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DESIGN AND MEASURED PERFORMANCE OF DUAL POLARISED APERTURE-COUPLED MICROSTRIP PATCH ANTENNAS

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Abstract

The paper discusses the design of aperture-coupled microstrip patch antennas for dual linear polarisation with special emphasis on achieving a wide bandwidth and a high polarisation isolation. The coupling behaviour of crossed slot and offset slot designs is discussed and compared to each other. Several design examples for aperture-coupled antenna elements illustrate how the polarisation isolation decreases as the bandwidth of the antenna elements is increasing. A design example for an 8x1 element subarray for a frequency in X-band (9.6 GHz) shows a standing wave ratio of VSWR<1.5 and a polarisation isolation of better than 25 dB within a bandwidth of 470 MHz.

Keywords: aperture-coupled antennas, dual linear polarisation.

1. Conventional Microstrip Patch Antennas

Microstrip patch antennas are very attractive candidates for versatile active phased array antennas offering several advantages such as low weight, low thickness as well as the integration of the feed network.

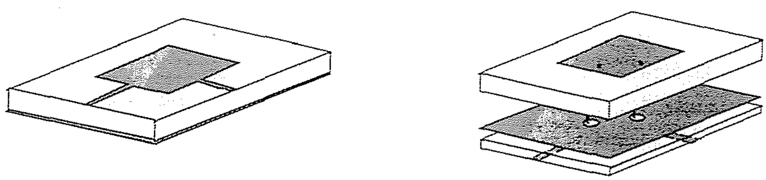

Fig. 1. Conventional dual polarised microstrip patch antennas
a) single layer microstrip-fed b) multilayer probe-fed

However, conventional microstrip antennas with a single substrate for both the radiating elements and the feed network as shown in *Fig. 1a* show only narrow bandwidth and low efficiency. Surface waves produce diffraction at the edges of the dielectric and coupling between the array elements. Parasitic radiation of the feed network produces higher sidelobes and cross-polarisation levels.

The described problems can be avoided, if the radiators and feed network are separated on different substrate layers. Using direct probe feeding as shown in *Fig. 1b*, however, solder connections are necessary on both the patch as well as the feed network layer. For large arrays this implies high production costs. In addition the solder connection itself can be problematic considering thermal and mechanical stresses.

2. Aperture-Coupled Microstrip Patch Antennas

Aperture-coupled microstrip antennas are very advantageous to overcome these difficulties. With a two layer design, the antenna elements and the feed network are located on separate substrates, which yield optimum array performance and eliminate the space competition between the antenna elements and the feed network (*Fig. 2*). Furthermore, the antenna elements can be mounted on a substrate with a low dielectric constant to increase the bandwidth and the radiation efficiency. Using slots as coupling apertures, the feed network is entirely shielded by the groundplane. With this configuration, excellent sidelobe performance and polarisation properties can be achieved.

2.1 Dual Polarisation Capability

There are several possibilities for the design of an aperture-coupled patch radiator for dual linear polarisation. As an example *Fig. 2* shows a crossed slot as well as an offset slot configuration.

For the crossed slot configuration two separate substrate layers are necessary for the two orthogonal feed lines, as it can be seen in *Fig. 2a*. This implies that additional care has to be taken for the exact positioning of this additional layer during the manufacturing process, resulting in higher production costs.

In order to avoid this additional layer two offset slots can be used as shown in *Fig. 2b*. With this approach both feed lines can be placed within the same layer.

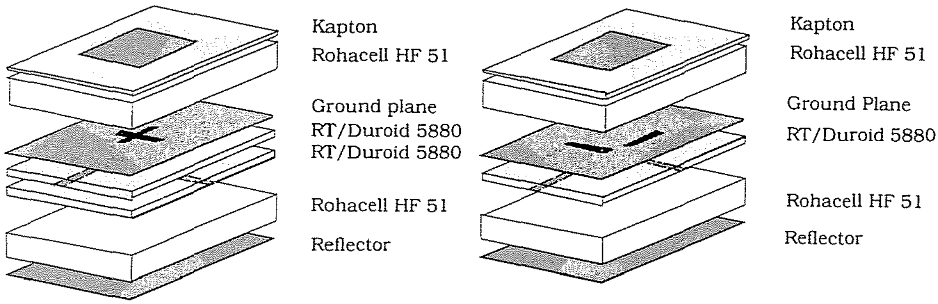

Fig. 2. Dual polarised aperture-coupled microstrip patch antennas with a) crossed slots and b) offset slots

In order to find the optimum feeding technique for a maximum bandwidth together with a minimum coupling level between both polarisations both designs were investigated theoretically and experimentally.

2.2 Modelling

The mathematical modelling of aperture-coupled antenna elements can be performed by an electric field integral equation method in combination with the reciprocity theorem as formulated by Pozar in [1]. In order to be able to account also for offset slots as well as for multiple dielectric layers the approach in [1] has to be extended as shown below.

In the analysis is assumed that all dielectric substrates below and above the ground plane extend to infinity in the x- and y-directions. The slot can be replaced by two equivalent magnetic currents M_y on both sides of the metallisation, having opposite direction. These unknown currents can be approximated by a single piecewise-sinusoidal (PWS) expansion mode:

$$M_y(x,y) = \begin{cases} \frac{1}{W_s} \cdot \frac{\sin(k_e(\frac{L_s}{2} - |y|))}{\sin(k_e \frac{L_s}{2})} & \text{for } \begin{array}{c} |y| < \frac{L_s}{2}\,, \\ |x| < \frac{W_s}{2} \end{array} \\ 0 & \text{elsewhere} \end{cases} \qquad (1)$$

with L_s the slot length, W_s the slot width and k_ϵ the effective wavenumber of the PWS mode. The current on the patch is expanded into a set of N entire domain modes for the x-variation and uniform (pulse) modes for the y-variation:

$$J_{sx}(x,y) = \sum_{n=1}^{N} I_n \cdot J^n(x,y) = \sum_{n=1}^{N} I_n \cdot J_0^n(x - S_x, y - S_y)\ . \qquad (2)$$

$$J_0^n(x,y) = \begin{cases} \frac{1}{W_p} \cdot \sin\left(\frac{n\pi}{L_p}\left(x + \frac{L_p}{2}\right)\right) & \text{for } \begin{array}{l} |x| < \frac{L_p}{2} \\ |y| < \frac{W_p}{2} \end{array} \\ 0 & \text{elsewhere} \end{cases} . \tag{3}$$

L_p denotes the patch length, W_p the patch width and I_n is the amplitude of the n-th mode on the patch. S_x and S_y represent the geometrical offset of the coupling slot from the center of the patch. The electric field integral equation can be used to enforce the boundary condition that the total tangential electric fields (due to the slot excitation and the currents excited on the patch) are zero over the patch surface. This results in the matrix equation

$$[Z] \cdot [I] = [V] , \tag{4}$$

where $[I]$ is the column vector of the unknown patch expansion mode coefficients, $[Z]$ is the moment method impedance matrix of the patch and $[V]$ is the voltage vector due to the excitation of the magnetic currents in the offset slot. The elements of the impedance matrices are given by

$$Z_{mn} = -\int_{Sp_0}\int_{S_p} J^m(x,y) \cdot G_{xx}^{EJ}(x,y,x_0,y_0) \cdot J^n(x_0,y_0) \cdot dx_0 dy_0 dx dy \tag{5}$$

and

$$V_m = \int_{Sa_0}\int_{S_p} J^m(x,y) \cdot G_{xy}^{EM}(x,y,x_0,y_0) \cdot M_y(x_0,y_0) \cdot dx_0 dy_0 dx dy , \tag{6}$$

where the integrations take place over the surface of the patch S_p and the surface of the coupling slot aperture S_a. G_{xx}^{EJ} and G_{xy}^{EM} are space domain Green's functions representing the transverse electric field caused by an electric current element on the patch or a magnetic current element in the slot, respectively.

Since the exact Green's functions for multilayered media cannot be directly derived in the space domain, *Eqs.* (5) and (6) are solved in the spectral domain. For this reason a two-dimensional Fourier transform pair is used to obtain the electromagnetic field components in the spectral domain.

$$\tilde{F}(k_x,k_y,z) = \iint_{\infty} f(x,y,z) \cdot e^{+j(k_x x + k_y y)} dx dy , \tag{7}$$

$$f(x,y,z) = \frac{1}{4\pi^2} \iint_{\infty} \tilde{F}(k_x,k_y,z) \cdot e^{-j(k_x x + k_y y)} dk_x dk_y . \tag{8}$$

Eqs. (5) and (6) can then be written as

$$Z_{mn} = -\frac{1}{4\pi^2} \iint\limits_{\infty} \tilde{J}_0^{m^*}(k_x, k_x) \cdot \tilde{G}_{xx}^{EJ}(k_x, k_x) \cdot \tilde{J}_0^{n}(k_x, k_x) \cdot dk_x dk_y \tag{9}$$

and

$$V_m = \frac{1}{4\pi^2} \iint\limits_{\infty} \tilde{J}_0^{m^*}(k_x, k_x) \cdot e^{jk_x S_x} \cdot e^{jk_y S_y} \cdot \tilde{G}_{xx}^{EM}(k_x, k_x) \cdot \tilde{M}_y(k_x, k_x) \cdot dk_x dk_y \ , \tag{10}$$

respectively, where $\tilde{G}_{xx}^{EJ}$ and $\tilde{G}_{xy}^{EM}$ represent the Fourier transformed Green's functions in the spectral domain.

Starting with the Maxwell equations in symmetrical form (including a fictive magnetic current density M) equivalent circuits can be derived for these spectral domain Green's functions for arbitrary stacked dielectric media [2].

In the spectral domain the fields can be decomposed into a spectrum of plane waves propagating along the interfaces of the dielectric layers in x-y-direction with the propagation vector $\mathbf{k} = k_x \cdot \vec{\mathbf{e}}_x + k_y \cdot \vec{\mathbf{e}}_y$.

By introducing a new u-v-coordinate system rotated by the angle $\phi = \arctan(\frac{k_y}{k_x})$ as given in *Fig. 3* a separation in TE- and TM-field components is possible.

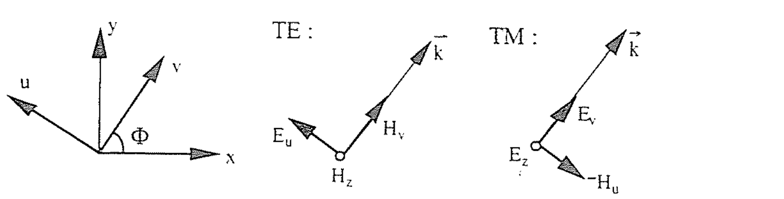

Fig. 3. Decomposition of the electromagnetic fields in TE- and TM-modes

The Maxwell equations applied to the multilayered structure can be modelled then by two equivalent circuits as shown in *Fig. 4*.

The electric current elements on the patch and on the feed line can be described by ideal current sources, whereas the magnetic current elements on the aperture can be replaced by an ideal voltage source.

Each dielectric layer of thickness d_i is represented by an ideal transmission line with the length d_i, the characteristic impedance $Z_{TE/TMi}$ and the propagation constant γ_i. The ground plane is modelled as a simple short circuit. The free space above the patch is replaced by an infinitely

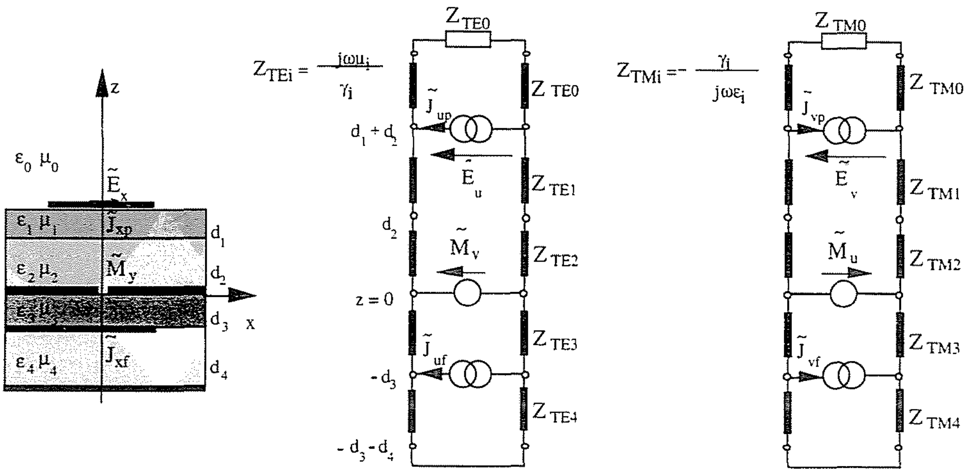

Fig. 4. Equivalent circuits to determine the Green's functions of the multilayered structure

long ideal transmission line whose input impedance is equal to its characteristic impedance.

$$\gamma_i = \sqrt{k_x^2 + k_y^2 - \varepsilon_i k_0^2}\ , \tag{11}$$

$$Z_{TEi} = \frac{j\omega\mu_i}{\gamma_i}\ , \quad Z_{TMi} = \frac{\gamma_i}{j\omega\varepsilon_i}\ . \tag{12}$$

Using transmission line theory the desired relationships between the electric and magnetic currents and the corresponding electric and magnetic fields can be determined for the TE- and TM-mode, respectively . The spectral domain Green's functions needed for the analysis of the aperture-coupled patch are obtained by rotating back to the original x-y-system.

When the patch currents have been calculated as solutions of the electric field integral *Eq.* (4) the coupling of the slot to the microstrip feedline is incorporated using the reciprocity theorem [1]. As all the electric and magnetic current distributions are known, all relevant antenna parameters such as the input impedance, the bandwidth, the antenna gain and the radiation patterns can directly be determined.

In order to verify the described method of moments approach a single aperture-coupled antenna element for a center frequency of 9.6 GHz was designed and the results compared with measurements. A 1.2 mm thick Rohacell HF 51 foam material was used as a substrate layer for the patches. Because the patches cannot be mounted directly on the foam an additional copper clad Kapton film of 90 μm thickness was used as a support layer for the patches. The patch dimensions were chosen to 12×12 mm^2 and the slot dimensions to 8.2×0.5 mm^2. The offset of the slot from the center of the patch was 4.0 mm in x-direction and 2.5 mm in y-direction.

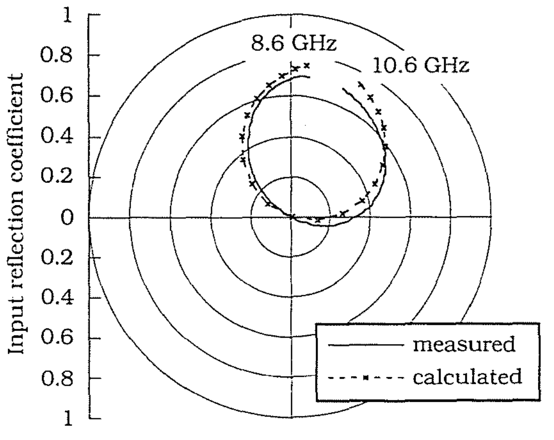

Fig. 5. Calculated and measured input reflection coefficient

Fig. 5 shows the measured and calculated input reflection coefficient for this antenna element in the frequency range from 8.6 to 9.6 GHz. The calculation was performed with 3 even and 5 odd sinusoidal entire basis functions on the patch and one PWS mode in the coupling slot. Excellent agreement between calculated and measured results have been achieved.

2.3 Practical Realisation

In order to show the correlation between the antenna bandwidth and the corresponding attainable polarisation isolation for the offset slot configuration as well as the crossed slot configuration several dual polarised antenna elements were designed with the method of moments approach [1,3], built and measured. All antenna elements were designed for a center frequency of 9.6 GHz. In order to achieve different bandwidths Rohacell HF 51 with a thickness of 1.4 mm, 1.6 mm, 1.8 mm and 2.0 mm was applied as a patch substrate. As a substrate for the feed lines Duroid 5880 with a thickness of 0.254 mm for the offset slot design and a thickness of 0.254 mm + 0.127 mm for the crossed slot design was used. The measured scattering parameters show a characteristic coupling behaviour between both polarisations for both feeding concepts over frequency. As an example *Fig. 6* shows the comparison of the measured scattering parameters for the built-up antenna elements with a patch substrate thickness of 1.6 mm. For the crossed slot design the coupling reaches its maximum at the center frequency and decreases for higher and lower frequencies. For the offset slot design the highest coupling occurs for frequencies below the center frequency, whereas the minimum of the coupling is achieved about 300 MHz above the center frequency. The measured bandwidth (VSWR < 1.5) for both designs is 470 MHz for the crossed slot design and 550 MHz for the offset slot design. A comparison of the measured bandwidth for all the investigated substrate thicknesses is shown in *Fig. 7.* The bandwidth of the offset fed antenna elements for a given substrate thickness is always larger than the bandwidth of the crossed slot fed antenna elements.

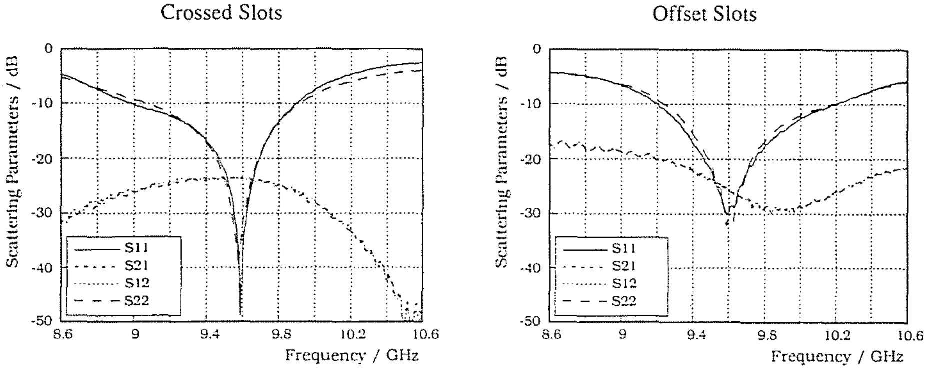

Fig. 6. Measured scattering parameters of dual polarised antenna elements (patch substrate: 1.6 mm Rohacell HF 51)

Due to this higher attainable bandwidth and the better coupling behaviour as well as due to the simpler and therefore cheaper mechanical structure only offset slots have been used for further investigations. With this offset slot design bandwidths of the dual polarised antenna elements up to 900 MHz (VSWR < 1.5) can be achieved. But the increase in bandwidth results also in an increasing coupling between both polarisations as can be seen in *Fig 8.*

3. Dual Polarised Aperture-Coupled 8x1 Element Subarray

Based on the obtained results for the antenna elements a dual polarised aperture-coupled 8 × 1 element subarray with offset slots for an active phased array antenna was designed.

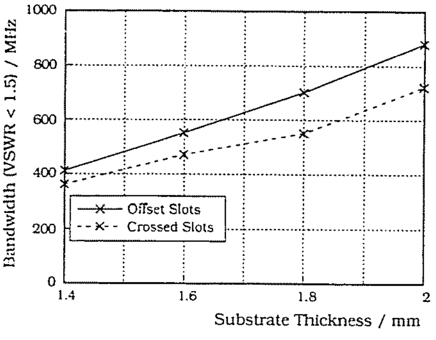

Fig. 7. Measured bandwidth for dual polarised antenna elements

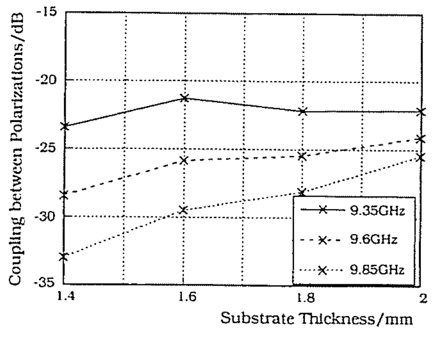

Fig. 8. Measured coupling for the offset slot antenna design

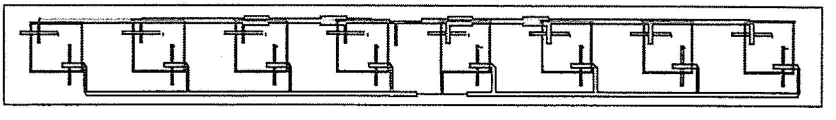

Fig. 9. Layout of an 8×1 element subarray with a series-parallel feed network

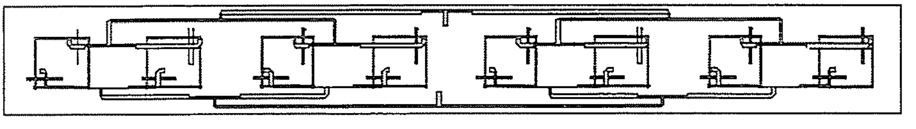

Fig. 10. Layout of an 8×1 element subarray with a corporate feed network

For the feed network either parallel or series-parallel feed arrangements have been investigated. The series-parallel feed network as shown in *Fig. 9* has the advantage of using short feed lines and has therefore low losses, but it is sensitive to frequency variations and substrate inhomogeneities. For broadband applications therefore a corporate feed network as shown in *Fig. 10* should be used. Due to the fact that for this feeding technique the line lengths from the subarray feed to all the subarray elements are equal, it is insensitive to frequency variations.

The measured scattering parameters for an 8×1 element subarray (patch substrate: 1.6 mm Rohacell 51 HF) with a corporate feed network

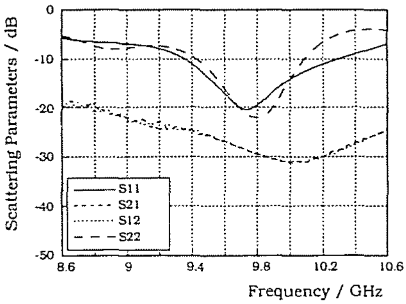

Fig. 11. Measured scattering parameters for the dual polarised 8×1 element subarray

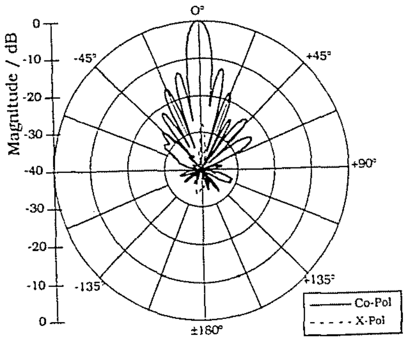

Fig. 12. Measured radiation pattern of the dual polarised 8×1 element subarray

are given in *Fig. 11.* The input reflection coefficients for both ports are better than −14 dB (VSWR < 1.5) over a bandwidth of 470 MHz and the isolation between both ports is less than −25 dB within this same range. The corresponding radiation pattern for horizontal polarisation at center frequency is given in *Fig. 12* for co- and cross-polarisation. Due to the use of aperture-coupling the cross-polarisation level is below −27 dB.

4. Conclusion

A detailed investigation of dual polarised aperture-coupled microstrip patch elements with crossed slots and offset slots was presented. The offset slot design offers the simpler mechanical and therefore cheaper structure as well as the better coupling behaviour and the higher attainable bandwidth.

Within a bandwidth of 470 MHz a realised 8x1 element subarray in X-band for a center frequency of 9.6 GHz shows a voltage standing wave ratio of less than 1.5 and a polarisation isolation of higher than 25 dB.

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