Leaky Wave Enhanced Feeds for Multi-Beam Reflectors to be used for Telecom Satellite Based Links

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Abstract—The use of dielectric super-layers for shaping the radiation pattern of focal plane feeds of a multi-beam reflector system is discussed. Using the super-layers, it is possible to reduce the spillover from the reflectors without increasing the dimension of each aperture. The effect has been demonstrated using a realistic array configuration. The experimental demonstration is obtained with configurations which are typical for satellite based multi beam telecommunication links. Thanks to the adoption of a ×1 reuse scheme, based on frequency and polarization orthogonal channels, the measured edge of coverage directivity has improved, with respect to a standard technology case, by 2 dB over an operational bandwidth of about 2%.

Index Terms—Leaky-wave antennas, leaky waves, reflector antenna feeds, reflector antennas.

I. INTRODUCTION

Present and next generation telecommunication satellite systems often require a multiple beam capability. A simple way to achieve high edge of coverage gain is to use a single aperture and a single feed per beam. However, as clarified in the introduction of [1], this method may lead to several dB of spillover losses. In the series of works [2], [3], [4] an approach based on the use of dielectric super-layers to enhance the radiation properties of small apertures in order to excite reflector antenna systems has been proposed. The same approach has also been applied to completely planar technology in [5]. The basis for the results presented in [2]-[5], can be traced back to the original paper of von Trentini [6]. In recent years several other authors have also followed a similar route [7]-[12]. This way of proceeding [2] makes use of the shaping of the radiation patterns via the excitation of a pair (TE/TM) of leaky waves, that radiate incrementally as they propagate between the ground plane and a super-layer.

In particular [4] demonstrated that an increase of edge of coverage gain of 1.3 dB over a bandwidth (BW) of about 6%, can be obtained for a dual reflector system characterized by moderate values of the F/D (Focal distance to Diameter ratios in the order of 0.85). The achievement was demonstrated only theoretically since it was assuming the presence of a filtering feeding network to facilitate frequency and polarization orthogonality of neighboring beams. In this contribution the same strategy based on the use of super-layers is first adapted to the design of reflector systems that use larger F/D ratio’s which are typical for satellite based communications links. Secondly, the new dedicated design is experimentally validated by hardware demonstrations of the dielectric super-layers, the feeds and the filtering structures.

A. Requirements

The starting point of this paper is a telecommunication satellite-based multi beam (≈ 10 × 10) reflector system for spot coverage of the earth. The beams will be arranged in an hexagonal lattice and separated by an angle Δθ in the order of ≈ 0.56°. To achieve such beam separation and negligible performance degradation for the outer beams, the F/D of the system should be in the order of 1.5 to 2. The separation between the feeds will be about 2.4λ0 (where λ0 is the wavelength at the central operating frequency f0) which is typical for existing telecom multi-beam systems. The selected focal distance is approximately F = 245.5λ0. The target relative frequency bandwidth will be in the order of 2%, and inside this band a minimal edge of coverage directivity of 44.5 dB should be guaranteed. Note that the edge of coverage gain/directivity is defined as the gain/directivity at the cross over between three adjacent beams: G_{eoc} = G(Δθ/√3).

Fig. 1. Hexagonal grid array with periodicity d of circular metallic waveguide Da mounted on a infinitely extended ground plane.
In order to achieve the orthogonality between neighboring beams a × 4 re-use scheme will be used that is very similar to the one that was described in [4]. The scheme is based on a hybrid frequency and polarization reuse. The total frequency band is divided in two sub-bands of widths $\Delta B = 0.9\%$ and centered at $f_{c1} = 0.992 f_0$ and $f_{c2} = 1.005 f_0$, with a roll off band $\Delta R = 0.45\%$ that separates the two useful bands. To separate the two bands, multiple cavity band pass filters are connected to the waveguide feeds. The feeds will be polarized either vertically or horizontally, by properly shaping the irises. After the implementation of the re-use scheme the isolation between neighboring beams will be at least 12 dB (roughly corresponding to −20 dB side lobe levels). The dimension of the feed array will be minimized in order to minimize the weight of the entire front end.

### B. Benchmark

An array based on standard waveguide horn technology cannot reach the target 44.5 dB edge of coverage directivity. However, in order to establish a benchmark to which to compare the performances of the leaky wave enhanced arrays, such a standard waveguide array is investigated first (Fig. 1). It consists of an hexagonal grid array composed of 19 circular waveguide horns mounted on a finite ground plane of dimensions $(20 \lambda_0 \times 20 \lambda_0)$. The reference system has the z axis normal to the ground plane. The periodicity $d$ of the array is set to $2.4 \lambda_0$. The circular apertures are obtained by flaring to a diameter of $D_a = 2.2 \lambda_0$ the feeding waveguides of diameter $D_w = 0.7 \lambda_0$, over a taper of length $t$. With this arrangement the apertures cover almost the entire central part of the ground plane. The length of the taper greatly influences the efficiency of the feeds. If the length can be made arbitrarily long (infinite taper) the performance of the feeds are optimal with each one of them presenting an aperture efficiency in the order of 80%. In that case the apertures are illuminated with essentially uniform phase [13]. Almost independently from the taper length the reflection coefficients, $S_{ii}$ and the mutual coupling coefficients, $S_{ij}$ of each of the waveguides can be designed to be lower than $−20$ dB and $−30$ dB over the entire BW, respectively. Such an array is appropriate to feed reflector systems characterized by large $F/D$ ratios.

To accurately evaluate the performance of a reflector system fed by an array of waveguides, the far fields after the reflector need to be calculated. Throughout this paper a dual-offset Gregorian antenna, satisfying the Mizugutchi [14] condition is assumed (to avoid the blockage of the feed array). As clarified in [4], in first approximation the design of the system is still based on equivalent diameter $D$ and focal distance $F$, especially when the $F/D$ is relatively large. The geometry of the system is shown in Fig. 2. Which also highlights the blockage that the over-dimensioned focal plane array manufactured in this activity would imply.

The secondary fields are obtained using a simplified Physical Optics code based on the equivalent parabola. The algorithm uses as inputs the far fields radiated by the reference array, which in turn have been calculated using Microwave Studio CST [15]. Finite array effects are negligible. Figure 3 shows among other curves the edge of coverage directivity as a function of the frequency associated to each element of the iris loaded waveguide array of Fig. 4. The selected antenna geometry has the diameter $D = 143.6 \lambda_0$ and focal distance $F = 245.5 \lambda_0$ (corresponding to $F/D = 1.71$). The three curves refer to the directivity simulated for the free space reference case with in phase field distribution, to the directivity simulated in double dielectric configuration and to the one from the prototype for the double dielectric configuration.
The two slabs of dielectric constant $\epsilon$.

The target configuration is described schematically in Fig. 6.

This is because the prototype demonstrator was manufactured to a single dielectric layer of $\epsilon$.

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In order to visualize the impact of the dielectric super-layers in Fig. 4a is electrically equivalent to the one in Fig. 4b. The closest effective (same frequency and polarization) waveguides are at a distance of $d_{eq} = 4.8\lambda_0$ which corresponds to mutual coupling in the order of $-30$ dB. Given that each iris-loaded waveguide is characterized by width $w = 0.667\lambda_0$, one can observe that eventually the area of each equivalent unit cell is about 50 times larger than the dimension of each waveguide.

The primary patterns from an array configuration as shown in Fig. 4b turn out to be clean and symmetric in all $\phi$ cuts. In order to visualize the impact of the dielectric super-layers in the present context, the predicted directivity patterns for a dielectric slab characterized by $\epsilon_r = 20.25$ in the four significant planes are reported in Fig. 5 at the frequency of 9.95 GHz. These patterns are also compared with the ones that would be achieved by the reference (in phase) circular apertures, and by the iris loaded waveguides without the super-layer enhancement.

It is apparent that the super-layer enhanced patterns are much more directive than the ones without super-layers and thus, over a small frequency range more suited to feed efficiently a reflector with large $F/D$.

In the rest of this paper a configuration with double dielectric slabs of lower dielectric constant electrically equivalent to a single dielectric layer of $\epsilon_r = 20.25$ will be considered. This is because the prototype demonstrator was manufactured using more readily available lower dielectric constant material. The target configuration is described schematically in Fig. 6. The two slabs of dielectric constant $\epsilon_r = 4.5$, and thickness $h_d = \lambda_d/4$, with $\lambda_d = \lambda_0/\sqrt{\epsilon_r}$, are located at a distance $h_s = \lambda_0/4$ one from the other. The lowest slab is at a distance $h = \lambda_0/2$ from the ground plane. All the dimensions are

II. PATTERN SYNTHESIS BASED ON LEAKY WAVE ENHANCEMENT

In [3], [4] it is explained how to use a super-layer configuration to enhance the radiation from each aperture in an imaging array configuration. In those articles it was clarified that the maximum achievable directivity enhancement was obtained when the mutual coupling with the closest feed was in the order of $-30$ dB (or lower). In fact, due to the first couple of TE/TM leaky waves, in leaky wave enhanced configurations the neighboring elements contribute out of phase with the center one. When applying the design considerations described in [3], [4] to the problem at hand, it follows that an appropriate single dielectric super-layer would be characterized by $\epsilon_r = 20.25$. The corresponding array configuration is shown in Fig. 4a, and it is very similar to the one that was presented in [3]. In the absence of a re-use scheme the mutual coupling between neighboring elements would be in the order of $-18$ dB, which is too high. Accordingly a $\times 4$ reuse scheme needs to be used. When the frequency and polarization orthogonality is introduced the low frequency and vertically polarized array in Fig. 4a is electrically equivalent to the one in Fig. 4b. The closest effective (same frequency and polarization) waveguides are at a distance of $d_{eq} = 4.8\lambda_0$ which corresponds to mutual coupling in the order of $-30$ dB. Given that each iris-loaded waveguide is characterized by width $w = 0.667\lambda_0$, one can observe that eventually the area of each equivalent unit cell is about 50 times larger than the dimension of each waveguide.

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![Fig. 4](image1)

![Fig. 5](image2)
evaluated at the central operating frequency assumed to be 10 GHz. Fig. 3 shows the edge of coverage directivity as a function of the frequency for this preferred twin dielectric slab configuration. Also reported in the same figure are the expected edge of coverage directivities for the reference in phase field horn array that fits in the array cell. This configuration is considered an idealized configuration against which to compare.

With respect to the free space reference case the dielectrically enhanced structure provides an increase in the edge of coverage directivity of more than 1.5 dB over a 2.1% BW (from 9.86 to 10.1 GHz).

III. FURTHER BW CONSTRAINTS

Once the edge of coverage directivity against bandwidth trade off has been investigated the actual useful BW evaluation will eventually be further reduced by the considerations on the impedance matching of the structure and the shape of the primary and the secondary patterns.

A. Impedance Matching

The highly resonant super-layer structure has a strong impact on the input impedance of each feed. The BW is reduced and this effect is directly related to the gain enhancement at broad side. This effect has been studied in [12] first and then by the authors of this paper with the specific reflector application [2]. The simulated S parameters associated to a

Fig. 6. Dielectric stratification corresponding to the double layer design.

Fig. 7. Simulated S-parameters pertinent to a single iris loaded waveguide in isolation and to a three element array where the elements are at distance 4.8λo one from the other.

Fig. 8. The simulated primary beam directivities generated by each iris loaded feed in array environment are plotted as a function of the frequency calculated using CST. a) E-plane. b) H-plane. c) D-plane.
double dielectric structure are shown in Fig. 7. The shown parameters are pertinent to a single waveguide in isolation (loaded by a double iris) and to a three waveguide array where the elements are at distance $4.8\lambda_0$ one from the other. One can see that the $S_{11}$ of the single element and the one in the array are essentially the same and that the mutual coupling between the neighboring elements are below $-30$ dB in the frequency interval dashed in grey indicating the desired operating bandwidth $9.875 - 10.1$ GHz. Note also that in this bandwidth the reflection coefficient of each aperture is below $-10$ dB. This implies a mismatch loss in the order of 0.5 dB at the beginning of the useful band.

B. Primary Patterns

The simulated primary radiation patterns (directivities) from each waveguide feed are shown in Fig. 8a), b), c) as a function of the frequency from 9.8 to 10.1 GHz, in the E-, H- and D-plane, respectively. Note that the directivities do not include the mismatch losses, which are the main cause of losses at lower frequencies. It is apparent that for higher frequencies the beams flatten and that the peak directivity is reduced. This is because the overall enhanced patterns are shaped by leaky waves whose pointing directions are further away from broadside as the frequency increases. This phenomenon is more evident at frequencies as high as 10.1 GHz, where the leaky wave contribution can be recognized in each $\phi$-cut around $\pm\theta_{LW} \approx 10^\circ$. This induces lower directivity at broad side. The fact that the primary beam degrades has a negative impact because the edge of the reflector, located at $17^\circ$, is illuminated with relatively higher field levels with respect to axial direction. This is not a big problem in terms of spill over, because the largest portion of the power is still intercepted by the reflector. In fact the edge of coverage remains high, as was apparent from Fig. 3. However this results in higher side-lobes in the secondary patterns. Fig. 8 also highlights the good pattern symmetry of the structure with cross polarized fields remaining low for all azimuthal angles [17].

C. Secondary Patterns

The simulated secondary beams generated by a reflector $(F/D = 1.71, \ F = 245.5 \lambda_0)$, when excited by the radiation patterns of Fig. 8 are plotted as a function of the frequency in Fig. 9. Apart from the edge of coverage gain, one of the key parameters that quantifies the quality of secondary patterns is the isolation between beams using the same polarization and same frequency. A first order approximation can be obtained assuming that all the beams are identical to the central one. In this case it is relevant to consider the isolation between the main beam and its first image centered at two angular periods of distance $2\Delta\theta = 1.12^\circ$. This exercise is also summarized in Fig. 9 for the highest frequency in the band which appears to be the one that presents relatively high side lobes associated with the beam degradation (splitting) that was observed in the primary beams. In Table I the most significant system parameters for the investigated structure are summarized, considering $F/D = 1.71$.

IV. Prototype Manufacturing

Even though the motivation of the work was a telecom scenario at Ka-band, a prototype at the scaled frequency of 10 GHz has been designed. A prototype array of 19 waveguide feeds was manufactured in the TNO internal workshop. A picture of the final hardware is shown in Fig. 10 which highlights the waveguide feeds, the ground plane, the super-layers and the filtering structures. The waveguide feed connecting the filters to the radiating ground plane, is only 17 mm long.

![Fig. 10. Back-view picture of the feed array manufactured and mounted. It includes the dielectric super-layers, the iris ground plane, the waveguide feeds and the band-pass filters.](image)

The dielectric super-layers $(60 \ \text{cm} \times 60 \ \text{cm})$ were custom made on borosilicate glass by ECN (http://www.ecn.nl/home/). The dielectric layers were over dimensioned in order for the edge effects to be completely negligible. Since the structure is resonant with overall relative bandwidth of about 2 $\%$, the separation between the ground plane and the first sheets, nominally $h = 15$ mm, should be accurate to 0.015 mm or less in order not to observe a significant frequency shift in the measured parameters. This accuracy is comparable to the
TABLE I
PARAMETERS CHARACTERIZING THE SECONDARY BEAMS FROM EACH FEED, $F/D=1.71$

<table>
<thead>
<tr>
<th>$f_{\text{req}}$ (GHz)</th>
<th>$D_{\text{max}}$ (dB)</th>
<th>$D_{\text{eoc}}$ (dB)</th>
<th>Roll - off (dB)</th>
<th>Isolation (dB)</th>
<th>X - Pol (dB)</th>
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<td>9.9</td>
<td>51</td>
<td>45.4</td>
<td>5.6</td>
<td>&gt;20</td>
<td>&lt;−30</td>
</tr>
<tr>
<td>10.0</td>
<td>51.5</td>
<td>45.7</td>
<td>5.8</td>
<td>16</td>
<td>&lt;−30</td>
</tr>
<tr>
<td>10.1</td>
<td>52</td>
<td>45.2</td>
<td>6.8</td>
<td>13</td>
<td>&lt;−30</td>
</tr>
</tbody>
</table>

Fig. 11. Front and side view of the feed array manufactured and mounted. On the top right the micro-metric tuning mechanism is visible, while in the bottom right the two dielectric layers are explicitly visible.

one required for the filters. The thickness of the slabs received was verified to be as requested, $h_d = \lambda_d/4 = 3.53$ mm. The separation, $h_s$, between the two dielectric slabs was frozen to its nominal value at the design frequency of 10 GHz, $h_s = \lambda_d/4 = 7.5$ mm. Details of the slabs are visible in Fig. 11. The dielectric constant of the received material was measured to be $\varepsilon_r = 4.5$, accurate to the second significant digit, with very small imaginary part. To allow for small variations in the dielectric constant, micro-metric screws were used to fine tune the distance between the ground plane and the ensemble of the two dielectric slabs so that the radiation frequency band would be in correspondence with the pass band frequencies of the filters.

A. Filtering Structures

The filters are realized in waveguide technology. The total frequency band is divided in two sub-bands of widths $\Delta B = 90$ MHz and centered at $f_{c1} = 9.92$ GHz and $f_{c2} = 10.055$ GHz, with a roll off band $\Delta R = 45$ MHz that separates the two useful bands.

A picture of one of the low frequency band pass filters before mounting is shown in Fig. 12. The filters are realized by cascading 5 cavities realized in WR 90 waveguides. The filters are obtained by machine milling solid steel blocks. Although this is not a manufacturing procedure that is considered for space flight hardware, it is representative for the RF prototype. Overall 19 filters were manufactured. A close view of the mounted filters is shown in Fig. 13. The measured $S$ parameters of both filters are shown in Fig. 14. The smooth variation of the phase of the reflected signals in the reflection bandwidths is also presented in Fig. 14b.

It is apparent that the band pass behavior of the two filters is very well behaved, with reflection coefficients in band lower than $-20$ dB and out off band rejections lower than $-25$ dB. It appears that with respect to the desired design bands, the present technological implementation of the lower band filter presents a 5 MHz shift toward higher frequencies. A relative shift of 0.05% appears unavoidable with the technology used at TNO and in absence of tuning screws. Overall the realization
of the filters was considered satisfactory.

V. MEASURED PERFORMANCES

After the filters were characterized as a self standing unit, the system (filters+feeds+super-layers) was assembled so that the performance of the entire primary feed structure could be evaluated. Before freezing the actual measurement geometry the separation between the metallic ground plane and the lowest slab, was fine tuned around \( h = 15 \) mm. In order to decide the most appropriate distance two sets of measurement of the copolarized radiation patterns at two frequencies were done. With respect to Fig. 15 the elements chosen for the fine tuning were the first (1) and the fourth (4).

The far field patterns were obtained by post processing after the near field was sampled with a commercial near field scanner.

Fig. 16 shows the low frequency (9.9 GHz) E-plane copolar primary feed radiation patterns associated to the central element for different tuning heights. It is apparent that for all heights the radiation patterns have very similar shapes. The measured gain, in this graph, includes the filters and the mismatch losses. For the smallest height, \( h = 14.93 \) mm the gain is almost 2 dB lower than for the higher values, highlighting the effects of the mismatch losses, and also the dependence of these losses as a function of frequency. The response of the higher frequency radiator (4) is complementary. The patterns at 10 GHz are shown in Fig. 17. It is apparent that the higher slabs give rise to much reduced gain at broadside which in turns implies higher side lobe from the secondary patterns as was discussed in section III.

Eventually it was decided that the best compromise distance to evaluate the system performance of the array was the one that was measured as 14.93 mm. This choice reduces slightly the actual gain in the lower band due to mismatch losses but guarantees that at the highest frequency the isolation of the beams is still within specification.
A. S-Parameters

Once the entire array was assembled and the $h$-value frozen the S parameters of the waveguide array, with the inclusion of the filters were measured. The measured $S_{ii}$ parameters for two feeding waveguides operating at two different central frequencies are reported in Fig. 18. The feeds considered are the ones indicated as "1" through "7" with reference to Fig. 15. The results indicate that on the lower band the reflection coefficients are approximately ($\Gamma_{dB} \approx -8.5$ dB) which correspond to about 15% of efficiency reduction ($1 - \Gamma^2$). In the high band the reflection coefficients are in the order of $-15$ dB, about 3% of efficiency reductions. The mutual coupling between ports (after filtering) was lower than $-50$ dB. This means mutual coupling can be neglected for the system.

Fig. 17. E-plane copolar primary feed radiation patterns associated to a lateral element (4), operating in the higher band, for different tuning heights.

Fig. 18. The measured S parameters pertinent to the array.

Fig. 19. The measured copolar and cross polar primary beams generated by the element 1 of the array plotted for three different frequencies within the lower frequency band on the: a) E-plane b) H-plane c) D-plane respectively. The scale in dB represents gain.
Fig. 19. The copolar radiation pattern for different azimuthal angles, $\phi = 0$, 30, 60, 90 are plotted for the frequencies: a) 9.875 GHz representing the lowest measured and useful frequency and b) 10.1 GHz representing the highest and useful frequency. The scale in dB represents gain.

**B. Measured Patterns**

Figures 19 and 20 show the measured primary co and cross polar radiation patterns from element 1 and 5, respectively. The Ludwig 3 definition has been used for defining co and cross polarization [18]. The two figures are pertinent to the E, H, D ($\phi = 45^\circ$) planes respectively and are shown for different frequencies all within the band allowed by the lower or higher frequency filter, respectively.

The figures show that the radiation patterns are very symmetrical, characterized by very low cross-polarization and that the frequency behavior is essentially as predicted. However, a difference of 1.5 dB in the gain at the lower frequencies is observed between Fig. 19 and the directivities calculated curves in Fig. 8. The present implementation in fact includes 1 dB of
losses associated to the ohmic losses in the filter and the return loss. The latter is frequency dependent and amount to up to 0.6 dB in the lower frequency band. Once the mismatch and ohmic losses are included in the calculations the agreement between calculation and measurements is within 0.5 dB in the lower frequency band. For the highest frequencies, since mismatch losses are negligible, the agreement between calculations and predictions is good. It is worth noting that the patterns from all copolarized and same frequency elements are very similar. This implicitly means that the interaction among neighboring elements can be neglected and the equivalent array in Fig. 4b was eventually achieved. The copolar, with respect to the Ludwig 3 definition, radiation pattern for different azimuthal angles, $\phi = 0^\circ, 30^\circ, 60^\circ, 90^\circ$ are plotted in Fig. 21 for the frequencies: a) 9.875 GHz representing the lowest measured and useful frequency and b) 10.1 GHz representing the highest measured and useful frequency.

The cross-polarization levels are also very low as anticipated in the calculations performed using CST [15]. The secondary beams generated by a reflector ($F/D = 1.71, F = 245.5A_0$), when excited by the measured primary patterns are plotted as a function of the frequency. Note that the isolation of $-16$ dB can be estimated by plotting the adjacent beam centered at $\theta_1 = 1.12^\circ$.

C. Comments on the Measurements

The comparison between measured and calculated directivities as a function of the frequency demonstrates the reproducibility of the calculations. Moreover the combined use of the calculated edge of coverage directivity in Fig. 3 and the inclusion of the ohmic and mismatch losses also allows one to obtain an accurate prediction of the edge of coverage gain. On top of these losses also the filter insertion loss will contribute for approximately a dB depending on the chosen material for fabricating the filters. Considering the complexity of the experiment, the agreement between measurements and simulation is good. The specific objective was to design a system meeting real mission specifications over a bandwidth of 2%. It has to be concluded that such a bandwidth is stretching the capabilities of such a resonant structure. In fact for the low frequency filter the overall matching is the order of $-8.5$ dB, which implies $-0.6$ dB of a reduction in gain. This gain reduction reduces partly the gain enhancement (in the order of $2$ dB) achieved thanks to the pattern shaping.

VI. CONCLUSIONS

A leaky wave enhanced focal plane feed array has been designed, manufactured and measured. The array is composed of 19 very compact iris-loaded waveguides operating in X band. It is a scaled version of a system that would be suited to feed a multi beam reflector system with the purpose to provide contour coverage of portions of the earth from a geostationary telecommunication satellite. The stringent manufacturing tolerances required to achieve $20$ dB gains from narrow band structures, resulted in a challenging engineering exercise.

Overall the measurements provided in this paper demonstrate that it is possible to use the design methodology described in [2]-[5] for reflector systems characterized by large $F/D$ (in the order of 2). Thus the performances of a multi beam single-feed-per-beam reflector system can be significantly enhanced to the level of becoming comparable to the ones of existing systems that rely on arrays of four reflectors, on a BW in the order of 2%.

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REFERENCES

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Dr. Ettorre received the Young Antenna Engineer Prize during the 30th ESA Antenna Workshop 2008 in Noordwijk, The Netherlands.