Investigation of Radiation Gain Enhancement Techniques for Microstrip Patch Antenna in Multilayer Superstrates

Michael F. Adamson[1,2], Ivor L. Morrow[2]
[1]DSTL, Porton Down, Salisbury, Wiltshire SP4 0JQ.
[2]EWIC, Cranfield University, Bedfordshire, MK43 0TR.

Abstract

An analytical methodology is described that gives insight into the behaviour of a microstrip patch antenna covered by multiple planar dielectric and magnetic superstrates. The expressions for the electric field in TM_{01} mode are combined with a multilayer media transmission line model to investigate the effect on the radiation pattern and gain. Design parameters such as relative permittivity or permeability value, thickness of the superstrate layer, and layer separation are optimised to maximise the antenna gain. Conditions for significant improvements in boresight gain were identified from the model and demonstrated by radiation pattern measurements on some multilayered prototype antennas.

I. Introduction

A previous paper by the authors [1] described a computational electromagnetic characteristic mode analysis for a microstrip patch antenna (MSA), covered by planar dielectric superstrates to enhance the radiation pattern gain. A traditional cavity mode model was used to determine the initial resonance and impedance bandwidth for the MSA. MSA with multiple planar superstrates were simulated using CST Microwave Studio and several prototype antennas were manufactured, of the general form shown schematically in Figure 1. Experimental measurements of impedance bandwidth, dielectric superstrate height, and radiation pattern showed reduced gain compared to that predicted by simulation. In the case of a double dielectric superstrate antenna configuration, the main lobe of the measured radiation pattern exhibited pronounced distortion, shown in Figure 2. The finite-difference time-domain CST model and experimental measurements gave little or no practical insight into the cause of the discrepancies between theory and measurement.

Previous researchers have also used other full-wave analysis methods to design and optimise material choice, thickness, and separation in these multilayered MSA[2]. This brute-force approach is unwieldy, providing limited understanding and often leads to time-consuming redesign if material layers are changed.

Here we describe the development of an equivalent transmission line model (TLM) that includes the patch antenna and the radiation model for an arbitrary dielectric/ferrite cascaded structure. The TLM enables rapid evaluation and understanding of the various effects that material selection and layer spacing have on the resultant radiated field pattern of the radiating system. A parametric investigation is conducted for a multilayer patch antenna operating at 10 GHz. The effect of key design parameters on the radiation pattern and the boresight gain is quantified numerically, and the effect of variation in the permittivity of the superstrate layer, the thickness of the superstrate layer(s), dimensions of the antenna, and air gaps are studied. Analytical results are presented illustrating these various effects, and some configurations of more practical use are highlighted for manufacturing and subsequent laboratory measurement.

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The results of CST full-wave simulations and laboratory measurements for prototype radiating antenna will be compared with the TLM results in the conference presentation.

II. Analysis of Patch Antenna and Cascaded Superstrate Structure

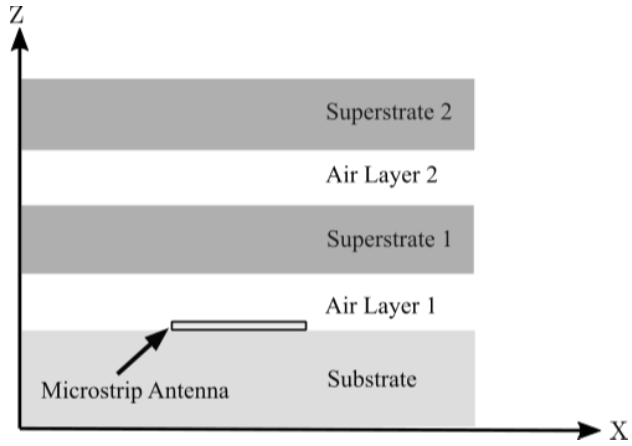

Fig. 1. A generalized configuration for the ABCD analysis of an MSA with superstrate and air layers

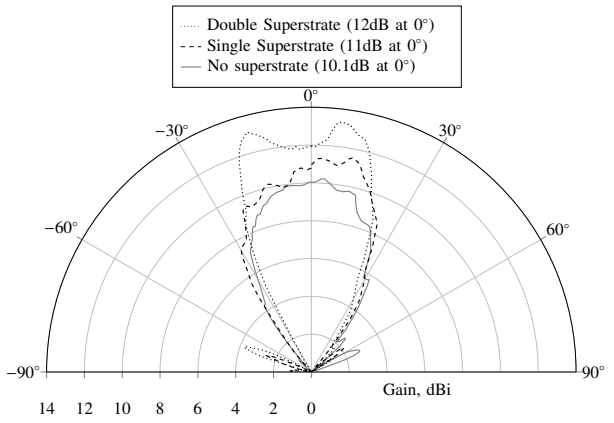

Fig. 2. E-plane radiation patterns for a patch antenna without and with single and double polycarbonate superstrates.

Intuitively, the modal current on the superstrate above the MSA should be a maximum for a maximum radiating field. This physically occurs when a single dielectric sheet is located approximately a quarter of guide wavelength $\frac{\lambda_g}{4}$ above the patch antenna resulting in a reinforcement of the radiated field. However the selection of the dielectric or ferrite materials and its thickness as well as the layer spacing must be fully taken into account. This situation becomes more complicated when multiple layers are considered.

A. Radiation form Microstrip Antenna

Figures 3 and 4 show a cavity model for the MSA and a coordinate system for computing the far field respectively. From the cavity model, the MSA at $z = h$ can be replaced by two magnetic current sources $\vec{M}_1$ and $\vec{M}_2$ along the y axis radiating an electric field of $\vec{E}_1$ and $\vec{E}_2$, respectively, at the observation point of $P(r, \theta, \phi)$.

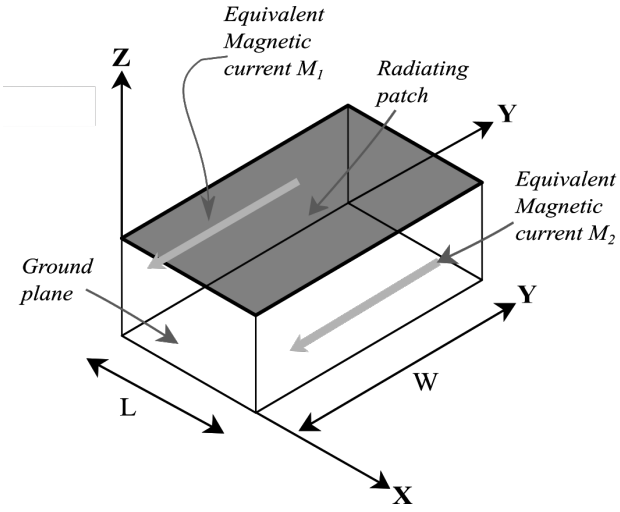

Fig. 3. Equivalent magnetic current sources, M_1 and M_2, representing the patch antenna.

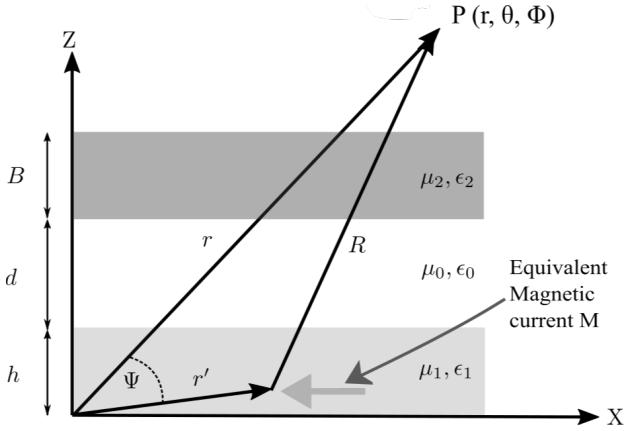

Fig. 4. Coordinate system for computing the far-field from the reciprocity electric source J2 at the patch antenna's radiation slot located at x = 0.

The ground plane is accounted for by image theory which results in two $-\hat{a_y}$ directed magnetic dipole sources of strength $2\vec{E_0}$. Considering the two magnetic sources as radiating slots of width W and height h, they form a two element array separated by L, a $\frac{\lambda}{2}$ spacing. Auxiliary potential functions, $\vec{A_p}$ and $\vec{F_p}$ can be used to derive the electric and magnetic fields from the equivalent current density sources[3]. Since there is no $\vec{J}$ present $\vec{A_p} = 0$.

$$\vec{F_p} = \frac{\epsilon}{4\pi}\oint \vec{M}\frac{e^{-jkR}}{R}ds = \frac{\epsilon c}{4\pi r}L, \quad \text{where } L = \oint \vec{M}e^{jk(r'Cos\Psi)} \tag{1}$$

$r'Cos\Psi = r' \cdot \hat{a_r}$ in cartesian coordinates becomes $r'Cos\Psi = y'Sin\theta Sin\phi + z'Cos\theta$ in spherical coordinates. Now, the far field radiated electric field $\vec{E_\phi}$ is:

$$\vec{E_\phi} = \frac{jke^{-jkr}}{4\pi r}(L_\theta - \eta N_\phi), \quad \text{with no } \vec{J} \text{ there is no } N_\phi \text{ component and with substitution for } L_\theta \tag{2}$$

$$\vec{E_\phi} = \frac{jke^{-jkr}}{4\pi r}(-2\vec{E_0})Cos\theta Sin\phi\left[\int_0^L\int_0^h e^{jk(y'Sin\theta Sin\phi + z'Cos\theta)}dz'dy'\right] \tag{3}$$

Equation 3 reduces to:

$$\vec{E_\phi} = \frac{jke^{-jkr}}{2\pi r}LV_0Cos\theta Sin\phi\left[\frac{SinY}{Y}\right]\left[\frac{SinZ}{Z}\right] \text{ where } \quad Y = \frac{kL}{2}Sin\theta Sin\phi \text{ and } Z = \frac{kh}{2}Cos\theta \tag{4}$$

A similar derivation can be made for $\vec{E_\theta}$. The MSA can be analysed as a two element array where a general result for the Array Factor(AF) is:

$$AF_n = \frac{1}{N}\left[\frac{Sin\left(\frac{N\psi}{2}\right)}{Sin\left(\frac{\psi}{2}\right)}\right] \text{ and the particular case becomes } \quad AF = \frac{Sin\left(\pi Sin\theta Cos\phi\right)}{2Sin\left(\frac{\pi}{2}Sin\theta Cos\phi\right)} \tag{5}$$

Hence the total $\vec{E_\phi}$ and $\vec{E_\theta}$ fields due to both $\vec{M_1}$ and $\vec{M_2}$ for the dominant TM_{01} mode are:

$$\vec{E_\phi} = \frac{jke^{-jkr}}{4\pi r}LV_0\left[Cos\theta Sin\phi\frac{SinY}{Y}\frac{SinZ}{Z}\right]\frac{Sin(\pi Sin\theta Cos\phi)}{2Sin\left(\frac{\pi}{2}Sin\theta Cos\phi\right)} \tag{6}$$

$$\vec{E_\theta} = \frac{-jke^{-jkr}}{4\pi r}\left[LV_0Cos\phi\left[\frac{SinY}{Y}\right]\left[\frac{SinZ}{Z}\right]\right]\frac{Sin(\pi Sin\theta Cos\phi)}{2Sin\left(\frac{\pi}{2}Sin\theta Cos\phi\right)} \tag{7}$$

B. TLM Model of the Multilayer Microstrip Antenna Structure

As the E_θ and E_ϕ are plane waves, the effect on them from the antenna structure can be found by the transmission line analogy by assuming the air and superstrate layers are interconnected transmission line segments terminated by a short circuit, as depicted in Figure5. The refractive indices, $n_1(\theta)$ and $n_2(\theta)$, and the corresponding propagation constants, β_0, β_1, and β_2, related to the multi-section transmission line equivalent model are given by:

$$n_1(\theta) = \sqrt{n_1^2 - \sin^2\theta}, \quad n_2(\theta) = \sqrt{n_2^2 - \sin^2\theta}, \quad \eta_0 = \sqrt{\frac{\mu_0}{\epsilon_0}} \tag{8}$$

$$\beta_0 = K_0\cos(\theta), \quad \beta_1 = K_0n_1(\theta), \quad \beta_2 = K_0n_2(\theta) \tag{9}$$

$$\tag{10}$$

The characteristic impedances for the TE wave or perpendicular polarization $Z_{c0_\perp}$, $Z_{c1_\perp}$, $Z_{c1_\perp}$ and the characteristic impedances for the TM wave or parallel polarization $Z_{c0_\parallel}$, $Z_{c1_\parallel}$, $Z_{c2_\parallel}$ are given by,

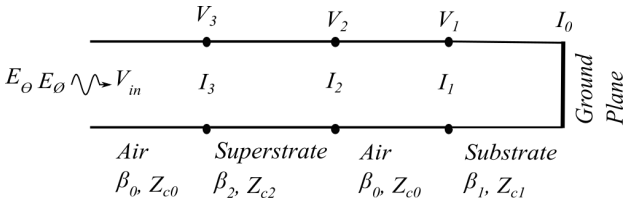

Fig. 5. Transmission line equivalent model of microstrip patch antenna with a superstrate cover.

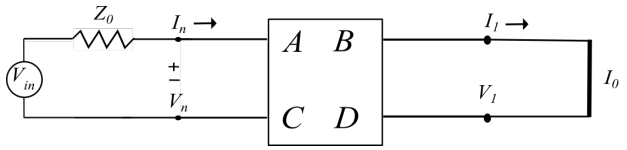

Fig. 6. Two port ABCD model.

$$Z_{c0_\perp} = \eta_0 \sec(\theta), \quad Z_{c1_\perp} = \frac{\eta_0 \mu_1}{n_1(\theta)}, \quad Z_{c2_\perp} = \frac{\eta_0 \mu_2}{n_2(\theta)} \tag{11}$$

$$Z_{c0_\parallel} = \eta_0 \cos(\theta), \quad Z_{c1_\parallel} = \frac{\eta_0 n_1(\theta)}{\epsilon_1}, \quad Z_{c2_\parallel} = \frac{\eta_0 n_2(\theta)}{\epsilon_2} \tag{12}$$

A lossless transmission line is assumed and since the transmission line model of the MSA and superstrate structure is a series of four terminal networks connected in cascade, ABCD analysis can be used to characterize the system [4]. The ABCD matrix for the structure is the product of the individual ABCD matrices for each four terminal network. Equation 13 shows the general form ABCD matrix for a transmission line, where l is transmission line length between terminals and Z_0 is the transmission line characteristic impedance:

$$\begin{pmatrix} cos(\beta l) & jZ_0 sin(\beta l) \\ j\frac{1}{Z_0} sin(\beta l) & cos(\beta l) \end{pmatrix} \tag{13}$$

In the two port network shown in Figure 6 the following expressions can be stated,

$$\begin{bmatrix} 0 \\ I_0 \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} V_n \\ I_n \end{bmatrix} \text{ , which gives: } 0 = AVn + BI_n \text{ and } I_0 = CV_n + DI_n \tag{14}$$

from which expressions for V_n, I_n and I_0 can be derived:

$$V_n = \frac{-BV_{in}}{(Z_0 A - B)} \tag{15}$$

$$I_n = \frac{AV_{in}}{(Z_0 A - B)} \tag{16}$$

$$I_0 = \frac{V_{in}}{(Z_0 A - B)} \quad \text{assuming a reciprocal network where } AD - BC = 1 \tag{17}$$

I_0 is the current flowing on the ground plane due to the far field radiation with V_{in} being either polarization E_θ or E_ϕ.

III. Results from ABCD Cascade Analysis

A parametric analysis of several multilayer configurations of the 10GHz MSA conforming to the generalized configuration shown in Figure 1 was performed where the effect of several factors on the antenna was examined. Insights from this examination were then used to optimize designs for increased boresight gain in this antenna structure.

A. Effect of Microstrip Antenna Width on Boresight Gain

The gain of a 10GHz MSA increases approximately linearly as the antenna width is increased as shown in Figure 12. This is explained by considering the cavity model where the distance between the two radiating elements of the MSA increases and so the AF increases and hence the gain will also increase.

B. Single Superstrate Configuration

The effect of changing the superstrate permittivity and thickness was investigated for a single superstrate suspended 1mm above a 10GHz MSA and the results in Figures 7 and 8 show very small increases in gain with thinner superstrates at higher permittivity.

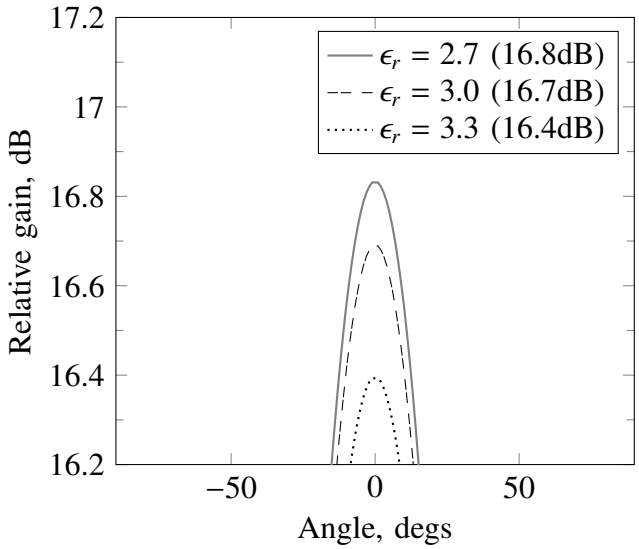

Fig. 7. ABCD calculated E-plane Gain of a 10GHz MSA with a 4.3mm thick single superstrate layer, with different permittivity, over a 1mm air layer

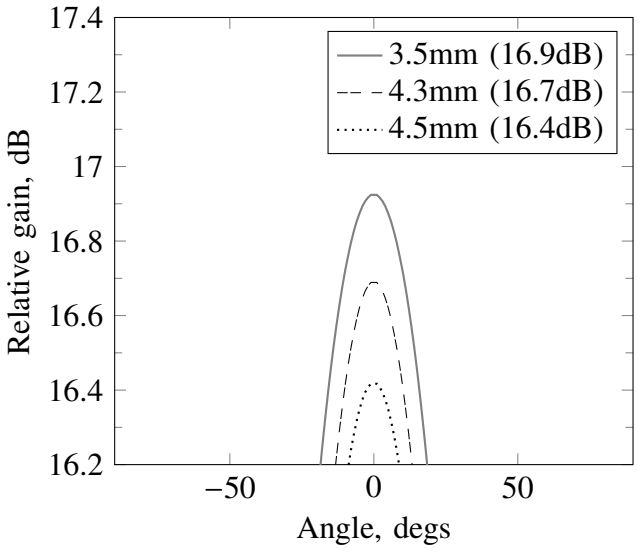

Fig. 8. ABCD calculated E-plane gain of a 10GHz MSA with a single ϵ_r = 3.0 superstrate layer with varying superstrate thicknesses over a 1mm air layer.

Two types of Taconic [5] substrates, RF-30 with ϵ_r = 3 and RF-10 with ϵ_r = 10, were considered for analysis and using their standard production thicknesses of 0.25mm, 0.5mm, 0.76mm and 1.52mm for RF-30 and 0.25mm, 0.5mm, 0.64mm, 1.52mm and 3.18mm for RF-10 to further investigate the effect of permittivity and superstrate thickness on the 10GHz MSA. Figures 9 and 10 show that for each thickness of superstrate a resonance gain condition is achieved with a greater air gap as the superstrate becomes thinner for both RF-30 and RF-10. The gains for RF-30 and RF-10 are comparable for the same superstrate thickness where for example 0.5mm thick superstrates produce boresight gains of 22.7dB and 22dB for RF-30 and RF-10 respectively however the air gap needed for resonance is 64mm for RF-30 but only 44mm for RF-10 which would mean a more compact antenna structure for the same performance. An important point is that superstrates much less than a quarter wavelength give much better performance for gain enhancement than for the quarter wavelength case as shown in Figure 9 where the 4.3mm thick superstrate is the worst performing for RF-30.

Figure 11 summarizes the the effect of superstrate thickness on boresight gain and shows that the resonance gain for both RF-30 and RF-10 increases approximately linearly with a decrease in superstrate thickness and that for a single superstrate a lower permittivity gives higher boresight gain.

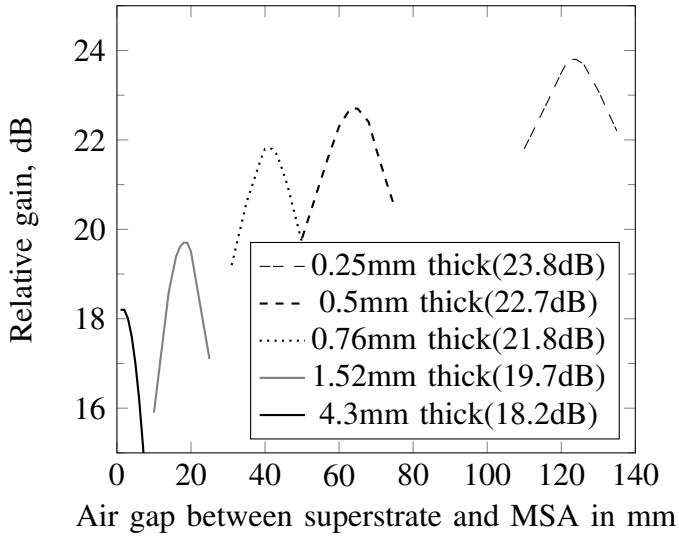

Fig. 9. ABCD cascade analysis of E-plane boresight gain for a 10GHz MSA with an ϵ_r = 3 superstrate of four standard production thicknesses with different air gaps between the superstrate and MSA.

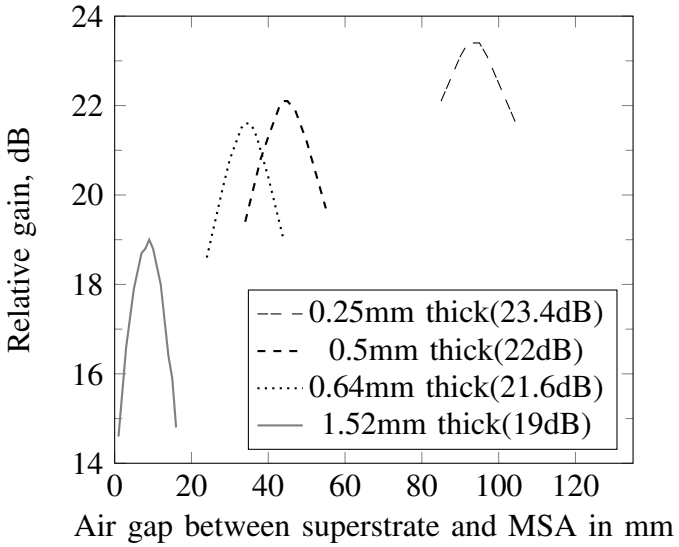

Fig. 10. ABCD cascade analysis of E-plane boresight gain for a 10GHz MSA with an ϵ_r = 10 superstrate of five standard production thicknesses with different air gaps between the superstrate and MSA.

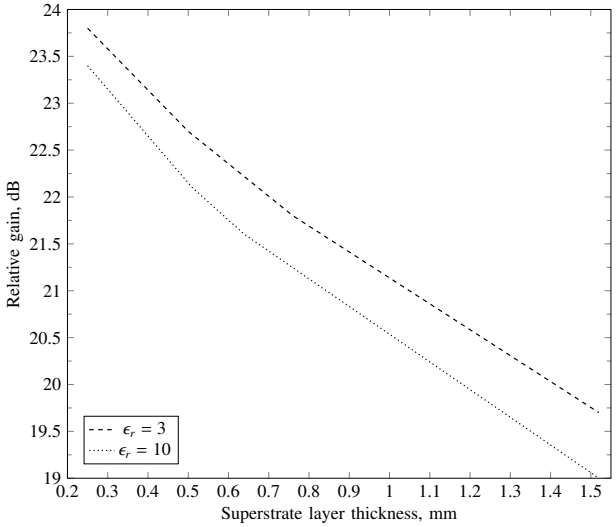

Fig. 11. ABCD cascade analysis of boresight resonance gain for varying superstrate thickness

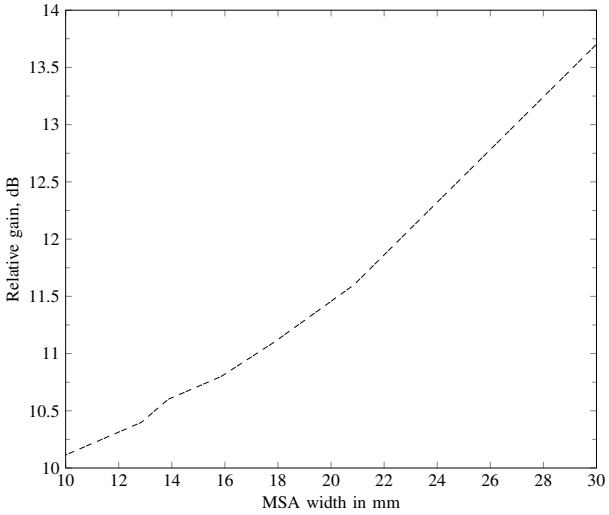

Fig. 12. ABCD cascade analysis of E-plane boresight gain of a 10GHz MSA with changing MSA width

C. Double Superstrate Configurations

Figure 13 shows the E-plane boresight gain of a 10GHz MSA when covered by a two-layer superstrate with no air layers comprising ϵ_r = 10 and ϵ_r = 3 for superstrate 1 and 2 respectively with each having several thicknesses of 0.25mm, 0.51mm, 0.64mm, 1.52mm, 3.18mm for superstrate 1 and 0.25mm, 0.5mm, 0.76mm and 1.52mm for superstrate 2. For 0.25mm, 0.51mm and 0.64mm thick superstrate 1 the gain increases by approximately 3dB per extra 1mm thickness of superstrate 2. As ϵ_r = 10 becomes thicker the increase in gain is affected and for 1.52mm thick the gain decreases when superstrate 2, ϵ_r = 3 is above 0.8mm thick. When ϵ_r = 10 is 3.18mm the gain decreases for all thicknesses of ϵ_r = 3. Figure 14 shows the opposite case where ϵ_r = 3 is superstrate 1 and ϵ_r = 10 is superstrate 2 and for all thicknesses

of superstrate layer 1 the gain has a peak particular to the thickness of superstrate layer 2 and the rise and fall in gain is 4.9dB per 1mm change in superstrate layer 2 thickness.

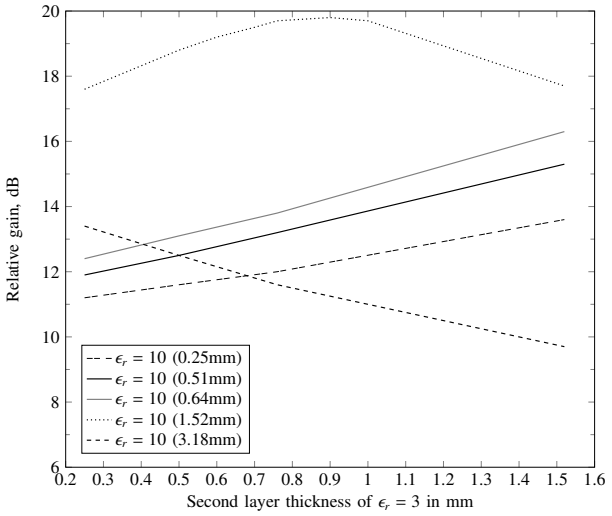

Fig. 13. ABCD cascade analysis of E-plane boresight gain for a 10GHz MSA with combinations of ϵ_r = 10 superstrate of five standard production thicknesses of 0.25mm, 0.51mm, 0.64mm, 1.52mm and 3.18mm as a first layer with a second layer of ϵ_r = 3 of four standard production thicknesses of 0.25mm, 0.5mm, 0.76mm and 1.52mm covering the MSA.

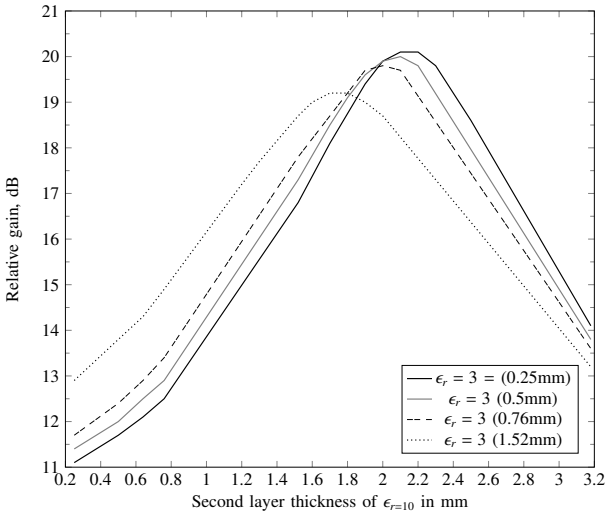

Fig. 14. ABCD cascade analysis of E-plane boresight gain for a 10GHz MSA with combinations of ϵ_r = 3 superstrate of four standard production thicknesses of 0.25mm, 0.5mm, 0.76mm and 1.52mm as a first layer with a second layer of ϵ_r = 10 of four standard production thicknesses of 0.25mm, 0.51mm, 0.64mm, 1.52mm and 3.18mm covering the MSA.

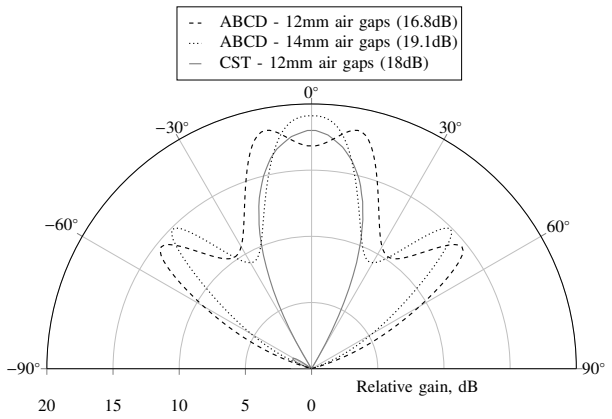

Fig. 15. Maximum E-plane gain of a 10GHz MSA with two 4.3mm thick, $\frac{\lambda_r}{4}$, superstrates of ϵ_r = 3 where CST had 12mm air gaps and ABCD method had 14mm air gaps

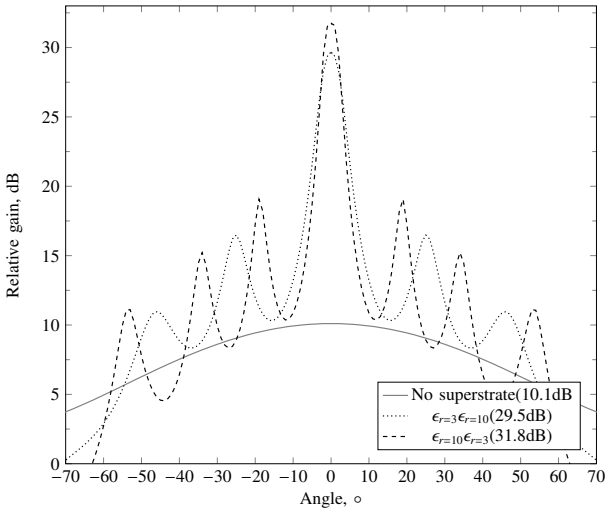

Fig. 16. 10GHz MSA resonance gain double superstrate configuration of a 0.5mm $\epsilon_{r=10}$ and $\epsilon_{r=3}$ superstrate in alternate superstrate 1 and 2 positions

Figure 15 shows the gain for a double superstrate configuration 10GHz MSA, illustrated in Figure 1, where each superstrate is ϵ_r = 3 and electrically a quarter wavelength thick, $\frac{1}{4}\lambda_R$, at 4.3mm and this was the configuration used in the authors previous work [1]. This shows the sensitivity of the boresight maximum gain to a slight change in air layers where it is reduced from 19.1dB to 16.8dB with an air gaps just 2mm smaller for air layer 1 and 2. A maximum boresight gain of 18dB, calculated by CST, was realized with air layers 1 and 2 of 12mm and the similarity in gain between the full-wave and analytical derivations

provide a verification of the ABCD cascade analysis. The difference is due to the ABCD cascade analysis using an expression for just the TM_{01} mode of the electric field for the MSA. This result indicates that the air gaps need to be sufficiently large to allow a plane wave to be incident on the superstrate and which will give the greatest diffraction according to Snell's law resulting in the resonance gain. The ABCD cascade analysis allows iterative tuning of the air gaps to find the resonance gain.

The resonance gain of a double superstrate configurations was investigated comprising one RF-30 and one RF-10 superstrate and comparing the effect of having them alternately in superstrate 1 and 2 positions. Both superstrates were 0.5mm thick since this gives good gain with the mid-size air gaps for the single superstrate case as shown in Figures 10 and 9. The results in Figure 16 show that having RF-10 as a first superstrate layer with RF-30 above it, with air layer 1 and 2 of 0mm and 58mm respectively, gives the best boresight gain of 31.8dB and a total antenna height of 59mm and the configuration with the superstrate positions swapped, with air layer 1 and 2 of 0mm and 48mm respectively, giving 29.5dB. Both plots show that the inclusion of air layers means the antenna pattern shows increased sidelobes

D. Double Superstrate Configurations Incorporating Magnetic Superstrates

Figure 17 shows the effect on the boresight gain when various thicknesses of a superstrate of RF-10 are covered by a 4.3mm thick superstrate of $\mu_r = 4$ with no air layers.

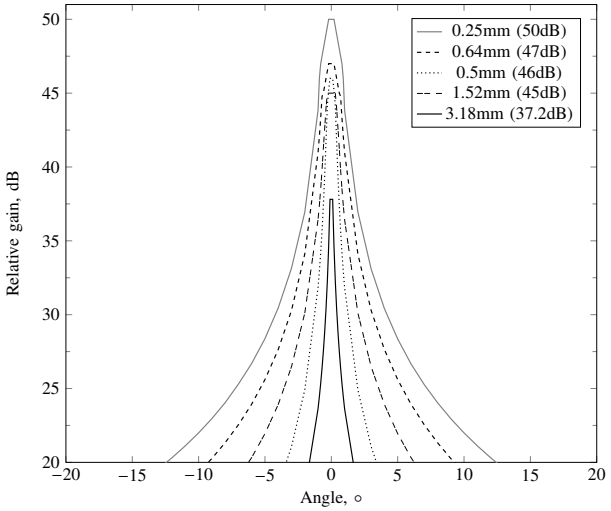

Fig. 17. 10GHz MSA resonance gain with a RF-10 superstrate of different thickness of 0.25mm, 0.51mm, 0.64mm, 1.52mm and 3.18mm as superstrate 1 and $\mu_r = 4$ of 4.3mm as superstrate layer 2

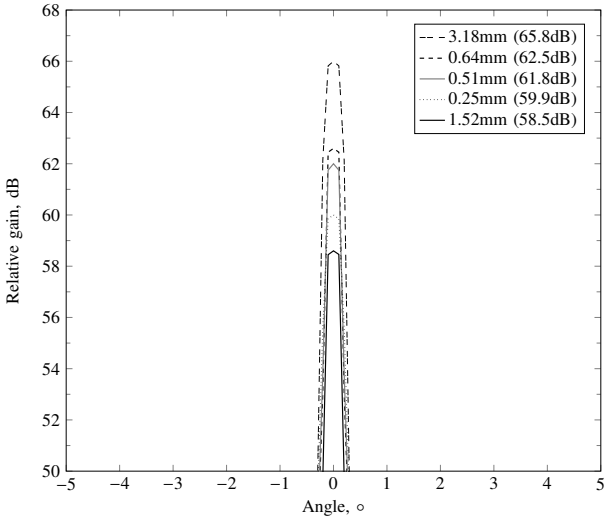

Fig. 18. 10GHz MSA resonance gain with RF-10 as superstrate 1 of different thickness of 0.25mm, 0.51mm, 0.64mm, 1.52mm and 3.18mm with 4.3mm thick $\mu_r = 4$ as superstrate 2 with air layers 1 and 2 of 0mm and 14mm respectively except for the configuration with 3.18mm thick superstrate 1 where air layer 1 was 29mm.

The highest gain of 50dB is achieved with a 0.25mm thick superstrate of RF-10 that gives a total structure height of 4.25mm. The gain reduces as RF-10 becomes thinner. The increased boresight gain from the inclusion of a magnetic superstrate as the top layer is due to the added gain it provides in the H plane. Adding air layer 1 and 2 of 0mm and 14mm respectively to the configuration further increases the gain as shown in Figure 18 where the highest gain of 65.8dB is achieved with superstrate layer 1 of RF-10 at 3.18mm thick. This makes the total height of the antenna structure 21.18mm. This very high gain is considered a mathematical artifice and thought not practically realisable due to the finite dielectric loss present and mechanical tolerances. Suitable magnetic superstrate with low values of permeability could be

microwave hexaferrites although these tend to be bulky and also possess loss. Alternatively, custom made permeability values can realised using metamaterials which are lightweight and permit planar integration but tend to be very narrowband.

IV. Conclusions

An analytic model was developed that allowed rapid numerical calculation of the radiation pattern and gain of MSA configurations comprising single and double layered superstrate composed of arbitrary dielectric and ferrite media. Generally, increases in MSA gain were feasible with a suitable choice of single superstrate; dielectric value, thickness and $\lambda_g/4$ resonant air gap above the antenna. Similar gains are also feasible with substitution of a magnetic superstrate. Stacking of multiple layers while observing the appropriate material resonant condition further increases the MSA gain. Stacking more than two superstrate layers may result in additional gain but the antenna size increase is perhaps undesirable. Dielectric substrates with low permittivity values are readily available; however, magnetic substrates of suitable low values are both difficult and very expensive to obtain.

A wide ranging study of practical MSA configurations with single and cascaded dielectric layers were investigated numerically. The study considered use of only available manufacturers production microwave laminates of different dielectric values and thickness. Several low cost single and double layer compact MSA configurations were identified that provided significant (typically $\geq$ 15 dBi) gain enhancement. The utility of the study also demonstrates that for little addition materials an MSA antenna can have its radiation directivity significantly increased.

Acknowledgment

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Adamson, Michael F.; Morrow, Ivor L.

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