An Equivalence Principle-Based Hybrid Method for Propagation Modeling in Radio Environments With Reconfigurable Intelligent Surfaces

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Abstract

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Index Terms—Electromagnetic propagation, numerical analysis, ray-tracing, reconfigurable intelligent surfaces, wireless communication.

I. INTRODUCTION

RECONFIGURABLE intelligent surfaces (RISs) can dynamically redirect electromagnetic (EM) waves in a radio environment to selected directions, improving the received signal strength (RSS) in non-line-of-sight (NLoS) environments [1]–[3]. Assisted by RISs, the performance of wireless communication networks can be improved in a controllable, programmable, and cost-effective manner [4], [5]. Thus, RISs are considered a promising technology for current, emerging, and future generations of communication systems [6], [7].

To characterize RIS-enabled links, several free-space path loss models that treat RISs as arrays of independent scattering elements have been proposed [8]–[11]. These models determine the scattered field of a single unit cell first and apply field superposition to compute the total scattered field of the RIS. By employing a phase difference between unit cells, the scattered fields of the RIS at different operation states are determined. However, due to the sub-wavelength spacing of the unit cells, there is strong mutual coupling between them, resulting in significant discrepancies between calculated and actual scattered electric fields [12], [13]. Consequently, these models have limited accuracy to compute the magnitude of the electric field and the path loss, even in free-space RIS channels [14]. Moreover, the performance of an RIS depends on the environment where it is deployed. Fig. 1 shows an indoor office geometry, where a transmitter (Tx) directly illuminates an RIS, to enhance the RSS at an NLoS receiver (Rx). In addition to the direct ray from the transmitter to the RIS, the transmitter also radiates fields in other directions, which interact with the environment, as illustrated by the red dotted rays in Fig. 1. The associated reflected, transmitted, and diffracted fields may impinge upon the RIS with non-negligible field levels compared to the direct wave, contributing to the scattered fields of the RIS. In fact, many wireless communication scenarios occur in highly reverberant environments, such as dense urban areas and indoor geometries (e.g., the office shown in Fig. 1) [15]. In these cases, rich scattering may cause the departure of an RIS from its intended behavior [16]. This can be examined by considering multiple incident waves on the RIS further produce multipath effects, as shown by the scattered fields.
of an RIS the blue dotted rays in Fig. 1.

Full-wave methods can model the above-mentioned effects that can potentially influence RIS performance. A Method of Moments (MoM) solution was presented in [17] to compute the path loss of a single-input-single-output (SISO) RIS-enabled link in free-space. Yet, full-wave analysis of 3-D geometries requires excessive computational resources. This motivated us to use ray-tracing to model wave propagation of RIS channels in [18], [19]. We computed the complex radar cross section (CRCS) of the RIS by full-wave analysis, thus determining the scattered fields in an indoor geometry. These fields were ray-traced and added, at intended receiver points, to the fields that existed in the environment without the RIS. Despite accounting for the mutual coupling between unit cells, [18], [19] did not include multipath components in the incident field on the RIS. Also, this method cannot produce accurate results near the RIS, since it relies on the CRCS, a far-field quantity.

Overall, existing free-space path loss models cannot be used to evaluate the site-specific performance of an RIS, as they inherently suffer from low accuracy, mainly because they disregard mutual coupling between unit cells. The method in [18], [19] ignores near-field scattering and only included a single incident ray when deriving site-specific propagation models of RIS channels. As a result, it is not suitable for many near-field applications of RISs [20]–[22] or cases where multiple waves illuminate the RIS. Full-wave methods offer high accuracy but their computational cost is prohibitive for realistic radio environments. This work aims to provide a new approach that overcomes these limitations, enabling accurate analysis in rich scattering environments, at any distance from the RIS. To this end, we propose a hybrid ray-tracing/full-wave method based on the equivalence principle, to model wave propagation in radio environments with RISs. The proposed method enables the analysis of the near- and far-field performance of an RIS in realistic environments, with accuracy that is comparable to full-wave analysis. We further validate our method by comparisons to measurement data for an indoor channel with an anomalous reflection metasurface, realizing a single operation state of an RIS. This is the first, experimentally validated method, that accurately and efficiently models wave propagation in realistic, 3-D radio environments with RISs, while considering near-field scattering.

The rest of this article is organized as follows: Section II introduces the modeling of an RIS via equivalent sources. Section III explains how these are combined with ray-tracing to derive a site-specific propagation model. Section IV presents two case studies to validate the proposed method by comparing simulation results with HFSS finite-element analysis (FEA). Experimental validation is presented in Section V. Section VI summarizes our contributions.

II. RIS MODELING IN RAY-TRACING VIA EQUIVALENT SOURCES

We focus on a scenario with a single transmitter and a single RIS, modeled as a secondary source in the radio environment of interest. We launch rays from the transmitter and the RIS and apply the shooting and bouncing rays (SBR) method with a ray path correction process based on the image method (IM) [23]–[25], to determine exact ray paths that exist in the environment. This ray path determination algorithm is explained in Section III. The electric fields associated with the rays from the transmitter are determined by the conventional field computation algorithm [25], with the known radiation pattern of the transmitter. This section details the determination of the scattered fields from the RIS, with the prerequisite that the incident rays on the RIS and all the ray paths from the RIS to the receiver points of interest have been beforehand determined. As explained in later sections, we determine these paths by ray-tracing.

A. Equivalent Surface Current Sources

We apply the Huygens’ principle [26] to find equivalent surface current sources that generate the scattered field of the RIS for incident fields \( \mathbf{E}_{\text{inc}} \) and \( \mathbf{H}_{\text{inc}} \). These equivalent sources will be integrated with ray-tracing to compute the scattered electric field of the RIS across the geometry. With this objective, we select a bounding surface (represented by the dashed line in Figs. 2(a) and (b)) that is at an infinitesimal distance above the RIS, considered at \( z = 0 \). The electric and magnetic fields below the bounding surface at \( z < 0 \) are outside the region of interest, and are set to zero (see Fig. 2(b)) [26]. According to the equivalence principle, we introduce equivalent electric and magnetic current densities on the bounding surface that satisfy the boundary conditions [27]:

\[
\mathbf{J}_s = \hat{a}_z \times \mathbf{H}_{\text{tot}},
\mathbf{M}_s = -\hat{a}_z \times \mathbf{E}_{\text{tot}},
\]

(1)

where \( \mathbf{E}_{\text{tot}} \) and \( \mathbf{H}_{\text{tot}} \) are the total electric and magnetic fields on the surface and can be computed using full-wave analysis with known incident waves. The total fields can be expressed as:

\[
\mathbf{H}_{\text{tot}} = \mathbf{H}_{\text{inc}} + \mathbf{H}_{\text{sc}},
\mathbf{E}_{\text{tot}} = \mathbf{E}_{\text{inc}} + \mathbf{E}_{\text{sc}},
\]

(2)
where \( \mathbf{E}_{\text{inc}} \) and \( \mathbf{H}_{\text{inc}} \) are the incident electric and magnetic fields, \( \mathbf{E}_{\text{sc}} \) and \( \mathbf{H}_{\text{sc}} \) are the scattered electric and magnetic fields.

We restrict the computation to the portion of the bounding surface that is over the RIS and coincides in size with the RIS (designated as surface \( S \) in Fig. 2(c)). As a result, the RIS operating in the state of interest is replaced by the surface \( S \) with current densities \( \mathbf{J}_s \) and \( \mathbf{M}_s \). In this work, we employ the Ansys HFSS finite-element solver with perfectly matched layer (PML) boundary conditions to simulate RISs and compute \( \mathbf{E}_{\text{tot}} \) and \( \mathbf{H}_{\text{tot}} \) as shown in Fig. 2(d).

To integrate these equivalent sources on \( S \) with ray-tracing, we need to compute the scattered electric field due to \( \mathbf{J}_s \) and \( \mathbf{M}_s \). This topic is addressed next.

**B. Near and Far Scattered Electric Fields of the Equivalent Sources**

By inspection of (1), both \( \mathbf{J}_s \) and \( \mathbf{M}_s \) have only \( x \) and \( y \) components (in the local coordinate system of the RIS):

\[
\mathbf{J}_s = \hat{a}_x J_x + \hat{a}_y J_y, \\
\mathbf{M}_s = \hat{a}_x M_x + \hat{a}_y M_y.
\]

Fig. 3 shows the surface \( S \), with a differential element \( ds' \) displayed. The \( x, y, \) and \( z \) components of the scattered electric field due to \( \mathbf{J}_s \) at the observation point \( (x, y, z) \) are [26]:

\[
E_{Ax} = -\frac{jk_0}{4\pi k_0} \int_S \left( G_1 J_x + (x-x') G_2 \right) e^{-jk_0 R ds'},
\]

\[
E_{Ay} = -\frac{jk_0}{4\pi k_0} \int_S \left( G_1 J_y + (y-y') G_2 \right) e^{-jk_0 R ds'},
\]

\[
E_{Az} = -\frac{jk_0}{4\pi k_0} \int_S \left( G_1 J_z + (z-z') G_2 \right) e^{-jk_0 R ds'},
\]

where \( \eta_0 \) is the impedance of free-space, \( k_0 \) is the wavenumber in free-space; \( x', y', \) and \( z' \) denote the coordinates of the differential element, \( R \) is the distance from each differential element to the observation point, \( ds' = dx'dy' \), and

\[
G_1 = -\frac{1 - jk_0 R + k_0^2 R^2}{R^3}, \\
G_2 = \frac{3 + j3k_0 R - k_0^2 R^2}{R^5}.
\]

The three components of the electric field due to \( \mathbf{M}_s \) are:

\[
E_{Fx} = -\frac{1}{4\pi} \int_S \left[ (z-z') M_y - (y-y') M_z \right] e^{-jk_0 R ds'},
\]

\[
E_{Fy} = -\frac{1}{4\pi} \int_S \left[ (x-x') M_z - (z-z') M_x \right] e^{-jk_0 R ds'},
\]

\[
E_{Fz} = -\frac{1}{4\pi} \int_S \left[ (y-y') M_x - (x-x') M_y \right] e^{-jk_0 R ds'}.
\]

Then, the total scattered electric field is:

\[
\mathbf{E}_{\text{sc}} = \hat{a}_x (E_{Ax} + E_{Fy}) + \hat{a}_y (E_{Ay} + E_{Fy}) + \hat{a}_z (E_{Az} + E_{Fz}).
\]

In its far-field region, the scattered electric field of \( S \) in its own local spherical coordinate system can be approximated as [26]:

\[
\mathbf{E}_{\text{sc}} = \hat{a}_\theta E_\theta + \hat{a}_\phi E_\phi
\]

\[
= -\hat{a}_\phi \frac{j k_0 e^{-j k_0 r'}}{4\pi r} \left( L_\theta + \eta N_\theta \right) + \hat{a}_\phi \frac{j k_0 e^{-j k_0 r}}{4\pi r} \left( L_\theta - \eta N_\phi \right),
\]

\[\text{with } \ r' \cos \psi = x' \sin \theta \cos \phi + y' \sin \theta \sin \phi.\]

The analytical expression of the scattered electric field (7) and the far-field approximation form (8) are utilized for the field computation process in ray-tracing, which is introduced in the following subsection. It is worth mentioning that (7) and (8) are computed numerically, with discrete, equal-size differential area elements. The total electric and magnetic fields at the differential elements are obtained by full-wave analysis. Over each element, the fields are considered constant; they are computed at the center of the element by full-wave analysis of the RIS scattering problem (Fig. 2(d)). Note that the size of the differential elements affects the accuracy of the computed scattered electric field. Therefore, we determine their size by studying the convergence of the computed scattered electric field, as we reduce \( ds' \).

**C. Embedding Equivalent Sources in Ray-Tracking**

To compute the scattered electric field due to the equivalent sources in a given environment with ray-tracing, the ray paths from the RIS that reach the receiver points are determined first.
In this work, only single diffraction is considered, assuming that multiple diffractions are negligible [28].

Let us consider a typical indoor RIS channel as shown in Fig. 4, where the role of the RIS is to redirect the direct wave from the transmitter to enhance the signal strength in room 2. As previously mentioned, the RIS is treated as a secondary source in the environment, and thus the computation of ray paths from the RIS is done in the same way as for a primary transmitter. We consider two receivers of interest: Rx-1 and Rx-2, which are respectively located in the far-field and near-field regions of the RIS. With the ray path determination algorithm that will be explained in the next section, rays that originate from the RIS and reach the two receiver points are determined. For simplicity, we display a selection of determined rays in Fig. 4, namely the direct ray $r_{11}^d$, the single-reflected ray $r_{12}$, and the diffracted ray $r_{13}$ that reach Rx-1, as well as the direct ray $r_{21}^d$, the single-reflected ray $r_{22}$, and the diffracted ray $r_{23}$ that reach Rx-2. The subscript $s$ denotes that the corresponding ray is directly scattered from the RIS. The rays $r_{12}$, $r_{13}$, $r_{22}$, $r_{23}$ are respectively linked to the rays $r_{11}^d$, $r_{13}$, $r_{22}^d$, $r_{23}^d$, as shown in the figure. The next task is to compute the electric field at the receiver points contributed by all rays. This is done by associating these fields with the scattered electric field of the RIS in (7) and (8).

In the near-field region of the RIS, the electric field of the direct ray from the RIS to a receiver point (e.g., the ray $r_{23}^d$) is computed using (7). The fields associated with non-direct rays (e.g., the rays $r_{22}^d$ and $r_{23}^d$) are computed with the same field computation algorithm as the receiver points in the far-field, which is introduced in the following.

The far-zone scattered electric field at a distance $d_{0}$, $E_{sc}^s (r = d_{0}, \theta, \phi)$, where $(r, \theta, \phi)$ are the spherical coordinates of the RIS, can be computed by (8). Then, the scattered electric field of the rays from the RIS at $(r, \theta, \phi)$ can be calculated using [25]

$$E_{sc}^s (r, \theta, \phi) = E_{sc}^s (r = d_{0}, \theta, \phi) d_{0} e^{jk_{d0}d_{0} \frac{e^{-jk_{d0}d_{0}}}{r}}. \quad (11)$$

With (11), the electric field of a direct ray that reaches a far receiver point (e.g., the ray $r_{11}^d$) can be directly computed. Note that the term $E_{sc}^s (r = d_{0}, \theta, \phi) d_{0} e^{jk_{d0}d_{0} \frac{e^{-jk_{d0}d_{0}}}{r}}$ describes the far-zone scattering characteristics of the RIS, which is combined with the reflection coefficients, the transmission coefficients, and the uniform geometric theory of diffraction (UTD) coefficients [25], [29], [30], to further compute the electric fields of non-direct rays. For example, the electric field of the ray $r_{12}$ is [25]:

$$E_{12} = \Gamma E_{sc} (r = d_{0}, \theta_{12}, \phi_{12}) d_{0} e^{jk_{d0}d_{0} \frac{e^{-jk_{d0}d_{0}}}{d_{12}}}. \quad (12)$$

In (12), $\Gamma$ is a diagonal $2 \times 2$ matrix consisting of TE and TM reflection coefficients in the facet-fixed coordinate system of the intersection facet (where the reflected ray is generated), $\theta_{12}$ and $\phi_{12}$ are angles in the local spherical coordinate system of the RIS for the ray $r_{12}^d$, and $d_{12}$ is the unfolded ray length from the RIS to Rx-1 (i.e., the sum of the ray lengths of $r_{12}^d$ and $r_{12}$). Note that $E_{sc}^s$ is expressed as a $2 \times 1$ matrix that contains the TE and TM components.

### III. SITE-SPECIFIC PROPAGATION MODEL FOR RIS CHANNELS

In this section, we present our end-to-end modeling method. We first introduce how the ray paths existing in the environment of interest are determined by combining the SBR method with the image method-based ray path correction process. Then, we discuss how ray-tracing is used to determine the incident waves on an RIS. Finally, we summarize our methodology in four steps.

#### A. Determination of Ray Paths

We derive the exact ray paths from a source point to a receiver point using the standard SBR method and a ray path correction algorithm [24], [25]. Fig. 5 shows selected ray paths from the transmitting existing in a geometry such as that of Fig. 4. Note that the Rx in Fig. 5 is placed at the location of the RIS in Fig. 4. This is done to determine the incident rays at the RIS, as explained next. In this work, we apply the widely used (nearly) uniform ray launching method by tessellating an icosahedron [24]. Depending on the angular separation of the launched rays and the ray lengths, reception spheres are used to capture rays that contribute to the total field at a receiver point. Then, the image method is applied to correct the ray paths. Fig. 5 displays how the ray $r_{3}^d$ is determined. The red and green dashed ray paths come from two adjacent launched rays (i.e., the red and green dotted rays), while only the red one is captured by the reception sphere. Then, the image source of the transmitter with respect to the intersection facet is determined, and exact ray paths and their lengths from the transmitter to the receiver are obtained (the black dotted and solid ones).

The algorithm involves correcting the ray paths computed by SBR. Therefore, the angular separation of the launched rays can affect the accuracy of the results. A large angular separation can result in some ray paths being lost during the SBR process, while a smaller angular separation can lead to increased computational cost due to more launched rays. Therefore, a convergence test of the results is applied to determine the optimal number of launched rays.

#### B. Incident Field at the RIS

In the channel geometry shown in Fig. 4, the RIS has an incident wave that directly comes from the transmitter. In reality,
the transmitter also radiates waves in other directions. These waves interact with the environment and may impinge upon the RIS. The field levels of these incident waves depend on both the transmitter’s radiation characteristics and placement, as well as the specific environment.

To determine incident waves on an RIS, we place a virtual receiver at the RIS. Then, all incident waves (including their directions, polarization, and field levels) are determined by ray-tracing the transmitter, as shown in Fig. 5. The incident field at an RIS is considered to be composed of multipath components determined by ray-tracing. For simplicity, only four incident rays (the direct ray $r_1$, two single-reflected rays $r_2$ and $r_3$, and a diffracted ray $r_4$) at the RIS are displayed in Fig. 5. All incident waves determined by ray-tracing can be included in the full-wave simulation round as wave excitations to compute $E_{\text{tot}}$ and $H_{\text{tot}}$. Then, the response of the RIS to these waves is modeled by the obtained equivalent surface sources.

The number of incident waves included in the computation of the equivalent sources can be reduced by disregarding rays of negligible amplitude. For example, let the ray-tracing of the transmitter return a total of $Q$ incident rays. The $q$-th ray is included in the full-wave analysis of RIS scattering as an incident wave if its magnitude at the RIS, $|E_{\text{inc}}^q|$ meets the condition:

$$\max_{q=1,2,...,Q} |E_{\text{inc}}^q| [\text{dBV/m}] - |E_{\text{inc}}^q| [\text{dBV/m}] \leq \epsilon,$$  \hspace{1cm} (13)

where $\epsilon$ is a set threshold.

C. Summary

The modeling steps are summarized as follows:

1) We first use ray-tracing to determine the incident waves on the RIS. In this round, a receiver point is placed at the RIS to compute all rays from the transmitter that reach the RIS. The number of rays can be reduced if needed, by applying (13).

2) We set the determined incident waves as excitations in the full-wave simulation to compute $E_{\text{tot}}$ and $H_{\text{tot}}$ of the RIS operating in a specific state.

3) We ray-trace the transmitter at first. The rays that impinge upon the RIS do not generate any reflected rays. Instead, new rays are launched and traced from the RIS, as a secondary source. This step returns $M$ rays from the transmitter and $N$ rays from the RIS that reach the receiver point of interest. Note that this receiver point is different than that used in step 1.

4) At the receiver point, we compute the fields associated with rays from the transmitter with the standard field computation algorithm. The fields of the rays launched from the RIS are computed by the field computation algorithm presented in Section II-C. Finally, we compute the total electric field at the receiver point as:

$$E_{\text{Rx}} = \sum_{m=1}^{M} E_{\text{Tx}}^m + \sum_{n=1}^{N} E_{\text{RIS}}^n,$$  \hspace{1cm} (14)

where $E_{\text{Tx}}^m$ and $E_{\text{RIS}}^n$ are the electric fields of the $m$-th ray from the transmitter and the $n$-th ray from the RIS, respectively. Note that the fields are expressed and added in the global coordinate system.

This process returns the electric field at one receiver point. Steps 3 and 4 can be repeated to compute the electric field at multiple receiver points.

IV. COMPARISON WITH FINITE-ELEMENT ANALYSIS

In this section, we validate the proposed method by comparing its results with HFSS finite-element analysis. Restricted by the intensive computational cost of finite-element analysis, we implemented two simple geometries that were manageable in HFSS and contained NLoS points. We investigated one scenario where the direct wave from the transmitter to the RIS dominated the incident field on the RIS and another scenario where the direct field impinging on the RIS along with multipath components of comparable strength. We used the unit
cell structure (Fig. 6) proposed in [31] to construct a $16 \times 16$ RIS with dimensions of $304 \times 304$ mm$^2$ ($5.87\lambda \times 5.87\lambda$, where $\lambda$ is the wavelength) that operated at 5.8 GHz. The unit cell features switch-ON and switch-OFF states, realizing $180^\circ$ phase variation of the reflected fields. All the data presented in this section were collected at 5.8 GHz. We studied a domain that included the near-field and far-field regions of the RIS. In each case study, we applied near-field computation in the region $r < 2$ m, and the value of the parameter $d_k$ in (11) for field computation was set to 10 m. With the test of the convergence of the modeling results, the dimensions of the differential elements were selected as $4 \text{ mm} \times 4 \text{ mm}$ ($\lambda/12.9 \times \lambda/12.9$). Also, the number of launched rays at both the transmitter and the RIS was 26,012 (achieving an angular separation between successive rays of approximately $1.52^\circ$).

A. Case Study 1: Direct Wave Dominates the Incident Field

We first study a simple RIS channel as shown in Fig. 7. A horizontally polarized rectangular horn antenna with a maximum gain of 18.04 dBi and transmit power of 30 dBm directly illuminated an RIS. The transmitter was centered at $(0 \text{ m}, 3 \text{ m}, 0.2 \text{ m})$, which was also selected as the phase reference point (i.e., the phase of the radiated field of the transmitter at this point was set to 0, for the determination of fields everywhere else). The RIS was mounted on the center of a wall with a relative permittivity $\varepsilon_r = 2.5$, a dielectric loss tangent $\tan \delta = 0.15$, and dimensions of $2 \times 0.4 \times 0.01$ m$^3$. A copper slab with conductivity $\sigma = 5.8 \times 10^7$ S/m and dimensions $1.5 \times 0.2 \times 0.01$ m$^3$ was placed at $(3 \text{ m}, 3.25 \text{ m}, 0.2 \text{ m})$ as an obstacle. The RIS was set to operate at a state (see Fig. 7) that could redirect an incident wave with an angle of $(\theta = 45^\circ, \phi = 180^\circ)$ to an angle of $(\theta = 22^\circ, \phi = 0^\circ)$. All angles are with respect to the local coordinate system of the RIS in Fig. 6. Hence, the RIS redirected the incident wave from the transmitter to the area that was obstructed by the copper slab. Due to the high memory requirements of finite-element analysis, its simulation domain was restricted to the region $2 \text{ m} \leq x \leq 5 \text{ m}$ and $0 \text{ m} \leq y \leq 4 \text{ m}$ on the $z = 0.2 \text{ m}$ plane. We uniformly placed 30,351 receiver points in this region for the ray-tracing simulation.

Due to the simplicity of the scenario, both the maximum numbers of reflections and transmissions allowed per ray in ray-tracing were set to 2 (in fact, only single-reflected and single-transmitted rays could exist in this geometry). Three incident rays were determined by ray-tracing as displayed in Fig. 7: the direct ray from the transmitter with a field level of $14.64$ V/m, two diffracted rays with field levels of $9.22 \times 10^{-5}$ V/m and $2.60 \times 10^{-5}$ V/m, respectively. These three rays were implemented as excitations in the full-wave simulation round. The interface between the near- and far-field computation domains ($r = 2$ m) is displayed in Fig. 7.

The electric field distribution obtained by the proposed method, shown in Fig. 7, illustrated that the RIS redirected the incident wave from the transmitter to the area blocked by the copper slab. This resulted in enhanced electric field levels at NLoS points (i.e., those in the shadow region of the copper slab). Fig. 8 compares the electric fields along several lines within the geometry, obtained by HFSS finite-element analysis, the proposed method, and the method presented in [19]. Note that only the direct wave was implemented as an excitation to model this scenario with the method of [19] because it was not able to account for multiple incident field components as discussed in Section I. The proposed method and the method of [19] produced very similar electric field levels in the far-field region of the RIS, as shown in Figs. 8(b) and (d). Since the direct wave from the transmitter was the dominant component of the incident field (as revealed by the field levels of the incident rays presented above), accurate far-field results were obtained by the method of [19]. Figs. 8(a) and (c) show that the proposed method provided significantly closer results to the finite-element analysis in the near-field region of the RIS compared to the method of [19].

![Fig. 7. The simple RIS channel used to validate the proposed method. The electric field obtained by the proposed method, the FEA simulation domain, the sampling points (indicated by red dotted lines) for comparison, the operation state of the RIS (yellow denotes switch-ON, blue denotes switch-OFF), and two incident rays with the highest and second highest field levels are displayed.](image-url)

![Fig. 8. Comparison of the results obtained by the three methods. (a) $x = 3$ m, (b) $x = 4$ m, (c) $y = 0.2$ m and (d) $y = 3$ m.](image-url)
Fig. 9. The RIS channel with multiple incident waves impinging upon the RIS used to validate the proposed method. The incident ray diagrams and the operation state of the RIS (yellow denotes switch-ON, blue denotes switch-OFF) are displayed.

Fig. 10. The electric field distribution obtained by the proposed method. The FEA simulation domain and the sampling points (indicated by red dotted lines) for comparison are displayed.

Fig. 11. Comparison of the results obtained by the two methods. (a) $x = 3 \text{ m}$, (b) $x = 4 \text{ m}$, (c) $y = 0.2 \text{ m}$ and (d) $y = 3 \text{ m}$.

**TABLE I**

| Electric Field Levels of Selected Incident Rays for Case Study 2 (Unit: V/m) |
|-------------------------------|-------|-------|-------|-------|-------|
| Ray $r_1$                  | $r_2$ | $r_3$ | $r_4$ | $r_5$ | $r_6$ |
| Ray $E$                     |       |       |       |       |       |
| 1.76                        | 1.57  | 0.39  | 0.39  | 0.097 |
| $r_6$                       |       |       |       |       |       |
| $E$                         | 0.094 | 0.0036| 0.0017| $2.37 \times 10^{-5}$| $1.19 \times 10^{-5}$ |

B. Case Study 2: Enhanced Multipath Propagation Scenario

The second scenario used for validation of our method is shown in Fig. 9. The transmitter was a finite length dipole, generating a richer set of multipath components in the geometry than that of Fig. 7. The dipole had a far electric field of $[32]$: 

$$E_{\text{dipole}} = \frac{a_0 j n_0 e^{-jk_0 r}}{2\pi r} \cos(k_0 l \cos \theta / 2) \cos (k_0 l / 2) \frac{\sin \theta}{\sin \theta},$$

(15)

where $I_0 = 1 \text{ A}$ and $l = 10 \text{ mm}$. The electric field is expressed in the local spherical coordinate system of the dipole (assuming that the dipole is along the $z$-axis), which is different from the global coordinate system defined in Fig. 9.

The dipole was placed parallel to the $xy$-plane at $(0 \text{ m}, 3 \text{ m}, 0.2 \text{ m})$. Two copper slabs with dimensions of $1.5 \times 0.2 \times 0.01 \text{ m}^3$ and $1.2 \times 0.2 \times 0.01 \text{ m}^3$ were placed at $(3 \text{ m}, 2.5 \text{ m}, 0.2 \text{ m})$ and $(0.6 \text{ m}, 4 \text{ m}, 0.2 \text{ m})$, respectively. We used the same RIS with the same location and orientation as in the geometry of Fig. 7. A different operation state was applied, as shown in Fig. 9. This state could redirect an incident wave with an angle of $\theta = 45^\circ$, $\phi = 180^\circ$ to an angle of $\theta = 29^\circ$, $\phi = 0^\circ$, in the local coordinate system of the RIS. This simple scenario was designed to examine the operation of an RIS in an enhanced multipath environment (as opposed to the first scenario where the direct ray from the transmitter to the RIS was dominant).

Both the maximum number of reflections and transmissions were set to 3. The first ray-tracing round returned 10 incident rays, namely the direct ray $r_1$, the single-reflected ray $r_2$, and the single-diffracted rays $r_3$, $r_4$, $r_5$, $r_6$, $r_7$, $r_8$, $r_9$, $r_{10}$, as displayed in Fig. 9. Particularly, $r_7$ and $r_8$ were reflected off copper slab 2 first and were respectively diffracted by different edges of copper slab 1, $r_5$ and $r_7$, $r_6$ and $r_8$, $r_9$ and $r_{10}$ respectively had the same ray paths. The electric field levels of these rays are listed in Table I. All these rays were implemented as excitations in the full-wave simulation round. The finite-element simulation domain and the number of receiver points in this region (for the simulation using the proposed method) were the same as in case study 1.

Then, we obtained the electric fields across the geometry, shown in Fig. 10. Fig. 11 compares the electric fields along several lines within the geometry, obtained by HFSS finite-element analysis, the proposed method, and the method presented in [19]. Only the direct wave was implemented as an excitation to model this scenario with the method of [19]. Fig. 11 shows that the proposed method achieved very close field level predictions to the HFSS finite-element analysis, as demonstrated by the comparison of the electric field levels.
The proposed method is orders of magnitude faster than full-wave analysis. Compared to the method of [19], it exhibits higher accuracy in the near-field region of the RIS, while consuming almost the same amount of time. Furthermore, it can be easily employed to model scenarios where an RIS is illuminated by multiple waves.

V. EXPERIMENTAL VALIDATION

In this section, we present modeling results of an actual indoor communication channel with an anomalous reflection metasurface, along with a measurement campaign as a means of validation of our method.

A. Channel Setup

In [33], we presented measurement data in an indoor radio environment with an anomalous reflection metasurface to validate the method presented in [18], [19]. The measurement data herein were collected using the same measurement setup as that in [33]. The geometry was a hallway junction located on the eighth floor of the Bahen building on the University of Toronto campus, as shown in Fig. 12. A global coordinate system is also defined and displayed in the same figure.

The metasurface consisted of $45 \times 33$ non-uniform unit cells with metallic rings printed on a full ground plane-backed substrate, enabling anomalous reflection [34], [35], as shown in Fig. 13. The coordinate system is the same as that in Fig. 12. The substrate was Rogers RO3010 with $\varepsilon_r = 10.2$, $\tan \delta = 0.0035$, and a thickness of $1.28 \text{ mm}$. The unit cells had dimensions of $12.93 \times 12.93 \text{ mm}^2 (\lambda/4 \times \lambda/4)$. The dimensions of the metallic rings were identical along the $z$-axis, but they varied along the $x$-axis, as listed in Table III (the unit cell numbers, from left to right, are indicated in Fig. 13). The metasurface was designed to redirect a vertically polarized (i.e., z-polarized) incident wave, with an incident angle of $26.6^\circ$ to the direction of $53^\circ$ at $5.8 \text{ GHz}$. Its dimensions were $609.6 \times 457.2 \text{ mm}^2 (11.8x \times 8.4\lambda)$. The role of the metasurface was to mimic an RIS operating at a specific state and to enable coverage within the shadow region of the hallway corner.

The transmitter, the metasurface, and the receiver were at the same height of $1.34 \text{ m}$. The transmitter was a vertically polarized rectangular horn antenna with a gain of $17.1 \text{ dBi}$ and was connected to a signal generator set to transmit power of 25 dBm. It was positioned at $(-2.5 \text{ m}, -6 \text{ m}, 1.34 \text{ m})$ and directly illuminated the metasurface, which was centered at $(0.5 \text{ m}, 0 \text{ m}, 1.34 \text{ m})$. The receiver was a vertically polarized omnidirectional whip antenna with a gain of $2 \text{ dBi}$ mounted on a tripod and connected to a spectrum analyzer. All facets including the walls, the floor, and the ceiling, were set as concrete with $\varepsilon_r = 5.31$ and $\tan \delta = 0.079$ in ray-tracing simulations. The RSS at multiple grid points (white points in Fig. 12) was measured. In this measurement, we focused on the far-field performance of the metasurface. To reduce the impact of fading, we took 12 RSS measurements at each point by rotating the receiver about the axis of the tripod. The radius of rotation was $2 \text{ cm}$, and the average RSS at each point was considered as the measured RSS at that point. We first calibrated the measurement setup by measuring the RSS versus distance along the main-lobe direction of the transmitter in the hallway, to evaluate polarization, impedance mismatch and other losses.

B. Numerical Modeling Results and Experimental Validation

The maximum number of reflections was set to 5 while transmitted rays were ignored in this environment. 222 incident rays were obtained by the first round of ray-tracing. In this case, we implemented the step discussed in Section III-B, to exclude rays of negligible magnitude, setting $e = 30 \text{ dB}$. The strongest component of the incident field corresponded to the direct ray from the transmitter, which was $4.65 \text{ V/m}$ ($13.35 \text{ dBV/m}$). The second highest field level among all the field components was only $0.11 \text{ V/m} (-19.17 \text{ dBV/m})$, contributed by the black dashed diffracted ray in Fig. 12. Based on (13), only one wave corresponding to the direct ray was included as an excitation in the full-wave simulation.

<table>
<thead>
<tr>
<th>Method</th>
<th>Case Study 1</th>
<th>Case Study 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>Proposed</td>
<td>28m 51s + 31s</td>
<td>30m 36s + 37s</td>
</tr>
<tr>
<td>Ref. [19]</td>
<td>29m 02s + 21s</td>
<td>27m 29s + 26s</td>
</tr>
<tr>
<td>HFSS FEA</td>
<td>27h 57m</td>
<td>26h 53m</td>
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</tbody>
</table>

*The time of the proposed method and the method of [19] is expressed as full-wave simulation time + ray-tracing time.*
Table III
RADIi OF THE METASURFACE UNIT CELLS ALONG THE x-AXIS

<table>
<thead>
<tr>
<th>cell #</th>
<th>radii</th>
<th>cell #</th>
<th>radii</th>
<th>cell #</th>
<th>radii</th>
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</thead>
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<tr>
<td>#1</td>
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<td>#2</td>
<td>3.60,2.70</td>
<td>#3</td>
<td>3.84,2.14</td>
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<tr>
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<td>4.00,1.90</td>
<td>#5</td>
<td>3.53,2.83</td>
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<td>4.20,1.60</td>
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<tr>
<td>#7</td>
<td>3.50,2.95</td>
<td>#8</td>
<td>3.50,3.00</td>
<td>#9</td>
<td>4.00,2.30</td>
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<tr>
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<td>#11</td>
<td>2.62,1.12</td>
<td>#12</td>
<td>3.50,2.47</td>
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<td>4.20,0.80</td>
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<td>3.84,2.14</td>
<td>#15</td>
<td>4.20,1.40</td>
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<td>#16</td>
<td>3.53,2.83</td>
<td>#17</td>
<td>3.47,2.98</td>
<td>#18</td>
<td>3.84,2.34</td>
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<td>4.20,2.00</td>
<td>#21</td>
<td>5.00,1.55</td>
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<tr>
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<td>#23</td>
<td>4.00,1.10</td>
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<td>#39</td>
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<td>#42</td>
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<tr>
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<td>#44</td>
<td>3.11,2.41</td>
<td>#45</td>
<td>3.83,1.82</td>
</tr>
</tbody>
</table>

round. The parameter $d_0$, the number of launched rays at the transmitter and the RIS, and the dimensions of the differential elements were also the same as in previous case studies. The near-field computation was applied in the region $r < 2$ m. The simulations were performed on the same desktop as previous case studies. 22,781 receiver points were uniformly distributed in the simulation domain to obtain the electric field distribution displayed in Fig. 12. The time required to execute the proposed method was 78m 49s (including 75m 14s of full-wave simulation time).

Fig. 12 shows the electric field distribution across the geometry obtained by our method, which demonstrates that the metasurface was successful in redirecting the incident wave from the transmitter to the NLoS locations. Fig. 14 compares the RSS obtained by the proposed method and measurements along three lines. Excellent agreement between the two sets of results can be observed. It is worth mentioning the simulated RSS at each point was an average of 10 uniformly distributed points on the same plane within a circular area with a radius of 2 cm, consistent with the measurement process.

VI. CONCLUSION
We presented a new hybrid ray-tracing/full-wave method based on the equivalence principle for rigorously modeling wave propagation in realistic RIS channels. Our method overcomes the limitations of existing methods for modeling propagation in RIS channels. For pure full-wave methods [17], the main limitation is the excessive computational cost to model realistic three-dimensional geometries. Our method reaches the accuracy of full-wave analysis at a fraction of the computational cost. Other methods [8]–[11] cannot be used to analyze the site-specific performance of RISs and disregard
mutual coupling between RIS unit cells, whereas we naturally account for it by analyzing the scattering of incident waves by the RIS in the full-wave portion of our approach. Finally, we obtain accurate field predictions in the near and far field region of the RIS, readily accounting for multiple incident waves, unlike [18], [19]. The accuracy of our method has been further verified by an indoor channel measurement.

As a result, this paper advances the state of the art in propagation modeling for RIS channels, contributing a novel methodology that is both accurate and efficient. Hence, it is a valuable tool for propagation studies in RIS channels, as well as the site-specific analysis, design, and optimization of RIS structures.

REFERENCES


