Impact of mutual coupling in leaky wave enhanced imaging arrays

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finite size ground plane effects on the radiation pattern consistency are also evaluated and a maximum pattern variation less than 3 dB from the ideal case are demonstrated.

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Impact of Mutual Coupling in Leaky Wave Enhanced Imaging Arrays

Nuria Llombart, Andrea Neto, Giampiero Gerini,
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Abstract—The impact of mutual coupling between neighboring radiators in an imaging array configuration in the presence of a dielectric super-layer is investigated. The super-layer generally aims at increasing the directivity of each element of the array. However, here it is shown that the directivity of the embedded element patterns are reduced by a high level of mutual coupling. Thus a trade off between directivity enhancement and close packing of the array elements must be found depending on the bandwidth and the pattern requirements.

Index Terms—Electromagnetic bandgap (EMB) materials, leaky waves (LWs), leaky wave (LW) antennas, reflector antenna feeds.

I. INTRODUCTION

The simplest multibeam focal plane imaging systems employ a single feed per beam configuration. In such systems the physical size of the feed elements should be small and the feeds should be

Fig. 1. Cross section of a single ground and superlayer stratification with the relevant geometrical parameters together with an active array element and immediate neighbors.

as close as possible to achieve small secondary beam separation and good crossover levels. Simultaneously, the patterns of each of the elements should be directive enough (or shaped in a certain way) to optimize the illumination of the reflector and to minimize spillover losses. An attractive approach to accommodate both these conflicting requirements could be to cover an array of compact radiators with a dielectric super-layer. The directivity enhancement of single planar radiators loaded by dielectric superlayers has already been demonstrated in recent articles [1]–[6]. Multibeam imaging arrays would benefit significantly if the demonstrated performance enhancement of a single feed could be achieved for several feeds in an array environment. However, this extension is not straightforward and no such system seems to have been realized to date.

In the present paper, dielectric super-layers supporting leaky waves are proposed in order to enhance the performance of imaging arrays that employ a single feed per beam. It is shown that the main limitation of using leaky waves in such arrays arises from mutual coupling effects. To quantify the mutual coupling, approximate analytical formulas based on the behavior of the leaky-wave poles of the pertinent Green's function are derived. Next the impact of this coupling on the radiation patterns is evaluated. As a study case a hexagonal array of square waveguides is considered initially. In the last section, a configuration is introduced that improves the quality of the radiation patterns having in mind beam shaping as opposed to achieving maximum gain. The results shown in this paper constitute the theoretical foundation and provide design guidelines for specific application driven designs.

II. FEED PATTERN ENHANCEMENT BASED ON LEAKY WAVES

A cross section of a single layer dielectric stratification is depicted in Fig. 1, together with the active element surrounded by the immediate neighbors. As explained in [1], the highest directivity at broadside is obtained when the slab thickness is $\lambda_d/4$ (with λ_d the wavelength in the dielectric at the frequency f_0) and $h_1 \approx \lambda_0/2$. These parameters lead to a resonance at f_0 . The zeroes of the denominators of the spectral Green's function applicable to this configuration represent the leaky-wave poles under investigation. These poles can be expressed as $k_{lw} = k_0(\sin \theta_{lw} + j\delta_{lw})$, which approximately define a pointing angle at which they are radiated and a radial attenuation constant along the ground plane. In [7] analytical approximations for the leaky-wave poles, valid over a broad frequency range, are provided.

The two graphs in Fig. 2 represent such poles for both TE and TM polarizations with respect to the z-coordinate. For several values of

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Fig. 2. Solutions of the dispersion equation for a single layer for several ϵ_r values, where $h_1 = \lambda_0/2$ and $h = \lambda_d/4$ at the central frequency. (a) The radiation angles θ_{LW} . (b) The attenuation constants δ_{LW} .

the dielectric constant ϵ_r , the figure shows θ_{LW} and δ_{LW} as a function of the normalized frequency f/f_0 . For lower dielectric constants, the leaky-wave beam is pointing towards larger angles. Furthermore, a lower value of the dielectric constant implies a larger amplitude for the attenuation constant, which results in a reduced amount of directivity enhancement with respect to the free-space case.

Most of the designs presented in [1]–[6] focus on operation at a specific resonance frequency, f_0 , and use high effective dielectric contrasts with the objective to obtain maximum directivity. Fig. 3 shows $|E|^2$ in the E- and H-planes radiated by an aperture of one wavelength at f_0 , operating in the presence of the dielectric slab. Clearly a higher dielectric constant leads to higher directivity.

III. LEAKY WAVE COUPLING

As demonstrated in [8], waveguide apertures operating in the presence of dielectric superstrates, which support leaky waves whose far-field radiation is pointing towards broadside, suffer from a reduced impedance bandwidth. This would lead to the selection of moderate values for the dielectric constant for the superstrate. It will be shown that for multibeam imaging systems based on a single feed per beam, a more appropriate selection of the dielectric constant is based on the level of the mutual coupling between the different elements.

Fig. 3. Directivity patterns $|E|^2$ in E and H-planes of the same structures analyzed in Fig. 2 excited by an aperture with dimension $w = \lambda_0$. Also the free space case is given for comparison.

The mutual coupling of the waveguide array in Fig. 1 can be expressed starting from the mutual admittances, which can be rigorously expressed in the spectral domain extending the procedure shown in [8]

$$Y = \frac{1}{(2\pi)^2} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} |M(k_x, k_y)|^2 G_{xx}^{hm}(k_x, k_y) e^{-jk_\rho d} dk_x dk_y \quad (1)$$

where $k_\rho = \sqrt{k_x^2 + k_y^2}$, $G_{xx}^{hm}(k_x, k_y) = -((k_x^2 I_{TE}(k_\rho) + k_y^2 I_{TM}(k_\rho))/k_\rho^2)$ is the spectral Green's function (GF) of the dielectric stratification that provides the magnetic field at $z = 0$ generated by a magnetic current at $z' = 0$, $M(k_x, k_y)$ is the Fourier transform of the equivalent magnetic current $m(x, y)$, and $d = \sqrt{d_x^2 + d_y^2}$ with d_x and d_y the distances between the waveguides in x and y , respectively. The integral can be evaluated numerically or asymptotically as shown in the appendix of [9], retaining only the leaky-wave contribution. The asymptotic evaluation will be provided in the following paragraph.

The mutual admittance, and corresponding mutual coupling, has two dominant contributions. One corresponds to the space wave and the other corresponds to the leaky modes propagating in the structure. When the waveguides have widths in the order of the free-space wavelength, the amplitude of the space-wave launched in directions tangent to the ground plane is very small, which results in a negligible space-wave coupling, lower than -30 dB. In this situation the mutual admittance is dominated by the leaky-wave contribution and one can refer to it as to the leaky-wave admittance: $Y \approx Y_{LW}$.

Focusing our attention to configurations of two waveguides that are either positioned in the E- ($\phi = \pi/2$) or in the H- ($\phi = 0$) plane, the leaky-wave admittances can be approximated as

$$Y_{LW}^{H/E} = j \frac{\sqrt{k_{LW(TE/TM)}} e^{j\pi/4}}{2\sqrt{2}\pi} |M(k_{LW(TE/TM)}, \phi)|^2 \times \text{Res}(I_{TE/TM}) \frac{e^{-jk_{LW(TE/TM)}d}}{\sqrt{d}} \quad (2)$$

where $d = d_x$ or $d = d_y$, $k_{LW(TE/TM)}$ is the first TE or TM leaky wave pole and Res indicates the residue of the electric-current solution of the pertinent transmission line as was indicated in [8]. $\text{Res}(I_{TE/TM}) \propto$

Fig. 4. Full-wave and leaky-wave admittances between two apertures $w = \lambda_0$ at a distance of $d = 1.2\lambda_0$, in both E and H planes, for the same structures as analyzed in Fig. 2. (a) Amplitude, (b) phase in the E plane, (c) phase in the H plane.

$1/k_{lw(TE/TM)}$, and therefore $Y_{lw(TE/TM)}^{H/E} \propto 1/\sqrt{k_{lw(TE/TM)}}$. Equation (2) can easily be generalized to waveguides oriented at arbitrary angles, using a combination of the TE and TM admittances.

Fig. 5. Mutual coupling between the TE_{10} modes of the same aperture of Fig. 4.

Even without considering the exponential attenuation, one observes clearly that for smaller values of $k_{lw(TE/TM)}$, the leaky wave contribution to the mutual admittance is larger. This observation is confirmed by the data of Fig. 4(a), which shows the amplitude of the mutual admittance between waveguides with sides of λ_0 and located at a distance $d = 1.2\lambda_0$, in both E and H planes, for the same dielectric super-layer configurations of Fig. 2. The admittance Y_{lw} is calculated using the approximate expression (2), while Y is calculated using the full-wave technique described in [10]. Fig. 4(b) and (c) show similar agreement between the phases of these two admittances. Finally the mutual coupling, calculated both using the approximate admittances and the full-wave tool, is shown in Fig. 5. The size of the dielectric panel and of the ground planes was assumed to be infinite in these preliminary results.

IV. IMPACT OF MUTUAL COUPLING

Considering the mutual coupling levels in Fig. 5, one can observe that the coupling is higher for higher dielectric constants. Higher dielectric constants give rise to poles associated with far fields that radiate towards broadside. In other words the directivity is intimately linked with the mutual coupling. It can be noted that this conclusion holds also for other types of superlayers, periodic dielectrics or metallic EBGs. The peak level of mutual coupling is about $|S_{12}| = -16$ dB and is reached for $\epsilon_r = 9$. The power coupled into the neighboring waveguides ($P_{12} = P_1 |S_{12}|^2$) has two main effects in an imaging array scenario:

- 1) The power P_{12} coupled to the neighboring waveguide can be considered dissipated as discussed in [11] and as illustrated in Fig. 6. The ultimate theoretical performance can be strongly degraded since, for example, $|S_{ij}| \approx -15, -18$ dB and -24 dB imply -0.91 dB, -0.43 dB and -0.08 dB of power lost in the matched loads of the 6 neighboring waveguides.
- 2) The embedded element pattern will be significantly different from the isolated element pattern even if all passive apertures are closed in perfect matched loads, $S_{load} = 0$. There will always be an additional directly scattered contribution to the pattern. The power associated to the scattered field is indicated as P_s .

While the power dissipated in the neighboring waveguides is a well-known drawback also for overlapping feed clusters and direct radiating arrays, the impact of mutual coupling on the radiation patterns of imaging arrays is far less understood.

The impact of the mutual coupling on the radiation patterns can be seen in Fig. 7. The graph presents the fields radiated in absence of the

Fig. 6. Power loss due to coupling with neighboring waveguide and power scattered from the non-perfect matched load.

dielectric layer (dotted lines) and those radiated in the presence of a dielectric layer characterized by $\epsilon_r = 4$ in isolation and in array configuration. Note that for the free space case the isolated and embedded patterns are essentially the same and thus only one representative curve has been presented. The patterns have been calculated using the commercial tool CST from MWS [12] and are shown in the main planes for the embedded and the isolated cases. In these simulations the size of the dielectric panel and of the ground plane was taken large enough to have small edge effects ($12\lambda_0 \times 12\lambda_0$). The maximum directivity in the array environments is lower than the one obtained in isolation and is essentially the same as the directivity obtained in case the dielectric slab is not there. That is because the distance between the array elements is such that the scattered field from the neighboring waveguides contributes almost out of phase with respect to the central element. A lower directivity will always occur when one tries to pack the elements as close as possible. This will now be explained in more detail. The leaky waves that contribute most to the mutual coupling do not result in a significant phase difference over a small distance. That is because the phase of the mutual coupling in (2) along the array is dominated by $e^{-jk_{lw}d}$, where d is defined in Fig. 1. For small separations d or for small values of θ_{lw} , $Re[k_{lw}]d \approx 0$.

If a waveguide is excited with a mode propagating from the bottom in the positive z -direction we can assume a phase equal to zero for the magnetic currents on the aperture. The forcing fields of the neighbors of the central waveguide are caused by the leaky waves that feed the aperture from the top (negative z -direction). As a result the equivalent magnetic currents on the neighboring apertures are all out of phase (180°) with respect to the magnetic current on the central aperture.

A schematic representation of the superposition between the field radiated by the central waveguide E_1 and the field scattered by the six surrounding waveguides E_{surr} is shown in Fig. 8. The thick curve represents the field E_1 , the thin curve with the dashed area represents E_{surr} , and the thin curve with the gray area represents the superposition of the two fields, E_{tot} . A realistic $S_{1j} \approx -20$ dB implies for all waveguides surrounding the central one that $m_j \approx -m_1/10$. Since the magnetic currents on the six surrounding apertures radiate in phase at broadside $E_{surr}(\theta = 0) \approx -0.6E_1(\theta = 0)$, which leads to $E_{tot} \approx 0.4E_1$. Given the periodicity, the scattered waves will contribute in phase with the radiation occurring from the central waveguide towards a grating-lobe angle $\theta_{gl} \approx 24^\circ$. This explains, in part, the widened embedded beams. Note that in this example $\theta_{gl} \approx \theta_{lw}$. The broader embedded beams are also clearly visible in the curves of Fig. 7. This example shows that in a single feed per beam scenario the mutual coupling between waveguides should be low, not only to avoid significant dissipated losses, but also to avoid the destructive interference from the neighboring waveguides. For example, $S_{1j} \approx -20$ dB results in an approximate 3 dB of loss in directivity with respect to

Fig. 7. Directivity of the calculated radiation patterns in the E and H-planes. The embedded patterns (central element of the array) present more flat and wider main beams than the isolated patterns. However also the embedded patterns present a more rapid drop-off for wider angles with respect to the free space cases.

the patterns in isolation. In case of $\epsilon_r = 9$ a peak mutual coupling of $S_{12} \approx -16$ dB would have been observed, which corresponds to $m_j \approx -0.16m_1$. Consequently the field radiated at broadside by the six surrounding waveguides would have been $E_{surr} \approx E_1$, i.e., as high as the one radiated by the central waveguide. Thus the gain enhancement achieved by the element in isolation would have been completely cancelled out by the mutual coupling. The embedded patterns would actually turn out to present a null at broadside. These observations reveal that array designs with large values of ϵ_r for the superstrate and small spacings d between the elements, have much worse performance than their single-element counterparts. As a design guideline 18 dB of mutual coupling in a hexagonal lattice can be considered the limit beyond which the effect of directivity enhancement due to the dielectric super-layer is essentially cancelled.

The embedded patterns in Fig. 7 are essentially flat until 20 degrees and then drop off fairly rapidly. Thus even if the embedded patterns are less directive at broadside than the one in the isolated case, they have been shaped to present better beam efficiencies than they would in the case of absence of the dielectric layer (indicated as free space case). Moreover, the drop-off can be even improved by optimizing the element shape of the elementary radiator as will be shown in the next section.

V. PATTERN SYNTHESIS

The patterns shown in Fig. 7 are far from optimum when one wishes to use them to feed a reflector antenna. To start with they are not symmetric in the two main planes. Moreover the considered superlayer configuration also excites a second TM mode with $\theta_{lw2}^{TM} \approx 70^\circ$ [8], which is responsible for the widening of the beam in the E-plane. This second TM leaky wave results in large spill-over losses since its radiation will not be intercepted by the reflector. This contribution can be cancelled by adopting small waveguides loaded with a double iris configuration as in Fig. 9 rather than using large waveguides. Each of the two pairs of slots is associated with one polarization so that each waveguide can be operated in circular polarization by properly phase shifting the two polarizations. The slots of each pair are excited in phase and are separated by a distance S such that their contributions cancel out exactly at θ_{lw2}^{TM} , leading to an $S = 0.53\lambda_0$. The slots are shaped as arcs in order to achieve the desired cancellation over the maximum azimuthal angle [13]. The array in Fig. 9 is composed of 19 waveguides of square cross section with width $w = 0.67\lambda_0$ and separation $d = 1.2\lambda_0$. In this case,

Fig. 8. Pictorial representation of the superposition between the field radiated by the central waveguide and the surrounding ones. Field radiated by the central waveguide, E_1 (thick curve), field scattered by surrounding waveguides E_{sur} (thin curve with dashed area), and superposition of the two fields, E_{tot} (thin curve with gray area).

Fig. 9. Final design of the prototype waveguide array. The area of each unit cell is significantly larger than the dimension of each waveguide.

the waveguide apertures are significantly smaller than the unit-cell dimension. From an electromagnetic point of view this is preferable since smaller waveguides support only the two fundamental and orthogonal TE waveguide modes, which simplifies the design of all the waveguide transitions in the front end. Since this iris-loaded design is fairly similar to the one shown in the previous section the mutual coupling between slots corresponding to the same polarization is also similar. The level is less than 20 dB over the operational bandwidth of about 10%. The impact of the mutual coupling on the radiation patterns can be seen in Fig. 10. The calculated patterns in four planes are shown for the embedded and the isolated waveguide. The pattern shapes are now very similar in all planes. The cross polarized field levels from the present radiating structure in the presence of the dielectric stratifications are very low (below -25 dB with respect to the maximum co-polarized field over the entire main beam in all ϕ cuts). This is consistent with the theoretical findings of [14].

The embedded patterns correspond to a scenario, in which all the waveguides surrounding the central one are closed in matched loads. These patterns are well suited for feeding center-fed reflectors characterized by moderate focal-distance-to-diameter ratios (F/D). The results in terms of edge-of-coverage gain that can be achieved using such patterns feeding a multibeam reflector system have been recently anticipated in [15]. They are very promising because, as demonstrated in [8], for low F/D ratios, the drop-off rate is driving the design instead of the broadside maximum directivity.

Fig. 10. Amplitude of the calculated radiation patterns of the designed array. (a) E and H planes and (b) 30° and 60° planes.

VI. CONCLUSION

In this paper it is shown that the limiting factor to directivity enhancement by means of dielectric superlayers in a multibeam imaging system is the mutual coupling between neighboring elements. A mutual coupling of -20 dB could imply as much as 3 dB degradation of the broadside directivity with respect to the case in which each of the radiators operates in isolation. A mutual coupling of -16 dB could even imply a null at broadside. This means that for closely packed imaging array configurations the performance enhancements should be found in the beam shaping (which can lead to improved secondary beam efficiencies) as opposed to the improvement of the primary gain at bore-sight. Moreover it has been shown how to design a compact waveguide radiator such as to eliminate the broadening of the beam due to a second TM mode, that always exists in dielectric super-layer configurations.

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On the Possibility of Interpreting Field Variations and Polarization in Arched Tunnels Using a Model for Propagation in Rectangular or Circular Tunnels

J. M. Molina-García-Pardo, M. Lienard, A. Nasr, and P. Degauque

Abstract—We investigate the possibility of using the modal theory of the electromagnetic propagation in rectangular or circular tunnels, to satisfactorily interpret experimental results, including polarization, in arched tunnels. This study is based on extensive measurement campaigns carried out in the 450 MHz–5 GHz frequency range.

Index Terms—Modal theory, polarization, propagation, tunnel.

I. INTRODUCTION

Wireless communications in confined environments, such as tunnels, have been widely studied for years, and a lot of experimental results have been presented in the literature, mainly to describe mean path loss versus frequency in environmental categories ranging from mine galleries and underground old quarries [1] to road and railway tunnels [2], [3]. In the last two cases, arc-shaped tunnels are quite usual. They have nearly the shape of a cylinder whose lower part is flat, supporting either rail tracks or a road. Predicting and interpreting the field distribution inside such tunnels, when the field is excited by an electric antenna, are important in the deployment of wireless communication systems.

This field distribution must take into consideration not only the mean path loss, which is often studied in the literature, but also the location and periodicity of the fading phenomenon and the co-polar/cross-polar ratio as a function of frequency, of the position of the antennas in the tunnel cross section and of the distance between the transmitter and the receiver. Indeed, polarization is an important parameter for optimizing the performance of a communication system if it is based on diversity and multiple input multiple output (MIMO) techniques. In this work, only a narrow band approach is considered. An example of practical application is the GSM-R, devoted to railway communication in Europe, and operating in the 900 MHz frequency band with an allocated bandwidth of few hundred kHz. For wide band transmission, additional characteristics as the delay spread and the coherence bandwidth, would be needed to characterize the channel [4].

From a mathematical point of view, the internal surface of an arched tunnel cannot be easily described using a canonical coordinate system and, consequently, no exact analytical formulation is currently available. However, an approximate approach based on an equivalent analysis method has been recently proposed [5] to predict the characteristic

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