An Efficient FDTD Algorithm Based on the Equivalence Principle for Analyzing Onbody Antenna Performance

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Abstract—In this paper, onbody antenna performance and its effect on the radio channel is analyzed. An efficient numerical technique based on the finite-difference time-domain technique and the equivalence principle is developed. The proposed technique begins with the problem decomposition by separately computing the wearable antennas and onbody propagation involving the digital human phantom. The equivalence principle is used as an interface between the two computational domains. We apply this technique to analyze onbody antenna and channel characteristics for three different planar body-worn antennas operating at the industrial-scientific-medical frequency band of 2.4 GHz. Simulated results are validated with measurement data with good agreement.

Index Terms—Body-area networks, body-worn antenna, equivalence principle, finite-difference time domain (FDTD).

I. INTRODUCTION

WIRELESS body-area networks (WBAN) have recently received increasing attention due to their promising applications in medical sensors and personal entertainment systems [1]–[4]. Ubiquitous and wearable computing is regarded as the central component of fourth-generation communication systems [5]. In order to design a spectrum and power-efficient onbody communication system, it is very important to understand radio channel characteristics, which was obtained by using experiments in the past [6]–[12]. A computational model based on the finite-difference time domain (FDTD) [13] was developed, aiming to provide a physical insight of an onbody radio channel [14], [15]; however, a point-source approximation was used in the simulation and, hence, it is not a true representation of body-worn antennas.

The FDTD technique [13] is a method based on the direct solution of the Maxwell’s curl equations, and is ideal to simulate stratified dielectric objects, such as a human body. However, if practical antennas are considered in the onbody radio channel, much higher spatial resolution is needed to accurately encounter small geometrical features in the antenna. This, in turn, increases the computation time and the memory requirements if a uniform mesh scheme is used. Although the use of a nonuniform mesh and a subriding scheme in FDTD can be applied, it may result in spurious solutions or even suffer from instability for the subgriding scheme [16]. It is also possible to combine the FDTD and other numerical schemes, for example, the frequency-domain method of moments (FD-MoM) to increase numerical efficiency. In [17] and [18], the equivalence principle has been used to divide the original problem into two subproblems: the radiating element, ideally, a metallic structure is modeled using the MoM, while the surrounding environment uses the FDTD. In [20] and [21], a time-domain MoM (TD-MoM) is used to analyze the antenna, but it suffers from numerical instability when applied to model planar antenna structures.

In this paper, in addition to our previous work [22], a numerical technique based on the surface equivalence theorem (Huygen’s principle) is developed to characterize onbody antennas. The method begins with a division of the original problem into two subproblems as shown in Fig. 1.

The antenna is analyzed in free space by using CST Microwave Studio. Near fields on a closed surface surrounding the
antenna are recorded. The human body is modeled in FDTD by using the HUGO digital human phantom [23]. Near fields obtained from the CST simulation are used as an input in the form of electrical and magnetic surface currents to the FDTD code [22]. The developed technique is applied to evaluate the onbody propagation, and the comparison between simulated results and measurement data shows good agreement. Furthermore, the importance of applying practical body-worn antennas (instead of simplified sources) to characterize the onbody radio channel has been demonstrated, and the variation of the path loss with the distance between transmitter and receiver is evaluated. The rest of this paper is organized as follows. Section II introduces the equivalence principle applied to the FDTD and explains the features of the developed algorithm. Section III presents some numerical results and their comparison with measurement. Section IV draws a conclusion.

II. DESCRIPTION OF THE NUMERICAL METHOD

The numerical problem is divided into two parts: 1) the antenna is first analyzed in free space with CST Microwave Studio and 2) a virtual box surrounding the antenna is set (as shown in Fig. 2) where near fields are sampled at the steady state for the desired frequency. Different from [22], in which indoor propagation scenarios were considered, the radiation pattern calculated with CST Microwave Studio cannot be directly applied to onbody propagation, and the radiation pattern must be modified due to backscattering and energy absorption when the antenna is placed close to the human body. The distance $d$ (as shown in Fig. 2) represents the spacing between the antenna and the human body; hence, its value has to be small when the antenna is body worn, and it is dependent on the FDTD spatial discretization. Second, recorded field components are converted into surface currents, to be used as an excitation source for the FDTD code [24] to evaluate onbody antennas and radio channels.

A. FDTD Formulation With the Equivalence Principle

The FDTD is based on a direct solution to Maxwell curl equations where the differential operator is approximated with the central finite differences

$$\begin{align*}
\nabla \times \mathbf{E} &= -\mu \frac{\partial \mathbf{H}}{\partial t} - \mathbf{J} \\
\nabla \times \mathbf{H} &= \varepsilon \frac{\partial \mathbf{E}}{\partial t} + \sigma \mathbf{E} + \mathbf{J}
\end{align*}$$

(1)

where $\mathbf{E}$, $\mathbf{H}$ are electric and magnetic fields, $\mu$ is the permeability, $\varepsilon$ is the permittivity, $\mathbf{J}$, $\mathbf{M}$ are electric and magnetic current density, and $\sigma$ is the electric conductivity. The fields on the equivalent surface, calculated with CST Microwave studio, are converted in surface currents using the relations

$$\mathbf{J} = \mathbf{n} \times \mathbf{H}$$

(2)

$$\mathbf{M} = \mathbf{E} \times \mathbf{n}$$

(3)

where $\mathbf{n}$ is the normal vector to the surface. The electric surface current $\mathbf{J}$ is used to update the electric field, while the magnetic surface current $\mathbf{M}$ is used to update the magnetic field as explained in [24].

B. Total-Field/Scattered-Field Implementation

In the classic equivalence principle, for the exterior region, only far-field information is of interest and, hence, the equivalent surface currents are set only to radiate outwards of the enclosed surface, and the electric and magnetic fields inside the surface are always assumed to be zero. However, if the surface current source radiates in proximity of an object (e.g., the human body), the usual condition that sets zero fields inside the equivalent surface prevents the reflected field from entering the surface, hence causing inaccuracy in FDTD simulations. To alleviate such a problem, we apply the total-field/scattered-field method within two different FDTD domains as follows.

1) Domain A [see Fig. 3(a)] contains only the equivalent surface in the free space.

2) Domain B [see Fig. 3(b)] is where surface currents radiate in the presence of the human body.

For both cases, the box size and current density of the virtual source are the same. Inside one iteration, operations are implemented as follows.

1) Fields are updated in the Domain A. The equivalent currents ($\mathbf{J}$, $\mathbf{M}$) only radiate outwards from the surface, and electric and magnetic fields are forced to zero inside the equivalent surface [as shown in Fig. 3(a)] in order to calculate the incident field components ($\mathbf{E}_{\text{inc}}$, $\mathbf{H}_{\text{inc}}$).

2) Fields are updated in Domain B. The equivalent currents ($\mathbf{J}$, $\mathbf{M}$) radiate outwards and inwards from the equivalent surface. The fields obtained ($\mathbf{E}_{\text{tot}}$, $\mathbf{H}_{\text{tot}}$) are a combination of incident fields (radiation of the equivalent currents) and back-scattered fields from the human body ($\mathbf{E}_{\text{scat}}$, $\mathbf{H}_{\text{scat}}$).

3) In the original domain, the electric-field components on the surface boundary are modified according to the following relation:

$$\mathbf{E}_{\text{scat}} = \mathbf{E}_{\text{tot}} - \mathbf{E}_{\text{inc}}$$

(4)

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Fig. 2. Equivalent surface surrounding the antenna: fields on this box are recorded; the distance $d$ between the antenna and the face of the box in proximity to the human body has to be small.
where $\Omega$ represents the equivalent surface surrounding the antenna. At each time step, the electric-field components, calculated from Domain A ($\mathbf{E}_{\text{inc}}$), are subtracted from the electric-field components calculated in Domain B ($\mathbf{E}_{\text{tot}}$) to allow only the back-scattered fields to propagate inside the equivalent surface. If there are no objects near the equivalent surface, the fields are zero inside, hence, the equivalence principle is satisfied.

Fig. 4(b) shows the locations of one of six faces of the equivalent surface, specifically, the one with its normal toward the outside surface $\mathbf{\hat{y}}$. In this paper, such surfaces are aligned with the magnetic-field components. In Domain B, the electric-field components [indicated by solid gray arrows in Fig. 4(b)] are modified inside the surface with those values of the same components just outside the equivalent surface which are obtained from the simulation of Domain A. Only the field components tangential to the surface are required to be modified, for example, inside the surfaces normal to the $x$-direction, $E_y$ and $E_z$ are modified; inside the surfaces normal to the $y$ direction, $E_x$ and $E_z$ are modified; and inside the surfaces normal to the $z$-direction, $E_x$ and $E_y$ are modified.

Fig. 5 shows the magnitude of the electric field considering the radiation of an inverted $I$ antenna [see Fig. 14(a)] in proximity to the human body. In Fig. 5(a), fields are forced to zero inside the box; while in Fig. 5(b), back-scattered fields are considered inside the box as explained before.

To further increase numerical accuracy of the equivalence principle, multiple reflections between the body-worn antennas and human body have to be taken into account. For example, when a body-worn microstrip-fed antenna is characterized in the presence of a human body, a metallic sheet with the same size as the antenna ground plane is placed inside the equivalent box [see Fig. 3(b)]. This would induce multiple reflections that are similar to those between the antenna and the body surface.

Fig. 6(a) shows the importance of implementing the total-field scattered-field technique. It is noted that if the field is set to zero inside the equivalent surface (dashed-dotted line), a large error is introduced in the calculation of the radiation pattern (approximately 15 dB). The introduction of a metallic sheet (representing the antenna ground plane) inside the box causes a variation on the antenna pattern of around 1–2 dB [see Fig. 6(b)].

Fig. 7 shows the permittivity and the conductivity values of the digital phantom sliced vertically at the frequency of 2.4 GHz. The electrical properties of the principal human
tissues at 2.4 GHz, are listed in Table I. Fig. 8(a) shows the permittivity value of the muscle and the skin in the frequency range of 0.1–4 GHz. It is clearly shown that the human tissues have strong frequency dispersion; however, as shown in Fig. 8(b), changes in the values of permittivity are small at the ISM frequency band (2.4–2.48 GHz).

III. NUMERICAL RESULTS

FDTD simulations are carried out in a 3-D domain (694 mm × 449 mm × 1972 mm) with a cell size of \(dx = dy = dz = 3\) mm. The time step \(dt\) is chosen to be \(dx/\sqrt{\varepsilon c}\) (where \(c\) is the speed of light) according to the stability criterion [13]. Simulations were executed for 7000 time steps in order to reach the steady state. A 10-cell Berenger’s perfect matched layer (PML) is used to truncate the simulation domain [26]. The excitation source is applied to the tangential components of the electric and magnetic currents [24] of each cell on the equivalent surface. In the case of Fig. 4 at the shaded face \(\hat{h} = \hat{\gamma}\), the transient response of equivalent currents \(J_{x}, J_{y}, M_{x}, M_{y}\) obtained from CST Microwave Studio simulations and the use of the equivalence principle are used as an excitation source. However, for the analysis of narrowband systems, it is possible to approximate the transient response of the equivalent currents with their spectral components at the central frequency (2.4 GHz) modulated with a narrowband pulse, as shown in the following:

\[
s(t) = S(f_{0})e^{-(\frac{\nu-\nu_{0}}{\tau})^2}e^{j2\pi f_{0}t}
\]  

(5)

where \(S(f_{0})\) is the value of the spectral component of the current at the central frequency \(f_{0}\) (calculated using CST Microwave Studio), \(t_{0}\) is the time delay (it is set to 6 ns), and \(\tau\) is the time duration of the signal (it is set to 5 ns to guarantee a bandwidth of 200 MHz). Fig. 10 shows the comparison between the transient response of the equivalent current \(M_{y}\) (directly obtained from the CST Microwave Studio in the frequency band of 2.3–2.5 GHz) and the one from (5). It can be seen that two waveforms resemble each other and either of them can be used in the simulation. Fig. 11 shows the amplitude of the magnetic currents at one face \(\hat{h} = \hat{\gamma}\) of the equivalent surfaces for the body-worn patch antenna of Fig. 9. The size of the equivalent surface (where the currents are imposed) is \(14 \times 9 \times 18\) cells.

The proposed algorithm has been validated by modeling three body-worn planar antennas operating at 2.4 GHz when placed in proximity of the human body. The spacing between the antenna and human body used in our simulation is approximately 3 mm in accordance with the measurement setup. The effect of the antenna-body spacing on the radiation characteristics was studied extensively and presented by authors in [27] and does not fall into the main scope of this paper.

<table>
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<tr>
<th>Tissue</th>
<th>(\sigma) (S/m)</th>
<th>(\varepsilon_{r})</th>
<th>(\delta) (mm)</th>
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<tr>
<td>Aorta</td>
<td>1.40</td>
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TABLE I

Electric Properties of Specific Human Tissues Used Within the Visible Man Model at 2.4 GHz [25] (\(\sigma\) is the tissue conductivity, \(\varepsilon_{r}\) is the dielectric constant, and \(\delta\) is the penetration depth in millimeters)
A. Antenna Radiation Pattern

The radiation pattern has been evaluated in free space and in presence of the human body with the antenna placed in proximity of the body trunk (see Fig. 9). Radiation patterns calculated from the proposed method have been validated with measurement results.

1) Microstrip Patch Antenna: The antenna is a conventional rectangular patch which is printed on an FR4 substrate of thickness $t = 1.5324$ mm and permittivity $\varepsilon_r = 3$. The patch is fed with a 50-\(\Omega\) transmission line and a quarter-wavelength impedance transformer. Dimensions are given in Fig. 12(a).

2) Planar Monopole Antenna: The planar monopole is printed on a board with a thickness of $t = 1.6$ mm and permittivity $\varepsilon_r = 4.6$. It has a microstrip line feed with 50-\(\Omega\) impedance to guarantee a good match. The antenna dimensions are given in Fig. 13(a). Fig. 13(b) and (c) shows the radiation pattern in free space and onbody, respectively. The omni-directional azimuthal pattern is strongly modified when the antenna is placed onbody. Different from the microstrip patch,
the small-size ground plane in this case does not provide full coverage and, hence, the effect of the human body in terms of wave absorption and reflection are visible with a front-to-back ratio of around 30 dB as shown in Fig. 13(c). At 2.4 GHz, the human tissues are very lossy, the energy present on the opposite side of the human body (with respect to the antenna) is mainly due to creeping wave propagation around the trunk.

3) Inverted L Antenna: The inverted L antenna is a modified monopole (bent monopole) aiming for antenna size reduction and omnidirectional radiation. The antenna is printed on an FR4 board with a substrate thickness of $t = 1.6$ mm and permittivity $\varepsilon_r = 4.6$ and the antenna dimensions are given in Fig. 14(a). Fig. 14(b) and (c) shows a good match between simulated and measured patterns. As for the planar monopole, the onbody distortion is significant, and the front-to-back ratio is around 30 dB [as shown in Fig. 14(c)]. The use of a monopole-like antenna (instead of a patch) increases the operational bandwidth; however, the high absorption from the human body drastically reduces the radiation efficiency.

B. Path Loss Versus Transmitter–Receiver Distance

The variation of the path loss with the transmitter–receiver distance is evaluated by placing the transmitting antenna on the bottom part of the trunk, and moving the receiver probe along the trunk in various positions as shown in Fig. 15. The path loss for a communication link is given by the ratio between the transmitted and received power

$$PL[dB] = 10\log\left(\frac{P_{RX}}{P_{TX}}\right)$$

(6)
where $P_{Rx} P_{Tx}$ are the transmitted and receiver power. It is well known that the variation of path loss with the transmitter–receiver distance follows this relation:

$$PL[dB] = \gamma 10 \log(d/d_0) + P_0[dB] \quad (7)$$

where $d$ is the distance between the transmitter and receiver, $d_0$ is a reference distance, and $P_0$ is the path loss value at the reference distance. The parameter $\gamma$ shows the dependence between path loss and distance, in particular $PL \propto (d/d_0)^{\gamma}$. In [28], a flat uniform dielectric phantom has been used to model the human body torso and the exponent $\gamma$ has been estimated to be around 3.5–4 at 2.4 GHz when the transmitting and receiving antenna are placed less than $\lambda/10$ away from the phantom. In reality, the human body torso presents a curvature along the longitudinal and horizontal directions and, hence, the use of flat or cylindrical uniform dielectric phantoms can lead to an inaccuracy in the estimation of the path-loss exponent. In [29], a measurement setup has been proposed, and the exponent is estimated as $\gamma = 3.23$ for propagation along the trunk. In this paper, the effect of body-worn antenna types on the path loss is investigated by using the simulation setup proposed in Fig. 15.

Fig. 16 shows the simulated path losses with linear fitting curves. The slope of each line represents the path-loss exponent $\gamma$ for three proposed body-worn antennas and an omnidirectional source, respectively. Results have proven that the antenna radiation pattern has a significant influence on the onbody propagation channel as shown in Table II, in particular, it is possible to appreciate in Fig. 16 that the use of a directive antenna for onbody communications produces path-loss data more spread around the linear fit line when compared with an omnidirectional source. For the setup proposed in Fig. 15, it is possible to consider the antenna radiation pattern as a function which weighs the path loss; the more directive the antenna is, the less valid the linear relation is between $PL[dB]$ and $\log(d/d_0)$.

Table II lists the simulated exponent $\gamma$; its values agree with the ones reported in [28]. The inverted $L$ antenna is quite omnidirectional in all planes, and the exponent obtained is $\gamma = 3.6$, which is the closest to $\gamma = 3.9$, for the point-source case. The simulated path-loss results for the monopole antenna [Fig. 16(b)] are the most spread around the linear fit line since the antenna has a more directive beam on the elevation plane $(plane \  z-y)$, and on the plane tangential to the body $(plane \  z-x)$ when compared with the other two antennas.

### IV. CONCLUSION

In this paper, an efficient numerical technique to evaluate the performance of wearable antennas has been presented. The proposed technique has been applied to investigate radiation pattern distortion for three types of body-worn antennas. The results...
indicate that the antenna with the ground plane has the least effect from the presence of the human body but suffers from a narrow bandwidth. Furthermore, the variation of path loss with the transmitter–receiver distance has been investigated for propagation along the human body torso. It is important to include the antenna radiation pattern in the estimation of the path-loss exponent $\gamma$. The simulation results demonstrate that an antenna (e.g., inverted $L$) close to the omnidirectional source has the least spread in a linear-fit path-loss curve.

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