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Design of Compact Logarithmically Periodic Antenna Structures for Polarization-Invariant UWB Communication

Oliver Klemp[a], Hermann Eul

[a]Department of High Frequency Technology and Radio Systems,
Hannover, Germany, e-mail: klemp@hft.uni-hannover.de

Abstract - Ultra Wideband (UWB) communication systems seem promising in order to establish a new standard for short-range data transmission with enhanced data rates in a frequency range from 3.1-10.6 GHz. The antenna design for UWB signal radiation is one of the main challenges, especially when low-cost geometrically small and efficients structures are required, serving as air-interfaces to the UWB transmission channel. Self-complementary antenna structures may be well suited in the design of UWB communication systems providing almost constant antenna parameters in the whole frequency range of operation. Adhering to certain geometrical delimitations, self-complementary, logarithmically periodic planar antennas can be realized as dual polarized antenna elements that may be applied in short-range communication systems with polarization diversity reception.

1 Introduction

UWB short-range wireless communication is about to become a global standard applying data transmission in the frequency band of 3.1-10.6 GHz. Suitable antenna elements that are able to cover the whole frequency band of operation are a key parameter in the successful implementation of UWB communication systems. Antenna parameters like input impedance, antenna gain and radiation pattern need to be almost constant with frequency to provide identical transmission conditions for the whole frequency band of operation ([1]). Therefore, in order to reduce transmission noise from dispersion of antenna parameters, a constant and frequency-independent position of the phase center of the antenna is recommended.
Following the well known design rules of frequency-independent antenna structures, the principle of self complementary metalization surfaces may be applied as well for the design of UWB antennas. Self-complementary, multi-terminal antenna structures provide purely real impedance processes ([2]) at the individual antenna ports.

Many broadband wireless systems require not only broadband antennas, but also dual polarization ([3], [4], [5]). In order to combat the effects of polarization fading ([4]) in portable radio environment, antennas are to be used that offer polarization-invariant transmission characteristics. Applying polarization invariant antenna structures in an environment with uncorrelated polarization fading, orthogonal transmission channels are formed, whose simultaneous use leads to an enhancement in quality of service. Therefore, dual-polarized broadband antennas can double the communication capacity over the same physical link in line-of-sight scenarios and can improve performance in multipath conditioned systems using a diversity scheme. Self-complementary, logarithmically periodic planar antennas ([6], [7]) serve as suitable representatives for the realization of broadband antenna structures. As shown in [8] they can be favorably used in polarization diverse transmission systems provided they adhere to a specific geometric structure. Using suitable basis geometry functions, log.-per. planar antennas can be realized with almost arbitrary polarization characteristics over a wide band of frequency use.

The article is organized as follows: The design of a planar, dual polarized log.-per. antenna element for UWB communications is presented in Section 3. It is deduced from the design of an adequate log.-per. trapezoid antenna as in [8] that was used in the field of mobile personal communications in a frequency range from 1.7 GHz to 6 GHz. The related basic principles in the design flow of log.-per. antenna modules and the initial antenna design are given in Section 2. The measurement results for this initial antenna design are summarized in Section 2.2.

2 Dual Polarized Broadband Antenna Design

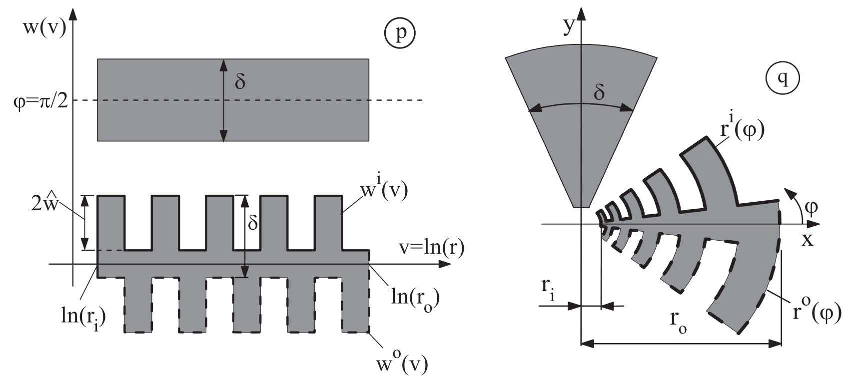


Figure 1: Design process of logarithmically-periodic antenna structures based upon a conformal transformation between the p- and the q-plane.

Subsequently planar, arbitrarily shaped metalization surfaces with infinite radial extent, an ideal electrical conductance and an infinitesimal geometrical thickness will be considered. Presumed the regarded geometries merge by exchanging conducting- with isolating surfaces, they provide self-complementary properties. The planar, self-complementary spiral- and logarithmically-periodic antennas are well known representatives to frequency-independent antenna designs ([9]). In terms of the desired polarization diversity reception, spiral antennas are not to be considered subsequently. Their distinct direction of rotation clearly assigns to a certain direction of circular polarization, and therefore is not adequate for the implementation in a polarization diversity transmission system.

In order to obtain compact antenna structures that can be easily integrated into short-range wireless communication devices, the design is restricted to planar log.-per. antenna structures. Considering the fact that each polarization state in a plane may be decomposed into a system of two orthogonal linear polarizations, the design of an antenna structure offering $N = 4$ antenna arms is an adequate choice. Antenna structures of that kind theoretically provide frequency-independent real input impedances of $133\,\Omega$ at the individual antenna ports ([10]).

The synthesis of log.-per. antenna structures starts with the choice of an adequate basis geometry function (so-called unit cell). From the literature e.g. sinusoidal basis functions, rectangular- and triangular basis functions are known ([6], [9]). In general, log.-per. antenna geometries are given by means of a conformal transformation between the corresponding basis geometry in the so-called p-plane and the complex q-plane according to:

$$q = x\,(v,w) + j\,y\,(v,w) = r\,(v)\;e^{j\,\varphi(v,w)} \qquad (1)$$
$$\text{with} \quad r\,(v) = e^{v} \quad \text{and} \quad \varphi\,(v,w) = w\,(v)\;.$$

In (1), v and w represent the cartesian coordinates in the p-plane, the complex q-plane is given in terms of x and y. Therefore, vertical sections of the unit cell are mapped to circular arc segments, horizontal sections lead to radial lines in the q-plane. Fig. 1 a), depicts two examples of periodic repetitions of unit cell geometries in the p-plane. The basis geometry functions shown provide a rectangular shape and a shape with constant amplitude. Applying the transformation given in (1), the basis geometry functions are mapped to the geometries in the image plane as shown in Fig. 1 b). δ is the angular spread of one antenna arm. In the radial direction, the antenna arm is delimited by the radii r_i and r_o. In Fig. 2, the respective symmetrical four-arm antenna with a rectangular shape is shown. Due to the functional principle of planar log.-per. antenna structures, each arm can be decomposed into two main operational units: A tapered, radial oriented line that distributes the transmission line power from the feeding point at the center of the antenna structure to the outer antenna region and azimuthally oriented resonators with different angular lengths l_q. Radiation occurs at radius positions r_rad, where the effective electrical lengths of the resonators are equal to a quarter of the operational wavelength λ ([7]) so that they are able to scatter electromagnetic energy. The theoretical boundaries of a quasi frequency-independent transmission behavior are limited by the azimuthal lengths of the resonators with the smallest- and with the largest angular lengths $l_{\mathrm{q,min}} = l_\mathrm{q}\,|_{r_\mathrm{i}}$ and $l_{\mathrm{q,max}} = l_\mathrm{q}\,|_{r_\mathrm{o}}$.

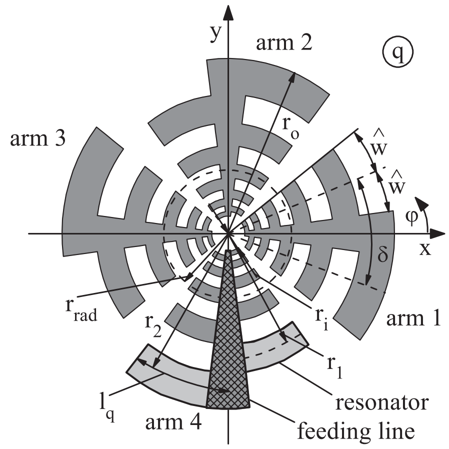


Figure 2: Symmetrical, log.-per. four-arm antenna structure based upon a rectangular basis geometry function.

The choice of the unit cell is arbitrary but influences to the cross-polarization decoupling between two orthogonal antenna arm pairs and the lower cut-off frequency. The trapezoidal basis geometry as reported in [8] was shown to provide the lowest cut-off frequency and the highest degree of cross-polarization decoupling among the basis geometry functions reported in the literature. In terms of an operation with dual linear polarization, each of the two pairs of adjacent antenna arms is related to a distinct linear polarization state that represents one diversity branch of the polarization diversity system. Therefore, differential voltages are applied at the feeding bridge of adjacent antenna arms. The trapezoid antenna consists of a planar, self-complementary ($\delta = \pi/4$) four-arm geometry, with $M = 5.5$ unit cell periods and an amplitude of $\hat{w} = \pi/4$. The percentage slope of the trapezoid is given as $s_\mathrm{s} = 50\%$. The antenna is extended in a radius interval of $r_\mathrm{i} = 5.0$ mm and $r_\mathrm{o} = 50.0$ mm in free-space.
Log.-per. antenna modules adhere to the principle of discrete scaling invariance. Therefore, the quotient of the angular lengths l_1 and l_2 of the unit cells of two subsequent resonators as shown in Fig. 2 equals to the inverse of the so-called self-similarity factor τ.

2.1 Specification of Operational Bandwidth

In practical realizations the frequency-independent impedance behavior of the antenna structure is limited due to the utilization of specialized feed assembly and metalization surfaces with finite extent. The radial extent of the metalization surfaces is defined by the outer delimitation radius r_o and the inner radius limit r_i which is determined by coaxial feeds at the center of the antenna structure. Although if the electromagnetically active regions are within the radial delimitations of the metalization surfaces and therefore offer self-complementary properties, almost constant characteristics of the feed impedance in a reduced frequency range of operation can be observed ([6]). From there the lower cut-off frequency for the quasi frequency-independent operational mode of the antenna structure is determined by the outer delimitation radius r_o. Generally, the theoretical lower- and upper cut off frequencies $f_\mathrm{l,c}$ and $f_\mathrm{u,c}$ of quasi frequency-independent impedance behavior can be formulated in accordance with [11] as follows:

$$f_\mathrm{l,c} = \frac{c_0}{4r_\mathrm{o}\left(\frac{\delta}{2}+\hat{w}\right)} \quad \text{and} \quad f_\mathrm{u,c} = \frac{c_0}{4r_\mathrm{i}\left(\frac{\delta}{2}+\hat{w}\right)} \,. \tag{2}$$

We will address our design restrictions to the definition of a lower cut-off frequency that specifies the lower bound of quasi frequency-independent impedance characteristics. Using a $\pm 25\%$ range of tolerance around the expected impedance value of $133\,\Omega$ which is representative for an ideal self complementary four-arm antenna structure in free space, the cut-off frequency is defined by the lowest frequency as of the impedance process remains in the tolerance range. As given in [8] the cut-off frequency of the trapezoid antenna amounts to $f_\mathrm{l,c} = 1.7$ GHz.
Subsequently the radiation properties of the planar, log.-per. four-arm trapezoid antenna will be summarized. Fig. 3 shows the radiated power in case of an excitation with dual linear polarization in a frequency interval from 1 GHz to 6 GHz. Apart from the lower cut-off frequency at $f_\mathrm{l,c} = 1.7$ GHz, the radiated power remains almost constant to be $P_\mathrm{rad} \simeq 8$ mW. In terms of antenna gain as shown in Fig. 4 a), a mean value of 5.7 dBi is observed with a slight variation of about 1.3 dBi in the interesting frequency range. As shown in Fig. 4 b) the cross polarization suppression in main beam direction for linear polarization, $\Gamma\left(\vartheta_0 = 0°, \varphi_0 = 0°\right)$, remains above approximately 20 dB in the entire frequency range of operation.

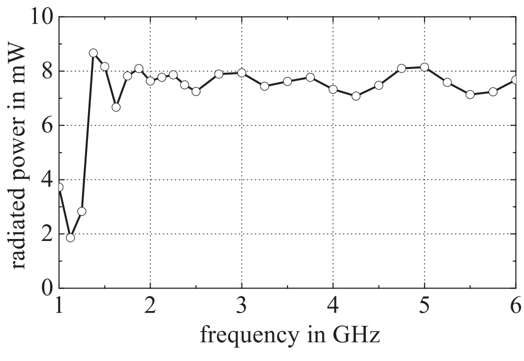
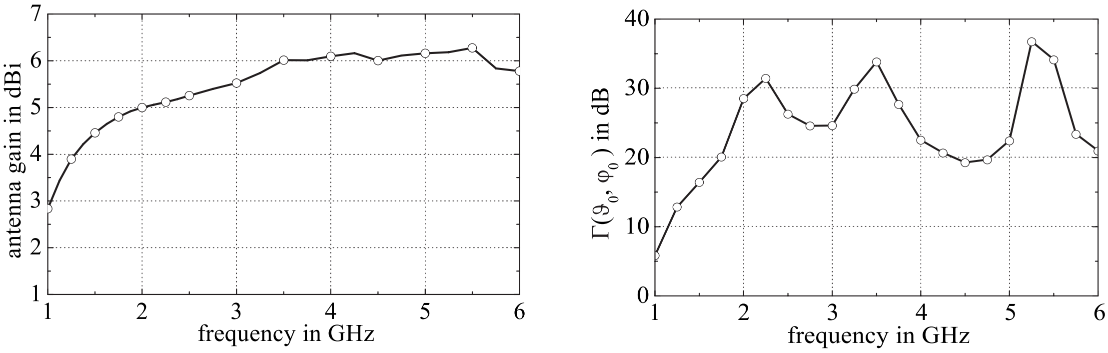


Figure 3: Radiated power in terms of an excitation with dual linear polarization of the trapezoid antenna as given in [8] in free-space.

2.2 Measurement Results

In order to verify the simulation results for the trapezoid antenna structure, the antenna element was developed on microwave substrate ($\epsilon_\mathrm{r} = 2.33$) and embedded between two substrate layers leading to a reduction of the theoretical impedance values at each antenna port to approximately $133\Omega/\sqrt{\epsilon_\mathrm{r}} \simeq 87\Omega$.



a) Antenna gain b) Cross-Polarization suppression

Figure 4: Radiation characteristics of trapezoid antenna as given in [8] in free-space.

Fig. 5 a) shows the results of input impedance behavior of the initial antenna design for simulation and measurement. In order to determine a lower cut-off frequency, the limits of the tolerance range were increased to account for the impairments introduced by the measurement setup. Reflections in the interface planes of the substrate material as well as interactions with the finite ground plane lead to an increased variance of the impedance chart. Therefore a $\pm 70\%$ range of tolerance is defined around the theoretically expected average value of approximately $87\,\Omega$. Simulation and measurement consistently yield to a cut-off frequency of quasi frequency-independent impedance behavior of approximately 1.7 GHz. In the frequency range from 1.7 GHz to 6GHz the measured impedance chart offers a mean value of $84.1\,\Omega$ and a resulting standard deviation of $32.9\,\Omega$.

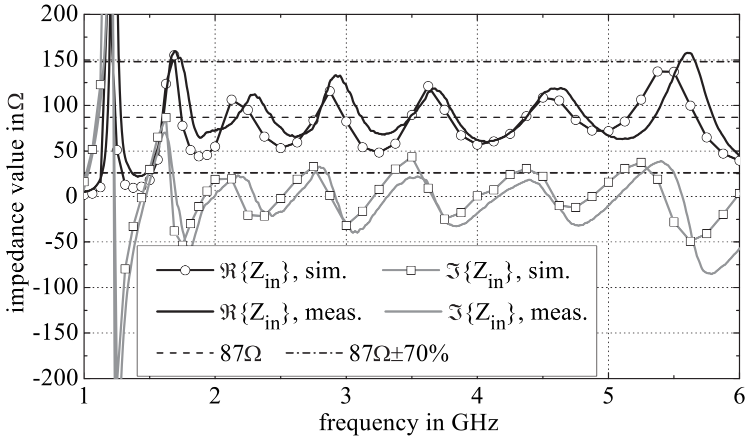


Figure 5: Simulation and measurement results of input impedance for one antenna arm of planar trapezoid antenna.

To evaluate the radiation behavior of the antenna structure, field measurements for a linear polarization state of the antenna module were performed at the frequency points $f_1 = 1.8\,\text{GHz}$ and $f_2 = 5.85\,\text{GHz}$. Numerical results from antenna far-field pattern simulation serve as a comparison. Fig. 5 a) and Fig. 5 b) show the normalized radiation pattern in the plane $\varphi = 90°$ for both frequencies. For $f_1 = 1.8\,\text{GHz}$, antenna main beam is directed to $\vartheta = 0°$ with a co-polar half-power beamwidth of $\vartheta_{\text{hbw}} = 92°$. At 5.85 GHz the co-polar pattern is also given by a global maximum at the desired angular position of $\vartheta = 0°$ with a slightly reduced co-polar half-power beamwidth of $\vartheta_{\mathrm{hbw}} = 69°$. The cross polarization discrimination results to $\Gamma(\vartheta_0, \varphi_0) = 18.5\,\mathrm{dB}$ at 1.8 GHz. According to the far field distribution at 5.85 GHz a cross-polarization discrimination of 18 dB is observed. Both, simulation and measurement are in excellent agreement.

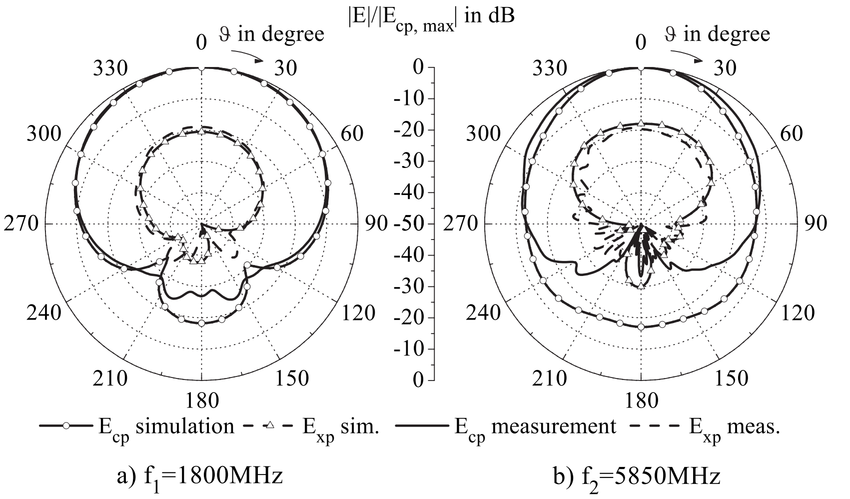
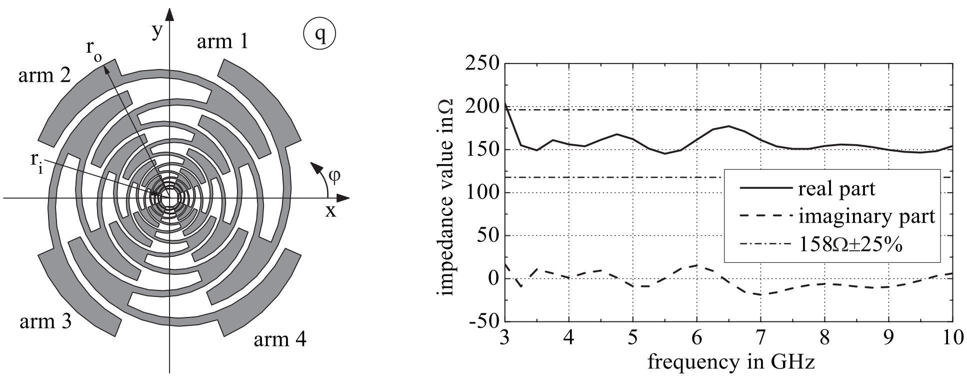


Figure 6: Simulation and measurement results of antenna radiation patterns for planar trapezoid antenna at $\varphi = 90°$.

3 Adapted Dual Polarized UWB Antenna Design

Subsequently the principle of discrete self-similarity ([6], [7]) is used to adapt the properties of the four-arm log.-per. trapezoid antenna given in Section 2 to the frequency range of UWB communications from 3.1 GHz to 10.6 GHz. Regarding the constant factor of discrete self-similarity, τ, the respective unit cell geometry of an antenna design as depicted in the p-plane of Fig. 1 a) is assigned to a periodic repetition of electromagnetic active regions with different geometrical dimensions in the q-plane. As shown in Sec. 2 radiation occurs if the operational wavelength of the antenna reaches a quarter in length of the azimuthally oriented resonators. Therefore by a scaling of τ the respective geometrical dimensions of the mapped unit cell sections may be easily matched to the desired frequency range of operation. Regarding the correspondence of angular lengths and antenna radii as in (2), log.-per. antenna structures can be designed in the desired frequency range of operation by fitting the radial delimitations of the antenna. It is therefore an elementary property of scaling-invariant antenna structures to adopt their frequency range of operation by matching the inner- and outer delimitation radii, respectively.

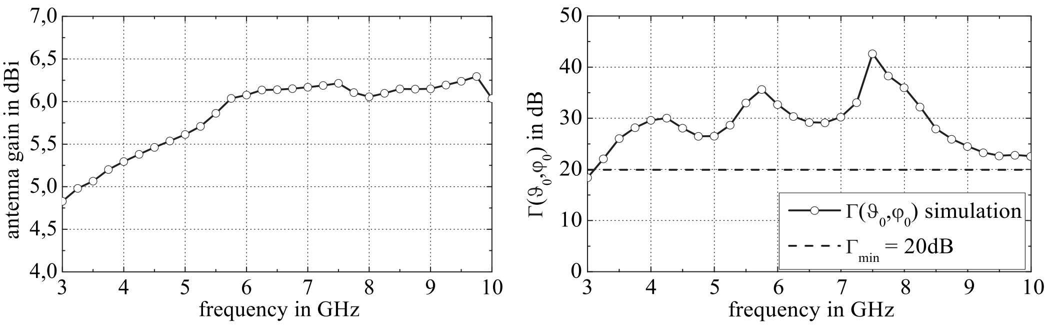
Applying the trapezoidal basis geometry as given in Section 2, the proposed design of a dual polarized four-arm UWB antenna is delimited by an inner radius of $r_{\mathrm{i}} = 2.0$ mm and an outer radius of $r_{\mathrm{o}} = 30.0$ mm. Using the design equations (2) in Sec. 2.1 and the geometry parameters from Section 2, the theoretical cut-off frequencies are specified by $f_{\mathrm{l,c}} = 2.1$ GHz and $f_{\mathrm{u,c}} = 31.8$ GHz, resulting in a relative operational bandwidth of $B_{\mathrm{rel}} = 15.1$.
Fig. 7 a) shows the dual-polarized four-arm trapezoid UWB antenna, that adheres to the simulated input impedance process, as given in Fig. 7 b). The observed mean value for the real part of the input impedance amounts to $157.9\,\Omega$ in free-space with a standard deviation of $11.96\,\Omega$ and therefore slightly deviates from the theoretically expected mean value of $133\,\Omega$. The imaginary part of the input impedance remains low so that a wide-band matching to the real value of the input impedance can be applied.



a) UWB antenna design

b) Simulated input impedance process

Figure 7: Geometry and simulated input impedance process of polarization-diversity UWB trapezoid antenna in free-space.



a) Antenna gain

b) Cross-polarization discrimination

Figure 8: Simulated results for antenna gain and cross-polarization discrimination $\Gamma(\vartheta_0, \varphi_0)$ dB for UWB antenna design in free-space.

Simulated properties of the UWB antenna design are given in Fig. 8 a) in terms of antenna gain. Antenna gain achieves a mean value of 5.85 dBi offering a standard deviation of 0.44 dBi in free-space. Fig. 8 b) shows the properties of cross polarization suppression for one adjacent pair of antenna arms in a linear polarized operational mode in main beam direction, $\Gamma(\vartheta_0 = 0°, \varphi_0 = 0°)$. The trapezoidal four-arm antenna design provides an outstanding degree of cross polarization discrimination in main beam direction better than 20 dB in the entire frequency range of operation.

4 Conclusion

Based upon the definition of trapezoidal unit cell geometries, log.-per. antenna structures with enhanced cross polarization decoupling in an increased frequency band of operation may be realized. According to the initial design of a log.-per. four-arm antenna structure that was introduced to cover the frequency range for mobile personal communications from 1.7 GHz to 6 GHz, the advantageous behavior of the trapezoid antenna resulting in an excellent polarization purity could be verified very well by means of simulation and measurement. Hence the self-complementary, log.-per. four-arm trapezoid antenna is in the best way suitable for the implementation in a dual polarized broadband antenna system. Using the principle of a discrete scaling invariance, the initial antenna design could be easily adapted to the frequency range of 3.1-10.6 GHz for UWB communications. The adopted antenna element can be matched to a constant value of input impedance and provides an excellent cross-polarization decoupling in the whole frequency band of operation. For that reason, the four-arm trapezoid UWB antenna may be favorably used for the integration into compact UWB communication systems applying polarization diversity reception.

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