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Dual-polarized antenna based on characteristic modes of conductive plate

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Abstract—In this paper, a dual-polarized antenna for small terminals is investigated. The antenna consists of a feed network that provides excitation to the terminal's metal chassis or the Printed Circuit Board (PCB) ground plane. Feed points are selected based on the Theory of Characteristic Modes. The proposed antenna is a 4-layer PCB with dimensions 80 x 80 mm. It operates within the bandwidth of 1820 – 1880 MHz with polar discrimination of 30dB, directivity of 4.4 dBi, and envelope correlation coefficient not exceeding 0.001.

Keywords—characteristic mode analysis, dual-polarized antenna, small terminal antenna

I. INTRODUCTION

Dual-polarized antennas are important elements in broadband transmission systems and radiolocation techniques. In wide bandwidth radio links, they are often used for polarization diversity or MIMO (multiple-input multiple-output) to increase the data throughput of the system or its range. Using antennas with strong polarization discrimination allows for low correlation between two data streams, which increases the efficiency of high-speed wireless transmission. An additional consideration is the available space for installing a particular antenna.

The research on characteristic mode antennas resulted in a significant number of works focusing on antennas for small terminals. Reviewing recent research, many implementations of dual-polarized or MIMO antennas have been found. In [1], a complete prototype of antennas for LTE Band 13 is presented, exploiting the design of the ground plane located in the horizontal plane. It uses additional radiating elements located in the vertical plane. Very high polarization purity of more than -30dB was achieved by using differential excitations and radiating elements located on five stacked laminates [2], resulting in a complex manufacturing process. It is also possible to use the distribution of characteristic currents to determine the shape of the board slots to obtain dual-circular polarized radiation, as shown in [3]. In the aforementioned work, this was obtained using two separate plates. The reference plane as a second antenna was also used in a small multi-antenna terminal reported in [4]. The terminal was equipped with one typical monopole-type antenna fed with a capacitive coupling element oriented vertically on a shorter edge, and a second antenna with inductive coupling elements on a longer edge. A similar method of capacitive excitation can be used by using a bezel with a suitable gap from the board, as shown in [5]. Obtaining a dual-polarized antenna with a single radiating element is presented in [6]. The antenna does not assume the use of a ground plane or metal enclosure for radiation, instead relying on a triangular patch

antenna design. The design from which it was decided to start in the creation of this article [7] is a plate with one slot located centrally.

This paper presents a dual-polarized antenna based on a PCB board modelled as a small terminal enclosure. The main design considerations are to leave full metallization on one side of the double-layer board with all the feeding and the matching network on the other. The design incorporates a matching network based on Surface-Mount Devices (SMD) components to tune to an expected frequency using a gamma matching circuit.

II. CHARACTERISTIC MODE ANALYSIS

In this paper, the characteristic mode analysis was performed for a solid metal plate of 80 x 80 mm, made out of Perfect Electric Conductor (PEC). The plate was analyzed using a CST Studio Suite characteristic modes solver, with six modes above the frequency of 1800 MHz. The Modal Significance (MS) parameter was obtained as:

$$MS = \left| \frac{1}{1 + j\lambda_n} \right| \quad (1)$$

Where λ_n is the eigenvalues and n corresponds to the index of the order of each mode. Referring to the theory of the characteristic modes [8], MS together with the modal excitation factor estimates the contribution of each mode to the total electromagnetic response to a given source. Fig. 1 shows the MS values as a function of frequency. As the intended bandwidth of operation is within the 1800 - 2200 MHz, modes 1 and 2 appear as dominant, i.e. their MS within this frequency range is close to unity. Furthermore, those modes exhibit a symmetrical performance, which is convenient for a dual-polarized antenna when both modes are used for orthogonal polarizations.

Fig. 2 presents the current distribution of characteristic modes ($J_1 - J_6$) for the frequency 1800 MHz. The corresponding patterns with directivity at 1800 MHz are shown in Fig. 3. Comparing the radiation patterns for different modes, it is noted that modes 1 and 2 exhibit a rotation-symmetrical performance, having a different dominant component of the electric field. This is further visualized in Fig. 4, which shows two main cross sections of radiation patterns for modes 1 and 2 in the ZX plane (constant $\varphi=0^\circ$) and the XY plane (constant $\theta=90^\circ$), and for two electric field components (E_{θ} or E_{ϕ}). It can be seen that the radiation patterns of the individual modes maintain rotational symmetry to each other. Please note that for the cross-section of the ZY

plane characteristics, the value for cross-polar reaches a null of the radiation pattern.

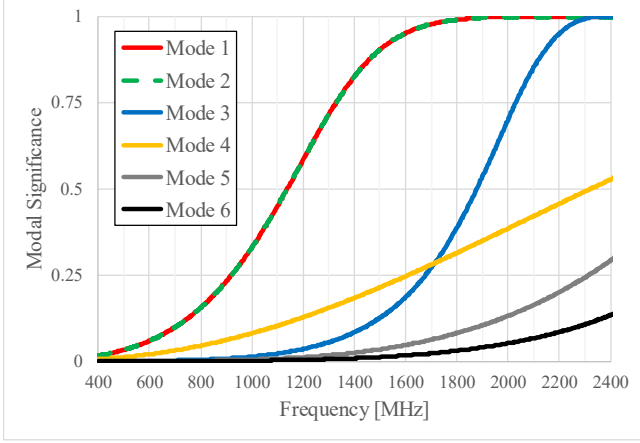


Fig. 1. Modal significance of 80x80mm PEC plate

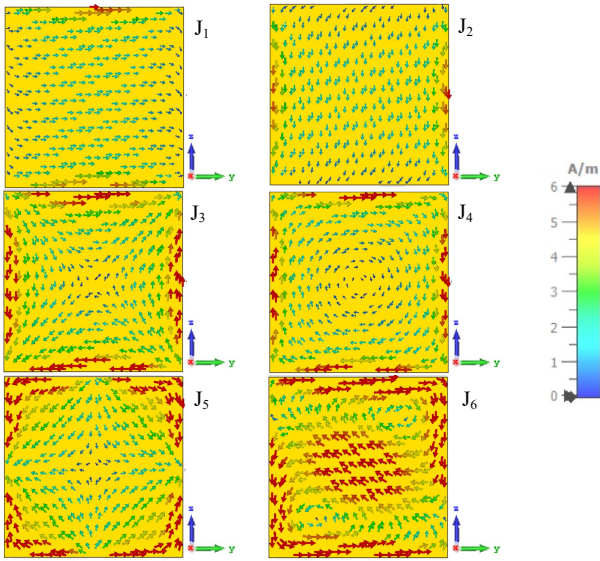


Fig. 2. Characteristic currents for 1800 MHz for modes 1-6

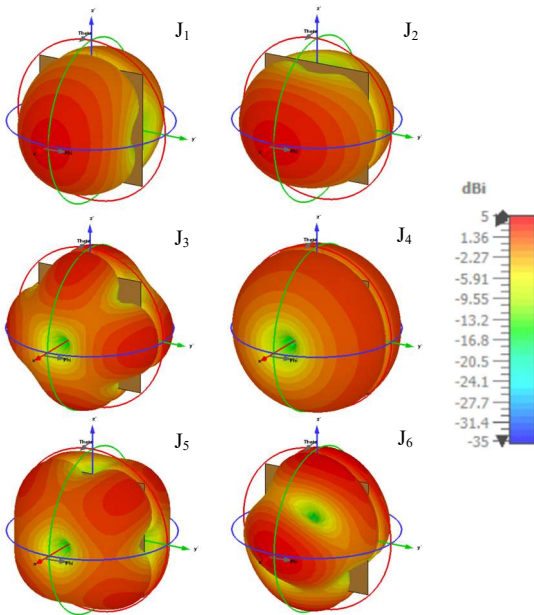


Fig. 3. Radiation pattern of directivity for 1800 MHz for modes 1-6

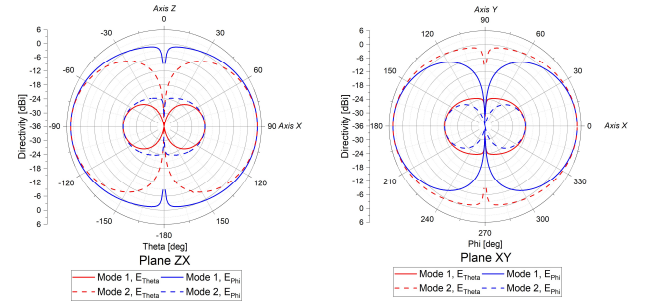


Fig. 4. Radiation pattern for main cut planes for modes 1-2

III. DESIGN OF SINGLE-POLARIZED PROTOTYPE

To excite energy radiation at a specific linear polarization, it is necessary to design appropriate excitations and a feeding network. Therefore this section presents an initial single-polarized design.

A. Selective excitation of the characteristic modes

Referring to [9], the specific distribution of characteristic currents can be enforced by using Inductive Couplers Elements (ICE) or Capacitive Couplers Elements (CCE). It was decided to use ICEs placed in the ground plane where the highest current density was observed for mode 1. The aforementioned inductive coupling elements are slots around which current distribution is forced by shorting the end of the transmission line to the ground plane. The length of the slots corresponds to $\lambda/6$ and with a width of 2 mm. These slots were offset from the edge of the board (as seen in Fig. 5), to allow space for a microstrip transmission line between the edge of the ground plane and the slot. The positions of the coupling elements along with the electrical connection of the transmission line are shown in Fig. 5.

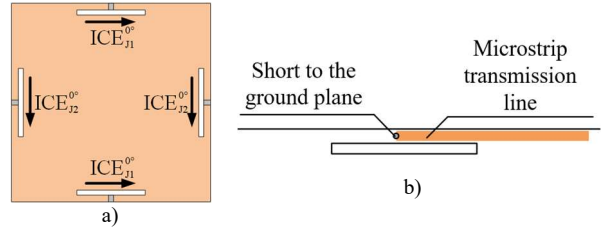


Fig. 5. Antenna model with inductive coupling elements: a) the ground side, b) the microstrip side

B. Antenna feed network design

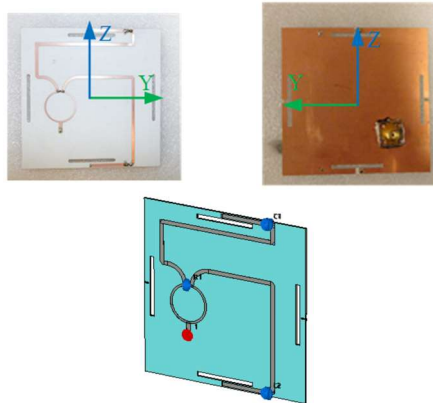
The antenna was designed on laminate Rogers 4003C with $\epsilon_r = 3.38$, substrate thickness of 0.032", and copper foil equal to 1/2 oz. All microstrip lines were designed for an impedance of 50Ω. Matching to each coupling element was done with a gamma circuit with a capacitor connected to the line in series and an inductor connected in parallel (see Fig. 6).

Due to the frequency range, the matching circuit was made on SMD elements, because tuning to the required frequency can be achieved easier than using sections of transmission lines. The line length from the input was chosen to have coupling elements excited in phase. The Wilkinson power divider was designed for a resonant frequency of 1800 MHz. The network was designed to excite mode 1. The Wilkinson power divider is fed by a center pin of the SMA connector whose body is soldered to the ground plane on the opposite side of the board. In the matching network area, the grounding of shunt inductors to the load is ensured by via-pins. Shorting the transmission lines at the ends is also realized by via-pins.

Technical drawing of a mechanical part, likely a bracket or support, showing dimensions in millimeters. The drawing includes a top view and a side view. Key dimensions are:

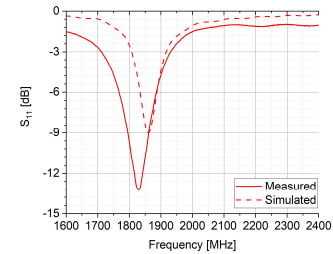
- Overall width: 80.00
- Overall height: 77.73
- Top horizontal distance: 53.24
- Top horizontal distance: 30.71
- Top horizontal distance: 2.00
- Top horizontal distance: 1.00
- Top horizontal distance: 1.60
- Top horizontal distance: 1.01
- Top horizontal distance: 1.82
- Top horizontal distance: 27.47
- Top horizontal distance: 2.00
- Top horizontal distance: .60
- Top horizontal distance: 26.27
- Top horizontal distance: 38.50
- Top horizontal distance: 39.00
- Top horizontal distance: 61.14
- Top horizontal distance: 62.92
- Top horizontal distance: 80.00
- Top horizontal distance: 51.07
- Top horizontal distance: 44.01
- Top horizontal distance: 35.27
- Top horizontal distance: 26.21
- Top horizontal distance: 20.40
- Top horizontal distance: 7.00
- Top horizontal distance: 1.00
- Top horizontal distance: 1.60
- Top horizontal distance: 1.01
- Top horizontal distance: 1.82
- Top horizontal distance: 27.47
- Top horizontal distance: 2.00
- Top horizontal distance: .60
- Top horizontal distance: 26.27
- Top horizontal distance: 38.50
- Top horizontal distance: 39.00
- Top horizontal distance: 61.14
- Top horizontal distance: 62.92
- Top horizontal distance: 80.00

IV. MEASUREMENTS OF THE SINGLE-POLARIZED ANTENNA PROTOTYPE



The prototype of PCB was prepared using LPKF Protomat D104. All measurements presented in this section were compared with the results of the simulation executed using the CST studio Time Domain solver. The feeding point was modelled as a $50\ \Omega$ discrete port in CST. All SMD components were modelled in CST using lumped element models.

circuit employs 1.1 pF capacitor connected in series and 1.6 nH inductor connected in parallel. A comparison between the results of the simulation and the measurements of the fabricated antenna is shown in Fig. 9. The measured bandwidth is 6.5% (1770-1890 MHz) and is two times wider in comparison to the calculated results equal 3.2% (1830-1890 MHz). This discrepancy is most likely due to laminate substrate losses, which may differ from those assumed in the simulation software.



The radiation characteristics were measured using a near-field spherical measuring system available at the Laboratory of Antenna Technology at Wrocław University of Science and Technology. The coordinate system of the antenna under test is shown in Fig. 8. To the comparison of the simulation model with the fabricated antenna the peak of the directivity and the peak of realized gain (see Fig. 10) was used. Moreover, the directivity patterns, for three principal cuts, i.e.: the ZX plane (constant $\varphi=0^\circ$), the ZY plane (constant $\varphi=90^\circ$), the XY plane (constant $\theta=90^\circ$), are shown in Fig. 11.

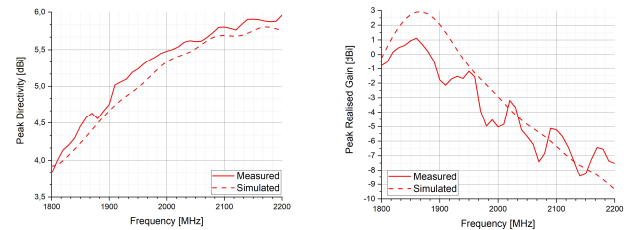


Fig. 11. Radiation pattern for 3 main cross sections and 3D pattern of the total value of directivity

Comparing the directivity patterns, it can be concluded that the results are in good agreement. The direction of maximum radiation of the fabricated prototype matches with the simulation model. The same dominant component of the electric field in the direction of maximum radiation was obtained as in the assumed analysis of characteristic mode 1. The nulls of the dominant component are also in the same direction as for the final PCB board simulation and CMA analysis. The largest discrepancies between measurement and simulation were at θ from -180° to -165° and from 165° to 180° . This is due to the fact, that the facility is not able to measure in these directions and the results are approximated by the far-filed transformation algorithm. As for the peak value of directivity, the difference between the milled prototype and the simulation result does not exceed 0.6 dB, which is considered acceptable. In contrast, the realized gain varies strongly as a function of frequency, and while at the beginning of the antenna's assumed bandwidth, this difference is 0.5 dB, it reaches almost 4 dB for the upper frequency. This may be the influence of the simulation software's unrealistic representation of all material losses. For this reason, it can be seen that measured radiated efficiency is more varied than in the simulation model (see Fig. 12).

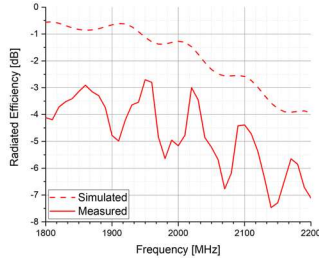


Fig. 12. Comparison of radiated efficiency

V. DESIGN OF DUAL-POLARIZED ANTENNA

The dual-polarized antenna is a prototype consisting of twin boards inverted vertically and rotated 90 degrees to each other in a back-to-back orientation. Assumptions about ports remained the same as in the single-polarized antenna model. The components of the matching network were modelled based on the values of the scattering matrix parameters provided by the SMD component manufacturer. Port 1 fed the ICES stimulating the J_1 characteristic currents associated with mode 1, while port 2 was responsible for the generation of the J_2 characteristic currents from the other side of the board. The prototype was manufactured in the same way as a single-polarization copy. The antenna was fed with coaxial cables instead of direct SMA connectors.

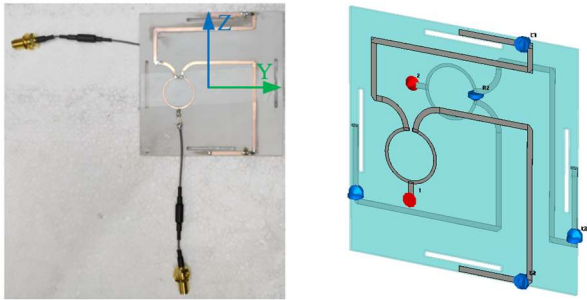


Fig. 13. Photo of the manufactured dual-polarized antenna prototype and simulation model

Based on the simulation results and prototype measurements, the reflection coefficient and isolation of ports were plotted (see Fig. 14). In Fig. 15 the directivity patterns for both modes of simulated model were compared for three principal cuts: the ZX plane (constant $\varphi=0^\circ$), the ZY plane (constant $\varphi=90^\circ$), the XY plane (constant $\theta=90^\circ$) and in Fig. 17 the directivity patterns of measured prototype were plotted.

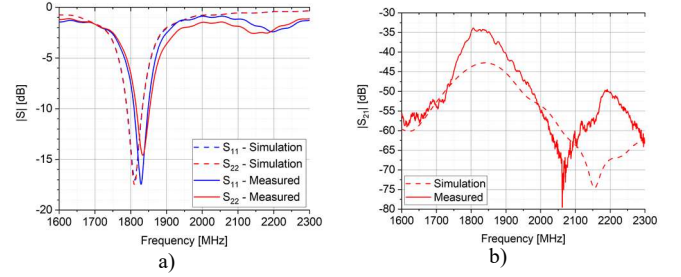


Fig. 14. Reflection coefficient (a) and isolation (b) of dual-polarized antenna ports

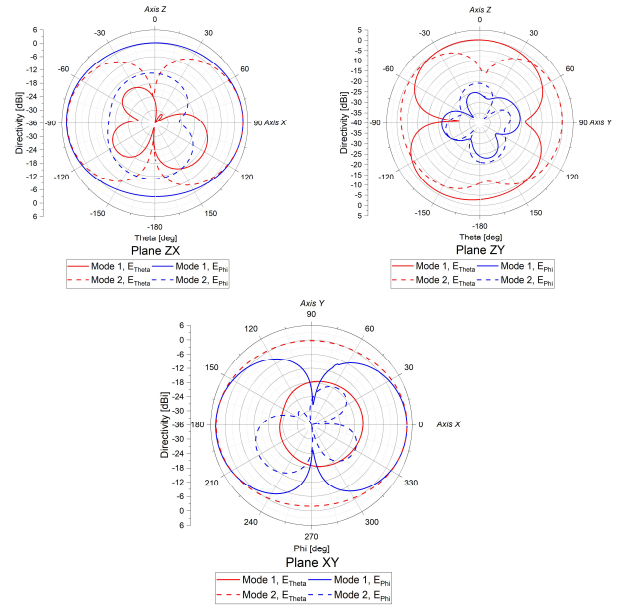


Fig. 15. Radiation pattern of the dual-polarized antenna simulation model

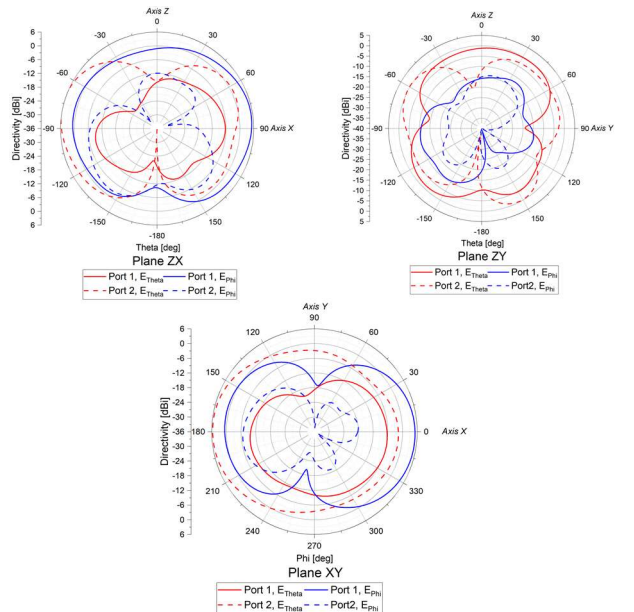


Fig. 16. Radiation pattern of measured dual-polarized antenna prototype

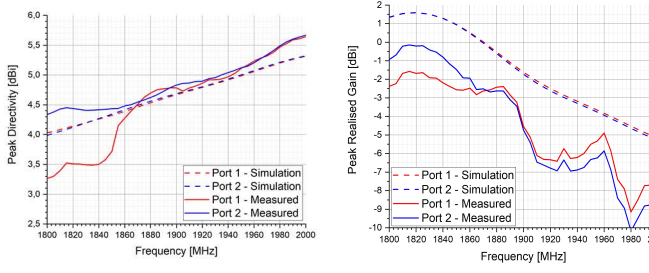


Fig. 17. Peak value of directivity and realised gain of the dual-polarized antenna

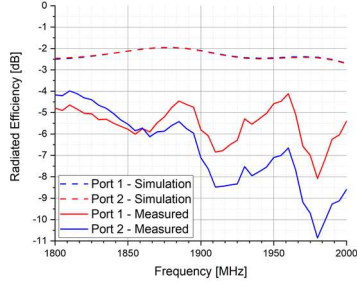


Fig. 18. Radiated efficiency of the dual-polarized antenna

Using a reflection coefficient $\leq -6\text{dB}$ as a reference, the operating bandwidth of the dual-polarized antenna simulation model is 4.4% (1770-1850 MHz) and 3.8% (1800-1870 MHz) for the manufactured prototype. For this band, the isolation between both simulated ports is $> 40\text{dB}$, compared to the measured antenna isolation of $> 35\text{dB}$. The difference between the peak directivity of the milled prototype and the simulation result reaches up to 1.2dB. In the operating band, the measured realized gain is 1.5dB lower than the simulation model for port 1 and 3dB for port 2. The radiated efficiency differs from the simulation value by 4dB. The calculated discrimination between ports in the X-axis direction is -18dB, while for the measurement of the manufactured prototype, it is -11 dB. The envelope correlation coefficient of the simulated model in the operating band of the antenna between modes does not exceed 0.001 (see Fig. 19).

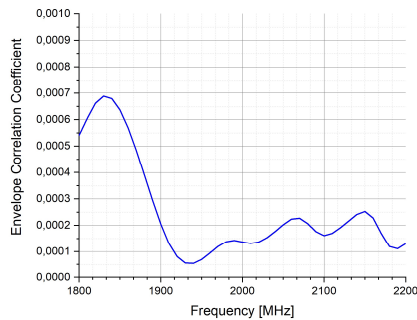


Fig. 19. Envelope Correlation Coefficient between modes of the simulation model

Given that the values of the aforementioned parameters differ significantly between the simulation and the made prototype, which was not the case for the single-polarized antenna, we suppose that the problem may be the technology of making this prototype. At the stage of assembling the prototype antenna, it was noted that the position of the coaxial cables relative to the board has a major impact on both impedance matching and also directivity. Making a prototype antenna using two PCBs connected to each other with the common ground layer is not the best technology due to the

inability to ensure continuity of conductivity across the surface.

VI. CONCLUSION

The implementation of the antenna feed network based on a microstrip line with a solid metallic plate as a radiating structure makes it possible to achieve omnidirectional radiation characteristics while simplifying the board fabrication technology. The implementation of a matching network based on a series capacitor and a parallel coil at each inductive coupling element makes it possible to tune the antenna to a given frequency with a bandwidth of 3.5-6%. The simulation model of the dual-polarized antenna achieves high isolation and good polarization discrimination in the direction of maximum radiation.

We assume that the antenna's performance can be improved by further refinement of the design and implementation of the final antenna on the 4-layer board. We also plan to work on expanding the antenna's operating bandwidth by making improvements to the matching network.

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