A Vectorial Analysis of UHF Propagation in Urban and Rural Environments

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ABSTRACT

In this work we present a vectorial analysis of the two-ray model, as well as a vectorial analysis of the whole street waveguide model in the UHF band. For both, an ideal dipole is assumed to be the transmitting aerial. For the analysis of the two-ray model, parallel polarization is investigated and a comparison between the approximated analysis (only algebraic sum of the fields) and non-approximated analysis (considering dipole's gain and vectorial analysis) is provided as well as an analysis of the error versus \( r \) distance. A comparison between vertical and horizontal components of the electric field is also carried out. Regarding the street model, the street is modeled as a multislit waveguide, and horizontal and vertical polarizations are investigated. As in the two-ray model, a comparison between the approximated and non-approximated analysis is provided as well as a comparison between the total field and the field due to lateral reflections only. Still for the street model, slits are inserted and comparisons and analyses are made. Comparisons provided show that a vectorial and gain analysis can be a very useful tool toward increasing theoretical analysis in spite of empirical analysis in the field of prediction.

Keywords: Urban propagation, channel characterization, street waveguide, ray-tracing.

1. INTRODUCTION

Mobile communication systems have a fundamental importance in areas like trade, business and public safety in many parts in the world [1]. Since the early 1990’s there has been an expressive increase in the mobile communications market [2], [3]. All these changes in the telecommunication scenario leave companies having to face up to a growing complexity of radio networks structure and design [4], [13], [15].

For frequencies in the UHF band (300 MHz - 3 GHz) the wavelength is small compared to geometrical dimensions of the scatterers (buildings and other structures along the streets), which makes ray optics methods convenient for analyzing or describing the behavior of electric fields in urban areas. The rays and the fields associated with them can be traced and evaluated, respectively, according to the characteristics of the medium. These considerations make the UHF band very attractive regarding the provision of mobile services making it scarce and expensive [2], [5], [6]. For a more effective utilization of this scarce spectrum, it is necessary to have at ones disposal accurate propagation models so that signal behavior can be predicted properly within urban and suburban environments [7]. In order to create a communication system model, mathematical descriptions of the transmitter, receptor and how the environment responds to the transmitted signal need to be determined. Once these descriptions are congregated into a theoretical model, it can be used to analyze the performance of a practical implementation before in fact implementing it [8], [13], [15]. Suggestions for personal communication services (PCS) have been based on the utilization of microcells with base station antennas in light posts, buildings and other structures along the streets, covering distances of at maximum 1 Km [6], [13], [15]. In urban environments it is usual to have buildings placed on both sides of the streets constituting a waveguide channel. For receiving and transmitting antennas below the rooftops there will be, besides direct and ground reflection, the presence of multiple lateral reflections on the faces of the buildings [6], [9]. This is described by MAZAR AND BRONSHTEIN [9], who model city streets as random multislit waveguides. For developing a street waveguide model the features encountered by rays, such as ground and buildings surface materials, should be taken into account. When dealing with reflection, the electric field of the incident wave is changed in intensity and phase by the complex reflection coefficient, which depends on these surface materials. For multiple lateral reflections on building faces, the resulting electric field is proportional to the power of the complex reflection coefficient of those surfaces [6], [9]-[10].

The purpose of this paper is to derive, through a vectorial and gain analysis, a two-ray model and a street waveguide model with urban characteristics as well as their variations and comparisons. Within this context, vectorial and gain analyses are compared to non-vectorial and gain analyses, i.e., approximated analyses with only an algebraic sum of the fields at the point of reception, as those described in [9] and [16].

In Section 1, we present the motivation, social rel-
evance, main objectives and the description of the work stages. In Section 2, we present the concept of the reflection coefficient in terms of parallel and perpendicular polarizations. In Section 3 we depict the two-ray model and perform a comparison between the approximated analysis (only algebraic sum of the fields) and non-approximated analysis (considering dipole's gain and vectorial analysis), and an error analysis in function of the r distance with regard to both models mentioned above is also provided. Still in this section, a time variation of the instantaneous electric field vector is also investigated. In Section 4, we depict a street waveguide regarding direct incidence, ground reflection and lateral reflections on the waveguide borders. Horizontal and vertical polarizations are analyzed. We also carry out a comparison between the main components of the electric field based on lateral reflections only and the total electric field. As in Section 2, we perform a comparison between approximated and non-approximated analysis. Finally, we compare the main components of the total electric field for different slit lengths and analyze the parameters involved.

2. REFLECTION COEFFICIENT

The complex reflection coefficient can be calculated only when the following parameters are known: the conductivity and relative permittivity of the reflecting surface, the incidence angle (angle that the incident ray makes with the reflecting surface), and the wavelength and polarization of the incident radiation [10].

2.1 Parallel Polarization

Parallel polarization occurs when the electric field is oriented parallel to the plane of incidence. The complex reflection coefficient for parallel polarization is given by [10],[11]:

\[ R_\parallel = \frac{\varepsilon_\perp \sin\alpha - \sqrt{\varepsilon_\perp \cos^2\alpha}}{\varepsilon_\perp \sin\alpha + \sqrt{\varepsilon_\perp \cos^2\alpha}} \]  \hspace{1cm} (1)

where the relative complex permittivity is given by:

\[ \varepsilon_\perp = \varepsilon - j60\lambda \sigma \]  \hspace{1cm} (2)

and where \( \sigma \) and \( \varepsilon \) are conductivity and relative dielectric constant of the reflecting surface, respectively, \( \lambda \) is the wavelength of the incident radiation and \( \alpha \) is the incidence angle.

2.2 Perpendicular Polarization

Perpendicular polarization occurs when the electric field is oriented perpendicular to the plane of incidence. The complex reflection coefficient for perpendicular polarization is given by [10], [11]:

\[ R_\perp = \frac{\sin\alpha - \sqrt{\varepsilon_\perp \cos^2\alpha}}{\sin\alpha + \sqrt{\varepsilon_\perp \cos^2\alpha}} \]  \hspace{1cm} (3)

where \( \varepsilon_\perp \) is given in Equation 2.

3. TWO-RAY MODEL

3.1 Using Approximations for Parallel Polarization

Considering \( h_A << r \) and \( h_B << r \) in Fig. 1, and therefore \( \alpha \) and \( \varphi \) very small, the resulting (total) electric field (\( E_T \)) in B will be given by an algebraic sum of (4) and (5) as follows:

\[ \hat{E}_1 = \frac{E_0}{r_1} e^{j\omega t} e^{-j\beta r_1}, \text{ due to ray 1} \]  \hspace{1cm} (4)

\[ \hat{E}_2 = \hat{R}_\parallel \frac{E_0}{r_2} e^{j\omega t} e^{-j\beta r_2}, \text{ due to ray 2} \]  \hspace{1cm} (5)

\[ r_1 = \sqrt{r^2 + (h_A - h_B)^2} \]  \hspace{1cm} (6)

where \( E_0 \) is a constant and \( \beta = 2\pi/\lambda \).

\( \sigma \), \( \varepsilon \) and \( \lambda \) are generally known, while \( \alpha \) has to be calculated according to the following parameters: \( h_A \), \( h_B \) and \( r \) [10]. According to Fig. 1:

\[ \tan\alpha = \frac{h_A + h_B}{r} \Rightarrow \alpha = \arctan\left(\frac{h_A + h_B}{r}\right) \]  \hspace{1cm} (7)

3.2 Without Approximations for Parallel Polarization (Ideal Dipole)

Now a more detailed model for parallel polarization is considered as illustrated in Fig. 2, where an ideal dipole with height \( h_A \) generates two electric field vectors (\( \hat{E}_1 \) and \( \hat{E}_2 \)) at the point of reception with height \( h_B \).

From Fig. 2 it can be seen that:

\[ \alpha = \arctan\left(\frac{h_A + h_B}{r}\right) \]  \hspace{1cm} (8)

\[ \varphi = \alpha - \psi \]  \hspace{1cm} (9)

The resulting (total) electric field vector at the point of reception is given by:

\[ \hat{E}_T = \hat{E}_1 + \hat{E}_2 \]  \hspace{1cm} (10)

The field vectors generated by rays 1 and 2 are given respectively by:

\[ \hat{E}_1 = \frac{E_0 e^{j\omega t} e^{-j\beta r_1}}{r_1} G(\theta) \hat{a}_\theta \]  \hspace{1cm} (11)

\[ \hat{E}_2 = \hat{R}_\parallel \frac{E_0 e^{j\omega t} e^{-j\beta r_2}}{r_2} G(\theta) \hat{a}_\theta \]  \hspace{1cm} (12)
where $G(\theta) = \sin\theta$ is the gain for an ideal dipole, and $r_1$ and $r_2$ are given by (6).

The total electric field vector from (10) has components either on vertical (z) and horizontal (y) directions. The vertical and horizontal components of this field are given respectively by:

$$\vec{E}_z = (\dot{E}_1 \cos\psi + \dot{E}_2 \cos\gamma) \vec{a}_z \quad (13)$$

$$\vec{E}_y = (\dot{E}_1 \sin\psi - \dot{E}_2 \sin\gamma) \vec{a}_y \quad (14)$$

where $\psi$ and $\gamma$ are the angles that vectors $\vec{E}_1$ and $\vec{E}_2$ make with vertical direction, respectively, and $\vec{E}_1$ and $\vec{E}_2$ are their respective magnitudes in the phasor form.

Thus we have,

$$\vec{E}_T = \dot{E}_y \vec{a}_y + \dot{E}_z \vec{a}_z \quad (15)$$

where $\dot{E}_y = E_y e^{j\theta_y}$ and $\dot{E}_z = E_z e^{j\theta_z}$.

Taking $\theta_y$ as reference,

$$\vec{E}_T = E_y \vec{a}_y + E_z e^{j\delta} \vec{a}_z \quad (16)$$

where $\delta = \theta_z - \theta_y$ and $180^\circ \leq \delta \leq 180^\circ$.

(16) is the phasor form of the instantaneous electric field given below:

$$\dot{E}_y(t) = E_y(t) \vec{a}_y + E_z(t) \vec{a}_z \quad (17)$$

$$\dot{E}_T(t) = E_y \cos(\omega t) \vec{a}_y + E_z \cos(\omega t + \delta) \vec{a}_z \quad (18)$$

whose variation in time describes an ellipse as represented in Fig. 3 [5], [12], [13].

As illustrated in Fig. 3, $\zeta$ describes the relationship between the peak components $E_y$ and $E_z$:

$$\zeta = \tan^{-1}\left(\frac{E_z}{E_y}\right) \quad 0^\circ \leq \zeta \leq 90^\circ \quad (19)$$

Also in Fig. 3, it can be noted that $\tau$ is the angle between the y-axis (horizontal) and the major axis of the ellipse.

Still from Fig. 3:

$$\zeta = \cot^{-1}(AR) \quad 1 \leq |AR| \leq \infty \quad -45^\circ \leq \xi \leq 45^\circ \quad (20)$$

where $|AR|$ is the axial ratio of the ellipse that describes the ratio of its major axis to its minor axis. The signal of AR is positive for clockwise and negative for counterclockwise sense of rotation.

$\xi, \delta, \tau$ are the parameters of the ellipse, where either the pair ($\xi, \tau$) or the pair ($\xi, \delta$) uniquely define the geometry of the ellipse. The interrelations of these angles are described below [5], [12], [13]:

$$\cos2\xi = \cos2\xi \cos2\xi \quad (21)$$

$$\tan\delta = \frac{\tan2\xi}{\sin2\tau} \quad (22)$$

$$\tan2\tau = \tan2\xi \cos\theta \quad (23)$$

$$\sin2\xi = \sin2\xi \sin\delta \quad (24)$$

According to (18) and Fig. 3, we present in Fig. 4 some of the above-mentioned ellipses traced out at constant distances:

Below in Figs. 5 and 6, graphics of the $\xi$ and $\tau$ angles in regard to distance $r$ on the ground, are presented for rural ground:

Below in Fig. 7, a comparison between vertical and horizontal components of the total electric field:

Comparing the two-ray model ellipses shown in Fig. 4 with their respective parameters indicated in Figs. 5 and 6, we see that the geometry of these ellipses reflects exactly the parameters calculated. One also notes that, for the ellipses, when $r$ rises, $\xi$ tends to go to $0^\circ$ and $\tau$ tends to go to $90^\circ$ since the vertical component tends to become much greater than the horizontal component (Figs. 5, 6 and 7). For the ellipses with greater axial rates (thinner ellipses) the $\xi$ parameters assume the lower values (Fig. 5). The parameter $\zeta$ assumes values lower than $45^\circ$ only when $E_z$ is lower than $E_y$ and like $\tau$, tends to go to $90^\circ$ as $r$ rises.
3.3 Error Analysis

Below in Fig. 8, a comparative graphic between non-approximated model (vectorial and gain analysis) and the approximated model (only algebraic sum of the fields):

Applying mathematical approximations in Equation 6-2:

\[ r_2 \approx r + \frac{(h_A + h_B)^2}{2r} \]  

(25)

From Fig. 2 and assuming the equality in (25), it follows that:

\[ \cos \alpha = \frac{r}{r_2} = \frac{r}{r + \frac{(h_A + h_B)^2}{2r}} = \frac{2r^2}{2r^2 + (h_A + h_B)^2} \]  

(26)

(27) provides a \( r \) value for an as small \( \alpha \) as desired, from which on the equality of the approximated and exact electric fields can be assumed, with an absolute error lower than \( | \varepsilon (2 - \varepsilon) | |E_1 + E_2| \) as demonstrated.
below:

\[ 1 - \cos \alpha = \varepsilon \Rightarrow 1 - \cos \psi < \varepsilon \Rightarrow \cos \psi > 1 - \varepsilon \quad (28) \]

\[ |\hat{E}_1 + \hat{E}_2| - |(1 - \varepsilon)^2 |\hat{E}_1 + \hat{E}_2| > error \]

\[ \varepsilon < [2 - \varepsilon]|\hat{E}_1 + \hat{E}_2| \quad (30) \]

where \( \hat{E}_1 \) and \( \hat{E}_2 \) are given in (4) and (5), respectively.

Considering the error in dB:

\[ 10 \log_{10} E_{\text{approx}} - 10 \log_{10} E_{\text{exact}} = error(dB) \quad (32) \]

\[ 10 \log_{10} \left( \frac{|\hat{E}_1 + \hat{E}_2|}{(1 - \varepsilon)^2 |\hat{E}_1 + \hat{E}_2|} \right) > error(dB) \quad (33) \]

\[ error(dB) < -20 \log_{10}(1 - \varepsilon) \quad (34) \]

Making \( \eta = -20 \log_{10}(1 - \varepsilon) \), it follows that:

\[ \varepsilon = 1 - 10^{-\frac{\eta}{20}} \quad (35) \]

Substituting (35) in (27), it can be seen that:

\[ r = (h_A + h_B) \sqrt{\frac{10^{-\frac{\eta}{20}}}{2.(1 - 10^{-\frac{\eta}{20}})}} \quad (36) \]

which is a \( r \) value for an as small \( \alpha \) as desired, from which one can see the equality of the approximated and exact electric fields can be assumed, with an error in dB lower than \( \eta \) (see Fig. 9).

As (25) is just an approximation, and the exact \( r_2 \) is always lower than this approximation, the above (31) and (34) are reinforced, since \( f(\varepsilon) = 2.\varepsilon - \varepsilon^2 \) grows with \( \varepsilon \) in the interval \([0,1)\) [13].

4. STREET WAVEGUIDE

Considering a street waveguide containing the transmitting and receiving antennas, indicated by the points S and P, respectively, and whose lateral borders and ground are constituted by the same material: typical for urban regions (\( \varepsilon = 3 \) and \( \sigma = 10^{-4} \) Mhos/m), one notes that rays from the source S, being S an ideal dipole, would hit P in several manners. In this work, we will consider three different ways: 1) directly; 2) with a ground reflection; and 3) with multiple lateral reflections on the borders of the waveguide [9], [15], [16].

4.1 Horizontal Polarization

Below, in Fig. 10, a sketch of electric field vectors originated by a horizontal dipole is shown. \( \vec{E}_D \) is the electric field vector due to the direct ray and \( \vec{E}_S \) is the electric field vector due to the ground reflected ray. From Fig. 10, it follows that:

\[ \hat{E}_D = \frac{E_0}{r_0} e^{j\omega t} e^{-j\beta r_0} G(\theta_0) \quad (37) \]

\[ \theta_0 = 90^\circ + \chi \quad (38) \]
\[
\chi = \arctg \left( \frac{y \_S - y \_Z}{z} \right) \tag{39}
\]
\[
r_0 = \sqrt{(y \_S - y)^2 + z^2} \tag{40}
\]
\[
\dot{E}_S = \hat{R}_1 \frac{E_0}{r_S} e^{j \omega t \_x} e^{-j \beta r \_S} G(90^\circ + \theta \_S) \tag{41}
\]
\[
\theta \_S = \arctg \left( \frac{y \_S - y \_C}{c} \right) \tag{42}
\]
\[
c = \sqrt{(x \_S + x)^2 + z^2} \tag{43}
\]
\[
r_S = \frac{2 \cdot x \_S}{\sin \alpha \_S} = \frac{2 \cdot x}{\sin \alpha} \tag{44}
\]
\[
\alpha \_S = \arctg \left( \frac{2 \cdot x \cdot \cos \chi \_z}{z} \right) = \arctg \left( \frac{2 \cdot x \cdot \cos \chi}{z} \right) \tag{45}
\]

One notes four kinds of generalization for reflections on the lateral borders of a plane waveguide:

- First upper reflection and an odd number of reflections sketched below in Fig. 11:

\[
\dot{E}_n = \hat{R}_n \frac{E_0}{r_n} e^{j \omega t \_x} e^{-j \beta r \_n} G(\theta) \tag{46}
\]
\[
r_n = \frac{z}{\cos \alpha} \tag{47}
\]
\[
\alpha = \arctg \left( \frac{(n+1) \cdot h - y \_S + y}{z} \right) \tag{48}
\]
\[
\theta = 90^\circ - \alpha \tag{49}
\]

- First upper reflection and an even number of reflections sketched below in Fig. 12:

\[
\dot{E}_n = \hat{R}_n \frac{E_0}{r_n} e^{j \omega t \_x} e^{-j \beta r \_n} G(\theta) \tag{50}
\]
\[
r_n = \frac{z}{\cos \alpha} \tag{51}
\]
\[
\alpha = \arctg \left( \frac{nh - y \_S + y}{z} \right) \tag{52}
\]
\[
\theta = 90^\circ + \alpha \tag{53}
\]

- First lower reflection and an odd number of reflections sketched below in Fig. 13:

\[
\dot{E}_n = \hat{R}_n \frac{E_0}{r_n} e^{j \omega t \_x} e^{-j \beta r \_n} G(\theta) \tag{54}
\]
\[
r_n = \frac{z}{\cos \alpha} \tag{55}
\]
\[
\alpha = \arctg \left( \frac{h \_S + y \_S - y}{z} \right) \tag{56}
\]
\[
\theta = 90^\circ + \alpha \tag{57}
\]

- First lower reflection and an even number of reflections sketched below in Fig. 14:

\[
\dot{E}_n = \hat{R}_n \frac{E_0}{r_n} e^{j \omega t \_x} e^{-j \beta r \_n} G(\theta) \tag{58}
\]
\[
r_n = \frac{z}{\cos \alpha} \tag{59}
\]
\[
\alpha = \arctg \left( \frac{nh + y \_S - y}{z} \right) \tag{60}
\]
\[
\theta = 90^\circ + \alpha \tag{61}
\]

From Figs. 10 - 14 it follows that:

\[
\vec{E}_T = \vec{E}_x + \vec{E}_y + \vec{E}_z \tag{62}
\]
\[
\vec{E}_T = \vec{E}_x e^{j \theta \_x} \vec{a}_x + \vec{E}_y e^{j \theta \_y} \vec{a}_y + \vec{E}_z e^{j \theta \_z} \vec{a}_z \tag{63}
\]
where:

\[
\vec{E}_y = (\hat{E}_T \cdot \sin \theta_S \cdot \sin \psi) \hat{a}_y
\]  
\[
\vec{E}_z = (\hat{E}_T \cdot \sin \theta_S \cdot \cos \psi) \hat{a}_z
\]  
\[
\vec{E}_x = (\hat{E}_x \cdot \cos \theta_S) \hat{a}_x + \hat{E}_y \cdot \cos (\omega t + \theta_y) \hat{a}_y + \hat{E}_z \cdot \cos (\omega t + \theta_z) \hat{a}_z
\]

4.2 Vertical Polarization

Below, in Fig. 15, a sketch of electric field vectors originated by a vertical dipole is shown. \( \vec{E}_D \) is the electric field vector due to the direct ray, \( \vec{E}_S \) is the electric field vector due to the ground reflected ray, and \( \vec{E}_L \) is the vector representing the resulting electric field regarding all lateral reflected rays.

From Fig. 15, it follows that:

\[
\hat{E}_D = \frac{E_0}{r_0} e^{j\omega t} e^{-j\beta t}
\]  
\[
\hat{E}_S = \hat{R}_S \frac{E_0}{r_S} e^{j\omega t} e^{-j\beta t} G(\theta_S)
\]  
\[
\theta_S = \theta_0 + \alpha_S \text{ where } \alpha_S \text{ is given by (45)}
\]  
\[
\hat{E}_T = \hat{E}_y + \hat{E}_z
\]  
\[
\hat{E}_T = \hat{E}_x e^{j\beta z} \hat{a}_x + \hat{E}_y e^{j\beta y} \hat{a}_y + \hat{E}_z e^{j\beta z} \hat{a}_z
\]  
\[
\hat{E}_x = (\hat{E}_D + \sum_{i=1}^{N_{\text{max}}} \hat{E}_{iup} + \sum_{i=1}^{N_{\text{max}}} \hat{E}_{ilow} + \hat{E}_S \cdot \cos \alpha_S) \hat{a}_x
\]  
\[
\hat{E}_y = (\hat{E}_S \cdot \sin \alpha_S \cdot \sin \psi) \hat{a}_y
\]  
\[
\hat{E}_z = -(\hat{E}_S \cdot \sin \alpha_S \cdot \cos \psi) \hat{a}_z
\]  
\[
\hat{E}_T(t) = \hat{E}_x \cos (\omega t + \theta_x) \hat{a}_x + \hat{E}_y \cos (\omega t + \theta_y) \hat{a}_y + \hat{E}_z \cos (\omega t + \theta_z) \hat{a}_z
\]

4.3 Results

According to (67), the tip of the resulting electric field vector traces out an ellipse in 3D space, as shown in Fig. 16:

Below in Fig. 17, a comparison between the main components of the total electric field and electric field due to lateral reflections only, for a frequency of 900 MHz, is presented.

Below in Fig. 18, a comparison between the three components of the total electric field, for a frequency of 900 MHz, is presented.

Below in Fig. 19, a comparison between the non-approximated model (vectorial and gain analysis) and the approximated model (only algebraic sum of the fields), for a frequency of 900 MHz, is presented.

For the street waveguide, we note from Fig. 16 that when \( z \) rises, the ellipses tend to become thin, tend to approximate their major axes from y-axis and tend to have \( E_y \) greater than \( E_z \). When we compare the main components of the electric field due to lateral reflections with the main components of the total electric field (Fig. 17), one notes that for higher distances (z) the lateral reflections field is a good approximation for the total field, while for lower distances a difference can be seen.
Fig. 17: Comparison between electric field due to lateral reflections only and total electric field, for \( y_s = 28 \ [m] \), \( y = 2 \ [m] \), \( x = x_s = 1.8 \ [m] \), \( N_{\text{max}} = 15 \) and \( h = 30 \ [m] \), versus \( z \) distance.

Fig. 18: Comparison between the three components of the total electric field, for urban ground, \( N_{\text{max}} = 15 \) reflections, \( y_s = 28 \ [m] \), \( y = 2 \ [m] \), \( x = x_s = 1.8 \ [m] \), \( h = 30 \ [m] \) and frequency of 900 MHz, versus \( z \) distance.

4.4 Multislit Waveguide

When dealing with a waveguide with slits as illustrated in Fig. 20, the ray reflections coinciding with the position occupied by the slits are not considered, i.e., the rays are lost through the slits. From Fig. 20, it follows that:

\[ qL + (q-1)l < z < qL + ql = qL + (q-1)l < z < q(L+l) \]

\[ (77) \]

Once the slit positions are identified, it is necessary that the points at which the rays will be reflected on to the lateral borders of the waveguide be known:

For all cases there are \( n \) reflections: in \( z_1, \ldots, \) and in \( z_1 + (n-1)\Delta \), where:

\[ 1\uparrow\downarrow \text{Upper Reflection (} n \text{ Reflections - } n \text{ Odd)}: \]

\[ z_1 = \frac{z_h(y_s)}{[(n+1)h - y_s - y]} \]

\[ \Delta = \frac{z_h}{[nh - y_s + y]} \]

\[ (78) \]

\[ (79) \]

\[ 1\uparrow\downarrow \text{Upper Reflection (} n \text{ Reflections - } n \text{ Even)}: \]

\[ z_1 = \frac{z_h(y_s)}{[(n+1)h + y_s + y]} \]

\[ \Delta = \frac{z_h}{[nh + y_s + y]} \]

\[ (80) \]

\[ (81) \]

\[ 1\downarrow\uparrow \text{Lower Reflection (} n \text{ Reflections - } n \text{ Odd)}: \]

\[ z_1 = \frac{z(y_s)}{[(n-1)h + y_s + y]} \]

\[ \Delta = \frac{z_h}{[nh + y_s - y]} \]

\[ (82) \]

\[ (83) \]

\[ 1\downarrow\uparrow \text{Lower Reflection (} n \text{ Reflections - } n \text{ Even)}: \]

\[ z_1 = \frac{z(y_s)}{[(n-1)h + y_s - y]} \]

\[ \Delta = \frac{z_h}{[nh + y_s - y]} \]

\[ (84) \]

\[ (85) \]
Below in Fig. 21, we provide a comparison of the main components of the total electric field for several slit lengths:

![Comparison of the main components of the total electric field for different slit lengths](image1)

Fig.21: Comparison of the main components of the total electric field for different slit lengths ($l$), for $y_s = 28 \, [m]$, $y = 2 \, [m]$, $x = x_s = 1.8 \, [m]$, $N_{\text{max}} = 15$ and $h = 30 \, [m]$, versus $z$ distance.

Regarding the multislit waveguide, one observes from Fig. 21 some sharp transitions in the field due to the lost rays through the slits, and the greater is the length of the slit ($l$), the greater is the width of these sharp transitions, making the signal brittle principally for horizontal polarization. As the slits are positioned periodically, one observes a periodicity in such transitions, in regard to $z$ variation, proportional to $L + l$. One can see that for $z \leq \kappa$ (where $\kappa \sim (L/|y_s - y|)$ for $y_s \neq y$, and $\kappa$ is maximum when $y_s = y$) the field remains unaltered.

5. CONCLUSIONS

For the two-ray model, when one compares the approximated model and the non-approximated model for parallel polarization, it can be seen a large difference between these two models, principally for lower distances ($r$) (Fig. 8). The relation seen in (36) can be an important tool for detecting if a vectorial analysis is necessary for a determined two-ray model planning, in accordance with an acceptable error.

For the street waveguide we can conclude that for higher distances the two-ray model field decays at a higher rate than the lateral reflections field, which is in good agreement with [16]. It asserts the extreme importance that lateral reflections have in a street waveguide model. As in the two-ray model, when one compares the approximated and the non-approximated model for horizontal polarization (Fig. 19), a difference can be seen between these two models, principally for lower distances ($z$). When we analyze reflections on the waveguide borders, a comparison is made such that a maximum number of reflections ($N_{\text{max}}$) necessary to offer a good estimate of the electric field is arrived at. Tests reveal that $N_{\text{max}} \sim z/h$.

This model was built with the idea of helping create the 3D street waveguide model, structuring the bases for its development.

A vectorial and gain analysis can be a very useful tool toward the increase of theoretical analysis in spite of empirical analysis in the field of prediction, emerging as an interesting instrument regarding the incessant search for better signal prediction [13], [14], [15].

References


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