Metamaterial Liner for MRI Excitation—Part 2: Design and Performance at 4.7T

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ABSTRACT The theoretical foundations of a metamaterial (MM) liner for the MRI bore that facilitates the propagation of reduced-cutoff cylindrical waveguide modes are presented in the Part 1 companion paper to this work. Here, in Part 2, the practical design and modelling of the novel MM liner is applied to a body MRI radio-frequency (RF) transmitter for 4.7T. An equivalent network mesh model is developed to reduce the distributed structure of the MM to one that uses lumped discrete tuning elements. The close match between full-wave simulation and the network model is demonstrated by comparison of the longitudinal propagation dispersion curves and input impedance. The application of the methods developed for analysis of the MM liner behavior are demonstrated in the design of a practical MM liner as an effective MRI RF transmitter. The liner’s simulated performance was evaluated using three key metrics of MR radio-frequency (RF) coil performance: transmit efficiency, the variation of the transmit field (a measure of homogeneity), and the 10g averaged local specific absorption rate (SAR). These metrics are compared to those of the commonly used birdcage (BC) coil. The main advantage of the MM liner is the 29.6% reduction in maximum SAR normalized to transmit field relative to the BC coil, with comparable transmit efficiency and homogeneity. Thus, the liner can potentially replace the BC coil as an RF transmitter and provide an enhanced safety margin, or allow the use of faster, more SAR-aggressive imaging sequences.

INDEX TERMS Circuit model, coil, dispersion, effective medium, magnetic resonance imaging, MRI, metamaterial, network, radiofrequency, ultra-high field, waveguide.

I. INTRODUCTION

The objective of transmit (Tx) radio-frequency (RF) coil design for MRI is to produce homogeneous transverse RF magnetic fields efficiently for MR excitation of a targeted imaging region of interest (ROI). The excitation ($B_1^T$) fields must be generated without creating concentrated electric fields, and thus, local specific absorption rate (SAR) that exceeds the IEC safety limits [1]. The transverse magnetic field for the 1st-order transverse electric (TE) mode of a circular waveguide is largely homogeneous, while producing electric fields that are also spread out more uniformly than conventional MRI coils, and therefore may produce lower local SAR for the same $B_1^T$. In the Part 1 companion paper to this work [2] the design of a metamaterial (MM) liner is presented, which consists of concentric rings, separated longitudinally and placed on the inside surface of the MRI bore (Fig. 1). The derivation of the MM properties and tuning for operation of the desired cylindrical waveguide mode at the Larmor frequency (4.7T–200MHz) demonstrates that such a MM liner structure could effectively produce a $B_1^T$ field suitable for MRI. Thus, this work details the practical design and excitation of a whole-body MM liner of finite length, comparing it to a conventional MRI RF excitation coil using the metrics described below.

The signal-to-noise received from a sample-dominated resonator/coil increases approximately proportionally to the static magnetic field strength ($B_0$). Additionally, the frequency of operation $f_0$ increases proportionally by the Larmor equation $f_0 = \gamma B_0$, where $\gamma$ is the gyromagnetic ratio of the nucleus (typically $^1H$, $\gamma = 42.7$ MHz/T). Interference patterns appear with increasing field strength due
regions of low B\textsubscript{1} intensity for the local SAR. For human body imaging
and dark areas of the transmit field and regions of high and
to wavelength effects in the sample, leading to constructive
directive field interactions that are observed as bright
and dark areas of the transmit field and regions of high and
low intensity for the local SAR. For human body imaging
this manifests in central brightening in the human head, and
regions of low B\textsubscript{1} in the anterior and posterior regions of the
human torso when imaged with a volume coil [3]. The field
strength of 4.7 T is chosen for this study, since it is the first
available field strength above the clinical 3T where local SAR
and homogeneity begins to severely constrain the application
of many MR imaging methods in the trunk and internal organs
(e.g., abdominal, thoracic, heart, lung). However, the methods
developed in this paper can be applied to any field strength
or bore size. The Larmor frequency greatly affects the key
performance metrics of transmit coils for MRI. The most
significant metrics are:

1. The transmission efficiency (\(\eta_{\text{Tx}}\)), which is a measure
   of the MR-relevant right circularly polarized magnetic
   field (\(B_1^+\)) produced for a given input power (units of
   \(\mu T/\sqrt{W}\)). The transmission efficiency is reported as
   the mean within a suitable region of interest (ROI).

2. The transmit field variation (\(\%\sigma_{\text{Tx}}\)), measured as the
   percent coefficient of variation (\(\%\text{CV}\)) of the \(\eta_{\text{Tx}}\)
   within the ROI.

3. The SAR. At high \(B_0\) MRI safety constraints are
   typically limited by the local \(10g\) averaged SAR (units of
   W/kg) [1]. The constraint imposed on the transmit
   field is expressed by reporting the mean transmit field
   produced relative to the 10g averaged local SAR, which
   is termed the safety excitation efficiency (\(\eta_{\text{SEE}}\), given
   in units of \(\mu T/\sqrt{W/kg}\)). The maximum 10g averaged
   SAR produced in the body is constrained by safety
   limits.

There are therefore three associated key challenges in RF
Tx coil design at field strengths \(\geq 3T\) that motivate this work:

1. The ohmic losses in conductive tissue are roughly
   proportional to the square of the frequency, while the
   field induced for a given current is roughly proportional
to the frequency: thus \(\eta_{\text{Tx}} \propto 1/\omega_0\).

2. For \(B_0 \geq 3T\) (\(f_0 \geq 128\) MHz) the wavelength (\(\lambda\)) inside
   the human body (with bulk relative permittivity \(\varepsilon_B = 40 - 60\)) is short relative to the dimensions of the human
   body (\(< 30\) cm). Thus, the regions of constructive and
destructive interference of the transmit field that arise
due to wavelength effects, termed interference patterns
[3], increases \(\sigma_{\text{Tx}}\).

3. The interference patterns and higher power require-
   ments to produce the same Tx field result in increased
   local 10g SAR, which constrain the Tx field amplitude
   and duty cycle that may be safely used in MRI scanning.

In addition, for whole-body excitation the access to the
center of the MRI scanner bore must not be impeded. Except
for dedicated localized Tx coils, the RF applicator is restricted
to a minimal lining around the scanner bore’s cylindrical
surface.

Numerous approaches have been taken in RF coil design
to address these challenges and a few popular volume coil
designs have emerged: the TEM coil, the birdcage coil
(BC), and transceive arrays using resonant elements such as
microstrips or dipoles. The SAR with the TEM coil and BC
is comparable, and studies comparing the two have shown
largely the same localized SAR and transmit field, with
sometimes minor disagreement between slightly different
studies [4]–[6]. In general, if optimal designs are used to
compare each Tx coil type, the \(\eta_{\text{Tx}}\) and \(\eta_{\text{SEE}}\) will be similar
when they have similar dimensions and/or the same \(B_1\)
coverage area, as the losses will be dominated by total losses
in the body. Thus, at high fields the efficiency is constrained
by the electric field associated with the desired \(B_1\) field,
which will be similar for all coil types [7].

Improvement in \(\eta_{\text{SEE}}\) can be achieved with transceive
arrays [5],[9],[10], where reducing the \(\%\sigma_{\text{Tx}}\) and SAR is
still an area of ongoing research [10], [11]. There are many
array element designs that attempt to improve metrics 1–3
without, or in addition to, high-density element shimming
(i.e., beamforming) techniques. In the head at 7T, the “tic-
tac-toe” coil [12] has improved \(\eta_{\text{SEE}}, \eta_{\text{Tx}}\) and \(\%\sigma_{\text{Tx}}\) than a
TEM coil with similar dimensions. In another approach, the
original travelling wave (TW) MRI [13], [14] was modified
to improve the efficiency while maintaining low SAR by the
use of stepped impedance resonators to focus the fields [15],
by using different exciting patch antenna or resonator designs
[16], or by modification of the waveguide structure itself
so that propagation is no longer dominated by the scanner
bore (e.g., by using a parallel plate waveguide structure in
reference ). In addition, TW structures like the helical antenna
have been shown to efficiently produce the desired transmit
field distribution, while also allowing the currents to be
distributed along the bore [17], [18], and at high frequencies
(\(> 7T\)), when operating in the fast-wave/axial mode regime,
circularly polarized fields are produced naturally [19], [20].

Recently, a theoretical [21], [22] description has been
developed for a thin lining of MM that reduces the cut-off
of hybrid transverse electric (HE) and hybrid transverse
magnetic (EH) modes in circular metallic waveguide envi-
ronments [21]. A fully anisotropic MM liner was produced.
using periodically reactively loaded and axially stacked rings, and was found to also improve Tx performance [23]. In the Part 1 companion paper to this work [2], an accurate effective medium model was presented for the MM liner and a method of design for a MM lined MRI bore was introduced. This work (Part 2) explores the practical design and implementation of the MM liner as an MRI coil. The practical implementation first requires developing a network model that can predict the current pattern, input impedance and tuning frequency of the MM liner. Subsequently, the application for MRI is investigated by comparing the MM liner to the conventional BC according to the Tx metrics described above.

The geometric description and network analysis of the MM liner is presented in section II, demonstrating three different reactive-loading scenarios resulting in a variety of useful propagation dispersion features suitable to create the desired current distribution and resonance mode for MRI at 4.7T (200 MHz). In section III full-wave simulations of the MM liner are compared for the three cases to results from network analysis, demonstrating the dispersion characteristics and the expected transverse field distributions. Finally, in section IV.B the operation of the full-scale MM liner is evaluated by simulation of the MR relevant Tx performance metrics (ηTx, %σTx and ηSEE) with a realistic human body model.

II. MM LINER STRUCTURE AND MODELS

A. GEOMETRY AND DESIGN SIMPLIFICATION

The MM liner developed in this study is a practical realization of the geometry shown in Fig. 1. A lining of anisotropic material with a thickness t, relative permittivity \( \varepsilon_r \) and relative permeability \( \mu_r \) on the inside of a circular waveguide bounded by a perfect electric conductor (PEC) of radius \( b \) alters the propagation characteristics of its waveguide modes [21], [23]. In Part I of this work, the dispersion of the effective material properties (assumed to be diagonal tensors \( \varepsilon_r \) and \( \mu_r \), \( \varepsilon_r = \varepsilon_0 \) diag(\( \varepsilon_1, \varepsilon_2, \varepsilon_3 \)), \( \mu_r = \mu_0 \) diag(1, 1, 1)), was derived for closely-spaced rings for EH01 and each HE2m1 mode (\( n \) is the azimuthal mode order). These rings included periodically-arranged radial inductors between rings and ground (PEC), and azimuthally directed capacitors that can be tuned to achieve the desired mode and longitudinal propagation at the desired frequency. Furthermore, it was demonstrated how different propagation and mode-spacing characteristics were observed for \( \varepsilon_r \) negative and near zero (ENNZ) with \( \mu_r \leq 1 - 5 \) compared to \( \mu_r \) negative and large (MNL), i.e., \( \mu_r \leq -10^2 \), with \( \varepsilon_r \) near zero but positive.

In this work, each constituent ring of the MM liner is realized using the structure shown in Fig. 2(a), consisting of \( 8 \) identical sections repeated azimuthally. The structure differs from that of Refs. [23], [24] by employing discrete longitudinal connections (reactances) that permit increased distance between rings and adds a degree of freedom for tuning the structure. Thus, a more practical alternative to the theory developed in Part 1 is required for analysis. The section close-up shown in Fig. 2(b) (a single segment of a ring) includes the components and dimensions that may be adjusted to tune the MM liner:

- Radially oriented (r-directed) capacitors (\( C_r \)) and inductors (\( L_r \)), with short connections between ground and the ring.
- Longitudinally oriented (z-directed) capacitors (\( C_z \)) and inductors (\( L_z \)), which connect adjacent rings together.
- Azimuthally oriented (\( \phi \)-directed) capacitors (\( C_{\phi} \)).

A \( L_{\phi z} \) is not represented explicitly in the diagram, since it would not be added to the large intrinsic inductance of the ring’s conductive segments. Practically, the inductors are lumped air-core inductors, while the capacitors are either high-quality-factor chip capacitors, or parallel-plate capacitors, as shown in Fig. 2(b). The parallel plate capacitor implementation consists of a length \( l_{\phi z} \) and width \( w_{\phi z} \) of overlapping conductor separated by the ring’s structural dielectric substrate, with thickness \( t \) and permittivity \( \varepsilon_s \).

The full-wave simulations in this study are performed with the finite element full-wave simulator HFSS.
(Ansys, Canonsburg, PA). A simplified model of Fig. 2(c), which is expected to be a close approximation of the physical model in Fig. 2(a-b). The physical components \( C^r, L^M_r, C^\phi, L^M_\phi, \) and \( C^\phi \) are replaced with lumped series reactances corresponding to the equivalent circuits of chip capacitors and air-core inductors (with inductance found from equations for single layer air core solenoid coils [25]). The network model developed in the next section applies to both physical and simplified models. Full-wave simulation, however, does not provide a comprehensive or intuitive understanding of the operation of the MM liner and requires substantial computing resources and time, thus precluding extensive exploration of the parameter space necessary for optimization. Therefore, a method of predicting the tuning frequency and input impedance from a few design parameters is a valuable tool to design the MM liner and compare its behaviour to the BC, the TEM coil [5], or other MRI transmit coils [12], [15], [18].

B. NETWORK MODEL

The propagation of travelling waves in an MM lined waveguide is dominated by the MM liner structure itself, which closely resembles a two-dimensional transmission line. Lumped element network models can accurately represent transmission lines (coplanar lines, coaxial lines, etc.) [26]. Thus, the network representation developed here mimics that of two-dimensional transmission-line matrix methods [27].

The model shown in Fig. 3(a) is a general lumped-circuit network representation of the interconnected ring structure of the MM liner and includes the tuning elements shown in Fig. 2. To allow for different tuning components on different rings and different segments of the same rings, the subscripts \( n_\phi \) and \( n_\phi \) are explicitly included. Thus, there are \( N_\phi \) meshes in the \( \phi \)- (azimuthal) direction associated with each ring, with the \( N_\phi \)-th mesh connected to the first. Similarly, the network model considers \( N_z \) rings spaced in the \( z \)- (longitudinal) direction.

The network model of Fig. 3(a) shows only the lumped impedances of the network, consisting of LC models of chip or parallel plate capacitors, stray capacitances and air-core inductors in the implementation presented here. It does not show the intrinsic inductances of the MM liner structure, which in this model are included as “geometric” inductances \((L^G_r, \phi, z)\) in series, so the total branch inductances \((L^r, \phi, z)\) are

\[
L^r, \phi, z = L^M_r, \phi, z + L^G_r, \phi, z. \tag{1}
\]

The mesh current paths of the ring structure are depicted in Fig. 3(b), each of which is associated with an intrinsic self-inductance that defines the values of \( L^G_r, \phi, z \) as well as having an associated mutual inductance with every other mesh. In the network model of the lumped impedances, there are \( N_\phi \times N_z \) of both the \( \phi \)-, and \( r \)-directed impedances \((Z_{n_\phi n_\phi}^r, Z_{n_\phi n_z}^r)\), but \( N_\phi \times (N_z - 1) \) \( z \)-directed impedances \((Z_{n_\phi n_z}^z)\).

The mesh currents and voltages are matrices expressed as

\[
\begin{bmatrix}
I^\phi_{1(1\rightarrow N_\phi)} & I^\phi_{2(1\rightarrow N_\phi)} & \cdots & I^\phi_{N_z(1\rightarrow N_\phi)}
\end{bmatrix}^\dagger,
\]

\[
\begin{bmatrix}
V^\phi_{1(1\rightarrow N_\phi)} & V^\phi_{2(1\rightarrow N_\phi)} & \cdots & V^\phi_{N_z(1\rightarrow N_\phi)}
\end{bmatrix}^\dagger. \tag{2a}
\]

\[
\begin{bmatrix}
V^z_{1(1\rightarrow N_\phi)} & V^z_{2(1\rightarrow N_\phi)} & \cdots & V^z_{N_z(1\rightarrow N_\phi)}
\end{bmatrix}^\dagger. \tag{2b}
\]

where \( \dagger \) indicates transpose, and the superscript indicates whether the currents or voltages correspond to the \( r, \phi, z \), or \( z \)-oriented meshes in Fig. 3. The general network equation that relates the matrices of currents and voltages is simply

\[
I = (Z^G + Z^M)^{-1} V = Z^{-1} V. \tag{3}
\]

The impedance matrix \( Z \) accounts for the lumped impedances in Fig. 3(a), as well as the geometrically dependent inductances of the current meshes in Fig. 3(b). To account for nearest neighbor mutual inductance between meshes of different rings, such as the meshes associated with \( I^\phi_{11} \) and \( I^\phi_{21} \) in Fig. 3(b), only three key mutual
inductance terms \((M^{r\phi}, M^{\phi r}, M^{\phi \phi})\) were included to account for the magnetic flux between the nearest-neighbor mesh currents. See the supplementary materials for details on how the elements of the impedance matrix \((Z^{r,\phi,z})\) and mutual inductance terms are assembled.\(^1\) Commercial circuit simulators can solve the network model, but the mutual impedance terms are not easily included. Implementation is far less cumbersome using the concise matrix method implemented here in MATLAB. The code employed to implement the network model for input impedance calculation and 3D model of the metamaterial liner eigenmode simulation and full-scale simulation have been included in the supplementary files. Results were validated by comparison (not included) to Electronics Desktop Circuit Simulator.

The case of two physically-separate \(C^\phi\) tuning capacitances between each radial impedance matches the case shown in Fig. 4, which best represents the distributed capacitance obtained by the overlapping strips in Fig. 2(a–b). Also, the distributed inductance and parasitic capacitances \((C^{dr}, C^{d\phi}\text{ and } C^{dz})\) are represented more faithfully by including multiple azimuthal sub-segments for each radial element. In this example, three mesh sub-segments are included for every radial element, so that \(N_\phi = 24\) in Fig. 4. All elements of \(Z^{n_z,n_\phi}_{r,\phi,z}\) are represented by series LC circuits in parallel with stray distributed capacitances between adjacent ring conductors \((C^{dz})\), between the ring conductor and ground \((C^{rd})\) and between azimuthal segments \((C^{d\phi})\). Additionally, for a symmetrically constructed MM liner all mesh tuning elements and geometric parameters will be equal for each \(n_z\) and \(n_\phi\). To exclude an \(n_z,n_\phi\) lumped inductor or capacitor we set \(j\omega L^{n_z,n_\phi}_{r,\phi,z} = 1/j\omega C^{n_z,n_\phi}_{r,\phi,z} = 0\).

\(^1\)Supporting document describes the derivation of the impedance matrix from the network model (Supplementary Material I. Derivation of Impedance Matrix)

Therefore, the elements of \(Z^{r,\phi,z}\) are

\[
Z^{n_z,n_\phi}_{r,\phi,z} = \begin{cases} 
\frac{(j\omega L^r + 1/j\omega C^r)}{j\omega C^{rd}}, & n_\phi = 3, 6, 9, \ldots N_\phi \\
\frac{1}{j\omega C^{rd}}, & \text{otherwise}
\end{cases}
\]

\(4\text{a})

\[
Z^{n_z,n_\phi}_{r,\phi,\phi} = \begin{cases} 
\frac{(j\omega L^\phi + 1/j\omega C^\phi)}{j\omega C^{\phi d}}, & n_\phi = 1, 3, 4, 6 \ldots N_\phi \\
L^\phi, & \text{otherwise}
\end{cases}
\]

\(4\text{b})

\[
Z^{n_z,n_\phi}_{r,\phi,z} = \begin{cases} 
\frac{(j\omega L^z + 1/j\omega C^z)}{j\omega C^{zd}}, & n_\phi = 3, 6, 9, \ldots N_\phi \\
\frac{1}{j\omega C^{zd}}, & \text{otherwise}
\end{cases}
\]

\(4\text{c})

In Fig. 4 the circuit model of the MM liner is presented, and the inductors and capacitors shown in green in this figure are the longitudinal electrical impedances between rings that consist of distributed stray capacitance \((C_{zd})\), inductances that are either lumped elements or are geometrically dependent as they result from the conductive traces \((L_z)\) and lumped capacitors placed in series with those conductive traces \((C_z)\). In MM liner designs such as that presented in Ref. [23], or the MM liner presented in Part 1 [2], there may be no conductive traces between rings, and the coupling between rings will be only due to \(C_{zd}\) and the mutual inductance between the azimuthally directed currents on the rings.
III. MODEL PARAMETERS AND ANALYSIS

The MM liner is a complex EM structure with many tuning and geometry-dependent parameters. Thus, this section explores the effects of each variable on the overall behavior of the structure. Unlike with other MRI resonators (loop coils, BC, etc.) the MM structure must be analyzed by considering the fields and current/voltage distributions as longitudinally propagating waveguide modes with complex propagation constants defined as \( \gamma_n(\omega) = \alpha_n(\omega) + j\beta_n(\omega) \), where the propagation constant is dependent on the mode order \( n \). The \( \beta_n(\omega) \) represents propagation and will be reported here in relation to the free-space wavenumber \( k_0 = \omega/c \). The \( \alpha_n(\omega) \) is associated with attenuation and decaying or evanescent modes. The propagation constant is used to determine the tuning values for a given length of MM liner. Specifically, for an efficient excitation in the middle of the MM liner a length \( (\Delta z N_z) \) of \( \beta_n(\omega)/\Delta z N_z/2\pi = m_z \sim 1/2 \) is required, where \( m_z \) is defined here as the longitudinal resonance mode order: \( m_z = \{0, 1/2, 1, 3/2 \ldots\} \).

A. DISPERSION ANALYSIS FROM NETWORK MODEL

The complex propagation constant \( \gamma_n(\omega) \) and the radial distribution of the fields defined by \( H_{r,\phi,z}(r) \) and \( E_{r,\phi,z}(r) \) have been rigorously derived in Refs. [21], [23]. The EM fields of the different modes of the structure shown in Fig. 1 are represented in a simplified form as

\[
H = e^{-\gamma_n(\omega)z} (H_r(r) \sin (n\phi) \hat{r} + H_\phi(r) \cos (n\phi) \hat{\phi}) + H_z(r) \sin (n\phi) \hat{z} \\
E = e^{-\gamma_n(\omega)z} (E_r(r) \cos (n\phi) \hat{r} + E_\phi(r) \sin (n\phi) \hat{\phi}) + E_z(r) \cos (n\phi) \hat{z}.
\]

They are dependent on the azimuthal mode order \( n \), angular frequency \( \omega \), \( \hat{\theta} \) and \( \hat{\phi} \). For each value of \( n \) there are two classes of solutions, the HE\(_{n1} \) and EH\(_{n1} \) modes, whose cutoff frequencies are most dependent on \( \{\epsilon_r, \mu_z\} \) and \( \{\epsilon_z, \mu_\phi\} \), respectively. Only the HE\(_{n1} \) modes (the second indices in HE\(_{n1} \) indicates radial mode order) and EH\(_{01} \) mode are investigated, since the ring structure under investigation supports primarily azimuthal currents that affect \( \mu_z \) and \( \epsilon_\phi \). To control the fields and prevent mode mixing, the MM liner is designed such that the propagation of the EH\(_{n1} \) and HE\(_{n1} \) modes do not occur over the same frequency range.

For the purposes of this work, we only focus on the real part of the propagation constant, \( \beta_n \), which determines the longitudinal resonance modes over the finite length of the MM liner. For an MM resonator to work efficiently it will also require \( \alpha_n(\omega) \ll \beta(\omega) \), so that over the length of the imaging region the field decay is small.

To relate the results of the network model to the propagation constant of the MM liner, the currents on the outer conductor (PEC boundary) of the bore are modelled as

\[
I(b) = \sum_n e^{-\gamma_n(\omega)z} \left( A_n \cos (n\phi) \hat{z} + B_n \hat{\phi} \sin (n\phi) \right).
\]
lattice pair boundary faces. For comparison, the network model requires the inversion of \( Z \) for each frequency point, which is a \( 4,608 \times 4,608 \) sparse matrix and requires much fewer computing resources than the HFSS simulation (can be performed on a desktop computer with < 2 GB RAM).

Part 1 [2, p. 1] demonstrates that, in the case that the MM structure includes radial inductors that are close to resonant with the stray capacitance between ring conductors segments and ground, the cutoff of HE\(_{n1}\) modes will occur below the EH\(_{01}\) mode. In the mode’s backward-propagating region the reactance of the azimuthal segments will be capacitive and the effective \( \varepsilon_{\phi} \) will be ENNZ. Additionally, close to the HE\(_{01}\) mode, where the series azimuthal reactance is near-zero, the cutoff of forward propagating HE\(_{n1}\) modes will occur at higher frequencies when the azimuthal reactances are inductive for radial capacitors (high-pass (HP)), or lower frequencies when the azimuthal reactances are capacitive for radial inductors (low-pass (LP)). The effective \( \mu_r \) will be MNL (\( \varepsilon_{\phi} \) near zero at cutoff and positive in propagating region). These three different cases indicated below are considered in this study:

- **ENNZ**, \( C^z = 1 \text{ pF}, L^M = 107.5 \text{ nH} \),
- **HP-MNL**, \( C^\phi = 15.8 \text{ pF}, C^i = 20 \text{ pF} \),
- **LP-MNL**, \( C^\phi = 10 \text{ pF}, L^M = 30 \text{ nH} \).

The geometric parameters were adjusted to achieve approximately the same cutoff values, mode spacing and dispersion slope with the network model as in the full-wave simulation, for all three cases, as described in the supplementary materials:

- \( L^{G^z} = 2.5nH, L^{G^\phi} = 19nH, L^{G^i} = 31nH, C^{\phi d} = 0.5 \text{ pF}, C^{id} = 3.3 \text{ pF}, C^{cd} = 0.5 \text{ pF}, M^{\phi\phi\phi} = M^{\phi\phi\phi} = -0.225, M^{\phi\phi\phi} = -0.01 \text{ nH} \).

Additionally, variation of the longitudinal tuning parameters \( C^z \) and \( L^z \), as described in the supplementary materials, was performed to determine the values that result in greatest slope in the dispersion diagrams (separation of longitudinal resonance modes, \( m_z \)) and prevent overlap of the HE\(_{n1}\) modes at the Larmor frequency (4.7T - 200 MHz), for the three cases:
- **ENNZ**, \( C^z = 6 \text{ pF}, L^z = 0 \text{ nH} \),
- **HP-MNL**, \( C^z = \infty \text{ pF}, L^z = 0 \text{ nH} \),
- **LP-MNL**, \( C^z = \infty \text{ pF}, L^z = 30 \text{ nH} \).

Without longitudinal conductive traces, the stray capacitance between rings (represented by \( C^{\phi d} \)) and the mutual inductance (primarily represented by \( M^{\phi\phi\phi} \)) dominate the coupling and propagation of the electromagnetic field. If \( \beta \ll \Delta z \) the dispersion vs. frequency will be nearly flat, resulting in a small frequency spacing between longitudinal mode resonances. Thus, the use of longitudinal conductive traces to introduce coupling between rings is highly advantageous by extending the allowable spacing between rings, while maintaining large frequency separation in the longitudinal modes.

Fig. 5 shows the resulting \( \beta/k_0 \) dispersion diagrams for the MM liner predicted using the network model, compared to those predicted using the HFSS eigenmode solver, with geometric and longitudinal tuning impedances adjusted manually to obtain the best match. The full-wave simulation is expected to be closer to the ground truth because it accounts for mutual impedance between all meshes, the actual propagation contribution from the scanner bore RF shield, as well as allowing more general distributions of current and voltage across the MM liner structure. However, the network model is abundantly accurate (within a few percent), making it valuable in practice for design optimization and to provide working estimates of the tuning values. The model has the greatest discrepancy with the HP-MNL case since it is the most dispersive and has a stronger dependence on the longitudinal impedance.

It was demonstrated in the Part 1 companion paper [2, p. 1] that the low Q-factor (< 200) of large radial inductances will result in significant loss of transmit efficiency. For MRI excitation, low losses are required and only a single HE\(_{11}\) longitudinal \( m_z = 1/2 \) resonance must be excited. Only the MNL high-pass case with direct longitudinal connections...
is free of other $\text{HE}_{n1}$ modes at 200 MHz and avoids the use of lossy inductors. Thus, the MNL high-pass case is best for the practical application envisioned here and considered for the rest of this study.

C. ELECTROMAGNETIC FIELD DISTRIBUTION

The longitudinal and transverse components of the simulated electric and magnetic field magnitudes in a central transverse plane of the eigenmode unit cell simulation (Fig. 2c) for the three cases are shown in Fig. 6. The $\text{HE}_{11}$ mode is the only one that provides the homogeneous transverse magnetic field desired for MRI excitation. For each $\text{HE}_{n1}$ mode the field distribution for the three cases is similar, with minor variations due to differences in the induced current distributions. The discrete structure of the MM liner approximates the continuous structure of the ideal MM liner depicted in Fig. 1(a), which results in a staircase or sampled approximation of the purely azimuthal sinusoidal distribution, and leads to regions of periodic nodes and peaks in the field magnitudes, especially near the circumference. The method of excitation and the boundaries at the ends of the liner influence the imposed current distribution; therefore, to take these effects into consideration it is necessary to evaluate the MM liner as a resonator with finite length, as in the next section.

IV. PRACTICAL METAMATERIAL LINER

A. METAMATERIAL LINER AND BIRDCAGE DESIGN

The 4.7 T scanner intended for experimental verification (Varian, University of Alberta, Peter S. Allen Research Center) produces image encoding magnetic field gradients with 4% linearity variation in a sphere with diameter of 25 cm, and thus the RF resonator length is designed to approximately match this effective longitudinal coverage ($N_z = 16$, 95 cm length). In our prior publication comparing the efficiency of a BC and MM liner without longitudinal connections it was observed that for the same longitudinal coverage the MM liner must be approximately double the length of the BC [28].

The full-scale MM liner model is shown in Fig. 7(a), and the BC used for comparison with comparable longitudinal coverage is shown in Fig. 7(b). The approximate electrical properties of the substrate for the MM liner rings (shown in green in Figure 2(c) and Figure 7) are those of a fiberglass reinforced low-loss substrate ($\varepsilon_r = 3.4$, and loss tangent of $\tan \delta = 0.001$). The values for $C^\phi$ and $C^r$ are those of the HP-MNL case of the MM liner. The $C^\text{ring}$ and $C^\text{leg}$ capacitor values for the hybrid BC (16.3 pF and 7.2 pF, respectively) were derived by algebraic methods considering the simulated impedance matrix of the meshes of the BC [29]. To include realistic losses, equivalent series resistances were added to the $C^\phi$ and $C^r$ capacitors of the HP-MNLMNL liner (0.0336Ω and 0.0265Ω, respectively) and $C^\text{ring}$ and $C^\text{leg}$ capacitors for the hybrid BC (0.0325Ω and 0.0737Ω, respectively), which correspond to a quality factor of 1500 at 200 MHz.

A hybrid BC topology is used, rather than a LP (where capacitors are placed only on the endrings) or a HP (where capacitors are placed only on the legs) or a HP (where capacitors are placed only on the legs) or a HP (where capacitors are placed only on the legs) or a HP (where capacitors are placed only on the legs) or a HP (where capacitors are placed only on the legs). For a shielded body coil design at 4.7T-200MHz the hybrid topology is necessary, otherwise the inductive reactance of the segments would become large relative to the stray capacitance between the conductor and the shield, and the desired birdcage mode could not be excited [30]. By distributing series capacitors on both end-ring and leg conductor segments this issue is avoided. In addition, the hybrid topology has a more
balanced field/voltage distribution across its structure leading
to improved insensitivity to loading and avoiding large
conservative electric fields that would increase the localized
SAR [30], [31]. A 1.4 m long conductive shield with 55 cm
diameter encompasses both designs in simulation to account
for the outer conductor (RF shield) of the bore. Both the MM
liner and BC coil are fed by lumped ports placed in series
with capacitors on the end-rings.

The input impedance of one port, with the other quadrature
port terminated in a short-circuit, was calculated using the
network model in Fig. 4 and with full wave simulation of
the structure shown in Fig. 7(a). The Visible Human Project female human body model 2.0 with standard human
tissue EM values was used as the numerical MRI phantom [32]–[34]. The body model was scaled to 92.5% of the
original in the transverse direction to fit within the available
diameter (matching that of the intended 4.7 T scanner).

To include realistic losses, the conductive elements of the
full-scale BC and MM liner, as well as RF shield, were
modelled as copper sheets (conductivity $\sigma = 5.8 \times 10^7 S/m$)
with a surface roughness of 1 $\mu$m. Simulation of the input
impedance was performed both with and without the human
body model centered in the liner. With the human body model
present, to solve a single frequency point at 200MHz the
direct solver used a mesh of 5236892 tetrahedra (first-order
basis functions with 33631292 matrix), requiring 57.6 GB RAM and 35-minute simulation time for each frequency point.

To emulate the effect of additional losses in the human
body, the input impedance is also calculated when the
equivalent series resistances added to the $C^\phi$ is increased by
a factor of 5 (to 0.168 $\Omega$).

The input resistance and reactance calculated using the
network model are provided in Fig. 8(a) and Fig. 8(b),
respectively, and simulated using HFSS in Fig. 8(c) and
Fig. 8(d), respectively. Since this is a truncated structure,
standing waves are formed leading to resonance peaks in the
reflection coefficient and input impedance associated with the
longitudinal resonance modes ($m_z$, see Section III). Note that
such resonances could not exist in the bore alone because it
is below cutoff, and are, in fact, enabled by the MM liner.

Below and near the cutoff of the $HE_{01}$ mode, interfering
resonances are observed due to propagating $EH_{11}$ modes,
which involve the longitudinal inductive connections and the
capacitances of the azimuthal sections. The propagation of
these modes terminates at the cutoff of the $HE_{01}$ mode. The
first few $m_z$ longitudinal standing wave modes for the $HE_{11}$
mode are indicated by arrows in Fig. 8(b) and Fig. 8(d).
Although the frequencies and input impedance for the full-
wave simulation and network model do not match exactly,
the trend and approximate spacing between resonances have
a similar level of agreement to analytical modeling vs.
measurement or simulation of BC modes.

In the presence of the body the frequency of the resonances
appears to be slightly increased (by less than 0.5%). This
small change may be a result of the currents induced on the
conductive human body effectively reducing the inductance
of the MM liner rings. This effect is observed in highly loaded
BC coils or local transmitters as well. In practice this effect
may be mitigated by adjustable tuning capacitances on the
first and last rings of the MM liner structure. As expected,
increased losses broaden the impedance peaks for both the
network model and full-wave simulation.

The azimuthally-directed current distributions for the first
four resonances of the $HE_{11}$ mode simulated from the
network model are compared to those on the cylindrical RF
shield from the full-wave model in Fig. 9(a) and Fig. 9(b),
respectively. The currents on the shield mirror those on the
rings, with nulls observed between the longitudinal spacings
of the ring segments. Peaks in the full-wave simulated current
distribution occur at the locations of the radial elements
and longitudinal wires. The network model only represents
the currents on the conducting wires of the MM structure
and does not fully reproduce all spatial EM aspects and the
distributed impedances of the structure; these include, for
example, mutual impedances between non-adjacent rings, the

![FIGURE 7. Diagram of (a) MM liner and (b) BC used for comparison of MRI metrics with dimensions and capacitors.](image-url)
FIGURE 8. The real and imaginary part of the input impedance for a port at the end-ring of a liner with $N_z = 16$ rings for the HP-MNL case simulated by network parameter analysis ((a) and (b)) or full wave analysis ((c) and (d)). The case with and without the body model (low loss vs. high loss case for the network model) is shown as green dashed lines or red solid lines, respectively. The first five longitudinal resonances ($m_z$) for the HE$_{11}$ mode are labelled along with the relative location of the HE$_{01}$, HE$_{11}$ and HE$_{21}$ mode propagation frequency ranges.

FIGURE 9. Currents ($I_\phi$) derived from (a) mesh network calculation and (b) full-wave simulation on the RF ground/shield for longitudinal resonance of the HE$_{11}$ mode for the radial capacitance case with ($L_{M} = 0$).

EM fields in the central free-space volume of the bore, and interaction of fringe fields at the terminating end rings with the MR bore. This may explain why the frequency spacing between the zeroth resonance and $1^{st}$ $m_z = 1/2$ resonance is smaller in the network model than in the simulation (Fig. 8 and Fig. 9).

Gradient-induced eddy currents are not expected to be an issue for this design because DC loops are absent. All azimuthal conductive segments are split up by the regularly spaced $C^\phi$ capacitors, which act as high impedance blocks for the low-frequency (<200 kHz) eddy-currents and greatly damp them. In the case of an added external shield other than the typical gradient shield, a slotted design and/or capacitive bridges would need to be employed to mitigate eddy currents.

The network model nevertheless captures the essential characteristics of the novel MM liner and thus is a useful tool for design optimization and comparison. Calculation of the EM fields and MR-relevant metrics (c.f., Section I) in presence of the complex geometry of the human body, however, requires numerical full-wave solvers.

B. COMPARISON OF TRANSMISSION METRICS

As discussed in Section I, the $\eta_{Tx}$, $\%\sigma_{Tx}$ and $\eta_{SEE}$ are the three essential metrics for comparison of transmit coils for MRI. Simulations of the transmit efficiency were performed with or without capacitor losses (zero equivalent series resistance) to quantify its impact. Simulations were also performed with the body (abdominal imaging) centered on the coil. Tuning was performed to achieve an isolation between quadrature ports of 25 dB or better with both the MM liner and BC designs by adjusting the reactance of lumped ports on the diametrically opposite sides of the excitation ports. The input power for excitation was adjusted to deliver 500 W into each excitation port with 90° phase difference.

The results for the simulated $B_1^+$ in the bore when it is empty are shown in Fig. 10. Both resonators have similar mean transmit efficiencies (within 3%) inside the outlined FOV (which covers a central section of the human body...
FIGURE 10. Central slices displaying the $B^+_1$ field produced inside the MRI bore when it is empty for the (a) MM liner and (b) BC designs at 4.7T-200MHz. The fields are normalized to the mean transmit efficiency ($\mu T/\sqrt{\text{KW}}$) in the outlined torso region.

TABLE 1. The mean transmit efficiency for the MM liner and BC designs in the previously described FOV. The cases for with and without the human body model are compared and with and without capacitor loss.

| Coils                  | Empty Bore | Body model |
|------------------------|------------|------------|
| MM – No Capacitor Loss | 8.39, 2.6% | N/A, N/A   |
| MM – With Capacitor Loss | 6.79, 4.2% | 2.36, 40.0% |
| BC – No Capacitor Loss | 9.61, 2.5% | N/A, N/A   |
| BC – With Capacitor Loss | 8.20, 2.1% | 2.74, 41.8% |

$\pm 10\text{cm}$ from the center longitudinally), as listed in Table 1. Additionally, the $\%\sigma_{TX}$ is very small ($<2.5\%$) for both designs within the outlined FOV, and they both have similar longitudinal coverage. near the current-carrying rungs the BC has regions of inhomogeneity.

Interestingly, the transverse distribution of the simulated $B^+_1$ within the human body model, shown in Fig. 11, is similar, showing an interference pattern with two nodes in the anterior-posterior direction due to the dominant wavelength effects in the body. However, the MM liner shows regions of much greater excitation near the head, that are not present in the empty bore simulation (Fig. 10). This is due to two factors: by acting as a waveguide there is a longitudinal reflection of the excited wave at the body boundary; and, there is a mixing of the nearby $0^{th}$ HE$_{11}$ longitudinal resonance due to the reduced $Q$ in presence of the body (bandwidth of the modes overlap). This should not negatively affect the performance in MRI excitation. In fact, a slightly lower $\%\sigma_{TX}$ is obtained for the MM liner vs. BC with the body included, but a slightly lower $\eta_{TX}$ is obtained, suggesting that the MM liner is more greatly affected by the loss in the capacitors. The overall large $\%\sigma_{TX}$ in the presence of the body is unavoidable with a single circularly polarized excitation but may be improved with additional excitation ports (maintaining the same CP excitation) or by using “beamforming” methods such as RF shimming, which may be implemented by placing input ports at different locations or combining the liner with resonators that have current distributions orthogonal to those of the MM liner.

The most important difference between the performance of the BC and MM liner for MRI is in the $1/\eta_{SEE}$ shown in Fig. 12. The maximum $1/\eta_{SEE}$ is reduced by approximately 29.6% (3.1 $\sqrt{\text{W/kg}/\mu T}$ vs. 4.4 $\sqrt{\text{W/kg}/\mu T}$), which is a remarkable advantage of the MM liner and in close agreement with our previous MM liner design simulation [28]. This is due to the distributed currents on the MM rings, as opposed to the currents concentrated on two end rings and longitudinal rungs of the BC, resulting in more homogeneously distributed E-fields. Thus, in the MM the SAR is redistributed towards the neck and legs and reduced at the periphery of the torso and arms. In high ($>1.5$ T) ultra-high ($>4.7$ T) field MRI the application of many rapid imaging sequences is
FIGURE 12. Maximum intensity projections of the $1/\eta_{\text{SEE}} (\sqrt{W/\text{kg}/\mu T})$ with human body model for the (a) MM liner and (b) BC designs (4.7T-200MHz). The maximum is displayed above the axial slice.

greatly limited by local SAR constraints [1], [35]. Thus, the significantly reduced local SAR of the MM liner design would improve the temporal imaging efficiency (i.e., allowing faster image acquisition).

V. DISCUSSION AND CONCLUSION

In this study, we have developed a robust methodology for designing and evaluating the MM liner as a whole-body transmit MRI coil. This follows the introduction of the theoretical underpinnings of the MM liner in the Part 1 companion paper [2]. Our main findings are as follows.

A. METAMATERIAL LINER VARIANTS

This study presents the first attempt of deriving a tuning method for the novel MM liner resonator structure. We have recently described the experimental validation of the MM liner for MRI with a head coil sized version [24], but until this study we did not describe the network model method we employed for its design. The prototype head coil differs from the whole-body design presented here in three ways: 1) the dimensions were smaller, 2) there were no longitudinal traces connecting the MM liner rings, and 3) each adjacent ring was tuned differently to enable the $\lambda/2$ longitudinal resonance of the HE$_{11}$ mode at both the $^{23}$Na-53MHz and $^1$H-200MHz imaging frequencies. The head coil MM liner can be evaluated and designed using the network model developed here, where, instead of using the same geometric impedance parameters for both rings, alternating lumped radial and azimuthal elements were employed. Since the head coil did not include longitudinal traces the ring thickness was larger ($t_r = 30\text{mm}$) and ring spacing shorter ($\Delta z = 20\text{mm}$), so that the mutual inductance between rings was the dominant mechanism of propagation along the longitudinal direction ($M_{\phi\phi}^{\text{HE}_{11}} \approx -0.125$ for the head coil vs. $M_{\phi\phi}^{\text{HE}_{11}} \approx -0.01$ for the whole-body version here). The spacing and presence of the HE$_{11}$ modes and the relative spacing between longitudinal resonances was predicted to the same level of accuracy as for the whole-body coil presented here and was also verified by measurement. The accuracy of simulation of the $B_1$ field distribution was also verified by imaging. The network model and the method of analysis described here will enable future experimental implementations and modifications with dimensions suited to any application. For example: using non-uniform ring spacing, non-uniform tuning on rings, combination and RF shimming with other transmission elements, coupling with additional resonators or the use of different numbers of azimuthal segments per ring.

B. TUNING METHOD AND DERIVATION

The description of the MM liner introduced here is different than the algebraic methods or transmission-line eigenmode descriptions employed previously for BCs [36], [37], [38], [39]. Exact relations for the spacing between longitudinal resonances are not obtained. Instead, a variation of the lumped element tuning parameters is used, and the mode and longitudinal resonances are identified from the calculated dispersion or currents. Ideally, the tuning values would be derived via analytical algebraic relations like the methods for the BC, but it is not clear at this time if the same methods used for the BC could be applied with the quasi-3D nature of the MM liner mesh structure and $N_z \times N_{\phi} \times 3$ impedance matrix ($Z$), vs the 1D periodic structure of the BC. To achieve this, it may be necessary to include further simplifications of the model, such as with the BC where it is often assumed the coil is symmetric and using purely reactive tuning impedances can achieve the ideal current distribution [36], [37].

C. ALTERNATIVE BC SAR REDUCTION STRATEGIES

The key benefit of the MM liner is a 29.6% decrease in the maximum $1/\eta_{\text{SEE}}$, with homogeneity and transmit efficiency similar to that of the conventional birdcage coil presented here. Other alternative coil designs have shown improved $\eta_{\text{SEE}}$, such as the tic-tac-toe coil (24%-35% improvement compared to a TEM coil [12]) or the microstrip coil array (24% improvement compared to a BC [6]). It is not possible to directly compare the results of those previous studies, or others, with those presented here for the MM liner and BC since those designs were for head imaging, as well as other coil geometry and/or bore dimensions and field strengths.

There are design strategies to optimize the $10g$ averaged $\eta_{\text{SEE}}$ and $\eta_{\text{Tx}}$ of the BC and they are not always followed. One of these strategies is the choice of the capacitor topology. Here, a hybrid topology was used, rather than a LP (where capacitors are placed only on the legs) or a HP (where capacitors are placed only on the end-rings). For a shielded body coil design at 4.7T-200MHz the hybrid topology is
necessary, since otherwise the inductive reactance of the segments would become large relative to the stray capacitance between the conductor and the shield, and the desired uniform, circularly polarized mode could not be excited [30]. By placing distributed series capacitors on both end-ring and leg conductor segments this mode can be excited despite the large dimensions of the coil. In addition, the hybrid topology has a more balanced field/voltage distribution across its structure leading to greater insensitivity to loading and minimizing large conservative electric fields that would increase the localized SAR [30], [31].

Another strategy is the optimization of the BC diameter to length ratio to improve the performance as demonstrated in Ref. [31]. However, Ref. [31] was not an exhaustive comparison of all the possible diameters, lengths, and frequencies, so it was not established if the optimal ratio holds for every specific diameter (550mm was not included), field strength (4.7T was not included), or coil positioning. Here we have chosen the BC length and diameter to fit the imaging field of view and diameter to match that of the available 4.7T system for future experimental validation, and the MTM liner was designed to match the same longitudinal FOV to make a fair comparison.

Another method of improving the transmission metrics of the BC coil is altering the geometry, e.g., by using an oval or asymmetric cross section that more closely matches the shape of the human body [40], or a design where the ends taper to a small diameter end-rings at one end [41]. For body coils these strategies are not possible to employ, as the design must come as close to the cylindrical bore geometry as possible to maximize the available space. It is also possible to increase the number of rungs to provide a more homogeneous current distribution and potentially improve the maximum \( \eta_{\text{SEE}} \). for body coil designs 16–24 rungs is normal and to the best of our knowledge there have not been studies on whether an increased number would provide any SAR benefit.

### D. ULTIMATE SAR

Recent methods derive the current patterns on a surface to approach the ultimate obtainable \( \eta_{\text{SEE}} \) and \( \%\sigma \eta_{\text{Tx}} \) [42], [43]. Translating the surface currents into a feasible resonator or transmit array design [11], [44] provides some insight into the potential ideal resonator design to minimize SAR. In general, for low-fields (<3T) and homogeneous phantoms the ideal BC circularly polarized field distribution is a close approximation of the optimum. However, at higher field strengths, and with asymmetric and heterogeneous imaging objects like the human body, an optimal current pattern becomes difficult to determine, and will differ depending on the target imaging region and individual in the population.

It should be noted that comparisons of SAR for MRI are done almost exclusively as simulation studies, since the ability to measure SAR distribution accurately, and build and test many different design variations, is largely unworkable. Thus, the comparisons in the literature are all numerical and analytical, which may neglect real world complications. For example, the issue of unbalanced voltages and large conservative electric fields with the different capacitor topologies for BCs is not well represented in numerical and analytical methods that calculate SAR based on the expected current distributions, rather than the finite element HFSS solver which includes the field distribution caused by large, concentrated voltages with lumped impedance elements.

### E. RESONATOR INDEPENDENT METHODS OF IMPROVING TRANSMIT PERFORMANCE

The use of independent phase and amplitude control of the two quadrature modes is one of several strategies that could be employed for both the MTM liner and the BC to optimize the \( \eta_{\text{SEE}} \) and \( \eta_{\text{Tx}} \). One study [45] has shown for an imaging field view targeting the spine, that a whole-body BC with independent phase and amplitude control produced the highest \( \eta_{\text{SEE}} \) in a body model when compared to different transmission arrays (up to a 24 element stripline array). Another study [46] showed that this technique could on average improve the \( \%\sigma \eta_{\text{Tx}} \) by 32% when targeting different imaging regions in different body models, while maintaining improved \( \eta_{\text{SEE}} \) compared to typical quadrature excitation.

High permittivity [47] pads can also be used to improve the three transmit metrics for both the MM liner and BC coil. More recently, resonant array structures, magnetic lenses, or metasurfaces have been employed in a manner similar to dielectric pads to improve the transmit metrics by an equivalent or greater extent, while being lighter and tunable [48]–[52]. Similarly, we have demonstrated that the same 2D transmission line MM structure used to construct the MM liner presented here can be used to construct a thin (2 cm) MM slab that when placed on the body improves all three transmit metrics (locally) at 3T [53]–[55]. The close match between the simulated and measured network parameters of this structure has also been verified [56].

A different strategy to improve the transmit performance of volume resonators is a structure that emulates a perfect magnetic conductor (termed a magnetic shield), which has shown some promise in improving the \( \eta_{\text{SEE}} \) (by 27% with tissues close to the BC [57]). However, this was at the expense of \( \eta_{\text{Tx}} \) and the structure takes up additional space in the bore. The magnetic shield may provide a particular improvement with the MM liner in which equal and opposite currents run along the shield and inner conductor. The magnetic shield attempts to alleviate the resulting field cancellation effects of these opposite currents.

Considering the promising improved \( \eta_{\text{SEE}} \) result of the MM liner in this simulation study, as well as for our previous head coil sized prototype [24], experimental comparison of a whole-body BC coil to the MM liner is a goal for future study.

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