Wide-Angle Scanning Cavity Antenna Element for Mobile Satcom Applications at Ka Band

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Abstract—A novel dual linearly polarized cavity antenna with excellent scan performance up to 60° in both E- and H-plane is presented. The highly integrated antenna element for Ka band Satcom on the move applications is operating in the uplink band at 30 GHz. Vertical stubs incorporated into the 3-D architecture are proposed to compensate parasitic effects limiting the scan volume of planar phased array antennas. To verify this approach, the vertical transition used in the design, and the entire antenna structure in single element configuration were fabricated and measured successfully.

Index Terms—dual-polarized; cavity antenna; planar antenna; phased array; scan blindness; vertical transition; quasi-coaxial-line; satcom.

I. INTRODUCTION

The promising perspective to provide high data rate communications links for a large variety of end users have become one of the main driver for new generation satellite communication systems at Ka band [1]. In mobile scenarios, where the antenna terminal is typically deployed on highly dynamic platforms such as airplanes, the system architecture needs to be low profile and lightweight to minimize its impact on the mechanical structure. Moreover, the antenna system in mobile scenarios has to track the satellite constantly in order to guarantee the service availability while moving. As a consequence of the indicated mechanical constraints, electrically steerable antennas are preferred over mechanical ones.

Therefore this paper reports on a dual linearly polarized cavity antenna for the uplink band at 30 GHz. It is shown that the investigated unit cell design is suitable for 2-D electronic beam steering up to 60° from boresight in all planes. In addition, high port-to-port decoupling is maintained within the entire scan volume which makes the element configuration suitable for integration into various architecture e.g. direct radiating [2] or reflectarray antennas [3].

II. UNIT CELL DESIGN

In this study the antenna element for Satcom on the move applications at Ka band is intended to operate in the uplink band from 29.5 GHz to 30.8 GHz. The unit cell lattice is kept reasonable small to cover a large scan volume down to 60° from boresight in all planes. As a trade-off between the scan performance and the integration level of the prospective antenna system, the element spacing is set to $l_{uc}=5$ mm using a non-skewed square lattice.

A 3-D multilayer architecture based on PTFE materials has been chosen to integrate all required components within the

Fig. 1: Exploded view of the proposed cavity antenna composed with the dielectric materials ①: Rogers RT Duroid 5880 ($h=1.524$ mm, $\epsilon_r=2.2$); ②: Rogers RT Duroid 5880 ($h=0.381$ mm, $\epsilon_r=2.2$); ③, ⑤: Rogers RO3003 ($h=0.127$ mm, $\epsilon_r=3$); ④: Rogers RO3003 ($h=0.508$ mm, $\epsilon_r=3$).

Fig. 2: Cross section view of the planar cavity structure including vertical stubs in the dielectric (3 unit cells).

very limited lateral extend. The exploded view of the proposed dual-polarized cavity antenna and its material composition is shown in Fig. 1.

The layer stack up is subdivided into 2 parts in order to reduce to complexity of the planar antenna structure. Subsequently, both boards are joined together using a conductive bonding foil. The cavity is formed within the top laminate ① by a series of surrounding through-hole vias. In a way different from recent published work on planar cavity antennas [4], [5] the upper rim is not congruent with the lower one. Following this novel approach, the proposed antenna structure provides additional vertical stubs in the dielectric (see Fig. 2). By adjusting its width and depth coupling between the radiating and evanescent modes in the aperture plane can be controlled

such that power is transferred to the dominant Floquet-Mode more efficiently, even for larger scan angles.

The lower multilayer PCB contains the ring slot aperture fed by a stripline for both polarizations. Typically the circumference of the ring slot is about one guided wavelength to obtain an inherently good isolation between both feedings as in this configuration the electrical field of the fundamental mode vanishes at the orthogonal feed position. On the other hand, coupling to the cavity tends to increase for smaller apertures. Hence, a reduced ring slot in conjunction with reactive tuning stubs opposite to the striplines have been introduced to enhance the port-to-port decoupling. The stripline is connected to the respective microstrip line on the bottom layer by a vertical RF-transition, which has been realized as quasi-coaxial line using 4 ground vias [6]. Additional shielding vias around the unit cell are employed to suppress parasitic coupling between the individual elements. Due to the discussed area constraints these are shared partially with the ground vias of the vertical RF-transitions.

A parametric analysis and optimization has been conducted with Ansys HFSS [7] assuming infinite periodic boundary conditions. The parameters of the proposed dual-polarized cavity antenna are shown in Fig. 3 and the corresponding dimensions are listed in Tab. I. Its noteworthy that the structure is symmetrical with respect to the both excitations.

Fig. 3: Parameters of the proposed dual-polarized cavity antenna (Dimensions of both ports are identical).

l_{uc}	5.0 mm	l_{cav1}	3.25 mm	w_{rim1}	0.6 mm
d_{via1}	0.30 mm	d_{via2}	0.40 mm	d_{via3}	0.4 mm
d_{via4}	0.40 mm	d_{gvia1}	0.625 mm	d_{pad1}	0.5 mm
d_{pad2}	0.60 mm	d_{slot1}	0.8 mm	d_{slot2}	1.0 mm
l_{slot1}	1.67 mm	w_{slot1}	0.42 mm	l_{slot1}	1.0 mm
w_{sl1}	0.4 mm	w_{sl2}	0.44 mm	l_{sl1}	0.8 mm
l_{sl2}	0.52 mm	w_{sl3}	0.75 mm	l_{sl4}	1.0 mm
w_{ms1}	0.30 mm	l_{ms1}	1.0 mm	-	-

TABLE I: Dimensions of the dual-polarized cavity antenna.

The simulated reflection coefficients of the presented dual-polarized cavity antenna are depicted in Fig. 4 for scanning in the E- and H-plane, respectively. In both principle planes the reflection coefficient remains below -10 dB from 29.2 GHz to 31.0 GHz for wide scanning up to 60° off-boresight. Due to the symmetrical arrangement of the two feeding structures their results are identical and results for the second port have been omitted in the following.

Fig. 4: Simulated reflection coefficient for various scan angles.

During the design process the emphasis was on the compensation of surface and leaky wave resonances as they typically appear in phased array antennas before the onset of grating lobes [8]. The impedance mismatch of those parasitic effects which manifests close to the operational band at 60° are apparently more improved in the E-plane by means of the vertical stub.

Fig. 5 illustrates the simulated transmission coefficients from port 1 to the dominant Floquet mode $TM(0,0)$ and $TE(0,0)$ for scanning in the E- and H-plane, respectively. Since the simulation model contains lossy materials and multiple radiating Floquet modes, the transmission coefficient provides a more precise evaluation criterion in terms of radiating efficiency [9]. The total scan degradation of the antenna can be

Fig. 5: Simulated transmission coefficient for various scan angles.

easily expressed by the normalized embedded element pattern as follows:

$$G_E(\vartheta) = |T(\vartheta)|^2 \cos(\vartheta), \quad (1)$$

where $T(\vartheta)$ is the transmission coefficient to the designated Floquet mode.

For scanning in the E-plane up to 60° from boresight the transmission coefficient is better than -0.75 dB in the operating bandwidth from 29.5 GHz to 30.8 GHz. In comparison, the scan performance in the H-plane is slightly lower with a minimum coupling coefficient to the $TE(0,0)$ Floquet mode of -1.0 dB at 30.8 GHz. Thus the proposed cavity antenna exhibits a maximum scan degradation of only 4 dB in the intended scan volume.

The port-to-port coupling $|S_{21}|$ of the dual linearly polarized cavity antenna element is depicted in Fig. 6. It remains below -20 dB in the frequency band from 28.6 GHz to 31.2 GHz when scanning up to 60° . The small dip at 32.15 GHz that determines the upper frequency bound is associated with a singularity in the admittance plane whenever the $TM(-1,0)$ or $TM(0,-1)$ Floquet mode enters the visible region [10]. In contrast, the lower frequency bound is found in the reduced

Fig. 6: Simulated port-to-port coupling coefficient for various scan angles.

isolation of the ring slot aperture.

III. EXPERIMENTAL RESULTS

The experimental characterization of the proposed cavity antenna has been performed in two parts. For the first tests, the vertical RF-transition has been assembled separately and the cavity antenna was realized in a single element configuration. The measurement results of both individual components are discussed in the following.

A. Vertical RF-transition

The vertical RF-transition has been fabricated and assembled in a back-to-back configuration as shown in Fig. 7. Two versions with different length of the stripline were realized to obtain the de-embedded S-parameter results of the single microstrip-to-stripline transition.

Fig. 7: Photograph of the vertical RF-transition in a back-to-back configuration.

The measured and simulated S-parameters of the vertical RF-transition based on a quasi-coaxial-line are presented in Fig. 8. The measurement results of the transmission structure

(a) Reflection coefficient

(b) Transmission coefficient

Fig. 8: Measured (de-embedded) and simulated S-parameters of the single vertical RF-transition.

show excellent agreement with the simulation over a large bandwidth, and the transition can be potentially used for uplink (30 GHz) as well as for downlink (20 GHz) architectures. In the band of interest (29.5 GHz–30.8 GHz) the measured reflection coefficient is strictly below -18.4 dB and the maximum transmission loss is 0.55 dB. In order to relax the area consumption of the vertical transition within the unit cell its ground vias are merged with those from neighboring elements (see inset in Fig. 7). In return this approach comes along with an increase of mutual coupling. However, the measured coupling coefficient $|S_{32}|$ to the adjacent transition is still lower than -32.5 dB at 30.8 GHz and indicates that the possible impact on the array performance is neglectable.

B. Single element configuration

In the next step, the proposed dual-polarized cavity antenna has been realized as a single element. Fig. 9 shows the measured reflection coefficient of both ports. The 10 dB return loss bandwidth is covered by both ports from 28.2 GHz to 33.2 GHz, and in the operational band the maximum reflection coefficient reaches -14.8 dB. In comparison to the simulations a slight shift of 800 MHz to lower frequencies is observed. Since the proposed antenna structures is composed of a relatively thick multilayer PCB, the reason may be found in the accumulation of material and fabrication tolerances.

The port-to-port coupling coefficient of the proposed cavity antenna is illustrated in Fig. 10. As can be seen from the measurement results the coupling level is well below -20 dB in the frequency band from 29.5 GHz to 30.8 GHz. Especially for the integration into active reflectarray antennas, which employ a unit cell topology of the cross-polarization type, this antenna parameter is crucial. Therefore the decoupling between both ports of the dual-polarized unit cell has to be sufficiently high because this defines the array's maximum permissible amplification [3], [11].

The characterization of the single dual-polarized cavity antenna in the far field was conducted for both ports in the entire uplink band. As an example, the radiation diagrams at

Fig. 9: Measured and simulated reflection coefficient of the dual-polarized cavity antenna element.

Fig. 10: Measured and simulated port-to-port coupling coefficient of the dual-polarized cavity antenna element.

the center frequency (30 GHz) for port 1 are shown in Fig. 11. The measured far field pattern exhibits ripples in both E- and

H-plane that can also be observed from the simulations. This common effect in small antenna configurations is originated from parasitic radiation at the edges of the finite substrate. Consequently, its impact on the radiation diagram degrades gradually with increasing number of elements and can be typically neglected in larger phased array antennas as considered in this work.

Fig. 11: Measured and simulated radiation diagram of the dual-polarized cavity antenna element at 30 GHz (port 1).

IV. CONCLUSION

A first prototype of a dual-polarized cavity antenna element for wide-angle scanning at 30 GHz is presented. Additional vertical stubs have been incorporated into the highly integrated planar structure to adjust the modes representing the electromagnetic field at the aperture plane. By the proposed design parasitic effects limiting the intended scan volume are suppressed effectively. As result, the unit cell element is able to operate efficiently from 29.2 GHz to 31 GHz with scan angles up to 60° off-boresight in both E-and H-plane.

The vertical RF-transition based on a quasi-coaxial line has been fabricated separately in a back-to-back configuration. The return and insertion loss obtained from the measurements are better than 18.4 dB and 0.55 dB, respectively. Furthermore, the proposed cavity antenna has been fabricated and assembled successfully in a single element configuration. The measured reflection coefficient remains below -14.8 dB for both ports and first measurements of the radiation diagram are presented. The experimental verification of the excellent scan performance will be evaluated in future work.

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