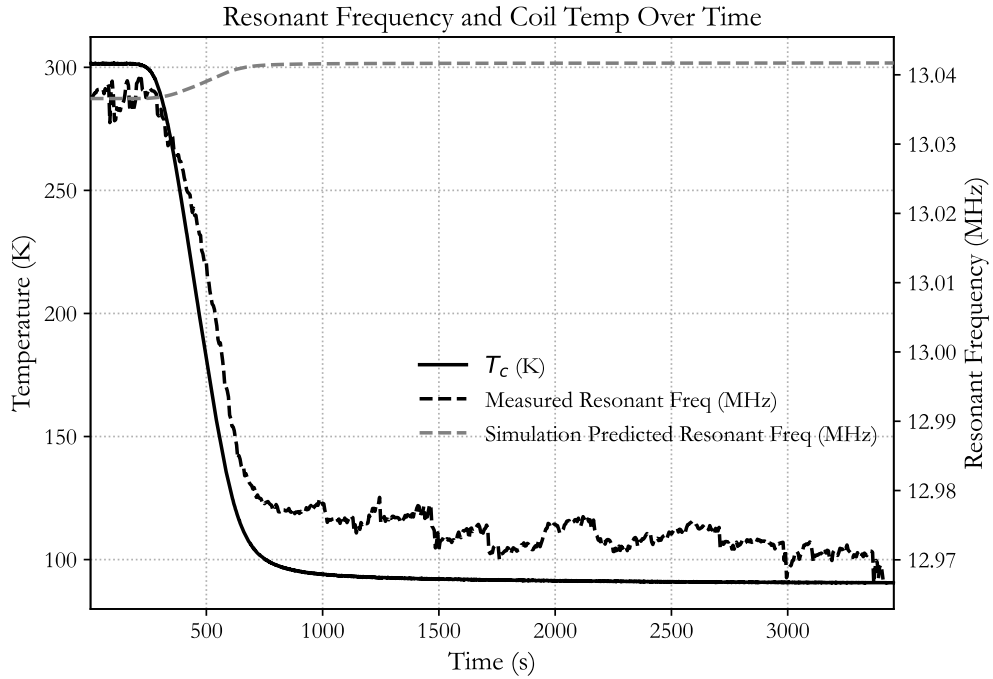


Figure 105: Measured vs. simulated resonant frequency shift during cooling. The primary y-axis shows the coil temperature (T_c , solid black line). The secondary y-axis displays the measured resonant frequency (dashed black line) and the simulation-predicted frequency (dashed gray line) in MHz.

This -53 kHz drop strongly indicates that the dominant factor influencing the frequency shift is not the coil's inductance, but a significant change in the total tuning capacitance. The resonance frequency of the coil is primarily determined by the relationship

$$f = \frac{1}{2\pi LC}. \quad (\text{VI-1})$$

While the small change in inductance predicted by the simulation is likely present, the much larger measured frequency drop must be caused by an increase in capacitance.

The tuning stage consists of thirteen 680 pF High-Q chip capacitors arranged in parallel, resulting in a total nominal capacitance of 8.84 nF at 300 K. According to the manufacturer's datasheet, these capacitors have a Temperature Coefficient of Capacitance (TCC) of ± 30 ppm/ $^{\circ}\text{C}$. This value enables the prediction of how the capacitance will change as the temperature drops by

210 K. Since the resonant frequency decreased, the total capacitance must have increased. Therefore, to calculate the expected shift, we use the most negative TCC value, $-30 \text{ ppm}/^{\circ}\text{C}$. This predicts a capacitance increase of 0.63%, leading to an expected frequency of 12.989 MHz, a downward shift of 41 kHz.

This calculated shift of -41 kHz is very close to the measured shift of -53 kHz. The minor difference can be explained by several real-world factors, such as component tolerances and the fact that the operating temperature of 90 K is outside the manufacturer's specified lower range ($\sim 218 \text{ K}$) for the TCC leading to unpredictable non-linear responses at lower temperatures.

The observed frequency drop, while contrary to initial simulations, can potentially be fully explained by the temperature coefficient of the tuning capacitors. The initial simulation accurately modeled the expected change in the copper coil but did not account for the much larger, opposing effect from the capacitors. The experimental results highlight that for this resonator design, capacitive temperature dependence is the dominant factor in determining the frequency shift at cryogenic temperatures. For further iterations, this drift should either be modeled and taken into account or high-end cryogenic temperature compatible MLLC's should be selected instead (e.g the Presidio Components ceramic capacitors (CRX)[183])

VI.C. Cryogenic Coil Performance

1. SNR Gain Measurements

To quantify the improvement in performance, the SNR was measured from images acquired at both room temperature (300K) and cryogenic temperature (90K) using the same method detailed in chapter III. The mean SNR was calculated from a region of interest (S) within the sample and compared to the standard deviation of the background noise (N). The results are summarized in Table 17 .

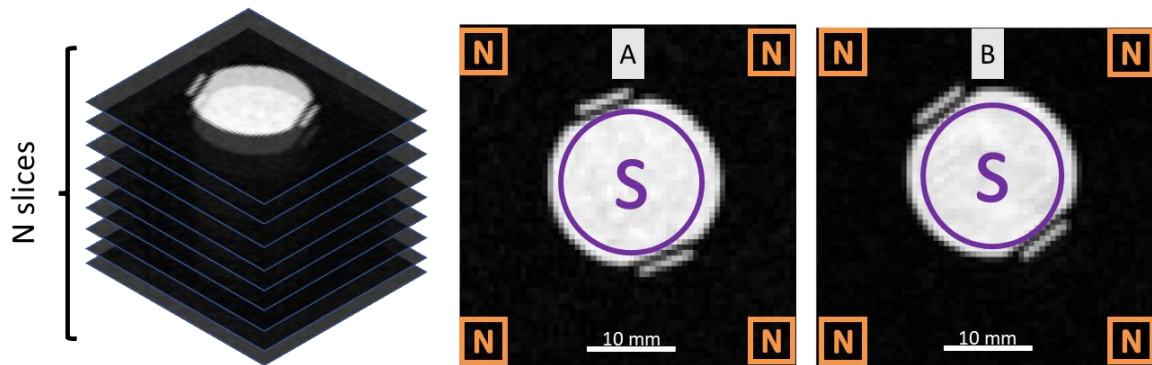


Figure 106: GRE acquisitions ($\text{FA} = 30^{\circ}$, $\text{TE} = 4\text{ms}$, $\text{TR}=12\text{ms}$, $\text{AQ time} = 1\text{m } 11\text{s}$) of 16.3 mM Copper (II) Nitrate Trihydrate solution filled chambers for SNR measurements of A) the coil at 300K B) the cooled coil at 90K.

Coil Temperature (K)	Mean Measured SNR from Images	Measured SNR Gain from Images
300	~ 36.78	$\sim 57\%$
90	~ 57.64	

Table 17 : Summary of SNR measurements and gain from cryogenic experiment

This measured gain, while substantial, is lower than the ~83% gain predicted by simulations for cooling to 90 K. The discrepancy can be attributed to several factors. Firstly, the simulation assumes a purely coil-noise-dominant regime. In a practical setup, there may be a non-negligible contribution from sample noise, which is not reduced by cooling the coil and thus limits the maximum achievement gain. Secondly, and most likely, factors such as imperfect impedance matching at cryogenic temperatures and minor losses in the cryostat materials can further reduce the measured performance compared to the ideal simulated case. This is further backed up by the discrepancies in quality factors with a simulated quality factors of around 342 at 300 K to 484 at 90 K for the simulation against the 240 at 300 K and close to 400 at 90K for the fabricated setup. Either way, to investigate this further, a complete unloaded cooling scenario should be performed to obtain unloaded quality factor profiles Q_u along with thorough material characterization and modeling specially for the Spaceloft Aerogel whose electrical properties are by far the biggest unknown.

Despite this, a 57% improvement represents a major enhancement in coil sensitivity which may rival the performance of the proposed 3D Plastronics setup in chapter III. These results are, however, insufficient to properly conclude on the better approach.

VII. Discussion and Conclusion

This work successfully demonstrated the proof-of-concept for a low-cost, cryogenically cooled RF coil system for benchtop MRI, validating it as a viable strategy for significant SNR enhancement. The primary goal was to develop an accessible system using readily available materials and fabrication methods such as 3D printing. The experimental results, which show a measured SNR gain of 57% upon cooling the coil to 90 K, confirm the fundamental hypothesis that a simple laboratory-made cooling system and cryogenic coil combination can lead to improved SNR performance.

However, while the predicted SNR gain at 90 K was approximately 83%, the measured gain was lower. This discrepancy can be attributed to several practical factors. The most significant is likely the difference between the simulated and measured quality factors; the fabricated coil exhibited a lower Q-factor at both room temperature (240 vs. 342) and cryogenic temperatures (400 vs. 484). This suggests that some losses are unaccounted-for in the proposed model, potentially from the adhesive copper foil, the dielectric properties of the Spaceloft Aerogel, or lumped element models. A thorough, unloaded characterization of the coil during a cooling cycle, and deeper material characterization under cryogenic conditions would be necessary to investigate these loss mechanisms further.

One of the biggest discrepancies was found in the resonant frequency shift was found during cooling. Simulations, which focused on the change in copper's skin effect, predicted a slight increase in frequency (~6 kHz). In contrast, the experiment revealed a significant decrease of 53 kHz. This result strongly indicates that the dominant effect was not the coil's inductance but rather the temperature-dependent characteristics of the MLCC tuning capacitors. A theoretical calculation based on the manufacturer's temperature coefficient of capacitance (TCC), rated for a minimum temperature of 218 K predicted a downward shift of 41 kHz, which tends to align with the measured value. This means that for this type of resonator design, the temperature stability of the lumped components is a more dominant factor than the changing properties of the conductor itself. Future simulations will have to incorporate accurate temperature-dependent models for all lumped components to yield better predictive results.

The project also highlighted significant engineering challenges that offer clear avenues for improvement. The most critical was the thermal insulation of the sample chamber. The gradual cooling of the sample, leading to the formation of ice, indicates that while the double-walled borosilicate tube effectively limited conductive heat transfer, it was insufficient to prevent heat

loss through thermal radiation. The radiation effects were severely underestimated when the choice was made to replace the internal opaque Aerogel insulation from the first preliminary design proposal with a transparent bore. This is a major design flaw that must be addressed to maintain sample viability and temperature stability, especially for live-cell imaging. Furthermore, the final coil temperature of 90 K, despite the 77 K liquid nitrogen source, points to an unavoidable temperature gradient inherent to the indirect cooling design which might to B_1 field inhomogeneities.

When comparing this cryogenic approach to the 3D Plastronics method detailed in Chapter III, a direct one-to-one conclusion is challenging. The baseline room-temperature SNR of the cryogenic coil (~ 37) was unexpectedly much lower than that of the manufacturer provided reference coil (~ 68), a difference not fully explained by the difference quality factors. A more controlled study is needed to investigate this baseline difference. Nevertheless, both techniques have demonstrated the potential to boost SNR, one by enabling better sensitivity through an optimized geometrical form factor, and the other relying on coil induced thermal noise reduction in a specific regime. These methods are, however, not mutually exclusive as there is a clear indication that finding a way to merge both approaches could lead to significant gains in the right regime. The deployment and investigation of such an approach will have to start with the development of cheaper and extremely thin MRI compatible insulation materials enabling a closer fit between the sample and the cryogenic coil.

Situating this work within the broader context of cryogenic MRI, the observed 57% SNR gain is a great achievement, consistent with the foundational principles of reducing thermal noise in a coil-noise-dominant regime. Other groups have reported similar or greater gains. For instance, Resmer et al. demonstrated a factor of 2.8 SNR improvement (a 180% gain) for a copper surface coil cooled to 77 K. These gains were obtained at a much lower field of 0.01 T where coil noise is even more pronounced and using Litz wires which are known to improve current distribution and reduce losses due to skin and proximity effects, maximizing the achievable Q-factor at cryogenic temperatures [178]. The performance of cryo-coils is often benchmarked against high-temperature superconductor (HTS) coils, which can achieve extremely high Q-factors and theoretically larger SNR gains [180]. However, HTS coils introduce substantial practical challenges, including high cost, mechanical fragility, limited compatibility with standard soldering and connection methods, and narrow operational bandwidths that can complicate imaging sequences [106]. Therefore, the approach presented here, a cryogenically cooled copper coil, represents good middle ground. It offers a performance boost over room-temperature coils while avoiding the prohibitive complexity and cost of HTS systems, aligning perfectly with the goal of creating an accessible, high-performance solution for widespread laboratory use. However, as the proposed cryostat design reached temperatures in the 90K range it can also be used with HTS coil. A complementary study can therefore be relatively easily conducted to investigate the gains from HTS in this specific context and elucidate if the potential benefits might outweigh the financial cost and other limitations.

The cryostat design remains a work in progress, but its successful initial demonstration opens up several exciting future directions. The most promising path forward involves integrating advanced fabrication techniques. For instance, Ceratronics, an extension of the 3D-Plastronics concept using ceramic materials, could be employed to create coil substrates with integrated cryogenic microfluidic channels. Such a design would allow liquid nitrogen to flow directly through the substrate, enabling more uniform and efficient cooling over the entire coil surface. This would mitigate the temperature gradients found in the current design and could potentially bring the coil temperature much closer to 77 K, unlocking even greater SNR gains.

General Conclusion

The central objective of this thesis was to address the lack of practical, non-invasive tools for the longitudinal monitoring of large, living, three-dimensional biological constructs in tissue engineering applications. While low-field benchtop Magnetic Resonance Imaging (MRI) presents an accessible and versatile solution, its utility is constrained by an inherently low SNR, which limits image resolution and prolongs acquisition times. This work successfully confronted this challenge by systematically investigating and developing two distinct, hardware-centric strategies for SNR enhancement. The first, relied on the design of optimized radiofrequency (RF) resonators fabricated via 3D Plastronics, and the second, on implementation of a proof-of-concept on an accessible and cost-effective cryogenic coil system.

The investigation began with a comprehensive, simulation-driven analysis to identify the most effective RF coil topology for generating a homogeneous magnetic field in a locally shielded, low-field environment. Through a combination of analytical modeling and full-wave electromagnetic simulations, the Bridged Loop-Gap Resonator (BLGR) was identified as the superior design. It demonstrated an optimal balance of high B1 field homogeneity, operational stability under varying sample loads, and intrinsic sensitivity, significantly outperforming classical solenoid, varying-pitch, and even sophisticated optimized spherical coil designs whose theoretical advantages were diminished by practical implementation constraints. To date, this work is one of the first to revisit classical coil geometry through numerical analysis to characterize their performance in terms of sensitivity and homogeneity, tailored specifically to locally shielded small sample imaging using a benchtop MRI. Although not yet published, a submission to a journal of magnetic resonance as an original paper is currently in preparation, detailing the full comparative study, validation methodology, and implications for benchtop MRI system design. The manuscript will emphasize how the presented optimized architectures can bridge the gap between theoretical coil performance and practical manufacturability, offering a scalable route toward high-sensitivity, homogeneous excitation for small to medium sized tissue-engineered constructs in low-field MRI environments.

The primary contribution of this thesis was the successful translation of this optimal BLGR design into a fully integrated, multifunctional culture chamber using a Rapid 3D Plastronics (R3DP) fabrication process. This approach enabled the creation of a perfusion culture chamber with a form-fitted RF coil and its matching circuit embedded directly onto its surface. This seamless integration of imaging coil and perfused culture chamber functionality yielded a significant, experimentally validated SNR gain of 60% and superior B1 field homogeneity when compared to a conventional, manufacturer-provided coil of the same kind. The system's efficacy was demonstrated through the high-resolution morphological assessment of a 3D bio-printed hydrogel scaffold *in situ*, validating 3D Plastronics not only as a fabrication method, but as a powerful tool for creating application-specific, high-performance integrated MRI setups bridging the gap between imaging and biological experimentation.

In parallel, a second strategy for SNR enhancement was explored through the design and implementation of a low-cost, proof-of-concept cryostat and cryogenic coil setup. The proposed cryostat successfully cooled a copper-based BLGR to approximately 90 K, resulting in a measured SNR gain of 57% compared to its room temperature performance. This result confirmed the viability of cryogenics as a powerful method for reducing thermal noise in a coil-noise-dominated regime. The proposed novel custom-made cost-effective and accessible cryostat design is particularly noteworthy for its approach, which innovatively bypasses the need for a complex vacuum system for thermal insulation. It relies on specifically selected, for the most part low-cost and easily accessible materials, while resolving significant assembly difficulties through the strategic use of custom 3D-printed components. This methodology not only simplifies construction but

also validates a practical and accessible pathway to achieving cryogenic cooling suitable for benchtop MRI environments in most laboratory settings. However, the study also illuminated significant engineering challenges, including the need for improved sample thermal insulation and the management of temperature-dependent component behavior, positioning this approach as a promising but less mature alternative requiring further refinement for routine laboratory use.

Comparing these two approaches, the 3D Plastronics approach offers the more immediate and versatile solution. It not only provides a boost in sensitivity but, more importantly, it redefines the experimental workflow by creating a cohesive, integrated system where the sample's "life-support" environment and the imaging sensor are one and the same. This work has demonstrated that the performance limitations of low-field MRI are not insurmountable but can be effectively addressed through targeted coil design and hardware innovation.

To ensure successful adoption of this tool by biologists, a crucial step will be to establish a common vocabulary between MRI specialists and tissue engineers. MRI physicists and engineers quantify performance in terms of markers like T1 and T2 relaxation times or the apparent diffusion coefficient (ADC), while biologists seek to evaluate functional tissue characteristics such as cell density, extracellular matrix composition, porosity, or viability. The full potential of this technology will only be realized when clear, quantitative correlations are established, translating MRI parameters into direct biological indicators. Creating this "translation lexicon" is essential for biologists to interpret MRI data autonomously and reliably, to integrate this non-invasive monitoring tool into their experimental protocols to accelerate the development of functional tissues.

Looking forward, the different approaches investigated here open numerous avenues for future research. The 3D Plastronics chamber can be further enhanced by integrating additional sensors for real-time monitoring of pH, oxygen, or temperature, creating a truly comprehensive "smart" culture chamber. The cryogenic system can be refined by leveraging advanced materials and fabrication techniques, such as Ceratronics with integrated microfluidic cooling channels, to achieve lower temperatures and more uniform thermal management. Ultimately, the tools and methodologies established in this thesis lay the groundwork for making quantitative, high-resolution MRI an indispensable and routine tool in the tissue engineering laboratory, accelerating the development of functional, clinically relevant biological tissues.

Building upon this foundation, a dedicated future research project would focus on merging the two successful strategies of this thesis (integrated design and cryogenic enhancement) into a single system. The key to this unification lies in exploiting the "Ceratronics" concept, moving beyond the proof-of-concept cryostat to develop a fully integrated cryogenic "smart" chamber. This project would abandon the parallel approach and instead use advanced, multi-material additive manufacturing of technical ceramics to fabricate a monolithic cryo-probe. Technical ceramics, such as alumina, offer a unique combination of high thermal conductivity, mechanical rigidity, and MRI compatibility. This strategy would leverage the high thermal conductivity of ceramics to create an ultra-efficient thermal link for cooling the RF coil while integrating the required functionalities through a 3D Ceratronics approach.

The core of this project would be to leverage Ceratronics to 3D print a cryostat body that integrates three key components. First, it would contain intricate microfluidic channels, as suggested, to circulate a cryogen (like liquid nitrogen) or cold gas. The high thermal conductivity of the alumina structure would ensure rapid and uniform cooling of the entire probe body. Second, this conductive ceramic body would serve as a highly efficient heat sink and precision formwork for a high-temperature superconducting (HTS) version of the optimized BLGR coil, which would be positioned in direct thermal contact with it. This would dramatically decrease the coil's thermal noise, pushing the SNR gain far beyond the 57% achieved with cooled copper and unlocking the true potential of cryogenics. Third, and most critically, the Ceratronic structure would be designed

to house a thermally isolated, independent perfusion chamber for the living biological construct. This "chamber-within-a-chamber" design, potentially using an integrated vacuum gap or insulating aerogels fabricated during the multi-material printing process, is essential for maintaining the sample at a stable temperature. It includes its own perfusion and the "smart" sensors (pH, O₂) envisioned for the Plastronics system., while the surrounding RF coil operates close 77 K.

This "Ceratron Cryo-Probe" would represent the ultimate synthesis of the thesis work. It would achieve the massive SNR enhancement of HTS cryogenics while simultaneously providing the seamless, "life-support" integration of the 3D Plastronics platform. By solving the engineering bottlenecks of thermal management using material conductivity to our advantage for the coil while creating a sophisticated insulation solution for the sample through adjustable agile geometries, this system is expected to deliver remarkable image quality. This would further empower biologists, transforming the benchtop MRI from a promising device into the gold-standard, non-invasive monitoring tool required to accelerate the development of large, thick, opaque tissue-engineered constructs.

List of publications

First Author Publications

2025 - *Low-field MRI as a multiparametric tool for large engineered tissue characterization* - Yilbert Gimenez*, Valernst Martial Gilmus*, Elliott Cowles, Lucie Essayan, Vincent Semet, Michel Cabrera, Christophe A. Marquette, Simon A. Lambert, Emma Petiot – Submission Ongoing (lien HAL à venir) *co-first authors

2025 - *3D Plastronics Culture Chambers and Benchtop MRI Setup Toward Tissue Engineering Applications* - Valernst Martial Gilmus, Yilbert Gimenez, Tony Gerges, Philippe Lombard, Vincent Semet, Christophe Marquette, Emma Petiot, Michel Cabrera, Simon Auguste Lambert – Submitted (lien HAL à venir)

2024 - *Low-cost 300 mT Benchtop MRI Scanner and Custom MR Bioreactor for real-time multi-parametric assessment of 3D-bioprinted biological tissue constructs* - Valernst Martial Gilmus, Yilbert Gimenez, Tony Gerges, Jean-Lynce Gnanago, Philippe Lombard, Vincent Semet, Michel Cabrera, Christophe Marquette, Emma Petiot, Simon Auguste Lambert - HAL : <https://hal.science/hal-05312609v1>

2022 - *Projet pratique de mise en œuvre de la plastronique* - Valernst Martial Gilmus, Tony Gerges, Philippe Lombard, Jean-Yves Charmeaux, Michel Cabrera, Bruno Allard - HAL : <https://hal.science/hal-05312581v1>

Other contributions

2025 - *Material Technical Guide for On-Demand AM Production of Bioreactors* - Christophe A. Marquette, Matthieu Koprivnik, Julien Le Boterff, Lucas Lemarié, Edwin-Joffrey Courtial, Valernst Martial Gilmus, Celine Thomann, Lukas Raddatz, Magali Barbaroux, Emma Petiot - Pre-print (lien HAL à venir)

2024 - *A Toolbox to Develop and Produce In-Mold Electronics Devices* - Lombard P., Cauchy-Clerc H., Semet V., Gerges T., Gilmus V. M., Arli M., Lakhdar H., Allard B., Charneau J.-Y., Cabrera M. - HAL : <https://hal.science/hal-04525630v1>

2023 - *Eco-conception de dispositifs plastroniques 3D : Électronique Structurale Surmoulée sur Acide PolyLActique* - Caroline Goument, Valernst Martial Gilmus, Philippe Lombard, Tony Gerges, Simon Lambert, Bruno Allard, Jean-Yves Charneau, Michel Cabrera - HAL : <https://hal.science/hal-04580726v1>

2023 - *In-Mold Electronics with Biosourced and Biodegradable Polymers* - Goument C., Gerges T., Lombard P., Lakhdar H., Arli M., Gilmus V. M., Lambert S. A., Charneau J.-Y., Cabrera M., Allard B. - HAL : <https://hal.science/hal-04069869v1>

2023 - *The MR-Bioreactor: Micro-MRI of thick living tissues to characterize 4-days old ex-ovo chick embryo morphology* - Jean-Lynce Gnanago, Sophie Calvet, Tony Gerges, Valernst Martial Gilmus, Camila Pereira Sousa, Michel Cabrera, Julien Falk, Simon Auguste Lambert - ISMRM 2023 Annual Meeting, Abstract #4555 - <https://archive.ismrm.org/2023/4555.html>

2023 - *In-Mold Electronics on Poly(Lactic Acid): towards a more sustainable mass production of plastronic devices* - Goument C., Gerges T., Lombard P., Lakhdar H., Arli M., Gilmus V. M., Lambert S. A., Allard B., Charneau J.-Y., Cabrera M. - HAL : <https://hal.science/hal-04069897v1>

2022 - *Combination of screen printed process with electrodeposition steps to drastically improve flexible MRI coils sensitivity* - Camila Sousa, Jean-Lynce Gnanago, Valernst Gilmus, Tony Gerges, Philippe Lombard, Michel Cabrera, Hugo Dorez, Simon Lambert - HAL : <https://hal.science/hal-04933822v1>

2022 - *Électronique Structurale Surmoulée sur acide PolyLActique (PLA)* - Goument C., Gerges T., Lombard P., Gilmus V., Lakhdar H., Arli M., Lambert S. A., Allard B., Charneau J.-Y., Cabrera M. - HAL : <https://hal.science/hal-04069768v1>

Bibliography

- [1] H. Kaul and Y. Ventikos, "On the Genealogy of Tissue Engineering and Regenerative Medicine," *Tissue Eng Part B Rev*, vol. 21, no. 2, pp. 203–217, Apr. 2015, doi: 10.1089/ten.teb.2014.0285.
- [2] S. A. Langhans, "Three-Dimensional in Vitro Cell Culture Models in Drug Discovery and Drug Repositioning," *Front Pharmacol*, vol. 9, p. 6, 2018, doi: 10.3389/fphar.2018.00006.
- [3] M. A. Lancaster and J. A. Knoblich, "Organogenesis in a dish: modeling development and disease using organoid technologies," *Science*, vol. 345, no. 6194, p. 1247125, July 2014, doi: 10.1126/science.1247125.
- [4] P. Bédard *et al.*, "Innovative Human Three-Dimensional Tissue-Engineered Models as an Alternative to Animal Testing," *Bioengineering*, vol. 7, no. 3, Sept. 2020, doi: 10.3390/bioengineering7030115.
- [5] Francesco Pampaloni, F. Pampaloni, Emmanuel G. Reynaud, E. G. Reynaud, Ernst H. K. Stelzer, and E. H. K. Stelzer, "The third dimension bridges the gap between cell culture and live tissue," *Nature Reviews Molecular Cell Biology*, vol. 8, no. 10, pp. 839–845, Oct. 2007, doi: 10.1038/nrm2236.
- [6] M. Ravi, V. Paramesh, S. R. Kaviya, E. Anuradha, and F. D. P. Solomon, "3D Cell Culture Systems: Advantages and Applications," *Journal of Cellular Physiology*, vol. 230, no. 1, pp. 16–26, Jan. 2015, doi: 10.1002/jcp.24683.
- [7] D. Egger and A. Lavrentieva, *Basic Concepts on 3D Cell Culture*, 1st ed. in Learning Materials in Biosciences. Springer, 2021. doi: 10.1007/978-3-030-66749-8.
- [8] T. J. Hinton, A. Lee, and A. W. Feinberg, "3D bioprinting from the micrometer to millimeter length scales: Size does matter," *Current Opinion in Biomedical Engineering*, vol. 1, pp. 31–37, Mar. 2017, doi: 10.1016/j.cobme.2017.02.004.
- [9] M. Sun *et al.*, "3D Cell Culture—Can It Be As Popular as 2D Cell Culture?," *Advanced NanoBiomed Research*, vol. 1, no. 5, p. 2000066, May 2021, doi: 10.1002/anbr.202000066.
- [10] S. Radi *et al.*, "The proteome of osteoblasts in a 3D culture perfusion bioreactor model compared with static conditions," *Scientific Reports*, vol. 15, no. 1, p. 12120, Apr. 2025, doi: 10.1038/s41598-025-96632-0.
- [11] A. M. Sillon, A. C. Allori, E. H. Davidson, D. D. Reformat, R. J. Allen Jr., and S. M. Warren, "A Novel Flow-Perfusion Bioreactor Supports 3D Dynamic Cell Culture," *BioMed Research International*, vol. 2009, no. 1, p. 873816, Jan. 2009, doi: 10.1155/2009/873816.
- [12] S. Maxson, D. Orr, and K. J. L. Burg, "Bioreactors for Tissue Engineering," in *Tissue Engineering: From Lab to Clinic*, N. Pallua and C. V. Suscheck, Eds., Berlin, Heidelberg: Springer Berlin Heidelberg, 2011, pp. 179–197. doi: 10.1007/978-3-642-02824-3_10.
- [13] S. Y. Nam, L. M. Ricles, L. J. Suggs, and S. Y. Emelianov, "Imaging Strategies for Tissue Engineering Applications," *Tissue Eng Part B Rev*, vol. 21, no. 1, pp. 88–102, Feb. 2015, doi: 10.1089/ten.teb.2014.0180.
- [14] A. Guagliardi, C. Giannini, A. Cedola, M. Mastrogiacomo, M. Ladisa, and R. Cancedda, "Toward the x-ray microdiffraction imaging of bone and tissue-engineered bone," *Tissue Eng Part B Rev*, vol. 15, no. 4, pp. 423–442, Dec. 2009, doi: 10.1089/ten.TEB.2009.0034.

- [15] Mrignayani Kotecha and M. Kotecha, "Principles and Applications of Quantitative Parametric MRI in Tissue Engineering," pp. 21–47, Feb. 2017, doi: 10.1002/9781119193272.ch2.
- [16] L. Chastagnier *et al.*, "Exploring multiple bioprocess monitoring modalities for Large-scale 3D Bioprinted Tissue Cultivation," Mar. 25, 2024. doi: 10.21203/rs.3.rs-4021195/v1.
- [17] S. F. Othman, H. Xu, and J. J. Mao, "Future role of MR elastography in tissue engineering and regenerative medicine," *Journal of Tissue Engineering and Regenerative Medicine*, vol. 9, no. 5, pp. 481–487, 2015, doi: 10.1002/term.1801.
- [18] M. Poirier-Quinot *et al.*, "High-Resolution 1.5-Tesla Magnetic Resonance Imaging for Tissue-Engineered Constructs: A Noninvasive Tool to Assess Three-Dimensional Scaffold Architecture and Cell Seeding," *Tissue Engineering Part C: Methods*, vol. 16, no. 2, pp. 185–200, Apr. 2010, doi: 10.1089/ten.tec.2009.0015.
- [19] E. Breton, C. Goetz, P. Choquet, and A. Constantinesco, "Low field magnetic resonance imaging in rat in vivo," *IRBM*, vol. 29, no. 6, pp. 366–374, Dec. 2008, doi: 10.1016/j.irbmret.2008.05.002.
- [20] C. Goetz, E. Breton, P. Choquet, V. Israel-Jost, and A. Constantinesco, "SPECT Low-Field MRI System for Small-Animal Imaging," *Journal of Nuclear Medicine*, vol. 49, no. 1, pp. 88–93, Jan. 2008, doi: 10.2967/jnumed.107.044313.
- [21] H. Nitzsche *et al.*, "Characterization of scaffolds for tissue engineering by benchtop-magnetic resonance imaging," *Tissue Eng Part C Methods*, vol. 15, no. 3, pp. 513–521, Sept. 2009, doi: 10.1089/ten.TEC.2008.0488.
- [22] H. Caysa, H. Metz, K. Mäder, and T. Mueller, "Application of Benchtop-magnetic resonance imaging in a nude mouse tumor model," *J Exp Clin Cancer Res*, vol. 30, no. 1, p. 69, July 2011, doi: 10.1186/1756-9966-30-69.
- [23] R. Ayde, M. Vornehm, Y. Zhao, F. Knoll, E. X. Wu, and M. Sarraçanie, "MRI at low field: A review of software solutions for improving SNR," *NMR in Biomedicine*, vol. 38, no. 1, p. e5268, Jan. 2025, doi: 10.1002/nbm.5268.
- [24] J. Hennig, "An evolution of low-field strength MRI," *Magnetic Resonance Materials in Physics, Biology and Medicine*, vol. 36, no. 3, pp. 335–346, July 2023, doi: 10.1007/s10334-023-01104-z.
- [25] N. Koonjoo, B. Zhu, G. C. Bagnall, D. Bhutto, and M. S. Rosen, "Boosting the signal-to-noise of low-field MRI with deep learning image reconstruction," *Scientific Reports*, vol. 11, no. 1, p. 8248, Apr. 2021, doi: 10.1038/s41598-021-87482-7.
- [26] M. Sarraçanie and N. Salameh, "Low-Field MRI: How Low Can We Go? A Fresh View on an Old Debate," *Front. Phys.*, vol. 8, June 2020, doi: 10.3389/fphy.2020.00172.
- [27] C. Z. Cooley *et al.*, "Design and implementation of a low-cost, tabletop MRI scanner for education and research prototyping," *Journal of Magnetic Resonance*, vol. 310, p. 106625, Jan. 2020, doi: 10.1016/j.jmr.2019.106625.
- [28] I. Georgakoudi, W. L. Rice, M. Hronik-Tupaj, and D. L. Kaplan, "Optical Spectroscopy and Imaging for the Noninvasive Evaluation of Engineered Tissues," *Tissue Engineering Part B: Reviews*, vol. 14, no. 4, pp. 321–340, Dec. 2008, doi: 10.1089/ten.teb.2008.0248.
- [29] Alyssa A. Appel *et al.*, "Imaging challenges in biomaterials and tissue engineering.," *Biomaterials*, vol. 34, no. 28, pp. 6615–6630, Sept. 2013, doi: 10.1016/j.biomaterials.2013.05.033.

- [30] D. M. Doroski, K. S. Brink, and J. S. Temenoff, "Techniques for biological characterization of tissue-engineered tendon and ligament," *Biomaterials*, vol. 28, no. 2, pp. 187–202, Jan. 2007, doi: 10.1016/j.biomaterials.2006.08.040.
- [31] M. Harling, P. Breeding, T. Haysley, M. Chesley, M. Mason, and K. Tilbury, "Multiphoton Microscopy for the Characterization of Cellular Behavior on Naturally Derived Polysaccharide Tissue Constructs With Irregular Surfaces for the Development of Platform Biomaterials," *Front Bioeng Biotechnol*, vol. 8, p. 802, July 2020, doi: 10.3389/fbioe.2020.00802.
- [32] S. J. Matcher, "Practical aspects of OCT imaging in tissue engineering," *Methods Mol Biol*, vol. 695, pp. 261–280, 2011, doi: 10.1007/978-1-60761-984-0_17.
- [33] P. A. Santi, "Light Sheet Fluorescence Microscopy: A Review," *J Histochem Cytochem.*, vol. 59, no. 2, pp. 129–138, Feb. 2011, doi: 10.1369/0022155410394857.
- [34] M. A. Pysz, S. S. Gambhir, and J. K. Willmann, "Molecular imaging: current status and emerging strategies," *Clinical Radiology*, vol. 65, no. 7, pp. 500–516, July 2010, doi: 10.1016/j.crad.2010.03.011.
- [35] T. F. Massoud and S. S. Gambhir, "Molecular imaging in living subjects: seeing fundamental biological processes in a new light," *Genes Dev*, vol. 17, no. 5, pp. 545–580, Mar. 2003, doi: 10.1101/gad.1047403.
- [36] F. Schick *et al.*, "1.5 vs 3 Tesla Magnetic Resonance Imaging: A Review of Favorite Clinical Applications for Both Field Strengths—Part 1," *Investigative Radiology*, vol. 56, no. 11, 2021, [Online]. Available: https://journals.lww.com/investigativeradiology/fulltext/2021/11000/1_5_vs_3_tesla_magnetic_resonance_imaging__a.2.aspx
- [37] N. H. M. F. Laboratory, "MRI: A Guided Tour - Magnet Academy." Accessed: May 05, 2025. [Online]. Available: <https://nationalmaglab.org/magnet-academy/read-science-stories/science-simplified/mri-a-guided-tour/>
- [38] R. R. Edelman, "The History of MR Imaging as Seen through the Pages of Radiology," *Radiology*, vol. 273, no. 2S, pp. S181–S200, Nov. 2014, doi: 10.1148/radiol.14140706.
- [39] C. P. Slichter, "Magnetic Interactions of Nuclei with Electrons," in *Principles of Magnetic Resonance*, C. P. Slichter, Ed., Berlin, Heidelberg: Springer, 1990, pp. 87–144. doi: 10.1007/978-3-662-09441-9_4.
- [40] J. E.- illustrated Jin, *Electromagnetic Analysis and Design in Magnetic Resonance Imaging*. in Biomedical Engineering; v. 1. CRC Press, 1998.
- [41] C. P. Slichter, "Basic Theory," in *Principles of Magnetic Resonance*, C. P. Slichter, Ed., Berlin, Heidelberg: Springer, 1990, pp. 11–64. doi: 10.1007/978-3-662-09441-9_2.
- [42] A. K. Singh, "Chapter 3 - Physicochemical, Electronic, and Mechanical Properties of Nanoparticles," in *Engineered Nanoparticles*, A. K. Singh, Ed., Boston: Academic Press, 2016, pp. 77–123. doi: 10.1016/B978-0-12-801406-6.00003-0.
- [43] C. M. Collins, *Electromagnetics in Magnetic Resonance Imaging*. in Physical Principles, Related Applications, and Ongoing Developments. Morgan & Claypool Publishers, 2016. doi: 10.1088/978-1-6817-4083-6.
- [44] John David Jackson, *Classical Electrodynamics (Third Edition)*. 1999. Accessed: Apr. 27, 2025. [Online]. Available: <http://archive.org/details/john-david-jackson-classical-electrodynamics-wiley-1999>
- [45] "Signal Detection Concepts," in *Magnetic Resonance Imaging*, John Wiley & Sons, Ltd, 2014, pp. 95–111. doi: 10.1002/9781118633953.ch7.

- [46] “One-Dimensional Fourier Imaging, k -Space, and Gradient Echoes,” in *Magnetic Resonance Imaging*, John Wiley & Sons, Ltd, 2014, pp. 141–164. doi: 10.1002/9781118633953.ch9.
- [47] “Multi-Dimensional Fourier Imaging and Slice Excitation,” in *Magnetic Resonance Imaging*, John Wiley & Sons, Ltd, 2014, pp. 165–206. doi: 10.1002/9781118633953.ch10.
- [48] K. P. Pruessmann, M. Weiger, M. B. Scheidegger, and P. Boesiger, “SENSE: sensitivity encoding for fast MRI,” *Magn Reson Med*, vol. 42, no. 5, pp. 952–962, Nov. 1999.
- [49] K. T. Block, M. Uecker, and J. Frahm, “Undersampled radial MRI with multiple coils. Iterative image reconstruction using a total variation constraint,” *Magnetic Resonance in Medicine*, vol. 57, no. 6, pp. 1086–1098, 2007, doi: 10.1002/mrm.21236.
- [50] M. Lustig, D. Donoho, and J. M. Pauly, “Sparse MRI: The application of compressed sensing for rapid MR imaging,” *Magnetic Resonance in Medicine*, vol. 58, no. 6, pp. 1182–1195, 2007, doi: 10.1002/mrm.21391.
- [51] J. A. Fessler and B. P. Sutton, “Nonuniform fast Fourier transforms using min-max interpolation,” *IEEE Transactions on Signal Processing*, vol. 51, no. 2, pp. 560–574, Feb. 2003, doi: 10.1109/TSP.2002.807005.
- [52] “Basics of Magnetic Resonance Imaging,” Barrow Neurological Institute. Accessed: Apr. 21, 2025. [Online]. Available: <https://www.barrowneuro.org/for-physicians-researchers/education/grand-rounds-publications-media/barrow-quarterly/volume-16-no-2-2000/basics-of-magnetic-resonance-imaging/>
- [53] “Sequence Design, Artifacts, and Nomenclature,” in *Magnetic Resonance Imaging*, John Wiley & Sons, Ltd, 2014, pp. 779–821. doi: 10.1002/9781118633953.ch26.
- [54] “Introductory Signal Acquisition Methods,” in *Magnetic Resonance Imaging*, John Wiley & Sons, Ltd, 2014, pp. 113–139. doi: 10.1002/9781118633953.ch8.
- [55] “Random Walks, Relaxation, and Diffusion,” in *Magnetic Resonance Imaging*, John Wiley & Sons, Ltd, 2014, pp. 619–636. doi: 10.1002/9781118633953.ch21.
- [56] A. Surov, H. J. Meyer, and A. Wienke, “Correlation between apparent diffusion coefficient (ADC) and cellularity is different in several tumors: a meta-analysis,” *Oncotarget; Vol 8, No 35*, 2017, Accessed: Jan. 01, 2017. [Online]. Available: <https://www.oncotarget.com/article/17752/text/>
- [57] B. J. Archer *et al.*, “Noninvasive Quantification of Cell Density in Three-Dimensional Gels by MRI,” *IEEE Transactions on Biomedical Engineering*, vol. 66, no. 3, pp. 821–830, Mar. 2019, doi: 10.1109/TBME.2018.2857443.
- [58] “MR Angiography and Flow Quantification,” in *Magnetic Resonance Imaging*, John Wiley & Sons, Ltd, 2014, pp. 701–737. doi: 10.1002/9781118633953.ch24.
- [59] M. Essig *et al.*, “Perfusion MRI: The Five Most Frequently Asked Technical Questions,” *American Journal of Roentgenology*, vol. 200, no. 1, pp. 24–34, Jan. 2013, doi: 10.2214/AJR.12.9543.
- [60] Shadi F. Othman, S. F. Othman, Huihui Xu, H. Xu, Jeremy J. Mao, and J. J. Mao, “Future role of MR elastography in tissue engineering and regenerative medicine,” *Journal of Tissue Engineering and Regenerative Medicine*, vol. 9, no. 5, pp. 481–487, May 2015, doi: 10.1002/term.1801.
- [61] P. Pothirajan, D. Dorcemus, S. Nukavarapu, and M. Kotecha, “True MRI assessment of stem cell chondrogenesis in a tissue engineered matrix,” in *2014 36th Annual International Conference of the IEEE Engineering in Medicine and Biology Society*, Aug. 2014, pp. 3933–3936. doi: 10.1109/EMBC.2014.6944484.

- [62] P. Pothirajan, S. Ravindran, A. George, R. L. Magin, and M. Kotecha, "Magnetic resonance spectroscopy and imaging can differentiate between engineered bone and engineered cartilage," in *2014 36th Annual International Conference of the IEEE Engineering in Medicine and Biology Society*, Aug. 2014, pp. 3929–3932. doi: 10.1109/EMBC.2014.6944483.
- [63] W. Gao *et al.*, "Application of medical imaging methods and artificial intelligence in tissue engineering and organ-on-a-chip," *Frontiers in Bioengineering and Biotechnology*, vol. Volume 10-2022, 2022, doi: 10.3389/fbioe.2022.985692.
- [64] K. Youssef, N. N. Jarenwattananon, B. J. Archer, J. Mack, M. L. Iruela-Arispe, and L. -S. Bouchard, "4-D Flow Control in Porous Scaffolds: Toward a Next Generation of Bioreactors," *IEEE Transactions on Biomedical Engineering*, vol. 64, no. 1, pp. 61–69, Jan. 2017, doi: 10.1109/TBME.2016.2537266.
- [65] Shadi F. Othman *et al.*, "MR elastography monitoring of tissue-engineered constructs.," *NMR in Biomedicine*, vol. 25, no. 3, pp. 452–463, Mar. 2012, doi: 10.1002/nbm.1663.
- [66] W. Li, L. Hong, L. Hu, and R. L. Magin, "Magnetization Transfer Imaging Provides a Quantitative Measure of Chondrogenic Differentiation and Tissue Development," *Tissue Engineering Part C: Methods*, vol. 16, no. 6, pp. 1407–1415, Dec. 2010, doi: 10.1089/ten.tec.2009.0777.
- [67] P. J. Basser and D. K. Jones, "Diffusion-tensor MRI: theory, experimental design and data analysis – a technical review," *NMR in Biomedicine*, vol. 15, no. 7–8, pp. 456–467, Nov. 2002, doi: 10.1002/nbm.783.
- [68] P. J. Basser and C. Pierpaoli, "Microstructural and Physiological Features of Tissues Elucidated by Quantitative-Diffusion-Tensor MRI," *Journal of Magnetic Resonance, Series B*, vol. 111, no. 3, pp. 209–219, June 1996, doi: 10.1006/jmrb.1996.0086.
- [69] M. Froeling, A. J. Nederveen, K. Nicolay, and G. J. Strijkers, "DTI of human skeletal muscle: the effects of diffusion encoding parameters, signal-to-noise ratio and T2 on tensor indices and fiber tracts," *NMR in Biomedicine*, vol. 26, no. 11, pp. 1339–1352, Nov. 2013, doi: 10.1002/nbm.2959.
- [70] S. Ramaswamy *et al.*, "Magnetic Resonance Imaging of Chondrocytes Labeled with Superparamagnetic Iron Oxide Nanoparticles in Tissue-Engineered Cartilage," *Tissue Engineering Part A*, vol. 15, no. 12, pp. 3899–3910, Dec. 2009, doi: 10.1089/ten.tea.2008.0677.
- [71] C. Obyn and I. Cleemput, "The capital cost and productivity of MRI in a Belgian setting," *JBR–BTR*, vol. 93, no. 2, pp. 92–96, 2010.
- [72] Aspect Imaging, "M-Series™ Compact MRI Systems." [Online]. Available: <https://aspectimaging.com/wp-content/uploads/2021/03/M-Series-Brochure-DOC10001019-Rev.3.pdf>
- [73] C. P. Bidinosti, I. S. Kravchuk, and M. E. Hayden, "Active shielding of cylindrical saddle-shaped coils: application to wire-wound RF coils for very low field NMR and MRI," *J Magn Reson*, vol. 177, no. 1, pp. 31–43, Nov. 2005, doi: 10.1016/j.jmr.2005.07.003.
- [74] Y. Liu *et al.*, "A low-cost and shielding-free ultra-low-field brain MRI scanner," *Nature Communications*, vol. 12, no. 1, p. 7238, Dec. 2021, doi: 10.1038/s41467-021-27317-1.
- [75] K. Selvaganesan, M. Sadinski, M. Gomes, A. Nacev, and S. B. King, "Self-shielded RF coils for imaging in unshielded environments at low-field," *ISMRM Annual Meeting*, [Online]. Available: <https://api.semanticscholar.org/CorpusID:274313319>
- [76] J. Vliem, C. Najac, L. Beatrice, A. Webb, and I. Zivkovic, "Design and performance of a Toroidal RF Volume Coil with Intrinsic Electromagnetic Interference Rejection for low-field

- Portable Halbach-Based MRI Systems,” *arXiv preprint arXiv:2505.21745*, 2025, [Online]. Available: <https://arxiv.org/abs/2505.21745>
- [77] “Tabletop MRI lab | Neurotechnology in Action | Brain and Cognitive Sciences,” MIT OpenCourseWare. Accessed: Apr. 28, 2025. [Online]. Available: https://ocw.mit.edu/courses/9-123-neurotechnology-in-action-fall-2014/resources/mit9_123f14_lab9/
- [78] M. Alecci and P. Jezzard, “Characterization and reduction of gradient-induced eddy currents in the RF shield of a TEM resonator,” *Magnetic Resonance in Medicine*, vol. 48, no. 2, pp. 404–407, 2002, doi: 10.1002/mrm.10226.
- [79] P. Jezzard, A. S. Barnett, and C. Pierpaoli, “Characterization of and correction for eddy current artifacts in echo planar diffusion imaging,” *Magnetic Resonance in Medicine*, vol. 39, no. 5, pp. 801–812, 1998, doi: 10.1002/mrm.1910390518.
- [80] Z. Yang *et al.*, “RF shielding designs for birdcage coils for preclinical MRI at 9.4 T,” *Magn Reson Imaging*, vol. 94, pp. 1–6, Dec. 2022, doi: 10.1016/j.mri.2022.08.018.
- [81] B. J. Lee, R. D. Watkins, C.-M. Chang, and C. S. Levin, “Low eddy current RF shielding enclosure designs for 3T MR applications,” *Magn Reson Med*, vol. 79, no. 3, pp. 1745–1752, Mar. 2018, doi: 10.1002/mrm.26766.
- [82] M. Yu *et al.*, “Ultra-low noise graphene/copper/nylon fabric for electromagnetic interference shielding in ultra-low field magnetic resonance imaging,” *Journal of Magnetic Resonance*, vol. 317, p. 106775, Aug. 2020, doi: 10.1016/j.jmr.2020.106775.
- [83] A. Besheer, H. Caysa, H. Metz, T. Mueller, J. Kressler, and K. Mäder, “Benchtop-MRI for in vivo imaging using a macromolecular contrast agent based on hydroxyethyl starch (HES),” *Int J Pharm*, vol. 417, no. 1–2, pp. 196–203, Sept. 2011, doi: 10.1016/j.ijpharm.2010.10.051.
- [84] J. P. Marques, F. F. J. Simonis, and A. G. Webb, “Low-field MRI: An MR physics perspective,” *Journal of Magnetic Resonance Imaging*, vol. 49, no. 6, pp. 1528–1542, 2019, doi: 10.1002/jmri.26637.
- [85] B. Blümich *et al.*, “The NMR-mouse: construction, excitation, and applications,” *Magnetic Resonance Imaging*, vol. 16, no. 5, pp. 479–484, June 1998, doi: 10.1016/S0730-725X(98)00069-1.
- [86] B. Blümich and K. Singh, “Desktop NMR and Its Applications From Materials Science To Organic Chemistry,” *Angewandte Chemie International Edition*, vol. 57, no. 24, pp. 6996–7010, 2018, doi: 10.1002/anie.201707084.
- [87] B. Blümich, F. Casanova, J. Perlo, F. Presciutti, C. Anselmi, and B. Doherty, “Noninvasive Testing of Art and Cultural Heritage by Mobile NMR,” *Acc. Chem. Res.*, vol. 43, no. 6, pp. 761–770, June 2010, doi: 10.1021/ar900277h.
- [88] L. Backhouse, M. Dias, J. P. Gorce, J. Hadgraft, P. J. McDonald, and J. W. Wiechers, “GARField magnetic resonance profiling of the ingress of model skin-care product ingredients into human skin in vitro,” *JPharmSci*, vol. 93, no. 9, pp. 2274–2283, Sept. 2004, doi: 10.1002/jps.20137.
- [89] Z. He, W. He, J. Wu, and Z. Xu, “The Novel Design of a Single-Sided MRI Probe for Assessing Burn Depth,” *Sensors*, vol. 17, no. 3, Art. no. 3, Mar. 2017, doi: 10.3390/s17030526.
- [90] “MR-CAT || Core Analyzing Tomograph.” Accessed: Apr. 22, 2025. [Online]. Available: <https://www.pure-devices.com/index.php/products/products-mr-cat.html>

- [91] J. Perlo, C. Mülder, E. Danieli, C. Hopmann, B. Blümich, and F. Casanova, “Digital processing of images of extruded rubber profiles for process control MRI,” *Measurement*, vol. 82, pp. 466–475, Mar. 2016, doi: 10.1016/j.measurement.2015.11.046.
- [92] A. Constantinesco *et al.*, “Low-field dedicated and desktop magnetic resonance imaging systems for agricultural and food applications,” *Magnetic Resonance in Chemistry*, vol. 35, no. 13, pp. S69–S75, 1997, doi: 10.1002/(SICI)1097-458X(199712)35:13%3CS69::AID-OMR198%3E3.0.CO;2-5.
- [93] S. Blystone, M. Nuix, A. S. Traoré, H. Cochard, C. Picon-Cochard, and G. Pagés, “Towards portable MRI in the plant sciences,” *Plant Methods*, vol. 20, no. 1, p. 31, Feb. 2024, doi: 10.1186/s13007-024-01152-z.
- [94] M. Jones, P. S. Aptaker, J. Cox, B. A. Gardiner, and P. J. McDonald, “A transportable magnetic resonance imaging system for *in situ* measurements of living trees: The Tree Hugger,” *Journal of Magnetic Resonance*, vol. 218, pp. 133–140, May 2012, doi: 10.1016/j.jmr.2012.02.019.
- [95] Pure Devices GmbH, “Benchtop MRI Spectrometers – High Performance, Low-Cost, Portable.” [Online]. Available: <https://www.pure-devices.com/index.php/products/products-research/products-research-overview.html>
- [96] “Education NMR NIUMAG EDUMR | Niumag Corporation.” Accessed: Apr. 23, 2025. [Online]. Available: <https://www.nmralyzer.com/education-nmr-niumag-edumr.html>
- [97] C. Kittel, *Introduction to solid state physics*, 8. ed., [Repr.]. Hoboken, NJ: Wiley, 20.
- [98] “Fast Imaging in the Steady State,” in *Magnetic Resonance Imaging*, John Wiley & Sons, Ltd, 2014, pp. 447–510. doi: 10.1002/9781118633953.ch18.
- [99] “Introduction to MRI Coils and Magnets,” in *Magnetic Resonance Imaging*, John Wiley & Sons, Ltd, 2014, pp. 823–857. doi: 10.1002/9781118633953.ch27.
- [100] L. Darrasse and J.-C. Ginefri, “Perspectives with cryogenic RF probes in biomedical MRI,” *Biochimie*, vol. 85, no. 9, pp. 915–937, Sept. 2003, doi: 10.1016/j.biochi.2003.09.016.
- [101] D. I. Hoult and R. Deslauriers, “A high-sensitivity, high-B1 homogeneity probe for quantitation of metabolites,” *Magnetic Resonance in Medicine*, vol. 16, no. 3, pp. 411–417, Dec. 1990, doi: 10.1002/mrm.1910160307.
- [102] M. Boudreau, C. L. Tardif, N. Stikov, J. G. Sled, W. Lee, and G. B. Pike, “B1 mapping for bias-correction in quantitative T1 imaging of the brain at 3T using standard pulse sequences,” *Journal of Magnetic Resonance Imaging*, vol. 46, no. 6, pp. 1673–1682, Dec. 2017, doi: 10.1002/jmri.25692.
- [103] S. Ropele *et al.*, “Assessment and correction of B1-induced errors in magnetization transfer ratio measurements,” *Magnetic Resonance in Medicine*, vol. 53, no. 1, pp. 134–140, Jan. 2005, doi: 10.1002/mrm.20310.
- [104] J. Mispelter, M. Lupu, and A. Briguet, *NMR Probeheads for Biophysical and Biomedical Experiments*. IMPERIAL COLLEGE PRESS, 2011. doi: 10.1142/p759.
- [105] D. D. Henkes, “A Design Tool for Improving the Input Match of Low Noise Amplifiers,” *High Frequency Electronics*, pp. 54–64, Feb. 2007.
- [106] A. Webb and T. O’Reilly, “Tackling SNR at low-field: a review of hardware approaches for point-of-care systems,” *MAGMA*, vol. 36, no. 3, pp. 375–393, 2023, doi: 10.1007/s10334-023-01100-3.

- [107] G. S. Smith, "Proximity Effect in Systems of Parallel Conductors," *Journal of Applied Physics*, vol. 43, no. 5, pp. 2196–2203, May 1972, doi: 10.1063/1.1661474.
- [108] P. L. Dowell, "Effects of eddy currents in transformer windings," *Proceedings of the Institution of Electrical Engineers*, vol. 113, no. 8, pp. 1387–1394, 1966, doi: 10.1049/piee.1966.0236.
- [109] A. Kumar, W. A. Edelstein, and P. A. Bottomley, "Noise figure limits for circular loop MR coils," *Magnetic Resonance in Medicine*, vol. 61, no. 5, pp. 1201–1209, 2009, doi: 10.1002/mrm.21948.
- [110] F. D. Doty, G. Entzminger, J. Kulkarni, K. Pamarthy, and J. P. Staab, "Radio frequency coil technology for small-animal MRI," *NMR in Biomedicine*, vol. 20, no. 3, pp. 304–325, 2007, doi: 10.1002/nbm.1149.
- [111] W. Mao, M. B. Smith, and C. M. Collins, "Exploring the limits of RF shimming for high-field MRI of the human head," *Magnetic Resonance in Medicine*, vol. 56, no. 4, pp. 918–922, Oct. 2006, doi: 10.1002/mrm.21013.
- [112] Q. X. Yang, S. H. Li, and M. B. Smith, "A Method for Evaluating the Magnetic Field Homogeneity of a Radiofrequency Coil by Its Field Histogram," *Journal of Magnetic Resonance, Series A*, vol. 108, no. 1, pp. 1–8, May 1994, doi: 10.1006/jmra.1994.1081.
- [113] J. Franke, "Three-Dimensional Molded Interconnect Devices (3D-MID)," in *Three-Dimensional Molded Interconnect Devices (3D-MID)*, örg F. J, Ed., Carl Hanser Verlag GmbH & Co. KG, 2014, p. I–XII. doi: 10.3139/9781569905524.fm.
- [114] X. V. L. Nguyen *et al.*, "Radio-Frequency Energy Harvesting Using Rapid 3D Plastronics Prototyping Approach: A Case Study," *Journal of Low Power Electronics and Applications*, vol. 13, no. 1, Art. no. 1, Mar. 2023, doi: 10.3390/jlpea13010019.
- [115] S. Kamotesov, P. Lombard, V. Semet, B. Allard, M. Moguedet, and M. Cabrera, "The Potential of 3D-MID Technology for Omnidirectional Inductive Wireless Power Transfer," in *2018 13th International Congress Molded Interconnect Devices (MID)*, in 2018 13th International Congress Molded Interconnect Devices (MID). Würzburg, Germany, Sept. 2018. doi: 10.1109/ICMID.2018.8526962.
- [116] "Warming Your Coffee with In-Mold Structural Electronics | Flexible Hybrid Electronics," Voltera Inc. Accessed: Apr. 29, 2025. [Online]. Available: <https://www.voltera.io/blog/warming-coffee-in-mold-structural-electronics>
- [117] O. Rusanen, T. Heikkilä, M. Heikkinen, and P. Niskala, "Smart Molded Structures Bring Surfaces to Life," 2019. Accessed: May 02, 2025. [Online]. Available: <https://www.semanticscholar.org/paper/Smart-Molded-Structures-Bring-Surfaces-to-Life-Rusanen-Heikkil%C3%A4/f9877d7778206333538ff9bcdd018bcd5058cfc8>
- [118] P. Lombard, T. Gerges, J.-Y. Charneau, B. Allard, and M. Cabrera, "Procédé plastronique Electronique Structurale Surmoulée (ESS, IME – In Mold Electronics) dans un projet de mise en œuvre pratique," *J3eA*, vol. 21, p. 1012, 2022, doi: 10.1051/j3ea/20221012.
- [119] A. M. Zamarayeva *et al.*, "Custom, spray coated receive coils for magnetic resonance imaging," *Sci Rep*, vol. 11, no. 1, p. 2635, Jan. 2021, doi: 10.1038/s41598-021-81833-0.
- [120] T. Gerges, V. Semet, P. Lombard, S. Gaillard, M. Cabrera, and S. A. Lambert, "3D Plastronics for Smartly Integrated Magnetic Resonance Imaging Coils," *Front. Phys.*, vol. 8, July 2020, doi: 10.3389/fphy.2020.00240.
- [121] Normandie AéroEspace, "Etude bibliographique / Etat de l'art – Plastronique," Normandie AéroEspace, France, 2022.

- [122] T. Gerges, V. Semet, P. Lombard, B. Allard, and M. Cabrera, "Rapid 3D-Plastronics prototyping by selective metallization of 3D printed parts," *Additive Manufacturing*, vol. 73, p. 103673, July 2023, doi: 10.1016/j.addma.2023.103673.
- [123] V. Dhinakaran, K. P. Manoj Kumar, P. M. Bupathi Ram, M. Ravichandran, and M. Vinayagamoorthy, "A review on recent advancements in fused deposition modeling," *Materials Today: Proceedings*, vol. 27, pp. 752–756, Jan. 2020, doi: 10.1016/j.matpr.2019.12.036.
- [124] F. Lupone, E. Padovano, F. Casamento, and C. Badini, "Process Phenomena and Material Properties in Selective Laser Sintering of Polymers: A Review," *Materials*, vol. 15, no. 1, Art. no. 1, Jan. 2022, doi: 10.3390/ma15010183.
- [125] "Pollen AM | Thermoplastiques élastomères." Accessed: Feb. 08, 2023. [Online]. Available: https://www.pollen.am/fr/thermoplastic_elasromers_general_introduction/
- [126] "Thermoplastics," Polymers. Accessed: May 02, 2025. [Online]. Available: <http://polymers.com.au/thermoplastics/>
- [127] D. Gerdil *et al.*, "A new perspective for the plastronic functionalization of 3D printed composites and high performances thermoplastic parts. From prototypes to final parts," Messe München and Organic and Printed Electronics Association, Feb. 2025. Accessed: May 02, 2025. [Online]. Available: <https://hal.science/hal-04995489>
- [128] M. Cabrera, J. Y. Jezequel, and J. C. Andre, "Three-dimensional machining by laser photopolymerization," in *Lasers in Polymer Science and Technology: Applications*, vol. III, J.-P. Fouassier and J. F. Rabek, Eds., Boca Raton, FL: CRC Press, 1990, pp. 73–95.
- [129] X. V. L. Nguyen *et al.*, "Accessing High Temperature Stereolithography Photopolymers Compabilities for Antennas," in *10th National Days on Energy Harvesting and Storage (JNRSE)*, Grenoble, France, June 2021. Accessed: Feb. 20, 2025. [Online]. Available: <https://hal.science/hal-03451946>
- [130] J.-L. Gnanago and et al, "Compact MRI bioreactor for real-time monitoring 3D printed tissue-engineered constructs. in Proceedings of the 29th International Society inMagnetic Resonance for Medecine Annual Meeting 3122," 2021. Accessed: Apr. 15, 2024. [Online]. Available: <https://archive.ismrm.org/2021/3122.html>
- [131] M. Abdelkader, S. Petrik, D. Nestler, and M. Fijalkowski, "Ceramics 3D Printing: A Comprehensive Overview and Applications, with Brief Insights into Industry and Market," *Ceramics*, vol. 7, no. 1, pp. 68–85, 2024, doi: 10.3390/ceramics7010006.
- [132] "Alumina 4N Resin," Formlabs. Accessed: July 26, 2024. [Online]. Available: <https://formlabs.com/store/materials/ceramic-alumina-4n-resin/?srsltid=AfmBOoredVFl-IBFNFZTCb-Bak11s1tau0A0x1JxmzdxmrgD3vTqOYHg>
- [133] "Modern Electroplating, 5th Edition | Wiley," Wiley.com. Accessed: Apr. 30, 2025. [Online]. Available: <https://www.wiley.com/en-us/Modern+Electroplating%2C+5th+Edition-p-9780470167786>
- [134] B. D. Barker, "Electroless deposition of metals," *Surface Technology*, vol. 12, no. 1, pp. 77–88, Jan. 1981, doi: 10.1016/0376-4583(81)90138-2.
- [135] S. Kamotesov, "Transmission d'énergie par induction électromagnétique en plastronique 3D," PhD Thesis, Université Claude Bernard Lyon 1, 2019. [Online]. Available: <https://theses.hal.science/tel-02496934v1/file/TH2019KAMOTESOV Sergkei.pdf>
- [136] X. V. L. Nguyen, "Investigation of ambient radio-frequency energy harvesting using a rapid prototyping plastronic approach," Theses, INSA de Lyon, 2023. [Online]. Available: <https://theses.hal.science/tel-04248664>

- [137] C. Goument, “Caractérisation, mise en forme et recyclage de polymères biosourcés pour le développement d’un procédé de fabrication plastronique respectueux de l’environnement,” These de doctorat, Lyon, INSA, 2023. Accessed: Apr. 29, 2025. [Online]. Available: <https://theses.fr/2023ISAL0109>
- [138] C. Coillot, E. Nativel, M. Zanca, and C. Goze-Bac, “The magnetic field homogeneity of coils by \hack\break means of the space harmonics suppression of \hack\break the current density distribution,” *Journal of Sensors and Sensor Systems*, vol. 5, no. 2, pp. 401–408, 2016, doi: 10.5194/jsss-5-401-2016.
- [139] D. I. Hoult and R. E. Richards, “The signal-to-noise ratio of the nuclear magnetic resonance experiment,” *Journal of Magnetic Resonance (1969)*, vol. 24, no. 1, pp. 71–85, Oct. 1976, doi: 10.1016/0022-2364(76)90233-X.
- [140] J. Mispelter and M. Lupu, “Homogeneous resonators for magnetic resonance: A review,” *Comptes Rendus. Chimie*, vol. 11, no. 4–5, pp. 340–355, 2008, doi: 10.1016/j.crci.2007.10.003.
- [141] C. Staat, M. Mützel, and A. Haase, “A bridged loop gap resonator (BLGR) for small animal imaging by 1.5 T MRI systems,” *Review of Scientific Instruments*, vol. 91, no. 3, p. 033704, Mar. 2020, doi: 10.1063/1.5128046.
- [142] B. de Vos *et al.*, “Design, Characterisation and Performance of an Improved Portable and Sustainable Low-Field MRI System,” *Front. Phys.*, vol. 9, July 2021, doi: 10.3389/fphy.2021.701157.
- [143] Z. H. Ren and S. Y. Huang, “The Design of A Short Solenoid with Homogeneous B1 for A Low-field Portable MRI Scanner Using Genetic Algorithm,” Paris, France: Proceedings of the 26th Annual Meeting of the ISMRM, 2018, p. 1720. [Online]. Available: <https://cds.ismrm.org/protected/18MProceedings/PDFfiles/1720.html>
- [144] A. Labiche, S. Kan, A. Leroy-Willig, and C. Wary, “Investigation of the radio frequency magnetic field inside a hollow cylinder used for nuclear magnetic resonance imaging,” *Review of Scientific Instruments*, vol. 70, no. 4, pp. 2113–2115, Apr. 1999, doi: 10.1063/1.1149722.
- [145] S. A. McNeill, P. L. Gor’kov, K. Shetty, W. W. Brey, and J. R. Long, “A low-E magic angle spinning probe for biological solid state NMR at 750MHz,” *Journal of Magnetic Resonance*, vol. 197, no. 2, pp. 135–144, Apr. 2009, doi: 10.1016/j.jmr.2008.12.008.
- [146] M. D. Harpen, “The spherical birdcage resonator,” *Journal of Magnetic Resonance (1969)*, vol. 94, no. 3, pp. 550–556, Oct. 1991, doi: 10.1016/0022-2364(91)90141-F.
- [147] J. W. Clark, “A new method for obtaining a uniform magnetic field,” *Review of Scientific Instruments*, vol. 9, no. 10, p. 320, 1938.
- [148] J. E. Everett and J. E. Osemeikhian, “Spherical coils for uniform magnetic fields,” *Journal of Scientific Instruments*, vol. 43, no. 7, p. 470, July 1966, doi: 10.1088/0950-7671/43/7/311.
- [149] L. Bolinger, M. G. Prammer, and J. S. Leigh, “A multiple-frequency coil with a highly uniform B1 field,” *Journal of Magnetic Resonance (1969)*, vol. 81, no. 1, pp. 162–166, Jan. 1989, doi: 10.1016/0022-2364(89)90274-6.
- [150] G. B. Arfken, H. J. Weber, and F. E. Harris, *Mathematical Methods for Physicists: A Comprehensive Guide*. Amsterdam: Academic Press, 2011.
- [151] A. Stogryn, “Equations for Calculating the Dielectric Constant of Saline Water (Correspondence),” *IEEE Transactions on Microwave Theory and Techniques*, vol. 19, no. 8, pp. 733–736, Aug. 1971, doi: 10.1109/TMTT.1971.1127617.

- [152] L. Klein and C. Swift, "An improved model for the dielectric constant of sea water at microwave frequencies," *IEEE Transactions on Antennas and Propagation*, vol. 25, no. 1, pp. 104–111, Jan. 1977, doi: 10.1109/TAP.1977.1141539.
- [153] "Electrical properties of selected Formlabs SLA resins," Formlabs Customer Support. Accessed: July 29, 2025. [Online]. Available: https://support.formlabs.com/s/article/Electrical-properties-of-selected-Formlabs-SLA-resins?language=en_US
- [154] "Dielectric Laboratories - CapCad." Accessed: July 12, 2025. [Online]. Available: <https://capcad.knowledspd.com/Default.aspx>
- [155] J. Murphy-Boesch and A. P. Koretsky, "An *in Vivo* NMR probe circuit for improved sensitivity," *Journal of Magnetic Resonance (1969)*, vol. 54, no. 3, pp. 526–532, Oct. 1983, doi: 10.1016/0022-2364(83)90333-5.
- [156] D. Kajfez and E. J. Hwan, "Q-Factor Measurement with Network Analyzer," *IEEE Transactions on Microwave Theory and Techniques*, vol. 32, no. 7, pp. 666–670, July 1984, doi: 10.1109/TMTT.1984.1132751.
- [157] A. Bitz, J. Felder, and T. Wittig, "Combined 3D Electromagnetic and Spin Response Simulation of MRI Systems," Dassault Systèmes, Vélizy-Villacoublay, France, 2019. [Online]. Available: <https://www.3ds.com/products-services/simulia/products/cst-studio-suite/>
- [158] A. F. Privalov, S. V. Dvinskikh, and H.-M. Vieth, "Coil Design for Large-Volume High-B1 Homogeneity for Solid-State NMR Applications," *Journal of Magnetic Resonance, Series A*, vol. 123, no. 2, pp. 157–160, 1996.
- [159] S. Kamotesov *et al.*, "Modelization and characterization of 2D and 3D mid inductors for multidirectional inductive proximity sensing," in *2016 12th International Congress Molded Interconnect Devices (MID)*, Sept. 2016, pp. 1–6. doi: 10.1109/ICMID.2016.7738936.
- [160] I. F. Ait-Ali, "Développement et intégration de microcapteurs de pH et de température dans des dispositifs microfluidiques polymères," Thèse de doctorat, Lyon 1, 2014. Accessed: Feb. 20, 2025. [Online]. Available: <https://theses.fr/2014LYO10003>
- [161] M.-P. Schmidt, A. Oseev, C. Engel, A. Brose, B. Schmidt, and S. Hirsch, "Flexible free-standing SU-8 microfluidic impedance spectroscopy sensor for 3-D molded interconnect devices application," *Journal of Sensors and Sensor Systems*, vol. 5, no. 1, pp. 55–61, Feb. 2016, doi: 10.5194/jsss-5-55-2016.
- [162] C. Goument *et al.*, "In-Mold Electronics on Poly(Lactic Acid): towards a more sustainable mass production of plastronic devices," *Int J Adv Manuf Technol*, vol. 125, no. 5, pp. 2643–2660, Mar. 2023, doi: 10.1007/s00170-023-10878-4.
- [163] J.-L. Gnanago and *et al.*, "(ISMRM 2022) 3D MRI characterization of 3D printed tumor tissue models using a plastronic MR-Bioreactor : Preliminary results. in Proceedings of the 30th International Society inMagnetic Resonance for Medecine Annual Meeting 2003," 2022. Accessed: Apr. 16, 2024. [Online]. Available: <https://archive.ismrm.org/2022/2003.html>
- [164] B.-N. B. Nguyen, H. Ko, R. A. Moriarty, J. M. Etheridge, and J. P. Fisher, "Dynamic Bioreactor Culture of High Volume Engineered Bone Tissue," *Tissue Engineering Part A*, vol. 22, no. 3–4, pp. 263–271, Feb. 2016, doi: 10.1089/ten.tea.2015.0395.
- [165] F. P. W. Melchels *et al.*, "The influence of the scaffold design on the distribution of adhering cells after perfusion cell seeding," *Biomaterials*, vol. 32, no. 11, pp. 2878–2884, Apr. 2011, doi: 10.1016/j.biomaterials.2011.01.023.

- [166] E. Petiot, “BIOPROCEDES DE CULTURE CELLULAIRE : APPLICATION A LA PRODUCTION DE VIRUS ET A LA MEDECINE REGENERATIVE,” Université Claude Bernard Lyon 1, Lyon, France, 2020. [Online]. Available: <https://tel.archives-ouvertes.fr/tel-04258278>
- [167] E. Capuana *et al.*, “Computational modeling and experimental characterization of fluid dynamics in micro-CT scanned scaffolds within a multiple-sample airlift perfusion bioreactor,” *Biochemical Engineering Journal*, vol. 191, p. 108797, Feb. 2023, doi: 10.1016/j.bej.2022.108797.
- [168] R. J. McCoy and F. J. O’Brien, “Influence of Shear Stress in Perfusion Bioreactor Cultures for the Development of Three-Dimensional Bone Tissue Constructs: A Review,” *Tissue Engineering Part B: Reviews*, vol. 16, no. 6, pp. 587–601, Dec. 2010, doi: 10.1089/ten.teb.2010.0370.
- [169] M. Yushchenko, P. Choquet, N. Salameh, and M. Sarraçanie, “Biplanar quadrature coil for versatile low-field extremity MRI,” *Front. Phys.*, vol. 11, Aug. 2023, doi: 10.3389/fphy.2023.987197.
- [170] “Specialty Components | Knowles Precision Devices.” Accessed: July 12, 2025. [Online]. Available: https://www.knowlescapacitors.com/Tools/TrimmerTool/PartDetail?Product_ID=132
- [171] Formlabs, “High Temp Resin Technical Data Sheet,” Formlabs Inc., Somerville, MA, USA, 2023. [Online]. Available: https://formlabs-media.formlabs.com/datasheets/High_Temp_Technical.pdf
- [172] J. P. Licata, H. Slupianek, S. Rizokulova, J. A. Gerstenhaber, and P. I. Lelkes, “Biocompatibility of 3D printed plastics for use in bioreactors,” *Bioprinting*, vol. 40, p. e00347, July 2024, doi: 10.1016/j.bprint.2024.e00347.
- [173] C. Hart, C. M. Didier, F. Sommerhage, and S. Rajaraman, “Biocompatibility of Blank, Post-Processed and Coated 3D Printed Resin Structures with Electrogenic Cells,” *Biosensors*, vol. 10, no. 11, 2020, doi: 10.3390/bios10110152.
- [174] G. O. Mallory and J. B. Hajdu, *Electroless Plating: Fundamentals and Applications*. Norwich, NY: William Andrew Publishing, 1990.
- [175] H. Gudbjartsson and S. Patz, “The rician distribution of noisy mri data,” *Magnetic Resonance in Medicine*, vol. 34, no. 6, pp. 910–914, Dec. 1995, doi: 10.1002/mrm.1910340618.
- [176] E. K. Insko and L. Bolinger, “Mapping of the Radiofrequency Field,” *Journal of Magnetic Resonance, Series A*, vol. 103, no. 1, pp. 82–85, June 1993, doi: 10.1006/jmra.1993.1133.
- [177] J.-L. Gnanago *et al.*, “Actuators for MRE: New Perspectives With Flexible Electroactive Materials,” *Frontiers in Physics*, vol. Volume 9-2021, 2021, doi: 10.3389/fphy.2021.633848.
- [178] F. Resmer, H. C. Seton, and J. M. S. Hutchison, “Cryogenic receive coil and low noise preamplifier for MRI at 0.01T,” *Journal of Magnetic Resonance*, vol. 203, no. 1, pp. 57–65, Mar. 2010, doi: 10.1016/j.jmr.2009.11.021.
- [179] M. Poirier-Quinot, J.-C. Ginefri, O. Girard, P. Robert, and L. Darrasse, “Performance of a miniature high-temperature superconducting (HTS) surface coil for in vivo microimaging of the mouse in a standard 1.5T clinical whole-body scanner,” *Magnetic Resonance in Medicine*, vol. 60, no. 4, pp. 917–927, Oct. 2008, doi: 10.1002/mrm.21605.
- [180] A. Labbé *et al.*, “Recent Advances and Challenges in the Development of Radiofrequency HTS Coil for MRI,” *Front. Phys.*, vol. 9, July 2021, doi: 10.3389/fphy.2021.705438.

- [181] R. A. Matula, “Electrical resistivity of copper, gold, palladium, and silver,” *Journal of Physical and Chemical Reference Data*, vol. 8, no. 4, pp. 1147–1298, Oct. 1979, doi: 10.1063/1.555614.
- [182] Inc. Aspen Aerogels, “Spaceloft Aerogel Blanket Insulation: Technical Guidance Document (ETA 11_0471, Issue 21),” Aspen Aerogels, Inc., Northborough, MA, USA, 2021. [Online]. Available: <https://www.aerogel.uk.com/spaceloft-tech-guide.pdf>
- [183] “Presidio Components Cryogenic Ceramic Capacitor Products.” Accessed: Sept. 07, 2025. [Online]. Available: <https://presidiocomponents.com/products/Cryogenic-Ceramic-Capacitor-Links.html>
- [184] P. Atkins, J. de Paula, and J. Keeler, *Atkins’ Physical Chemistry*. Oxford University Press, 2022. doi: 10.1093/hesc/9780198847816.001.0001.

Annex I

Detailed Derivation of the Magnetic Vector Potential $\vec{A}_{in}$

The goal here is to derive the expression for the magnetic vector potential, $\vec{A}$, inside a spherical shell of radius R that carries a surface current density $\vec{K} = K \sin\theta' \vec{e}_\phi'$.

The fundamental definition of the magnetic vector potential $\vec{A}$ at a position $\vec{r}$ due to a surface current $\vec{K}$ flowing on a surface S is given by:

$$\vec{A}(\vec{r}) = \frac{\mu_0}{4\pi} \int_S \frac{\vec{K}(\vec{r}')}{|\vec{r} - \vec{r}'|} da'$$

Where μ_0 is the permeability of free space, $\vec{r}'$ is the position vector to a point on the surface S where the current element is, $\vec{r}$ is the position vector to the point where the potential is calculated and da' is the differential surface area element on the sphere.

The main difficulty in solving this integral lies in the term $|\vec{r} - \vec{r}'|$.

To handle the distance term, a multipole expansion for $1/|\vec{r} - \vec{r}'|$ in terms of Legendre polynomials can be used. Since only the potential inside the sphere is considered, the expansion valid for $r < R$ is written:

$$\frac{1}{|\vec{r} - \vec{r}'|} = \sum_{l=0}^{\infty} \frac{r^l}{R^{l+1}} P_l(\cos\gamma)$$

Here, γ is the angle between the vectors $\vec{r}$ and $\vec{r}'$. A crucial insight for this specific problem is that the current distribution $\vec{K} = K \sin\theta' \vec{e}_\phi'$ can be considered as a pure **dipole source** [cite]. This means that all terms in the multipole expansion except for the $l = 1$ term will integrate to zero due to symmetry and orthogonality.

Therefore, only the $l = 1$ term needs to be considered:

$$\frac{1}{|\vec{r} - \vec{r}'|} \approx \frac{r}{R^2} P_1(\cos\gamma)$$

Since $P_1(x) = x$, $P_1(\cos\gamma) = \cos\gamma$. The term $\cos\gamma$ is simply the dot product of the unit vectors: $\cos\gamma = \vec{e}_r \cdot \vec{e}_r'$.

Substituting the simplified $l = 1$ term into the main equation for $\vec{A}$ becomes:

$$\vec{A}(\vec{r}) = \frac{\mu_0}{4\pi} \int_S \frac{\vec{K}(\vec{r}')}{R^2} r (\vec{e}_r \cdot \vec{e}_r') da'$$

Expressing the vectors and differential area in coordinates gives:

- Current Density: $\vec{K}(\vec{r}') = K \sin\theta' \vec{e}_\phi'$
- Position Vectors: $\begin{cases} \vec{e}_r = \sin\theta \cos\phi \vec{e}_x + \sin\theta \sin\phi \vec{e}_y + \cos\theta \vec{e}_z \\ \vec{e}_r' = \sin\theta' \cos\phi' \vec{e}_x + \sin\theta' \sin\phi' \vec{e}_y + \cos\theta' \vec{e}_z \end{cases}$
- Dot Product: $\vec{e}_r \cdot \vec{e}_r' = \sin\theta \sin\theta' \cos(\phi - \phi') + \cos\theta \cos\theta'$

- Differential Area: $da' = R^2 \sin\theta' d\theta' d\phi'$

To make the integration manageable, the current vector $\vec{K}$ is resolved into its Cartesian components:

$$\begin{aligned}\vec{e}_{\phi'} &= -\sin\phi' \vec{e}_x + \cos\phi' \vec{e}_y \\ \vec{K}(\vec{r}') &= K \sin\theta' (-\sin\phi' \vec{e}_x + \cos\phi' \vec{e}_y)\end{aligned}$$

A component-wise integration can therefore be performed by computing the integral for each Cartesian component of $\vec{A} = A_x \vec{e}_x + A_y \vec{e}_y + A_z \vec{e}_z$.

- **z-component (A_z)**

The current $\vec{K}$ has no z-component, so $A_z = 0$.

- **x-component (A_x)**

$$\begin{aligned}A_x &= \frac{\mu_0 r}{4\pi} \int_0^{2\pi} \int_0^\pi \overbrace{(-K_0 \sin\theta' \sin\phi')}^{K_x} \overbrace{(\vec{e}_r \cdot \vec{e}_r')}^{\text{dot product}} \overbrace{\sin\theta' d\theta' d\phi'}^{\text{area}/R^2} \\ A_x &= \frac{-\mu_0 r K}{4\pi} \int_0^{2\pi} \int_0^\pi \sin^2 \theta' \sin\phi' (\sin\theta \sin\theta' \cos(\phi - \phi') + \cos\theta \cos\theta') d\theta' d\phi'\end{aligned}$$

Integrate over ϕ' first. The integral contains two terms:

$$\begin{aligned}\int_0^{2\pi} \sin\phi' \cos(\phi - \phi') d\phi' &= \int_0^{2\pi} \sin\phi' (\cos\phi \cos\phi' + \sin\phi \sin\phi') d\phi' \\ &= \sin\phi \int_0^{2\pi} \sin^2 \phi' d\phi' = \pi \sin\phi \\ \int_0^{2\pi} \sin\phi' d\phi' &= 0\end{aligned}$$

Substituting this back, the second part of the main integral disappears:

$$A_x = \frac{-\mu_0 r K}{4\pi} (\pi \sin\phi) \int_0^\pi \sin\theta \sin^3 \theta' d\theta'$$

Evaluate the θ' integral gives: $\int_0^\pi \sin^3 \theta' d\theta' = 4/3$.

$$A_x = \frac{-\mu_0 r K}{4\pi} (\pi \sin\phi) (\sin\theta) \left(\frac{4}{3}\right) = -\frac{\mu_0 K}{3} r \sin\theta \sin\phi$$

- **y-component (A_y)**

The process is identical for A_y :

$$A_y = \frac{\mu_0 r K}{4\pi} \int_0^{2\pi} \int_0^\pi \sin^2 \theta' \cos\phi' (\sin\theta \sin\theta' \cos(\phi - \phi') + \cos\theta \cos\theta') d\theta' d\phi'$$

The relevant ϕ' integrals are:

$$\begin{aligned}
 \int_0^{2\pi} \cos \phi' \cos(\phi - \phi') d\phi' &= \int_0^{2\pi} \cos \phi' (\cos \phi \cos \phi' + \sin \phi \sin \phi') d\phi' \\
 &= \cos \phi \int_0^{2\pi} \cos^2 \phi' d\phi' = \pi \cos \phi \\
 \int_0^{2\pi} \cos \phi' d\phi' &= 0
 \end{aligned}$$

Substituting this back:

$$\begin{aligned}
 A_y &= \frac{\mu_0 r K}{4\pi} (\pi \cos \phi) \int_0^\pi \sin \theta \sin^3 \theta' d\theta' \\
 A_y &= \frac{\mu_0 r K}{4\pi} (\pi \cos \phi) (\sin \theta) \left(\frac{4}{3}\right) = \frac{\mu_0 K}{3} r \sin \theta \cos \phi
 \end{aligned}$$

- **Recombination and Final Result**

The components to form the vector $\vec{A}$ are recombined to give:

$$\begin{aligned}
 \vec{A} &= A_x \vec{e}_x + A_y \vec{e}_y = \frac{\mu_0 K}{3} r \sin \theta (-\sin \phi \vec{e}_x + \cos \phi \vec{e}_y) \\
 \boxed{\vec{A}_{in}(r, \theta, \phi) &= \frac{\mu_0 K}{3} r \sin \theta \vec{e}_\phi}
 \end{aligned}$$

Annex II

Varying-Pitch Solenoid Trajectory (end-tapered)

The solenoid trajectory is kept circular in (x, y) while the axial coordinate is generated by a pitch-per-radian function $c(t) > 0$:

$$\begin{cases} x(t) = R \cos(t) \\ y(t) = R \sin(t) \\ z(t) = Z(t) - \frac{L}{2}, \quad Z'(t) = c(t) \end{cases} \quad (\text{VII-1})$$

with end constraints

$$Z(0) = 0, \quad Z(2\pi N) = L \Leftrightarrow \int_0^{2\pi N} c(t) dt = L.$$

Introduce the normalized winding coordinate $s := t/(2\pi N) \in [0, 1]$ and select a smooth, symmetric end-taper profile that increases turn density near $z = \pm L/2$:

$$c(s) = \bar{c} [1 - \alpha \cos^2(\pi s)], \quad 0 \leq \alpha < 1,$$

where α controls the end loading ($\alpha = 0$ gives a uniform pitch). The normalization $\bar{c}$ follows from EQ:

$$\bar{c} = \frac{L}{(2\pi N)(1 - \alpha/2)}.$$

A closed form for $Z(t)$ is obtained by integration. With $s = t/(2\pi N)$,

$$Z(t) = L \left[s - \frac{\alpha}{4\pi(1 - \alpha/2)} \sin(2\pi s) \right], \quad z(t) = Z(t) - \frac{L}{2}.$$

This end-tapered parametrization preserves the circular path in (x, y) while re-parametrizing $z(t)$ to obtain a higher turn density at the coil ends, compensating “edge effect” and improving B_1 homogeneity in the central region.

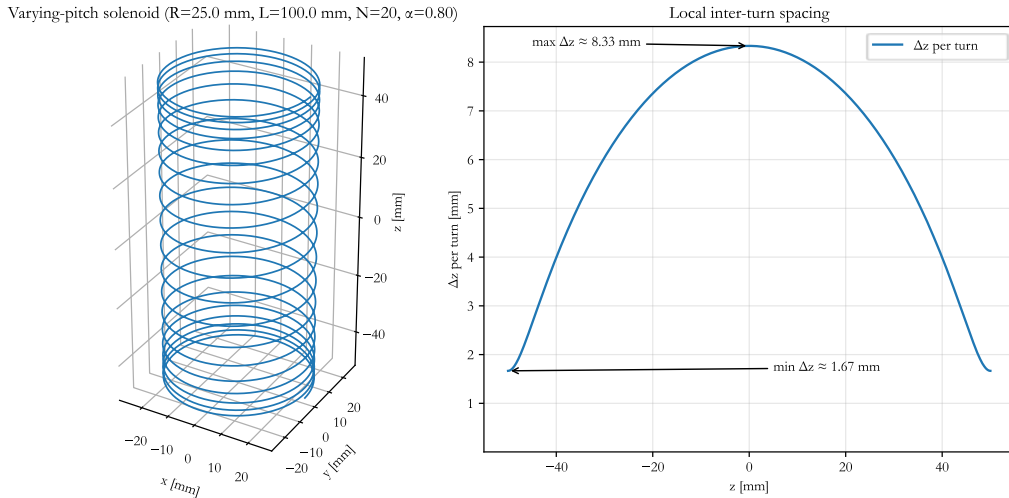


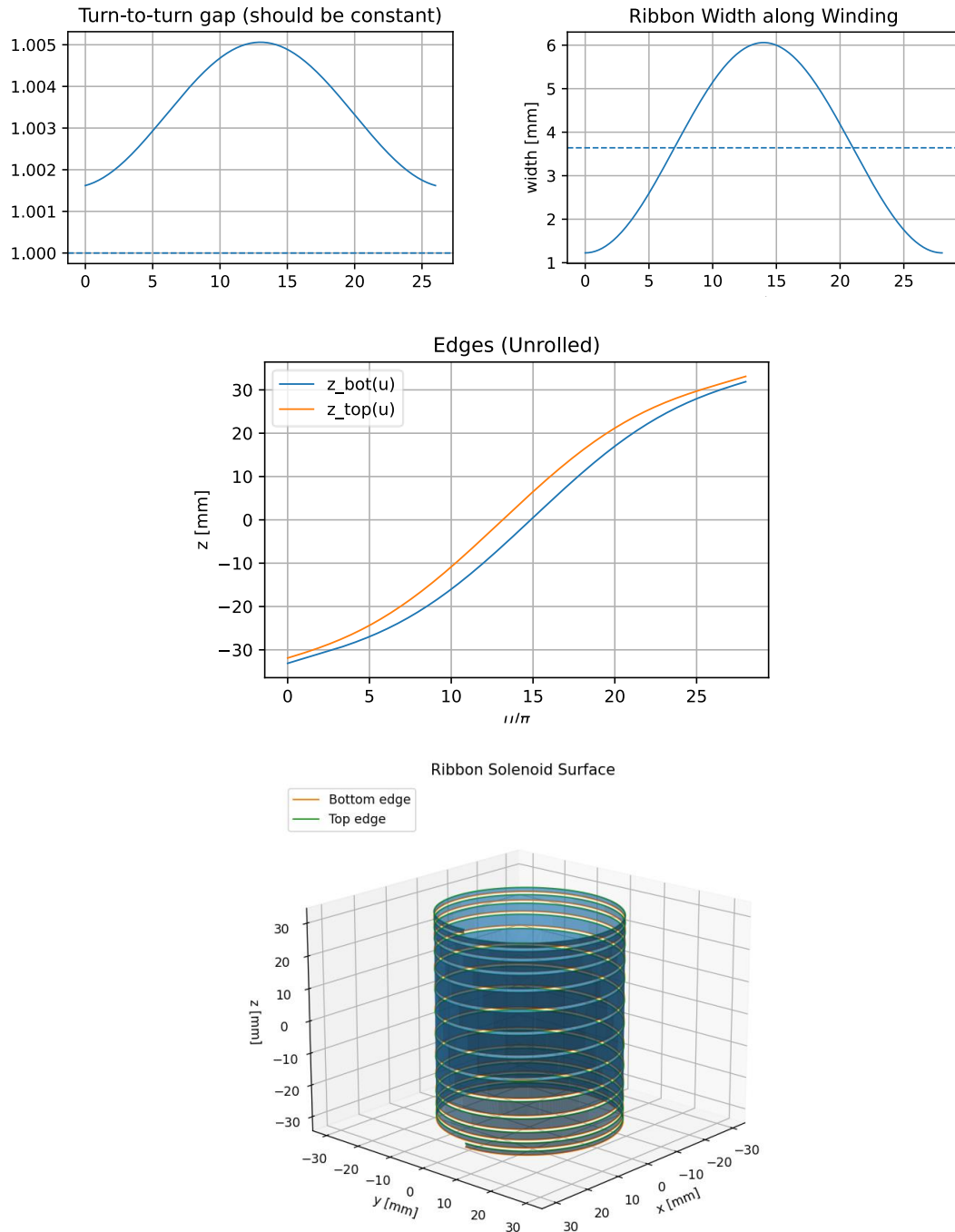
Figure 107: Varying-Pitch Solenoid Trajectory and Spacing

Useful Quantities	
Local pitch (per radian)	$c(t) = \frac{L}{2\pi N} \frac{1 - \alpha \cos^2(\pi s)}{1 - \alpha/2}, \quad s = \frac{t}{2\pi N}$
Local turns per unit length (approx.)	$n(z) \approx \frac{1}{2\pi c(t)}$
Minimum inter-turn spacing (near the ends)	$\Delta z_{\min} \approx 2\pi c_{\min} = \frac{L}{N} \frac{1 - \alpha}{1 - \alpha/2}$

Table 18: Varying-Pitch Solenoid Trajectory Useful Quantities

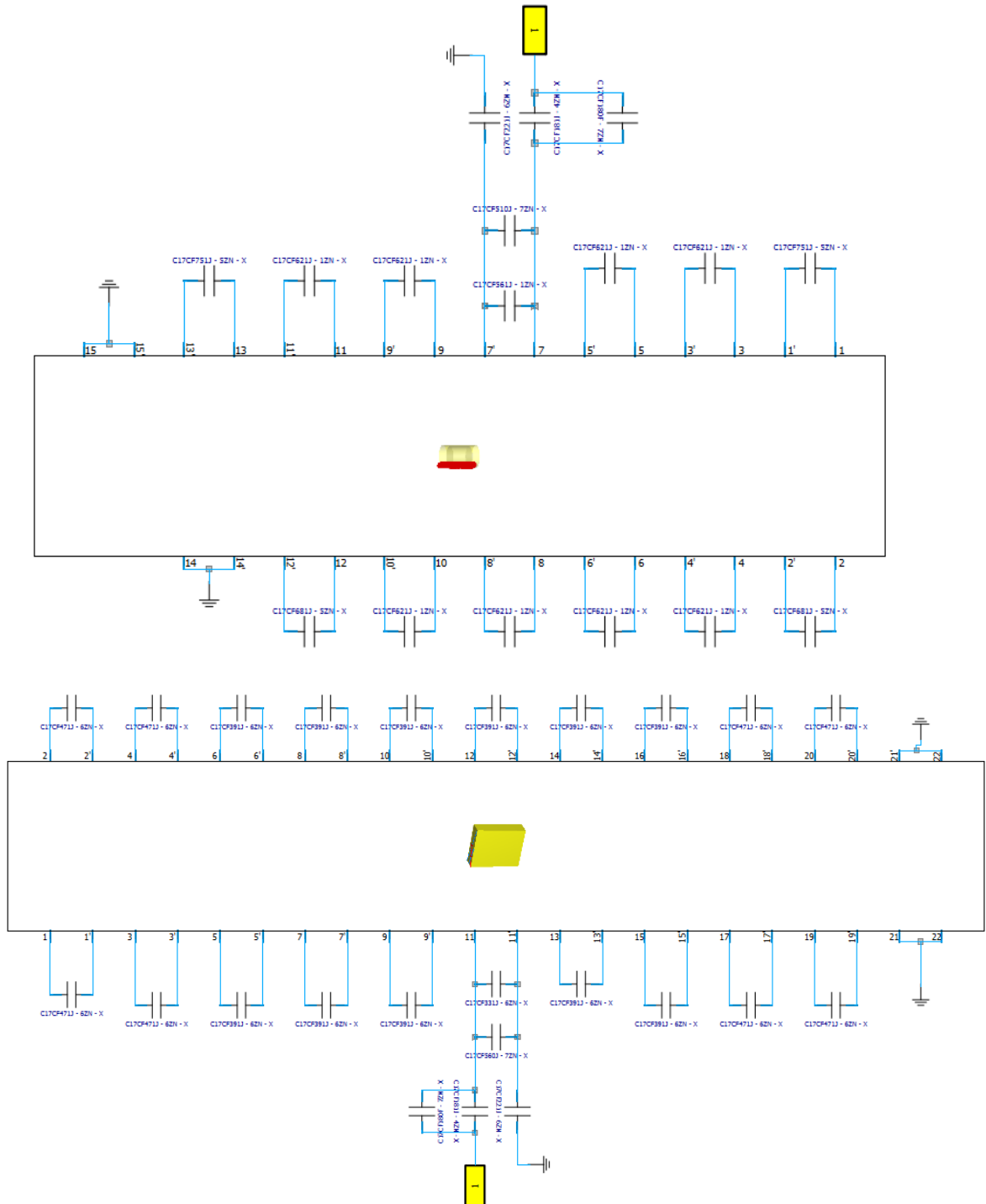
Annex III

Ribbon-Solenoid Surface (Parametric Definition and Constant-Gap Construction)



Annex IV

Detailed Co-Simulation Schematics



Annex V

Evidencing the Impact of The Shielded Enclosure

To evidence the impact of the shielded enclosure, a short experiment was performed. A 3D printed test tube (D 20mm L 60mm) filled with saline (0.9% NaCl) was placed in the provided benchmark coil and the unshielded fabricated 3D Plastronics culture chamber. Acquisitions were performed using low resolution Spin Echo sequences:

- **Ultra-Low Resolution (1.88 mm isotropic):** TE 8, TR 100, Acquisition Time 26s
- **Low Resolution (94 mm isotropic):** TE 20, TR 100, Acquisition Time 3m 25s

Noise Figures and SNR values were estimated using the method described in Chapter III Section V.D.

Results:

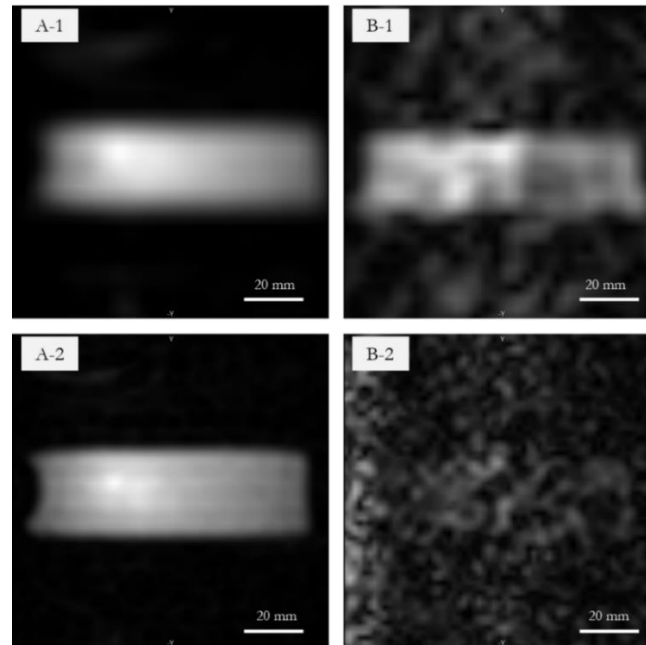


Figure 108: Spin Echo MRI acquisitions comparing signal quality and noise performance between a shielded commercial benchmark coil (left) and an unshielded 3D Plastronics coil (right). A cylindrical saline-filled phantom ($\varnothing$ 20 mm, L 60 mm) was imaged using ultra-low resolution (A-1, B-1) and low resolution (A-2, B-2) SE sequences. The shielded coil yields markedly higher SNR and lower noise background across both protocols. Imaging with the unshielded coil at low resolution yielded excessive noise and barely noticeable phantom noticeable.

Coil Type	Measured Noise Factor (dB)	Mean SNR (Ultra-Low Res)	Mean SNR (Low Res)	Estimated SNR Loss (Unshielded)
Shielded Commercial Coil	2.30 dB ($F = 1.7$)	~ 31	~ 44	Reference
Unshielded 3D Plastronics Coil	42.38 dB ($F = 17,337.5$)	~ 6	Measurement Impossible	$\sim \times 5$ lower

Table 19: Quantitative Comparison of Measured Noise and SNR Performance with and without Shielded Enclosure

Annex VI

Theoretical Estimation of Electrical Properties of 0.9% NaCl and 16.3 mM Copper(II) Nitrate Solution (25 °C)

1. Scope

This annex summarizes the electrical and dielectric properties of two representative aqueous electrolytes used for coil loading and tissue-mimicking calibration: (i) a 0.9 % w/v sodium chloride solution (154 mM, 1:1 electrolyte), and (ii) a 16.3 mM copper(II) nitrate trihydrate solution (2:1 electrolyte). All quantities are evaluated at 25 °C using the extended Onsager relation with parameters tabulated by *Atkins et al.* (2022) and standard solvent properties of water at 298 K.

2. Theoretical background

For strong electrolytes at low to moderate concentrations, the molar conductivity decreases with the square root of concentration according to the Debye–Hückel–Onsager law:

$$\Lambda = \Lambda^0 - (B_1 \Lambda^0 + B_2) \sqrt{c},$$

where Λ is the molar conductivity ($\text{S cm}^2 \text{ mol}^{-1}$), Λ^0 the limiting molar conductivity, c the molarity (mol L^{-1}), and B_1, B_2 solvent-dependent coefficients. For water at 25 °C, *Atkins' Physical Chemistry* [184] reports the following constants:

- 1:1 electrolytes (e.g. NaCl): $B_1 = 0.229$, $B_2 = 60.2 \text{ S cm}^2 \text{ mol}^{-1} \text{ M}^{-1/2}$;
- 2:1 electrolytes (e.g. $\text{Cu}(\text{NO}_3)_2$): $B_1 = 0.778$, $B_2 = 69.8 \text{ S cm}^2 \text{ mol}^{-1} \text{ M}^{-1/2}$.

These constants incorporate the viscosity and permittivity of water and are directly applicable at 298 K. The electrical conductivity (S m^{-1}) then follows from

$$\kappa = \Lambda c = \Lambda \times 10^{-4} \times 1000 c = 0.1 \Lambda c,$$

where Λ is in $\text{S cm}^2 \text{ mol}^{-1}$ and c in mol L^{-1} .

The loss tangent in the radio-frequency range dominated by conduction is given by

$$\tan \delta = \frac{\sigma}{\omega \epsilon_0 \epsilon'},$$

with $\omega = 2\pi f$, $\epsilon_0 = 8.854 \times 10^{-12} \text{ F m}^{-1}$, and ϵ' the relative permittivity of the solution (≈ 78 for dilute aqueous media).

3. Numerical evaluation

Property	Symbol	NaCl (0.9 %)	$\text{Cu}(\text{NO}_3)_2$ (16.3 mM)	Units
Electrolyte type		1:1	2:1	
B_1		0.229	0.778	—
B_2		60.2	69.8	$\text{S cm}^2 \text{ mol}^{-1} \text{ M}^{-1/2}$
Λ^0		126.4	250.0	$\text{S cm}^2 \text{ mol}^{-1}$
Concentration	c	0.154	0.0163	mol L^{-1}

$\sqrt{c}$		0.393	0.128	$M^{1/2}$
Molar conductivity	Λ	91.4	216.0	$S\text{ cm}^2\text{ mol}^{-1}$
Conductivity	κ	1.41	0.35	$S\text{ m}^{-1}$
Relative permittivity	ε'	77.8	77.9	—
Loss tangent (12.77 MHz)	$\tan\delta$	25.6	6.3	—

Table 20 :Electrical properties of the two electrolyte solutions at 25 °C and $f = 13$ MHz.

Annex VII

Proposed Improved 3D Plastronics Culture Chamber Design for SNR Maximization

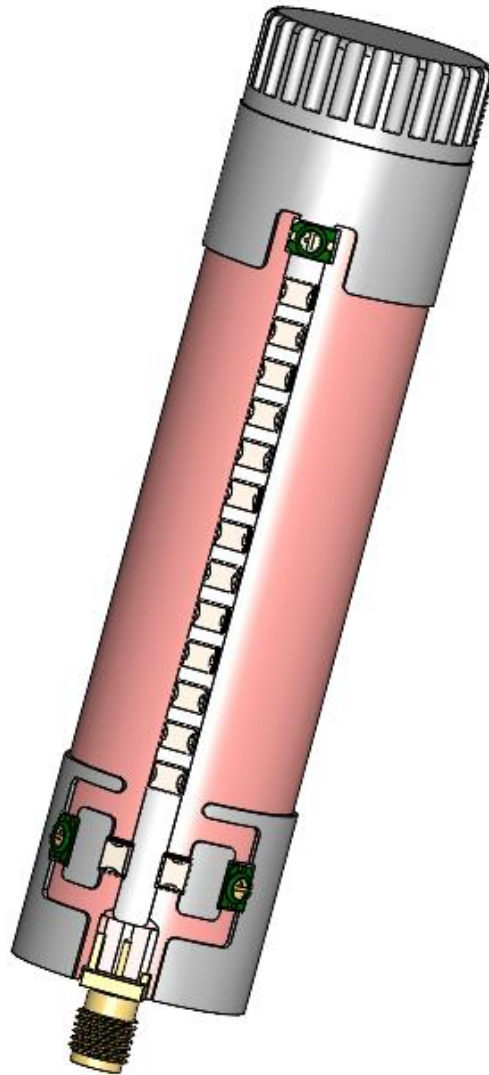


Figure 109 : Proposed Improved 3D Plastronics Bioreactor Design

Annex VIII: Résumé étendu en Français

1. Chapitre I : Etat de l'art et Motivation

1.1. Introduction

La complexité croissante des tissus 3D de grande taille pose des défis importants en matière de caractérisation et d'observation. En effet, les modalités optiques actuelles (par exemple, un microscope) fournissent des informations moléculaires et cellulaires à haute résolution, mais sont limitées à des échantillons de petites tailles sans aucune profondeur de pénétration. Ces limites ont rendu nécessaire la mise au point de nouveaux outils dédiés à l'ingénierie tissulaire 3D. Ces derniers doivent être capables d'effectuer une caractérisation complexe et détaillée, y compris une résolution subcellulaire à macroscopique, et d'évaluer toutes les informations pertinentes couvrant les changements morphologiques, physiologiques et moléculaires au sein de l'échantillon. Ils doivent également permettre une observation sûre, quantitative et non invasive avec des profondeurs de pénétration adaptées à l'épaisseur de l'échantillon. Jusqu'à présent, quelques candidats prometteurs ont été proposés [13]. Par exemple, les rayons X et la micro-tomographie par ordinateur offrent une excellente résolution et un excellent contraste pour les structures minéralisées. Cependant, ils restent limités dans l'imagerie des tissus mous de taille moyenne en raison de l'exposition aux rayonnements et du faible contraste [15], [16]. Les ultrasons offrent une imagerie en temps réel des propriétés mécaniques et de l'intégrité des scaffolds, mais ils sont limités par le contraste restreint des tissus mous [17], [18]. La TEP (tomographie par émission de positons) et la SPECT (tomographie par émission monophotonique) se concentrent sur l'imagerie fonctionnelle et biochimique pour une large gamme de tailles d'échantillons. Elles offrent une bonne sensibilité moléculaire, mais manquent de résolution spatiale et nécessitent des marqueurs ionisants [19], [20]. En revanche, l'imagerie par résonance magnétique (IRM) se distingue par sa capacité unique d'imagerie non invasive à haute résolution de toutes les tailles d'échantillons, avec un excellent contraste des tissus mous, tout en fournissant simultanément des informations anatomiques, physiologiques et microstructurelles détaillées [15].

1.2. L'IRM, une modalité de choix pour l'ingénierie tissulaire

Parmi les méthodes d'imagerie disponibles, l'imagerie par résonance magnétique (IRM), bien qu'elle ne dispose pas de la résolution au niveau cellulaire de la microscopie optique ni de la haute sensibilité de l'imagerie nucléaire, est la seule technique qui permet une évaluation tridimensionnelle non invasive de la structure et de la fonction des tissus situés en profondeur dans des constructions opaques à la lumière. Cela la rend particulièrement adaptée aux applications d'ingénierie tissulaire 3D. Bien que les méthodes optiques fournissent des détails subcellulaires inégalés et que la TEP permet de suivre des voies moléculaires spécifiques, elles ne peuvent toutefois pas fournir une vue complète des tissus denses et de grande taille. L'IRM surmonte ce défi critique de la profondeur de pénétration dans l'échantillon sans utiliser de rayonnements ionisants, ce qui permet d'étudier longitudinalement la même structure tissulaire tout au long de son développement. De plus, la puissance de l'IRM réside dans sa nature multiparamétrique. À partir d'une seule session d'imagerie, elle peut fournir des cartes anatomiques détaillées révélant la distribution des cellules et l'intégrité de la structure, tandis que des techniques avancées telles que l'IRM de diffusion et l'IRM de perfusion peuvent fournir des informations fonctionnelles et quantitatives cruciales, respectivement, sur la cellularité et la gestion des flux de nutriments/déchets. Cet ensemble de données quantitatives riches permet d'établir une corrélation entre la croissance structurelle et la viabilité fonctionnelle. Il offre une compréhension complète de la santé et du développement du tissu artificiel. Par conséquent, malgré ses limites en termes de résolution et de coût, l'IRM est un outil indispensable pour caractériser et optimiser les processus biologiques complexes qui entrent en jeu dans l'ingénierie tissulaire à grande échelle.

1.3. Évaluation des propriétés des structures tissulaires complexes à l'aide d'une boîte à outils IRM multiparamétrique

Il existe plusieurs techniques IRM qui peuvent être regroupées dans une « boîte à outils IRM » et ainsi adaptées aux différents besoins d'observation en matière d'ingénierie tissulaire. Le Table 3 récapitule certaines techniques IRM. La technique d'analyse IRM est choisie en fonction des propriétés tissulaires à identifier et à analyser.

Propriété tissulaire évaluée	Technique IRM	Références
Teneur en eau des tissus, maturité et structure des tissus. Le rapport T2/T1 peut être un indicateur de la rigidité des tissus.	Cartographie T1 et T2	[15], [61], [62]
Restriction du mouvement de l'eau par la microstructure tissulaire, la croissance tissulaire, la porosité, le flux nutritif, la cellularité et la tension d'oxygène.	Imagerie pondérée en diffusion	[15], [56], [57]
Vascularisation, distribution et gestion du flux nutritif, contrainte de cisaillement	Perfusion	[63], [64]
Propriétés mécaniques, notamment la rigidité et la viscoélasticité des tissus.	MRE (élastographie par résonance magnétique)	[60], [65]

Tableau 21 : Techniques IRM et propriétés tissulaires correspondantes pour de l'observation en ingénierie tissulaire

Bien qu'elles ne soient pas présentées ici, il existe de nombreuses autres techniques. De manière non exhaustive, il est possible de citer les méthodes par contrastes $T_{1\rho}$ et les mesures de transfert de magnétisation (MT) pour évaluer le contenu macromoléculaire de la matrice [15], [66], l'imagerie par tenseur de diffusion (DTI) pour évaluer l'anisotropie des tissus et l'organisation des fibres [67], [68], [69], ou encore le marquage par oxyde de fer superparamagnétique (SPIO- Labelling) pour la localisation et le suivi des cellules [70].

1.4. Surmonter les limites de l'IRM pour une utilisation en laboratoire

L'une des principales limites qui empêchent une utilisation plus large de l'IRM dans l'ingénierie tissulaire en laboratoire tient à la nature même des systèmes d'IRM. Les scanners IRM conventionnels sont coûteux, avec des frais d'acquisition et de maintenance élevés, atteignant souvent plusieurs millions d'euros [71]. Ce poids financier limite l'accès à l'IRM aux centres d'imagerie spécialisés et aux grands hôpitaux, ce qui la rend peu pratique pour les environnements de laboratoire courants. De plus, les systèmes IRM nécessitent généralement une infrastructure dédiée : des salles blindées, des équipements cryogéniques pour les aimants supraconducteurs et des protocoles de sécurité stricts pour gérer les champs magnétiques puissants. L'encombrement physique important et la complexité technique de ces installations constituent des obstacles importants à leur intégration dans les laboratoires d'ingénierie tissulaire, où l'espace est souvent limité et où les flux de travail exigent une grande flexibilité. Les questions de sécurité compliquent encore davantage la situation. Les champs magnétiques statiques élevés présentent des risques tant pour les opérateurs que pour les équipements, ce qui nécessite un contrôle rigoureux et limite la

compatibilité avec de nombreux appareils de laboratoire. En conséquence, malgré ses puissantes capacités, l'IRM reste sous-utilisée.

1.5. L'IRM de paillasse comme solution pratique

Les scanners IRM de table (ou de bureau) à faible champ sont apparus comme une catégorie d'instruments polyvalents. Conçus pour s'intégrer dans les environnements de laboratoire ou de salle de classe, ces systèmes compacts donnent à l'IRM de nombreuses possibilités au-delà des installations cliniques et de recherche traditionnelle. En assouplissant les exigences en matière de taille, de blindage et de cryogénie, les scanners de table offrent une solution pratique pour une variété d'applications spécialisées et émergentes.

Les scanners de table fonctionnent généralement dans une plage de 0,01 à 1 Tesla, ce qui est inférieur à l'IRM clinique standard (1,5 à 3 T). Ces systèmes compacts permettent d'utiliser l'IRM dans les laboratoires, les salles de classe et les points de soins en évitant l'infrastructure encombrante des machines à champ élevé. La réduction de l'intensité du champ s'accompagne d'un rapport Signal sur Bruit plus faible, mais les progrès modernes en matière de conception des aimants et d'électronique ont permis d'obtenir des images exploitables même à des intensités de quelques dizaines ou centaines de milli-tesla [26]. Les appareils IRM de table sont généralement autonomes et plus faciles à installer, ne nécessitant souvent pas de salles blindées spéciales ni de systèmes cryogéniques complexes. Beaucoup de ces systèmes présentent également des configurations ouvertes, qui améliorent leur polyvalence en facilitant l'accès aux échantillons, en s'adaptant à divers types de spécimens et en permettant l'intégration d'outils analytiques complémentaires.

1.6. Le défi du rapport Signal sur Bruit (SNR)

En IRM, la qualité des images obtenues est fondamentalement dictée par le signal de résonance magnétique (RM) et la présence de bruit. Si de nombreux facteurs, notamment la résolution spatiale, la pondération du contraste et la vitesse d'imagerie, influencent l'acquisition IRM, aucun indicateur ne reflète plus directement la qualité de l'image acquise que le rapport Signal sur Bruit (SNR). Le SNR reflète l'équilibre entre l'amplitude du signal RM induit et le bruit de fond aléatoire provenant à la fois de l'échantillon et du système de détection.

1.6.1 Bruit en IRM

Les charges dans un milieu conducteur ne sont jamais complètement immobiles. À toute température différente de zéro, les collisions aléatoires avec le réseau cristallin environnant entraînent les charges dans un mouvement de type brownien. L'application de l'équation stochastique de Langevin à une seule charge révèle que ce mouvement microscopique donne lieu à deux effets indissociables : les « collisions » aléatoires des atomes du réseau cristallin produisent des fluctuations (la force de Langevin), tandis que les « frottements thermiques » dans l'environnement entraîne une dissipation d'énergie. Le théorème de fluctuation-dissipation explique qu'à l'équilibre thermodynamique, les fluctuations spontanées doivent s'accompagner d'un mécanisme de dissipation équivalent. L'ouvrage *Introduction to Solid State Physics* de Charles Kittel présente un aperçu complet de ce résultat [97].

Expérimentalement, Nyquist et Johnson ont confirmé que l'agitation thermique dans une résistance génère un bruit de tension et de courant mesurable dans des conditions stables. Ils ont démontré que le bruit de tension efficace (σ_{noise}) à travers une résistance (R) à la température (T), mesuré sur une bande passante Δf , obéit à la relation suivante :

$$\sigma_{noise} = \sqrt{4 k_B R T \Delta f} \quad (0.1)$$

Où k_B est la constante de Boltzmann et σ_{noise}^2 représente la variance de la tension du bruit.

Le bruit capté par un résonateur RF dans une caractérisation IRM est régi par le même principe. En tenant compte des différents mécanismes de pertes en jeu, il peut être exprimé comme suit :

$$\sigma_{noise} = \sqrt{4 k_B R_{eff} T_{eff} \Delta f} \quad (0.2)$$

Où R_{eff} est la résistance effective observée aux bornes du résonateur, qui inclut toutes les pertes du résonateur telles que les pertes ohmiques intrinsèques du conducteur et les pertes dépendantes de la fréquence résultant des effets diélectriques et de rayonnement dans l'échantillon ; T_{eff} est la température de bruit effective, une moyenne pondérée des températures physiques du résonateur et de l'échantillon ; Δf est la bande passante de bruit effective de la chaîne d'acquisition.

Cependant, comme la chaîne d'acquisition n'est pas idéale, elle ajoute son propre bruit caractérisé par un facteur de bruit F , qui multiplie la puissance du bruit par F (et la tension efficace par $\sqrt{F}$). Le bruit efficace total est donc :

$$\sigma_{noise} = \sqrt{F} \cdot \sqrt{4 k_B R_{eff} T_{eff} \Delta f}, \quad (0.3)$$

1.6.2 Définition du rapport Signal sur Bruit

Par définition, le SNR pour un seul voxel est :

$$SNR = \frac{V_{signal}}{\sigma_{noise}} \quad (0.4)$$

Où (V_{signal}) est la tension de signal induite maximale par unité de courant (I.34), et (σ_{noise}) est le bruit provenant principalement des contributions thermiques du résonateur et de l'échantillon.

Par conséquent, le rapport Signal sur Bruit pour un seul balayage donne :

$$SNR_{single} = \frac{V_{signal}}{\sigma_{noise}} = F^{-1/2} \frac{\omega_0 (B_1/I) M_{\perp} \delta V}{\sqrt{4 k_B R_{eff} T_{eff} \Delta f}}. \quad (0.5)$$

De plus, le calcul de la moyenne de N scans identiques améliore le rapport Signal sur Bruit de ($\sqrt{N}$). Pour un temps de répétition T_R et un temps d'acquisition total (t_{acq}), ($N = t_{acq}/T_R$) ce qui donne :

$$\text{SNR} = F^{-1/2} \frac{\omega_0 (B_1/I) M_{\perp} \delta V}{\sqrt{4 k_B R_{\text{eff}} T_{\text{eff}} \Delta f}} \sqrt{\frac{t_{\text{acq}}}{T_R}}. \quad (0.6)$$

Enfin, Δf représente la bande passante totale de réception divisée par l'intervalle d'échantillonnage dans l'espace fréquentiel. Dans le domaine temporel, ce concept se traduit par le temps d'observation, ce qui correspond à la durée pendant laquelle le signal est acquis. Selon la dualité temps-fréquence de la transformée de Fourier, la résolution fréquentielle est inversement proportionnelle à la fenêtre d'acquisition. Par conséquent, la bande passante peut être exprimée de manière équivalente comme suit $\Delta f = 1/t_o$ où (t_o) est le temps d'observation par pixel :

$$\text{SNR} = F^{-1/2} \frac{\omega_0 (B_1/I) M_{\perp} \delta V}{\sqrt{4 k_B R_{\text{eff}} T_{\text{eff}}}} \sqrt{\frac{t_o \cdot t_{\text{acq}}}{T_R}}. \quad (0.7)$$

1.6.3 Optimisation du rapport Signal sur Bruit pour la microscopie par résonance magnétique

Par rapport à d'autres modalités d'imagerie telles que la microscopie optique, l'imagerie par résonance magnétique (IRM) est fondamentalement limitée par des niveaux de signal intrinsèquement faibles. Seule une petite fraction des spins nucléaires contribue à la magnétisation nette détectable par le résonateur RF, car la différence de population entre les états de spin à l'équilibre thermique est minime. En conséquence, le signal RM est intrinsèquement faible, tandis que le bruit thermique provenant du résonateur et de l'échantillon reste important.

Cette limite inhérente devient encore plus prononcée en microscopie à résonance magnétique (MRM), où de petits volumes d'échantillons et une résolution spatiale extrêmement élevée sont requis. Dans ce contexte, le signal détectable est proportionnel au volume du voxel, qui doit être considérablement réduit pour obtenir une résolution à l'échelle du micron, tandis que le bruit reste inchangé. À mesure que la résolution augmente, le rapport Signal sur Bruit se détériore rapidement, ce qui constitue une contrainte fondamentale pour la qualité de l'image.

Le principal défi du rapport Signal sur Bruit en microscopie RM est d'obtenir un signal suffisant pour générer des images informatives et haute résolution tout en préservant les détails structuraux fins nécessaires à la caractérisation quantitative des tissus artificiels et des petits échantillons biologiques.

La relation SNR présentée ci-dessus peut s'écrire comme suit :

$$\text{SNR} \propto \delta V \times S_{\text{eq}} \times \sqrt{t_{\text{acq}}} \quad (0.8)$$

Cela montre que l'augmentation du temps d'acquisition total (t_{acq}), de la taille des voxels (δV) ou de la sensibilité de l'équipement (S_{eq}) améliorera le rapport Signal sur Bruit global. L'optimisation du rapport Signal sur Bruit est une tâche multiforme qui implique des compromis entre les capacités matérielles, les paramètres d'acquisition et les besoins expérimentaux.

- Augmentation du temps d'acquisition : la moyenne de plusieurs scans réduit le bruit aléatoire grâce à $(\sqrt{N})$, ce qui améliore le rapport Signal sur Bruit. Cette approche est simple et n'affecte pas la résolution spatiale ni la pondération du contraste. Cependant, l'augmentation de la durée du scan n'est pas toujours pratique, car les considérations de débit et les contraintes de temps spécifiques à l'application imposent souvent des limites dans les contextes cliniques et expérimentaux.
- Augmentation de la taille des voxels : l'augmentation du volume des voxels améliore directement le rapport Signal sur Bruit, car le signal RM évolue de manière linéaire avec (δV) . Cependant, cela se fait au détriment de la résolution spatiale. Dans le contexte de l'imagerie microscopique de tissus artificiels ou lorsque les détails structuraux sont essentiels, l'augmentation de la taille des voxels compromet la capacité à distinguer les caractéristiques fines.
- Augmentation de la sensibilité de l'équipement :
 - Intensité du champ statique : le signal RM évolue de manière globalement linéaire avec l'intensité du champ magnétique statique ($\mathbf{B}_0$), tandis que le bruit augmente plus lentement, ce qui se traduit par un meilleur rapport Signal sur Bruit à des intensités plus élevées ($\mathbf{B}_0$). Cependant, l'augmentation de l'intensité du champ n'est pas sans inconvénient. Le coût et la complexité des aimants à champ plus élevé augmentent fortement avec ($\mathbf{B}_0$), rendant ces mises à niveau économiquement inatteignables. De plus, des champs plus puissants nécessitent un blindage plus sophistiqué, des infrastructures plus conséquentes et des protocoles de sécurité plus stricts. Au-delà des considérations financières, une ($\mathbf{B}_0$) plus élevée modifie également les temps de relaxation et peut exacerber les artefacts de susceptibilité. Dans de nombreuses configurations d'IRM à bas champ, y compris celles conçues pour l'ingénierie tissulaire ou l'imagerie des petits animaux, maximiser la sensibilité du résonateur et l'efficacité de l'acquisition s'avère être une voie plus pratique et plus rentable pour améliorer le rapport Signal sur Bruit que d'augmenter ($\mathbf{B}_0$).
 - Sensibilité du résonateur : étant l'un des moyens les plus efficaces pour optimiser le rapport Signal sur Bruit, la sensibilité du résonateur joue un rôle crucial. Un résonateur efficace améliore l'induction du signal grâce à (B_1/I) , tout en minimisant les pertes perçues grâce à la réduction des contributions du bruit (σ_{noise}), ce qui améliore efficacement le rapport Signal sur Bruit. Une conception minutieuse du résonateur grâce à une géométrie sur mesure, une distribution optimisée du courant et une optimisation des performances peuvent considérablement augmenter (S_{eq}) en améliorant localement la sensibilité. Dans les systèmes à bas champ ou miniatures, l'optimisation des résonateurs permet souvent d'obtenir les gains les plus importants par rapport à d'autres stratégies.
 - Paramètres d'acquisition : le rapport Signal sur Bruit est également modulé par des paramètres d'acquisition tels que la bande passante, le temps de répétition (TR), le temps d'écho (TE) et l'angle de basculement. Par exemple, la réduction de la bande passante du récepteur améliore le rapport Signal sur Bruit en réduisant les contributions du bruit, mais risque d'introduire des artefacts dus au décalage chimique et à la susceptibilité. Les angles de basculement peuvent être optimisés (par exemple, en utilisant les conditions d'angle d'Ernst [98]) afin de maximiser le rapport Signal sur Bruit compte tenu de contraintes spécifiques (T_1) et de TR. Chaque paramètre doit être soigneusement sélectionné afin d'équilibrer le rapport Signal sur Bruit par rapport au contraste, à la résolution et au temps d'acquisition.

Pour relever le défi central de la microscopie par RM en produisant un signal suffisant tout en conservant la résolution microscopique nécessaire à l'imagerie quantitative des tissus artificiels, des améliorations stratégiques sont nécessaires. Si de nombreux paramètres peuvent être ajustés pour améliorer le rapport Signal sur Bruit, toutes les stratégies ne sont pas aussi pratiques ou efficaces les unes que les autres. Le simple fait d'augmenter le temps d'acquisition est universellement applicable, mais limité par des contraintes de temps dans de nombreuses configurations expérimentales. L'augmentation de la taille des voxels permet d'améliorer directement le rapport Signal sur Bruit, mais nuit considérablement à la résolution spatiale, ce qui est inacceptable pour les applications nécessitant un niveau de détail élevé. L'augmentation de l'intensité du champ statique offre une solution puissante mais coûteuse, avec des obstacles pratiques et d'infrastructures importants qui limitent l'accessibilité et l'intégration globales dans un laboratoire biologique standard. En revanche, l'amélioration de la sensibilité du résonateur RF grâce à une conception et une optimisation minutieuse semble offrir le moyen le plus efficace, le plus évolutif et le plus rentable d'améliorer le rapport Signal sur Bruit, ce qui en fait la stratégie privilégiée.

2. Chapitre II : Etude analytique et simulations

2.1. Introduction

Le principal obstacle à l'utilisation généralisée des systèmes IRM de table à bas champ dans les laboratoires est le compromis inhérent entre leur accessibilité et leurs performances. Si ces systèmes offrent une solution pratique et rentable pour les laboratoires, leur faible intensité de champ magnétique entraîne intrinsèquement une diminution du rapport Signal sur Bruit (SNR), ce qui compromet la qualité de l'image et allonge les temps d'acquisition, en particulier pour les applications à haute résolution. Le résonateur RF a été identifiée comme le composant le plus critique et le plus accessible pour améliorer considérablement le rapport Signal sur Bruit, mais sa conception pour les systèmes de table à faible champ magnétique présente des défis uniques.

Contrairement aux systèmes cliniques à haut champ, où la source principale de pertes est en général l'échantillon biologique, ou aux systèmes à micro résonateurs, où les pertes conductrices et diélectriques sont minimales [110], les configurations de table pour petits échantillons fonctionnent souvent dans un régime électromagnétique de moyenne portée (caractérisé par des produits fréquence-dimension $f \cdot d \approx 2\text{-}30 \text{ MHz} \cdot m$). Dans ce régime intermédiaire, toutes les sources de pertes, y compris la résistance du conducteur du résonateur, la dissipation diélectrique capacitive, les pertes de conduction induites par l'échantillon et l'amortissement des courants de Foucault provenant du blindage voisin, contribuent de manière significative à la dégradation globale du signal.

Un aspect critique de la conception des résonateurs RF pour les systèmes IRM de table est l'intégration d'un blindage RF. Le blindage est essentiel pour protéger la chaîne de réception RM sensible contre les interférences électromagnétiques externes (IEM) et pour minimiser les interactions indésirables entre le résonateur RF et d'autres composants métalliques du scanner, tels que les résonateurs magnétiques ou à gradient [40], [99]. Dans les systèmes compacts de table, ce blindage est généralement « local », c'est-à-dire qu'il est placé à proximité du résonateur RF, faisant souvent partie du boîtier du résonateur lui-même. Si les blindages locaux sont efficaces pour contenir les champs RF, leur proximité avec le résonateur peut représenter un défi important. Le matériau conducteur du blindage peut entraîner une induction de courants de Foucault, qui à son tour réduit le facteur de qualité du résonateur (Q), modifie sa fréquence de résonance et son impédance, et peut perturber la distribution du champ $\mathbf{B}_1$. Par conséquent, le résonateur RF et son blindage local ne peuvent pas être conçus séparément. Ils constituent un système couplé électromagnétiquement qui nécessite une co-optimisation minutieuse afin d'équilibrer un blindage efficace avec une dégradation minimale des performances du résonateur. Par conséquent, pour obtenir des performances optimales du résonateur RF, visant principalement à maximiser le rapport Signal sur Bruit, il est nécessaire de mettre en place une stratégie d'optimisation complète qui tienne compte de manière méticuleuse de cette interaction complexe entre les mécanismes de pertes.

Considérant la complexité de cette conception, la simulation numérique offre le moyen le plus efficace et le plus répétable d'optimiser la géométrie des résonateurs. Les modèles analytiques et les formules mathématiques fournissent des estimations initiales précieuses, mais ils reposent sur des hypothèses, telles que des conducteurs infiniment longs ou fins, des symétries géométriques parfaites et l'absence d'interactions environnementales [104]. Ils modélisent avec difficulté et manque de précision les détails complexes des résonateurs réelles avec des géométries 3D complexes. Ces modèles ne capturent généralement que certains effets dominants tels que les modèles de courant et de champ, tout en négligeant d'autres interactions et phénomènes d'ordre supérieur tels que les effets de proximité entre des conducteurs rapprochés et le couplage électromagnétique avec des échantillons ou des blindages à proximité. Bien que ceux-ci puissent également être modélisés de manière analytique, ils nécessitent généralement des études au cas par cas rigoureuses et fastidieuses. De ce fait, leur pouvoir prédictif se limite souvent à des

approximations de premier ordre où les géométries sont simplifiées, et ils sont insuffisants pour l'optimisation fine de résonateurs complexes et réalistes.

L'évaluation quantitative d'échantillons biologiques de composition hétérogène et de formes irrégulières a été reconnue comme très difficile par Hoult et Deslauriers [101]. Comme exposé au chapitre I, la puissance du signal provenant de chaque petit élément de l'échantillon dépend du champ B_1 local et, si le champ varie dans l'espace, différentes régions subissent des angles de basculement différents, ce qui crée une incertitude dans l'interprétation du signal perçu. À moins que la distribution spatiale de B_1 et les propriétés intrinsèques de l'échantillon ne soient connues avec précision à un endroit donné, il est difficile d'obtenir une interprétation précise et physiologiquement pertinente. Ces incertitudes sont aggravées lorsque l'échantillon se déplace ou change de taille, ou lorsque le facteur Q du résonateur et sa calibration varient en fonction de la taille et de la composition de l'échantillon. Ceci est inévitable dans les applications d'ingénierie tissulaire où les échantillons passent de cellules souches à des structures 3D complexes. L'utilisation de résonateur présentant une excellente homogénéité du champ B_1 , une sensibilité élevée et un comportement résonant stable peuvent contribuer à atténuer considérablement ces difficultés. Par conséquent, pour l'imagerie quantitative d'échantillons de grandes tailles, l'homogénéité du champ B_1 sera essentielle pour obtenir des acquisitions quantitatives précises et pertinentes sur le plan physiologique.

En combinant calculs analytiques et simulations numériques, ce chapitre se concentre sur l'optimisation et la comparaison de différentes topologies de résonateurs volumétriques homogènes. Bien que les conceptions de résonateurs de surface hautement sensibles puissent offrir une sensibilité locale bien supérieure [104], leur utilisation dans le contexte des applications d'ingénierie tissulaire sera limitée aux échantillons et aux constructions minces (~ 5 mm) nécessitant une très faible pénétration (peau, organoïdes 2D, etc.). Comme le B_1 des résonateurs de surface diminue rapidement avec la profondeur (environ la moitié de leur diamètre ou plus), ils sont intrinsèquement mal adaptés à la caractérisation de grands échantillons à volume complet (~ 40 mm). De même, les résonateurs en réseau ne seront pas pris en compte dans cette étude, car le scanner de paillasse ciblé est un système à canal unique et la mise en œuvre d'un réseau nécessiterait des modifications matérielles importantes, telles que l'ajout d'un récepteur multicanal ou d'un banc de commutateurs multiplexeurs.

Chaque résonateur sera optimisé afin de maximiser le rapport Signal sur Bruit et l'homogénéité du champ B_1 , soit sur une région d'intérêt cylindrique de 38×38 mm, soit sur une région d'intérêt sphérique de 19 mm de rayon (en fonction de la topologie du résonateur). Ils seront spécialement conçus pour un scanner IRM à bas champ de table de 0,3 T de Pure Devices GmbH (gamme Research Lab), fonctionnant à une fréquence de Larmor ^1H de 13,03 MHz avec une région d'intérêt utilisable nominale de $40 \times 40 \times 60$ mm. La conception est contrainte par les dimensions physiques du scanner, qui limitent la taille maximale du boîtier blindé du résonateur à $180 \times 55 \times 220$ mm. Dans ce contexte, l'objectif principal de l'optimisation du résonateur RF sera de maximiser les performances du résonateur tout en maintenant une homogénéité élevée du champ B_1 dans la région d'intérêt.

Comme exposé dans le chapitre précédent, du point de vue du résonateur, le rapport Signal sur Bruit est amélioré principalement en maximisant le facteur de qualité (Q) et le facteur de remplissage (η), qui quantifient ensemble l'efficacité du couplage magnétique entre le résonateur et l'échantillon, tout en maintenant le bruit provenant des interférences externes à un minimum. Cependant, ces mesures n'auront un impact significatif sur le rapport Signal sur Bruit que dans un domaine où le bruit du résonateur est dominant et où la plupart des pertes dépendent de ce même résonateur. L'homogénéité du champ est toutefois régie par la géométrie du résonateur et la topologie de l'enroulement, qui doivent être sélectionnées et ajustées avec précision afin de fournir des champs RF uniformes. L'homogénéité du champ généralement requise dans la conception

d'un résonateur RF est définie comme ayant soit des variations de champ inférieures à $\pm 5\%$ [104], soit un coefficient de variation (CoV) inférieur à (5%) [111].

L'objectif de ce chapitre est d'identifier la topologie optimale de résonateur RF blindé localement pour les applications d'imagerie de petits échantillons sur le scanner IRM de table 0,3 T mentionné précédemment ci-dessus.

2.2. Résultats

Une analyse préliminaire a permis d'identifier les topologies de résonateurs RF les plus solides sur le plan théorique sur la base de modèles idéalisés en espace libre, c'est-à-dire que les interactions avec l'environnement ne sont pas considérées. Bien qu'elles soient essentielles pour comprendre les principes fondamentaux de conception, ces approximations de premier ordre ne tiennent pas compte des interactions électromagnétiques complexes présentes dans un environnement IRM réel. Pour obtenir une prévision plus précise des performances, il est nécessaire de passer des équations analytiques à des simulations électromagnétiques (EM) 3D complètes qui intègrent la géométrie des conducteurs, le blindage RF et le chargement des échantillons diélectriques.

Sachant qu'elles présentent une bonne homogénéité et les principes de conception identifiés précédemment, trois topologies ont été sélectionnées pour une phase de modélisation avancée : le résonateur à boucle pontée (BLGR) et le solénoïde à pas variable (VPS) comme conceptions cylindriques optimales, et le résonateur N-Ring Extended Hoult & Deslauriers (H&D) comme conception sphérique optimale. Cette section décrit la méthodologie utilisée pour construire et évaluer des modèles réalistes de ces candidats avant de les comparer entre eux.

Le VPS améliore le solénoïde classique en modulant l'espacement entre ses enroulements, les rendant plus serrés aux extrémités et plus lâches au centre afin d'augmenter la densité de courant aux bords du résonateur pour contrer la baisse naturelle de l'intensité du champ, améliorant ainsi considérablement l'homogénéité dans un format compact.

Cependant, pour maintenir une longueur de résonateur fixe tout en relâchant les enroulements centraux, l'angle de pas au centre doit augmenter. À mesure que le pas central s'élargit, la composante axiale du courant au centre du résonateur ne peut plus être considérée comme négligeable par rapport à la composante azimutale, ce qui entraîne une baisse de l'homogénéité près des enroulements centraux et réduit la largeur de la région homogène disponible.

Pour remédier à cela, Privalov et al. [158] ont proposé une conception qui utilise un ruban mince de largeur variable en continu. Cette méthode permet d'augmenter la densité des spires aux extrémités du résonateur, ce qui corrige l'inhomogénéité dans la direction axiale, comme dans les conceptions à pas variables. Cependant, en faisant varier la largeur du ruban plutôt que l'écart entre les spires, l'écart peut être maintenu constant et étroit. Cela minimise les variations de champ à proximité du conducteur, ce qui améliore également l'hétérogénéité dans la direction radiale.

2.3. Conclusion

Cette étude a permis une évaluation de trois topologies avancées de résonateurs RF, à savoir le solénoïde à ruban à largeur variable (VWRS), le résonateur à boucle pontée (BLGR) et le résonateur sphérique étendu Hoult & Deslauriers (H&D), afin d'identifier la conception optimale pour l'imagerie quantitative de petits échantillons sur un scanner IRM de table de 0,3 T. Les résultats révèlent que le choix du résonateur optimal implique un compromis nuancé entre la maximisation de la sensibilité brute et l'obtention d'un profil de performance équilibré et robuste.

Les résonateurs sphériques H&D, malgré leur élégance théorique, se sont révélées moins efficaces dans un environnement simulé réaliste. Les contraintes pratiques liées à la taille finie des conducteurs et à l'intégration des composants ont considérablement dégradé leur homogénéité, tandis que leurs facteurs Q chargés sont restés modestes. Cela souligne une conclusion essentielle :

pour les résonateurs blindés localement dans ce régime opérationnel, les conceptions intrinsèquement robustes et possédant des facteurs Q intrinsèques élevés, comme le BLGR, sont plus efficaces que celles basées uniquement sur des approximations idéales du schéma de courant.

L'analyse identifie deux candidats exceptionnels parmi la famille BLGR, le BLGR_20C et le BLGR_13C, chacun présentant son propre ensemble d'avantages.

Le BLGR_20C est le leader incontesté en matière de sensibilité brute. Grâce à son facteur Q chargé remarquablement élevé de 374, sa sensibilité relative est supérieure de plus de 30 % à celle de toute autre conception, ce qui constitue un avantage décisif dans l'environnement à faible signal de l'IRM de paillasse. Bien que son homogénéité (1,63 % CoV) soit légèrement inférieure à celle des modèles les plus performants, elle reste excellente et convient parfaitement aux applications quantitatives exigeantes. Son principal inconvénient est qu'il fonctionne dans un régime dominé par le bruit de du résonateur ($Q_u/Q_\ell = 1,21$), ce qui, bien qu'inévitable pour certains modèles à Q élevé, signifie qu'une partie importante du bruit provient du résonateur lui-même.

Le BLGR_13C apparaît comme une alternative très viable et robuste qui offre un profil de performances plus équilibré. Il atteint la plus grande homogénéité de champ B1 de toutes les conceptions testées, avec un CoV de 1,58 %, ce qui le rend idéal pour les applications nécessitant une précision quantitative maximale. Il fonctionne dans le régime souhaitable dominé par le bruit de l'échantillon ($Q_u/Q_\ell = 2$), ce qui garantit que la qualité de la mesure est limitée par l'échantillon plutôt que par les imperfections du matériel. Bien que sa sensibilité soit inférieure à celle du BLGR_20C, elle reste la deuxième plus élevée et supérieure à toutes les conceptions non BLGR. De plus, sa complexité réduite (13 condensateurs contre 20) simplifie la construction, le réglage et la reproduction. Pour les applications nécessitant la fabrication d'un grand nombre de résonateurs, la variante à 13 condensateurs est l'option la plus économique, les MLCC non magnétiques coûtant entre 1,5 et 2 euros par unité (jusqu'à 15 à 20 euros dans des cas particuliers), contre quelques centimes par unité pour les condensateurs standards.

Enfin, le choix de la conception optimale dépend de l'application. Pour les expériences où la maximisation du rapport Signal sur Bruit est le facteur le plus critique, le résonateur BLGR_20C est la meilleure option. Cependant, pour les applications exigeant la plus grande fidélité quantitative combinée à des performances robustes en matière de bruit et une mise en œuvre plus simple, le BLGR_13C représente une alternative excellente et convaincante.

4. Chapitre III : Fabrication d'un dispositif à l'aide de la méthode « Rapid-3D Plastronics »

4.1. Introduction

Pour tirer pleinement parti de l'IRM en tant que modalité efficace pour la caractérisation 3D d'échantillons vivants, il est nécessaire de concevoir une configuration combinant une chambre perfusée 3D avec un résonateur blindé ajusté dans un seul et même dispositif. De plus, la configuration doit s'adapter à l'espace limité disponible propre aux systèmes d'IRM de paillasse. Les procédés de fabrication actuels reposant sur l'électronique conventionnelle sont insuffisants et ne permettent pas d'intégrer de manière transparente les composants mécaniques et électriques. En effet, l'utilisation des technologies de fabrication électronique conventionnelles, principalement basées sur des cartes de circuits imprimés 2D rigides, n'est pas adaptée à la conception des géométries 3D complexes nécessaires pour optimiser la sensibilité et garantir des champs magnétiques radiofréquences homogènes. Actuellement, les configurations IRM reposent en grande partie sur l'utilisation de résonateurs RF disponibles dans le commerce, ce qui conduit à des assemblages sous-optimaux. La fixation des échantillons et l'intégration des périphériques nécessaires dans ces résonateurs RF impliquent souvent des méthodes peu fiables, telles que des rubans adhésifs et des supports de fortune, qui manquent de reproductibilité et de précision.

L'approche plastronique 3D, une technique qui permet l'intégration complète de circuits et de composants électroniques sur des surfaces 3D, offre une solution prometteuse. Les méthodes de fabrication plastronique 3D reposent sur l'intégration de circuits sur des surfaces polymères 3D, avec des pistes conductrices servant d'interconnexions ou d'éléments fonctionnels tels que des antennes [114] ou des inductances [159]. Elles permettent également de mettre en place des fonctionnalités actives telles que des capteurs de pH et des systèmes microfluidiques [160], [161], avec des choix de matériaux adaptés à l'application (d'origine biologique/biodégradables [162], résistants à la température [129], biocompatibles, etc.).

Les méthodes plastronique 3D ont déjà démontré leur capacité à intégrer de manière transparente des résonateurs RF dans des configurations IRM complexes [120], [130], [163]. L'utilisation de méthodes similaires, intégrant des résonateurs bien dimensionnés directement à la surface de la chambre de culture, permettrait de créer un système cohérent combinant les fonctionnalités d'imagerie et de culture tissulaire. Cette approche garantit un positionnement optimal du résonateur par rapport à l'échantillon, maximisant la sensibilité et l'homogénéité du champ tout en conservant la capacité de la chambre de culture à fournir un environnement contrôlé, similaire à celui *in vivo*. De plus, la flexibilité de la fabrication plastronique 3D permet de concevoir des géométries de chambres de culture et des configurations de résonateurs personnalisées, s'adaptant à une large gamme de tailles d'échantillons.

Ce chapitre vise à présenter un dispositif évolutif et rentable pour caractériser les constructions tissulaires 3D issues de l'ingénierie tissulaire grâce à une approche qui améliore les capacités d'imagerie, mais offre également de nouvelles possibilités d'explorer les mécanismes, la dynamique et les interactions qui sous-tendent le développement fonctionnel des tissus dans les systèmes bio-conçus, en combinant l'adaptabilité de la fabrication plastronique 3D avec un scanner IRM à bas champ rentable, accessible et portable pour une intégration en ligne dans le processus de fabrication des tissus.

Afin de fournir une comparaison basée sur les performances avec une autre approche plus « classique », un résonateur volumique optimisé fourni par le fabricant sera utilisée comme référence, facilitant une évaluation quantitative des performances (rapport Signal sur Bruit (SNR), homogénéité du champ, praticité de l'assemblage, etc.). Plus précisément, le résonateur volumique optimisé fourni par le fabricant servira de norme de référence par rapport à laquelle la configuration du résonateur plastronique 3D fabriqué sur mesure sera évaluée selon des protocoles

identiques de balayage. Les deux configurations utiliseront la même topologie de résonateur afin de garantir une comparaison équitable et axée sur les performances. Comme indiqué au chapitre II, la topologie BLGR a démontré la plus grande stabilité dans des conditions variables de charge tout en conservant une homogénéité et une sensibilité élevées. Elle sera donc mise en œuvre à la fois dans le résonateur de référence du fabricant et dans notre configuration personnalisée fabriquée par plastronique 3D.

4.2. Résultats et Conclusion

Il a été démontré que l'intégration d'un résonateur RF, conçu sur mesure et fabriqué à l'aide des techniques plastroniques 3D directement sur la chambre de culture, présente certains avantages par rapport à un système de résonateur blindé classique. Dans notre évaluation comparative de deux configurations expérimentales, l'approche Plastronique 3D s'est avérée avantageuse dans plusieurs domaines critiques.

Tout d'abord, la capacité de la technologie Plastronique 3D à intégrer le résonateur RF plus près de la zone d'intérêt a permis d'obtenir un gain global de 60 % du rapport Signal sur Bruit mesuré et une légère amélioration de l'homogénéité du champ B_1^+ . Les cartes de champ B_1^+ dérivées de la méthode à double angles ont montré à la fois une augmentation de l'amplitude moyenne de B_1^+ (de $1,78 \times 10^{-4}$ T à $2,79 \times 10^{-4}$ T) et une réduction du coefficient de variation (de 2,15 % à 1,53 %) dans la configuration Plastronique 3D par rapport à l'alternative commerciale.

Le gain SNR mesuré est conforme aux prévisions théoriques dérivées de notre modèle SNR (55%), où l'amélioration des facteurs de qualité, la couverture volumique effective et la réduction du facteur de bruit contribuent à l'amélioration de la sensibilité observée. La principale source d'amélioration du rapport Signal sur Bruit est le facteur de remplissage amélioré, obtenu en réduisant le rayon intérieur du résonateur de 21 mm à 17 mm, ce qui rapproche le capteur de l'échantillon. Dans la conception de cette chambre, il a été décidé de faire passer le chemin de retour du système fluidique par le résonateur intégré et d'utiliser la configuration à 13 condensateurs présentée au chapitre II. L'intégration d'un chemin de retour circulant à l'extérieur du résonateur intégrée dans les conceptions futures devrait permettre une plus grande proximité entre le résonateur et l'échantillon (rayon intérieur d'environ 12 mm et longueur de 58 mm) et pourrait offrir une sensibilité encore meilleure du résonateur, pouvant atteindre environ 120 % avec la conception à 13 condensateurs (sur la base du facteur Q estimé à partir de simulations numériques) ou même plus avec une conception à 20 condensateurs. L'homogénéité améliorée est le résultat d'un ajustement et d'un positionnement plus serrés et plus conformes, tout en minimisant les effets de bord, et en s'assurant que la longueur du résonateur est plus grande que son diamètre [104] avec un facteur de forme $FF \approx 0,53$ par rapport à $FF \approx 0,65$ pour le résonateur fourni par le fabricant.

De plus, la configuration Plastronique 3D a présenté un facteur Q chargé légèrement plus élevé (370 contre 322) et un facteur de bruit nettement plus faible (1,48 contre 2,01), ce qui indique un stockage d'énergie plus efficace et un blindage supérieur contre les interférences RF externes. Ces deux aspects peuvent être attribués à un blindage plus efficace. La différence de facteur Q provient de la plus grande surface blindée par rapport aux dimensions du résonateur dans la configuration Plastronique 3D ($178 \times 52 \times 218$ mm pour une bobine de 34 mm de diamètre) par opposition à la configuration commerciale ($90 \times 51 \times 170$ mm pour une bobine de 42 mm de diamètre), ce qui réduit considérablement les effets de proximité du blindage. Le boîtier imprimé en 3D dans la configuration Plastronique 3D a permis une meilleure couverture du résonateur. La chambre était presque entièrement fermée, ne laissant qu'un petit espace au fond pour accueillir les tubes fluidiques.

Enfin, ces résultats confirment que l'approche Plastronique 3D représente une alternative viable et avantageuse aux configurations conventionnelles de résonateur IRM de paillasse, en

particulier pour les applications nécessitant une intégration étroite avec des chambres de culture fluidiques. La combinaison d'un facteur de remplissage amélioré, d'une efficacité de blindage accrue et d'une adaptation optimisée entre le résonateur et l'échantillon s'est traduite par des gains mesurables en termes de sensibilité, d'homogénéité du champ et de suppression du bruit. Comme indiqué précédemment, cette amélioration est particulièrement pertinente dans les contextes précliniques et d'ingénierie tissulaire, où un rapport Signal sur Bruit élevé et une excitation uniforme sont essentiels pour résoudre les détails structurels fins et obtenir des mesures IRM quantitatives. À l'avenir, l'optimisation de la géométrie du résonateur et de la conception de la chambre offre la possibilité d'obtenir des gains de sensibilité encore plus importants, ouvrant la voie à des solutions matérielles IRM spécifiques de nouvelle génération pour certaines applications, rendues possibles par Plastronique 3D.

Par rapport à d'autres initiatives dans le domaine de l'IRM bas champ pour l'ingénierie tissulaire, notre configuration affiche des performances compétitives. Par exemple, si Nitzsche et al. [21] ont utilisé avec succès un système de paillasse de 0,5 T pour caractériser des scaffolds, leur approche reposait sur des résonateurs standards et 128 moyennes de balayage unique, ce qui a nécessité un temps de mesure total d'environ 90 minutes. Un gain de rapport Signal sur Bruit de 60% à l'aide d'une configuration Plastronique 3D similaire à celle proposée dans ce travail permettrait de remédier aux longs temps d'acquisition : une image nécessitant 90 minutes avec le résonateur d'origine pourrait être acquise en seulement 35 minutes environ avec une qualité équivalente. Dans l'étude originale, les scaffolds étaient transférés des plaques de culture vers des tubes RMN pour l'imagerie, ce qui présentait un risque de perturbation mécanique, de contamination ou de désalignement. Leur configuration nécessitait d'interrompre la culture pour réaliser l'imagerie, sans compatibilité avec la perfusion pendant le processus d'imagerie, ce qui rendait impossible l'évaluation du transport des nutriments et de la prolifération cellulaire dans des conditions de croissance. Une chambre d'imagerie de culture intégrée Plastronique 3D pourrait permettre une observation non invasive et *in situ* sans interrompre les conditions stériles, tout en améliorant la reproductibilité, en maintenant un alignement spatial précis et en prenant en charge la culture basée sur la perfusion, permettant ainsi l'imagerie dans des conditions physiologiquement pertinentes. Nos travaux s'appuient sur les résonateurs Plastronique 3D précédents de Gerges et al. [120] et Gnanago et al. [130] en intégrant pleinement le résonateur à une chambre de perfusion, créant ainsi un dispositif unique et optimisé spécialement conçu pour l'observation *in situ* des tissus.

Ce travail démontre la faisabilité et les avantages d'une configuration IRM entièrement intégrée basée sur la technologie Plastronique 3D, spécialement conçue pour l'imagerie de structures bio-imprimées perfusées. En intégrant des résonateurs RF conformes directement sur la chambre de culture et en incorporant un blindage local dans une conception mécanique modulaire, le système a permis d'améliorer l'homogénéité du champ B_1^+ , d'obtenir un rapport Signal sur Bruit plus élevé et d'obtenir des performances d'imagerie reproductibles dans des conditions de bas champ. La capacité de la configuration à capturer des structures internes détaillées sous perfusion offre une valeur significative pour l'évaluation non destructive en ingénierie tissulaire.

Au-delà de l'amélioration des performances d'imagerie, **une configuration IRM basée sur la technologie** Plastronique 3D offre des avantages pratiques en termes de flexibilité de conception et de prototypage rapide. Le processus de fabrication permet l'intégration directe de résonateurs RF et **de points d'accès fluidiques personnalisés**, tels que des entrées et sorties de perfusion, dans le corps de la chambre de culture. Cela permettrait d'adapter chaque système à des géométries de structure, des exigences de débit et des conditions expérimentales spécifiques, améliorant ainsi la reproductibilité et l'adaptabilité. En intégrant le résonateur directement dans la structure imprimée, la conception évite les limites courantes des résonateurs traditionnels, telles qu'un alignement sous-optimal, une distance accrue par rapport à l'échantillon et un couplage électromagnétique dégradé.

La polyvalence du procédé Plastronique 3D va au-delà des résonateurs RF. Comme l'ont démontré d'autres études utilisant des techniques de fabrication similaires, il est possible d'intégrer divers capteurs (température, pH [160], [161], etc.) ou même des actionneurs piézoélectriques [177] pour l'élastographie par résonance magnétique (MRE). Cela transformerait le dispositif en une plateforme d'observation multimodale, fournissant des données physiologiques en temps réel parallèlement aux scans IRM anatomiques, offrant ainsi une vision plus holistique du développement des tissus.

De plus, cette approche de fabrication est intrinsèquement évolutive. Le flux de travail de conception numérique (CAO, CFD et analyse EM jusqu'à l'impression 3D et la fonctionnalisation) permet une itération et une personnalisation rapides de la chambre de culture et de la géométrie du résonateur afin de s'adapter à différentes tailles de structures, différents types de cellules, ou même à l'intégration avec différents systèmes IRM de paillasse. Contrairement à la fabrication conventionnelle qui nécessite un réoutillage coûteux et fastidieux, les nouvelles conceptions peuvent être imprimées et métallisées en quelques jours, ce qui rend cette technologie très adaptable à divers besoins de recherche et ouvre la voie à des chambres de culture standardisées, spécifiques à certaines applications et compatibles avec l'IRM.

Cependant, certaines limites subsistent. Le contraste entre le support et le milieu environnant dépend du contenu biochimique de ces milieux, qui peut évoluer avec la maturation des échantillons et varier d'un échantillon à l'autre. Dans les situations de très faible contraste, des gains supplémentaires de rapport Signal sur Bruit peuvent être nécessaires pour optimiser les stratégies d'acquisition IRM afin d'obtenir une différenciation maximale.

En outre, si le blindage réduit efficacement le bruit, il limite également l'accès physique à la chambre et nécessite des étapes d'assemblage manuel qui peuvent ralentir la répétition des expériences. Cependant, de nouvelles améliorations de la géométrie et de la conception du blindage pourraient aider à surmonter cette difficulté. Le besoin actuel dans le montage proposé pour visser et dévisser l'enceinte peut être résolu dans les prochaines itérations en adoptant un système de verrouillage mécanique. Pour remédier davantage à la nature encombrante du blindage passif actuel, d'autres stratégies peuvent également être explorées (par exemple, un blindage actif utilisant des résonateurs secondaires pour annuler les interférences externes ou des topologies de résonateurs auto-blindés avec des motifs conçus pour confiner intrinsèquement le champ magnétique [73], [74], [75], [76]). Cette nouvelle approche améliorerait considérablement l'accès des utilisateurs et la quantité d'images réalisables.

Dans l'ensemble, ces résultats soulignent le potentiel de la Plastronique 3D en tant que solution évolutive et personnalisable pour intégrer des capacités d'imagerie dans des systèmes bio-conçus, ce qui en fait une solution très intéressante pour remédier aux limites de sensibilité d'une configuration de table à bas champ.

5. Chapitre IV : Preuve de concept d'une antenne cryogénique

5.1. Introduction

Une autre stratégie convaincante pour améliorer le rapport Signal sur Bruit consiste à refroidir cryogéniquement le résonateur RF. Cette technique vise à réduire considérablement la résistance électrique des conducteurs du résonateur et à réduire le bruit du système, ce qui se traduit par d'importants gains en sensibilité. Cette approche a été étudiée et mise en œuvre avec succès dans diverses applications de RMN et d'IRM à petite échelle [100], [178], [179], [180].

Bien que les matériaux supraconducteurs aient été considérés ces dernières années comme la solution ultime pour les applications de résonateurs refroidis cryogéniquement [180], leur utilisation dans les IRM de table peut s'avérer peu pratique. Les supraconducteurs à haute température (HTS) souffrent de contraintes sévères en termes de géométrie et de flexibilité de réglage, d'un coût excessif et d'une fiabilité médiocre sous cycle thermique [178]. De plus, leurs facteurs de qualité très élevés, bien que théoriquement attrayants, conduisent à des bandes passantes étroites de résonance qui sont souvent plus petites que la bande passante d'imagerie. Cela peut entraîner plusieurs problèmes critiques. En mode émission, les temps de décroissance longs peuvent déformer les profils d'impulsions RF et compliquer la synchronisation des séquences, tandis qu'en mode réception, la bande passante limitée peut produire une sensibilité dépendante de la fréquence et des artefacts de bandes d'image [106]. Ces effets nécessitent des stratégies supplémentaires d'amortissement ou de surcouplage, qui peuvent annuler une grande partie de l'avantage théorique du rapport Signal sur Bruit.

De plus, les résonateurs en cuivre sont robustes, peu coûteux et faciles à fabriquer dans une variété de géométries, y compris des résonateurs volumétriques et superficiels qui peuvent être adaptés à des tailles d'échantillons spécifiques. Bien que la résistance électrique du cuivre génère plus de bruit qu'un supraconducteur, en particulier à température ambiante, ce bruit est modélisable et peut être atténué en refroidissant le résonateur. Cependant, même avec un refroidissement, la réduction du bruit est limitée par les propriétés intrinsèques du matériau, atteignant un plateau indépendant de la température définie par le rapport de résistance résiduelle (RRR) et l'effet de peau anormal qui limite la diminution de la résistance électrique des conducteurs normaux à basse température et restreint les performances ultimes des résonateurs en cuivre cryogéniques [180]. Néanmoins, pour la plupart des applications, les résonateurs en cuivre offrent un équilibre fiable entre performances, coût et facilité d'utilisation, ce qui en fait une solution attrayante. De plus, dans le cadre de ce travail, l'objectif est d'améliorer les performances d'imagerie à moindre coût et de fournir une première preuve de concept. Pour cela, les résonateurs refroidis au cuivre constitueront un bon démonstrateur initial.

Afin de rester fidèle à l'idée de fournir une configuration peu coûteuse et personnalisable pour les applications d'ingénierie tissulaire, une conception de cryostat évitant les matériaux coûteux et mettant l'accent sur un processus de fabrication rationalisé et accessible a été proposée. Cette stratégie privilégie la facilité de fabrication à l'aide de matériaux disponibles dans le commerce, ce qui permet à tout laboratoire de biologie de construire un système sur mesure sans équipement ni expertise spécialisée. Il a été décidé que le cœur de cette conception s'appuierait sur l'impression 3D pour créer un boîtier sur mesure, une simple bouteille en polyéthylène haute densité (PEHD) pour former le contenant central et des matériaux aérogels courants, évitant ainsi le recours à des systèmes complexes de vide. Pour le refroidissement, l'azote liquide (LN_2) a été choisi comme cryogène largement accessible et rentable. L'espoir est que cette combinaison fournisse une méthode robuste et fiable pour refroidir le résonateur de cuivre à des températures cryogéniques, permettant ainsi d'améliorer considérablement le rapport Signal sur Bruit (SNR). Cette approche peu coûteuse et « à faire soi-même » devrait contribuer à démontrer la validité du concept, en prouvant que des résonateurs RF haute performance pour l'IRM de paillasse peuvent être réalisées

avec un investissement minimal, supprimant ainsi un obstacle important à l'adoption généralisée de cette technologie dans l'ingénierie tissulaire et d'autres domaines de la recherche biologique.

5.2. Résultats et Conclusion

Ce travail a démontré la validité du concept d'un système de résonateurs RF à refroidissement cryogénique et à faible coût pour l'IRM de paillasse, le confirmant comme une stratégie viable pour une amélioration significative du rapport Signal sur Bruit. L'objectif principal est de développer un système accessible utilisant des matériaux et des méthodes de fabrication facilement disponibles, tels que l'impression 3D. Les résultats expérimentaux montrent un gain de rapport Signal sur Bruit mesuré de 57% après refroidissement du résonateur à 90 K. Cela confirme l'hypothèse fondamentale selon laquelle un simple système de refroidissement fabriqué en laboratoire et une combinaison de résonateurs cryogéniques peuvent améliorer les performances du rapport Signal sur Bruit.

Cependant, alors que le gain simulé du rapport Signal sur Bruit prévu à 90 K est d'environ 83%, le gain mesuré est inférieur. Cet écart peut être attribué à plusieurs facteurs pratiques. Le plus important est probablement la différence entre les facteurs de qualité simulés et mesurés : le résonateur fabriqué présente un facteur de qualité Q inférieur aux prédictions à température ambiante (240 au lieu de 342) et à celles à température cryogénique (400 au lieu de 484). Cela suggère que certaines pertes ne sont pas prises en compte dans le modèle proposé, potentiellement dues à la feuille de cuivre adhésive, aux propriétés diélectriques du Spaceloft Aérolog ou aux modèles de composants discrets. Une caractérisation approfondie et sans charge du résonateur pendant un cycle de refroidissement, ainsi qu'une caractérisation plus approfondie des matériaux dans des conditions cryogéniques, seraient nécessaires pour étudier plus en détails ces mécanismes de pertes.

L'une des plus grandes différences a été observée dans le décalage de la fréquence de résonance pendant le refroidissement. Les simulations, qui se sont concentrées sur la modification de l'effet de peau du cuivre, prédisent une légère augmentation de la fréquence (~ 6 kHz). En revanche, l'expérience a révélé une diminution significative de 53 kHz. Ce résultat indique clairement que l'effet dominant n'était pas l'inductance du résonateur, mais plutôt les caractéristiques dépendantes de la température des condensateurs d'accord MLCC. Un calcul théorique basé sur le coefficient de température de capacité (TCC) du fabricant, évalué pour une température minimale de 218 K, prédit un décalage vers le bas de 41 kHz, ce qui tend à correspondre à la valeur mesurée. Cela signifie que pour ce type de conception de résonateur, la stabilité thermique des composants discrets est un facteur plus important que les propriétés changeantes du conducteur lui-même. Les simulations futures devront intégrer des modèles précis dépendants de la température pour tous les composants discrets afin d'obtenir de meilleurs résultats prédictifs.

Le projet a également mis en évidence d'importants défis techniques qui offrent des pistes d'amélioration évidentes. Le plus critique est l'isolation thermique de la chambre d'échantillonnage. Le refroidissement progressif de l'échantillon, conduisant à la formation de glace, indique que si le tube en borosilicate à double paroi limite efficacement le transfert de chaleur par conduction, il est insuffisant pour empêcher la perte de chaleur par rayonnement thermique. Les effets du rayonnement ont été fortement sous-estimés lorsque le choix a été fait de remplacer l'isolation interne opaque en aérolog de la première proposition de conception préliminaire par un tube transparent. Il s'agit d'un défaut de conception majeur qui doit être corrigé afin de maintenir la viabilité des échantillons et la stabilité de la température, en particulier pour l'imagerie de cellules vivantes. En outre, la température finale du résonateur de 90 K, malgré la source d'azote liquide à 77 K, indique un gradient de température inévitable inhérent à la conception du refroidissement indirect, qui pourrait entraîner des hétérogénéités du champ B_1 .

Il est difficile de tirer une conclusion directe en comparant cette approche cryogénique à la méthode Plastronique 3D détaillée au chapitre III. Le rapport Signal sur Bruit initial à température ambiante du résonateur cryogénique (~ 37) est étonnement beaucoup plus faible que celui du résonateur de référence fourni par le fabricant (~ 68), une différence qui ne s'explique pas entièrement par les facteurs de qualité différents. Une étude plus contrôlée est nécessaire pour examiner cette différence de base. Néanmoins, les deux techniques ont démontré leur potentiel pour améliorer le rapport Signal sur Bruit, l'une en permettant une meilleure sensibilité grâce à un facteur de forme géométrique optimisé, l'autre en s'appuyant sur la réduction du bruit thermique induit par le résonateur dans un régime spécifique. Ces méthodes ne s'excluent toutefois pas mutuellement : trouver un moyen de fusionner les deux approches pourrait conduire à des gains significatifs dans le bon régime. Le déploiement d'une telle approche et l'étude devront commencer par le développement de matériaux isolants compatibles avec l'IRM, moins coûteux et extrêmement fins, permettant un ajustement plus précis entre l'échantillon et le résonateur cryogénique.

Dans le contexte plus large de l'IRM cryogénique, le gain de 57 % du rapport Signal sur Bruit observé est une grande réussite, conforme aux principes fondamentaux de réduction du bruit thermique dans un régime où le bruit du résonateur est dominant. D'autres groupes ont rapporté des gains similaires ou supérieurs. Par exemple, Resmer et al. ont démontré une amélioration du rapport Signal sur Bruit d'un facteur 2,8 (un gain de 180 %) pour un résonateur en cuivre refroidi à 77 K, bien qu'à un champ beaucoup plus faible de 0,01 T où le bruit du résonateur est encore plus prononcé et où l'utilisation de fils Litz est connue pour améliorer la distribution du courant et réduire les pertes dues aux effets de peau et de proximité, maximisant ainsi le facteur Q réalisable à des températures cryogéniques [178]. Les performances des résonateurs cryogéniques sont souvent comparées à celles des résonateurs à supraconducteurs à haute température (HTS), qui peuvent atteindre des facteurs Q extrêmement élevés et, en théorie, des gains de rapport Signal sur Bruit plus importants [180]. Cependant, comme indiqué dans l'introduction, les résonateurs HTS présentent des défis pratiques importants, notamment un coût élevé, une fragilité mécanique et des bandes passantes opérationnelles étroites qui peuvent compliquer les séquences d'imagerie [106]. Par conséquent, l'approche présentée ici, à savoir un résonateur en cuivre refroidi cryogéniquement, représente un bon compromis. Elle offre des performances supérieures à celles des résonateurs à température ambiante tout en évitant la complexité et le coût prohibitifs des systèmes HTS, ce qui correspond parfaitement à l'objectif de créer une solution accessible et performante pour une utilisation généralisée en laboratoire. Cependant, comme la conception proposée du cryostat atteint des températures de l'ordre de 90 K, elle peut également être utilisée avec un résonateur HTS. Une étude complémentaire peut donc être menée relativement facilement pour examiner les avantages du HTS dans ce contexte spécifique et déterminer si les avantages potentiels l'emportent sur le coût financier et d'autres limites.

La conception du cryostat est toujours en cours, mais sa démonstration initiale réussie ouvre plusieurs perspectives d'avenir passionnantes. La voie la plus prometteuse consiste à intégrer des techniques de fabrication avancées. Par exemple, Ceratronics, une extension du concept de Plastronique 3D utilisant des matériaux céramiques, pourrait être utilisé pour créer des substrats de résonateur avec des canaux microfluidiques cryogéniques intégrés. Une telle conception permettrait à l'azote liquide de s'écouler directement à travers le substrat, ce qui conduirait à un refroidissement plus uniforme et plus efficace sur toute la surface du résonateur. Cela atténuerait les gradients de température observés dans la conception actuelle et pourrait potentiellement rapprocher la température du résonateur de 77 K, ce qui permettrait d'obtenir des gains de SNR encore plus importants.

Ce projet a permis de concevoir, de fabriquer et de valider avec succès un résonateur RF cryogénique comme preuve de concept pour un scanner IRM de table. Une amélioration de 57 % du rapport Signal sur Bruit a été obtenue expérimentalement, confirmant que le refroidissement

cryogénique est un outil puissant pour surmonter les limites de sensibilité inhérentes aux systèmes à bas champ. Bien que des divergences entre la simulation et l'expérience aient été constatées, elles ont fourni des informations essentielles sur le comportement réel des composants à des températures cryogéniques et ont mis en évidence les domaines clés à améliorer dans la conception. Les limites, notamment en ce qui concerne l'isolation des échantillons et les gradients thermiques, ne sont pas insurmontables et constituent une feuille de route claire pour les itérations futures. La voie à suivre consiste à combiner les avantages de cette approche cryogénique avec des méthodes de fabrication avancées telles que Ceratronics afin de développer une plateforme d'imagerie robuste, performante et accessible pour l'ingénierie tissulaire et d'autres applications biomédicales.

Conclusion générale

L'objectif principal de cette thèse est de pallier le manque d'outils pratiques et non invasifs pour la caractérisation longitudinale de grandes structures biologiques tridimensionnelles dans les applications d'ingénierie tissulaire. Si l'imagerie par résonance magnétique (IRM) à bas champ offre une solution accessible et polyvalente, son utilité est limitée par un rapport Signal sur Bruit (SNR) intrinsèquement faible, qui réduit la résolution des images et allonge les temps d'acquisition. Ce travail a permis de relever ce défi en étudiant et en développant systématiquement deux stratégies distinctes, axées sur le matériel, pour améliorer le rapport Signal sur Bruit. La première reposait sur la conception de résonateurs radiofréquence (RF) optimisés fabriqués via Plastronique 3D, et la seconde sur la mise en œuvre d'une preuve de concept avec un système accessible et rentable de résonateurs cryogéniques.

L'étude a débuté par une analyse complète, basée sur la simulation, afin d'identifier la topologie de résonateur RF la plus efficace pour générer un champ magnétique homogène dans un environnement localement blindé et à bas champ. Grâce à une combinaison de modélisation analytique et de simulations électromagnétiques, le résonateur à boucle pontée (BLGR) a été identifié comme la conception la plus performante. Il a démontré un équilibre optimal entre une homogénéité élevée du champ B1, une stabilité opérationnelle sous des charges d'échantillons variables et une sensibilité intrinsèque, surpassant de manière significative les conceptions classiques de solénoïdes à pas variable et même les conceptions sophistiquées de résonateurs sphériques optimisées dont les avantages théoriques étaient diminués par les contraintes de mise en œuvre pratique.

La principale contribution de cette thèse a été la transformation et l'intégration réussies de cette conception BLGR optimale dans une chambre de culture multifonctionnelle entièrement intégrée utilisant un processus de fabrication « Rapid 3D Plastronics (R3DP) ». Cette approche a permis de créer une chambre de culture par perfusion avec un résonateur RF ajusté et son circuit d'adaptation intégré directement à sa surface. Cette intégration transparente de la fonctionnalité du résonateur d'imagerie et de la chambre de culture par perfusion a permis d'obtenir un gain SNR significatif et validé expérimentalement de 60 % et une homogénéité supérieure du champ B1 par rapport à un résonateur conventionnel du même type fourni par le fabricant. L'efficacité du système a été démontrée par l'évaluation morphologique haute résolution d'un scaffold hydrogel bio-imprimé en 3D *in situ*, validant ainsi la technologie Plastronique 3D non seulement comme méthode de fabrication, mais aussi comme outil puissant pour créer des configurations IRM intégrées haute performance spécifiques à une application, comblant ainsi le fossé entre l'imagerie et l'expérimentation biologique.

Parallèlement, une deuxième stratégie d'amélioration du rapport Signal sur Bruit a été explorée grâce à la conception et à la mise en œuvre d'un cryostat et d'une configuration de résonateur cryogénique à faible coût. Le cryostat proposé a réussi à refroidir un BLGR à base de cuivre à environ 90 K, ce qui a permis d'obtenir un gain de rapport Signal sur Bruit mesuré de 57 %. Ce résultat a confirmé la viabilité de la cryogénie comme méthode puissante pour réduire le bruit thermique dans un régime dominé par le bruit du résonateur. Cependant, l'étude a également mis

en évidence d'importants défis techniques, notamment la nécessité d'améliorer l'isolation thermique des échantillons et la gestion du comportement des composants en fonction de la température, positionnant cette approche comme une alternative prometteuse mais moins mature, nécessitant d'être affinée pour une utilisation courante en laboratoire.

En comparant ces deux approches, celle de la Plastronique 3D offre la solution la plus immédiate et la plus polyvalente. Elle permet non seulement d'augmenter la sensibilité, mais surtout de redéfinir le déroulement des expériences en créant un système cohérent et intégré où l'environnement de « survie » de l'échantillon et le capteur d'imagerie ne font qu'un. Ces travaux ont démontré que les limites de performance de l'IRM à bas champ ne sont pas insurmontables, mais peuvent être efficacement surmontées grâce à une conception ciblée des résonateurs et à des innovations matérielles.

À l'avenir, les différentes approches étudiées ici ouvrent de nombreuses perspectives pour la recherche future. La chambre de culture fabriquée en Plastronique 3D peut être encore améliorée en intégrant des capteurs supplémentaires pour l'observation en temps réel du pH, de l'oxygène ou de la température, créant ainsi une chambre de culture « intelligente » véritablement complète. Le système cryogénique peut être perfectionné en tirant parti de matériaux et de techniques de fabrication avancés, tels que Ceratronics avec des canaux de refroidissement microfluidiques intégrés, afin d'atteindre des températures plus basses et une gestion thermique plus uniforme. Enfin, les outils et les méthodologies établis dans cette thèse jettent les bases pour faire de l'IRM quantitative à haute résolution un outil indispensable et courant dans les laboratoires d'ingénierie tissulaire, accélérant ainsi le développement de tissus biologiques fonctionnels et cliniquement pertinents.