Improved Antenna Isolation in Transmit/Receive Applications

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***Short Abstract* — This paper shows how TX-RX-isolation can be significantly improved if two-port antennas with passive cross-talk cancellation are employed. An antenna structure composed of two half-wave spaced monopole antennas serves as an example to demonstrate how the frequency response of the residual antenna cross-talk can be adjusted to meet various system demands.**

Keywords* — *antennas; mutual coupling; duplexing; radio-repeater

I. INTRODUCTION

Systems with simultaneously operating transmitters and receivers require an excellent isolation between receive and transmit paths for preventing the transmit signal to cause saturation and intermodulation distortion in the sensitive receiver part. Furthermore, masking of the receive signal by spectral transmit signal components falling into the receive band has to be avoided. In case that frequency bands for transmit (TX) and receive (RX) are sufficiently wide separated in frequency, bandpass filters and/or filter duplexers are used to provide the required isolation.

On the other hand, there are applications e.g. radio repeaters where RX and TX bands have to be closely spaced (or even have to occupy the same band), where in principle the demanded isolation cannot be achieved with microwave filters or where isolation provided by filters is not sufficient. The following discussions are focused on this type of applications.

In case of a shared antenna nonreciprocal ferrite circulators are used to isolate RX and TX ports. Besides the limited isolation of circulators (< 35 dB) a very high return loss (matching) is required at the port of the shared antenna. Alternatively, two-port antennas with isolated ports can be used (self-duplexing antennas). High antenna-port isolation is well-known to be available from widely spaced antennas or – in case of narrow spacing- if they possess mutually orthogonal polarization (dual-polarized antennas, see e.g. [1]). However, in contrast to common belief high isolation between antenna ports can also be achieved for relatively closely spaced co-polar antennas.

From the most general point of view a two-port antenna can be considered as an arbitrary open structure with 2 ports. Both ports are associated with radiated far fields. Their angular dependence can (for a given frequency) be characterized by complex-valued vector radiation patterns (short port patterns)

$$\vec{C}_m(\Theta,\Phi) = \sqrt{g_m(\Theta,\Phi)}\exp\left(j\varphi_m(\Theta,\Phi)\right)\vec{p}_m(\Theta,\Phi) \qquad (1)$$

with $m = 1,2$ as the port number, g_m as the antenna gain function and φ_m as the phase function, both corresponding to port *m*. The phase functions φ_m are referenced for a common point (usually center of antenna structure). Polarization is characterized by the unit phasor $\vec{p}_m$ with $|\vec{p}_m| = 1$. For two-port antennas without dissipative losses and perfectly isolated ports the two port-patterns have to be mutually orthogonal, which means that

$$\oint \vec{C}_1^*(\Theta,\Phi)\cdot\vec{C}_2(\Theta,\Phi)\,\mathrm{d}\Omega = 0 \qquad (2)$$

holds. The integral in eq.(2) is carried out over the unit sphere and dΩ denotes the element of solid angle. The physical meaning of orthogonality in conjunction with a pair of port patterns can be explained by considering the totally radiated power in case of a two-port antenna where both ports are excited simultaneously. This totally radiated power can be compared with the radiated power values in case of separately excited ports. In general, the totally radiated power differs from the sum of the radiated power values for separate port excitation, but in the special case of mutually orthogonal patterns radiated power superposition holds. In case of dual-polarized antennas, eq.(2) is met due to mutually orthogonal polarization with

$$\vec{p}_1^*\cdot\vec{p}_2 = 0\,.$$

However, in case of port patterns with identical polarization eq.(2) is met by different angular dependence of the far field phases φ_1 and φ_2 (common reference point) and/or antenna gain functions g_1 and g_2. Based on this fact, it was shown in previous publications [2,3] that isolated ports can even be achieved if the total size of the antenna structure is significantly smaller than half a free space wavelength (e.g. $\lambda_0/10$. In case of these very compact antennas for diversity and MIMO applications, isolation is obtained by cost of a considerably reduced frequency bandwidth and/or efficiency [2]. In contrast to these previous publications, the present paper deals with two-port antennas with an element spacing on the order of about $\lambda_0/2$ and is focused on high antenna port isolation in a considerably wide frequency band for transmit/receive applications (self-duplexing two-port antenna).

II. COMBINATION WITH FILTER FUNCTION IN CASE OF NARROW-SPACED RX AND TX BANDS

Due to the physical origin of TX-RX-decoupling via isolated antenna ports it (of course) does not rely on the utilization of different non-overlapping frequency bands for TX and RX. Therefore, it could in principle even be used in case of simultaneous TX and RX operation in an identical frequency band (e.g. for on-frequency radio repeater). In this section a different situation is considered, namely the case where TX and RX occupy non-overlapping frequency bands, but the guard band between these bands is much to small to allow TX/RX isolation exclusively via microwave bandpass filters. By considering this special case, the combined effect of antenna port isolation and bandpass filtering can readily be discussed (see Fig. 1).

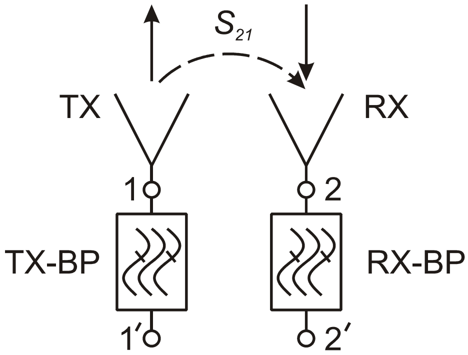

Figure 1. Transmit/receive operation in non-overlapping TX and RX frequency bands with transmitter bandpass TX-BP, receiver bandpass RX-BP and non-vanishing cross-talk (S21) between the antenna ports.

Fig. 2 schematically shows the situation one has to deal with if due to system demands the width of the guard band between TX and RX bands surpasses the selectivity properties available from realistic bandpass (BP) filters. Starting with the given TX and RX bands the insertion loss response $L_{TX}(f)$ and $L_{RX}(f)$ of TX-BP and RX-BP are depicted in Fig. 2. The path from TX-port $1'$ to RX-port $2'$ is given by the chain composed of TX-BP, (unwanted) transmission link between the antennas (cross-talk) and RX-BP. In a rigorous evaluation of the overall insertion loss L of this chain, multiple reflections at these chain components had to be taken into account. However, if the cross-talk attenuation $L_A/\text{dB}=-20\log|S_{21}|$ is assumed to exceed 10 dB, multiple reflection can be neglected. Thus, the combined filter insertion loss can be estimated by the sum $L_{TX}(f)+L_{RX}(f)$. The curve for this combined filter insertion loss in Fig. 2 (solid line) clearly indicates that due to finite filter selectivity and a relatively narrow guard band, the available attenuation can drop to unacceptable low values (22 dB in the example) at frequencies close to the inner edges of the bands. If an overall insertion loss L is demanded (70 dB in the example) the remaining attenuation $L-L_{TX}-L_{RX}$ has to be provided by the antenna isolation $L_A/\text{dB}=-20\log|S_{21}|$. The curve for this required cross-talk attenuation L_A is also shown in Fig. 2 and serves as an representative example for systems requirements.

III. BASIC PRINCIPLE OF CROSS-TALK COMPENSATION

A pair of co-polarized antennas with a spacing of about $\lambda_0/3$ to $2\lambda_0/3$ exhibits a cross-talk attenuation on the order of 10 to 20 dB. This level of isolation turns out to be not sufficient for the class of application addressed in section II.

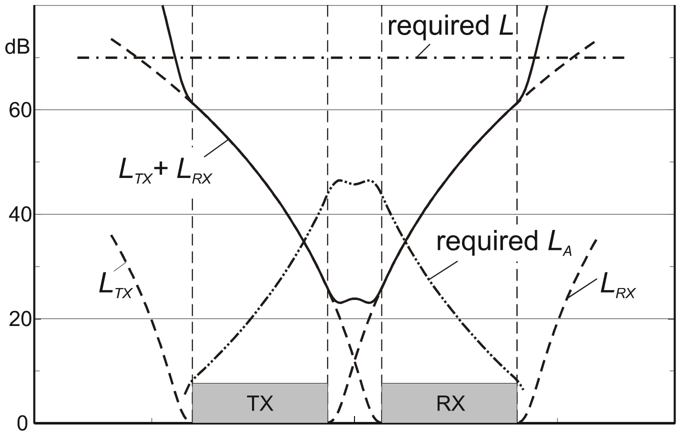

Figure 2. Determination of the required cross-talk attenuation $L_A/\text{dB}=-20\log|S_{21}|$ from TX and RX filter response L_{TX} and L_{RX} and the required overall TX-RX isolation L.

Isolation between the antenna ports can be enhanced by employing cancellation techniques. These techniques require a modification of the antenna structure such that in addition to the already existing "propagation path A" (Fig. 3, left) a second path B is created (Fig. 3, right) whose parameters can be adjusted to cancel the unwanted transmission via path A. If the paths are modeled via linear passive two-ports (lower part of Fig. 3) the perfect cancellation (perfect port isolation) requires that for shorted port 2 the two output currents are directly opposed, i.e. $I_{2,B}=-I_{2,A}$.

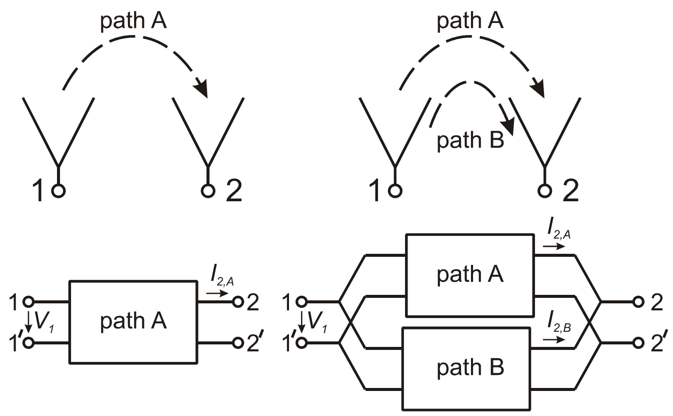

Figure 3. Left: Two-port antenna without cross-talk cancellation and corresponding model. Right: Cancellation of path A by a second path B and corresponding model.

There are different options for the realization of the second path. In [2,3] we have shown that an external passive (ideally loss-less) reciprocal and non-radiating decoupling-network can be placed between the antenna ports. However, the second path can also be created within the radiating structure [4]. It should be noted that in [5] a cancellation techniques ("feedforward isolation enhancement") is proposed for increasing the isolation in case of dual-polarized microstrip patch antennas. In contrast to our approach the output signals of the two paths are summed up with a resistive power combiner preventing the output signal of path A to flow back into path B. Since the principle of this alternative concept relies on components with dissipative loss, the simplicity of the circuit is achieved at the expense of an increased attenuation of the received signal.

In contrast to our previous papers on this subject special emphasis will be devoted to structures which allow variations in the frequency response of the cross-talk attenuation. In particular, it will be demonstrated that with path B being itself composed of multiple paths perfect isolation can be achieved at multiple frequencies.

IV. DESIGN EXAMPLE

A. *Antenna Structure*

The concept of cross-talk cancellation as described in section III can in principle be applied to any two-port antenna. A structure with an element spacing of about a half free space wavelength (λ_0) was chosen in order to demonstrate that high isolation (> 40 dB) in a sufficiently wide frequency regime can be obtained even if a stringent restriction for the spacing between TX element and RX element applies and, furthermore, both elements are using identical polarization.

A configuration composed of two 2.45-GHz monopoles with about half-wave spacing was chosen as a starting structure. It possesses a crosstalk $|S_{21}|/\mathrm{dB} = -L_A/\mathrm{dB}$ of about 16 dB. A "second path" (see Fig. 3, right) was realized by placing a microstrip structure in between the two monopole antennas (Fig. 4). The planar conductors are separated from the metallic ground plane by means of a 1 mm thick dielectric layer with $\varepsilon_r = 4$. In a first step the lateral strip width *w* was chosen small enough to ensure that no propagating mode except the quasi-TEM mode is supported. In order to keep the transmission coefficient of this second path sufficiently small (about – 16 dB) both ends of the transmission line were separated from the metallic rods of the monopoles by a gap of width *s*. Furthermore, another gap of width *g* is placed in the center (see Fig. 4). This microstrip structure acts as a second path (see Fig. 3, right) which together with the path due to mutual coupling between the two monopoles forms a two-path circuit. By a proper choice of the geometric parameters of the microstrip structure a transmission null (practically isolation > 50 dB) can be achieved at one prescribed frequency.

In a second step the width *w* of the microstrip sections was significantly increased in order to allow a dual-mode propagation along these line sections. For $w > \lambda_0 / 2\sqrt{\varepsilon_r}$ a propagating quasi-TE10 mode (with maximum electric field strength at the lateral edges) can be supported, but is not excited due to symmetry properties. For $w > \lambda_0 / \sqrt{\varepsilon_r}$ the quasi-TE20 mode begins to propagate. In this case the microstrip structure supports two different transmission paths between the monopole antennas, one path due to the quasi-TEM mode and the second path due to the quasi-TE20 mode. Together with the path due to mutual coupling between the two monopoles a three-path behavior occurs. In principle, a three-path structure allows cancellation and thus perfect transmission nulls (ideal port isolation) at two different frequencies which can be chosen independently from each other. This latter case, namely a structure with dual-mode operation was studied in details by means of numerical simulations (section B), followed by some experiments (section C).

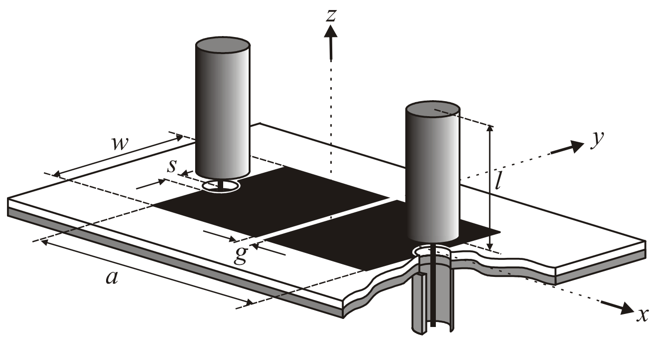

Figure 4. Antenna structure under consideration. Two coaxially fed monopole antennas with planar microstrip structure for cross-talk cancellation. Port 1 on the left side ($x<0$) of the structure.

B. *Numerical Simulations*

By means of numerical simulations (CST MICROWAVE STUDIO [6]) the properties of the antenna structure according to Fig. 4 with width *w* large enough to allow quasi-TEM and quasi-TE20 propagation were extensively studied. Fig. 5 depicts the frequency response of $|S_{21}|$ (cross-talk) for differently chosen geometric parameter of the microstrip structure (curves b, c and d) in comparison to the case without microstrip cancellation structure (curve a). Fig. 5 clearly indicates that the cross-talk level is significantly reduced in comparison to curve a) with about 16 dB isolation. Furthermore, as a novel result, it is seen that due to the 3-path structure minima occur at two different frequencies and the allocation of these frequencies can be varied to a large extend by means of changes in the geometrical parameters. This allows the frequency response of the residual cross-talk to be tailored to specific requirements. Fig. 2 can serve as an example for such a required response. Figs. 6 and 7 depict different cuts through the radiation pattern corresponding to the case that port 1 is fed and port 2 is terminated by a matched load. Additionally, the port pattern of port 2 is added to the azimutal pattern in Fig. 6. Data were taken for case d) in Fig. 5 and frequency at the second minimum of $|S_{21}|$. Due to the symmetry of the structure the patterns for ports 1 and 2 follow from each other by mirroring at the symmetry plane between both monopoles. Figs. 6 and 7 indicate that the magnitude pattern in the plane $\Theta = \pi/2$ deviates from a circle (about 3 dB variations). This is due to the fact that - even if port 2 is isolated from port 1 - the second monopole and also the planar microstructure contribute to the pattern.

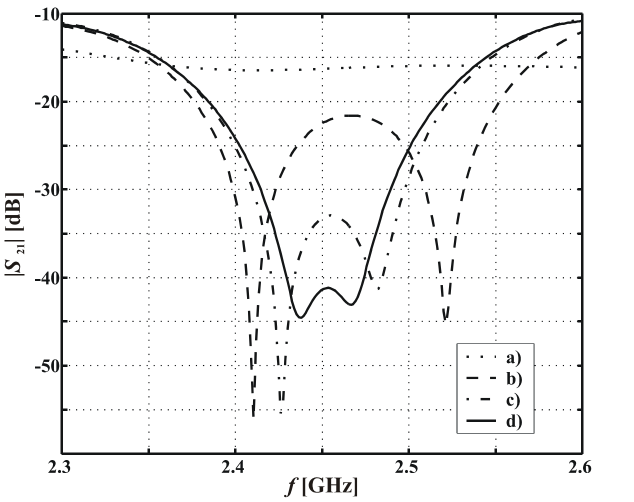

Figure 5. Frequency response of crosstalk for different choice of geometry parameter (curves b,c and d) of planar structure in Fig. 4 in comparison to the case without planar cancellation structure (curve a).

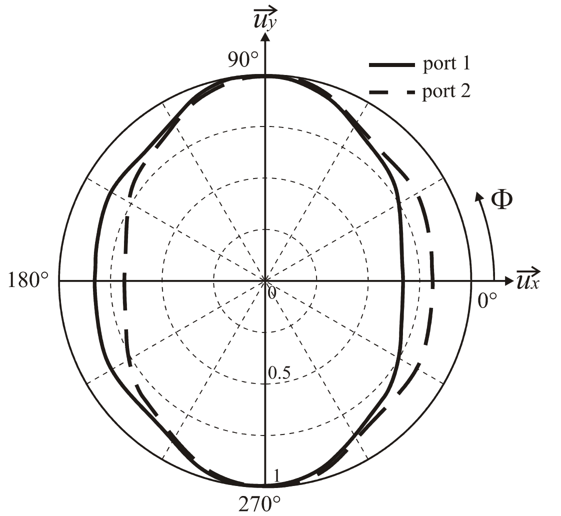

Figure 6. Azimutal magnitude pattern for port 1 and 2 (relative field strength E_Θ) at Θ = 90° (xy-plane).

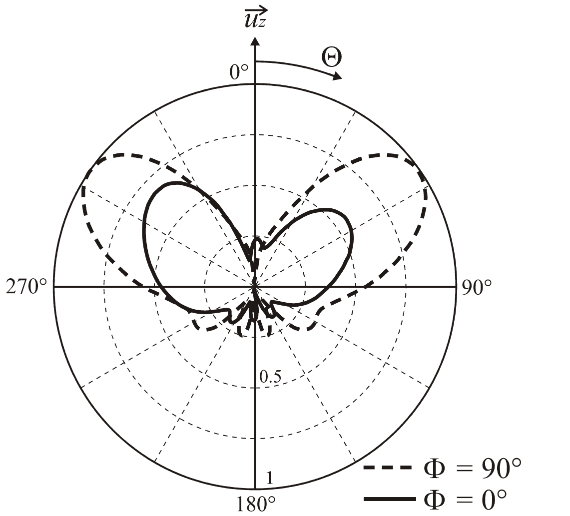

Figure 7. Elevation magnitude pattern of E_Θ for port 1 at Φ = 0 and Φ = 90°.

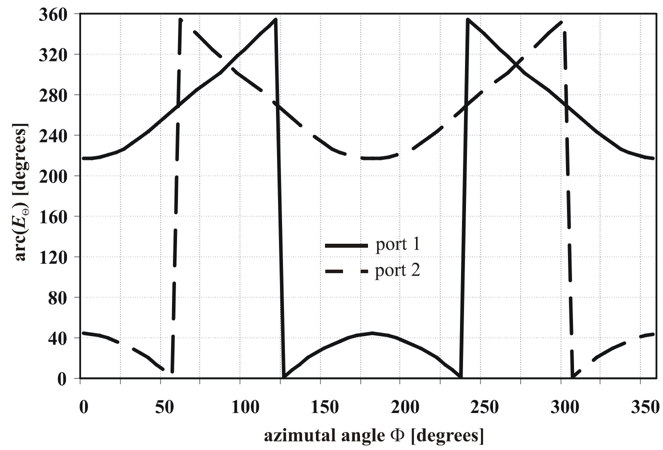

Figure 8. Azimutal angular dependence of the phase of the electrical field strength E_Θ for port 1 and 2 corresponding to Fig. 6.

Since the ports are isolated from each other the corresponding complex-valued port patterns of both ports are mutually orthogonal in the sense of (2). This is mainly due to the difference in the angular phase-variations of the far field for port 1 and 2, demonstrated with the azimutal angular dependence of the phase angle of the electrical field (Θ-component) in Fig. 8.

C. *Experimental Results*

In order to verify the results from computer simulations the antenna structures with differently chosen parameters of the microstrip structure were manufactured (see Fig. 8) and experimentally analyzed by means of S-parameter measurements with a network analyzer (Agilent E 8363A).

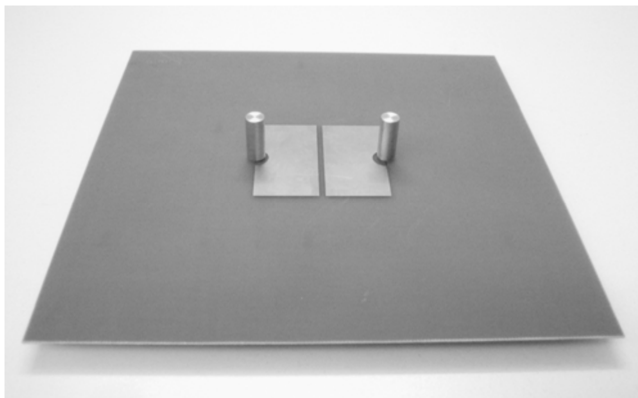

Figure 8. Photograph of the experimentally investigated antenna structure. Port 1 on the left side.

The result for one particular choice of these parameters (monopoles: diameter 9 mm, length l = 23.2 mm, distance a = 63.3 mm; microstrip structure: width w = 58.8 mm, gap width g = 2.9 mm) is shown in Fig. 9. The experimentally determined port isolation exceeds 42 dB and 30 dB at a fractional bandwidth of 1.2 % and 3.3 %, respectively.

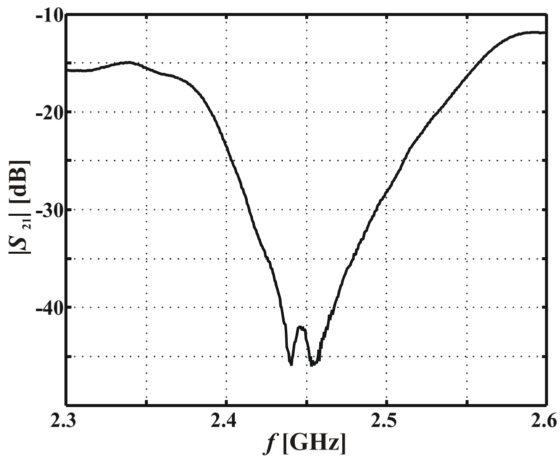

Figure 9. Experimental results for the crosstalk between ports 1 und 2 for antenna structure according to Fig. 8 (particular choice of geometric parameters). VSWR<2 for frequency range from 2.2 GHz to 2.8 GHz.

V. CONCLUSIONS

A pair of half-wave spaced monopole antennas was used as an example to demonstrate that high isolation between antenna ports can be achieved via a passive cross-talk cancellation technique. The proposed antenna structure provides multiple-path propagation (M paths) between the antenna ports which (theoretically) allows perfect isolation at M-1 frequency points. The presented approach can be exploited where due to a narrow frequency spacing filter properties are not sufficient for providing the demanded TX-RX isolation.

It was experimentally shown that in spite of the narrow element spacing an isolation of more than 45 dB could be achieved.

The usefulness of an even higher isolation between the antenna ports becomes questionable if the antenna system suffers from additional indirect coupling due to time-varying scattering from nearby objects.

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