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Electrodynamic Analysis of Antennas in Multipath Conditions

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1. Introduction

During the last years, a sustained growth in the number of users in mobile communication systems has led the necessity to increase their capacity. With the imminent arrival of the fourth generation and the deployment of new high data rate services, the increasing of system capacity is essential. Adaptive arrays and MIMO antennas (multiple input and multiple output) become a promising affordable solution.. The use of these techniques will not only increase the capacity but also they improve the signal quality, coverage range and they simplify the application of new services by exploiting the specific characteristics of these antenna systems. However, the introduction of these technologies involves changes in network planning and deployment, in addition, the increased complexity of both the tranceptors as the radio resource management. In these new radio technologies, antenna system devices shall be designed to get advantage from the multipath. Moreover antennas cannot be considered apart from the radio channel to define the whole structure of radio-communication system, as well as, its management and planning. This is why this chapter addresses the analysis of antennas in a multipath environment. The chapter consists of four more sections. The following one is a comprehensive and detailed analysis from the point of view of the electromagnetic theory of a receiving antenna. It will be the basis for the third section, which presents the time analysis of the output antenna signals with deep fading, oriented to a better understanding of more complex systems as the MIMO ones. In this case, the procedure is used to evaluate the spatial correlation between radiators. The fourth section provides a brief description of fading signals from a statistical point of view, deepening into the classification of radio channels. The fifth section presents some conclusions and a list of references.

2. Receiving antenna

When a wave arrives to an antenna, it will circulate a current on the antenna conductors; inducing an electromotive force (*e.m.f*) (at the input terminals of the antenna, so it excites a guided wave through the receiver input. In Figure 1 this situation is outlined, where Z_R is the input impedance of the receiver (neglecting the input transmission line, that is to say, the

receiver input impedance reflected at the terminals of the line connected to the antenna). This situation is quite common using transmission lines matched to the receiver impedance.
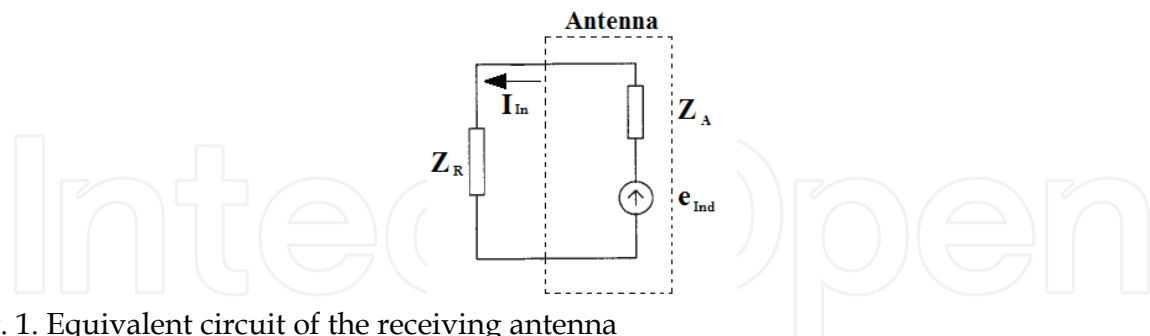

Fig. 1. Equivalent circuit of the receiving antenna

Please, note that under these conditions, the antenna behaves as a generator with an input impedance equal to the output transmitting antenna impedance. The remaining problem to solve is the ratio of effective field incident on the antenna and induced e.m.f on its terminals. In order to study this problem, the reciprocity theorem shall be applied to the situations that are shown in Figure 2.
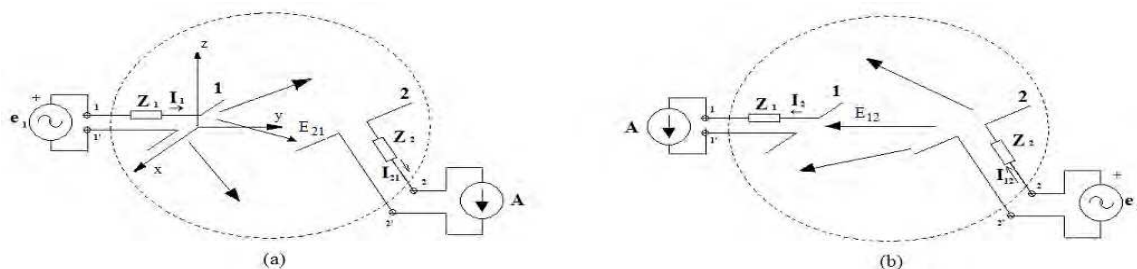

Fig. 2. Equivalent quadripole, linear and reciprocal, to study the interaction between a transmitter and a receiver antenna

Consider any two antennas, placed arbitrarily in free space mutually in far zone, in order to obtain the voltages and currents on their terminals, the previous proposed model that describes a link between two antennas as a linear and reciprocal quadripole will be used. If the terminals 1-1'of the quadripole of Figure 2a is connected to a generator the *e.m.f* e_1 and impedance Z_1, at Terminal 2-2', connected to a load Z_2, it circulates a current I_{21}. If a generator, with an *e.m.f* e_2 and output impedance Z_2, is connected to the terminals 2-2' (Figure 2b), and a load Z_1 is connected to terminals 1-1', it circulates a current I_{12}. According to the reciprocity theorem (Monson, 1996):

$$\frac{e_1}{I_{21}} = \frac{e_2}{I_{12}}$$

(1)

and following the formula of an antenna in far field, the situation shown in Figure 2a the actual value of the field produced by the antenna 1 in the location of the antenna 2 can be determined (Dolujanov, 1965):

$$E_{21} = \frac{\sqrt{30 \cdot D_{Max1} \cdot P_{Rad1}}}{r} \cdot F_{C1}(\theta_{21}, \phi_{21}) \tag{2}$$

where the subscript 1 refers to the first antenna; θ_{21} y ϕ_{21} are the angles that define the direction in which the antenna 2, and F_{C1}. is the radiation pattern (directional characteristic). The coordinate system is placed on the antenna 1. Every one of the parameters included in the formula (2) refers to the resulting field of the antenna (superposition of the main polarization and the cross-polarization). At the input of the antenna 1, the effective value of current will be:

$$I_1 = \frac{e_1}{|Z_{A1} + Z_1|}$$

Likewise, according to the expression of the radiation resistance of an antenna (Nikolski, 1976), this is:

$$P_{Rad1} = I_1^2 \cdot R_{Rad} = \left| \frac{e_1}{Z_{A1} + Z_1} \right| \cdot R_{Rad1}$$

Substituting this result in (2) and clearing e_1, it gets:

$$e_1 = \frac{E_{21} \cdot |Z_{A1} + Z_1| \cdot r}{\sqrt{30 \cdot R_{Rad1} \cdot D_{Max1}} \cdot F_{C1}(\theta_{21}, \phi_{21})}. \tag{3}$$

When repeating the above analysis to the situation shown in Figure 2b, it gets:

$$e_2 = \frac{E_{12} \cdot |Z_{A2} + Z_2| \cdot r}{\sqrt{30 \cdot R_{Rad2} \cdot D_{Max2}} \cdot F_{C2}(\theta_{12}, \phi_{12})}. \tag{4}$$

According to the Reciprocity Theorem (1), the (3) y (4) can be related:

$$\frac{E_{21}|Z_{A1} + Z_1| \cdot r}{I_{21} \cdot \sqrt{30 \cdot R_{Rad1} \cdot D_{Max1}} \cdot F_{C1}(\theta_{21}, \phi_{21})} = \frac{E_{12}|Z_{A2} + Z_2| \cdot r}{I_{12} \cdot \sqrt{30 \cdot R_{Rad2} \cdot D_{Max2}} \cdot F_{C2}(\theta_{12}, \phi_{12})}$$

And it can also be written as:

$$\left[\frac{I_{12}}{E_{12}}\right] \cdot \frac{E_{21}|Z_{A1} + Z_1| \cdot r}{\sqrt{30 \cdot R_{Rad1} \cdot D_{Max1}} \cdot F_{C1}(\theta_{21}, \phi_{21})} = \left[\frac{I_{21}}{E_{21}}\right] \cdot \frac{E_{12}|Z_{A2} + Z_2| \cdot r}{\sqrt{30 \cdot R_{Rad2} \cdot D_{Max2}} \cdot F_{C2}(\theta_{12}, \phi_{12})} \tag{5}$$

Up to now the polarization of the incident wave has not be considered. However, the induced current in the surface of the receiving antenna is determined by those incident wave components "parallel" to the polarization of the receiving antenna[1]. That is to say, the wave field produced by the antenna 2 in the location of the antenna 1 can be expressed as:

$$\vec{E}_{12} = E_{12} \cdot \vec{e}_{2T}\left(\theta_{12}, \phi_{12}\right)$$

where E_{12} is the actual value of the field and $\vec{e}_{2T}$ is normalized vector indicating the polarization of the wave transmitted by the antenna 2. When identifying the type of polarization of the antenna 1 itself in reception for the normalized vector: $\vec{e}_{2R}\left(\theta_{21}, \phi_{21}\right)$, thus, the effective value of the component of the incident field that matches the type of polarization of the antenna 1 at the reception will be as (Márkov & Sazónov, 1978):

$$E_{R12} = E_{12} \cdot \left| \vec{e}_{2T}\left(\theta_{12}, \phi_{12}\right) \cdot \vec{e}_{1R}\left(\theta_{21}, \phi_{21}\right)\right| \qquad (6)$$

Note that for standard polarization vectors is satisfied that:

$$\vec{e}_{2T}\left(\theta_{12}, \phi_{12}\right) \cdot \vec{e}_{2T}^{*}\left(\theta_{12}, \phi_{12}\right) = 1; \quad \vec{e}_{2R}\left(\theta_{21}, \phi_{21}\right) \cdot \vec{e}_{2R}^{*}\left(\theta_{21}, \phi_{21}\right) = 1$$

Therefore E_{R12} (6) is the real value of electric involved in the process of reception. Note that if the antennas 1 and 2 were equal and with identical directions of pointing ($\theta_{12} = \theta_{21}$ y $\phi_{12} = \phi_{21}$), thus:

$$\left| \vec{e}_{2T}\left(\theta_{12}, \phi_{12}\right) \cdot \vec{e}_{2R}^{*}\left(\theta_{12}, \phi_{12}\right)\right| = \left| \vec{e}_{1T}\left(\theta_{21}, \phi_{21}\right) \cdot \vec{e}_{1R}^{*}\left(\theta_{21}, \phi_{21}\right)\right|$$

where the maximum value will happen when: $\vec{e}_{1T}\left(\theta_{21}, \phi_{21}\right) = \vec{e}_{1R}^{*}\left(\theta_{21}, \phi_{21}\right)$, which the polarization of the transmitting antenna and the receiving one are the same but with opposite sense (seen from a common reference system), that is to say, an identical polarization seen from transmitting point of view. Therefore, we can write the expression (6) as:

$$E_{12} = \frac{E_{R12}}{\left| \vec{e}_{2}\left(\theta_{12}, \phi_{12}\right) \cdot \vec{e}_{1}^{*}\left(\theta_{21}, \phi_{21}\right)\right|} \qquad (7)$$

---

[1] Polarization of an antenna is defined from transmission point of view. However, although the reception point of view is opposite to the transmission one, the polarization of an antenna is defined equally.

without distinction in the polarization vectors whether it is an antenna transmission or reception. Similarly we can write the actual value of the field incident at the antenna 2 from the antenna 1:

$$E_{21} = \frac{E_{R21}}{\left| \vec{e_1}(\theta_{21},\phi_{21}) \cdot \vec{e_2^*}(\theta_{12},\phi_{12}) \right|} \tag{8}$$

Given that the denominators of expressions (7) and (8) are equal, and substituting these in (5), we obtain:

$$\left[\frac{I_{12}}{E_{R12}}\right] \cdot \frac{|Z_{A1} + Z_1|}{\sqrt{R_{Rad1} \cdot D_{Max1}} \cdot F_{C1}(\theta_{21},\phi_{21})} = \left[\frac{I_{21}}{E_{R21}}\right] \cdot \frac{|Z_{A2} + Z_2|}{\sqrt{R_{Rad2} \cdot D_{Max2}} \cdot F_{C2}(\theta_{12},\phi_{12})} \tag{9}$$

Analysing both members; the relationship (I_{12}/E_{R12}) depend only on the characteristics of the antenna. The field incident on the antenna with the same polarization induced currents on the antenna. On the other hand the factors of the left member of (9) depend exclusively on the characteristics of the antenna 1. Similarly, it appears that all the factors of the right-hand side of (9) depend exclusively on the characteristics of the antenna 2. Since the above analysis there is anyone restriction on the type of antenna used (in general, antennas 1 and 2 are different), the obvious conclusion is:

$$\left[\frac{I_{12}}{E_{R12}}\right] \cdot \frac{|Z_{A1} + Z_1|}{\sqrt{R_{Rad1} \cdot D_{Max1}} \cdot F_{C1}(\theta_{21},\phi_{21})} = C \tag{10}$$

where C is a constant that has the same value for all antennas. Therefore, C can be obtained by replacing in (10) the values of the parameters of any antenna; in particular the Hertz's dipole. Figure 3 shows a dipole antenna formed by a thin conductor of length L, with an impedance Z_R connected at its terminals, impinging a wave of linear polarization parallel to the dipole.
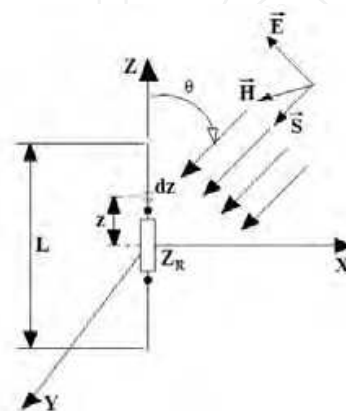

Fig. 3. Antenna type dipole in reception

This will induce a current $I(z)$ along the conductor. The *e.m.f de*, induced a small segment of length dz will be:

$$de = E \cdot sen(\theta) \cdot dz \qquad (11)$$

As for the conductor circulate the current $I(z)$, a power will be delivered to the antenna:

$$dP = de \cdot I^*(z) = E \cdot sen(\theta) \cdot I^*(z) \cdot dz$$

The total power delivered to the antenna is:

$$P = \int_L E \cdot sen(\theta) \cdot I^*(z) \cdot dz$$

In the terminals (see Figure 1) is obtained:

$$P = I_{In}^2 \cdot (Z_A + Z_R)$$

And thus:

$$I_{In}^2 \cdot (Z_A + Z_R) = \int_L E \cdot sen(\theta) \cdot I^*(z) \cdot dz$$

In the case of a Hertz dipole with length $\Delta l \ll \lambda$ we can presume the current uniform: $I(z) = I_{Ent}$ and, therefore:

$$I_{In}^2 \cdot (Z_A + Z_R) = E \cdot sen(\theta) \cdot I_{Ent}^* \cdot \Delta l$$

Or what is the same: $\dfrac{I_{Ent}}{E} = \dfrac{I_{12}}{E_{R12}} = \dfrac{sen(\theta) \cdot \Delta l}{|Z_A + Z_R|}$

The radiation pattern of Hertz's dipole is known $F_C(\theta) = sen(\theta)$, its directivity is of: $D_{Max} = 1.5$. Moreover, the radiation resistance is of: $R_{Rad} = 80\pi^2(\Delta l/\lambda)^2$; so, substituting these data in (10):

$$C = \frac{\lambda}{\sqrt{120 \cdot \pi}}$$

where the value *C* obtained for the dipole Hertz is unique for all antennas, and it can be replaced in (10) to obtain the expression of the current at the terminals of the any antenna

(I_{Ind}), induced by that component of the field of the incident wave (E_R) with a polarization equal to the receiving antenna itself and arriving at the antenna in the direction defined by angles θ and ϕ :

$$I_{Ind} = E_R \cdot \frac{\lambda}{\pi \cdot |Z_A + Z_R|} \cdot \sqrt{\frac{R_{Rad} \cdot D_{Max}}{120}} \cdot F_C(\theta,\phi) \qquad (12)$$

According to Figure 1 the induced *e.m.f* at the terminals of the antenna can be determined by:

$$e_{Ind} = I_{Ind} \cdot |Z_A + Z_R| = E_R \cdot \frac{\lambda}{\pi} \cdot \sqrt{\frac{R_{Rad} \cdot D_{Max}}{120}} \cdot F_C(\theta,\phi) \qquad (13)$$

Taking into account the expressions (12) and (13), the values of the current and the induced *e.m.f* at the terminals of the antenna have a dependency on the direction of arrival of the incident wave, expressed by $F_C(\theta,\phi)$, which is the radiation pattern of the antenna in transmission. The expression (13) can be written like this:

$$e_{Ind}(\theta,\phi) = e_{Ind\,Max} \cdot F_C(\theta,\phi)$$

Then $e_{Ind\,Max}$ expresses the value of induced *e.m.f* when the wave arrives from the direction of maximum reception of the antenna, and $F_C(\theta,\phi)$ represents the normalized radiation pattern of the field of the antenna in reception mode, which is equal to that characteristic of their antenna transmission. On the other hand, the coefficient of directivity is function of the radiation pattern. Therefore, it confirms that its value is the same regardless of the antenna works as transmitters or receivers. Similarly, in the case of linear antennas, the effective length:

$$l_{Ef} = \frac{\lambda}{\pi} \cdot \sqrt{\frac{D_{Max} \cdot R_{Rad}}{120}} \qquad (14)$$

that depends on the coefficient of maximum directivity, the radiation resistance, and will have the same value in transmission and reception. Substituting in expression (13), it becomes:

$$e_{Ind} = E_R \cdot l_{Ef} \cdot F_C(\theta,\phi)$$

where the product: $E_R \cdot l_{Ef} = e_{Ind\,Max}$ is the maximum value of the induced *e.m.f* (when $F_C = 1$). It is important to stress the significance of the effective length of the antenna; that is to say, this is a length such, when multiplied with the incident field intensity (the polarization equal to the antenna itself, incident in the direction of maximum reception),

gives us the maximum value of the induced *e.m.f* . Known the induced *e.m.f* in the antenna, with reference to Figure 1 we can determine the voltage at the input terminals of the receiver, supposed this is connected directly to the antenna, by:

$$V_R = \left| \frac{Z_R}{Z_A + Z_R} \right| \cdot e_{Ind} \qquad (15)$$

Considering the presence of a transmission line (the characteristic impedance Z_0) between the antenna (impedance Z_A) and receiver (impedance Z_R), it defines the reflection coefficient at the antenna input (Pozar, 2004):

$$\rho_A = -\frac{Z_A - Z_0}{Z_A + Z_0}$$

and at the receiver input: $\rho_R = \frac{Z_R - Z_0}{Z_R + Z_0}$; the expression (15) will transformed:

$$V_R = \frac{|1 + \rho_R| \cdot |1 + \rho_A| \cdot \exp(-\alpha \cdot L)}{2 \cdot |1 + \rho_R \cdot \rho_A \cdot \exp(-j2 \cdot \Gamma \cdot L)|} \cdot e_{Ind} \qquad (16)$$

where $\Gamma = \alpha + j\beta$ is the propagation constant of the transmission line (which includes the attenuation constant α and the phase constant β), and L is the length of the line. In case of low frequency receivers ($f < 30\,MHz$) the condition: $|Z_R| >> |Z_A|$ is usually applied; thus (replacing in 15) is obtained: $V_R \approx e_{Ind}$; so, maximum voltage to the receiver's input. Moreover, matching the transmission line to the antenna ($\rho_A \approx 0$) and $|Z_R| >> Z_0$, $\rho_R \approx 1$ and expression (16) also gives us the maximum input voltage of the receiver: $V_R = \exp(-\alpha \cdot L) \cdot e_{Ind}$, where the factor: $\exp(-\alpha \cdot L) < 1$ takes into account transmission losses along the line. On the other hand, in case of higher frequency it is more difficult to provide high power amplifiers, so the purpose is to maximize the real power delivered by the antenna to receiver. If the receiver is directly connected to the antenna, the power supplied to the receiver is:

$$P_R = I_{Ind}^2 \cdot R_R = E_R^2 \cdot \frac{\lambda^2 \cdot R_R}{\pi^2 \cdot |Z_A + Z_R|} \cdot \frac{R_{Rad} \cdot D_{Max}}{120} \cdot F_P(\theta, \phi)$$

where R_R is the real part of the input impedance of the receiver. Maximum transmitting power forces to an impedance matching between the receiver and antenna ($Z_R = Z_A^*$). Applying this condition in the above expression and using the classical expression of the efficiency of an antenna: $R_{Rad} = \eta_A \cdot R_{In}$, we obtain (Balanis, 1982):

$$P_R = E_R^2 \cdot \frac{\lambda^2}{4 \cdot \pi^2} \cdot \frac{\eta_A \cdot D_{Max}}{120} \cdot F_P(\theta, \phi)$$

When using a transmission line of low losses between the receiver and the antenna and impedance matching between the receiver and the line, the power supplied by the antenna to the line will be:

$$P_L = E_R^2 \cdot \frac{\lambda}{\pi^2} \cdot \frac{Z_0 \cdot R_A}{\left| Z_A + Z_0 \right|^2} \cdot \frac{\eta_A \cdot D_{Max}}{120} \cdot F_p(\theta, \phi) = E_R^2 \cdot \frac{\lambda^2}{4\pi^2} \cdot \left(1 - \left| \rho \right|^2 \right) \cdot \frac{\eta_A \cdot D_{Max}}{120} \cdot F_p(\theta, \phi)$$

A part of power will flow to the receiver:

$$P_R = P_L \cdot \exp(-2 \cdot \alpha L) = E_R^2 \cdot \frac{\lambda^2}{4\pi^2} \cdot \left(1 - \left| \rho \right|^2 \right) \cdot \exp(-2 \cdot \alpha L) \cdot \frac{\eta_A \cdot D_{Max}}{120} \cdot F_p(\theta, \phi) \qquad (17)$$

and the other part: $P_{Cons.L} = P_L \cdot \left[1 - \exp(-2\alpha L) \right]$, will be consumed by the line. In expression (17) we can see that through the appropriate orientation of the antenna, by matching the direction of maximum reception of the antenna with the direction of arrival of the wave we get $F_P = 1$ and the received power reaches its maximum value by adjusting the direction:

$$P_{RMax} = \left(1 - \left| \rho_A \right|^2 \right) \cdot \exp(-2\alpha L) \cdot \eta_A \cdot \frac{E_R^2}{120\pi} \cdot \frac{\lambda^2}{4\pi} \cdot D_{Max} \qquad (18)$$

Considering this expression; factor $E_R^2 / 120\pi$ represents the module of the Poynting vector of the incident wave to the antenna. If we multiply this factor by the physical area of the antenna, the power incident on the antenna is obtained:

$$P_{Inc} = \frac{E_R^2}{120\pi} \cdot A_{Geom} \qquad (19)$$

The antenna is not able to fully grasp the incident power that really is:

$$P_{Cap} = \frac{E_R^2}{120\pi} \cdot \left(\frac{\lambda}{4\pi} \cdot D_{Max} \right) = \frac{E_R^2}{120\pi} \cdot A_{Ef} \qquad (20)$$

where:

$$A_{Ef} = \frac{\lambda^2}{4\pi} \cdot D_{Max} \qquad (21)$$

It will be called the effective area of the antenna; namely the area through which the antenna fully captures the incident power density. The relationship:

$$\xi_A = \frac{P_{CAP}}{P_{Inc}} = \frac{A_{Ef}}{A_{Geom}} \qquad (22)$$

is called the coefficient of the utilization of the surface of the antenna. The expression (18) now we can write:

$$P_{R\,Max} = \left(1 - |\rho_A|^2\right) \cdot \exp\left(-2\alpha L\right) \cdot \eta_A \cdot \xi_A \cdot P_{Inc} = \left(1 - |\rho_A|^2\right) \cdot \exp\left(-2\alpha L\right) \cdot \eta_A \cdot P_{Cap}$$

Note that the antenna does not use all the power captured, but a fraction called <u>useful captured power</u>:

$$P_{Cap\,Use} = \eta_A \cdot P_{Cap} \qquad (23)$$

the other part: $\left(1 - \eta_A\right) \cdot P_{Cap}$ is the power losses in the antenna as heat during the reception. Between the antenna and the transmission line is not always met the condition of impedance matching, therefore only part of the useful captured power useful is delivered to the line:

$$P_L = \left(1 - |\rho_A|^2\right) \cdot P_{Cap\,Útil} \qquad (24)$$

and, only the fraction given by (17) is delivered to the receiver. In Figure 4 shows schematically the flow of power from the wave that propagates in free space and arrives to the antenna, to the receiver. From the analysis we can summarize the conditions for optimal reception: coincidence of the polarization itself of the antenna with the polarization of the incident wave (This ensures E_R maximum); orientation the antenna to the direction of arrival of the wave ($F_P = 1$); effective area (or effective length) maximum of the antenna (which depends on its physical characteristics); high efficiency ($\eta_A = 1$); proper impedances matching between the antenna and feed line ($\rho_A \approx 0$); low losses of the feed line ($\alpha_L \approx 0$). In practice these conditions are met satisfactorily, so the power at the receiver is close to optimal value:

$$P_{Opt} = \frac{E_R^2}{120\pi} \cdot \eta_A \cdot A_{Ef} = \frac{E_R^2}{120\pi} \cdot \frac{\lambda^2}{4\pi} \cdot G_{Max} \qquad (25)$$

This expression clearly reveals the role of the maximum gain at the reception. The value of G_{Max} of any antenna indicates either the number of times that the power delivered to the receiver exceeds that delivered by an isotropic radiator ($G_{Max} = 1$) under the same conditions of external excitation, coupling and losses of the transmission line. In a similar

way we can say that the coefficient of directivity D_{Max} of an antenna (as was noted earlier, has the same value in transmission and reception), at the receiving antenna indicates the number of times that the power captured by the antenna exceeds that delivered by an isotropic radiator. Finally, keep in mind that the presence of induced current in the receiving antenna also determines an effect known as secondary radiation. We must emphasize the fact that, in general, the directional characteristic of the secondary radiation does not match the directional characteristic for the transmitting antenna.
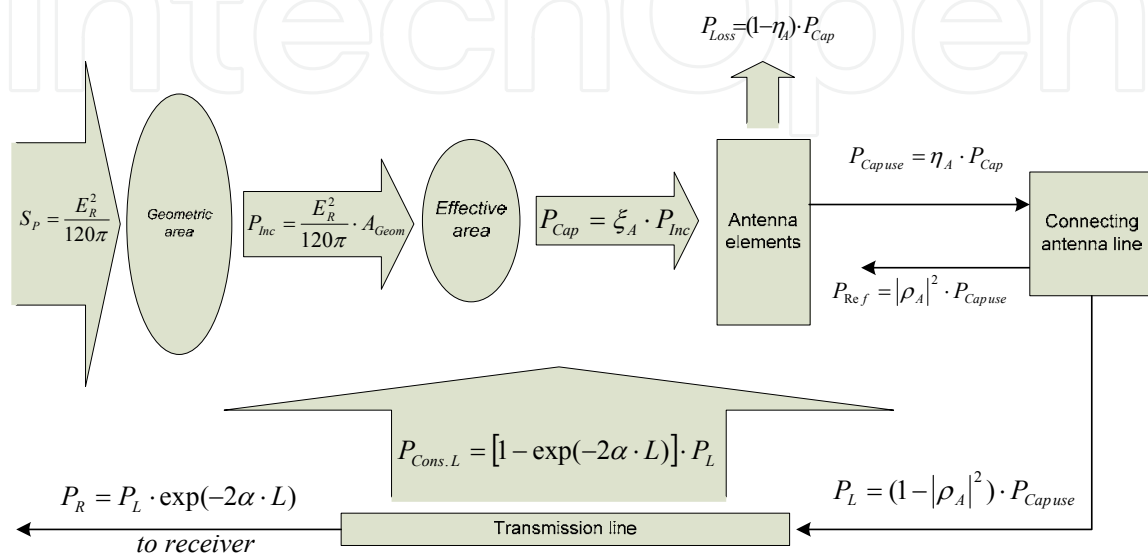
Fig. 4. Power flow from the wave in free space to the receiver

This is explained by that shape of the induced current distribution on the elements of the antenna is not equal to that found when the excitation takes place at the terminals of the antenna. However, the total power of secondary radiation can be calculated from:

$$P_{Sec\,Rad} = I_{Ind}^2 \cdot P_{Rad} = E_R^2 \frac{\lambda^2}{\pi^2} \cdot \frac{R_{Rad}^2}{\left|Z_A + Z_R\right|^2} \cdot \frac{D_{Max}}{120} \cdot F_p\left(\theta, \varphi\right) \tag{26}$$

In this expression the factor: $F_P(\theta, \phi)$ defines the directional pattern of the antenna during the reception; while $P_{Sec\,Rad}$ is the total power of secondary radiation in all directions of space. This phenomenon has a special interest in antennas that act as passive elements, where generally Z_R is the impedance of a ($Z_R = 0$) or a pure reactive element ($Z_R = jX_R$). Particularly this treatment can be extended to objects that do not really fulfil the mission of the antennas, that serving (intentionally or not) as reflectors of radio waves. In these cases, as can be shown easily from (26), it is possible to reach a power of secondary radiation which is 4 times larger than the optimum power of reception given by the formula (25). The analysis of antennas in reception mode, leads to a set of conclusions of great importance. First we establish that many of the properties of the antennas are the same as transmission as reception, which simplifies its research, since it is not necessary to determine these

properties in both regimes. Thus, the impedance of the antenna, its directional pattern, its directivity, efficiency, and gain are the same in both schemes of work. The expressions obtained (mainly induced *e.m.f*) in the receiving antenna (13), are useful in tasks of calculation and design of antennas in general. There are two parameters that are used in the study of the receiving antennas (aperture antennas mainly); the coefficient of utilization of the surface of the antenna and the effective area.

3. Antennas receiving mode in multipath conditions

It is said that an antenna operates under multipath conditions when in it impinge radio waves arriving from different directions. Figure 5 show the multipath phenomenon.
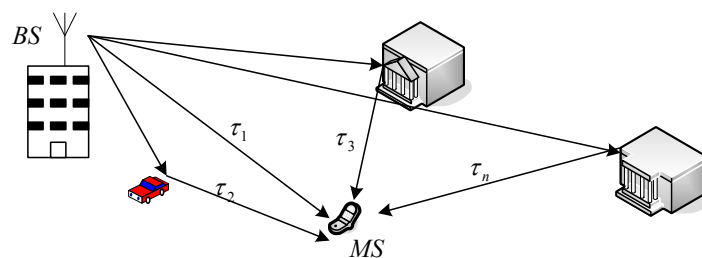
Fig. 5. Phenomenon of multipath propagation

In Figure 5 it can be seen: transmitter and receiver antennas, rays that define the different propagation paths from transmitter to receiver antenna, and the scattering elements (buildings and cars), which are called scatterer. Propagation environments, together with the communications system can be divided into: indoor and outdoor. The theory of radio channels is a rather broad topic not covered here, but from point of view of the antenna, we will present (only from the point of view spatial) similar to the patterns that characterize the radiation of the antennas (Rogier, 2006). This way, according to the angular distribution of power that reaches the antenna, we can present them as omnidirectional, and with some directionality. Then, the shape of the angular distribution of power that characterizes the channel depends on the position of the antenna inside the environment of multipath propagation. Figure 6 shows some examples.
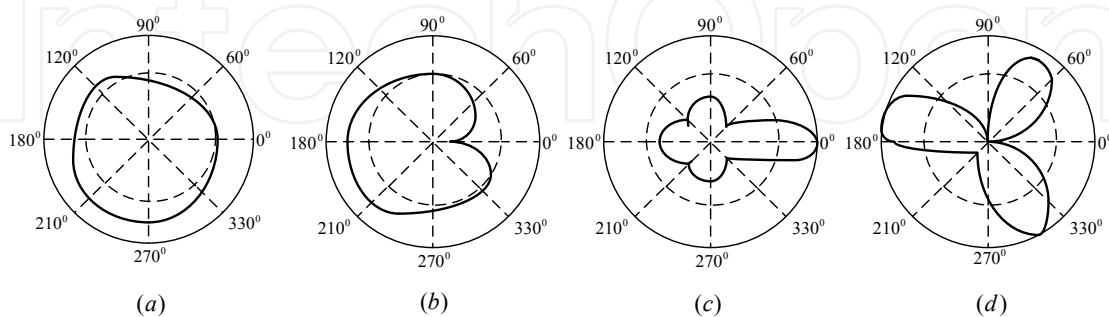

Fig. 6. Angular distribution of the power reaching the antenna by multiple pathways, a) omni directional channel, b) dead zone channel, c) directional channel, d) multidirectional channel.

Figure 6 shows patterns corresponding to the measured power at the terminals of a high directivity antenna used to sample the channel performance in time. The graphs that are shown, they correspond themselves only to the azimuthal plane, often with higher importance as in case of mobile communications.. Figure 6a shows some omni-directional angular distribution, where the incoming waves reach the antenna with a similar intensity from all directions from statistical point of view. Figure 6b presents a channel with an angular distribution indicating some directional properties, in which the waves impinge the antenna from all directions, except from one sector, normally called dead zone. Figure 6c shows the case where the waves reach the antenna from a defined direction. Figure 6d is a typical situation when waves reach the antenna from some well defined directions, (in this case three), in most cases are caused by discrete clusters of scatterer, as in the mobile communications enabling the use of smart antenna systems. In practice there may be all kinds of Multipath described in Figure 6 on a single antenna. This is the true of the antenna is in a mobile terminal that changes its spatial position over time. Induced *e.m.f* at the antenna terminals, which has been described in terms of the angular power spectrum in the plots of Figure 6, is the statistical average of the amplitude of the signal at the antenna terminals. In fact, the resulting signal has a fading performance, due to the fasorial summation of all the waves arriving to the antenna with different amplitudes and phases, due to the difference in the delay associated with each propagation path (Blaunstein et al., 2002. Under this situation induced *e.m.f* in the terminals of an antenna has a fading nature, as it is shown in Figure 7.
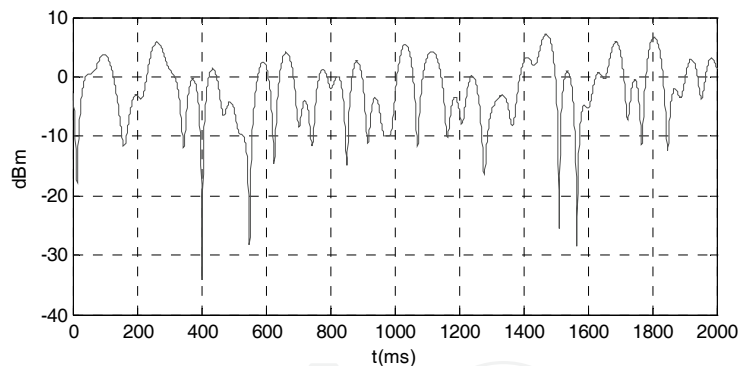

Fig. 7. Signal at the antenna terminals under multipath

The fading performance of the signal can be, explained by the multipath phenomenon using a ray model at the plane. In a first approximation, one considers n waves coming through L_n different paths to a Q- point, in which there is no antenna. The spreading angle associated with each beam is zero, so $\sigma_n = 0$, so it is valid to propose that the resulting signal $u(t)$ in Q point is given by (27):

$$u(t) = \sum_{n=1}^{L_n} a(\phi_n) \cdot s_n(t - \tau_n) + N(t) \tag{27}$$

where $a(\phi_n) = \exp(j \cdot K \cdot \cos(\phi_n))$, is the phase that comes the n signal to the Q point for the ϕ_n direction, $K = 2\pi/\lambda$ is the constant propagation wave to the working frequency, whose wavelength is λ, τ_n is the delay associated to S_n signal. It has also introduced additive noise $N(t)$ in the point. The expression (26) accurately describes the fading nature associated to the multipath. However, the expression (26) does not include the antenna. A computational procedure based on (26), that considers the presence of the antenna and its parameters to obtain the fading signals at their terminals, when it does interact with virtual radio channels is described out below (Molina & De Haro, 2007). The philosophy is based on the creating an effect similar to that described by the equation (27), and simultaneously introduce in the equation that describes the induced *e.m.f* in an antenna in the reception mode (equation (13) of the previous section). Using as antenna a half wavelength dipole, whose radiation pattern is as shown in Figure 8.
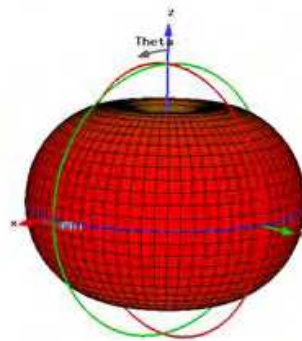
Fig. 8. Radiation pattern of half-wave dipole

The radiation pattern of dipole is displayed using a two dimensional plot of the directional characteristics of amplitude, phase and polarization. The polarization follows the definitions of the main polarization and cross polarization given by (Ludwig 1973 and Markov & Sazonov, 1978). Figure 9 shows two-dimensional pattern of amplitude and phase in the main and crossed polarizations of the dipole. This form of plot allows a faster execution due to the use of matrices in the procedure.
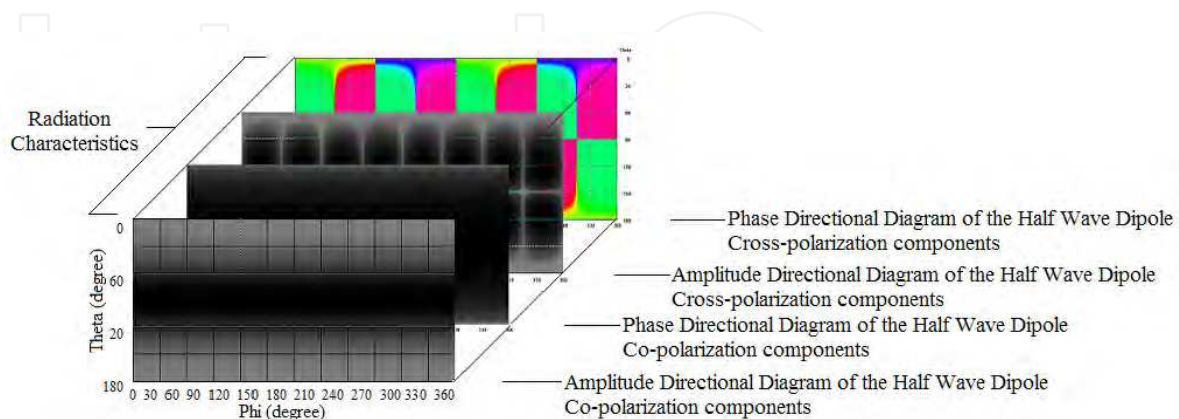

Fig. 9. Radiation characteristics of the half-wave dipole in 2-D.

Moreover, one defines the multipath radio channel as a 360 x 180 matrix that represents all possible directions of space from where can reach the waves to the antenna. It is generated dynamically by the pseudo random number of waves, their amplitudes, their phases, and the coordinates of its angular direction of arrival. In the same way you generate the angular spread associated to each ray.. Figure 10 shows the results of a simulation of the virtual channel multipath where the probability distribution associated with the generation of the angles of arrival was uniform in the 4π radians of the sphere and the probability distribution of the amplitude of the signals was uniform too
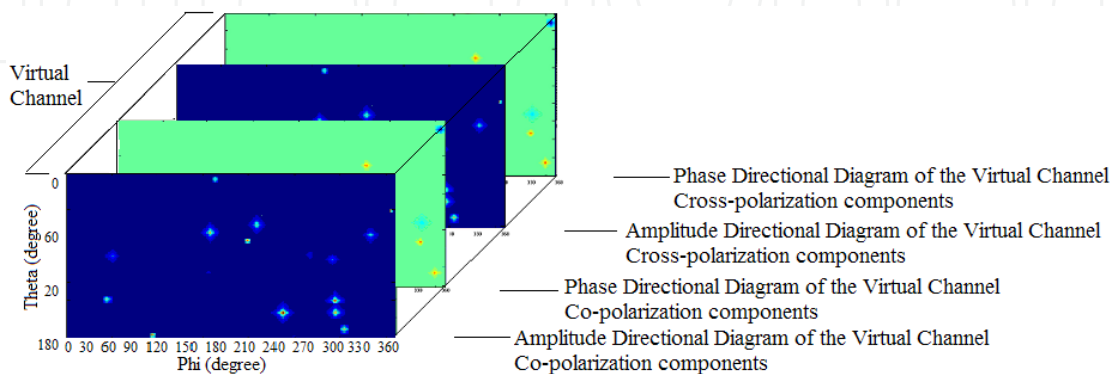
Fig. 10. Virtual multipath radio channel in 2-D.

Note in Figure 10, that as in the case of the antenna, the colour scale indicates the intensity with which the signal arrives. The channel has been defined by analogy with the antennas: amplitude and phase pattern in the main and cross polarization, corresponding to the signals that incident in the antenna. Now, the $\vec{H}(\theta,\phi,t)$ function, models the dynamic performance of space time channel, which, by analogy with the antennas, are contained all the characteristics of amplitude, phase and polarization of the channel:

$$\vec{H}(\theta,\phi,t) = H_\theta(\theta,\phi,t)\vec{i}_\theta + H_\phi(\theta,\phi,t)\vec{i}_\phi \tag{28}$$

where H_θ and H_ϕ are the functions in the principals planes of channel (in main and cross polarization). They, $\vec{i}_\theta$, $\vec{i}_\phi$ are unit vectors indicating the orientation of the electric field incident. Each one of the functions that are part of the right-hand side of (28) is defined as follows:

$$\vec{H}_\theta(\theta,\phi,t) = h_\theta(\theta,\phi,t) \cdot e^{j\psi_\theta(\theta,\phi,t)} \vec{i}_\theta \tag{29a}$$

and

$$\vec{H}_\phi(\theta,\phi,t) = h_\phi(\theta,\phi,t) \cdot e^{j\psi_\phi(\theta,\phi,t)} \vec{i}_\phi \tag{29b}$$

They, h_θ and h_ϕ are the directional pattern associated to the amplitude component of the main and cross polarization in the channel. ψ_θ and ψ_ϕ are also the patterns of the phase associated to the main and cross polarization components of the channel. With this description, and adapting the equation (13) the previous section to the present situation; induced *e.m.f* at the terminals of the dipole can be raised by the following expression in scalar form:

$$u(t) = l_{Ef} \cdot \int_0^{2\pi}\int_0^{\pi} H(\theta,\phi,t) \cdot F(\theta,\phi) \cdot sen(\theta)d\theta d\phi \quad [V] \tag{30}$$

The components for the main and cross polarization, equation (30) are:

$$\begin{aligned} u(t) = l_{Ef} \cdot \int_0^{2\pi}\int_0^{\pi} h_\theta(\theta,\phi,t) \cdot e^{j\Psi_\theta(\theta,\phi,t)} \cdot f_\theta(\theta,\phi) \cdot e^{j\Phi_\theta(\theta,\phi)} sen(\theta)d\theta d\phi + \\ + l_{Ef} \cdot \int_0^{2\pi}\int_0^{\pi} h_\phi(\theta,\phi,t) \cdot e^{j\Psi_\phi(\theta,\phi,t)} \cdot f_\phi(\theta,\phi) \cdot e^{j\Phi_\phi(\theta,\phi)} sen(\theta)d\theta d\phi \end{aligned} \quad [V] \tag{31}$$

where we recall that the functions f_θ and f_ϕ are the directional characteristics amplitude at the main and cross polarization and the functions Φ_θ and Φ_ϕ are the phase patterns of the antenna in these polarizations. The geometrical interaction between the antenna and the multipath radio-channel can be shown in Figure 11.
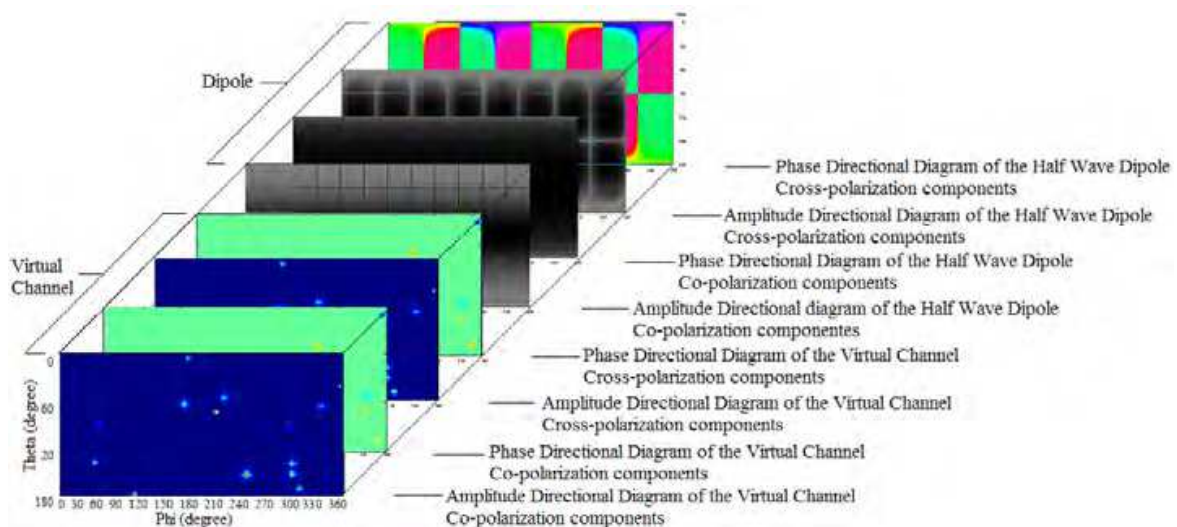

Fig. 11. Geometry of the interaction between the antenna and the channel

Last expressions can be treated in a discrete way, tabulating the functions required by the equation (31):

$$u(t) = l_{Ef} \cdot \sum_{\phi=0}^{N} \sum_{\theta=0}^{M} h_\theta(\theta,\phi,t) \cdot e^{j\Psi_\theta(\theta,\phi,t)} \cdot f_\theta(\theta,\phi) \cdot e^{j\Phi_\theta(\theta,\phi)} sen(\theta) \Delta\theta \Delta\phi +$$
$$+ l_{Ef} \cdot \sum_{\phi=0}^{N} \sum_{\theta=0}^{M} h_\phi(\theta,\phi,t) \cdot e^{j\Psi_\phi(\theta,\phi,t)} \cdot f_\phi(\theta,\phi) \cdot e^{j\Phi_\phi(\theta,\phi)} sen(\theta) \Delta\theta \Delta\phi \qquad [V] \qquad (32)$$

In (32) equation, $\Delta\theta = \pi/N$ y $\Delta\phi = 2\pi/M$, represent the angular resolution step in coordinates θ and ϕ respectively, with a value of one degree, so $N = 180$ and $M = 360$. The induced *e.m.f* obtained on the terminals of the dipole due to its interaction with the virtual channel is shown below:
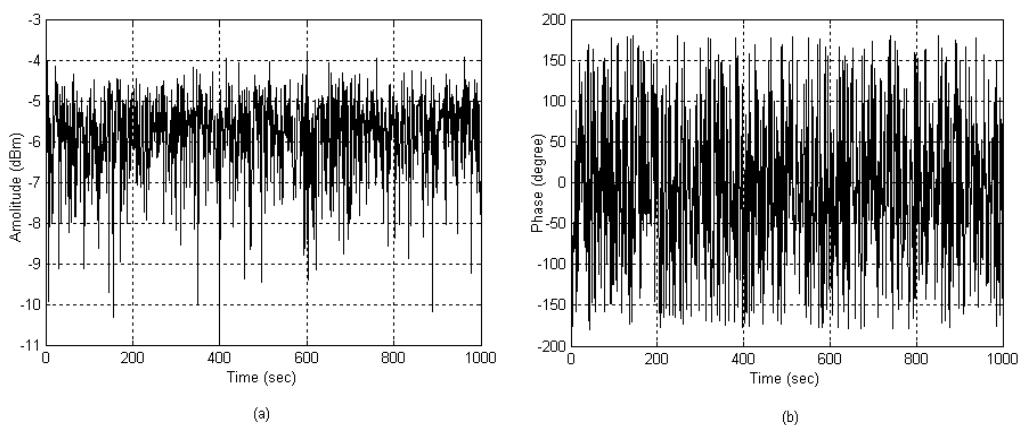
Fig. 12. Signal at the antenna terminals. a) Amplitude fading, b) phase fading.

As seen in Figure 12, the induced *e.m.f.* in the antenna terminals, obtained by the procedure has a fading performance. Moreover, amplitude fading have a statistical distribution of *Rayleigh* type and the phase is uniformly distributed in the range $[0-2\pi]$, (Molina & De Haro, 2008). Figure 13 shows the statistical adjustment of the amplitude and phase fluctuations of this signal.
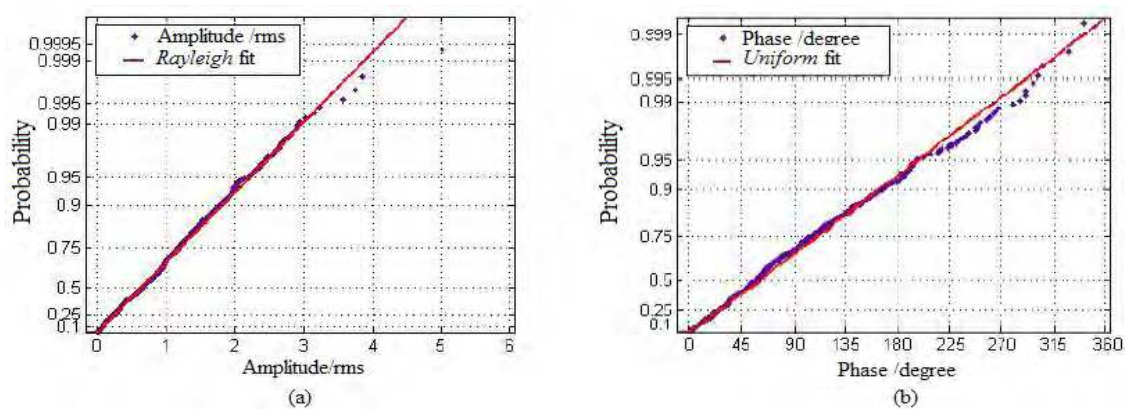

Fig. 13. Statistical adjustment of signal fading. a) Amplitude distribution of type *Rayleigh*, b) phase *uniform* in the interval $[0-2\pi]$.

An important outcome of the simulation is the fact that the signal statistics obtained corresponds closely to real signal channels (Blaunstein et al., 2002). This also provides an explanation of the operation of the receiving antennas under multipath environments, the idea can be reused for practical purposes. So, it is possible the measurement of spatial correlation in systems multi-antennas, as Kildal & Rosengren (2002) has proposed to evaluate the performance of. The evaluation of multi-antennas systems from the point of view of their spatial correlation is based on the implementation of the virtual radio channel which interact between various antennas simultaneously. The correlation between the envelopes of the induced *e.m f.* between the different antennas is calculated including the interaction with the virtual radio channel. Measured of the radiation patterns of the different antennas are obtained in an anechoic chamber, or they are simulated by means of electromagnetic simulation. It's important to emphasize that it takes into account the mutual coupling between antennas, which are implicit in the tabulated measured of the radiation characteristics (radiation patterns of antennas, directivity, and radiation resistance). The equation of the correlation coefficient between the envelopes associated with the signal terminals of any two antennas of the system is as follows (Hill, 2002):

$$\rho = \frac{s_A(t) \cdot s_B^*(t)}{\sqrt{\left| s_A(t) \right|^2 + \left| s_B(t) \right|^2}} \quad [V] \tag{33}$$

Figure 14 shows the problem of evaluation is a system of two separate and parallel dipoles, please note that both measures are related to the origin of the same coordinate system.
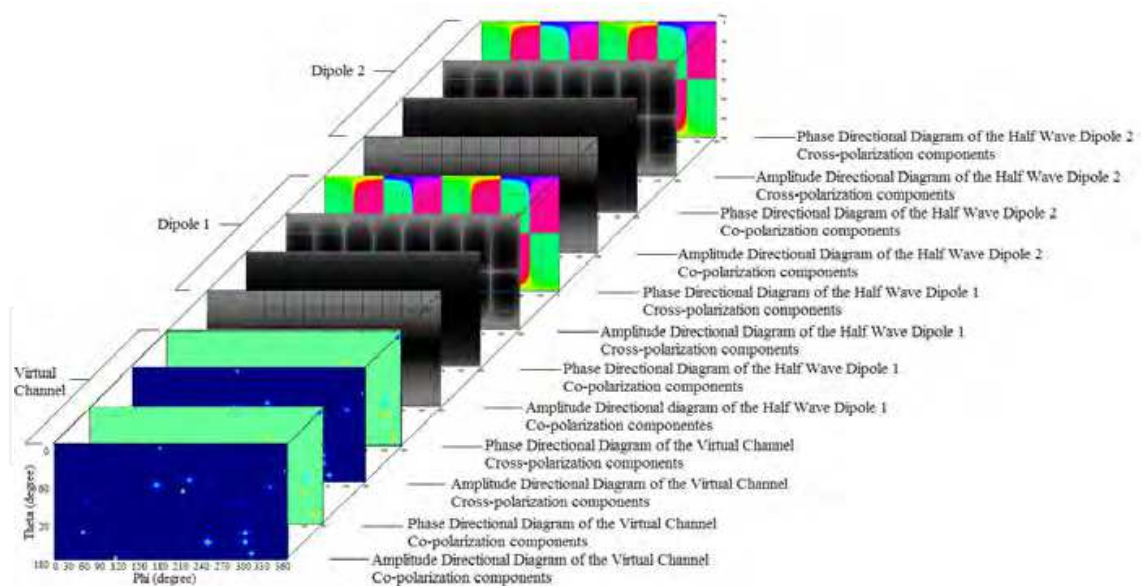

Fig. 14. Geometry of the problem with two antennas under test.

As an example the spatial correlation coefficient as a function of separation between the electric dipoles has been computed and the obtained results are shown in Figure 15.
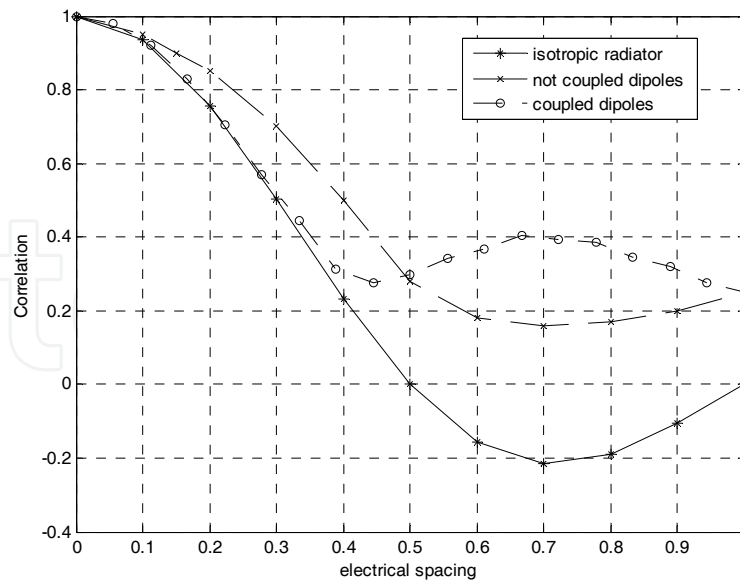
Fig. 15. Correlation coefficients   as a function of spatial separation between antenna elements.

In this figure, there are three graphs. The first one shows the spatial correlation function between two isotropic radiators, without mutual coupling; The second one presents two dipoles without mutual coupling; and third plot shows the correlation between two real dipoles taking into account mutual coupling. The first case of isotropic radiators, shows a *sinc shape*, which corresponds to the well known theoretical studies. For dipole antennas, which were not taken into account the mutual coupling, the plot is nearly a *sinc*. The changes are due the radiation pattern of dipoles are not isotropic, but toroidal, with zones with high radiation and zones where the radiation is entirely null. The plot does not reach the minimum at zero or negative but the general shape of the graph is still approaching the *sinc*. The third plot is linked to the situation of coupled dipoles, is the real case and the explanation of how they are less defined exponentially decreasing is in two fundamental reasons: The first is that the radiators are not isotropic, as explained in the previous situation; the second is that when taking into account the mutual coupling, radiation patterns of the two dipoles are changing because of the electromagnetic interaction between them to vary the separation, so changing the directivity D_{Max} , the radiation resistance R_{Rad} of both dipoles, and their effective lengths l_{Ef} (14). The computations have been performed using a sampled radiation pattern matrix of dimension of 360 x 180. This matrix can include not only simulated values but measured patterns providing an early measurement of the correlation coefficients. Moreover it can evaluate systems built without having to measure them in a reverberation chamber with problems that this implies (time, complexity of the measures, special tools, etc).

4. Radio channels classification

Experience has shown that information about their average values is not sufficient to ensure the quality of performance of radio-communications systems, which takes into account a

number of parameters of great importance to the design, operates and manage the radio system. Moreover, some concepts have been used in the previous sections but they need deep explanations..:

Fading is the sudden variation and reduction of signal received power with respect to its nominal value. This is due to the superposition of waves that arrive by different path. The phenomenon has a basically spatial nature, but the spatial variations of the signal are experienced as temporal variations when the receiver or transmitter they move through the dispersive channel. Figure.16 shows the parameters of the interest for the signal characterization, as: the nominal power received $P_N(dBm)$, the depth of the fading $P_F(dBm)$, and the duration of the fading τ.
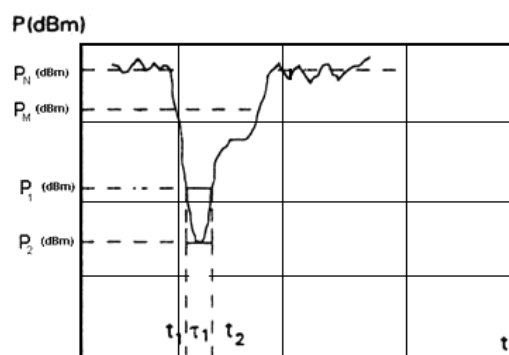
Fig. 16. Fading parameters

Fading performances produces changes in the the spectral characteristics, probability distributions of radio channel. Table 1 shows a classification according to the fading parameters according to the mentioned parameters.

Characteristics	Fading Type	
Depth	Deep	Very Deep
Duration	slow	Fast
Spectral characteristics	flat	Selective
Generating mechanism	k factor	Multipath
Probability distribution	Gaussian	Rayleigh, Rice

Table 1. Fading classification

Table 1, provides various kinds of fading in two columns, within which there is some relationships. Deep fade is usually selective and caused by multipath interference. A flat fading plane appears normally in case of narrow bandwidth producing the same distortion along the carrier spectrum. On the other hand, the selective fading produces different distortions along the spectrum of the modulated signal. Time variation of the desired signal and interference, plays a crucial role in the reliability analysis of a system imposing requirements to the type of modulation, transmission power, protection ratio against interference, diversity techniques, and coding method. That is why; output signal from a radio channel is studied as a random process using statistical methods to characterize them. The radio channels are classified taking the name of the statistical distribution function that

describes the signal obtained. In this way, some typical channels are: *Normal, Gaussian, Rayleigh, Rice, and Nakagami*. In the case of the fading signals at the terminals of the antennas that operate under multipath, Practice has shown that the probability distributions that best fit are: *Rayleigh*, and *Nakagami-Rice*. Then will analyze these distributions (Rec. ITU-R P. 1057-1, 2001).

Rayleigh, when several multipath components with an angle of arrival that are uniformly distributed in the in the range $[0-2\pi]$, the *Rayleigh* distribution describes the fading fast of the signal envelope, both spatial and temporal. Therefore, it can be obtained mathematically as envelope limit of the sum of two noise signals in quadrature with Gaussians distributions. The probability density function, *PDF*, is expressed as follows:

$$PDF = \begin{cases} \dfrac{x}{\sigma^2} \exp\left\{ -\dfrac{x^2}{2\sigma^2} \right\} & x \geq 0 \\ 0 & x < 0 \end{cases} \tag{34}$$

Equation (34), x is the random variable and σ^2 variance or average voltage of the envelope of the received signal. Its maximum value is $\exp(-0.5)/\sigma = 0.6065/\sigma$ and it corresponds to the random variable $x = \sigma$. The cumulative distribution function *CDF* which is given by:

$$CDF = \Pr(x \leq X) = \int_0^X PDF(x)dx = 1 - \exp\left\{ -\frac{X^2}{2\sigma^2} \right\} \tag{35}$$

The average value x_{mean} of the *Rayleigh* distribution can be obtained from the condition:

$$x_{mean} \equiv \int_0^\infty x \cdot PDF(x)dx = \sigma \cdot \sqrt{\frac{\pi}{2}} \approx 1.253 \cdot \sigma \tag{36}$$

while the variance or average power of the signal envelope of the *Rayleigh* distribution can be determined as:

$$\sigma_x^2 = \int_0^\infty x^2 \cdot PDF(x)dx - \frac{\pi\sigma^2}{2} = \sigma^2\left(2 - \frac{\pi}{2} \right) \approx 0.429\sigma^2 \tag{37}$$

The *rms* value of the envelope signal is defined by the square root of $2\sigma^2$, is :

$$rms = \sqrt{2} \cdot \sigma = 1.414 \cdot \sigma \tag{38}$$

The median of the envelope of this signal is defined from the following condition:

$$\frac{1}{2} = \int_0^{x_{median}} PDF(x)dx \tag{39}$$

and it is obtained:

$$x_{median} = 1.177\sigma \tag{40}$$

All these parameters are presented in the x-axis of Figure 17.

Rice distribution appears when several multipath components and a line of sight component are added between the antennas of the transmitter and receiver. A parameter, known as *K factor,* is introduced, which is the rate between the following components:

$$K = \frac{\text{Power of the dominant component}}{\text{Power of the multipath components}} \tag{41}$$

Usually, the *PDF* and *CDF* functions of this distribution are expressed in terms of the *K factor*, as shown:

$$PDF(x) = \frac{x}{\sigma^2} \exp\left\{-\frac{x^2}{2\sigma^2}\right\} \cdot \exp(-K) \cdot I_0\left(\frac{x}{\sigma}\sqrt{2K}\right) \tag{42}$$

and

$$CDF(x) = 1 - \exp\left\{-\left(K + \frac{x^2}{2\sigma^2}\right)\right\} \cdot \sum_{m=0}^{\infty}\left(\frac{\sigma\sqrt{2K}}{x}\right)^m \cdot I_m\left(\frac{x}{\sigma}\sqrt{2K}\right) \tag{43}$$

The *K factor* is represented by the following ratio $K = A^2/2\sigma^2$, where A is the peak voltage of the power or envelope of the dominant component, $I_o(\bullet)$ and $I_m(\bullet)$ are the modified Bessel function of first kind and zero order and m respectively. Figure 17 shows the *Rayleigh PDF* graph and some *PDF* graphs of the several K values. See in this graph (figure 17a) that is asymmetrical bell-shaped, and that with increasing the K value (figure 17b), the graphics are changing so. In the case where $K = 0$, the *Rice* distribution becomes a *Rayleigh* distribution, this is perfectly understandable, because for this value is not presence of dominant component of signal, and is only in the presence of multipath components. When K is increasing the graphics are starting to be tighten and tend to a *Gaussian* distribution. This is a result of the increase in signal level associated with the dominant component. This is the real situation when exist the line-of-sight between the antennas of the transmitter and receiver.
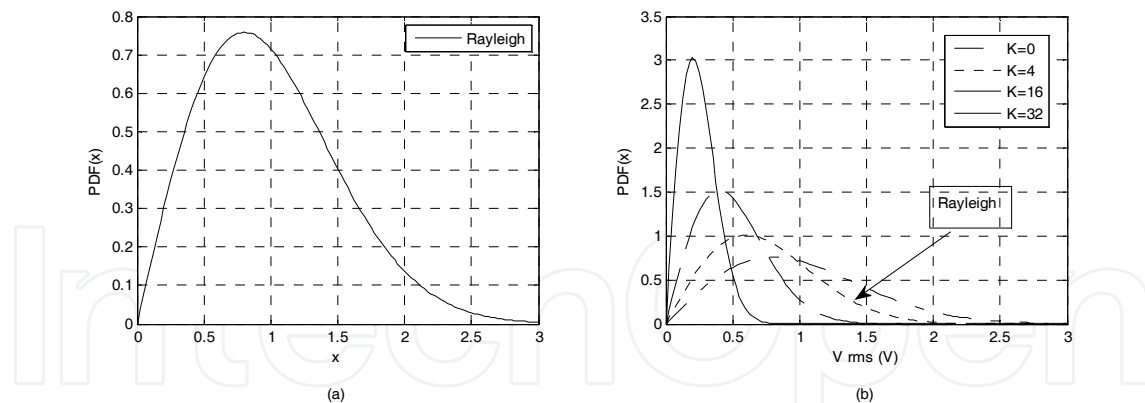
Fig. 17. a) Rayleigh density function and b) *Rayleigh PDF* for the several *K* values

In practice usually identified mixed distributions. Identification of the type given channel fluctuations at the terminals of an antenna probe is of great importance. Knowledge of the channel parameters associated allows selecting the most suitable antenna system, both in the transmitter as the receiver system, allows the configuration of the all signal processing. Knowing the parameter of the channel allow to define the all the structure of the radio system.

5. Conclusions

In this chapter an analysis on the behaviour of antennas in the reception system has been performed. The conclusions are the basis for the approach to the problem of antennas in receiving mode operating in multipath conditions. It was shown that many properties of the antennas are the same in transmission and in reception, which simplifies its study, since it is not necessary to determine these properties in both regimes. Thus, the impedance of the antenna, directional properties, its directivity, efficiency, and gain are the same in both ways of work. Several expressions (mainly the induced *e.m.f*), at the receiving antenna has been obtained. An analysis of the flow of electromagnetic power from the wave travelling in free space and incident at the antenna until you arrive to the receiver, several causes of loss that can occur in this tract were shown, allowing to define conditions to be met to achieve optimal reception: coincidence of the polarization of the antenna with of the incident wave, orientation of the antenna with the direction of arrival of the wave, maximum effective area antenna (length), which depends on its building characteristics, high efficiency, impedances matching between the antenna and feed line. Once the theory and basis on the antenna in reception mode has been stated, the operation under multipath conditions has been covered. A computational procedure that allows see the behaviour an antenna when waves impinge from all directions of space has been presented. The presented procedure has been proved to be useful not only from an educational point of view, but also for assessment multi-antennas system from the viewpoint of spatial correlation, which generalizes its range of applicability. Finally, the importance of classification and identification of radio channels has been explained through a statistical analysis of the fading signal at its terminals. *Rice* and *Rayleigh* channels models were used that best describe the main essence of the multipath. It was observed that when decreasing the signal level of the dominant path this one tends to *Rayleigh*. Knowing the behaviour of antennas under multipath conditions is of significant

importance to the engineers who design, plan and operate radio systems, in order to configure, optimize and select proper system elements, which are ultimately defined by the set-antenna channel.

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