Circularly Polarized Antenna With Steerable Dipole-Like Radiation Pattern

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Circularly polarized antenna with steerable dipole-like radiation pattern

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Abstract—An omnidirectional circularly-polarized antenna with a rotatable dipole-shaped radiation pattern is proposed. The antenna is realized using a back-to-back coupled microstrip patch arrangement. The pattern is rotated by means of a phase shift, enabling reception (or transmission) of signals from all angles around a sphere. The method enables continuous pattern steering without the need for semiconductor or MEMS components integrated into the antenna. It also allows the use of more than one radiation pattern simultaneously, potentially providing coverage for any spherical angle.

The theory is supported by simulation and measurement of four typical antenna configurations. The maximum gains are between 1.9 and 4 dBi and average axial-ratio varies between 2.5 and 3.65 dB at 2.47 GHz. The impedance bandwidth is from 2.4 to 2.51 GHz and the reconfigurable circular polarization is achieved from 2.464 to 2.484 GHz.

Index Terms—Adaptive antennas, Circular polarization, Microstrip antennas, Omnidirectional antennas

I. INTRODUCTION

OMNIDIRECTIONAL circularly-polarized antennas (OCPA) provide benefits in several applications including telemetry, tracking & command systems, satellite navigation, WLAN and RFID. Although basic linearly-polarized radiators (dipoles or monopoles) are omnidirectional, it was not until the 1950's that circularly-polarized (CP) antennas reached comparable omnidirectional capability [1]. Since then, OCPA's became more compact and easier to manufacture. They can be generally summarized into four categories, based on the principles of operation:

- a linearly-polarized antenna surrounded by a polarizer [1]-[3]. This approach provides good omnidirectionality. As a minimum spacing is usually required between radiator and polarizer, the manufacture is complex and the solution is more practical for higher frequencies.
- an array of CP radiators wrapped around a common center [4-7]. This is conceptually the simplest solution, however the antenna performance depends on the number of radiators. It also requires a feed network to distribute the signal to the array elements.
- a centrally located vertically-polarized radiator surrounded by a set of horizontally-polarized radiators [8]-[9]. The horizontal elements are usually realized as quarter wavelength stubs. The vertically-polarized element may vary depending on the design. Recently the use of zeroth order resonance combined with a reduced groundplane size, enabled these OCPAs to be made planar.
- two back-to-back coupled patches [10]-[12]. This solution is entirely planar, although requires multilayered PCB technology. Some decrease in performance may be observed in the plane of the substrate, however both the axial-ratio (AR) and CP gain can be kept within reasonable limits.

These approaches exhibit dipole-like radiation patterns, which are omnidirectional in one plane and have a figure-of-eight shape in the orthogonal principal cuts. There are two nulls in the pattern, which usually correspond to minimum radiated power and worst axial ratio.

The majority of CP reconfigurable antennas in the literature deal with polarization reconfigurability [13]-[14], with few investigations on switched radiation patterns [15]. With regard to OCPA's, this is the first paper providing the methodology to change the plane of the omnidirectionality. In this work a reconfigurable OCPA capable of rotating the plane of omnidirectionality over 180° is proposed. The concept derives from the back-to-back coupled class of OCPA's and the reconfiguration is achieved using two ports, fed with different phases. The use of phase shifters allows a continuous transition between various configurations. No semiconductor or MEMS switching components are required, improving reliability and intermodulation performance, while reducing manufacturing costs.

The proposed steerable omnidirectional CP antenna is able to receive or transmit signals from any arbitrary angle, subject to proper steering. It is well accepted, that an isotropic antenna
is not realizable [16]. However in the proposed solution, two or more radiation patterns can be used simultaneously (for example through rat-race-coupler or Rotman lens) in a MIMO configuration, providing full spherical coverage.

This paper is organized as follows: section II describes the proposed antenna; section III describes the general principles of operation; section IV reports the simulated and measured results for various configurations and conclusions are made in section V.

II. DESIGN

The antenna consists of two layers of substrate (Taconic™ RF 35, $\varepsilon_r = 3.47$ and height $h = 3$ mm each) placed between three layers of metallization (as seen in Fig. 1). The middle metallization layer (layer B) forms a ground plane, whereas layers A and C form back-to-back coupled circular patches and a feed network. Two SMA connectors (feed A and feed C) are located on opposite sides of the antenna: SMA grounds are connected to the groundplane (on opposite side) and the feed probes are connected through a groundplane hole to short sections of $50 \, \Omega$ microstrip line. Part of the substrate ($L_{\text{hole}} \times W_{\text{hole}}$) is cut away to accommodate the connector flanges.

The 6.8 mm wide $50 \, \Omega$ microstrip lines on layers A and C connect to quarter wavelength transformers ($68 \, \Omega$ impedance, 4 mm wide, $L_1 = 18.5$ mm) and then to 94 $\Omega$ microstrip sections (2 mm wide). A microstrip power splitter connects to two 186 $\Omega$ lines (0.2 mm wide). On each side, one line connects directly to the patch ($A_1$ and $C_1$), while the other connects through a 90° phase shifter ($A_2$ and $C_4$).

It should be noted that the groundplane is smaller (by $S_{\text{ground}}$ in $\pm x$ direction and $H_{\text{ground}}$ in $y$ direction) than the outer perimeter of the substrate because omnidirectional CP performance requires a reduced groundplane [11].

III. PRINCIPLES OF OPERATION

The omnidirectional CP performance is achieved by using a back-to-back coupled arrangement of dual-fed circular microstrip patches. To achieve omnidirectional performance, the ground plane size is significantly reduced, allowing the patches to couple and thus preventing degradation of the CP gain and AR in the plane of the substrate. The proposed antenna offers a wider AR bandwidth and increased CP gain compared to the CPW-fed antenna in [11], however the physical principles for fixed pattern omnidirectional CP are the same and hence won't be discussed further here.

The following innovative mechanism allows reconfigurability of the omnidirectional CP plane. To explain the principles, the phase difference at the SMA input connectors (feed A and feed C in Fig 1) is investigated. The antenna exhibits omnidirectional CP performance when these ports are fed with equal amplitude. However the plane of omnidirectionality is linearly proportional to the phase difference $\Delta_{\text{ph}}$ between feed A and feed C; this plane can be rotated around the $\pm z$ axis by varying $\Delta_{\text{ph}}$. This relation can be approximated by:

$$\gamma = \frac{1}{2} \Delta_{\text{ph}}$$

(1)

where $\gamma$ is the angle between the plane of omnidirectionality and the $xz$-plane. To demonstrate this, the patch phase relationships are investigated (described in Table I). The study is made for the cases of $\Delta_{\text{ph}} = 0^\circ$ and $\Delta_{\text{ph}} = 180^\circ$.

The plane of omnidirectionality rotates around the $\pm z$ axis, so there will always be strong CP radiation in the directions
normal to the patches (±z directions). For other angles, it will depend on $\Delta_{ph}$. Table I describes the phase at quadrature points ($A_n \/ C_n$) for both patches: for $\Delta_{ph} = 0^\circ$ the phase at patch feed points $A_1$ and $C_1$ (shown in Fig. 1) is equal, for reference, say $0^\circ$. Due to the dual-feed phase shift, the phase at points $A_2$ and $C_4$ is $90^\circ$ delayed. However, for $\Delta_{ph} = 180^\circ$, $A_1$ is $180^\circ$ out of phase with $C_1$ and $A_2$ is $90^\circ$ delayed with respect to $A_1$. A similar relationship holds for $C_1$ and $C_4$. This is illustrated in Table I.

Since the patches are oriented back-to-back, for the case of $0^\circ$ phase difference between points $A_n \/ C_n$ (i.e. $\pm x$ for $\Delta_{ph} = 0^\circ$ and $\pm x$ for $\Delta_{ph} = 180^\circ$) the electric field is oriented in opposing directions (Fig. 2a) and hence produces a null in the radiation pattern. However, for $180^\circ$ phase difference between points $A_n \/ C_n$, the electric fields are aligned (Fig. 2b) and strong radiation is produced. Thus, it is seen from Table I, that for $\Delta_{ph} = 0^\circ$, points $A_2 / C_2$ and $A_4 / C_4$, are out-of-phase hence providing omnidirectional radiation in the azimuth plane ($xz$-plane with $\gamma = 0^\circ$); For $\Delta_{ph} = 180^\circ$ points $A_2 / C_2$ and $A_4 / C_4$ are out of phase, hence omnidirectional radiation in elevation plane ($yz$-plane with $\gamma = 90^\circ$).

Although only two configurations are calculated in Table I, the proposed methodology can be extended to any other phase shift $\Delta_{ph}$ in the range of $0^\circ$ to $360^\circ$. This offers smooth and continuous transition between different radiation patterns.

The described theory refers to an ideal case, with the antenna surrounded by free space and fed by a perfect feed. However in practice, the antenna is cable fed and mounted on a vehicle or device, which is usually bigger than the antenna itself. This may perturb the relationship given in (1) for most configurations of $\Delta_{ph}$ because of reflections from these structures. A consequential reduction in gain in the direction in which the reflecting object is located may be observed and a degraded AR in the opposite direction, because reflected RHCP signals become counter-polarized. Since the proposed antenna can radiate in all spherical angles the measurement set-up (cables, SMA connectors etc.) will impact upon the measured radiation pattern. Nevertheless, it is considered that these effects correspond to real life scenarios, as the commercial version of the proposed antenna will most likely be mounted in a similar configuration.

### IV. SIMULATION AND MEASUREMENT

The antenna was designed for the WLAN band 2.4 - 2.51 GHz and the measured and simulated S11 $\leq -10$ dB over the whole band as seen in Fig. 3. The small discrepancy is due to manufacturing tolerance: the fabrication technique was constrained by a resolution of $0.2$ mm, which corresponds to the width of the $186 \, \Omega$ microstrip lines. An additional simulation was performed (results not shown for brevity), in which the line width was varied by $\pm 0.1$ mm. The resultant S11 varied by up to $2.5$ dB, which is similar to the discrepancy visible in Fig. 3. All simulations were made using the time domain solver of CST Microwave Studio.

In order to validate the theory described in section III both SMA feeds A and C were connected by phase-matched cables to a rat race coupler providing $\Delta_{ph} = 0^\circ$ and $\Delta_{ph} = 180^\circ$. A pair of ferrites were positioned on the cables, to minimize current flow on the outside of the cable (a problem expected due to the small groundplane). As the antenna radiates in practically all directions (depending on the configuration), the rat race coupler was separated from the antenna by a distance.
of 30 cm using a specially designed PVC holder (which also increased the positioning accuracy) and covered by a block of absorbing material. Fig. 4 shows a photo of the measurement arrangement as seen in the anechoic chamber [17].

A. Configuration with $\Delta_{\phi} = 0^\circ$

The configuration was achieved by feeding antenna through a rat-race coupler. The in-phase port was connected to measurement equipment and the out-of-phase port terminated with a match. In the antenna impedance bandwidth (2.4 - 2.51 GHz) the difference in amplitude between two output ports was below 0.1 dB and phase varied between 0.5° and 6°.

Figs. 5-8 show the realized gains and AR for the $\Delta_{\phi} = 0^\circ$ configuration. The optimum frequency (in terms of AR in the omnidirectional plane) is 2.47 GHz. Fig. 5 shows the omnidirectional radiation pattern in the azimuth plane ($xz$-plane), in accordance with (1). Fig. 6 shows a figure of eight shape in the elevation plane ($yz$-plane). The shouldering visible in the pattern are the effects of the feed network. This effect is visible mainly in the $yz$-plane, in which the support structure is located. The average measured AR in the $xz$-plane is 3 dB and for all angles it is better than 5.68 dB, which corresponds to -10 dB cross-polarization [18]. Fig. 7 shows the AR measured around a full sphere, as a function of $\theta$ and $\phi$ (in accordance with coordinate system on Fig. 1). It can be seen, that this is a pattern of a CP dipole, with an omnidirectional azimuth plane at $\theta = 90^\circ$ and two nulls at $\theta = 0^\circ$ and $\theta = 180^\circ$.

Fig 8 shows the measured AR for the azimuth plane ($\theta = 90^\circ$) as a function of $\phi$ and frequency. The AR remains below 5.68 dB for all angles in the $\theta = 90^\circ$ plane, which is maintained across a bandwidth of 2.464 - 2.494 GHz, (30 MHz, 1.2 %). The measured RHCP gain varies from +2 to -2 dBic, which is comparable to the gain of a dipole.
B. Configuration with $\Delta_{\text{ph}} = 180^\circ$

The configuration was realized using a rat-race coupler. The out-of-phase port was connected to the measurement equipment and the in-phase port terminated. The difference in amplitude between two output ports was less than 0.6 dB and phase varied between 185° and 188°.

Figs. 9-12 demonstrate the radiation patterns and AR for the $\Delta_{\text{ph}} = 180^\circ$ configuration. Fig. 9 shows that the measured RHCP realized gain (at 2.47 GHz) agrees well with simulation, with the exception of angles $120^\circ \leq \theta \leq 240^\circ$, which are obstructed by the rat-race-coupler and covering absorber. The figure of eight shape, expected in the azimuth plane ($xz$-plane), is slightly tilted in both simulation and measurement, shown in Fig 10.

Fig. 11 illustrates the measured AR around a full sphere. The CP dipole pattern is also visible, however the pattern is omnidirectional in the elevation plane ($\phi = 90^\circ$ and $\phi = 270^\circ$). Two nulls are visible at $\phi = 0^\circ$; $\theta = 90^\circ$ and $\phi = 180^\circ$; $\theta = 90^\circ$. An additional null around $\theta = 170^\circ$ is due to the absorber and supporting structure.

Fig. 12 shows the measured AR for the elevation plane as a function of $\theta$ and frequency. Three regions can be clearly distinguished for both AR and gain performance (as seen in Fig. 9-12):

- For angles $150^\circ \leq \theta \leq 235^\circ$, the AR is strongly varying from 0.4 to 11 dB, with three noticeable peaks exceeding the 5.68 dB limit. A drop in RHCP gain is also observed for this region in Fig. 9. These effects are expected because the feed circuit/absorber/support structure are in this direction.
- For angles $10^\circ \leq \theta \leq 150^\circ$ and $235^\circ \leq \theta \leq 350^\circ$ the view is unobstructed and a good CP performance is visible, with an average AR of 2.5 dB. For this region, the bandwidth for which the AR < 5.68 dB is from 2.434 to 2.484 GHz, (50 MHz, 2%). This is more wideband than the $\Delta_{\text{ph}} = 0^\circ$ case and exhibits a slight downward frequency shift. This is due to the additional dielectric in $yz$-plane [19], required for the quarter wavelength transformer. The overlapping band of both configurations is 2.466 - 2.484 GHz (0.8 %). The maximum gain is 2.6 dBi.
- For angles $350^\circ \leq \theta \leq 10^\circ$ there is no visible drop in RHCP gain, however the AR deteriorates up to 8.5 dB. This is because of support structure reflections (generating counter-polarization and hence degrading cross-polarization). Although the reflected LHCP signal is small, it is sufficient to strongly degrade the AR [18]. This effect was validated by an additional simulation, in which a brick of ECCOSORB LS-26 absorber was placed in the same location as in measurement configuration. The results showed strongly decreased AR for angles $350^\circ \leq \theta \leq 10^\circ$ and the RCHP gain was decreased down to -4 dBi for $\theta = 0^\circ$ and to -9 dBi for angles $\theta = 160^\circ$ and $\theta = 200^\circ$. Also some ripples in AR up to 6 dB occur across the full omnidirectional plane. These effects are similar in measurement, however do not occur in simulation without the absorber.

Fig. 8. Measured axial ratio (dB) for $\Delta_{\text{ph}} = 0^\circ$ as a function of angles $\phi$ and frequency.

Fig. 9. Measured and simulated realized gains for $\Delta_{\text{ph}} = 180^\circ$ in the $yz$-plane (for $\phi = 90^\circ$ and varying $\theta$) at 2.47 GHz measured and 2.46 GHz simulated.

Fig. 10. Measured and simulated realized gains for $\Delta_{\text{ph}} = 180^\circ$ in the $xz$-plane (for $\phi = 90^\circ$ and varying $\theta$) at 2.47 GHz measured and 2.46 GHz simulated.
C. Configuration with $\Delta \phi = 90^\circ$

For this configuration, the rat race coupler was replaced by a classical hybrid coupler in order to generate $\Delta \phi = 90^\circ$. One port was connected to the measurement equipment and the other terminated. The difference in amplitude between two output ports was $\leq 1.7$ dB and phase varied from $88^\circ$ to $90^\circ$.

According to (1) the plane of omnidirectionality will be tilted by $45^\circ$ with respect to the $xz$-plane ($\gamma = 45^\circ$). This plane is referred to as $\gamma_{45}$ for brevity and is defined in spherical coordinates as:

$$\theta = \arccotg(\cos \phi)$$  \hspace{1cm} (2)

The data presented in Figs. 13 and 15 are for this plane.

An orthogonal plane tilted $-45^\circ$ with respect to the $xz$-plane is denoted as $\gamma_{-45}$ and defined as:

$$\theta = \arccotg(-\cos \phi)$$  \hspace{1cm} (3)

The RHCP gain in Fig. 13 shows good performance (up to 4 dBi) from $220^\circ \leq \phi \leq 130^\circ$. A drop in gain is observed for $130^\circ \leq \phi \leq 220^\circ$, which is the direction in which absorber is located. In the $\gamma_{-45}$ plane (Fig. 14), the figure-of-eight shape is less prominent in measurement than simulation. Notably, the null expected for $\phi = 0^\circ$ is not visible in measurement, which is due to support structure reflections. As seen in Fig. 15, there is also an increase in AR for $\phi = 250^\circ$ and $\phi = 350^\circ$ due to the antenna microstrip feedline located on this side in the plane of omnidirectionality. This effect is also observed for simulation (with AR up to 8.3 dB). The average AR over the full $\gamma_{45}$ plane is 3.19 dB.
D. Configuration with $\Delta \phi = -90^\circ$

For the $\Delta \phi = -90^\circ$ configuration, an omnidirectional pattern should occur in the $\gamma-45$ plane (Fig. 17), as defined by (3) and a figure-of-eight shape in $\gamma_{45}$ plane (Fig. 18), defined by (2). For this configuration, the absorber and supporting structure is located close to $\phi = 0^\circ$ in the $\gamma-45$ plane. As seen in Fig. 19, the AR performance is below 4 dB in the region $160^\circ \leq \phi \leq 310^\circ$ (layer C side) and it decreases around $\phi = 125^\circ$. This is flipped with respect to the $\Delta \phi = 90^\circ$ map in Fig. 15. Also the measured null in radiation pattern on Fig. 18 is less pronounced here for $\phi = 180^\circ$, which for this cut is the direction of the absorber. The maximum RHCP realized gain is 3.3 dBic and average AR in $\gamma_{45}$ plane is 3.65 dB.

V. CONCLUSION

The proposed antenna allows reconfiguration of the omnidirectional circularly polarized radiation pattern by control of phase between feed ports. This technique allows continuous pattern steering, with a theoretical infinite number of configurations. No semiconductor components or MEMS switches are directly integrated into the antenna, which is desirable for good intermodulation performance, power...
handling and manufacturability. The phase shifting can be implemented by any means convenient to the antenna manufacturer. Additionally, more than one pattern can be used simultaneously, hence the antenna can transmit/receive into any spherical angle. By augmenting this property with diversity methods, a full CP spherical coverage is possible with a single antenna.

The underlying principles of operation are described and validated by both simulation and measurement. For the basic configurations presented, the gains and AR's are within acceptable limits for such antennas, although the impact of the supporting structure is visible throughout measurements.

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technological strategy reporting and assessment of compliance with international standards on human exposure to electromagnetic energy. The industrial contacts also stem from several successful transfers of fundamental design research into applied solutions.

Dr. Ammann received a best paper award at the 2006 Loughborough Antennas and Propagation Conference, the 2009 SFI best paper award in the China Ireland International Conference in ICT and commercialization awards for work with DecaWave Ltd, Taoglas Ltd and Sequoia Smart Solutions. He also received a 2008 CST University Publication Award for work on a “Wideband Reconfigurable Rolled Planar Monopole Antenna” and a 2011 CST Award for work on “Miniature Ceramic Dual-PIFA Antenna to Support Band Group 1 UWB Functionality in Mobile Handset”. He sits on the management committee of the EU COST Action IC1102, “Versatile, Integrated, and Signal-aware Technologies for Antennas (VISTA)” and is active in the EurAAP working group on small antennas. As a member of the IEEE International Committee for Electromagnetic Safety, he participated in the revision of the IEEE Std. C95.1, 2005 & 2012 standards for Safety Levels with Respect to Human Exposure to Radio Frequency Electromagnetic Fields, 3 kHz to 300 GHz. He is also a member of the URSI Committee for Communications and Radio Science within the Royal Irish Academy and official member of URSI Commission K: Electromagnetics in Biology and Medicine. He has chaired and organized special sessions on small antennas, UWB antennas and UWB Wireless Communication Systems at EuCAP and IEEE APS and chaired the Antennas and Propagation Track for the 65th IEEE VTC, Dublin 2007. He was the local chair for the October 2008 EU COST IC0603 workshop and meeting in Dublin. He is currently associate editor for the IEEE Antennas & Wireless Propagation Letters.

Hamam Shakhtour was born in Bethlehem, Palestine, in 1982. He received the B.Sc. degree in electrical engineering from Birzeit University, Birzeit, Palestine, in 2005 and the M. Sc. degree in electrical engineering and information technologies from Karlsruhe Institute of Technology (KIT), Karlsruhe, Germany, in 2009. He is currently at the Institute of High Frequency Technology (IHF), Aachen, Germany, working on near-field measurement techniques for active antenna characterization as part of his Ph.D. degree. Additionally of his current interests are antenna design and materials characterization.

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Dirk Heberling is a member of VDE and IEEE and since 1998 has been a member of the ITG expert group 7.1 "Antennen" which he directed as a chairman from 2002 - 2003. During this time he was responsible as General Chairman and organizer of the international antenna conference INICA 2003, September 2003, Berlin. Since 1998 he has been a member of the European competence projects for Antennas COST 260, COST 284, IC0603 and IC1102, from 2002 - 2007 he was the German delegate of COST 284 and now he is the German delegate of IC1102. From 2002 - 2003 he was co-organizer of the European network of excellence on Antennas ACE. He is member of the steering committee and organizing committee for the European Conference on Antennas and Propagation, EuCAP.