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Diagonal Polarized Pin-Based Cavity Antenna Array in Planar Technology

R.J. Bolt^{#1}, S. Bruni^{*2}, A. Neto^{#3}, G. Gerini^{#4}

[#]TNO Defence, Security and Safety

Oude Waalsdorperweg 63, 2597 AK Den Haag, The Netherlands

¹roland.bolt@tno.nl

³andrea.neto@tno.nl

⁴giampiero.gerini@tno.nl

^{*}Formerly at TNO now with Calearo Antenna S.p.A.
Via Bacchiglione 49, 36033 Isola Vicentina (Vi), Italy

²sbruni@calearo.com

Abstract— This contribution describes the design and characterization of a very compact 8-element proximity fed cavity antenna array realized completely in PCB technology. The cavities are formed by the placement of a series of through pins (vias) along a rectangular outline. Micro-strip lines excite the structure, and slot coupling is used to feed the individual radiating cavities. The elements are linearly polarized, but are arrayed with a diagonal orientation. The configuration allows for defining a single multifunction system that combines functionalities as radar and ESM in X-band. A -10 dB bandwidth of near 20% is achieved with the antenna and -10 dB for $|\theta_s| < 50^\circ$ over the majority of the design band. The main achievement of the array is the extreme efficiency, also in a design that resorts to very dense dielectrics.

I. INTRODUCTION

An ever increasing amount of sensors are placed on military platforms. For successful missions these sensors are indispensable in the operating environments of today. Such sensors, amongst others, enable functionalities like navigation, communication and observation. The complex ensemble of individual systems requires a large portion of the available platform area. Apart from this, it will inherently increase system weight and power consumption, and additionally, the probability of EMI will grow larger.

It is therefore straightforward to pursue the integration of functionalities wherever possible, so that the aforementioned aspects may be significantly reduced. A possible combination is found in ESM (0.5 – 18 GHz) and X-band volume search radar. The remainder of this contribution is focused on the design and characterisation of an antenna for this combination. From both functionalities a set of requirements can be derived that is necessary for the design of the antenna; this will be treated in the next section. The design of an antenna for the conceptual combination will give an idea of the level of integration that is achievable with the proposed technology.

II. ANTENNA REQUIREMENTS

For the realisation of a multi-function system that combines FMCW radar with ESM, a wide-band antenna array needs to be designed. With the radar operating in the lower section of the X-band (9 – 10.5 GHz) and the ESM throughout it, we can

readily identify a number of requirements. Due to the ESM application, the antenna is a linear and diagonally polarized array. Since components used in front-ends become smaller, aspects like compactness and weight of the antenna had to be addressed as well. It is recognised that demands on the antenna gain (efficiency) put a physical minimum to its radiating aperture size. However, the longitudinal dimension, from front-end output to aperture, mostly allows for a degree of freedom.

For the current application there is no need to limit the beamwidth in the vertical direction due to the volume search type of operation. An opening angle in the order of 90° is sufficient and therefore a linear array can be considered. Eight transmit modules and an equal number of receive modules will perform the separate modes of operation hence two arrays of 8 elements are needed. The system should be able to locate a source at an angle of 60° offset from the array normal.

III. ANTENNA ELEMENT DESIGN

The challenging aim is to design a linear antenna array with a bandwidth larger than 15% which achieves scanning up to 60° throughout the frequency range. The FMCW radar band will be located around $f = 9.4$ GHz. To make the antenna design compact and simple to build it is designed in planar technology. The envisaged configuration is shown in Fig. 1.

Fig. 1 Cross section and top view of the radiating element. Slot dimensions: l_s , w_s ; cavity dimensions: l_{cav} , w_{cav} ; feeding micro-strip line width: w_m and its stub length: l_{stub} . Feed line substrate thickness: h_m , and superstrate thickness: h . Two ground planes are present, one on top loaded with an aperture and another one loaded with a slot.

The lower substrate carries the feeding micro-strip line. Energy from the feed line is coupled through a slot etched in the ground plane in the direction of the superstrate. From there the transmitted energy is redirected by the structure on top for radiation into free space.

To excite the radiating elements, proximity feeding was used. This exhibits in general good performance over a larger frequency band when compared to direct coupling via a probe or micro-strip line [1], [2]. We have used a two substrate configuration separated by a ground plane rather than a classical tri-plate feed built-up. This eliminates the extra ground plane (backing reflector) and limits mutual coupling at the level of the feeding network. The omitting of the backing reflector can be justified when: the substrate material is relatively dense and the slot non-resonant, i.e.: this will increase the front-to-back ratio (FTBR).

The choice of the material permittivity is, apart from the FTBR, also dictated by the bandwidth and scan angle requirements. The FTBR should be sufficiently high and reasonably $\text{FTBR} > 15$ dB. Higher permittivities would result in a higher FTBR, provided that its thickness is maintained at a fixed electric height. It is well known that increasing the permittivity will be at the cost of bandwidth. For our case a good and pragmatic value of the superstrate permittivity is: $\epsilon_r = 9.8$ [3]. To achieve a high transmission of the power from the micro-strip line, through the slot into the superstrate, it should be relatively thick, in the order of $h \sim 0.25\lambda_d$, introducing a good match to free space. λ_d is the wave length in the dielectric. The requirement on the scan angle will be treated in the next section.

The indicated thickness results in a good transmission over a large bandwidth, i.e.: $> 40\%$ [3]. However, it will be at the cost of surface waves (SWs) being readily introduced in the superstrate. To render the antenna array efficient, as a next step, cavities have been introduced to prevent SW propagation. The cavities are created by the placement of vias (Fig. 1) placed along a rectangular outline. The inter-via spacing is chosen small enough ($< \lambda_d/10$) to imitate a solid electric wall.

IV. ANTENNA ARRAY DESIGN

In the previous section a third aspect was mentioned regarding the permittivity of the superstrate. It referred to the maximum scan angle the array has to achieve. From the requirements of section II, this should be $\theta_s = 60^\circ$. It must be recognized that this is a stringent requirement for arrays based on planar technology, and inherently demands the taken precautionary measures to limit SW propagation.

Standard evaluation of the maximum inter-element distance (dx) results in a 14 mm spacing ($f_{\max} = 11.5$ GHz). The given spacing puts a hard limit to the size of the cavity. Mind that each single cavity design must also incorporate the extra thickness of two times the diameter of a via to introduce its walls, and an additional spacing between two subsequent cavities with respect to the manufacturability. Since the antenna should facilitate the integration of ESM, implying a diagonal polarization, somewhat alleviates this problem as shown in Fig. 2. The expression for the maximum possible

width w_{cav} of the cavity, without obstructing its neighbors is given in (1).

Fig. 2 Considered geometry for the radar-ESM array. The cavities are oriented at $\phi_c = 45^\circ$ for diagonal polarization

$$dx = [w_{cav} + \Delta w_{cav} + dy'] \sqrt{2} < \frac{\lambda_0}{1 + \sin \theta_s} \quad (1)$$

The dimension Δw_{cav} represents the extra spacing required for the vias, w_{cav} is the cavity width. The spacing between the cavity walls is dy' . The resulting allowable width becomes: $w_{cav} < 8.7$ mm.

For the final design of the antenna elements and array the commercial tools [4] and [5] were used. In Fig. 3 the effect of introducing the cavity is shown. Its presence makes the antenna element much more efficient. The efficiency ϵ_{rad} is defined through: $G = \epsilon_{rad} D$ with G and D the gain and directivity at broadside respectively. The gain for the slot alone is on average below 0 dB, the directivity is on average 6 dB, which means that only $1/4$ -th of the power is actually entering free space while the rest is confined to the superstrate. Still the FTBR is on average 10 dB over the band. Introducing the cavity, the effect of “capturing” the surface waves is readily seen: the gain and the directivity curves are almost superimposed, meaning an efficiency of $\epsilon_{rad} \approx 1$. Since now the majority of the power is available for radiation into free space, the FBTR has been greatly improved.

Fig. 3 [5] The effect of loading the slot with the cavity: efficiency and front-to-back ratio.

V. HARDWARE

The antenna is based on Rogers TMM10i ($\epsilon_r = 9.8$). As bonding material Rogers 4450B-DX ($\epsilon_r = 3.3$) was used. This material was proposed by the manufacturer for assembling multilayer TMM10i. This thermoset material ensures no

delaminating effects during the process of soldering the connectors. A photograph of the hardware is shown in Fig. 4. The dimensions (nomenclature Figure 1) are: $l_s = 9$ mm, $w_s = 1.2$ mm; $l_{cav} = 17$ mm, $w_{cav} = 8.5$ mm; $w_m = 0.8$ mm and $l_{stub} = 3$ mm. Total thickness of $h_m = 0.381$ mm + 0.2 mm = 0.581 mm and $h = 2.54$ mm. The vias are of a diameter $d_{via} = 0.4$ mm ($\frac{1}{2}\Delta w_{cav}$).

Fig. 4 The produced 8-element cavity array. Back- and front view. The inset shows clearly the vias which form the cavities. The apertures are numbered 1 to 8 from left to right.

VI. CHARACTERIZATION OF THE ARRAY

The realized planar array antenna has been characterized by measuring the scattering parameters and the embedded element patterns. The measured scattering parameters are shown in Fig. 5.

Fig. 5 [4] Plots of the different measured and simulated scattering parameters. Simulated curves: S_{11} (solid black), S_{55} (solid light grey), S_{88} (solid grey), S_{43} (solid thick black), S_{41} (lower solid thick grey). Measured data: broken lines, match with color.

To get an idea of the scanning behavior of the linear array, the active reflection (AR) coefficient of an array element has been considered. The AR at any antenna element input can be found through:

$$\Gamma_n(f, \theta_s) = \sum_{m=1}^M S_{nm} |a_m| |a_n|^{-1} e^{j2\pi f c_0^{-1} (n-m) d_x \sin \theta_s} \quad (2)$$

where $n, m \in \{1, 2, \dots, M\}$ with $M = 8$, c_0 is the free space velocity, and a_p presents the incident signal amplitude at port p . Then, using (2) the active n -th element input impedance is found as:

$$Z_{in;n}^{act}(f, \theta_s) = Z_0 \frac{1 + \Gamma_n(f, \theta_s)}{1 - \Gamma_n(f, \theta_s)} \quad (3)$$

For several frequencies throughout the band (3) has been computed. Subsequently, the active element has been matched to the broadside impedance, i.e.:

$$\bar{\Gamma}_n(f, \theta_s) = \frac{Z_{in;n}^{act}(f, \theta_s) - Z_{in;n}^{act}(f, 0)}{Z_{in;n}^{act}(f, \theta_s) + Z_{in;n}^{act*}(f, 0)} \quad (4)$$

In (4) the asterisk represents the complex conjugate. The resulting curves are shown in Fig. 6, and regard the fourth element (see figure's inset) of the array. These AR's are based on calculated data only because not all parameters have been measured.

Fig. 6 Graph showing for several frequencies the active reflection coefficient related to element 4 of the array (simulated data).

As a final consideration the radiation patterns are treated. Each individual radiating element has been tested in a near-field measurement range. For brevity only the embedded patterns of element 4 are shown for $f = 9.4$ GHz, Fig. 7. Experimental and computed co- and cross-patterns are plotted for both the horizontal (H-) and vertical (V-) plane (inset).

Fig. 7 [4] Graph showing the measured and computed embedded element (nr. 4) patterns at $f = 9.4$ GHz. Solid is V-plane, dash-dot is H-plane (see inset). The thick curves represent the co-polarized patterns; the thin curves represent the cross-polarized patterns. Black: computed, grey: experimental.

For the cross-polarized patterns, Ludwig's third definition has been used.

Since we were not able to measure a fully excited array, the radiation patterns under those conditions have been computed. The related patterns for both broadside and scanned radiation are shown in Fig. 8 for a single frequency ($f = 9.4$ GHz). The radiation patterns computed with use of the measured embedded element patterns at the same frequency show very comparable results for broadside radiation. In case of scanning the main beam's shoulder is absent and the side lobes are a few decibels higher.

Fig. 8 [4] Graph showing the computed gain patterns for a fully active array at 9.4 GHz. Both broadside and scanned ($\theta_s = -60^\circ$) patterns are presented for the H- and V-plane. Solid is co-polarized; dash-dot is cross-polarized.

For the considered case of extreme scanning, i.e.: $\theta_s = -60^\circ$, the radiation patterns show an appreciable change in shape as a function of the operating frequency. This change is mainly experienced in the vicinity of the main beam. In the graph of Fig. 9 the co-polarized far-field patterns in the horizontal plane, under scanning conditions, have been plotted for a

Fig. 9 [4] Graph showing the computed gain patterns (co-polarized) for a fully active array under scanning conditions ($\theta_s = -60^\circ$) for several frequencies. Patterns extracted from the H-plane.

number of frequencies (see figure's inset).

For the higher frequencies the radiation pattern becomes clearly more directive due to a diminishing amount of power being spilled over the edge of the array. The increase in effectively radiating aperture and the array's edge being relatively farther away with frequency increase, result in a more distinguishable main beam.

VII. DISCUSSION

A compact planar array of slot coupled radiating cavities has been designed and tested. The array achieves a -10 dB bandwidth of near 20%. Under scanning conditions the considered active reflection coefficient remains below -10 dB over an average scan range of $|\theta_s| < 50^\circ$ for a very large portion of the frequency band. Only for a very small band around $f = 10.4$ GHz this scan range becomes more limited: $|\theta_s| < 30^\circ$, only to increase again for the higher frequencies. For this frequency an increase in the mutual couplings is indeed observed. A tolerance study pointed out that this deviation is due to a change in the applied glue layer thickness.

The average front-to-back ratio is 17 dB which was sufficient for the current application. The relative high level of cross-polarization (13 dB), has a maximum in the direction of the main beam, and is most probably due to the existence of (symmetric) higher order modes. These modes result from the asymmetric positioning of the apertures in the array. However, the cross-polarization purity was not a driver in this design.

The patterns for the fully excited array show the array end-effect for extreme $\theta_s = 60^\circ$, which can also be seen from the embedded patterns. Gradually with increase in frequency the main beam becomes more pronounced. For $f > 10.2$ GHz eventually the pattern shoulder disappears and a clear main beam can be distinguished.

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