PROXIMITY EFFECTS ON DIVERSITY ANTENNAS

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ABSTRACT Proximity effects constitute a threat to the performance of a diversity antenna when it is operating under physically constricted, limited-area conditions which are typical of indoor environments. Two factors contribute to the degradation of the antenna performance: the reduction of the branch mean signal-to-noise ratios and the increase of the branch signal correlation. The overall loss of diversity gain can be evaluated from these factors. Theoretical investigations of a two-element space diversity antenna mounted on an infinite ground plane and an infinite corner reflector show that proximity effects (i.e. proximity to the ground plane or reflector) can result in a significant loss of diversity gain. The severity of the proximity effects depends on the spacing between the two elements, and on the location and orientation of the pair of elements.

1. Introduction

Antenna diversity has been widely used to combat multipath fading in wireless mobile communications. Basically, a diversity antenna consists of two or more elements, or "branches", that ideally have independent fading characteristics. These branches, all carrying the same message, when properly combined will yield a resultant with greatly reduced severity of fading. Independent fadings can be achieved by making use of phase, polarization or amplitude differences among the branches. In practice, due to lack of independent propagation paths or constraints on the antenna design, it can sometimes be difficult to obtain independent fadings. Furthermore, proximity effects of surrounding objects can become a threat to the performance of a diversity antenna when it is operating under constricted, limit-area conditions, which are typical for indoor environments. In general, compact diversity antennas are desirable for base and remote stations in an indoor environment because of ease of deployment and portability. When the antenna elements are placed close together, mutual coupling increases and the proximity effects of a nearby object are likely to degrade the diversity performance of the antenna. Proximity effects on a diversity antenna can be identified as primarily causing: 1) Distortion of element field patterns, 2) Change of mutual coupling between elements, and 3) Change of element self-impedance. The subsequent effects are reduction of branch mean signal-to-noise ratios (SNR) and increase of signal correlation, which result in an overall loss of diversity gain. In this paper, the proximity effects of an infinite ground plane and an infinite corner reflector on a two-branch space diversity antenna will be investigated.

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2. Loss of diversity gain

The performance of a diversity antenna is most often given in terms of diversity gain, which is the improvement of the SNR of the combined signal in comparison with that of a single branch [1]. The definition of diversity gain is associated with the statistical model of the branch signals and the method of combining them. The Rayleigh model is probably the most commonly used one for describing multipath fading in a mobile communication environment. The cumulative probability (CMP) of the SNR γ of a signal having Rayleigh fading is given by

$$P(\gamma) = 1 - e^{-\frac{\gamma}{\Gamma}} \qquad (1)$$

where Γ is the signal mean SNR.

There are several ways to combine the branch signals of a diversity antenna. The two most popular schemes are perhaps selection diversity and maximum ratio combining. While selection diversity offers practical advantages, maximum ratio combining gives the best statistical reduction of fading of any known linear diversity combiner. As a result, it is usually used as a performance benchmark for diversity systems. Knowing the CMP of the combined signal, its SNR at a given probability level and hence the diversity gain can be determined. When a diversity antenna is operating close to a disturbing object, its diversity gain can be affected in two ways. First, the branch signals may become highly correlated making the characteristics of the combined signal look like that of a single branch. Second, the mean power level of one or more branches may be reduced drastically, bringing down the mean SNR of the combined signal to an unacceptable level. For a two-branch maximum ratio combiner, the reduction of diversity gain due to non-zero branch correlation and changes in branch mean SNR is given approximately as $1/\sqrt{\Gamma_1\Gamma_2(1-\rho^2)}$, (see [2]), where ρ is the correlation coefficient between branch 1 and 2, and Γ_1, Γ_2 are their mean SNR's, respectively. In the extreme cases, where high correlation and large differences among the branch mean SNR's are present, it can be shown that this approximation can yield large errors. Following the general results of Pierce and Stein [3], the exact expression for the CMP of the output SNR can be shown to be

$$P(\gamma) = \frac{c_1}{u_1}[e^{u_1\gamma}-1] + \frac{c_2}{u_2}[e^{u_2\gamma}-1] \quad (2)$$

where

$$u_{1,2} = \frac{-(\Gamma_1+\Gamma_2) \pm \sqrt{(\Gamma_1-\Gamma_2)^2+4\Gamma_1\Gamma_2\rho^2}}{2\Gamma_1\Gamma_2(1-\rho^2)} \quad (3)$$

$$c_{1,2} = \pm\frac{1}{\sqrt{(\Gamma_1-\Gamma_2)^2+4\Gamma_1\Gamma_2\rho^2}} \quad (4)$$

The expression of $P(\gamma)$ for some special cases which leave (2) undetermined, if evaluated directly, are:

(a) $\rho = 1$:

$$1 - e^{-\frac{\gamma}{\Gamma_1+\Gamma_2}} \quad (5)$$

(b) $\rho = 0$ and $\Gamma_1 = \Gamma_2 = \Gamma$:

$$1 - e^{-\frac{\gamma}{\Gamma}}\left(1+\frac{\gamma}{\Gamma}\right) \quad (6)$$

Diversity gain, at a given probability level P (usually chosen to be 0.01) and with the reference to a single branch with Rayleigh fading, is defined as

$$\mathrm{DG} = \frac{\gamma}{\gamma_s} \quad (7)$$

where γ, γ_s are the roots of (1) and (2), respectively. Suppose, due to proximity effects, the branch mean SNR's and correlation coefficient are changed, resulting in a new diversity gain $\mathrm{DG}' = \gamma'/\gamma_s$. Then, the loss of diversity gain (LDG) can be defined as

$$\mathrm{LDG} = \frac{\mathrm{DG}}{\mathrm{DG}'} = \frac{\gamma}{\gamma'} \quad (8)$$

Diversity gain, in a strict sense, should be defined as the improvement in the output SNR with reference to a given branch, when more branches are added without changing the mean SNR of the first one. However, the definition in (7) is more general and its related definition of LDG given by (8) will be used in this paper to give a figure of merit for proximity effects on a diversity antenna.

3. Proximity effects on a two-branch space diversity antenna

Consider a two-branch diversity antenna as shown in Fig. 1 with impedance and load matrix given by

$$\mathbf{Z}_A = \begin{bmatrix} Z_{11} & Z_{12} \\ Z_{21} & Z_{22} \end{bmatrix}, \qquad \mathbf{Z}_L = \begin{bmatrix} Z_1 & 0 \\ 0 & Z_2 \end{bmatrix} \quad (9)$$

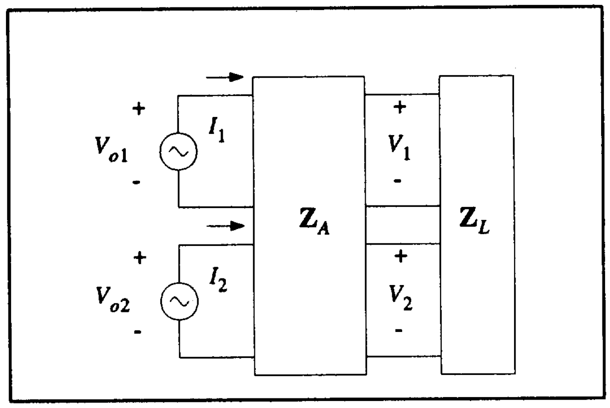

Figure 1. Equivalent circuit of a two-branch diversity antenna.

Denoting the open circuit voltage vector with $\mathbf{V}_o^T = (V_{o1}, V_{o2})$, one can define the open circuit voltage correlation matrix as

$$\mathbf{L}_o \equiv \langle \mathbf{V}_o\mathbf{V}_o^H \rangle = \begin{bmatrix} \langle V_{o1}V_{o1}^* \rangle & \langle V_{o1}V_{o2}^* \rangle \\ \langle V_{o2}V_{o1}^* \rangle & \langle V_{o2}V_{o2}^* \rangle \end{bmatrix} \quad (10)$$

where T denotes transpose, * complex conjugate, H Hermitian transpose and $\langle\ \rangle$ ensemble average. The loaded circuit voltage vector $\mathbf{V}$ and its corresponding correlation matrix $\mathbf{L}$ are defined in a similar manner. It can be shown easily that

$$\mathbf{L} \equiv \langle \mathbf{V}\mathbf{V}^H \rangle = \mathbf{Z}_L\mathbf{Y}\mathbf{L}_o\mathbf{Y}^H\mathbf{Z}_L^H \quad (11)$$

where $\mathbf{Y} = (\mathbf{Z}_A + \mathbf{Z}_L)^{-1}$. The correlation coefficient of the loaded circuit voltages is

$$\rho = \frac{\langle V_1V_2^* \rangle}{\sqrt{\langle V_1V_1^* \rangle\langle V_2V_2^* \rangle}} \quad (12)$$

Assuming the loads have the same noise power N, the branch mean SNR's are given by

$$\Gamma_m = \mathrm{Re}\,\{\frac{1}{2N}\langle V_m I_m^* \rangle\} = \mathrm{Re}\,\{\frac{1}{2NZ_m^*}\langle V_m V_m^* \rangle\}, \quad m = 1, 2 \qquad (13)$$

In a multipath environment, usually it can be assumed that the amplitude and phase of the incident waves are spatially uncorrelated. For a diversity antenna with only θ-polarization as in the following analysis, it can be shown that [4]

$$\langle V_{om} V_{on}^* \rangle = C \oint E_{\theta m}(\Omega) E_{\theta n}^*(\Omega) S_\theta(\Omega)\, d\Omega, \quad m, n = 1, 2 \qquad (14)$$

where C is a proportionality constant, $\Omega = (\theta, \varphi)$, $E_{\theta m}(\Omega)$ is the far field pattern of the mth element, and $S_\theta(\Omega)$ is the angular distribution of the incident waves.

3.1 Proximity effects of an infinite ground plane

Consider a space diversity antenna consisting of two vertical half-wave dipoles separated by a spacing s as shown in Fig. 2(a). For simplicity, assume the incident waves are omnidirectional and confined to the horizontal plane. For each dipole loaded with a 75 ohm impedance, the envelope correlation coefficient ρ_e, being approximately equal to $|\rho|^2$ for Rayleigh fading [5], is calculated as function of dipole spacing and plotted in Fig. 3 as the solid curve. It is seen that a low value of $\rho_e = 0.27$ can be achieved for $s = 0.1\lambda$. Compared with the case of zero correlation, this corresponds to a LDG of 0.6 dB. As a rule of thumb, only values of $\rho_e > 0.6$ give rise to a significant loss.

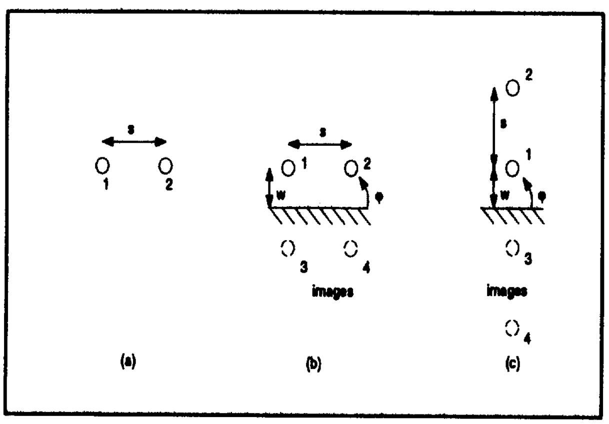

Figure 2. (a) Two-branch space diversity antenna. (b), (c) Different geometries for ground-plane proximity to the same antenna.

The antenna is now mounted on an infinite ground plane as shown in Fig. 2(b). The configuration, by using image theory, is equivalent to an array of 4 dipoles. The impedance matrix and the far field patterns of the branches are modified to

$$Z_A' = \begin{bmatrix} Z_{11} - Z_{13} & Z_{12} - Z_{14} \\ Z_{21} - Z_{23} & Z_{22} - Z_{24} \end{bmatrix} \qquad (15)$$

$$E_{\theta 1}'(\Omega) = E_{\theta 1}(\Omega) - E_{\theta 3}(\Omega), \qquad 0 \le \varphi \le \pi \qquad (16)$$

$$E_{\theta 2}'(\Omega) = E_{\theta 2}(\Omega) - E_{\theta 4}(\Omega), \qquad 0 \le \varphi \le \pi \qquad (17)$$

where Z_{mn} is the mutual impedance between dipoles m and n, and $E_{\theta m}(\Omega)$ is the far field pattern of dipole m. For a fixed dipoles-to-plane distance of $w = \lambda/8$, ρ_e is recalculated and plotted in Fig. 3 as the dashed curve. It is seen that $\rho_e = 0.61$ for $s = 0.1\lambda$ corresponding to a LDG of 1.9 dB. Another mounting configuration is shown in Fig. 2(c), where the array axis of the dipoles is perpendicular to the ground plane. For a fixed distance from the first dipole to the ground plane $w = \lambda/8$, ρ_e is recalculated and plotted in Fig. 3 as the dotted curve. The proximity effects are seen to be even stronger than in the previous case with $\rho_e = 0.97$ for the same value of s, which corresponds to a LDG of 6.5 dB. The significant increase in the signal correlation is due to the increasing mutual coupling and co-phasing effect of the ground plane. For this orientation of the antenna, the phase differences in the field patterns disappear and their amplitudes have similar shapes when the antenna is close to the ground plane.

Up to this point, only losses due to nonzero signal correlation have been discussed. The contribution of power reduction in the branches to the total LDG is also significant. This is particularly true for many situations where proximity effects may cause a strong mismatching to one or more branches of the diversity antenna. For the 3 cases considered above, the branch mean SNR's are calculated as functions of dipole spacing and normalized to the mean SNR of a dipole in free space. The results are shown in Fig. 4. In case (a), $\Gamma_1 = \Gamma_2 = -2.1$ dB. This reduction is solely due to mutual coupling between the dipoles since no ground plane is present. In case (b), the proximity effects of the ground plane reduce the branch mean SNR's to -3.5 dB. As can be seen from (15), the self-impedance of dipole 1 has been reduced from Z_{11} to $Z_{11} - Z_{13}$, worsening the mismatching. In case (c), $\Gamma_1 = -6.2$ dB and $\Gamma_2 = -1.9$ dB. Even though dipole 1 is at the same distance from the ground plane as in case (b), Γ_1 is greatly reduced because most of the power of the incident waves has been absorbed by dipole 2 which is located at a larger distance from the ground plane.

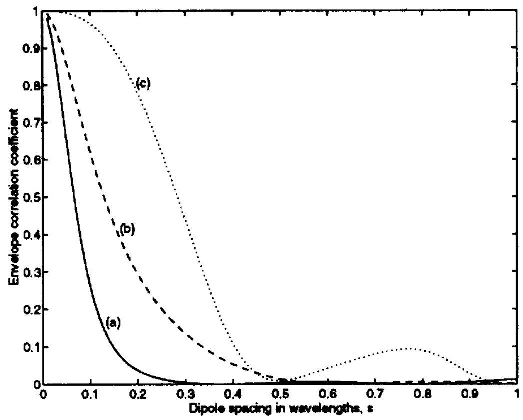

Figure 3. Envelope correlation coefficient as a function of dipole spacing for cases (a), (b) and (c) in Fig. 2. $w = \lambda/8$

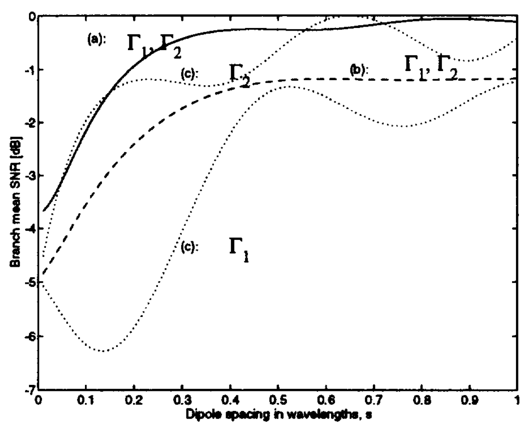

Figure 4. Branch mean SNR's as functions of dipole spacing for cases (a), (b) and (c) in Fig. 2. $w = \lambda/8$.

The total LDG for each of the above 3 cases is plotted in Fig. 5 with reference to a space diversity antenna with infinite dipole spacing. As expected, case (c) is worst with an LDG of more than 10 dB at $s = 0.1\lambda$. The key contribution to this loss is the high signal correlation. It should be noted here that the contributions to the total loss from the signal correlation increase and the branch mean SNR reductions are not additive (in dB), especially when the signal correlation is high and there is a large difference in the branch mean SNR's. As the dipole spacing increases to infinity, the losses approach 1.2 dB in cases (b) and (c). This is no surprise since the dipoles are still close to the ground plane so mismatching remains while mutual coupling and signal correlation vanish.

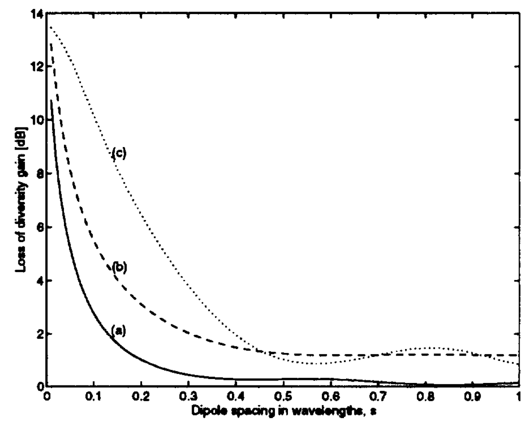

Figure 5. Loss of diversity gain as a function of dipole spacing for cases (a), (b) and (c) in Fig. 2. $w = \lambda/8$.

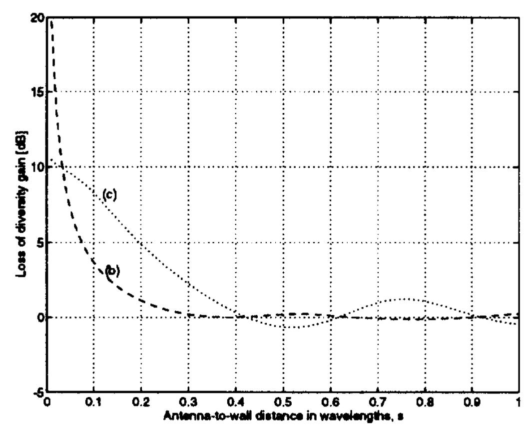

Figure 6. Loss of diversity gain as a function of dipole-to-ground plane distance for cases (b) and (c) in Fig. 2. $s = 0.1\lambda$.

It is seen that a space diversity antenna with 0.1λ dipole spacing, while working reasonably well in free space conditions, may have its performance severely degraded near a ground plane. For $s = 0.1\lambda$, the variation of the LDG with respect to the dipole-to-ground plane distance is shown in Fig. 6. Here, the LDG for cases (b) and (c) is plotted with reference to case (a), where the antenna is in free space, to illustrate the proximity effects of the ground plane. Up to the ground plane, the loss in case (b) is extremely large since both dipoles are strongly mismatched resulting in vanishing power reception. However, the loss decays quickly away from the ground plane and becomes negligible for $w > 0.3\lambda$. In case (c), only dipole 1 can approach the ground plane, whereas dipole 2 is always at least 0.1λ away, so for small w the loss is not as large as in case (b). On the other hand, it decreases more slowly due to the high signal correlation. It is interesting to note that the loss is negative at $w = 0.5\lambda$, implying an improvement in diversity gain. The oscillatory behavior of the loss is largely due to the infinite dimension of the ground plane. In practice, for finite ground planes, the loss will probably decay more quickly, with or without oscillation. One can define a *proximity distance* at which the loss is less than a given small value. This is the distance the antenna should be kept away from the ground plane so that no significant proximity effects are felt.

3.2 Proximity effects of an infinite corner reflector

Similar to the idea of the infinite ground plane, an infinite right-angled corner reflector has the simple geometry of two infinite ground planes intersecting at a right angle, but nevertheless presents an approximation to practical configurations. For instance, it can be approximately the corner of an office at which the diversity antenna of an indoor base station is mounted. The infinite dimension of the reflector is adequate in simulating the office's corner since the proximity effects are negligible just a few wavelengths from the corner, so that the far walls of the office do not significantly influence the antenna performance.

Consider the same diversity antenna with 0.1λ dipole spacing as above. The antenna is mounted in an infinite corner reflector with the general geometry shown in Fig. 7. By image theory, the impedance matrix and the far field patterns of the branches are

$$\mathbf{Z}_A' = \begin{bmatrix} Z_{11}-Z_{13}+Z_{15}-Z_{17} & Z_{12}-Z_{14}+Z_{16}-Z_{18} \\ Z_{21}-Z_{23}+Z_{25}-Z_{27} & Z_{22}-Z_{24}+Z_{26}-Z_{28} \end{bmatrix} \quad (18)$$

$$E_{\theta 1}'(\Omega) = E_{\theta 1}(\Omega) - E_{\theta 3}(\Omega) + E_{\theta 5}(\Omega) - E_{\theta 7}(\Omega), \quad 0 \le \varphi \le \frac{\pi}{2} \quad (19)$$

$$E_{\theta 2}'(\Omega) = E_{\theta 2}(\Omega) - E_{\theta 4}(\Omega) + E_{\theta 6}(\Omega) - E_{\theta 8}(\Omega), \quad 0 \le \varphi \le \frac{\pi}{2} \quad (20)$$

where Z_{mn} is the mutual impedance between dipoles m and n, and $E_{\theta m}(\Omega)$ is the far field pattern of dipole m. In order to get a general picture of the proximity effects of the reflector, we look at three different orientations of the antenna: $\alpha = 0°$, $\alpha = 45°$, and $\alpha = 135°$. For each case, the variation of the LDG on the antenna position is calculated and shown as a contour plot in Figs. 8, 9 and 10, respectively.

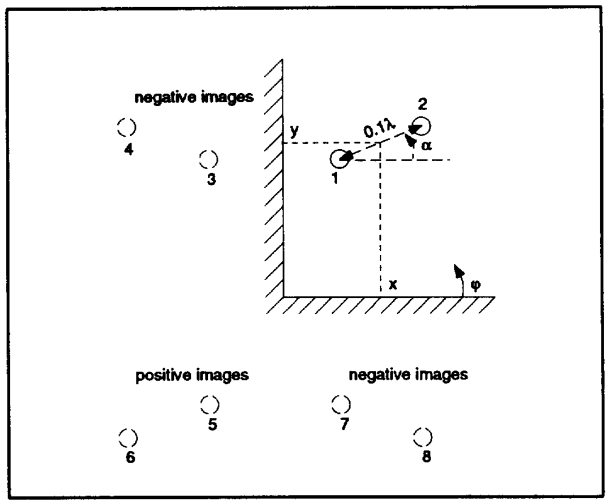

Figure 7. Geometry of space diversity antenna in infinite right-angled corner reflector.

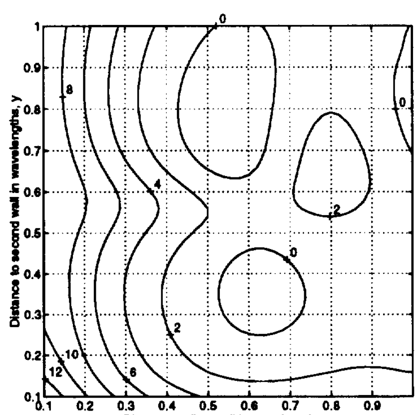

Figure 8. Contour plot of loss of diversity gain [dB] for space diversity antenna in corner reflector (see Fig. 7). Antenna midpoint is varied and dipole spacing is fixed at 0.1λ. $\alpha = 0°$.

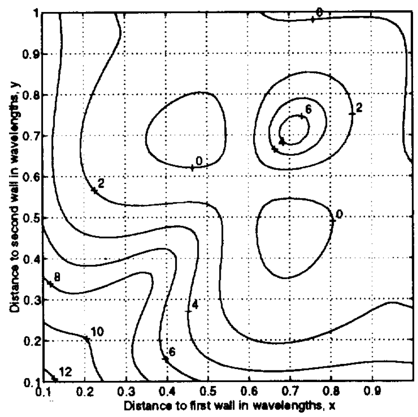

Figure 9. Same as in Fig. 8. $\alpha = 45°$.

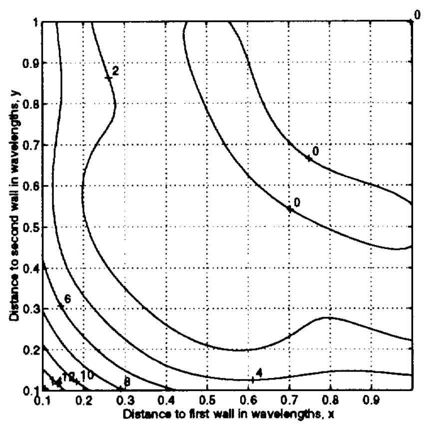

Figure 10. Same as in Fig. 8. $\alpha = 135°$.

In the first case, Fig. 8 shows a strip of high loss along the vertical wall. This is similar to the proximity effects of an infinite ground plane shown in Fig. 2(c). The loss variation is more complicated due to the proximity effects of the horizontal wall. In the second case shown in Fig. 9, the loss is very high at the corner of the reflector and at a region about three quarters wavelength from the walls, which is largely due to the very high signal correlation (not shown) in these regions, especially along the bisector of the reflector where the field patterns are highly similar. The proximity effects are weakest in the third case as shown in Fig. 10, where only a small region at the corner has LDG > 6 dB. The signal correlation in this case is low due to the dissimilar asymmetry of the field patterns around the bisector. In all three cases, the loss is generally higher than in the case of the infinite ground plane.

4. Conclusions

The proximity effects of an infinite ground plane and an infinite corner reflector on a two-branch space diversity antenna have been investigated theoretically. It was shown that the diversity performance of an antenna was severely degraded when operating close to these disturbing objects. The severity of the proximity effects depends on the antenna orientation, dipole spacing and distance from the disturbing object. In general, it was found that the proximity effects are significant when the diversity antenna with 0.1λ dipole spacing is within a distance of 0.3λ from the ground plane and 0.5λ from the walls of the corner reflector. Better performance can be achieved by increasing the dipole spacing. In this investigation, the loss of diversity gain, determined from the exact CMP of a maximum ratio combiner, has been used as a figure of merit for the proximity effects.

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