Antenna Arrays for RF-MEMS based 77 GHz On-Board Wake Vortex Detection Sensor

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Short Abstract—**In this paper the first generation of transmit and receive antennas for an on-board wake vortex detection front end module at 76.5GHz is presented. The antennas are realized with serially fed patch antenna arrays connected in parallel by a corporate feed network. The use of RF-MEMS phase shifters between the feed network and the 2D patch antenna arrays will, in a future integration step, allow for steering the receive antenna beam in discrete steps in the horizontal plane. The principle of operation of the front end module and the design of both antennas are described. Simulated and measured results are presented to validate the design.**

Keywords: wake vortex detection, microstrip patch antenna array, millimetre wave radar, RF-MEMS

I. INTRODUCTION

The fast growth of air traffic observed in the last few years has highlighted the necessary increase of landing and take-off sequences in a majority of the existing airports worldwide. In order to increase the occupation rate without compromising the high security level, a better understanding of the aircraft's environment and more particularly of the risks associated with increased air traffic is necessary. Among the different risky phenomena to be identified are wake vortices which are generated by airplanes while landing or taking-off. These turbulences present a risk for following airplanes and limit the maximum throughput of arriving and departing airplaines. If it were possible to detect the wake vortices from an airplane, the starting and landing time sequence could be reduced and air traffic could be relaxed. The on-board detection of wake vortices, however, represents a significant challenge.

A number of sensors are under development for the on-board detection of wake vortices. Most of these systems are based on optical detection of air turbulences (e.g. LIDAR, Specles) and are limited mainly to the clean air case and are less efficient in rainy or dusty air. These conditions, however, represent the ideal configuration for a microwave sensor because of the presence of particles as scatterers of electromagnetic radiation in the air. Therefore, the development of a dual sensor system (optical and radar) is a promising approach for all-weather operability [1].

The primary specifications required for the on-board detection of wake vortices such as the high angular resolution, while maintaining a small overall system size and the sufficient scattering echo (even for a low density of the hydrometeors), leads obviously to the implementation of a millimetre wave sensor. The attenuation of the high radar frequencies used in those systems allows for a limited range of operation preventing interference with other systems but is sufficient to detect the turbulences present directly in front of the aircraft. To provide a low cost solution to the proposed aeronautical application and due to the similarity of the system specifications and architecture, the wake vortex detection sensor is based on the same principle as the next generation of automotive cruise control radar [2]. The main differences between both applications are the lower detection threshold of the wake vortex radar and the radiation characteristic of the antennas.

In this paper the transmit and receive antennas of a wake vortex detection radar are presented. The model used for the design is described and the simulated and measured far field patterns and return loss of both antennas are shown.

II. SPECIFICATIONS OF THE ANTENNAS

The radar under consideration in this paper is an FMCW radar aiming at the detection of the position and moving direction of small particles, fog, rain or snow, present in the air and carried by the air flow. The reflections from the small scatterers present in front of the aircraft allow for the detection of the rotating movement of the particles and thus for the visualisation of wake vortices.

In order to detect the position of the small particles, a fixed beam transmit antenna and a horizontally steerable receive antenna are designed at 76.5 GHz. The fixed beam antenna presented in this paper constitutes the first step to an RF-MEMS based antenna, electrically steerable in the horizontal plane. The radiation pattern of the receive antenna is a narrow pencil beam, which has to be steered in 21 discrete positions between ± 30°. The main lobe of the transmit antenna must be wide enough to cover the entire area scanned by the receive antenna: ± 30°.

In the vertical plane both far field patterns are broadside and quite narrow to prevent the presence of parasitic reflections. These specifications in the vertical plane are achieved by implementing a 16 serially fed patch antenna array (Fig. 1).

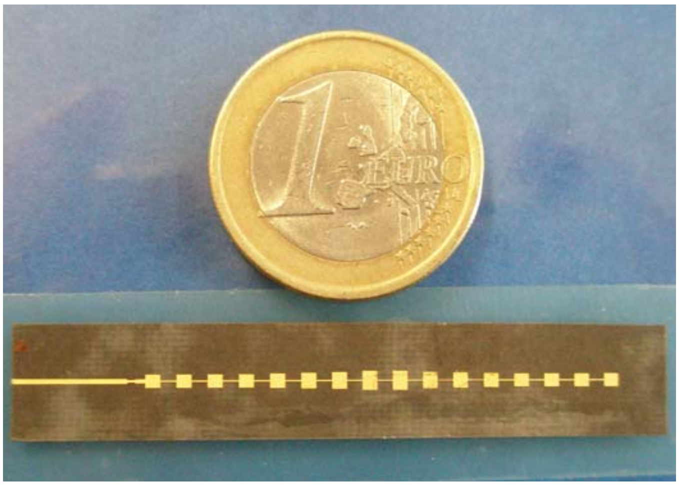

Figure 1. Photo of the fabricated serially fed patch antenna array, 16 patches

III. TRANSMISSION LINE MODEL

The design of the serially fed patch antenna array is achieved by using the Extended Transmission Line Model (ETLM) presented in [3]. This model takes the effect of the increased current density observed in the surroundings of the discontinuity between the microstrip line and the patch into account and is therefore much more accurate than the first model proposed in [4].

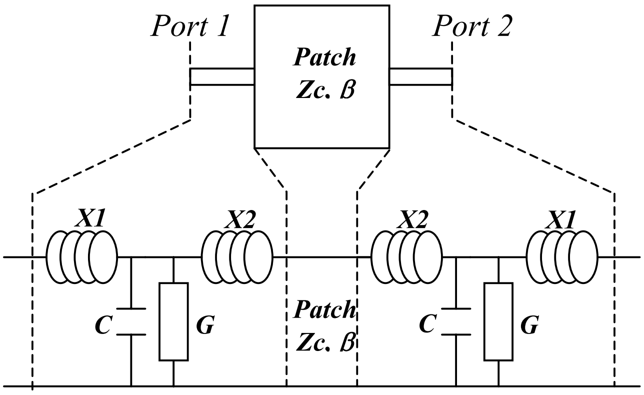

Figure 2. Extended Transmission Line Model

According to the ETLM model a single radiating slot is modelled by 4 lumped elements:

- ***X1***, ***X2***: Inductances modeling the increased current density due to the discontinuity in width between the feeding microstrip line and the patch.
- ***C***: Capacitance modeling the additional electrical length of the patch due to the fringing fields.
- ***G***: Conductance modeling the power radiated by the slot.

The patch itself is modelled by a simple transmission line having a low characteristic impedance ***Zc*** and a given propagation constant β. Both are determined by the patch dimensions. The value of the lumped elements are found either by using the usual theoretical formula given in [3] and [5] or from the simulated S-parameters S_{11} and S_{21} of the 2 ports discontinuity.

The realization of a broadside pattern is achieved by choosing the appropriate length for the microstrip line joining two patches so that the electrical length between them is λg/2 at the design frequency. Such an antenna array is said to be resonant because of the 180° electrical length observed between two side by side radiating slots. An equivalent circuit of such an array is given in Fig. 3.

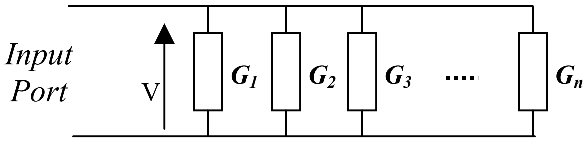

Figure 3. Equivalent circuit of a resonant array having n patches

In Fig. 3, the conductances G_i modeling the different slots in the array are excited by the same voltage V - same amplitude and phase - as the voltage applied at the input port.

In order to maximize the antenna gain and lower the side lobe level in the vertical plane (E-plane), an aperture distribution is implemented in the same way as presented in [6]. The different patches in the antenna array being fed by an identical voltage V, the power radiated by a slot is directly proportional to its radiating conductance G_i. It is given by

$$P_i = \frac{1}{2}.\mathrm{Re}\left\{V^2 * G_i\right\} \quad (1)$$

The radiating conductance G_i of the slot is determined by the width of the patch. It is given by

$$\frac{1}{G} = \frac{\varepsilon_{r,eff} * Z_0^{\,2}}{120 * \left[(k_0 h)^2 \left[0.53 - 0.03795 * \left(\frac{k_0 W}{2} \right)^2 - \left(\frac{0.03553}{\varepsilon_{r,eff}} \right) \right] \right]} \quad (2)$$

Z_0 is the characteristic impedance of the patch, W is the physical width of the patch, k_0 is the wave number in air, h is the substrate thickness, and $\varepsilon_{r,eff}$ is the effective dielectric constant.

Solving (1) and (2) for a given aimed aperture distribution allows for the determination of the widths of the different patches in the array. The length of the different microstrip lines is then determined by implementing the previously presented ETLM model.

IV. SIMULATION AND MEASUREMENT OF THE 1D ARRAY

The 16-patch 1D antenna array shown in Fig. 1 was simulated with the commercial 3D FDTD software EMPIRE. It was realized on 5 mil thick Rogers R/T Duroid 5880 substrate and was measured in an anechoic chamber. The simulated and measured vertical far field patterns are shown in Fig. 4. The side lobe level is below -14dB, the -3dB beam width is 5° as required and the measured gain is 15dB. The radiation pattern is broadside at 76.5GHz. Above and below this design frequency a small off-axis angle is observed for the main lobe direction. This angle is about 1° per GHz and is due to the intrinsic design principle of the antenna.

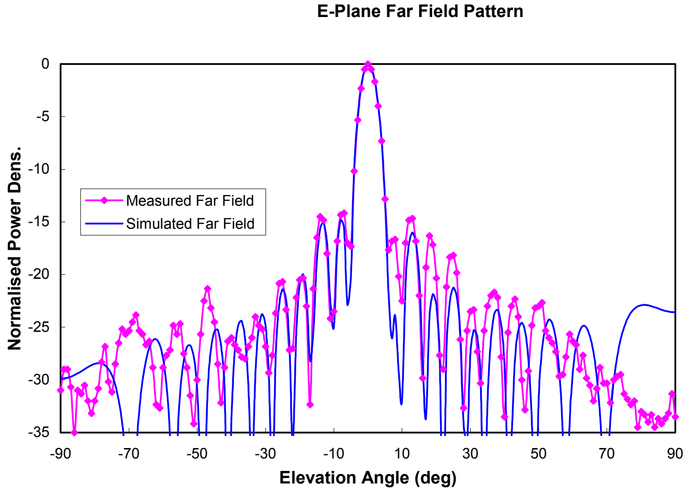

Figure 4. Normalised simulated and measured far field patterns - 1D array, vertical plane -

The measured return loss of the 1D antenna array is shown in Fig. 5. The input matching is better than -10dB from below 75 GHz up to 77GHz. The measured frequency range was limited by the measurement equipment.

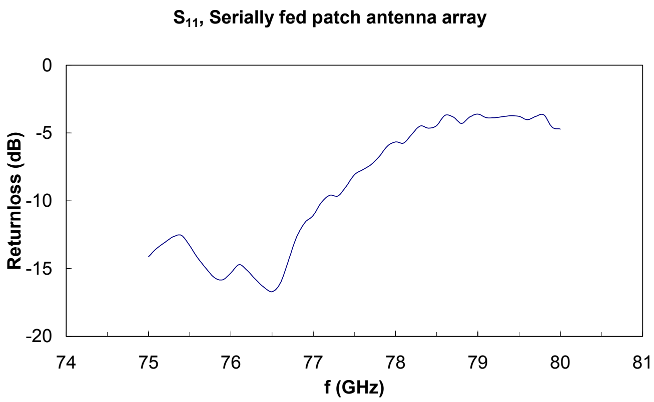

Figure 5. Measured return loss of the serially fed patch antenna array

V. SIMULATION AND MEASUREMENT OF THE 2D ARRAY

The previously presented serially fed 1D-array is the starting point for realizing the complete transmit and receive antennas. The receive antenna is built with 28 such 1D-antenna arrays fed in parallel by a corporate feed network. The transmit antenna is realized with 3 1D-arrays. Both antennas are shown in Fig. 6. The location of the RF-MEMS phase shifters, which will be implemented in the next integration step, allowing for a steerable beam of the receive antenna in the horizontal plane is also indicated in Fig. 6.

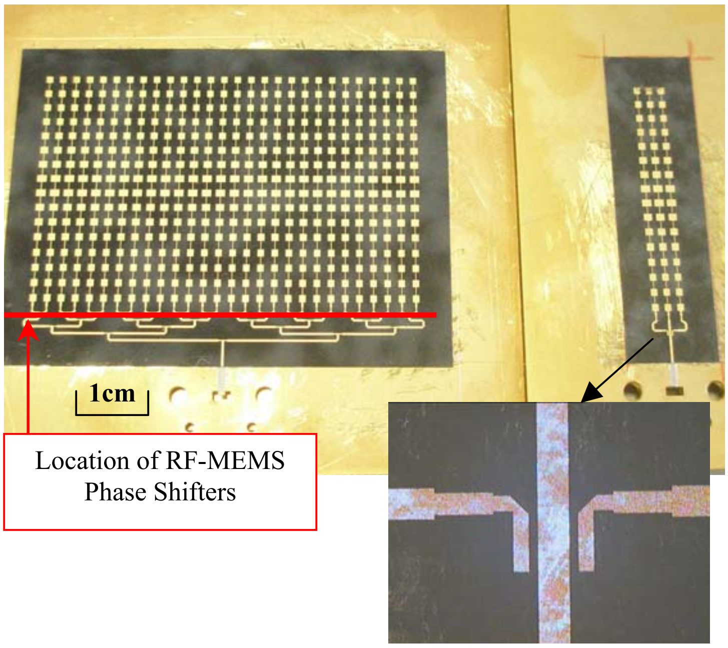

Figure 6. Receive (left) and transmit (right) antenna arrays

The excitation coefficients used to feed the 2D arrays are determined by a so-called Remez-Type algorithm. This algorithm computes the best approximation in the Tchebycheff sense of the ideal radiation characteristic [7]. This aperture distribution ensures the realization of the minimum side lobe level while achieving a certain prescribed beam width. The corporate feed network is fabricated with asymmetrical T-junction power dividers dimensioned to achieve the Tchebycheff aperture distribution.

The feed network of the transmit antenna is realized with a coupled transmission line directional coupler. This coupling structure allows for a weak feeding of the two 1D side arrays necessary to achieve the required wide 60°-beamwidth for the transmit antenna.

The simulated far field pattern was computed without feed network by using a multi-port excitation to assess the Tchebycheff aperture distribution especially in terms of side lobe level. The radiation pattern measurement was realized with an absorber covering the corporate feed network to prevent an increased side lobe level due to its radiation. The corporate feed network of the next generation of antennas will be realized on LTCC substrate, having a higher dielectric constant than the Rogers substrate currently used in order to minimize the radiation of the feed network. The simulated and measured H-plane far field patterns of the receive antenna are shown in Fig. 7. The measured E-plane pattern is identical to the results of the 1D array shown in Fig. 4.

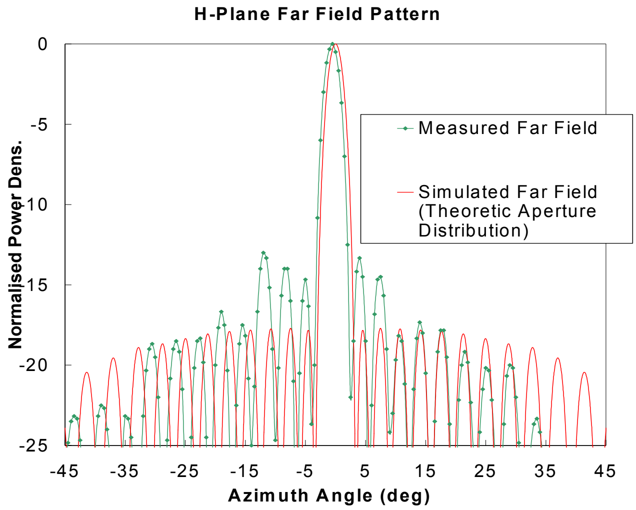

Figure 7. Normalised simulated and measured far field patterns - receive antenna, horizontal plane -

The differences observed between the simulated far field pattern and the measured one can be explained by an inaccuracy in the excitation coefficients achieved by the corporate feed network. The finite minimal width of the microstrip lines used in the T-junction power dividers make the exact achievement of the ideal Tchebycheff aperture distribution impossible and the 28 1D arrays are fed by an approximation of that distribution. The measured antenna gain of the receive antenna is 25.5dB. The attenuation of the signal in the corporate feed network and losses in the patch array are estimated to 5.5dB in order to explain the difference to the simulated gain of 31dB. The -3dB beam width is 3°.

The measured return loss of the receive antenna is better than –10dB from below 75 GHz up to 77.3 GHz (Fig. 8).

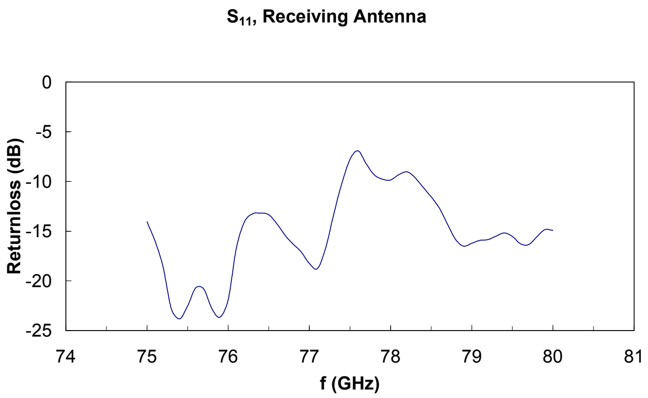

Figure 8. Measured return loss of the receive antenna

The simulated and measured far field patterns of the transmit antenna match quite well and are shown in Fig. 9.

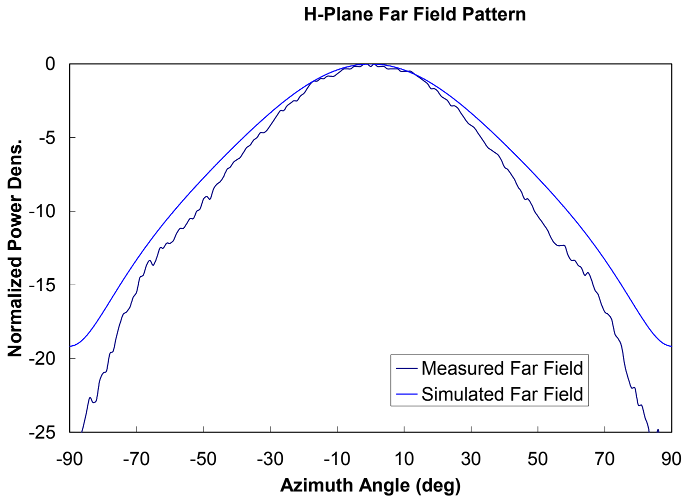

Figure 9. Normalised simulated and measured far field patterns - transmit antenna, horizontal plane -

The slightly reduced beamwidth as compared to the simulation is believed to be due to a too strong coupling in the directional coupler used to realize the feed network of the transmit antenna.

VI. CONCLUSION

The designs of two fixed beam antenna arrays were demonstrated and the model used for the antenna design was validated by simulated and measured results. The receive antenna realized with 28 1D serially fed patch antenna arrays assumes a measured gain of 25.5dB. The side lobe level of this antenna is below -14dB and the -3dB-beam width of the measured pencil beam is 5° in the E-plane and 3° in the H-plane. The transmit antenna has the same radiation characteristic in the E-plane but assumes a main beam about 60° wide in the H-plane, which allows for covering the desired detection area. This study constitutes the first step of the development of an RF-MEMS steerable antenna, which will be designed for a wake vortex detection front end module based on automotive radar technology.

ACKNOWLEDGMENTS

The work is supported by the German Bundesministerium für Bildung und Forschung (BMBF) under the contract “RadarAuge” (#16SV2078). The authors would like to thank Mr. R. Körber of Astyx GmbH for performing the antenna measurements and Ms. S. Hoyler of Reinhardt Microtech for supplying the mask sets for the antenna fabrication. Many thanks to Christian M. Siegel for the helpful discussions on feed networks.

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