A Wide-Band Propagation Model Based on UTD for Cellular Mobile Radio Communications

Wei Zhang

Abstract—The present wide-band propagation model based on uniform geometrical theory of diffraction (UTD) for cellular mobile radio communications includes two major contributions. First, a UTD-based narrow-band channel transfer function containing both the diffracted electric field and the reflection of diffracted electric fields is derived. It not only is an important element of the wide-band modeling method here, but also leads to a total path-loss prediction model verified by comparisons with previously published theoretical and experimental results. In particular, the distance for horizontal placement on the street allows one to calculate the ray-path length difference (used in wide-band modeling) for the diffracted field and the reflection. Second, new refinements (including a number of explicit-form expressions to an existing method experimentally confirmed, simulating wide-band radiowave propagation for rural environments including terrain profiles) are added, making it applicable here. The method generates time-domain path loss, wide-band path loss, and relative power in the frequency domain. The time-domain path loss physically interprets and reasonably predicts the power-delay profiles. The presence of this and similar power delay profiles, as well as the behavior of relative power in the frequency domain, has been confirmed by existing wide-band propagation measurements. The value of wide-band path loss is on the order of the total path loss at the carrier frequency.

Index Terms—Land mobile radio cellular systems, propagation.

I. INTRODUCTION

The present work is an extension of previous investigations [1], [2]. It develops and addresses a wide-band propagation model based on uniform geometrical theory of diffraction (UTD) [3] for cellular mobile radio (CMR) communications [4]. The purpose of [2], as well as those of many other propagation models addressed in the open literature [4]–[13], is to predict the path loss for continuous wave (CW) or narrow-band transmission for mobile radio communications. In particular, considerable extensions to the Walfisch and Bertoni model [8] for Fig. 1 were obtained in [13], by making variable building heights and spacings, where comparisons with measurements were presented as well. It is important and necessary to study wide-band radio-channel propagation characteristics for present and future mobile communications [4], [14]–[18]. Descriptions of narrow and wide band, referring to the finite bandwidth $W$ of a receiver and to the maximum observable delay $\tau_{\text{max}}$ of the channel impulse response, were presented in [16]. Accordingly, $\tau_{\text{max}} \leq W^{-1}$ and $\tau_{\text{max}} \gg W^{-1}$ define narrow- and wide-band systems, respectively.

Manuscript received July 22, 1996; revised July 8, 1997. This work was supported by the Academy of Finland, Helsinki, Finland.

The author is with the Radio Laboratory, Institute of Radio Communications, Helsinki University of Technology, FIN 02150, Otakaari, Finland.

Publisher Item Identifier S 0018-926X(97)07994-5.

Fig. 1. Radiowave propagation in presence of buildings.

About 6000 power-delay profile measurements were made for United States digital cellular mobile radio systems [18]. A method previously developed in [15], simulating wide-band radio-channel propagation characteristics for rural environments, including terrain profiles, generates time-domain path loss that physically interprets and reasonably predicts power-delay profiles. The present model extensively utilizes the method [15] but includes two major contributions. First, a UTD-based narrow-band channel transfer function [15], which is an explicit-form expression containing the diffracted electric field and the reflection of diffracted electric fields, is obtained. Notations 1 and 2 in Fig. 1 indicate the paths for the diffracted field and the reflection, respectively. The total path loss all along path 1 was modeled in the absence of reflecting buildings in [2], where a more rigorous UTD-based expression for multiple diffractions by buildings addressed in [1] was applied. The transfer function not only is an important element of the wide-band modeling method here, but also leads to an improved path-loss prediction model. Second, refinements including a number of other explicit-form expressions to the method [15] are added, making it applicable to Fig. 1.

The narrow-band channel transfer function is first addressed in Section II. The path-loss prediction model derived from this transfer function is then presented in Section III, as well as comparisons with published results [5], [8], [19]. The band-limited impulse response (BIR) [15] is formulated in Section IV. Parameters describing wide-band characteristics are presented and calculated in Section V, including time-domain path loss, wide-band path loss, and relative power in frequency domain. The entire work is concluded in Section VI, along with discussions on future work. Finally, the Appendix derives some equations presented in Section II.

II. NARROW-BAND FORMULATION

This section addresses the narrow-band channel-transfer function that is a very important element of the method [15] and discusses its relevance to cellular mobile radio communications. The transfer function is used in the total path-
loss prediction in Section III, as well as in the utilization of the method [15] modeling wide-band radio-channel propagation characteristics in Sections IV and V.

The modeling of urban cellular mobile radio propagation in the presence of buildings (as implemented in the present work) is shown in Fig. 1, similar to illustrations presented in [4], [5], and [8]. As seen in Fig. 1, there are a number of rows of buildings between the base station and mobile. The rows of buildings have been appropriately modeled as rows of knife edges or half screens with an equal spacing d in the range of 30–60 m [7], [8]. The validity of such treatments has been addressed in the original work [8].

Let $H(f)$ be the narrow-band channel transfer function and $f$ be the frequency. Moreover, $P_r$ and $P_t$ are the radiated and received powers from the isotropic transmitting antenna and by the isotropic receiving antenna, respectively. Specifically, $H(f)$ is defined as $|H(f)|^2 = P_r/P_t$. Thus, $H(f)$, for the two rays indicated in Fig. 1, can be expressed as

$$H(f) = H_1(f) + H_2(f)$$ (1)

where $H_1(f)$ and $H_2(f)$ relate to the diffracted electric field, and to the reflection of diffracted electric fields from the wall of the building next to the mobile, respectively. The amplitude of $H_2(f)$ nearly equals that of $H_1(f)$. The phase difference between them can thus cause deep fading in the total received signal [8] indicating multipath effects. These fading effects correspond to path-length differences ranging from 10 to 105 m or to time delays $\tau$ from 35 to 350 ns. The available bandwidth $W = 100$ MHz [15] results in $W^{-1} = 10$ ns and in $350$ ns $\approx W^{-1}$. The present model, therefore, focuses on the basic two-ray case which can result in a power-delay profile and may adequately describe wide-band characteristics where $H_2(f)$ would become a distinct reflection [15], [18].

Functions $H_1(f)$ and $H_2(f)$ can be derived as

$$H_1(f) = \frac{\lambda}{4\pi D_1} \left| \frac{E_{\eta+1}}{E_0} \right| D_{\text{sh},h} \frac{D_1}{D_2(D_1 + D_2)} e^{-jk(D_1 + D_2)}$$ (2)

$$H_2(f) = R_{H,V} \frac{\lambda}{4\pi D_1} \left| \frac{E_{\eta+1}}{E_0} \right| D_{\text{sh},h} \frac{D_1}{r_2(D_1 + r_2)} e^{-jk(D_1 + r_2)}$$ (3)

where $|E_{\eta+1}/E_0|$ is the electric-field amplitude function [1], [7] for multiple diffractions caused by buildings. Here, $E_0$ and $E_{\eta+1}$ are the electric-field magnitude of an incident plane wave and its corresponding total electric field at screen position $\eta + 1$, respectively. The corresponding geometry is shown in Fig. 2. The integer $\eta$ is determined by $\eta \approx d_\text{h}/d$ and the local screen numbers $\eta + 1$ where $d_\text{h}$ is the horizontal range between the base station and local screen. In (2) and (3), $D_1$, $D_2$, and $r_2$ are the path lengths between the base-station antenna and local knife edge, between the local knife edge and mobile antenna, and for the reflection by the wall of the building next to the mobile, respectively. These path lengths indicated in Fig. 3, showing images for the local screen and mobile to determine reflection path length $r_2$ and angle $\theta$, are expressed as

$$D_1 = \sqrt{d_r^2 + (h_t - h_m)^2}$$ (4)

$$D_2 = \sqrt{d_r^2 + (h_b - h_m)^2}$$ (5)

$$r_2 = \sqrt{(2d - d_r)^2 + (h_b - h_m)^2}$$ (6)

Specifically, $h_t$ and $h_b$ are the base station and mobile antenna heights, respectively, $h_m$ is the average height of rows of buildings, and $d_r$ is defined as the horizontal range between the local screen and mobile. Fig. 3 clearly shows that $D_1 + D_2$ and $D_1 + r_2$ are, respectively, the path lengths for the diffracted field and for the reflection. Moreover, $D_{\text{sh},h}$ and $D_{\text{sh},v}$ are the UTD [3] diffraction coefficients for a spherical field from the base station transmitting antenna. Specifically, “$s$” and “$v$” denote the soft and hard boundaries, respectively. The soft and hard boundaries correspond, respectively, to the horizontally and vertically polarized electric fields [3] here. In addition, $R_{H,V}$ is the reflection coefficient [20]. Here, “$H$” and “$V$” correspond, respectively, to the horizontal and vertical polarizations. Finally, $k$ is the wavenumber and $\lambda$ is the wavelength.

The reflection angle $\theta$ (indicated in Fig. 1) can be determined by the images for a local screen and mobile (shown in Fig. 3) and is derived from

$$\tan \theta = \frac{h_b - h_m}{2d - d_r},$$ (7)
Let $\varepsilon_r$ be the relative permittivity of the reflecting building wall. Thus, $R_{HV}$ and $R_{VH}$ can be calculated as

$$R_{HV, V} = \frac{\cos \theta - a_{HV} \sqrt{\varepsilon_r - \sin^2 \theta}}{\cos \theta + a_{VH} \sqrt{\varepsilon_r - \sin^2 \theta}}$$

(8)

where $a_{HV} = 1$ and $a_{VH} = 1/\varepsilon_r$ correspond to $R_{HV}$ and $R_{VH}$, respectively. As $\varepsilon_r \to \infty$, implying a perfectly conducting wall, one obtains $R_{HV} = -1$ and $R_{VH} = 1$. Values of $\varepsilon_r$ in the range of about 5–15 were used for calculating reflections in the UHF (300 MHz to 3 GHz) band, e.g., [4], [10]–[12], and [15]. A value of $\varepsilon_r = 5$ [11] is selected for concrete walls at the 2-GHz band in the later numerical calculations. Since $\varepsilon_r$ is an input parameter, calculations at other $\varepsilon_r$ values can also be done using (3), (7), and (8). Results published in the literature on mobile radio communications indicate that the value of $\varepsilon_r$ makes negligible differences to the propagation measurements and calculations.

Let the angle $\alpha$ (shown in Fig. 1) be derived from

$$\tan \alpha = \frac{h_1 - h_2}{d_1}.$$  

(9)

Thus, $D_{s,h}^{\Pi}$ can be calculated by

$$D_{s,h}^{\Pi} = \frac{-e^{-j\pi/N}}{2\sqrt{2\pi k}} \left[ F(X_1) - \sin(\theta - \alpha)/2 \right]$$

$$+ \frac{F(X_2)}{-\cos(\theta + \alpha)/2}$$

(10)

where $\mp$ corresponds to the soft "—" and hard "++] boundaries, respectively. Moreover, $X_1$ and $X_2$ are defined by

$$X_1 = \sqrt{2kL_{\Pi}} |\sin(\theta - \alpha)/2|$$

(11)

$$X_2 = \sqrt{2kL_{\Pi}} |\cos(\theta + \alpha)/2|$$

(12)

$$L_{\Pi} = \frac{r_2}{1 + r_2/D_1}.$$  

(13)

Furthermore, $F(X)$ is the transition function defined in [3]. In the actual computation, it is appropriate to write $F(X) \approx 1$. This follows from $X < 10^{-3}$. Also, $F(X)$ can be approximated as $F(X) \approx 1$ for $X > 10$. In other cases, it is advantageous to write

$$F(X) = \sqrt{\pi X} e^{jX}$$

$$- 2j \sqrt{\pi e^{2X}} \int_0^X e^{-j\tau^2} d\tau.$$  

(14)

Specifically, (14) relates to typical radio geometry for cellular mobile communications, as defined in [1]. The UTD diffraction coefficient $D_{s,h}^{\Pi}$ can also be expressed by (10)–(13) provided that $\beta$ and $L_{\Pi}$ are replaced by $\beta_1$ and $L_{\Pi}$, respectively. The latter are written as

$$\beta = \arctan \frac{h_1 - h_2}{d_2}.$$  

(15)

$$L_{\Pi} = \frac{D_2}{1 + D_2/D_1}.$$  

(16)

The more rigorous expression of $E_{n+1}/E_0$, as it applies to the present work, is derived and analyzed in [1]. It has been expressed [1], [2] as

$$E_{n+1}/E_0 = e^{-jkr_1} c_{\cos \alpha} + \frac{D_{s,h}(\alpha)}{\sqrt{d}} e^{-jkr_1} [1 + (n-1) c_{\cos \alpha}]$$

$$\times \left[ 1 - \frac{D_{s,h}(\pi/2) e^{jkr_1} (\cos \alpha - 1) \sqrt{d}}{1 + D_{s,h}(\pi/2) e^{jkr_1} (\cos \alpha - 1) \sqrt{d}} \right],$$

(17)

The UTD diffraction coefficients $D_{s,h}(\pi/2)$ and $D_{s,h}(\alpha)$ can be calculated by

$$D_{s,h}(\pi/2) = \frac{-e^{-j\pi/4}}{2 \sqrt{2 \pi k}} \left[ \sqrt{\pi k} d_2 \cosh \alpha/2 \pm 1 \right]$$

$$D_{s,h}(\alpha) = \frac{-e^{-j\pi/4}}{2 \sqrt{2 \pi k}} \left[ \frac{F(X_3)}{\sin(\alpha/2)} \mp \cos(\alpha/2) \right]$$

(18)

where $\sqrt{X_3} = \sqrt{2kL_{\Pi}} \sin(\alpha/2)$. The minus sign in front of $\sqrt{X_3}$ on the right-hand side of (18) ensures that the diffracted field point is located in the shadow zone [1].

The present model applies only to the mobile in the shadow zone. In other words, $\alpha$, $\beta$, and $\theta$ must satisfy $\beta \geq \theta > \alpha$. Naturally, (7) and (15) ensure that $\beta \geq \theta$ is satisfied. Thus, (7) and (9) lead to

$$\frac{h_1 - h_2}{2d_2} > \frac{h_1 - h_2}{d_1}.$$  

(20)

This defines the region of applicability for the present model. Taking $h_1 = 50$ m, $h_2 = 10$ m, $h_3 = 1.6$ m, $d = 60$ m, and $d_r \geq 5$ m, (20) leads to $d_r > (h_1 - h_2)(2d - d_r)/(h_1 - h_2)$ < 548 m. In fact, $d_r > 548$ m, together with other its corresponding parameters, applies to cellular mobile radio communications in most cases. At $d_r = 1000$ m, (20) results in $h_1 < h_2 + d_r(h_1 - h_2)/(2d - d_r) > 383$ m. Also, $h_1 < 83$ m corresponds well to practical situations for the CMR communications. The derivation of (2) and (3) is presented in the Appendix.

III. Path-Loss Prediction Model

This section applies results presented in Section II to a total path-loss prediction model. It also makes comparisons of the present model with the model addressed in [8] and with a study based on measurements reported in [19]. Neither [8] nor [19] included or presented results of total path loss as a function of $d_r$. Both presented results of the total path loss only for the vertically polarized transmission, which practically applies to mobile radio communications.

A. Explicit-Form Expressions

The total path loss $L_t = -10 \log_{10}(P_r/P_i) = -20 \log_{10}[H(f)]$ in dB corresponding to Fig. 1 can be derived from (1) through (3) as

$$L_t = L_0 + L_1 + L_{md} + L_r$$  

(21)

where $L_0$, $L_1$, and $L_{md}$ are the losses for free-space radiowave propagation, due to the local row numbering $n+1$ of buildings in the vicinity of the mobile, and due to multiple diffractions caused by rows of buildings between the base station and local row of buildings, respectively. These losses correspond only to path 1 for the diffracted electric field [2] without considering the building next to the mobile [7]. To include the contribution
of path 2 in the present model, $L_r$ in dB is the loss caused by the reflection of the diffracted electric fields from the wall of the building next to the mobile. Specifically, $L_0 + L_1 + L_r$ in dB is the path loss for the geometry shown in Fig. 3, developed from illustrations presented in [2], [7], and [8], including this reflection.

The free-space loss $L_0$ referring to the model in [6] has been written [2] as

$$L_0 = 20 \log_{10} f (\text{GHz}) + 20 \log_{10} D (\text{km}) \quad (22)$$

$$D = \sqrt{(d_t + d_r)^2 + (h_t - h_r)^2} \quad (23)$$

where $D$ in kilometers is the path length between the base station and mobile antennas in free-space. The loss $L_4$ due to the local screen has been derived [2] as

$$L_4 = -20 \log_{10} \left[ \frac{D_1(D_t + D_2)}{D^2} \right] + 10 \log_{10} D_2 (\text{m}) \quad (24)$$

The loss $L_{md}$ is also termed additional path loss and has been calculated as

$$L_{md} = -20 \log_{10} \left| \frac{E_{n+1}}{E_0} \right| \quad (25)$$

The loss $L_r$ caused by the reflection can be written as

$$L_r = -20 \log_{10} \left[ 1 + \frac{R_{HH}}{r_2} \right] \quad (26)$$

Clearly, $L_r$ depends on the polarization of transmissions, as well as on the phase difference denoted by $k(r_2 - D_2)$ for the two rays. Also, it accounts for multipath effects that would result in deep fading of the total received signal.

B. Numerical Results

Predictions of $L_t$ for both vertical and horizontal polarizations can be performed for given parameters $f, h_t, h_r, d_t, d_r,$ and $d$. Examples of $L_t$ at 2.154 GHz as functions of $d_t, h_t,$ and $d_r$ are presented in Figs. 4–6, respectively. The solid and dashed lines are, respectively, for vertically and horizontally polarized transmissions. The plots of $L_n, L_{md}, L_4,$ and $L_t$ are indicated by “reflection,” “multiple,” “local screen,” “free space,” and “total,” respectively. Clearly, $L_0 > L_4 > L_{md},$ together with the behavior of $L_n,$ shows their relative contributions to $L_t$ in Figs. 4 and 5. It can be seen in Figs. 4(b) and 5(b) that $L_t$ for horizontal polarization is about 7–8 dB larger than that for vertical polarization at $d_r = 55$ m. At $d_r = 5$ m, however, $L_t$ in Figs. 4(a) and 5(a) is much less sensitive to the polarization. The polarization dependence of $L_t$ is mainly caused by $L_4$ and $L_r.$ In fact, $L_0$ does not depend on, and $L_{md}$ is insensitive [1] to, the polarization. The free-space loss $L_0$ increases with $d_t$ and is relatively insensitive to $h_t$ at $d_t = 1000$ m, but $L_4$ is less sensitive to $d_t$ and to $h_t.$ Both $L_0 < 0$ and $L_r > 0$ occurred, where $L_r < 0$ enhances the received power. Ripples of the $L_t$ curves in the small $d_t$ range are caused by $L_{md}$ in Fig. 4. In the region where $L_{md}$ is negligible, $L_t$ can be approximated as $L_t \approx L_0 + L_1 + L_r.$ In the region where $L_{md}$ becomes considerable, however, $L_t$ is calculated by (21). The breakpoint between the two regions occurs in the vicinity of $\alpha \sqrt{d/\lambda} = 0.4 [7], [1],$ where $\alpha \sqrt{d/\lambda}$ in radians is a parameter introduced in [8]. In both regions, the behavior of $L_t$ depends essentially on $\alpha \sqrt{d/\lambda}.$ Obviously, $L_t$ as well as $L_{md}$ increases as $\alpha \sqrt{d/\lambda}$ decreases.
The Walfisch and Bertoni model reported in [5] and [8] has been experimentally confirmed. Dashed dot (---) lines included in Figs. 4 and 5 present results of the Walfisch and Bertoni model computed by [8, Eqs. (15), (16), (19), and (20)]. It is shown that two models agree well at the common applicable range. The Walfisch and Bertoni model results correspond approximately to a received power range dependence of \(1/d^2\), applying to the parameter \(\alpha \sqrt{d/\lambda}\) at the interval of \(0.02 \leq \alpha \sqrt{d/\lambda} \leq 0.4\). In the \(\alpha \sqrt{d/\lambda} \leq 0.4\) region of Figs. 4 and 5, \(-L_p\) for the vertical polarization is on the order of 2–3 dB. This also agrees with the prediction by Walfisch and Bertoni [8]. Their model reduced the path loss by \(10 \log_{10} 2 \approx 3\) (dB) to include reflections from buildings next to the mobile and other sources, which may result in an amplitude of electric fields nearly equal to that of the primary diffracted field [8].

Since \(L_{md}\) is negligibly small in the region of \(\alpha \sqrt{d/\lambda} > 0.4\), \(L_t\) can be approximated as \(L_t \approx L_0 + L_1 + L_p\). In actual mobile radio communication situations, \(L_p\) appears to be a path-loss deviation. Such a trend shown in Fig. 4 reasonably agrees with measurements performed at 927 MHz for a base-station antenna height \(h_t = 50\) m [19]. Based on their measurement results, the authors of [19] state that the propagation loss may be modeled as free-space propagation loss plus a nearly constant value (37.9 dB) in the \(d_t\) range of 100–800 m. This corresponds to a received power range dependence of \(1/d^2\). Their experimental breakpoint between received power range dependences \(1/d_1^2\) and \(1/d_2^2\) took place at \(d_t = 800\) m. As an estimate, an average building height of \(h_b = 15\) m, together with (9), leads to the \(\alpha \sqrt{d/\lambda}\) values of 0.42–4.58 for the breakpoint, where \(d\) in the range of 30–60 m was applied. In Fig. 5, the region of \(\alpha \sqrt{d/\lambda} > 0.4\) covers \(h_t > 50\) m, where \(L_t\) is relatively insensitive to \(h_t\).

Curves of \(L_p\) overlap those of \(L_{md} \approx 0\) in Fig. 6 and, therefore, the \(L_{md}\) curves are omitted for clarity. Here, \(L_0\) is relatively insensitive to \(d_t\) at a \(\Delta d_t\) increment of 0.1 m for \(d_t = 1000\) m. At small values of \(d_t\), \(L_t\) for the vertical polarization is larger than \(L_t\) for horizontal polarization. Such a relation will not be valid once \(d_t\) becomes larger than about 10 m. Also, \(L_t\), together with \(L_1\), causes \(L_t\) to be sensitive to \(d_t\) as well as to the polarization. Moreover, \(L_1\) decreases and depends less and less on the polarization as \(d_t\) increases.
IV. BAND-LIMITED IMPULSE RESPONSE

This section addresses band-limited impulse response and related terms introduced in [15]. As a second major contribution of the present work, a number of explicit-form expressions are introduced, making them applicable to modeling wide-band propagation for the CMR communications.

Let \( \hat{h}(t) \) be the band-limited impulse response where \( t \) denotes the time. The approximation to \( \hat{h}(t) \) is written in [15] as

\[
\hat{h}(t) = \hat{h}_1(t) \cos(2\pi f_0 t) + \hat{h}_Q(t) \sin(2\pi f_0 t) \tag{27}
\]

where \( \hat{h}_1(t) \) and \( \hat{h}_Q(t) \) are the in-phase and quadrature components [15], respectively, and \( f_0 \) is the carrier frequency. In this work, \( \hat{h}_1(t) \) and \( \hat{h}_Q(t) \) are derived from \( \hat{h}_{3P}(t) = \hat{h}_1(t) + \mathbf{i}\hat{h}_Q(t) \). The complex-channel impulse response \( \hat{h}_{3P}(t) \) is written as

\[
\hat{h}_{3P}(t) = \left[ H(f_0) + \sum_{n=1}^{N/4-1} X_T(n\Delta f) \right] \Delta f \tag{28}
\]

where \( X_T(n\Delta f) \) [15] is the truncated frequency spectrum of the actual frequency spectrum \( X(f) \) of a periodically repeated pseudonoise (PN) sequence, \( f \) denotes the radio frequency (RF) of the baseband, \( \Delta f \) is the frequency increment, and \( N\Delta f/4 = f_c \) where \( f_c \) is the shift register clock frequency.

The truncated frequency spectrum \( X_T(f) \) [15] has been approximated as

\[
X_T(f) = \frac{\sin^2(\pi f/f_c)}{(\pi f/f_c)^2} \tag{29}
\]

for \( |f| < f_c \). In other cases, it is written as \( X_T(f) = 0 \). The present work also explicitly writes the autocorrelation function \( R_{xx}(t) \) of the input signal [15], [21] as

\[
R_{xx}(t) = \left[ 1 + 2 \times \sum_{n=1}^{N/4-1} X_T(n\Delta f) \cos(2\pi n\Delta ft) \right] \Delta f \tag{30}
\]

Fig. 7 presents the behavior of \( R_{xx}(t) \) calculated at \( f_c = 50 \) MHz and \( N = 4096 \) where \( R_{xx}(t) \) has a maximum value of 0.09028230 \( f_c \), which applies to the later numerical calculation. The number of \( N = 4096 \) corresponds to a maximal PN sequence, i.e., \( m \) sequence with 1023 (= \( N/4-1 \)) chips [15]. The corresponding frequency increment \( \Delta f \) needed in the fast Fourier transform (FFT) calculation, i.e., (28) and (30), is calculated by \( \Delta f = 4f_c/N = 0.048828125 \) MHz. The resulting time increment is \( \Delta t = 1/(4f_c) = 5 \) ns, corresponding to a path length difference of 1.5 m. The total time \( T \) displayed is thus calculated by \( T_0 = N\Delta x = N/(4f_c) = 20.48 \) µs. The inverse fast Fourier transform (IFFT) calculation could span a frequency range of \( FB = N\Delta f = 4f_c = 200 \) MHz.

The first main spectral lobe \( f_0 \neq f_c \) results of \( H(f) \) are used in the \( \hat{h}(t) \) calculation. To distinguish from \( L_t \) defined in Section III, let \( L_n \) in decibels be the narrow-band path loss defined by

\[
L_n = 20 \log_{10} |H(f)| = -L_t \tag{31}
\]

Figs. 8 and 9 present the \( L_n \) results calculated at \( f_0 = 2.154 \) GHz and \( f_c = 50 \) MHz for the frequency range of \( f_0 \neq f_c \) corresponding to Fig. 1. The interference between the diffracted field and its reflection causes oscillations of the \( L_n \) curves. Such multipath effects correspond to deep fading of the total received signal. The phase difference between two neighboring extremes satisfies

\[
2\pi = \frac{2\pi}{c} \delta f(r_2 - D_2) \tag{32}
\]

where \( \delta f \) is the frequency difference and \( c \) is the speed of light in free-space. Thus, \( \delta f \) observed in Figs. 8 and 9 can be calculated as

\[
\delta f = \frac{c}{r_2 - D_2} \tag{33}
\]

In Fig. 8, \( d_e \) results in \( r_2 \approx 115 \) m, \( D_2 \approx 10 \) m, and \( \delta f \approx 2.857 \) MHz. The number of extremes is thus determined by \( 2f_c/\delta f \approx 35 \). Indeed, 35 extremes appear in Fig. 8. Similarly, \( d_e \approx 55 \) m results in \( r_2 \approx 66 \) m, \( D_2 \approx 56 \) m, and \( \delta f \approx 30 \) MHz. As a result, the number of extremes is determined by \( 2f_c/\delta f \approx 3 \) and three extremes are clearly observed in Fig. 9.

V. WIDE-BAND CHARACTERISTICS

This section calculates time-domain path loss, wide-band path loss, and relative power in frequency-domain for the CMR communications. These parameters are introduced in [15], however, for completeness, definitions, in addition to some explicit-form expressions, are presented here.

A. Time-Domain Path Loss

The band-limited impulse response \( \hat{h}(t) \) can also be expressed as \( \hat{h}(t) = \hat{E}(t) \cos(2\pi f_0 t - \phi(t)) \) where \( \hat{E}(t) = \sqrt{\hat{i}_1^2(t) + \hat{i}_Q^2(t)} \) and \( \tan \phi(t) = \hat{h}_Q(t)/\hat{h}_1(t) \). Specifically, the square of the envelope \( \hat{E}(t) \) represents the time-dependant
ZHANG: MODEL BASED ON UTD FOR CELLULAR MOBILE RADIO COMMUNICATIONS

Fig. 8. Narrow-band path loss $L_n$ calculated at $d_f = 1000$ m, $h_t = 50$ m, $h_b = 10$ m, $h_r = 1.6$ m, $d = 60$ m, and $d_r = 5$ m.

Fig. 9. Narrow-band path loss $L_n$ calculated at $d_f = 1000$ m, $h_t = 50$ m, $h_b = 10$ m, $h_r = 1.6$ m, $d = 60$ m, and $d_r = 55$ m.

The $L_p(t)$ results corresponding to calculations presented in Figs. 8 and 9 are presented in Figs. 10 and 11, respectively. The first and second peaks in Fig. 10 take place at $t_1 = 3370$ ns and $t_2 = 3720$ ns, corresponding to path lengths $D_1 + D_2 = 1011$ m and $D_1 + \gamma_2 = 1116$ m for the diffracted field and for the reflection, respectively. Clearly, the second peak is larger than the first. The presence of such peaks, i.e., power delay profiles, has been confirmed by wide-band propagation measurements [15]–[18]. The time-domain path loss would be used in considering the performance of digital communication systems. Although the first and second peaks in Fig. 11 overlap, maxima appear at $t_1 = 3520$ ns and $t_2 = 3555$ ns, corresponding to path lengths $D_1 + D_2 = 1056$ m and $D_1 + \gamma_2 = 1066$ m for the diffracted field and for the reflection, respectively.

The received power. If $L_p(t)$ is the time-domain path loss in decibels, it can be calculated [15] by

$$L_p(t) = 20\log_{10} \left[ \frac{E(t)}{E_{max}} \right].$$  \hspace{1cm} (34)
reflection, respectively. Unlike the peaks evident in Fig. 10, the first peak is now larger than the second one. The entire time-domain range of $T_t = 20,480$ ns contains 4096 values of $L_p(t)$.

The maximum of $L_p(t)$ could be a reasonable estimate of the wide-band path loss [15]. Let $L_{pvm}$ and $L_{pdm}$ be the $L_p(t)$ maxima for the vertically and horizontally polarized transmissions, respectively. They satisfy a relation as $|L_{pvm}| < |L_{pdm}|$ in Figs. 10 and 11. In the latter, the first $L_p(t)$ peaks corresponding to the diffracted field take the values of $L_{pvm}$ and $L_{pdm}$. In Fig. 10, however, the second $L_p(t)$ peaks corresponding to the reflection represent $L_{pvm}$ and $L_{pdm}$ positions. Clearly, the reflection from buildings existing next to the mobile should be considered in the total path-loss calculation.

B. Wide-Band Path Loss

The wide-band path loss [15] $L_w$ in dB is now written as

$$L_w = 10 \log_{10} \left[ \frac{P_R}{P_T} \right]$$

$$P_T = 1 + 2 \times \sum_{n=1}^{N/4} |X_T(n\Delta f)|^2$$

$$P_R = |H(f_0)|^2 + \sum_{n=1}^{N/4} |H(f_0 - n\Delta f)|^2$$

$$+ |H(f_0 + n\Delta f)|^2 |X_T(n\Delta f)|^2$$

where $X_T(-f) = X_T(f)$ is used in rewriting $P_T$, and $P_R$ defined, respectively, by [15, Eqs. (22) and (23)]. Let $L_{wvh}$ and $L_{whv}$ in decibels be the wide-band path losses of $L_w$ for the vertically and horizontally polarized transmissions, respectively. Values of $L_{wvh}$ and $L_{whv}$ corresponding to results shown in Figs. 8 and 9 are presented in Table I. The maxima of $L_{pvm}$ and $L_{pdm}$ obtained from Figs. 10 and 11 are also included. The corresponding narrow-band path loss $L_n = -L_t$ at $f_0 = 2,154$ GHz is presented as well. Here, $L_{wvh}$ and $L_{whv}$ are the $L_n$ values for the horizontally and vertically polarized transmissions, respectively.

Results presented in Table I clearly satisfy the relation $|L_{wvh}| < |L_{whv}|$. In general, $L_{pvm}$ and $L_{pdm}$ are approximations to $L_{wvh}$ and $L_{whv}$, respectively. They satisfy the relations $|L_{wvh}| < |L_{pvm}|$ and $|L_{whv}| < |L_{pdm}|$. Moreover, the contribution by the diffracted field to the received power is dominant for a large horizontally local range $d_r = 55$ m. In particular, the difference of $|L_{pvm} - L_{whv}|$ is about $10 \log_{10} 2 \approx 3$ (dB). This agrees with the discussions presented in [8] where the 3-dB received power enhancement is considered and included in the path-loss calculation for vertical polarization. For a small horizontally local range $d_r = 5$ m, however, the contribution by the reflection to the received power becomes dominant. The difference of $|L_{pvm} - L_{whv}|$ considerably deviates from 3 dB. The difference of $|L_{wvh} - L_{whv}|$ is larger than that of $|L_{pvm} - L_{pdm}|$ for the specific CMR communications. To avoid larger errors, $L_{wvh}$ and $L_{whv}$ may not always be used to evaluate the wide-band path losses $L_{wvh}$ and $L_{whv}$, respectively. The technique measuring $L_{wvh}$ was described in [15]. It can be seen that $|L_{wvh}| < |L_{whv}|$ appears for $d_r = 55$ m, although $|L_{wvh}| > |L_{whv}|$ occurs at $d_r = 5$ m.

C. Relative Power in Frequency Domain

Let $\hat{P}(f)$ be the Fourier transform of the magnitude $\hat{E}(t)$ calculated by

$$\hat{P}(f) = \sum_{n=1}^{N} \hat{E}(n\Delta t)e^{-j2\pi fn\Delta t}.$$

All 4096 $\hat{E}(t)$ values in the entire time-domain range of $T_t = 20,480$ ns are used in this Fourier transform calculation. Clearly, $\hat{P}(-f) = \hat{P}^*(f)$ is satisfied where $\hat{P}^*(f)$ is the conjugate of $\hat{P}(f)$. This leads to $|\hat{P}(f)| = |\hat{P}(f)|$. Let $P(f)$ in decibels be relative power in the frequency domain defined by

$$P(f) = 20 \log_{10}|\hat{P}(f)|.$$

Similar calculated and measured results were presented in [15]. As a refinement to the term introduced in [15], (38), and (39) are explicitly addressed here. Figs. 12 and 13 present $P(f)$ for both vertically and horizontally polarized transmissions, corresponding to results shown in Figs. 10 and 11, respectively. Indeed, $P(f)$ spans a frequency range $f_B = 4 f_c = 200$ MHz.
It is anticipated and reasonable that the main spectral lobe of $P(f)$ appears in the finite bandwidth $W = 2f_c = 100$ MHz, i.e., from $-50$ to $50$ MHz. Such a frequency-range behavior has been confirmed by wide-band propagation measurements [15].

VI. CONCLUSION

The present work includes two major contributions to the modeling of wide-band radiowave propagation in urban areas.

First, a UTD-based narrow-band channel transfer function, containing the diffracted electric field and the reflection of diffracted electric fields, is derived. It not only is an important element of the wide-band modeling method here, but also leads to a total path-loss prediction model for the specific CMR communications. The model has been verified by comparisons with the Walfisch and Bertoni model and with a study based on measurements. In particular, the distance between the mobile and local row of buildings allows one to calculate the ray path-length difference, used in wide-band modeling, for the diffracted field and the reflection. Second, modifications to an existing and validated method enable the modeling of wide-band radiowave propagation here. The method generates time-domain path loss, wide-band path loss, and relative power in frequency domain. The time-domain path loss physically interprets and reasonably predicts power delay profiles. Existing wide-band propagation measurements have confirmed the presence of this and similar power-delay profiles as well as the reasonable frequency-range behavior of relative power in frequency domain. It was expected that the value of wide-band path loss is on the order of the total path loss at the carrier frequency.

Future work should extend the path-loss model to three-dimensional situation treating the more realistic case of transmitter and receiver obliquely intersecting the building walls. It should also contain experimental and theoretical investigation of the relative importance of multiple reflections, usually with smaller field amplitudes, within street canyons for digital communication systems. The entire wide-band modeling method presented in the paper can be extended and applied to other site specific radio propagation predictions [4], [22] for wireless personal communications. In other words, once the narrow-band channel transfer function for a site-specific case is obtained, the corresponding wide-band propagation characteristics can be predicted by using the explicit-form expressions defined here. It is important that future wide-band propagation experiments for the specific CMR communications (as well as for other site specific cases) be performed. These experiments should be complemented by presenting necessary parameters describing the corresponding environments. This would allow not only testing and refinement of the present model, but also the extension to other realistic cases.

APPENDIX

DERIVATION OF (2) AND (3)

To derive (2) and (3), let $E_i$ and $E_d$ be the incident electric field of spherical waves and the total electric field received, respectively. These electric fields for Fig. 3 can be written [3] as

$$E_d = E_i D D_{s,h} A^1 e^{-jkD_2} + R_{h,v} E_i D_{s,h} A^{II} e^{-jkD_2}$$  \hspace{1cm} (A.1)

$$A^1 = \sqrt{\frac{D_1}{D_2(D_1 + D_2)}}$$  \hspace{1cm} (A.2)

$$A^{II} = \sqrt{\frac{D_1}{r_2(D_1 + r_2)}}$$  \hspace{1cm} (A.3)
On the right-hand side of (A.1), the first and second terms represent the diffracted electric field and the reflection of diffracted electric fields, respectively. Here, \( E^d \) can be expressed as

\[
\frac{E^d}{\sqrt{2}} = \frac{\sqrt{\lambda D_1}}{D_1} e^{-j k D_1}.
\] (A.4)

Letting \( 0 < d_r \to d_r \), one obtains \( r_2 \to D_2, A^H \to A^I, \) and \( \bar{D}_{s,h}^\Pi \to \bar{D}_{s,h}^\Iota \). It is instructive to notice that a perfectly conducting wall corresponds to \( R_H = 1 \) and \( R_V = 1 \). In this case, the boundary condition of \( \nabla \times \vec{E} = 0 \) is satisfied; the total tangential component of electric fields appears to vanish near the perfectly conducting wall surface. The received power \( P_r \) by the isotropic receiving antenna can be written as

\[
P_r = \frac{\lambda^2 |E|^2}{4\pi} \frac{P_{\text{t}}}{2\eta_0}
\] (A.5)

where \( \eta_0 \approx 120\pi \) is the free-space impedance. To include multiple diffractions by rows of buildings present indicated in Fig. 2, \( E^d \) in (A.1) should be replaced by \( E^d[I_n+1/E_0] \). For Fig. 1, (A.1), (A.4), and (A.5) result in

\[
P_r = \left( \frac{\lambda}{4\pi D_1} \right)^2 \left| E_{n+1} \right|^2
\cdot \left| D_{s,h}^I A^I e^{-j k (D_1 + D_2)} + R_{HV} D_{s,h}^\Pi A^H e^{-j k (D_1 + r_2)} \right|^2,
\] (A.6)

Let \( H(f) = H_1(f) + H_2(f) \) be explained as \( |H(f)|^2 = P_r/P_t \). Thus, (A.2) and (A.3) lead to (2) and (3). Both \( H_1(f) \) and \( H_2(f) \) can be written as \( |H_1,2(f)|^2 = P_{1,2} / P_t \), where received powers \( P_{1,2} \) and \( P_2 \) correspond to \( H_1(f) \) and \( H_2(f) \), respectively. Therefore, \( H(f) \) is also explained as \( |H(f)|^2 = P_r / P_t \) for each of the two rays.

ACKNOWLEDGMENT

The author would like to thank Prof. A. Räissänen, Associate Prof. P. Vainikainen, and Prof. V. Porra for their encouragement. He would also like to thank the reviewers for their comments, which improved the paper.

REFERENCES


Wei Zhang received the Dr. Tech. degree in electrical engineering from Helsinki University of Technology (HUT), Espoo, Finland, in 1994. From August 1986 to October 1989, he worked at the Qingdao Research Center of China Research Institute of Radiowave Propagation, Qingdao, China, doing research mainly on radar remote sensing of precipitation, attenuation, and depolarizations due to sand storm and smoke, and incoherent scattering by layered medium composed of spheroidal particles. During November 1989 to December 1993, he worked as well as studied at the Radio Laboratory of HUT, investigating a model of a melting layer of precipitation, attenuation and depolarizations due to the melting layer and rain, scattering properties of the melting layer, incoherent scattering evaluation for layers of melting layer and rain, and single scattering by spheroidal hail and sleet particles in the resonance region. Since January 1994, he has been with the Institute of Radio Communications of HUT. His current research interest includes investigating both narrow- and wide-band propagation characteristics for mobile radio communications.