Research Article

A Unified Geolocation Channel Model—Part I (Path Loss)

Ilir F. Progri

Gifet Inc., 5 Euclid Ave. #3, Worcester, MA 01610, USA

ORCID: 0000-0001-5197-1278

Correspondence should be addressed to Ilir Progri; iprogr@gifet.com

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Indoor geolocation systems have emerged and have attracted a wide audience recently. There are three important components that constitute an indoor geolocation system: transmitter, receiver, and indoor channel model. To some extent the primary transmitter design requirements are transmit power and a frequency or a set of frequencies of operation. The receiver is designed to operate at a certain distance from the transmitter and at the same frequency or set of frequencies of operation. Hence, any indoor channel model must take into account these two important design requirements.

The indoor channel is perhaps very intriguing because multipath is ubiquitous due to high signal scattering, reflection, and refraction. It is been widely accepted in the literature that the two primary channel effects are: path loss and multipath distribution. First, the path loss model is currently accepted to be a function of the transmitter and receiver geometry and frequency of operation. Second, the most widely used and accepted indoor channel multipath distribution models are Raleigh, Rician, and lognormal. As indicated in the second paper (or part II) the Rayleigh distribution is wider than Rician and Rician is wider than lognormal. This implies that the Rayleigh fading channel is the most severe and Rician fading is more severe than lognormal and lognormal is the least severe channel model.

Unfortunately, the dependency of the path factor on the frequency of operation in the context of a unified channel model is overlooked in the literature. Furthermore, there is no agreement on an indoor multipath distribution channel model. Although a unified channel model has not yet been found, in this paper we make a first attempt to present a unified channel model which consists of a unified path-loss model and a unified multipath distribution model.

The unified path loss model consists of an approach for linking together the path loss models of the three geolocation systems (macro-outdoor, micro-outdoor, and indoor) with the distance between the transmitter and receiver, R , and the frequency of operation, f . Although there are several parameters that affect the power loss factor, we consider R and f as the most important parameters for two reasons. Most of the geolocation systems presented in the literature are based on a direct measure of the time of travel; i.e., distance between the transmitter and receiver. The frequency of these systems varies; hence, the path loss factor varies as well.

While unifying the path loss model was to some extent initiated in the literature, unifying the multipath distribution model is currently a silent quest in the literature. It appears that the tendency is to come up with newer and more

sophisticated models that would explain the characteristics of the old models. The approach that we present in part II (on in a separate paper in the future) to unify the multipath distribution models is rather simple.

The theoretical performance results are validated from the measurements currently reported in the literature.

Index Terms— Indoor, geolocation, systems, transmitter, receiver, channel, path-loss, multipath, Raleigh, Rician, lognormal.

1 Introduction

The communication channel is defined as the medium (or the environment) between the transmitting antenna and the receiving antenna [1]-[25], as depicted in Fig. 1. An electromagnetic wave propagating through a communications channel undergoes *loss* in power and *dispersion* in direction.

The ability of the medium to absorb an electromagnetic wave depends on its physical properties, which is known as propagation power loss. The power loss (in dB) can be inversely proportional with twice, three times, 4 times, or n times the transmitter receiver distance or with m times the signal frequency where n and m are real; therefore, the power loss can be of a quadratic, cubic, biquadratic, or n^{th} order law of the inverse distance and m^{th} order law of the inverse signal frequency.

The *dispersion* of an electromagnetic wave results from non-uniformity of the geometry and physical properties of the environment. The three most common phenomena that perturb an electromagnetic wave are *reflection*, *refraction*, and *scattering* (or *diffraction*). When there is a direct path between the transmitter and the receiver, the signal is received through the line-of-sight (LOS) path. A signal that is received through the LOS path loses power as a result of the lack of conductivity of the medium and through refraction which occurs as a result of the existence of physical layers with different refractive coefficients. Part of the LOS signal may also be reflected and scattered. When a signal is received through paths different from the LOS path it is said that the signal is received through *non LOS* (NLOS) paths. Like the LOS path, these paths also undergo reflection, refraction, and diffraction. Generically, the outcome of the signal dispersion is called *multipath*.

This paper is organized as follows: First, we consider the path loss wireless communication channel models. In this consideration the idea is to systematize the path-loss wireless channel model and come up with a unified path loss model with is considered next. Third, the paper is concluded with a summary and conclusions section.

2 Geolocation Channel Models

The properties of geolocation system (or wireless communications) channels have been the focus of research performed by many communication and fields engineers for many years [1]-[25]. This has led to combined efforts from scientists and engineers to come up with channel models that are easy to model, allow for accurate predictions, and are computationally efficient. It is important to emphasize here that although there is yet to be found a unique and complete channel model, there are several models that offer good to very accurate prediction of channel behavior under certain conditions. Many channel models have been validated through experimental measurements. Nevertheless, there is a need to conduct more measurements and it is desirable to refine current analytical models in a way that leads to a unified channel model. We would like to address to some extent an approach that might lead to a unified wireless communication model. First, we start with the path loss model.

Based on the current conventions in the wireless community [6]-[17], and [25], channel models are classified into three categories: macro-outdoor geolocation systems, micro-outdoor geolocation systems, and indoor geolocation systems.

3 Macro-outdoor Geolocation Path Loss Model

A macro-outdoor geolocation system consists of a network of transmitters and receivers in which the transmitting antenna has a coverage radius going from approximately 1 km to 20 km [13], [15]. As we have discussed in the introduction, there are two primary channel effects: path loss and multipath distribution (which is discussed in a separate publication or part II [2]).

The propagation path between a transmitter (TX) and a receiver (RX) is illustrated in Fig. 2 (a) horizontal view (or looking from the top) and (b) vertical view (or looking from forward). There are four buildings, one transmitter, and one receiver as illustrated in Fig. 2(a). As shown in Fig. 2 (b) the distance between two consecutive buildings is d and the height

of each building is z_b . The transmitter's height is z_t and the receiver's height is z_r . The ray coming from the transmitter is refracted from the rooftop of the third building; therefore, the receiver receives one refracted path whose length is r_1 and one reflected path whose length is r_2 as shown in Fig. 2 (b).

The length of the direct path from the transmitter to the rooftop of the third building is R . The incident angle of the direct ray with the horizontal plane is ϕ Fig. 2 (a) and the vertical phase is θ Fig. 2 (b). Next, we shall see how to assess the path loss based on the model shown in Fig. 2 (a) and (b).

The free-space path loss, Q_0 , is the ratio of received power, P_r , to radiated power, P_t , for isotropic antennae in free-space. Assume that the receiving antenna with gain, G_r , is located at a distance, R , from the transmitting antenna with gain, G_t . The sector average power, P_r , from the receiving antenna is, according to [3], equal to

$$P_r = \left(\frac{c}{4\pi f R}\right)^2 G_t G_r P_t = Q_0 G_t G_r P_t \quad (1)$$

where the free-space path loss, Q_0 , is determined from [3]

$$Q_0 = \left(\frac{\lambda}{4\pi R}\right)^2 = \left(\frac{c}{4\pi f R}\right)^2 \quad (2)$$

Similarly, the average *path loss*, Q , is defined as the ratio of the sector average received power, P_r , to the radiated (or transmitted) power, P_t , times the factor $G_t G_r$ [15]. Employing Bertoni's definition [15], the path loss can be written as a product of three components: (1) free-space path loss, Q_0 ; (2) reduction factor due to previous rows propagation, Q_1^2 ; and (3) reduction due to diffraction, Q_2

$$Q = \frac{P_r}{G_t G_r P_t} = Q_0 Q_1^2 Q_2 \quad (3)$$

The free-space path loss, Q_0 , corresponds to the path loss from the transmitter to the rooftop of the third building as shown in Figs. 2(a) and 2(b).

The field propagating through the rooftop of a building is approximated as a plane wave propagating parallel to the ground with an azimuth angle, $90^\circ - \phi$, where ϕ is the angle between the plane propagation vector and the horizontal direction x (see Figs. 2(a) and 2(b)). The receiver receives two paths of the same original plane wave: one diffracted path and the other reflected path, which are function of the elevation angle θ . The signal received at the receiver is reduced by the factor, Q_2 , given by [15]

$$Q_2 = \frac{1}{2\pi k \cos \phi} \left[\frac{1}{r_1} D^2(\theta_1) + \frac{1}{r_2} \Gamma^2 D^2(\theta_2) \right] \quad (4)$$

where Γ is the reflection coefficient at the building face. The wavelength λ is defined as the distance for which $k\lambda = 2\pi$, that yields the known expression for the wavenumber, k , equal to $k = 2\pi/\lambda$. The wavenumber k is interpreted as the number of wavelengths in a distance of 2π and its units are m^{-1} . The diffraction coefficient $D(\theta_i)$, $i = \{1, 2\}$ is determined from [15]

$$D(\theta_i) = \frac{1}{\theta_i} - \frac{1}{\theta_i + 2\pi}; \quad i = \{1, 2\} \quad (5)$$

The expressions for the elevation angle θ_i and distance r_i are simply

$$\theta_i = \tan^{-1} \left(\frac{\Delta z_i}{\Delta x_i} \right); \quad r_i = \sqrt{(\Delta x_i)^2 + (\Delta z_i)^2} \quad (6)$$

Signals reflected from next row of buildings have amplitude nearly equal to the diffracted path; therefore, the factor, Q_2 , is simply [15]

$$Q_2 = \frac{1}{\pi k \cos \phi} \frac{1}{r_1} D^2(\theta_1) \quad (7)$$

Further this relation is observed as deep fast fading and Q_2 given by (4) is twice the first term of Q_2 given by (7).

The field reaching the rooftop before the receiver is reduced by the factor, Q_1 . This factor depends on the row spacing, frequency, and according to [15] is determined from

$$Q_1(g_p) \approx 2.35 \left(\frac{z_t - z_b}{R} \sqrt{\frac{d}{\lambda} \cos \phi} \right)^{0.9} \quad (8)$$

Combining the three factors together, (2), (4), (or (7)), and (8), into (3), yields, Q , given by

$$Q = \frac{2.35^{1.8}}{4} \frac{\lambda^{2.1}}{(2\pi)^3} \frac{(z_t - z_b)^{1.8}}{R^{3.8}} \frac{d^{0.9}}{(\cos \phi)^{0.1}} \frac{1}{r_1} D^2(\theta_1) \quad (9)$$

Based on (9), a range index 3.8 is close to what is reported in the literature about the measurements performed in North American cities [8], [15]. The path loss, Q , varies proportionally to the wave length power of 2.1; thus, inverse proportionally to the frequency power of 2.1.

Let R_0 denote the known reference distance which is in the far field of the transmitting antenna (for example 1 km for the macro-outdoor systems) then the total path loss factor, Q , can be written as

$$Q(R_0) = \alpha_1(f) \beta_1(R_0) R_0^{-3.8} \quad (10)$$

where α_1 is a coefficient that depends only on the operating frequency and β_1 is a coefficient that indirectly depends on the distance R_0 .

FIGURE 1: A block diagram of a generic channel.

FIGURE 2: the propagation path between the transmitter and the receiver—taken and modified from [15].

Suppose that for some distance $R > R_0$ the total path loss factor, Q , is given by

$$Q(R) = \alpha_1(f)\beta_1(R)R^{-3.8} = Q(R_0)\left(\frac{R_0}{R}\right)^{3.8}\frac{\beta_1(R)}{\beta_1(R_0)} \quad (11)$$

The above equation is written in dB, $\tilde{Q}(R) = 10\log_{10}Q(R)$,

as follows

$$\tilde{Q}(R) = \tilde{Q}(R_0) + 38\log_{10}\frac{R_0}{R} + 10\log_{10}\frac{\beta_1(R)}{\beta_1(R_0)} \quad (12)$$

Let f_0 denote the reference operation frequency of system. The total power loss factor, Q , (see (9)) can be expressed as

$$Q(f_0) = \delta_1(R)\varepsilon_1(f_0)f_0^{-2.1} \quad (13)$$

Suppose that for some other frequency, $f > f_0$, the total path loss factor, Q , is given by

$$Q(f) = \delta_1(R)\varepsilon_1(f)f^{-2.1} = Q(f_0)\left(\frac{f_0}{f}\right)^{2.1}\frac{\varepsilon_1(f)}{\varepsilon_1(f_0)} \quad (14)$$

Thus, the total power loss factor in dB, $\tilde{Q}(f) = 10\log_{10}Q(f)$, is given by

$$\tilde{Q}(f) = \tilde{Q}(f_0) + 21\log_{10}\frac{f_0}{f} + 10\log_{10}\frac{\varepsilon_1(f)}{\varepsilon_1(f_0)} \quad (15)$$

This concludes the discussion on macro-outdoor geolocation path loss model. Next, we continue the description of the micro-outdoor geolocation path loss model.

4 Micro-outdoor Geolocation Path Loss Model

A micro-outdoor geolocation system consists of a network of transmitters and receivers in which the transmitting antenna has a coverage radius range from 100 m up to 1 km [12]–[15]. Similar to the macro-outdoor geolocation systems, physics of propagation for micro-outdoor geolocation systems can be classified into two groups: (1) path loss and (2) multipath distribution (discussed in a future publication).

The path loss model for the micro-outdoor geolocation systems is different from the path loss model of the macro-outdoor geolocation systems. The discussion presented here is verified with experimental results, which confirms the accuracy of the prediction models.

Consider a simple two-ray model as illustrated in Fig. 3. The receiver receives two rays: the LOS component and the NLOS component. For isotropic antennae the LOS path loss factor is given by [15]

$$Q_{LOS} = \left(\frac{\lambda}{4\pi}\right)\left|\frac{e^{-jkr_1}}{r_1} + \Gamma\frac{e^{-jkr_2}}{r_2}\right|^2 \quad (16)$$

where r_1 and r_2 and the direct and ground reflected ray paths and Γ if the ground reflection coefficient defined as [15]

$$\Gamma = \frac{\cos\theta - a\sqrt{\varepsilon - \sin^2\theta}}{\cos\theta + a\sqrt{\varepsilon - \sin^2\theta}} \quad (17)$$

FIGURE 3: Two ray model showing the Fresnel ellipse about the direct ray—taken and modified from [15].

(a) Footprints of four rectangular multiple story buildings

(b) Two turn routes from the transmitter to the receiver

FIGURE 4: The propagation path between the transmitter and the receiver for rectangular buildings—taken and modified from [15].

where $a = 1$ for horizontal polarization and $a = 1/\epsilon$ for vertical polarization. For typical ground surfaces ϵ is equal to $\epsilon = 15 - j90/f$, where f is the frequency of operation in MHz [15].

It is suggested that the Fresnel radius, R_b , is an important parameter that determines the degree of the logarithmic power slope defined as [15], [19]

$$R_b = \frac{4z_t z_r}{\lambda} \quad (18)$$

The propagation over buildings for low antennae is treated next. The reduction Q_M in the rooftop field at the M^{th} row past the base station due to propagation past the previous rows depends on the signal frequency and path geometry. The last two parameters conspire the dimensionless parameter, g_c , given by [15], [17]

$$g_c = (z_t - z_b) \sqrt{\frac{\cos \phi}{\lambda d}} \quad (19)$$

The reduction, Q_M [15], [17], can be computed from the Boersma functions $I_{n,q}$ [18] given by

$$Q_M = \sqrt{M} \left| \sum_{q=0}^{\infty} \frac{(2g_c \sqrt{j\pi})^q}{q!} I_{M-1,q} \right| \quad (20)$$

where the recursion relation for the Boersma function is [15], [17], and [18]

$$I_{M-1,q} = \frac{(M-1)(q-1)}{2M} I_{M-1,q-2} + \frac{1}{2\sqrt{\pi}M} \sum_{n=1}^{M-2} \frac{I_{M-1,q-1}}{\sqrt{M-1-n}} \quad (21)$$

with initial terms

$$I_{M-1,0} = M^{-3/2} \quad (22)$$

and

$$I_{M-1,1} = \frac{1}{4\sqrt{\pi}} \sum_{n=0}^{M-1} [n(M-n)]^{-3/2} \quad (23)$$

If $g_c = 0$ then $Q_M(g_c = 0) = 1/M$, which is equivalent to $\log[Q_M(g_c = 0)] = -\log M$; the log of Q_M decreases linearly with the log of M . If $z_t < z_b$ then the $\log Q_M$ (i.e., the slope of the curve of $\log Q_M$) decreases initially more rapidly than $\log M$ but it quickly approaches the $\log M$ variation. Conversely, if $z_t > z_b$ then the $\log Q_M$ decreases initially less than $\log M$ but it quickly approaches the $\log M$ variation [15], [17]. Quantitatively, the slope of the curves is given by

$$s = -\frac{\log(Q_{M+1}/Q_M)}{\log(M+1/M)} = \frac{\log(Q_M) - \log(Q_{M+1})}{\log(M+1) - \log(M)} \quad (24)$$

The range index, n , is computed in terms of the slope, s , as follows:

$$n = 2(1 + s) \quad (25)$$

The approach presented here has a limitation because it does not account for crossing streets, which can form a significant fraction of all paths over a small area (see Fig. 4(a)). Also, the present approach does not account for high (or very tall) buildings (see Fig. 4(b)), for which, the propagation is not carried out over the buildings but through the streets and around

the corners [15].

It is found that the signal power level decreases by about 20 dB when the signal propagation path turns a corner. Thus, the signal received by receivers on streets that cross the 2nd street on which the transmitter is located, such as RX1 is essentially due to signals that make a single turn off the 2nd street, as indicated by route 1. More turns are required to reach locations on streets parallel to the 2nd street, such as RX2, RX3, and RX4 which are all located in 3rd street [15].

Each one turn route is composed of an infinite number of two-dimensional (2D) ray paths that make m reflections at the buildings on the 2nd street followed by n reflections at the buildings on the 3rd avenue (RmDRn rays). Rays that are multiply diffracted at the corners of one intersection are ignored since they are significantly weaker. Note that each 2D ray is composed of two rays, one of which is reflected from the ground and appears to come from the image of the transmitter in the ground plane [15].

On one hand, the path loss associated with an RmRn ray for vertically polarized antennae is given by [15]

$$Q = |\Gamma(\phi_1)|^{2m} |\Gamma(\phi_2)|^{2n} Q_{LOS} \quad (26)$$

where $\Gamma(\phi_i)$, $\forall i = \{1,2\}$, are the reflection coefficient at the building faces from the 2nd street and 3rd avenue. The factor Q_{LOS} is computed from (16) and r_1 and r_2 are the total unfolded path lengths of the two three-dimensional (3D) rays. On the other hand, the path loss associated with an RmDRn ray is computed from [15]

$$Q = \frac{[|\Gamma(\phi_1)|^{2m} |\Gamma(\phi_2)|^{2n} Q_{LOS} \equiv Q_{mn}] D^2(\psi) (\rho_1 + \rho_2)}{2\pi k \rho_1 \rho_2} \quad (27)$$

where ρ_1 and ρ_2 are the unfolded 2D ray lengths between the diffracting edge and the transmitter or receiver respectively. For an absorbing boundary condition, the diffraction coefficient, $D(\psi)$, is given by (5). Also, the diffraction angel ψ is negative in the illuminated region and positive in the shadow region.

We propose to model the path loss given by (26) and (27) in the form of

$$Q(R) = \alpha_2(f) \beta_2(R) R^{-n} \quad (28)$$

where $\alpha_2(f)$ and $\beta_2(R)$ are different from $\alpha_1(f)$ and $\beta_1(R)$ corresponding to the macro-outdoor systems. This is going to yield the following

$$\tilde{Q}(R) = \tilde{Q}(R_0) + 10n \log_{10} \frac{R_0}{R} + 10 \log_{10} \frac{\beta_2(R)}{\beta_2(R_0)} \quad (29)$$

R_0 is the reference distance equal to 1 km.

Similarly, this model given by (26) and (27) can be transformed as follows

$$Q(R) = \delta_2(R) \varepsilon_2(f) f^{-m} \quad (30)$$

where $\delta_2(R)$ and $\varepsilon_2(f)$ are different from $\delta_1(R)$ and $\varepsilon_1(f)$ corresponding to the macro-outdoor geolocation systems. This is going to yield the following

$$\tilde{Q}(f) = \tilde{Q}(f_0) + 10m \log_{10} \frac{f_0}{f} + 10 \log_{10} \frac{\varepsilon_2(f)}{\varepsilon_2(f_0)} \quad (31)$$

Note the n and m for micro-outdoor geolocation systems can be different from n and m corresponding to macro-outdoor geolocation systems. For example, based on the measurements performed in five German cities [10] the slope n varies from -2 to -5 .

5 Indoor Geolocation Path Loss Model

An indoor geolocation system may be pictured as having one or several transmitters and receivers in which the transmitter antenna has a coverage radius ranges from 1 m up to 150 m [12]-[15]. As we are now already familiar with, the physics of propagation for indoor geolocation systems can be classified into two groups: (1) path loss model and (2) multipath distribution model [2], [25], [26], [29].

The indoor propagation includes scattering inside rooms, refraction between rooms, penetration between floors, refraction from inside to outside, and refraction from outside to inside. It is suggested that either 2-D or 3-D ray tracing models can be applied to yield accurate indoor propagation predictions [15]. Due to the complexity associated with the analysis for the indoor channel model, we consider only the indoor propagation model due to scattering inside rooms and the most important one. Refractions between rooms, penetration between floors, and refractions from inside to outside and from outside to inside have, to some extent, a similar behavior. Depending on the wall or floor material, thickness, and surface, a propagation loss of 10-15 dB is reported for these types of indoor propagation channel models. The alternative to ray-tracing models is to create a model based on indoor experiments [14], [22], and [24] or on statistical models.

The propagation inside rooms includes the influence of room furniture and ceiling fixtures. Due to the typical construction of modern buildings, the ray incident on the ceiling and on the floor will be scattered rather than reflected inside rooms [15],

which is illustrated in Fig. 5(a). As indicated in Fig. 5(a), when the transmitter and receiver are placed in clear space it is suggested that the propagation can be explained through Fresnel ellipse mechanism [15]. For small distances between the transmitter and receiver the ellipse lies entirely within the clear space, and the presence of scattering will not affect the fields associated with the direct ray; therefore, the path loss will have the $1/R^2$ free-space dependence [15]. As we increase the separation between transmitter and receiver the ellipse becomes larger, and the scattering is contained within the ellipse (see Fig. 5(a)), which produces a path loss greater than that of the free-space [15].

The ellipse first encounters the scattering at a distance h^2/λ . The path loss in excess to the free space is computed at 900 and 1800 MHz for $h = 1.5$ m and is plotted in Fig. 36 of [15] as a function of the transmitter receiver separation R . The path loss is small for small distances up to 40 m and then increases dramatically. According to Bertoni, the theoretical approach finds good agreement with the measurements made in an office building having very large open areas furnished with 1.57-m high cubicles but with no floor to ceiling walls [15].

Ray procedures have been used to account for reflection and transmission at interior and exterior walls, treating the interaction as a specularⁱ process. Use of the specular approach involves two approximations, the first being that the linear extent of the wall is large enough to act as a planar reflector, and that it is electrically smooth so that the scattering does not dominate. For ray paths whose unfolded length is up to 100 m, the maximum width of the Fresnel ellipse in the horizontal plane is less than 4.1 m at 900 MHz and 2.9 m at 1.8 GHz. Since the length of walls is commonly 4 m or more, they span most of the Fresnel zone, or several zones, so that they can reasonably be considered large [15].

Figure 5(b) shows the ray model dealing with wall reflections [15]. Five rays emanated from the transmitter reach the receiver after the transmission through and reflection from the walls. The total path length L is the sum of the path length of specular components. For example, the path loss of the signal associated with the r_1 and r_2 specular components, whose path length equal to $(r_1 + r_2)$, is proportional with the factor $1/(r_1 + r_2)^2$ is addition to the excess path loss of Fig. 36 of [15]. Nevertheless, scattering, associated with segments s_1 and s_2 , exhibits a multiplicative factor of $1/(s_1 s_2)^2$ [15]. Scattering influences the signal in the vicinity of the walls and its amplitude decreases more rapidly with distance than the

amplitude of the reflected rays; therefore, it is neglected [15].

Ray shooting approach or *image theory* are usually suggested to determinate all possible 2D ray paths. In the first approach rays start off from the transmitter at one degree angular interval. Each ray is traced through its interaction with the first wall, where it generates a transmitted and reflected ray [15]. Both rays are then traced to the next interaction, and so on, building a binary tree of rays, which continues through some present number of interactions [15].

It is known that discrete rays have zero probability of intercepting a given point; thus, a circle of finite radius proportional to the ray length is used to represent the receiver [15]. In the second approach the exact ray path between points is determined by imaging the source in the plane of each wall, one at a time, and checking that the plane between the image and the receiver intersects the wall in the physically existing segments. The same process is repeated for double/triple imaging of all combinations of the two/three walls etc [15]. It is reported that triple imaging is adequate for most cases [15].

Assuming isotropic antennae and having found the contributing rays, the path loss associated with the i th ray is computed from

$$Q_i = Q_0 E(L_i) \prod_n |\Gamma_n(\phi_{ni})|^2 \prod_m |\Gamma_m(\phi_{mi})|^2 \quad (32)$$

where L_i is the total unfolded path length of the ray and Q_0 is the free-space factor given by (2). The coefficient, $E(L_i)$, includes the excess path loss and the coefficients, $\Gamma_n(\phi_{ni})$ and $\Gamma_m(\phi_{mi})$, are the reflection and refraction coefficients at the walls that the ray encounters [15].

The path loss associated with segments, $L_0, L_1, \dots, L_M$ separated by diffraction at edges through angles, $\beta_0, \beta_1, \dots, \beta_M$ provided that each edge lies outside the shadow boundary of the previous edge is determined from

$$Q_i = Q_0 L L_0^{-1} \left[\prod_j \frac{D^2(\beta_j)}{2\pi k L_j} \equiv \frac{1}{(2\pi k)^M} \prod_j \frac{D^2(\beta_j)}{L_j} \right] \quad (33)$$

where L is the total path length; i.e., $L = \sum_j L_j$, Q_0 is the free-space factor given by (2) for a distance L , and $D(\beta_j)$ is the diffraction coefficient given by (5).

We propose to model the total path loss given by (33) in the form of

$$Q(R) = \alpha_3(f) \beta_3(R) R^{-n} \quad (34)$$

where $\alpha_3(f)$ and $\beta_3(R)$ are different from $\alpha_1(f)$ and $\beta_1(R)$ corresponding to the macro-outdoor geolocation systems. This is going to yield the following

(a) The Fresnel zone for propagation between the TX and RX in clear space between building furnishings and ceiling fixtures.

(b) The two dimensional ray-tracing accounting for specular reflection and transmission at walls.

FIGURE 5: Four GPS satellite CC function (top) vs. ML function (bottom).

$$\tilde{Q}(R) = \tilde{Q}(R_0) + 10n \log_{10} \frac{R_0}{R} + 10 \log_{10} \frac{\beta_3(R)}{\beta_3(R_0)} \quad (35)$$

R_0 is the reference distance equal to 1 m.

Similarly, the indoor propagation model given by (33) can be transformed in the form of

$$Q(R) = \delta_3(R) \varepsilon_3(f) f^{-m} \quad (36)$$

where $\delta_3(R)$ and $\varepsilon_3(f)$ are different from $\delta_1(R)$ and $\varepsilon_1(f)$ corresponding to the macro-outdoor geolocation systems. This is going to yield the following

$$\tilde{Q}(f) = \tilde{Q}(f_0) + 10m \log_{10} \frac{f_0}{f} + 10 \log_{10} \frac{\varepsilon_3(f)}{\varepsilon_3(f_0)} \quad (37)$$

Note the n and m for indoor geolocation systems can be different from n and m corresponding to macro-outdoor geolocation systems. For example based on the measurements summarized in [10] the slope n varies from -2 to -4 .

6 Unified Path Loss Geolocation Channel Model

Although a unified channel model has not yet been found, in this section we make a first attempt to present a unified channel model which consists of a unified path loss model and a unified multipath distribution model (which will be discussed in a future publication).

The unified path loss model consists of an approach for linking together the path loss models of the three geolocation systems (macro-outdoor, micro-outdoor, and indoor) with the distance between the transmitter and receiver, R , and the frequency of operation, f . Although there are several parameters that affect the power loss factor, we consider R and f as the most important parameters for two reasons. Most of the geolocation systems discussed in [25] are based on a direct measure of the time of travel; i.e., distance between the transmitter and receiver. The frequency of these systems varies; hence, the path loss factor varies as a function of the frequency as well.

Equations (12), (29), and (35) can be generalized as

$$\tilde{Q}(R) = \tilde{Q}(R_0) + 10n \log_{10} \frac{R_0}{R} + 10 \log_{10} \frac{\beta_i(R)}{\beta_i(R_0)} \quad (38)$$

where R_0 is the reference distance equal to 1 km (for macro-outdoor geolocation systems), 100 m (for micro-outdoor geolocation systems), and 1 m (for indoor geolocation systems). Also, β_i is equal to β_1 (for macro-outdoor systems), β_2 (for micro-outdoor systems), and β_3 (for indoor geolocation systems).

The term $10 \log_{10} \beta_i(R)/\beta_i(R_0)$ is very complicated and not very easy to be assessed. Nevertheless, it is proposed that a zero mean Gaussian noise model fit reasonably well [10]

$$10 \log_{10} \beta_i(R)/\beta_i(R_0) \cong \chi_\sigma \quad (39)$$

where χ_σ denotes a zero mean Gaussian random variable that reflects the variations in average received power that naturally occurs as well as indirect dependence of the distance. The standard deviation σ plays an important role on the accuracy of the propagation loss model [10].

Equation (38) is generalized by substituting (39) into (38) as [10]

$$\tilde{Q}(R) = \tilde{Q}(R_0) + 10n \log_{10} \frac{R_0}{R} + \chi_\sigma \quad (40)$$

where n denotes the power law relationship between distance and received power.

The reader is reminded that (40) shows the dependency between the power loss factor (in dB) and the distance between the transmitting antenna and the receiving antenna. Another important parameter that affects the power loss factor is the frequency of operation. Unfortunately, the dependency of the power loss factor on the frequency of operation was overlooked in the literature until 2003 [25].

Similarly, (15), (31), and (37) can be generalized as

$$\tilde{Q}(f) = \tilde{Q}(f_0) + 10m \log_{10} \frac{f_0}{f} + 10 \log_{10} \frac{\varepsilon_i(f)}{\varepsilon_i(f_0)} \quad (41)$$

where f_0 is the reference frequency, ε_i is equal to ε_1 (for macro-outdoor systems), ε_2 (for micro-outdoor systems), and ε_3 (for indoor geolocation systems).

The term $10 \log_{10} \varepsilon_i(f)/\varepsilon_i(f_0)$ is very complicated and not very easy to be assessed. Nevertheless, we model this term as a zero mean Gaussian process; hence, we have

$$10 \log_{10} \frac{\varepsilon_i(f)}{\varepsilon_i(f_0)} \cong \gamma_\sigma \quad (42)$$

where γ_σ denotes a zero mean Gaussian random variable that reflects the variations in average received power that naturally occur and indirect dependence of the frequency. Similarly, we anticipate that the standard deviation σ will greatly influence the accuracy of the prediction of the path loss propagation model.

We can generalize (41) by substituting (42) in into (41) as follows

$$\tilde{Q}(f) = \tilde{Q}(f_0) + 10m \log_{10} \frac{f_0}{f} + \gamma_\sigma \quad (43)$$

where m denotes the power law relationship between the frequency and the power loss factor.

Sometimes we desire to have the path loss model be a function of both f and R . In this case, (40) and (43) are combined as follows:

$$\tilde{Q}(R, f) = \tilde{Q}(R_0, f_0) + 10n \log_{10} \frac{R_0}{R} + 10m \log_{10} \frac{f_0}{f} + \eta_\sigma \quad (44)$$

where $\tilde{Q}(R_0, f_0)$ is the path loss corresponding to the frequency f_0 and distance R_0 and η_σ is a normal random variable. Clearly, if $R = R_0$ then (44) is the same as (42) as long as η_σ is the same as χ_σ . On the other hand if $f = f_0$ then (44) is the same as (43) as long as η_σ is the same as γ_σ . So then in general η_σ can be model as a sum of χ_σ and γ_σ . And since χ_σ and γ_σ are zero mean random variables then η_σ is also zero mean and with variance the sum of the variance of χ_σ and γ_σ .

7 Simulation Results

To give an intuitive explanation of the free space path loss we have considered the following example. Suppose that operating carrier frequency of the TX-RX is taken from the set $f = \{0.9, 1.17642, 1.2276, 1.57542, 20\}$ GHz. Suppose that the relative distance between the TX-RX is taken from the set $R = \{1, 2, \dots, 20\}$ km.

Based on these assumptions we have computed the free space path loss Q_0 (dB) (see (2)) versus $R = \{1, 2, \dots, 20\}$ km and the result is plotted in Fig. 6(a). As shown from Fig. 6(a) free space path losses can be as high as -91 dB ($R = 1$ km and $f = 900$ MHz) and as low as -145 dB ($R = 20$ km and $f = 20$ GHz). Suppose that the distance from the transmitter to the receiver is taken from the set $R = \{1, 5, 10, 15, 20\}$ km and that the operation frequency changes from $f = \{1, 2, \dots, 20\}$ GHz. Based on these assumptions we have computed the free space path loss Q_0 (dB) (see (2)) versus $f = \{1, 2, \dots, 20\}$ GHz and the result is plotted in Fig. 6(b). Similar to Fig. 6(a) Q_0 (dB) can be as high as -92 dB and as low as -145 dB.

7.1 Macro-outdoor Geolocation Path Loss Model Simulations

In order to illustrate the total path loss, Q , based on the macro-outdoor geolocation path loss model, we have considered the following setup. Suppose that (the azimuth angle) $\phi = 60^\circ$, (the consecutive building distance) $d = 4$ m, (the transmitter height) $z_t = 20$ m, (the building height) $z_b = 12.5$ m, (the antenna height) $z_a = 1$ and $f = \{0.9, 1.17642, 1.2276, 1.57542, 20\}$ GHz. Assuming that $R = \{1, 2, \dots, 20\}$ km we compute the path loss values Q for every frequency f as indicated in Fig. 7(a).

Next, assuming that frequency changes from $f = \{1, 2, \dots, 20\}$ GHz we compute the values of Q for every $R = \{1, 5, 10, 15, 20\}$ km as shown in Fig. 7(b).

7.2 Micro-outdoor Geolocation Path Loss Model Simulations

In order to assess the LOS path loss, Q_{LOS} (dB), as a function of the frequency and distance between the transmitting antenna and receiving antenna, R , we consider two examples in which the height of the transmitting/receiving antenna is 20/1.5 m.

(a) Free space path loss Q_0 (dB) model vs. R (km). (b) Free space path loss Q_0 (dB) vs. f (GHz).FIGURE 6: Free space path loss Q_0 (dB) (a) vs. R (km) and (b) vs. f (GHz) for a macro-outdoor geolocation system. (a) Total path loss Q (dB) vs. R (km) (b) Total path loss Q (dB) vs. f (GHz)FIGURE 7: Total path loss Q (dB) vs. (a) R (km) and (b) f (GHz) for a macro-outdoor geolocation system. (a) Total path loss Q (dB) vs. R (km) horizontal polarization (b) Total path loss Q (dB) vs. R (km) vertical polarizationFIGURE 8: Total path loss Q (dB) vs. (a) R (km) and (b) f (GHz) for a micro-outdoor geolocation system.

(a) Horizontal polarization

(b) Vertical polarization

FIGURE 9: Total path loss Q (dB) vs. f (GHz) (a) horizontal and (b) vertical for micro-outdoor geolocation systems. (b) Unified path loss model for a micro-outdoor geolocation system, $\sigma = 2$. (c) Unified path loss model for an IGS, $\sigma = 1$.FIGURE 10: Total path loss Q (dB) vs. R (m) (a) macro-outdoor, (b) micro-outdoor, and (c) indoor geolocation systems and $n = \{2, 3, 4, 5, 6\}$. (a) Unified path loss model for a macro-outdoor geolocation system, $\sigma = 0.5$. (a) Unified path loss model for a macro-outdoor geolocation system, $\sigma = 0.5$.

(b) Unified path loss model for a micro-outdoor geolocation system, $\sigma = 2$.

(c) Unified path loss model for an indoor geolocation system, $\sigma = 1$.

FIGURE 11: Total path loss Q (dB) vs. f (GHz) (a) macro-outdoor, (b) micro-outdoor, and (c) indoor geolocation systems and $m = \{2, 2.1, 2.2, 2.3, 2.4\}$.

Figures 8(a) and 8(b) illustrate the LOS path loss, Q_{LOS} (dB), versus the distance between the transmitter and the receiver, R , for every $f = \{0.9, 1.17642, 1.2276, 1.57542, 10\}$ GHz, for horizontal polarization (a) and vertical polarization (b). The distance R changes from 10 m to 1000 m. The LOS path loss at 900 MHz for $R = 10$ m is about $-47/-51$ dB and for $R = 1000$ m is about $-92/-89$ dB for horizontal/vertical polarization. The path loss corresponding to L_5 is almost the same as the path loss corresponding to L_2 and somewhat different from the one corresponding to the L_1 frequency. The LOS path loss at 10 GHz for $R = 10$ m is about $-80/-69$ dB and for $R = 1000$ m is about $-120/-108$ dB for horizontal/vertical polarization.

The LOS path loss numbers for R between 10 and 100 m are

comparable with the numbers provided in Fig. 20 of [15]. The LOS path loss at 900 MHz is about 20/15 dB higher than the path loss at 10 GHz for horizontal/vertical polarization. The LOS path loss drops somewhat faster for horizontal polarization than for vertical polarization; thus, vertical polarization would be a better choice than horizontal polarization.

Figures 9(a) and 9(b) illustrate the LOS path loss, Q_{LOS} (dB), versus the frequency, f , for micro-outdoor geolocation systems for every $R = \{10, 250, 500, 750, 1000\}$ m, for horizontal polarization (a) and vertical polarization (b). The frequency f changes from 1 to 10 GHz. The LOS path loss for $R = 10$ m and $f = 1$ GHz is about -55 dB and for $f = 10$ GHz is about -75 dB for either horizontal or vertical polarization.

The LOS path loss for $R = 250$ m and $f = 1$ GHz is about $-80/-78$ dB and for $f = 10$ GHz is about $-108/-96$ dB for horizontal/vertical polarization. The LOS path loss for $R = 500$ m and $f = 1$ GHz is about $-83/-90$ dB and for $f = 10$ GHz is about $-113/-102$ dB for horizontal/vertical polarization. The LOS path loss for $R = 750$ m and $f = 1$ GHz is about $-88/90$ dB and for $f = 10$ GHz is about $-107/-111$ dB for horizontal/vertical polarization. The LOS path loss for $R = 1000$ m and $f = 1$ GHz is about $-92/-90$ dB and for $f = 10$ GHz is about $-120/-108$ dB for horizontal/vertical polarization. The LOS path loss at 10 m is about 40 dB higher than the path loss at 1000 m for either horizontal or vertical polarization.

The minimums and maximums of the LOS path loss depend on the distance as well as the frequency. The LOS path loss drops somewhat faster for horizontal polarization than for vertical polarization. Henceforth, vertical polarization would be a better choice than horizontal polarization.

Based on the data that is shown in Fig. 8, we find the logarithmic power slope of the Q_{LOS}

$$\tilde{Q}_{LOS}(R = 10\text{m}) = \tilde{Q}(R_0) + 10n = -50 \text{ dB} \quad (45)$$

$$\tilde{Q}_{LOS}(R = 100\text{m}) = \tilde{Q}(R_0) + 20n = -70 \text{ dB} \quad (46)$$

Solving the above system produces, $\tilde{Q}(R_0) = -30$ dB and $n = -2$; hence, the logarithmic power slope for the $\tilde{Q}_{LOS}$ curves shown in Fig. 8 is -2 . Similarly, we can find that the logarithmic power slope for the curves shown in Fig. 9 is $m = -2$.

The Fresnel distances for the first experiment are $R_b = \{360, 470, 491, 630, 4000\}$ m which correspond to the frequencies $f = \{0.9, 1.17642, 1.2276, 1.57542, 10\}$ GHz. As shown in Figs. 8 (a) and (b) the logarithmic power slope is

near 2 for $x_r < R_b$ or equal to 2 for $x_r > R_b$.

For a given transmitting/receiving antenna height of 13.4/1.6 m, measuring the normalized signal strength (dB) at 800-MHz frequency for a transmitter receiver distance varying from 1 to 1000 m on a logarithmic scale produces a regression line slope less than 2 for $x_r < R_b$ and near 4 for $x_r > R_b$ [16], [20]. Nevertheless, the effect of traffic on all the ground-reflected rays is not fully understood. Measurements made in Manhattan and two experiments shown in Figs. 8 and 9 show a logarithmic power slope is near 2 at distances beyond, R_b [21]. Therefore, this is currently a controversial issue to be resolved in the future.

7.3 Unified Path Loss Geolocation Channel Model Simulations

In order to illustrate the unified path loss model given by (40) we consider the following example.

Figure 10(a) depicts the unified path loss, Q , versus the relative distance between the transmitter and the receiver, R , going from 1 km to 20 km and for $n = \{2, 3, \dots, 6\}$.

In the first example we consider a macro-outdoor geolocation system with the following parameters $\tilde{Q}(R_0) = -140$ dB, $R_0 = 1$ km, and $\sigma = 0.5$.

In the second example we consider a micro-outdoor geolocation system with the following parameters $\tilde{Q}(R_0) = -70$, $R_0 = 100$ m, and $\sigma = 2.5$. Figure 10(b) depicts the unified path loss, Q , versus relative the distance between the transmitter and the receiver, R , going from 100 m to 1 km and for $n = \{2, 3, \dots, 6\}$.

In the third example we consider an indoor geolocation system with the following parameters $\tilde{Q}(R_0) = 0$ dB, $R_0 = 1$ m, and $\sigma = 1$. Figure 10(c) shows the unified path loss, Q , versus relative the distance between the transmitter and the receiver, R , going from 1 m to 100 m and for $n = \{2, 3, \dots, 6\}$.

In order to illustrate the unified path loss model given by (43) we consider the following example.

In the first example we consider a macro-outdoor geolocation system with the following parameters $\tilde{Q}(f_0) = -140$, $f_0 = 1$ GHz, and $\sigma = 0.5$. Figure 11(a) depicts the unified path loss, Q , versus the frequency, f , going from 1 GHz to 10 GHz and for $m = \{2, 2.1, \dots, 2.4\}$.

In the second example we consider a micro-outdoor geolocation system with the following parameters $\tilde{Q}(f_0) = -70$, $f_0 = 1$ GHz, and $\sigma = 2.5$. Figure 11(b) depicts the unified path loss, Q , versus the frequency, f , going from 1

GHz to 10 GHz and for $m = \{2, 2.1, \dots, 2.4\}$.

In the third example we consider an indoor geolocation system with the following parameters $\tilde{Q}(f_0) = 0$ dB, $f_0 = 1$ GHz, and $\sigma = 1$. Figure 11(c) depicts the unified path loss, Q , versus the frequency, f , going from 1 GHz to 10 GHz and for $m = \{2, 2.1, \dots, 2.4\}$.

8 Conclusions

As a summary, we have provided a rather simple way to model the path loss of the communication channel of an indoor geolocation system as a function of the transmitter and receiver distance, R , and operating frequency, f . Some of the models presented in this paper are validated with experimental results reposted in literature. There is, however, a need to perform more measurements and validate and improve our models. The second part of the unified communications channel model, in which will discuss the unified multi-path model was presented in Proгри 2006 [2].

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10 References

- [1] I. Proгри, "A unified geolocation channel model—part I (path loss)," in *Proc. ION-NTM 2005*, San Diego, CA, pp. 1148-1161, Jan. 2005, http://giftet.com/Proгри/Proгри_2005_01_01_ION_NTM.pdf.
- [2] I. Proгри, "A unified geolocation channel model—part II (multipath distribution)," in *Proc. WTS-2006*, Pomona, CA, pp. 1-8, Apr. 2006, DOI: <http://doi.org/10.1109/WTS.2006.334542>, URL: http://giftet.com/Proгри/Proгри_2006_04_01_IEEE_WTS.pdf.
- [3] D.E. Kerr, *Propagation of Short Radio Waves*, 1st ed. New York, NY: McGraw-Hill, 1951.
- [4] R.S. Kennedy, *Fading Dispersive Communication Channel*. New York, NY: Wiley-Interscience, 1969.
- [5] J.G. Proakis, *Digital Communications*. New York: McGraw-Hill, 1983.

- [6] J.D. Parson, *The Mobile Radio Propagation Channel*. New York: Halsted Press, 1992.
- [7] K. Siwiak, *Radio Wave Propagation and Antennas for Personal Communications*. New York: Artech House, 1995.
- [8] G.D. Ott, A. Plitkins, "Urban path-loss characteristics at 820 MHz," *IEEE Trans. Veh. Technol.*, vol. VT-27, pp. 189-197, 1978, DOI: <https://doi.org/10.1109/T-VT.1978.23748>.
- [9] T.A. Sexton, K. Pahlavan, "Channel modeling and adaptive equalization of indoor radio channels," *IEEE J. Select. Areas Commun.*, vol. SAC-5, pp. 128-137, Feb. 1987, DOI: <https://doi.org/10.1109/49.16852>.
- [10] T.S. Rappaport, "Characterization of UHF multipath radio channels in factory buildings," *IEEE Trans. Ant. Prop.*, vol. 37, no. 8, pp. 1058-1069, Aug. 1989, DOI: <https://doi.org/10.1109/8.34144>.
- [11] S.J. Howard, K. Pahlavan, "Autoregressive modeling of wide-band indoor radio propagation," *IEEE Trans. Comm.*, vol. 40, no. 9, pp. 1540-1552, Sep. 1992, DOI: <https://doi.org/10.1109/26.163575>.
- [12] L.J. Greenstein, N. Amitay, Ta-Shing Chu, L.J. Cimini, G.J. Foschini, M.J. Gans, I. Chih-Lin, A.J. Rustako, R.A. Valenzuela, G. Vannucci, "Microcells in personal communications systems," *IEEE Comm. Mag.*, vol. 30, no. 12, pp. 76-88, Dec. 1992, DOI: <https://doi.org/10.1109/35.210359>.
- [13] J.B. Andersen, T.S. Rappaport, S. Yoshida, "Propagation measurements and models for wireless communications channels," *IEEE Comm. Mag.*, vol. 33, no. 1, pp. 42-49, Jan. 1995, DOI: <https://doi.org/10.1109/35.339880>.
- [14] K. Pahlavan, P. Krishnamurthy, A. Beneat, "Wideband radio propagation modeling for indoor geolocation applications," *IEEE Comm. Mag.*, vol. 36, no. 4, pp. 60-65, Apr. 1998, DOI: <https://doi.org/10.1109/35.667414>.
- [15] H.L. Bertoni, W. Honcharenko, L.R. Macel, H.H. Xia, "UHF propagation prediction for wireless personal communications," in *Proc. IEEE*, vol. 82, no. 9, pp. 1333-1359, Sep. 1994, DOI: <https://doi.org/10.1109/5.317081>.
- [16] L.R. Macel, H.L. Bertoni, H.H. Xia, "Unified approach to prediction of propagation over building of all ranges of base station antenna height," *IEEE Trans. Veh. Technol.*, vol. 42, pp. 41-45, Feb. 1993, DOI: <https://doi.org/10.1109/25.192385>.
- [17] H.H. Xia, H.L. Bertoni, "Diffraction of cylindrical and plane waves by an array of absorbing half screens," *IEEE Trans. Ant. Prop.*, vol. 40, pp. 170-177, Feb. 1992.
- [18] J. Boersma, "On certain multiple integrals occurring in wave-guide scattering problem," *SIAM J. Math. Anal.*, vol. 9, no. 2, pp. 377-393, 1978.
- [19] T. Iwama, E. Moriyama, H. Ryuko, S. Sekizawa, T. Sarauwatari, "Investigation of propagation characteristics above 1 GHz for microcell land mobile radio," in *Proc. IEEE Veh. Technol. Conf.*, pp. 396-400, 1990, DOI: <https://doi.org/10.1109/VETEC.1990.110354>.
- [20] D.M.J. Devasirvatham, C. Banerjee, R.R. Murray, D.A. Rappaport, "Two frequency radio-wave propagation measurements in Brooklyn," in *Proc. IEEE Int. Conf. Univ. Pers. Commun.*, pp. 23-27, Dallas, TX, 1992, DOI: <https://doi.org/10.1109/ICUPC.1992.240827>.
- [21] A.J. Rustako, Jr., N. Amitay, G.J. Owens, R.S. Roman, "Radio propagation at microwave frequencies for line-of-sight microcellular mobile and personal communications," *IEEE Trans. Veh. Technol.*, vol. 40, pp. 203-210, 1991, DOI: <https://doi.org/10.1109/25.69989>.
- [22] T.S. Rappaport, D.A. Hawbaker, "Wide-band microwave propagation parameters using circular and linear polarized antennas for indoor wireless channels," *IEEE Trans. Comm.*, vol. 40, no. 2, pp. 240-245, Feb. 1992, DOI: <https://doi.org/10.1109/26.129185>.
- [23] T. Lo, J. Litva, H. Leung, "A new approach for estimating indoor radio propagation characteristics," *IEEE Trans. Ant. Prop.*, vol. 42, no. 10, pp. 1369-1376, Oct. 1994, DOI: <https://doi.org/10.1109/8.320742>.
- [24] T.K., Blankenship, T.S. Rappaport, "Characteristics of impulsive noise in the 450-MHz band in hospitals and clinics," *IEEE Trans. Ant. Prop.*, vol. 46, no. 2, pp. 194-203, Feb. 1998, DOI: <https://doi.org/10.1109/8.660963>.
- [25] I.F. Proгри, "An assessment of indoor geolocation systems," Ph.D. Dissertation, Worcester Polytechnic Institute, May 2003, URL: http://giftet.com/Proгри/Proгри_2003_05_Ph.D._Dissertation.pdf.
- [26] I. Proгри, *Indoor Geolocation Systems—Theory and Applications. Vol. I*, 1st ed., Worcester, MA: Giftet Inc., ~800 pp., ~2017 (not yet available in print).
- [27] I. Proгри, W.R. Michalson, "An alternative approach to multipath and near-far problem for indoor geolocation systems," in *Proc. 14th Inter. Tech. Mtg. Sat. Div. ION*

(*ION GPS 2001*), Salt Lake City, UT, pp. 1434-1443, Sep. 2001.

- [28] I. Progri, W.R. Michalson, J. Wang, M.C. Bromberg, R.J. Duckworth, "Requirements of a C-CDMA pseudolite indoor geolocation system," *J. Geol. Geoinfo. Geointel.*, vol. 2017, article ID 2017071605, 5 pg., Nov. 2017. DOI: <http://doi.org/10.18610/JG3.2017.071605>, <http://www.giftet.com/JG3/2017/071605.pdf>.

- [29] I. Progri, W.R. Michalson, J. Wang, M.C. Bromberg, "Indoor geolocation using FCDMA pseudolites: signal structure and performance analysis," *Navigation*, vol. 54, no. 3, pp. 242-256, fall 2007, DOI: <http://doi.org/10.1002/j.2161-4296.2007.tb00407.x>, http://giftet.com/Progr/Progr_2007_fall_Navigation.pdf.

ⁱ Relating to or having the properties of a mirror.