

Coupling Modeling and Reduction Techniques of Cavity-Backed Slot Antennas: FDTD Versus Measurements

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Abstract—Cavity-backed slot (CBS) antennas are commonly used in earth-based and space-borne applications. In this manuscript, the finite-difference time-domain (FDTD) method is used to analyze different antenna characteristics of such elements including input impedance, coupling, and radiation patterns. Emphasis is given on coupling calculations especially on different methods to reduce coupling. Specifically, coupling reduction is attempted using lossy material superstrates or ground plane discontinuities, such as slits. Numerical issues concerning the analysis of CBS antennas in the context of FDTD are also discussed.

Index Terms—Cavity-backed slot antennas, coupling modeling, coupling reduction, FDTD.

I. INTRODUCTION

C AVITY-BACKED slot (CBS) antennas have been widely used in different types of applications within the microwave band, including radars, satellite communications, mobile telephony, broadcast TV, and aircraft/spacecraft communications. They are relatively easy to manufacture, lightweight and often small in size. Their low profile is an important characteristic, especially for aircraft, missile and spacecraft applications, because they can be flush-mounted on the surface of airborne vehicles without affecting the vehicle's aerodynamic profile.

Coupling is an important factor in today's communication systems which have become more complex and use a large number of antennas to support all required services. The reliability and integrity of communication systems can be significantly affected by coupling between transmitting and/or receiving elements that are mounted on the same structure such as a helicopter or an aircraft airframe. Especially when several antennas are collocated on the same structure of finite dimensions, interaction, referred to as cosite interference, can deteriorate the quality of communications and corrupt the signals with noise. This leads to an extremely critical EMI/EMC problem. In addition, airborne communications require extremely reliable systems that sustain continuous operation without jamming or interruption. In such airborne communication systems, the installation of a new antenna appears to be a

challenging task. Its mounting location is influenced not only by the type of service the antenna will provide, but also by the mounting locations of the existing antennas on the airframe. Therefore, the interactions between the new antenna and other radiating elements should be analyzed in order to optimize its position, such that coupling satisfies specific requirements. It can be concluded that coupling mechanisms are very important in the design of reliable communication systems.

CBS antennas have been analyzed in numerous papers using different methods. Two of the early efforts to analyze such antennas were presented by Calejs in [1] and by Adams in [2]. The equivalence principle along with the generalized network formulation have been used in [3] to form the integral equation for a CBS antenna which was then solved by a method of moments formulation. The finite-difference time-domain (FDTD) method was applied to the analysis of a single CBS antenna by Omiya *et al.* [4]. Furthermore, coupling between two CBS antennas was examined in [5] using FDTD combined with the Kirchhoff transformation. However, in this paper, although using FDTD, the various issues related to computational time and accuracy are discussed and resolved using appropriate modeling. Furthermore, calculations are performed to examine the dependence of coupling on the frequency of operation. In addition, coupling-reduction techniques are also presented including incorporation of lossy superstrates and ground plane slits.

Lossy materials have been used to suppress the surface currents induced on conducting bodies and reduce the radar cross section of structures, such as airplanes or antennas. In [6], a lossy coating was used to suppress the undesirable side-lobes of a paraboloidal reflector antenna. Also, lossy materials have been used in the design of horn antennas to improve side-lobe levels, equalize the E - and H -plane radiation patterns, and reduce the cross-polarized radiation [7]–[10]. In [11], the contour path FDTD was used to analyze pyramidal horn antennas with and without composite E -plane inner walls. Planar and cylindrical arrays of open-ended waveguides have been analyzed in the presence of a lossy ground in [12]. In [13], the radiation of a rectangular waveguide mounted on a lossy flange was examined, and it was shown that the lossy flange improves the E -plane radiation pattern as well as the cross-polarized radiation. Here, lossy superstrates are used to suppress surface currents and reduce coupling between antennas that are flush-mounted on a ground plane. Moreover, the effect of distance and orientation on coupling is examined through FDTD simulations and measurements. All the different coupling-reduction

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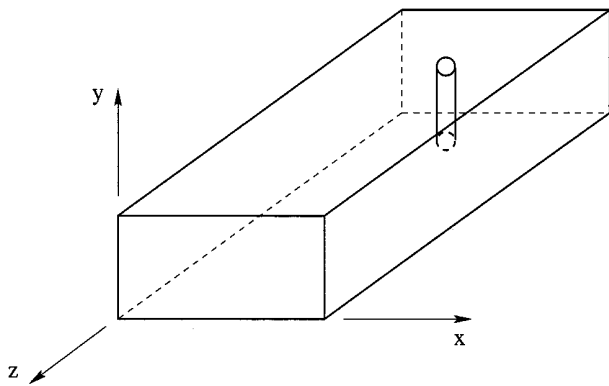


Fig. 1. A 3-D view of an air-filled rectangular CBS antenna fed with a probe oriented in the y -direction.

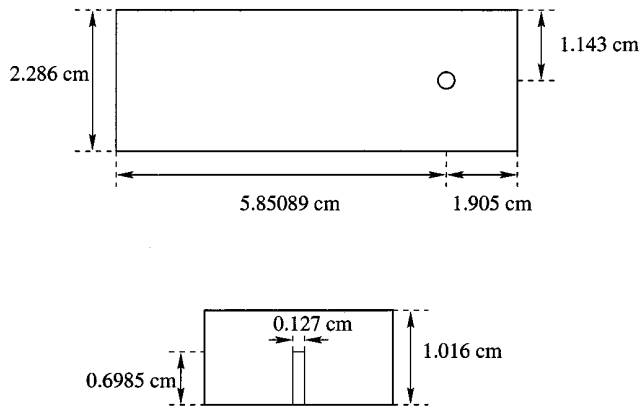


Fig. 2. A two-dimensional view of an air-filled rectangular CBS fed with a probe oriented in the y -direction.

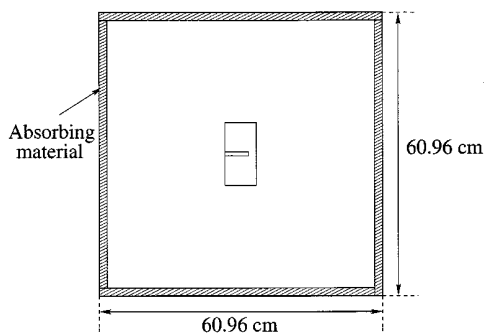


Fig. 3. A top view of a CBS mounted on a ground plane (experimental configuration).

techniques are compared through their radiation characteristics, such as gain, efficiency and coupling reduction. The numerical results are validated by comparison with measurements and a few available finite-element method (FEM) calculations.

II. FDTD MODELING

FDTD analysis of electromagnetic problems usually involves post-processing that transforms the time-domain data to the frequency domain. This transformation is performed using a discrete Fourier transform (DFT) procedure. However, in order for the DFT to give accurate results, the FDTD simulation time must be chosen long enough for all the transient phenomena

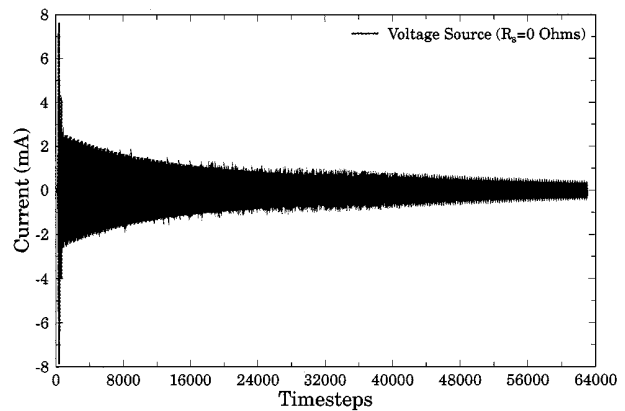


Fig. 4. Current of the probe exciting the CBS antenna for $R_s = 0$.

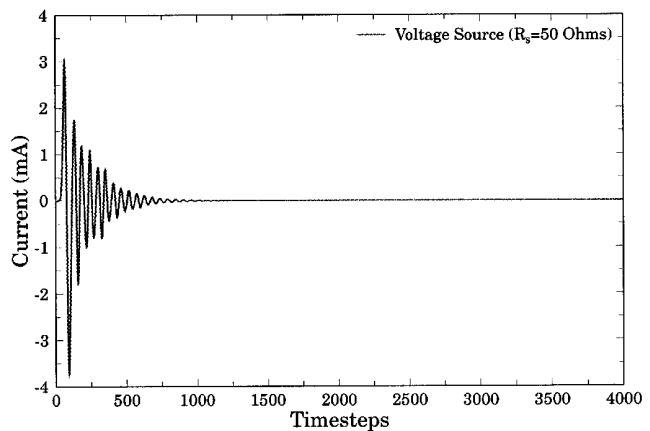


Fig. 5. Current of the probe exciting the CBS antenna for $R_s = 50 \Omega$.

to decay. This requirement can be restrictive, especially in highly resonant and high-quality factor (Q) structures where the transients may need a prohibitive simulation time to decay. A very effective technique that resolves this problem is based on a source with an internal resistance which provides the excitation [14]–[16]. The basic characteristics of FDTD are well documented and will not be repeated here. Instead, the emphasis will be on the modeling of the CBS antennas, their radiation characteristics, and coupling-reduction techniques.

In order to illustrate the effectiveness of a voltage source with an internal resistance, it was decided to compute the input impedance of an air-filled CBS antenna originally analyzed in [17]. A three-dimensional (3-D) view of the cavity under consideration is exhibited in Fig. 1 and a detailed description of the geometry is shown in Fig. 2. Notice that the CBS antenna is formed by an open-ended standard X-band waveguide. The input impedance of the CBS antenna was measured in the ElectroMagnetic Anechoic Chamber facility at Arizona State University. In the experiment, the aperture antenna was mounted on a finite ground plane of dimensions 24×24 in (60.96×60.96 cm), and the edges were covered with a strip of absorbing material to reduce diffractions (see Fig. 3). Additionally, the antenna was rotated with respect to the principal ground plane axes and offset relative to the center of the ground, to direct the edge diffractions away from the aperture. Moreover, the input

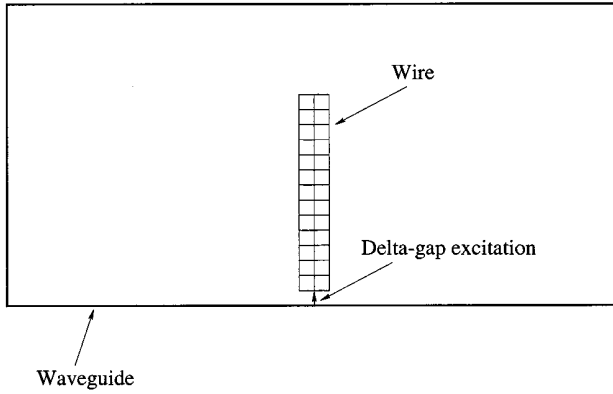


Fig. 6. Cross-section of the FDTD mesh of the CBS antenna for a cell size of 0.6 mm.

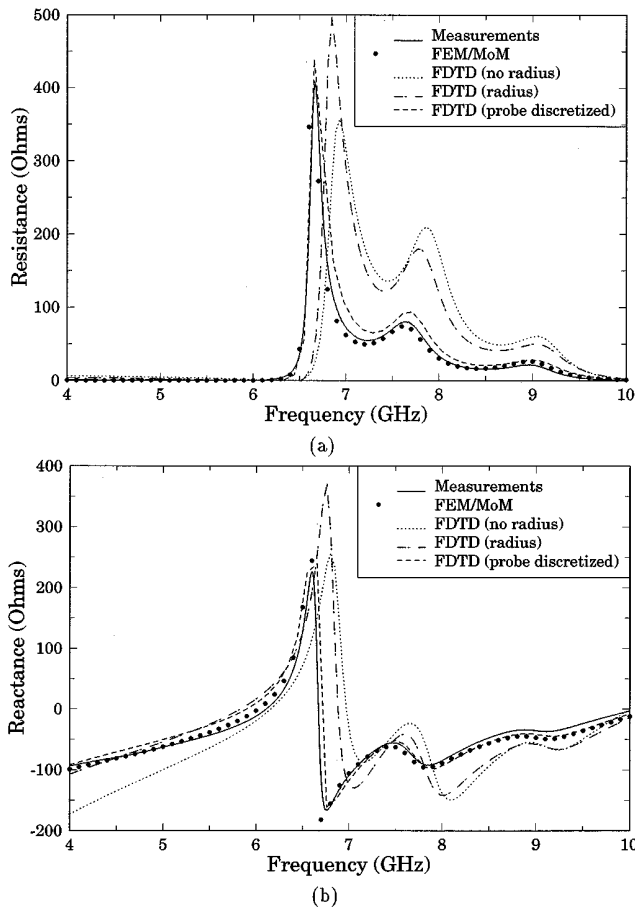


Fig. 7. Input impedance of a CBS antenna. (a) Resistance. (b) Reactance.

impedance of the same aperture mounted on an infinite ground plane was calculated in [17] using a hybridization of the FEM with the moment method (MoM) and compared very well with measurements. In this paper, it was shown that the dimensions of the ground plane do not have a profound effect on the value of the input impedance.

In the FDTD simulations, this aperture antenna was mounted on a 9×9 cm finite ground plane, which is smaller than the one used in the measurements, to reduce the size of the computational space. In order to determine the input impedance of the

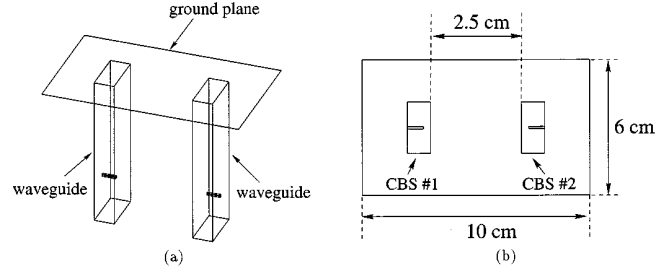


Fig. 8. Geometry of two identical CBS antennas mounted on a rectangular ground plane (for antenna specifications see Fig. 2). (a) 3-D view. (b) Top view.

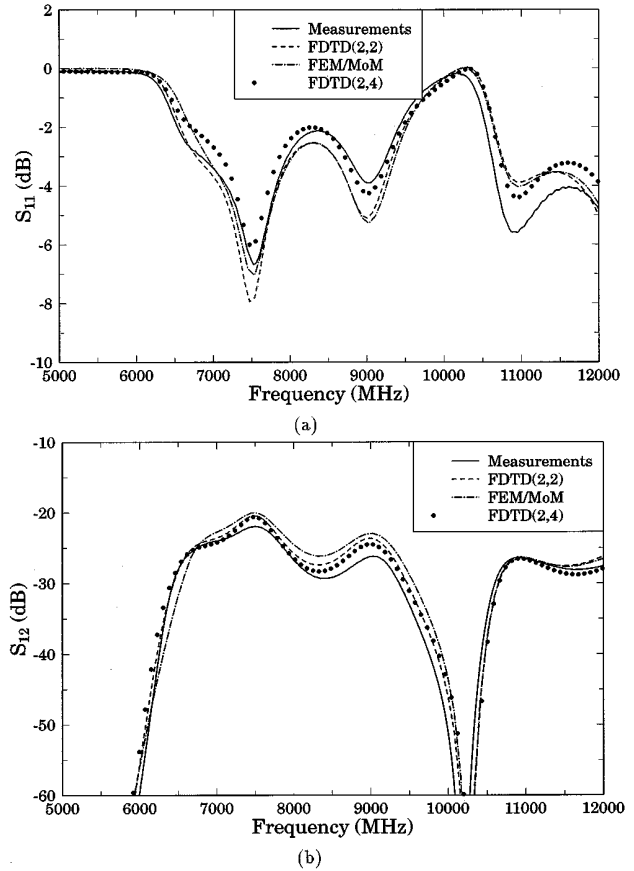


Fig. 9. S -parameters of two identical CBS antennas mounted on a rectangular ground plane (for antenna specifications see Fig. 1.4).

antenna, the voltage and current at the feeding probe of the CBS antenna have to be computed. The simulation time should allow the transient effects in both the voltage and current to decay, so that they can be accurately Fourier transformed. The voltage is a user defined function of time that decays quickly to zero, whereas the current may need a long time to decay. Two FDTD simulations were performed, and for both, the cell size was 1.5 mm. In the first case, the voltage source had zero internal resistance ($R_s = 0$), and in the second case, R_s was set equal to 50Ω . Figs. 4 and 5 show the computed current at the probe exciting the cavity for the two cases, respectively. Obviously, the current in the case with no internal resistance ($R_s = 0$) did not decay to zero even after 64 000 time steps, indicating the resonant behavior and the high-quality factor, Q , of the antenna. On the

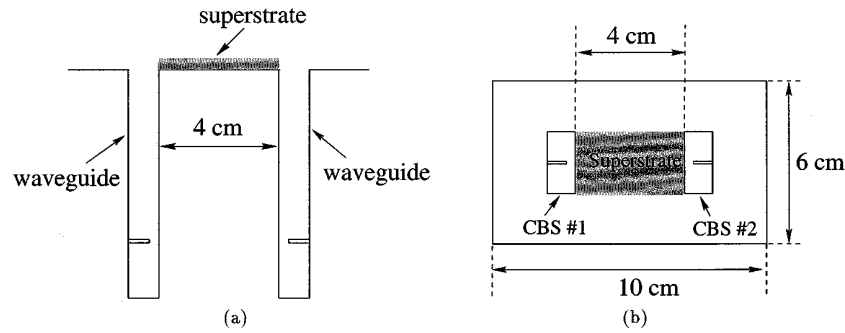


Fig. 10. Geometry of two identical CBS antennas mounted on a rectangular ground plane (for antenna specifications see Fig. 2). (a) Side view. (b) Top view.

contrary, in the second simulation ($R_s = 50 \Omega$), the source current converged to zero very fast and the FDTD calculation time reduced significantly. Based on our experience, the use of an internal resistance is essential for efficient simulations of CBS antennas. The internal resistance is usually chosen close to the characteristic impedance of the transmission line and the larger it is, the faster the transient phenomena decay. However, the internal resistance cannot be chosen to be very large since it may lead to instabilities due to the interaction of the source with the numerical scheme or due to the neglect of the displacement current through the FDTD cell which includes the source [16].

III. IMPEDANCE

In order to examine the accuracy of FDTD results which are computed using a source with an internal resistance, the input impedance of the CBS antenna examined in Section 2 was computed for three different cases. In all the cases, the feeding probe was excited by a voltage source with R_s equal to 50Ω . In the first case, the radius of the probe (0.0635 cm) was not modeled and the cell size was 1.5 mm. In the second case, the radius was taken into account by using the thin-wire model and the cell size was 1.5 mm. In the third case, the cell size was 0.6 mm and the probe itself was discretized along with the remaining geometry (see Fig. 6). Fig. 7 illustrates the computed input resistance and reactance of the aperture antenna for the three different cases. Also, the FDTD calculations are compared with measurements as well as with results obtained by the FEM/MoM hybrid formulation reported in [17]. Evidently, the accuracy of the FDTD predictions depends greatly on the wire modeling of the probe that excites the antenna. Excellent agreement between the FDTD computations and measurements is observed in the case where the probe was discretized (third case). The improved accuracy of this third FDTD simulation can be attributed to the finer discretization and the enhanced modeling of the probe.

IV. COUPLING

The coupling between two identical CBS antennas (whose specifications are defined in Fig. 2) mounted on a rectangular 10×6 cm ground plane, 25 mm apart from each other, was also analyzed. The geometry of the two cavities along with the dimension specifications are shown in Fig. 8. FDTD simulated the physical dimensions of the finite ground plane whereas the FEM/MoM hybrid method simulated an infinite ground plane.

The S parameters of the two antennas were calculated using both methods: FDTD and FEM. S_{11} and S_{22} correspond to the reflection coefficient of each antenna, and they are equal because the two antennas are identical and the geometry is symmetrical. Furthermore, S_{12} and S_{21} correspond to the coupling between the antennas, and they are also identical due to reciprocity. The numerical results of both methods are compared with measurements in Fig. 9, and they all exhibit very good agreement. The discrepancies at the higher end of the band are attributed to discretization errors. In these simulations the FEM mesh consisted of 75 874 elements with an average edge size 0.16 cm, whereas the FDTD used a cell size of 0.6 mm and a computational domain of $160 \times 160 \times 142$ cells. Notice that two different FDTD methods were used. The first one, denoted as FDTD(2, 2), represents a second-order accurate FDTD, both in-time and in-space. The second one, denoted as FDTD(2, 4), represents a second-order accurate in-time and fourth-order accurate in-space FDTD. It can be observed that the FDTD(2, 4) scheme does not provide a dramatic improvement in accuracy compared to the FDTD(2, 2) scheme. However, it is expected that as the distance between the two apertures becomes large in terms of the wavelength, FDTD(2, 4) should outperform FDTD(2, 2).

V. REDUCTION OF COUPLING

In this section, different methods of coupling reduction are examined. The geometry under examination is shown in Fig. 10. The separation between the two apertures of the CBS antennas is 4 cm, and the space between the two apertures is covered with a superstrate. Initially, coupling was calculated using FDTD without any superstrate, and this case is labeled "blank" for reference purposes. Then, a superstrate of lossy material was placed on top of the ground plane, filling the space between the two apertures. The superstrate was constructed using a number of layers of 1.5-mm thick ECCOSORB GDS electric/magnetic composite material. Three different superstrates were used with thicknesses of 1.5, 3.0, and 4.5 mm consisting of one, two or three GDS material layers, respectively. Coupling computations were again performed by FDTD for all three superstrates. Although the GDS material does not have constant material properties along the entire band of interest, its properties do not exhibit large variations. Therefore, in the FDTD simulations, constant material properties (found by averaging the maximum and minimum values) were used. These properties are: $\epsilon_r =$

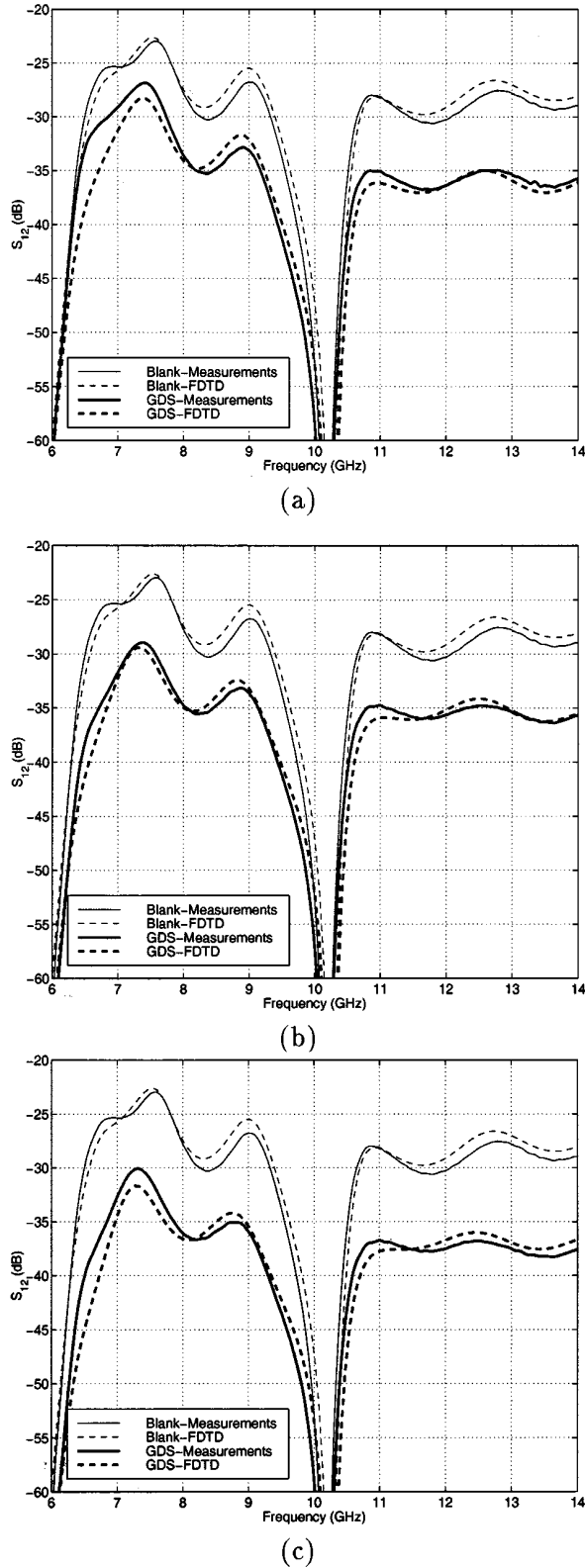


Fig. 11. Coupling with and without lossy superstrate. (a) 1.5-mm thick GDS. (b) 3.0-mm thick GDS. (c) 4.5-mm thick GDS.

13.4, $\sigma = 0.17$ S/m, $\mu_r = 1.5$, and $\sigma^* = 95\,000$ Ω /m, where σ and σ^* are the electrical and magnetic conductivities, respectively.

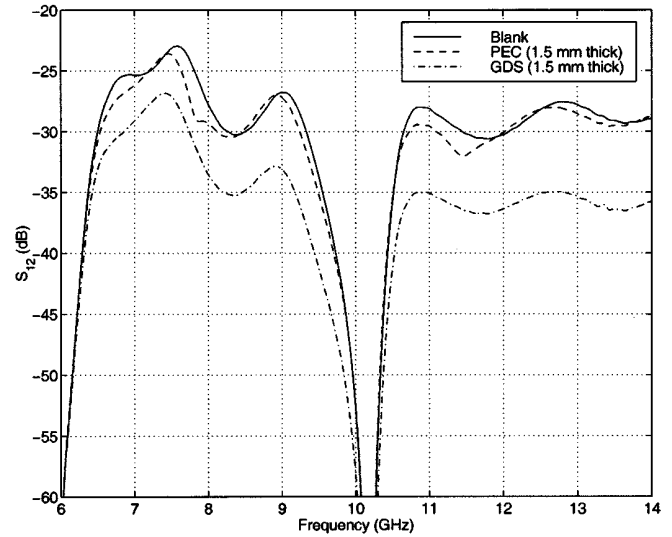


Fig. 12. Measured coupling for three different cases.

The FDTD predictions of coupling for the GDS material are illustrated in Fig. 11 along with the predictions for the case with no superstrate (blank case), and they are also compared with the corresponding measurements. The FDTD calculations for all cases compare very well with measurements, even at levels below -30 dB with maximum discrepancies ranging from 1 to 2 dB. It is also observed that adding the lossy superstrate substantially reduces coupling. Specifically, the maximum coupling decreased 4 dB for the 1.5-mm thick GDS, 6 dB for the 3.0-mm thick GDS, and 7 dB for the 4.5-mm thick GDS (the maximum coupling for each case is defined as the maximum over the entire band of frequencies). This reduction of coupling can be attributed to the minimization of surface currents on the ground plane which couple energy from one CBS antenna to the other. In addition, to examine if the blockage created by the physical presence of the superstrate contributes to the coupling reduction, an 1.5-mm thick piece of PEC (perfect electric conductor) was placed between the apertures, replacing the lossy superstrate. The measured coupling for this case is plotted in Fig. 12 and compared with measurements of the equal thickness GDS superstrate and the case with no superstrate (blank). Apparently, the inclusion of the PEC piece does not reduce the maximum coupling, even though it introduces maximum blockage. Therefore, it can be concluded that the reduction of coupling in the case of the lossy superstrate is mainly attributed to the minimization of the surface currents and not the blockage.

To better understand the mechanism of coupling reduction in the case of lossy superstrates, the induced surface current densities on the ground plane of the two CBS antennas were computed for two cases: (a) with a one-layer (1.5-mm thick) GDS superstrate, and (b) no superstrate, at 11.92 GHz. In both cases only CBS #1 was excited (see Fig. 10). Figs. 13 and 14 illustrate the normalized total surface current densities on the ground plane in dB. It is apparent that, in the case with no loss (see Fig. 13), the surface current densities in the area between the two apertures are intense (as high as -15 dB), thereby establishing a strong coupling mechanism. However, in the case

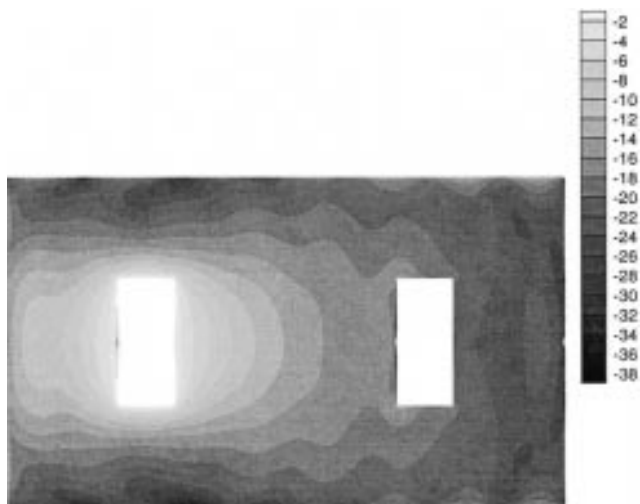


Fig. 13. Total surface current density J_{tot} in decibels, with no lossy superstrate at 11.92 GHz.

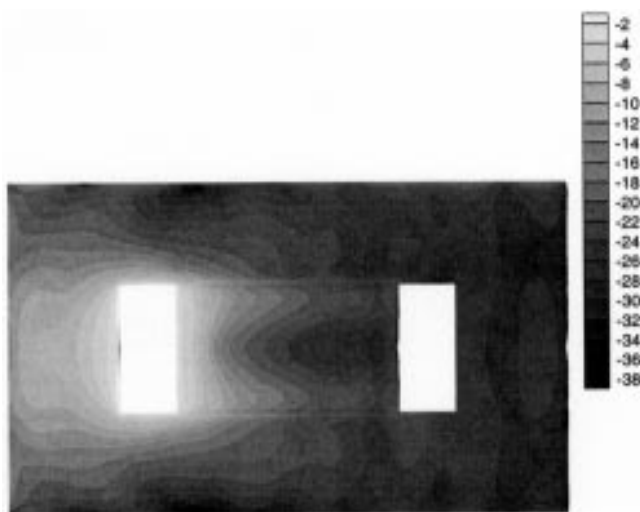


Fig. 14. Total surface current density J_{tot} in decibels, with a 1.5-mm thick GDS superstrate at 11.92 GHz.

with the lossy superstrate (see Fig. 14), the surface current densities in the area between the two apertures have been significantly reduced, exhibiting levels as low as -25 dB. Therefore, it is clear that lossy superstrates weaken the coupling mechanism supported by surface current densities.

Another method of reducing coupling due to surface currents is to introduce a discontinuity, such as a thin slit, on the ground plane between the two apertures. A 2-mm wide slit was symmetrically cut at the center of the ground plane, as shown in Fig. 15. The FDTD calculations for this case are plotted against measurements in Fig. 16(a), and they exhibit excellent agreement. During our coupling studies, the width of the original slit (2 mm) was increased to 4 mm and 8 mm, sequentially. The FDTD predictions for these cases are illustrated and compared in Fig. 16(b). It is observed that the wider the slit is, the smaller the coupling becomes. However, the reduction of coupling did not change dramatically when the width of the slit became two or four times larger than the original 2-mm wide slit. Moreover,

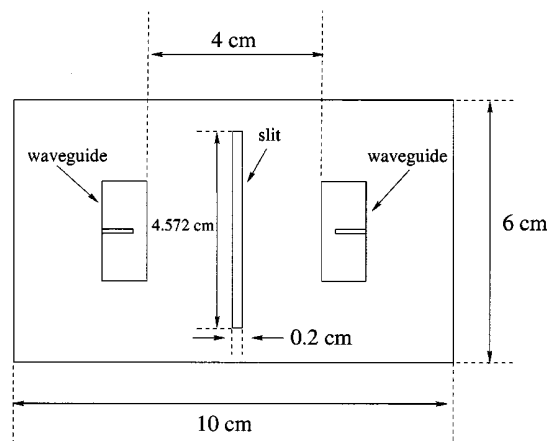


Fig. 15. Geometry of two CBS antennas on a ground plane with a slit.

it appears that as the slit becomes wider, the reduction of coupling is larger at higher frequencies. This can be possibly attributed to the fact that the slit is electrically larger at higher frequencies thereby disrupting the surface currents on the ground plane for a larger electrical distance. Specifically, the 2-mm and 8-mm slits are $\lambda/20$ and $\lambda/5$ wide, respectively, at 7.5 GHz. However, the same slits, 2 mm and 8 mm, are approximately $\lambda/12$ and $\lambda/3$ wide, respectively, at 12 GHz. The incorporation of a 2-mm slit caused a 1-dB reduction of the maximum coupling, as illustrated in Fig. 17, which presents the comparison of four different cases of measured [see Fig. 17(a)] or calculated [see Fig. 17(b)] coupling. Conclusively, the largest reduction of coupling occurs for the thickest GDS superstrate of 4.5 mm, whereas the smallest reduction happens for the slit. Therefore, a lossy superstrate can be very effective in minimizing the coupling between aperture antennas by reducing the surface currents on the ground plane. Moreover, superstrates are very practical because they can be flush mounted, and they do not affect the mechanical strength of the ground plane as slits do.

Disorienting the two CBS antennas is another technique that can be used to reduce coupling. The configuration of the aperture antennas examined above (seen in Fig. 10) is called the E -plane configuration. The H -plane configuration is formed by rotating both aperture antennas 90° (see Fig. 18). Pozar in [18] examined the mutual coupling between two rectangular microstrip patches for both E -plane and H -plane configurations. He demonstrated that the E -plane configuration exhibits the smallest coupling for electrically-small separation distances, while the H -plane configuration exhibits the smallest coupling for electrically large separation distances. Therefore, it is expected that coupling of the H -plane configuration of two CBS antennas will be much smaller than the coupling of the E -plane configuration, especially for separation distances that are not very small in terms of wavelength. This mechanism of coupling reduction is also discussed in [19], and it can be explained by comparing the coupling of two side-by-side versus two collinear dipoles. Alternatively, the pattern of a CBS antenna in the H -plane nearly vanishes along the ground plane due to boundary conditions. In the E -plane, the pattern does not have to vanish along the ground plane unless the dimension of the aperture in that plane is a multiple of a wavelength [19].

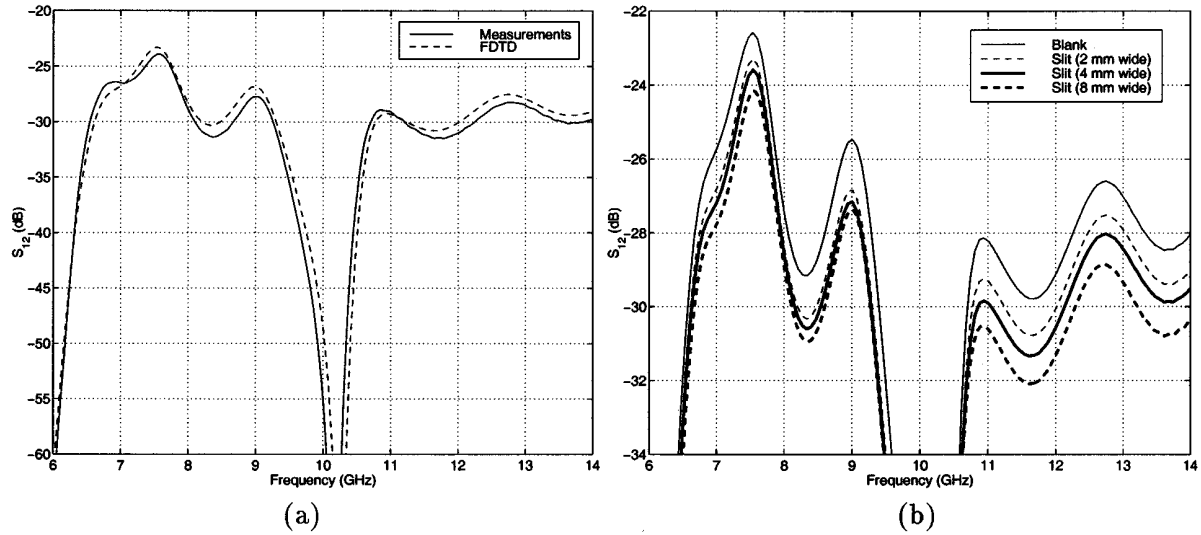


Fig. 16. Coupling with a slit on the ground plane. (a) 2-mm wide slit. (b) FDTD predictions for different widths of the slits.

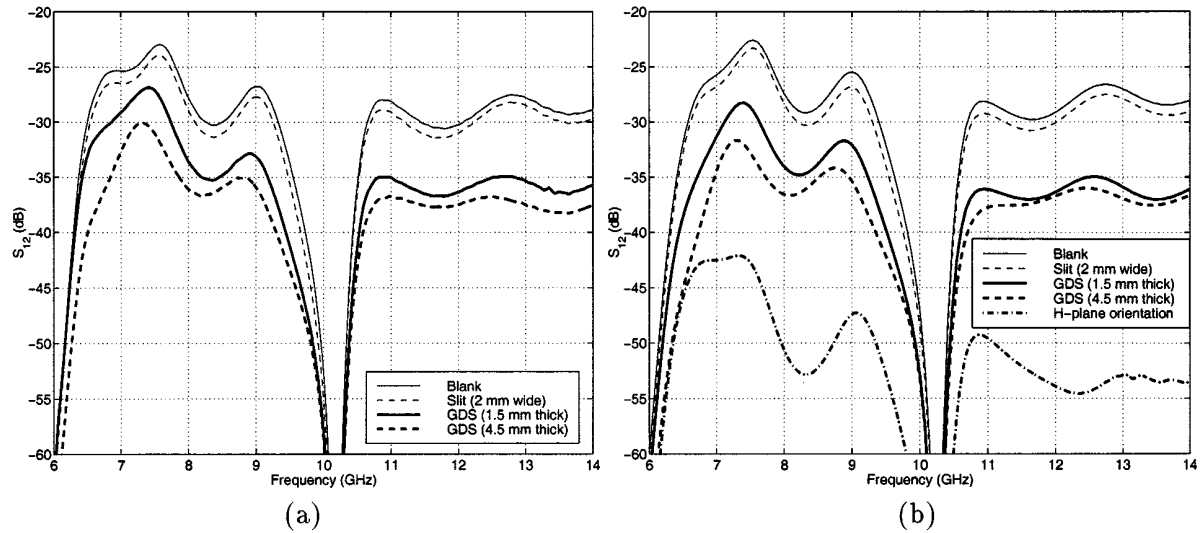


Fig. 17. Coupling between two CBS antennas. (a) All the measurements. (b) All the FDTD predictions.

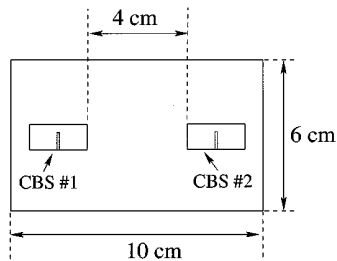


Fig. 18. Top view of the two CBS antennas oriented along the H -plane.

Hence, the mutual effects between two CBS antennas which are oriented along the H -plane are less intense because the antennas are placed in the direction of minimum radiation. One FDTD simulation was performed to verify this line of thought by calculating the coupling between two CBS antennas oriented along the H -plane (see Fig. 18). The results of this simulation are illustrated in Fig. 17(b). It is observed that the

coupling of the H -plane configuration is 20 dB less than the coupling of the E -plane configuration (labeled “blank”), which is a substantial reduction.

Finally, dependence of coupling on the separation distance between two CBS antennas is examined. Coupling was measured for various separation distances by placing the two antennas on a larger ground plane (1×1 m). The measured results are illustrated in Fig. 19(a) for a range of distances from 25 to 160 mm at increments of 15 mm. As expected, coupling decreases as the separation between the two antennas becomes larger. FDTD simulations were also performed for a few selected cases of separation distances (25, 40, and 55 mm), and the computed results are illustrated in Fig. 19(b). The predictions agreed very well with the measurements and exhibited similar variation. Also, the measured coupling was plotted at four selected frequencies versus the separation distance in Fig. 20. The coupling decreased following a variation similar to $1/r$ at all four frequencies, exhibiting a total reduction of about 14 dB

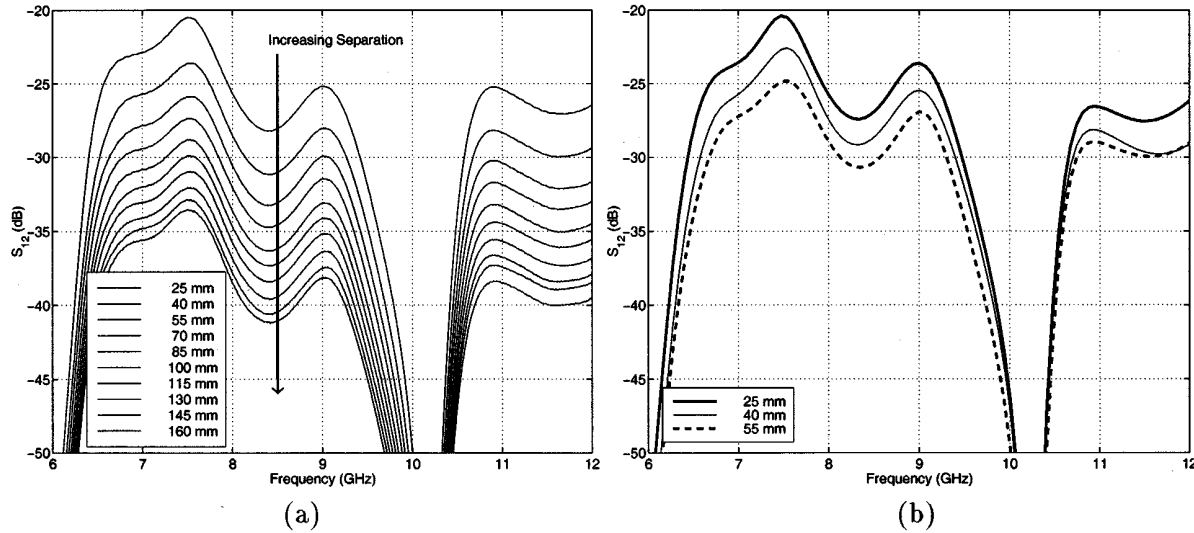


Fig. 19. Coupling between two CBS antennas for different separation distances. (a) Measurements. (b) FDTD predictions.

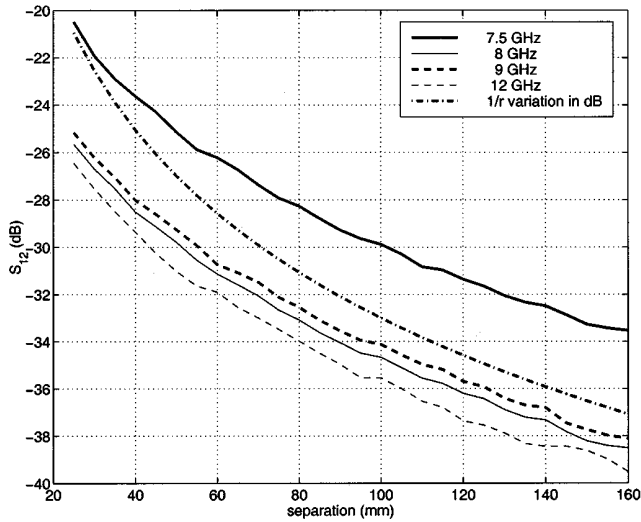


Fig. 20. Measured coupling versus separation distance at different frequencies.

when the separation was increased from 25 to 160 mm (see Fig. 20).

VI. PATTERNS

In the previous section, various methods of minimizing coupling between two CBS antennas were investigated. Lossy superstrates, as well as slits, on the ground plane were implemented and analyzed. However, these alterations of the physical geometry also change the radiation characteristics of the antennas. Especially, the presence of composite superstrates introduces losses which can potentially reduce the antenna gain. Radiation patterns in the E -plane were computed using FDTD for two cases: a) with a one-layer (1.5-mm thick) GDS superstrate and b) no superstrate, at two frequencies of 7.5 and 11.92 GHz. In both cases, only CBS #1 was excited (see Fig. 10). The GDS material properties used in the single-frequency FDTD

simulations were picked to be equal to the measured material properties at the corresponding frequency. These properties are: 1) at 7.5 GHz, $\epsilon_r = 13.4$, $\sigma = 0.09$ S/m, $\mu_r = 1.75$, and $\sigma^* = 74\,000\,\Omega/\text{m}$, and 2) at 11.92 GHz, $\epsilon_r = 13.4$, $\sigma = 0.25$ S/m, $\mu_r = 1.35$, and $\sigma^* = 110\,000\,\Omega/\text{m}$. The numerical results are illustrated in Figs. 21 and 22, and the predictions are compared with measurements. All radiation patterns were calculated as well as measured by exciting only one of the two CBS antennas. The agreement between the measurements and the numerical calculations is very satisfactory and the maximum discrepancy ranges from 1 to 2 dB. Finally, the measured gain of the antennas for the cases with and without the lossy superstrate are compared for both frequencies of 7.5 and 11.92 GHz in Fig. 23. It is observed that the introduction of a lossy superstrate causes a reduction of the antenna gain which can be as high as 5 dB for some observation angles. Consequently, when incorporating lossy materials, there is a trade-off between minimizing the coupling and maintaining high efficiency. Furthermore, the E -plane radiation patterns of the two CBS antennas on the ground with and without the 2-mm slit are essentially identical (especially when the observation angle is on the top of the ground plane, i.e., for elevation angles $-90^\circ \leq \theta \leq 90^\circ$) and, therefore, they are not illustrated and compared through a figure.

VII. SUMMARY OF COUPLING REDUCTION METHODS

Different coupling reduction techniques were presented and analyzed. Traditionally, the most intuitive way to reduce coupling between two antennas is to increase their separation distance. Measurements and simulations illustrated the decreasing behavior of coupling with increasing separation distances (see Section V) which was very similar to a $1/r$ variation (see Fig. 20).

Other more novel techniques of coupling reduction were also examined, such as the introduction of ground discontinuities, e.g., slits, and the use of lossy superstrates between the two antennas. Slits were found to be less effective in reducing coupling

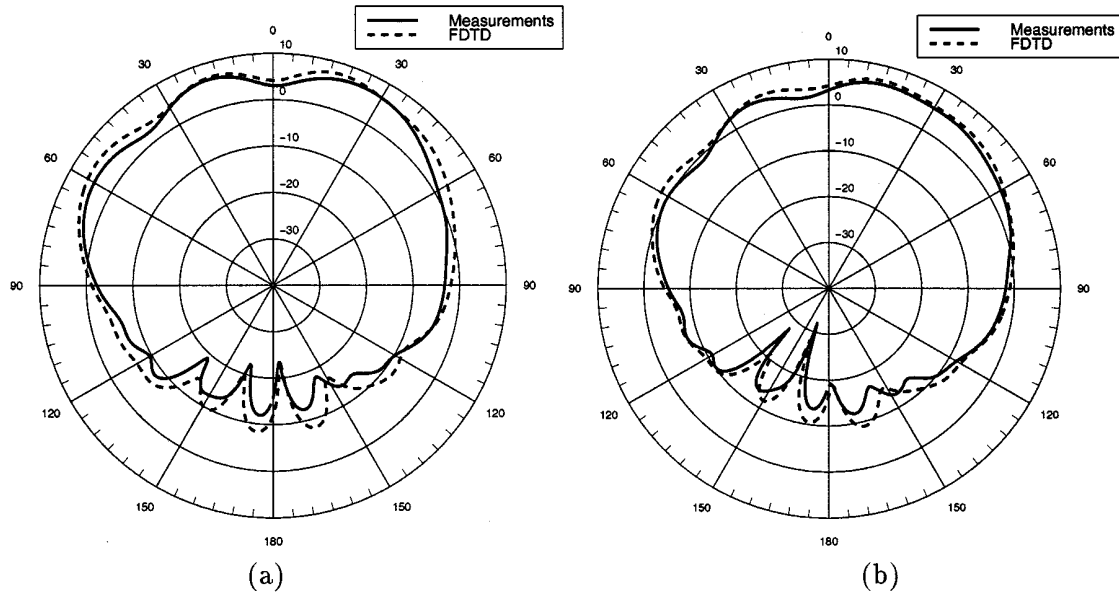


Fig. 21. Patterns at 7.5 GHz with (a) no superstrate and (b) 1.5-mm GDS thick superstrate.

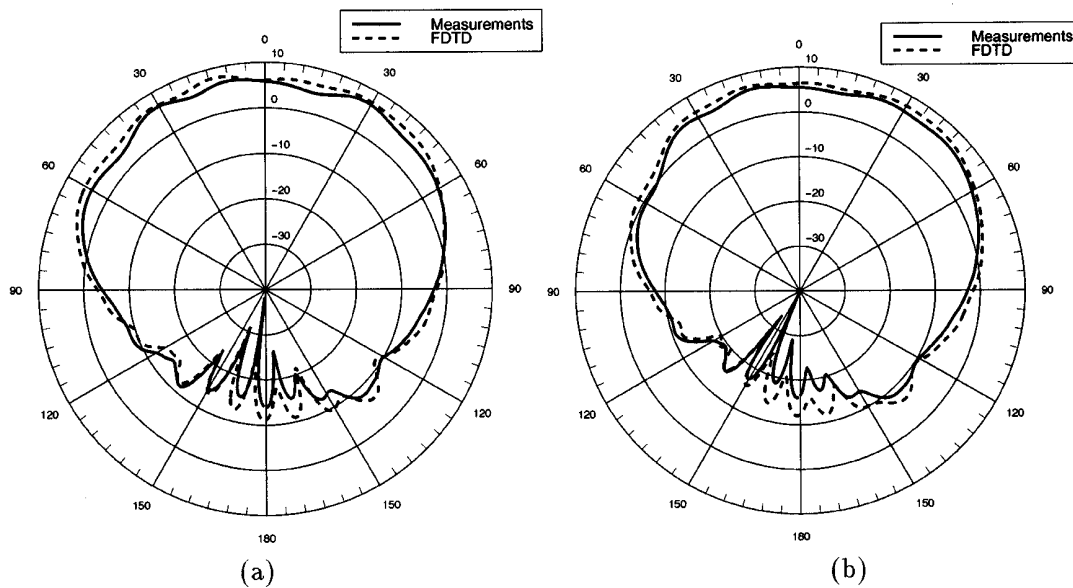


Fig. 22. Patterns at 11.92 GHz with (a) no superstrate and (b) 1.5-mm GDS thick superstrate.

than lossy superstrates. However, the incorporation of lossy materials significantly influences the radiation characteristics of an antenna. To illustrate the different advantages and disadvantages of using lossy superstrates, some of the results are tabulated and they correspond to the configuration of the two CBS antennas examined in the previous sections. Gain and efficiency computations were performed by exciting only CBS #1 (see Fig. 10).

In Table I, the antenna efficiency is shown for the three different thicknesses of GDS material used in our above analyses at 7.5 and 11.92 GHz. At 7.5 GHz, the efficiency ranges from 70%–74%, and at 11.92 GHz it ranges from 83%–88%. In Table II, the antenna gain at an elevation angle of $\theta = -60^\circ$

(where the minus corresponds to the left side of the elevation pattern) is shown for different superstrate thicknesses. It is also compared to the antenna gain with no lossy superstrate present at 7.5 and 11.92 GHz. The elevation angle of $\theta = -60^\circ$ was chosen for our comparison because, at this angle, the gain exhibited the larger decrease when lossy materials were used (see Fig. 23). When the thickest GDS superstrate (4.5 mm thick) was used, the gain decreased 4.6 dB at 7.5 GHz and 6.1 dB at 11.92 GHz compared to the gain with no superstrate. Notice though that the gain toward zenith was almost the same for the cases with or without lossy superstrates (see for example Fig. 23).

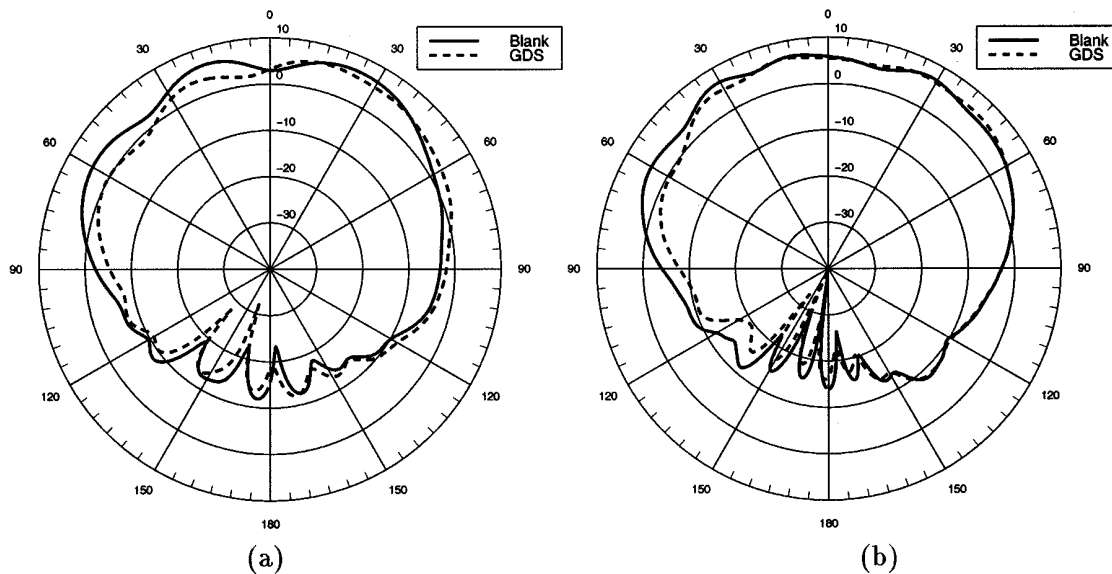


Fig. 23. Measured patterns. (a) At 7.5 GHz. (b) At 11.92 GHz.

TABLE I
EFFICIENCY OF THE TWO CBS ANTENNAS CONFIGURATION FOR DIFFERENT THICKNESSES OF THE GDS SUPERSTRATE

Frequency	Thickness of GDS superstrate		
	1.5 mm	3.0 mm	4.5 mm
7.5 GHz	74%	73%	70%
11.92 GHz	88%	85%	83%

TABLE II
GAIN OF THE TWO CBS ANTENNAS CONFIGURATION AT ELEVATION ANGLE $\theta = -60^\circ$

Frequency	No Loss	Thickness of GDS superstrate		
		1.5 mm	3.0 mm	4.5 mm
7.5 GHz	5.8 dB	1.9 dB	2.0 dB	1.2 dB
11.92 GHz	5.6 dB	1.2 dB	0.8 dB	-0.5 dB

TABLE III
COUPLING REDUCTION OF THE TWO CBS ANTENNAS CONFIGURATION

Frequency	Thickness of GDS superstrate		
	1.5 mm	3.0 mm	4.5 mm
7.5 GHz	3.9 dB	6.2 dB	7.8 dB
11.92 GHz	6.1 dB	5.4 dB	7.2 dB
Average Coupling Reduction over the band 6-14 GHz	5.8 dB	6.4 dB	8.2 dB

Apparently, the use of lossy superstrates yields a decrease in the antenna gain and efficiency but provides a coupling reduction in return. The achieved coupling reduction for the different superstrates is depicted in Table III. These coupling reductions were obtained by comparing the cases with and without lossy superstrates. At 7.5 GHz, the coupling was reduced by 3.9 dB for the thinnest superstrate of 1.5 mm, and by 7.8 dB for the thickest superstrate of 4.5 mm. At 11.92 GHz, the coupling was reduced by 6.1 dB for the thinnest superstrate of 1.5 mm, and by 7.2 dB for the thickest superstrate of 4.5 mm. Additionally, the average coupling reduction over the entire band of 6–14 GHz

TABLE IV
COUPLING REDUCTION OF THE TWO CBS ANTENNAS CONFIGURATION

Frequency	Width of the ground plane slit		
	2.0 mm	4.0 mm	8.0 mm
7.5 GHz	0.7 dB	1.0 dB	1.5 dB
11.92 GHz	0.9 dB	1.5 dB	2.4 dB
Average Coupling Reduction over the band 6-14 GHz	1.0 dB	1.4 dB	2.5 dB

was 5.8 dB for the thinnest superstrate of 1.5 mm, and 8.2 dB for the thickest superstrate of 4.5 mm. A similar table is constructed for the cases that incorporated a slit in the ground plane (see Table IV). As mentioned earlier, the coupling reduction in the cases of the slits is not as significant as in the cases of the lossy superstrates.

Radiation characteristics, such as coupling, efficiency and gain, are critical to the design of an antenna and must be taken into consideration. In a practical configuration there is always a compromise between coupling reduction and maintenance of high efficiency and gain. Decisions regarding a design should take into account the communication system operation requirements.

VIII. CONCLUSION

In this paper, the analysis of CBS antennas in the context of FDTD was presented. Different numerical issues related to the modeling of such antennas were described. It was found that the use of a voltage source with an internal resistance in FDTD is indispensable for efficient computations. Both FDTD and FEM/MoM were used to compute the input impedance of a single CBS antenna and coupling between two elements, and they compared very well with measurements. However, it was found that for coupling versus distance calculations, the hybrid FEM/MoM is faster than the pure FDTD. This is attributed

to the fact that FEM/MoM does not discretize the space between the two cavities, but rather treats the open-space interaction within the MoM formulation. Therefore, the FEM/MoM computational space remains constant because only the CBS antennas need to be discretized in the hybrid methodology. On the contrary, FDTD has to also account for the space between the two antennas thereby yielding very large domains, especially for large distances.

A very comprehensive study of coupling between two CBS antennas was also performed through measurements and simulations. The dependence of coupling on the separation distance between the two antennas was investigated. Disorienting the antennas was also found to be a very effective method of coupling reduction. Specifically, the H -plane configuration exhibited significantly smaller levels of coupling than the E -plane configuration. Moreover, other alternative methods of reducing coupling were examined including incorporation of lossy superstrates as well as ground plane discontinuities, such as slits. It was illustrated that lossy material superstrates can be very effective in decreasing the coupling between two antennas, but they cause a reduction of the antenna gain. In the summary section, some of these trends were reported. Therefore, in a practical design there is always a compromise between reduction of coupling and maintenance of a sufficiently high gain.

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