

EVALUATION AND COMPENSATION OF MUTUAL COUPLING AND OTHER NON-IDEALITIES IN SMALL ANTENNA ARRAYS

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Preface

This thesis work was carried out at IDC, Radio Laboratory/SMARAD of Helsinki University of Technology (TKK) during 1999–2005. The work is based mostly on the project work in the finalized SYTE and BROCOM projects of the Institute of Digital Communications (IDC), which is an inter-laboratory organization at TKK. These project works have been financed by National Technology Agency TEKES and industry of Finnish mobile operators (Nokia, Radiolinja, Elisa and Sonera). The funding of the Radio Laboratory projects by the Academy of Finland was also important to complete the thesis work. The postgraduate work has been financially supported by Finnish foundations ‘HPY:n tutkimussäätiö’ and ‘EIS:n Säätiö’ (Foundation of the Finnish Society of Electronics Engineers). I am grateful for all the financial support I have received during the thesis work.

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I wish to thank all personnel I have worked with in the Radio Laboratory. Especially warm thanks deserve my co-authors in the publications included in this thesis: Dr. Anssi Toropainen, Pasi Suvikunnas and Dr. Jarmo Kivinen. I would also like to thank technicians Eino Kahra and Lauri Laakso for their help in research antenna manufacturing and Lorenz Schmuckli for the help in computer problems. Viktor Sibakov I like to thank for his support in the laboratory and companionship during the laboratory leisure activities.

Finally, and most of all, I want to thank my wife Irina and our children Kuisma and Leevi for their patience during the work.

Espoo, April 28, 2006

Ilkka Salonen

Abstract

Smart antenna technology is a challenging area in the development of wireless communications. Using smart antennas the quality of a radio link can be improved by many ways. Smart antennas are active antenna arrays or groups with changeable complex-valued weights at inputs and outputs. Good electrical matching of the array and the similarity and ideality of element patterns is usually expected. This dissertation focuses on the problems in the smart antenna arrays caused mainly by mutual coupling. Mutual coupling causes reflected power in the feeding system, input/output signal correlation and corruption of the element patterns. The arrays used in this thesis are small microstrip arrays. The used frequency is about 5.3 GHz. For several arrays the element patterns and scattering matrices are measured and used in calculations and measurements. Also simulated patterns and scattering matrices are used.

Due to mutual coupling the element patterns in an array are usually corrupted and therefore pattern correction should be used in smart antennas to improve the use of adaptive algorithms. In linear pattern correction the element patterns are reshaped using all antenna elements in the array. It is a computational method using a correction matrix between true and idealized inputs/outputs of array branches. For this pattern correction two basically different methods are used. The least squares error method can be used to find the correction matrix if the actual element patterns and the wanted element patterns are known, whereas in the scattering matrix method the correction matrix is defined only with the scattering matrix. These methods are compared in this thesis and the least squares error method is found to result in clearly better array patterns. The disadvantage of the scattering matrix method is that it does not compensate ground plate diffraction. However, the scattering matrix is easier to obtain than the element patterns and its use can give better understanding of the coupling mechanisms and therefore help the antenna design. Thus its use in pattern correction is examined more accurately. An extension of the least squares pattern correction method is done by correcting the array to a virtual array with different element spacing. The results show, that the element spacing in the virtual array should not differ significantly from the spacing in the real array.

In addition to the pattern correction with a correction matrix the use of the real patterns for beamforming is examined. In a modified least squares method for beamforming the weighting (cost function) is used. The beamforming with and without robust weighting is compared on the relative scale and the use of weighting give better results.

When antenna elements in an array are placed closer to each other, mutual coupling increases. At the same time the correlation between received signals increases. However, the signal correlation is usually caused by the signal propagation, and the effect of mutual coupling is minor. But, when signals arrive from many different directions, the pattern correlation caused by mutual coupling gives a realistic estimate of the signal correlation. The pattern correlation is a pure array characteristic and can be found easily. In this thesis the connection between pattern correlation and mutual coupling is examined. Equations are derived for this connection using scattering parameters or reflected power. These equations allow estimate mutual coupling from pattern correlation and vice versa, which is important for antenna array development. A more detailed formulation of the connection is done for lossless two-element arrays. In practice, when there are losses in the array, mutual coupling is not necessarily usable in estimation of pattern correlation.

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List of publications

- [P1] I. Salonen, A. Toropainen, and P. Vainikainen, "Linear pattern correction in a small microstrip antenna array," *IEEE Transactions on Antennas and Propagation*, vol. 52, no. 2, pp. 578–586, Feb. 2004.
- [P2] I. Salonen and P. Vainikainen, "Optimal virtual element patterns for adaptive arrays," *IEEE Transactions on Antennas and Propagation*, vol. 54, no. 1, pp. 204–210, Jan. 2006.
- [P3] I. Salonen, C. Icheln, and P. Vainikainen, "Beamforming with wide null sectors for realistic arrays using directional weighting," *Helsinki University of Technology Radio Laboratory Publications*, Report S 274, (ISBN 951-22-7973-8)(ISSN 1456-3835), Espoo, Finland, Dec. 2005, 24p.
- [P4] I. Salonen, C. Icheln, and P. Vainikainen, "Microstrip antenna circuit model and linear pattern correction," in *Proceedings of 11th International Symposium on Antenna Technology and Applied Electromagnetics (ANTEM 2005)*, Saint-Malo, France, June 15-17, 2005, 4 downsized pages on pp. 182–183.
- [P5] I. Salonen, P. Vainikainen, "Estimation of signal correlation in antenna arrays," in *Proceedings of JINA 2002 International Symposium on Antennas*, Nice, France, 2002, vol. 2, (ISSN 1168-3848), pp. 383–386.
- [P6] I. Salonen, C. Icheln, and P. Vainikainen, "Antenna array pattern correlation and input type," *Electronics Letters*, vol. 41, no. 10, pp. 575–577, May 2005.
- [P7] I. Salonen, C. Icheln, and P. Vainikainen, "Pattern correlation and mismatch in two-element antenna arrays," *Microwave and Optical Technology Letters*, vol. 48, no. 1, pp. 41–43, Jan. 2006.
- [P8] I. Salonen, C. Icheln, and P. Vainikainen, "The dependency of pattern correlation on mutual coupling and losses in antenna arrays," *Microwave and Optical Technology Letters*, vol. 47, no. 2, pp. 145–147, Oct. 2005.
- [P9] P. Suvikunnas, I. Salonen, J. Kivinen, and P. Vainikainen, "A novel MIMO antenna for laptop type device," in *Proceedings of Antenna Measurement Techniques Association (AMTA) 26th annual Meeting and Symposium*, Stone Mountain, Georgia, USA, Oct. 17-22 2004, pp. 118–123.

In publications [P1]-[P8], the author of the thesis did all the reported work and had the main responsibility for preparing the papers. Professor Pertti Vainikainen supervised all the papers. In [P1] Anssi Toropainen was supervising the work. In [P3], [P4], [P6], [P7] and [P8] Clemens Icheln undertook supervisory work.

In [P9] the antenna design concerning the basic idea of use, structure, antenna simulations, manufacturing, and antenna measurements was performed by the author of this thesis, while the use of simulated antenna patterns in MIMO capacity simulations with measured channel responses was carried out by Pasi Suvikunnas. The antenna design part in the paper is written by the author of this thesis. The channel simulation part and the general documentation were done by Pasi Suvikunnas.

Symbols

$\bar{\mathbf{a}}$	dimensionless array input voltage wave
C	capacitance
d	element spacing in array
\mathbf{D}	diagonal matrix containing scaling factors
f	frequency
f_n	element pattern of n^{th} element in array
\mathbf{F}	matrix of element patterns
g_n	array antenna element pattern with reference point in its own center
H	Hermitian (conjugate) transpose, \mathbf{A}^H is Hermitian transpose of matrix \mathbf{A} , the Hermitian transpose is often denoted also \mathbf{A}^{*T}
$\text{Im}(x)$	imaginary part of x
$\bar{\mathbf{I}}$	vector of currents
j	index of a vector or matrix or notation for the imaginary unit i
k	wave number
\mathbf{K}	correction matrix
L	inductance
P	power
\mathbf{P}	matrix of powers
$\mathbf{P}_{rad, 0}$	matrix of relative radiated power, radiated power rate matrix
\mathbf{P}_{refl}	reflected power matrix, matrix of relative reflected powers, reflected power rate matrix
r	pattern correlation used in the comparison of array patterns
r_{12}	pattern correlation in a two-element antenna array
\mathbf{R}	correlation matrix
$\text{Re}(x)$	real part of x
\mathbf{S}	scattering matrix
S_{ij}	component of scattering matrix
T	transpose, \mathbf{A}^T is a transpose of matrix \mathbf{A}
$\vec{\mathbf{V}}^+$	vector of input voltage waves of the array
$\vec{\mathbf{V}}^-$	vector of output voltage waves of the array
$\bar{\mathbf{x}}$	eigenvector
\mathbf{y}	normalized admittance matrix
\mathbf{Y}	admittance matrix
\mathbf{z}	normalized impedance matrix
Z_0	system impedance
\mathbf{Z}	impedance matrix
λ	wavelength
λ	eigenvalue
$\vec{\Psi}$	array pattern
φ	phase angle
ρ_e	envelope correlation
θ	direction angle (azimuth)
\otimes	component-wise multiplication
$*$	complex conjugate, c^* is the complex conjugate of c

Abbreviations

AoA	<u>a</u> ngle <u>o</u> f <u>a</u> rrival
DBF	<u>d</u> igital <u>b</u> eam <u>f</u> orming
DOA	<u>d</u> irection <u>o</u> f <u>a</u> rrival
LSE	<u>l</u> east <u>s</u> quares <u>e</u> rror
MIMO	<u>m</u> ultiple <u>i</u> n <u>m</u> ultiple <u>o</u> ut, a radio link between two smart antennas (adaptive arrays)
PDA	<u>p</u> ersonal <u>d</u> igital <u>a</u> ssistant, a handheld computer
SMA	<u>s</u> mall <u>m</u> icrowave <u>a</u> dapter, standard for RF cables and devices
SNR	<u>s</u> ignal to <u>n</u> oise <u>r</u> atio
TKK	Helsinki University of Technology, <u>T</u> eknillinen <u>k</u> orkeak <u>o</u> ulu.

1 Introduction

The effect of mutual coupling and other non-idealities in antenna arrays is examined in this thesis. Different methods to compensate for the effect of mutual coupling on the element patterns of an antenna array are examined. In addition the connection between mutual coupling and pattern correlation is examined in detail and is applied in simulations that are useful for smart antenna development.

1.1 Background

A smart antenna is typically an antenna array with a signal combining unit where the RF signals received or transmitted by the antenna elements can be combined in different manners. In the most general case the amplitude and phase (delay) can be arbitrary for each antenna element. Smart antenna technology is developing and it is expected to be in wide use in the near future. MIMO links between smart antenna arrays are nowadays under active development. The theoretical background is well established [1].

Mutual coupling in antenna arrays has been investigated and discussed several decades. It complicates the use of antenna arrays. It causes mismatch (reflected power), signal correlation, and corruption of the element patterns. Usually, mutual coupling is not taken into account perfectly and simplified patterns are used in algorithms. In traditional arrays with tens of elements under the effect of mutual coupling, only the elements near the array edges “see” the environment differently than other elements with neighboring elements at both sides. In small arrays with 2-8 elements each antenna element is in a different environment. The only realistic possibility for simplification of antenna patterns is in the symmetry of the array. One example for the usual simplification for antenna array is the Toeplitz structure (equal subdiagonal elements) of the correlation matrix [2], [3], [4], [5], [6], which is not accurate due to the differences between the antenna element patterns [7], [8], [9], [10], caused by the different surrounding for each element in the array.

The effect of mutual coupling on antenna patterns can be compensated partially with different methods. In [11] the least squares error (LSE) method is used to correct the mutual coupling effect in element patterns. Usually, in adaptive algorithms equal and ideal element patterns are expected [12]. The array correction/calibration gives a way to use these algorithms with better accuracy.

Mutual coupling and pattern correlation are connected to each other. The correlation of signals received by the antenna elements is caused both by the correlation of the incoming signals and by mutual coupling. When signal is coming from many different directions, signal correlation can be estimated with mutual coupling [13].

1.2 Objectives of the work

The objective of the thesis is to find countermeasures for non-idealities in the antenna array and to increase the understanding of the effect of mutual coupling. Pattern correction provides the possibility to improve the performance of the antenna array in adaptive use. The corrected arrays can be directly used in the adaptive algorithms, where ideal behavior of the element patterns is assumed. On the other hand, pattern correlation gives an estimate of signal correlation, e.g. if signals arrive

from many different directions. Scattering parameters can be easily measured. Thus the connection between mutual coupling and pattern correlation is interesting, when the correlation properties of an array need to be determined. Also the array matching is important and in this thesis it is shown, for example, that the pattern correlation defines the lowest limit for mutual coupling.

1.3 Contents of the thesis

This work consists of a summary of a work presented in papers [P1]-[P9]. Element patterns and scattering matrices are measured for microstrip arrays with different element spacings. The pattern correction is examined in papers [P1]-[P4]. In [P1] the basic LSE method is used and compared with the correction with the scattering matrix. In [P2] the LSE method is developed to find with an iteration procedure a corrected array (virtual array) with arbitrary element spacing and high similarity of the element patterns. In [P3] the measured, perturbed patterns are used in beamforming of an array power pattern and a weighting function is developed to find a LSE solution on the relative scale. In [P4] the scattering matrix correction method is used with an input circuit extraction procedure and compared with the use of the reference plane shift done in [P1]. In [P5]-[P8] the connection between reflected signal/power (scattering matrix, reflected power matrix) and pattern correlation (correlation matrix) is presented and used in different estimations, examining the connection between mismatch and pattern correlation. In [P9] an adaptive antenna group with four inputs is developed as a prototype of a smart antenna, which can be placed in a corner of a laptop-type device. The goal was to develop a wide band antenna group. In the development process the use of the connection between scattering matrix and pattern correlation was in a practical test; i.e. can the connection help the practical antenna development process.

The summary part is organized as follows: Chapter 2 contains basics of adaptive array, mutual coupling and pattern correlation and describes the basic equations for the connection between mutual coupling and pattern correlation. Chapter 3 considers pattern correction and beamforming with real element patterns. Pattern correction has two modifications; the use of measured patterns and the use of measured scattering matrix. In Chapter 4 the connection of mutual coupling and pattern correlation is analyzed with different examples and finally the development of a new multiantenna structure for mobile communications is described.

2 Adaptive array and mutual coupling

2.1 Adaptive arrays

Adaptive arrays are arrays with adjustable inputs or outputs [14]. In the most general case, the weights (amplitude and phase) of each antenna element can be set to any value. Traditionally adaptivity is used to combine the signals received with different antenna elements in an optimal manner to reach a better signal to noise ratio (SNR) [15], [16]. An adaptive array is one type of smart antenna [17]. Smart antennas enable signal processing. The usual algorithms for signal processing in a smart antenna are direction of arrival (DOA) detection algorithms (named also angle of arrival detection (AoA)) and digital beamforming (DBF). In adaptive beamforming a desired array pattern is formed with maximum radiation towards the signal of interest and nulls towards the signal not of interest [14], [15], [17]. In Fig. 1 we see an example of an array with changeable weights w_i . A new category of the use of smart antenna technology using adaptive array is a link between two multielement arrays, named a MIMO link [14], [18].

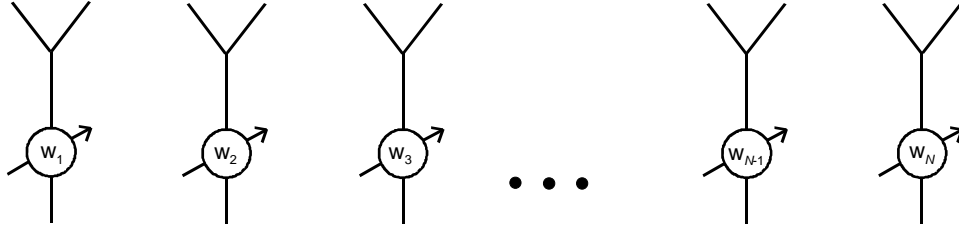


Figure 1. *Adaptive array with N antenna elements.*

Mutual coupling is a problem in antenna arrays. It causes, for small antenna arrays, element pattern distortion, so that the element patterns become different. Mutual coupling can be compensated with a matrix method [12]. Another problem caused by mutual coupling is the reflected power. Furthermore, mutual coupling causes correlation between signals received by different antenna elements.

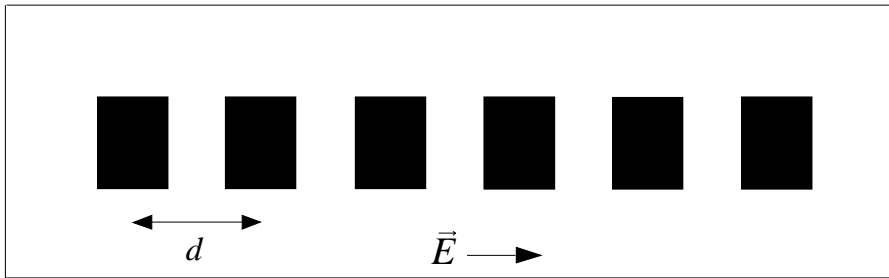


Figure 2. *A microstrip antenna array used in [P1]-[P6]. Element spacing d and orientation of the induced electric far field E are shown for the given microstrip element orientation. Metallization is black and the substrate is white. The backbone metallization is with the substrate dimensions.*

2.2 Experimental arrays and other antenna groups

One type of antenna array used in this work is microstrip array with six antenna elements. A principal explanation of the seven used microstrip arrays with different element spacings and antenna element orientations is in [7], [11] and [P1]. Microstrip arrays were used in [P1]-[P6]. In Fig. 2 is shown the general structure of a microstrip array. Another multiantenna system used in this thesis is a compact antenna group with two stacked dual-polarized antennas used in [P9]. Its principal structure is described in more detail in Chapter 4.4. For all used antenna arrays, the scattering matrix and the element patterns are measured. For the antenna group, the scattering matrix and the element patterns are measured and also simulated. The measurements of the antenna patterns were carried out in the large anechoic chamber of Radio Laboratory of TKK, for the main polarization in the frequency range around 5.3 GHz.

2.3 Mutual coupling

Mutual coupling in an antenna array causes the input signal at one array port appear at the other ports as reflected power and as apart of the output signal and in the case of reception the received signals are correlated.

2.3.1 Scattering, impedance, and admittance matrices

Because the input to an antenna port can be a voltage wave, a voltage or a current, there are three different definitions of mutual coupling [19]:

$$\vec{V}^- = \mathbf{S} \vec{V}^+ \quad (2.2.1)$$

$$\vec{V} = \mathbf{Z} \vec{I} \quad (2.2.2)$$

$$\vec{I} = \mathbf{Y} \vec{V} \quad (2.2.3)$$

Scattering matrix \mathbf{S} gives a linear relation between incoming voltage waves \vec{V}^+ and outgoing, reflecting voltage waves \vec{V}^- . The impedance matrix \mathbf{Z} and admittance matrix \mathbf{Y} give both linear relation between the port voltages and the port currents and they are inverse matrices of each other. The scattering matrix depends on the system impedances, which can be given as a diagonal matrix \mathbf{Z}_g . The system impedances are the impedances of the loads or generators connected to the antenna ports. In the measurements for this work the system impedances are always equal to the standard impedance $Z_0 = 50 \Omega$, and the system impedance matrix \mathbf{Z}_g is denoted as \mathbf{Z}_0 . In this case the scattering matrix, the impedance matrix and the admittance matrix are related with equations [19]

$$\mathbf{z} = (\mathbf{I} + \mathbf{S}) \cdot (\mathbf{I} - \mathbf{S})^{-1} \quad (2.2.4)$$

$$\mathbf{y} = Z_0 \mathbf{Y} = Z_0 \mathbf{Z}^{-1} = \mathbf{z}^{-1} \quad (2.2.5)$$

The matrices \mathbf{z} and \mathbf{y} are the relative impedance and admittance matrices. An important matrix related to the scattering matrix is the matrix of reflected power

$$\mathbf{P}_{refl,0} = \mathbf{S}^H \mathbf{S}. \quad (2.2.6)$$

Matrix of reflected power is a dimensionless Hermitian matrix and it has been named in this thesis and in the included papers also unusually as reflected power rate matrix and the matrix of relative reflected powers. The eigenvalues λ_s and the eigenvectors $\bar{\mathbf{x}}_s$ of the scattering matrix are defined by

$$\mathbf{S}\bar{\mathbf{x}}_s = \lambda_s \bar{\mathbf{x}}_s, \quad (2.2.7)$$

where the eigenvalues are complex numbers. The corresponding eigenvalues λ_p of the relative reflected power matrix $\mathbf{S}^H \mathbf{S}$ are defined

$$\mathbf{S}^H \mathbf{S} \bar{\mathbf{x}}_p = \lambda_p \bar{\mathbf{x}}_p, \quad (2.2.8)$$

where the eigenvalues have real positive values equal to the relative reflected power, the amount of the reflected power, when using the corresponding eigenvector as the array input. The basic property of the eigenvectors of the relative reflected power matrix eigenvalues is their orthogonality [7]

$$\bar{\mathbf{x}}_{p,i}^H \bar{\mathbf{x}}_{p,j} = \sigma_{ij}, \quad (2.2.9)$$

where σ_{ij} is the Kronecker symbol equal to unity if $i = j$ and otherwise equal to 0. It is important to note, that for a lossless array the corresponding eigenbeams or –patterns are also mutually orthogonal [7]. The mean relative reflected power $\langle P_{refl} / P_{in} \rangle$ can be calculated

$$P \langle P_{refl} / P_{in} \rangle = \langle \lambda_p \rangle \quad (2.2.10)$$

or

$$\langle P_{refl} / P_{in} \rangle = \frac{1}{N_{el}} \sum_{i=1}^{N_{el}} |S_{ij}|^2, \quad (2.2.11)$$

but only when there is no preferred input at the array ports [7], [P7]. This is not the usual case. This is the case of the general adaptive use of the array when the mean relative reflected power does not depend on any special propagation situation and can be easily calculated and used as an array characteristic.

2.4 Pattern correlation in an antenna group and mutual coupling

Pattern correlation in the antenna array or in the antenna group is a quantitative characteristic of the correlation between antenna elements. Pattern correlation gives the signal correlation in the case when signals are arriving from many different directions. Basically, the signal arrival directions should cover the antenna element patterns [20]. For the future MIMO antenna links the multipath propagation caused by a rich scattering environment is a precondition for high data rates [1]. Thus, there is an interest in pattern correlation for diversity and MIMO applications [21].

The pattern correlation matrix \mathbf{R} is defined as

$$\mathbf{R} = \mathbf{F}_0 \mathbf{F}_0^H, \quad (2.3.1)$$

where \mathbf{F}_0 is the matrix of the normalized element patterns. In the case of the used six-element microstrip arrays, \mathbf{R} is a 6x6-element matrix with unity values in its diagonal and \mathbf{F}_0 contains discretized, vectorized element patterns in its rows. Each element pattern is defined with an input voltage wave present at one array port, when the other ports are terminated with matched loads. These are so called active element patterns, element patterns in the array environment, corrupted due to mutual coupling [22], [23], [24], [25], [26]. Each row $\vec{\mathbf{f}}_{0,i}$, $i = 1$ to N_{el} , in the matrix \mathbf{F}_0 is normalized to unity norm

$$\vec{\mathbf{f}}_{0,i} = \frac{\vec{\mathbf{f}}_i}{\sqrt{\vec{\mathbf{f}}_i^H \vec{\mathbf{f}}_i}}, \quad (2.3.2)$$

where the row vectors $\vec{\mathbf{f}}_i$ are the rows of the measured pattern matrix \mathbf{F} . In the case of the used six-element microstrip arrays \mathbf{F} is a 6x359 matrix with a 1° increment in the direction angle. With completely defined array element patterns (all directions, both polarizations) the radiated power matrix is defined as

$$\mathbf{P}_{rad} = \mathbf{F} \mathbf{F}^H. \quad (2.3.3)$$

The matrix \mathbf{F} can be scaled with the input power, and then \mathbf{P}_{rad} changes to relative radiated power matrix $\mathbf{P}_{rad,0}$. In the general case the relation between input power, radiated power and lost power can be presented in matrix form [P7]

$$\mathbf{I} - \mathbf{P}_{refl,0} = \mathbf{I} - \mathbf{S}_L^H \mathbf{S}_L = \mathbf{P}_{diss,0} + \mathbf{P}_{rad,0} = \mathbf{L}^H \mathbf{L} + \mathbf{F}_L \mathbf{F}_L^H, \quad (2.3.4)$$

where $\mathbf{P}_{refl,0}$, $\mathbf{P}_{diss,0}$ and $\mathbf{P}_{rad,0}$ are the dimensionless matrices of reflected power, dissipated power and radiated power, respectively. The subscript L means losses. In certain cases the losses can be ignored and we get the lossless case

$$\mathbf{I} - \mathbf{P}_{refl,0} = \mathbf{I} - \mathbf{S}^H \mathbf{S} = \mathbf{P}_{rad,0} = \mathbf{F} \mathbf{F}^H. \quad (2.3.5)$$

The connection between mutual coupling and pattern correlation is presented in matrix form in [7] and in [P5]

$$\mathbf{R}_{pat} = \mathbf{D}^{-1} (\mathbf{I} - \mathbf{S}^H \mathbf{S}) \mathbf{D}^{-1} = \mathbf{D}^{-1} \mathbf{F}^H \mathbf{F} \mathbf{D}^{-1} = \mathbf{F}_0 \mathbf{F}_0^H, \quad (2.3.6)$$

where the diagonal matrix \mathbf{D} contains the square roots of the diagonals of matrix $\mathbf{F} \mathbf{F}^H$ or $\mathbf{I} - \mathbf{S}^H \mathbf{S}$. For a two-element array with reciprocity we can write using (2.3.6) the dependency of pattern correlation r_{12} on scattering parameters

$$r_{12} = -\frac{S_{11}^* S_{12} + S_{12}^* S_{22}}{\sqrt{1 - |S_{11}|^2 - |S_{12}|^2} \cdot \sqrt{1 - |S_{22}|^2 - |S_{12}|^2}} = r_{21}^* . \quad (2.3.7)$$

This is the basic equation for the connection between pattern correlation and scattering parameters in lossless two-element arrays derived in [7] and presented in [P5]. In [P5] a symmetric antenna pair is considered. Equation (2.3.7) is presented also in [27] for the envelope correlation ρ_e , for which we can write

$$\rho_e \approx |r_{12}|^2 . \quad (2.3.8)$$

Equation (2.3.8) is based on the well-known relation between signal envelope correlation and cross-correlation [28], which are time-dependent. According to [13], [29] it has some requirements of ideality for the antennas, which are fulfilled, for example, for a small dipole in free space. Also the signal should arrive from many directions in the antenna beam region. Equation (2.3.5) for power balance between input power, reflected power and radiated power seems trivial. However, it was derived in [7]. The rescaling of the radiation patterns to unity norm done in (2.3.2) giving pattern correlation matrix with unity autocorrelation components on the diagonals in (2.3.7) is trivial. Often the mathematical expression of correlation contains also the mean value [13]. In the case of arbitrary phase the mean complex value is zero and can be omitted [7], [30]. Equation (2.3.7) is used in [P5] and [P7] where pattern correlation in two-element arrays is considered. In [P5] and [P6] Equation (2.3.6) is used with measured data for microstrip arrays. In [P6] also the corresponding equations of pattern correlation for voltage-driven and current driven arrays are given.

3 Mutual coupling compensation and beamforming with real patterns

The array element patterns can be corrected computationally before they are used in signal processing algorithms [12]. The accurate placement of array pattern nulls is important. The basic method of pattern correction used in this thesis is the LSE method presented in [P1], [31], [32], [33], [34], [35], [36]. It requires the measured element patterns. The scattering matrix method on the other hand does not require measured element patterns and is thus interesting and also examined in this thesis. In the third presented method of virtual element patterns the idea is to find array element patterns that are as similar as possible. Those are searched using the LSE method and iteration without any predetermined desired element patterns, but only the measured patterns. The beamforming with measured element patterns is presented here together with the pattern correction, because the element patterns are typically perturbed by mutual coupling. Usually, when mutual coupling is not taken into account, idealized patterns are used in beamforming algorithms and in this case with real patterns we can view the correction be included in the beamforming.

3.1 Mutual coupling compensation with the LSE method

In the LSE method of the linear pattern correction that correction matrix is searched, which gives the minimum squared error between wanted and corrected array patterns. In the ideal case the correction is exact

$$\mathbf{F}_{wanted} = \mathbf{K} \mathbf{F}_{meas} . \quad (3.1.1)$$

The measured element patterns \mathbf{F}_{meas} used in this thesis are the so called active element patterns [22], [23], [24], [25], [26], measured for each element in the array, when the others are terminated with matched loads. In practice, solution can be found only approximately. Therefore, with the pseudoinverse LSE method we get

$$\mathbf{F}_{wanted} \approx \mathbf{F}_{corrected} = \left[\mathbf{F}_{wanted} \mathbf{F}_{meas}^H \left(\mathbf{F}_{meas} \mathbf{F}_{meas}^H \right)^{-1} \right] \mathbf{F}_{meas} = \mathbf{K}_{LSE} \mathbf{F}_{meas} . \quad (3.1.2)$$

The pattern correction undertaken using (3.1.2) is examined in [P1] for seven different array configurations with varying spacings and element orientations. The results show, that after linear pattern correction the array patterns are very close to ideal array patterns: They are smooth and with exact positions of the nulls. In the case of greatest element spacing the correction is less optimal due to more pronounced ground plate edge diffraction. In other cases with smaller substrate plate and especially with stronger coupling the correction results are very good.

3.2 Virtual element pattern iteration in mutual coupling compensation

When we apply pattern correction, the usual goal is to get identical element patterns [12]. If there is no other requirements for the corrected element patterns, the method of virtual elements can be used. In this method the searched ideal and similar element patterns can be presented as

$$f_n^{id} = g(\theta, d_{virt}) \cdot e^{-j\left(\frac{N_{el}-n}{2}\right)kd_{virt}\sin(\theta)}, n = 1, N_{el} \quad (3.2.1)$$

where $g(\theta)$ is the common element pattern, also called the element factor [37], identical for all elements and the exponent term is the array factor for element number n in an array with N_{el} elements. Array factor depends on the displacement of the element from the array center. Unfortunately the element factors in a real array differ from each other due to mutual coupling [12], [38], and also the corrected element patterns differ from each other [P1]. An optimal solution is searched using iteration. The virtual spacing d_{virt} is not necessarily the same as the real spacing in the array. With a fixed virtual distance d_{virt} the element factor $g(\theta)$ is the centralized element pattern, where the element is placed at the center of the virtual array. For a given $g(\theta)$ with fixed d_{virt} the idealized, wanted element patterns $f_n^{id}(\theta)$ can be calculated using (3.2.1) and the corresponding correction matrix can be found with (3.1.2). The wanted element factor $g(\theta)$ and the corresponding wanted element patterns $f_n^{id}(\theta)$ change during iteration. The current element factor is in each iteration cycle the complex-valued mean $\langle g_n(\theta) \rangle$ of the element factors of the last corrected element patterns. For element patterns $f_n(\theta)$ the corresponding different $g_n(\theta)$:s can be found with (3.2.1). According to [P2] the iteration converges well and leads finally to a good agreement between the final centralized corrected element patterns $g_n(\theta)$. For an iteration round $i \rightarrow i+1$ we can write

$$\begin{aligned} f_n(\theta, i)_{n=1, N_{el}} &\xrightarrow{(3.2.1)} g_n(\theta, i)_{n=1, N_{el}} \longrightarrow g(\theta, i+1) = \left\langle g_n(\theta, i)_{n=1, N_{el}} \right\rangle \xrightarrow{(3.2.1)} \\ f_n^{id}(\theta, i+1)_{n=1, N_{el}} &\xrightarrow{(3.1.2)} f_n(\theta, i+1)_{n=1, N_{el}} \end{aligned} \quad (3.2.2)$$

It is shown in [P2] that good correction results can be achieved even in cases when the virtual element spacing differs significantly from the real one. However, a decreased virtual spacing changes the element patterns to more directive ones so, that in an array scan the beam is not moving as quickly as expected by the direction angle of the array factor. Another point with practical importance mentioned in [P2] is that the little larger metallic ground plate can allow an increase in the virtual array spacing. This is interesting for conformal arrays mounted on devices with a metallic cover; a virtual array with greater spacing can have better resolution.

3.3 Beamforming with a real array

The practical application of beamforming is examined with a real array in [P3]. The real array is an uncorrected array with measured element patterns. The beamforming is performed for an array with the typical spacing of about 0.5λ . In the beamforming an array pattern

$$\bar{\Psi} = \bar{\mathbf{a}}^T \mathbf{F}_{meas} \quad (3.3.1)$$

is generated. The problem is to find the corresponding input coefficients $\bar{\mathbf{a}}$. If the desired complex valued array pattern $\bar{\Psi}_{desired}$ is known, then the LSE solution can be found with the pseudoinverse of the measured array pattern matrix [P3], [39], [40]

$$\bar{\mathbf{a}}_{opt}^T = \bar{\Psi}_{desired} \left\{ \mathbf{F}_{meas}^H [\mathbf{F}_{meas} \mathbf{F}_{meas}^H]^{-1} \right\}. \quad (3.3.2)$$

The different directions are not always equally important in array patterns and thus a direction-dependent cost function, directional weighting, is used. In [P3] the optimal input coefficients in the case of the directional weighting are found with

$$\bar{\mathbf{a}}_{opt}^T = \left(\bar{\Psi}_w \otimes \bar{\Psi}_{desired} \right) \left\{ \left(\mathbf{F}_w \otimes \mathbf{F}_{meas} \right)^H \left[\left(\mathbf{F}_w \otimes \mathbf{F}_{meas} \right) \left(\mathbf{F}_w \otimes \mathbf{F}_{meas} \right)^H \right]^{-1} \right\}, \quad (3.3.3)$$

where the direction-dependent weighting function $\bar{\Psi}_w$ is a cost function, a real-valued vector with the same number of components as the pattern vectors, and called the weighting pattern in [P3]. Matrix \mathbf{F}_w contains in all its rows the same weighting pattern $\bar{\Psi}_w$. Equation (3.3.3) used in [P3] is in accordance with the more usual power weighting in [41]. Equation (3.3.3) written with amplitude weighting shows simply, that the method is the same LSE, as in (3.3.2). In [P3] the inverse of the wanted amplitude pattern of the array is used as the weighting pattern. In this case of weighting the LSE method results in complex-valued array pattern with the LSE error in phase direction and relative error in the amplitude direction.

Often only the amplitude value of the array pattern, the power pattern, is of interest. In this case the best solution is searched among array patterns with different phase patterns. Iteration is used in [P3] to find the optimum amplitude pattern. In the iteration, the phases are allowed to change freely. A basic case with box-type array amplitude patterns is examined in [P3] focusing mainly in wide null generation. Preliminary work on box-type array pattern generation was published in [42] and [43]. Wide nulls are important when interfering signals are arriving from a sector, which is typical case in multipath propagation scenario.

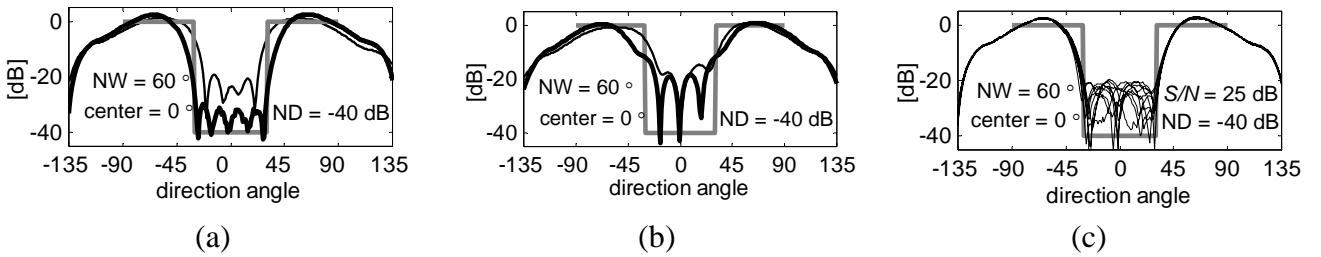


Fig. 3. In a) is presented array patterns with a wide null using input coefficients obtained for a realistic array with measured element patterns. In (b) array patterns with the same input coefficients are presented for an array with idealized element patterns. In c) are array patterns perturbed with noise at the antenna ports. The desired array pattern is denoted by thick gray solid line and the array pattern obtained using weighting is denoted in a) and b) by thick black solid line. By a thin black solid line in a) and b) is denoted the case without weighting and in c) the cases with array pattern perturbations caused by Gaussian noise added independently to antenna port, when the unperturbed input coefficients are found using weighting.

The information on perturbations and uncertainties in the element patterns and input port values is important for beamforming. In Fig. 3a) we see an example, when accurate measured element patterns are used. The width of the wide null is 60° and its depth is -40 dB. That kind of deep wide null is possible for the given array only in the forward direction. In Fig. 3b) the perturbations in

element patterns caused by mutual coupling are not taken into account and in Fig. 3c) the effect of noise at the antenna ports is presented. The noise level is taken to give perturbations in the array pattern of about the same magnitude as mutual coupling. Fig. 3 demonstrates well the need of pattern correction or the use of realistic element patterns. Fig. 3c) shows that the desired wide null level should be not lower than the noise level, as pointed out in [P3]. As well the other perturbations caused by near field effects affect the choice of the desired null depth.

3.3.1 Comparison of beamforming with weighting and without it

The weighting method is compared with finding the input coefficients without weighting in [P3]. The wanted array amplitude patterns are box-type patterns with wide nulls used for interference cancellation of widely located interferers. The basic difference between array pattern generation with weighting using the inverses of the wanted amplitude pattern values and the method without weighting is that weighting gives a solution with minimum relative (i.e. in dB's) amplitude error, while the usual LSE method gives the minimum error of amplitude on the linear scale.

In [P3] the methods are compared in the relative scale. It is not surprising, that the method with weighting is in general better on dB scale. In a wide region of array pattern parameters (null width, depth and position) it behaves better than the pure LSE solution. However, in some extreme cases it is not better: the solution, even though it is the best in the LSE sense, it is not usable in interference reduction. Thus a modified LSE criterion in relative scale is used in [P3] for the comparison of the final results. It takes into account the high and low levels in the radiation patterns separately. If the fitting is bad for the high or low region, the result is not good, even if the fitting is good on the other region. This can happen for example when a null or a beam is close to the limit of the calculation area and also when there is a very high narrow beam (which means that there is a very wide null as well). In the case of extreme depths of null or sidelobe level the lack of the limitation of the relative scale for the values near to zero is one explanation for the instability of the method with weighting. Some modifications to the weighting function in respect to the extreme cases are presented in [P3] and they show that the method can be developed further with more sensitive weighting.

3.4 Mutual coupling compensation using scattering matrix

The method using the scattering matrix in pattern correction is evaluated in [12]. The same correction method can be found in [44] using the impedance matrix presentation of mutual coupling, and also as a minor result in [45], where the element pattern in open circuit condition is defined as the original, unperturbed pattern. In [P1] the use of the scattering matrix in pattern correction is examined by comparing it with the basic LSE method discussed in Chapter 3.1. The scattering matrix is easier to be obtained than the measured element patterns and therefore the information on its validity in pattern correction is important. The correction in [P1] is based on the finding of the suitable reference plane for scattering matrix. In [P4] the extraction of the equivalent input circuit was examined in the scattering matrix-based correction.

3.4.1 The correction matrix in the scattering matrix-based correction

For a voltage driven array the radiation is defined with input voltages and in the case of mutual coupling compensation the input voltage waves are manipulated to give the wanted input voltages. The scattering matrix method of pattern correction is used in [P1] and [P4]. The used correction matrix is

$$\mathbf{K}_{S,V} = (\mathbf{I} + \mathbf{S})^{-1}, \quad (3.4.1)$$

where subscript S refers to pattern correction using scattering matrix, V means that the array is a voltage driven array, for which the array input voltages defines the radiation and are thus the correct input type. For a current driven array we can write respectively

$$\mathbf{K}_{S,I} = (\mathbf{I} - \mathbf{S})^{-1}, \quad (3.4.2)$$

where the subscribe I denotes a current-driven array, as for example a dipole array, for which the current feed is the correct input, which defines the radiation of the single element. When the wanted feeds for the ideal antenna elements are multiplied with the correction matrix we get the required input voltage wave vector.

The corrections with inverses of $\mathbf{I} + \mathbf{S}$ and $\mathbf{I} - \mathbf{S}$ are the same as the corrections with $(\mathbf{I} + \mathbf{y})/2$ and $(\mathbf{I} + \mathbf{z})/2$, respectively [7], [46]. Papers using correction with $\mathbf{I} + \mathbf{z}$ are for example [44], [46], and [47]. In the widely cited case of correction in [44] the correction is with $-\mathbf{S}$. The case of correction with $+\mathbf{S}$ is used in [44]. The simple array correction with the impedance matrix has been shown to improve the signal-to-noise-ratio in [44], [47] and the MIMO link capacity in [48].

3.4.2 Reference plane adjustment

The reference plane adjustment is important in the scattering matrix method of pattern correction. In (3.4.2) we see, that the correction matrix changes, when the scattering matrix phases change with changing reference plane. As well change the impedance and admittance matrices. In [P1] the scattering matrix was measured with the reference plane adjusted to the free side of the SMA connector of the antenna element. Further it was moved to the antenna side of the SMA connector, to the lower end of the feed probe of the patch antenna. In Fig. 4 we see the measured resonators (S_{jj} :s)

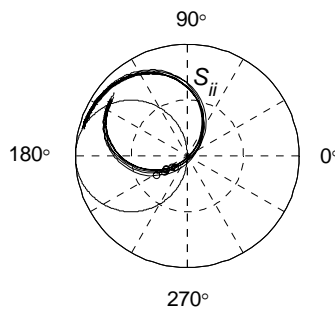


Figure 4. *Measured S_{jj} :s of 6-element microstrip array with element spacing of 0.4λ . The perfect circle is for ideal parallel resonator with constant capacitance and inductance and with a resistance matched to 50Ω .*

for the array with spacing $d = 0.4\lambda$ and the ideal admittance circle when the reference plane is in the lower end of the feed probe. The phase change with reference plane change is frequency dependent. According to [P1] it should be moved further 23° in the direction of the antenna at the center frequency, which gives a 45° counterclockwise rotation on the Smith diagram (see errata in this

thesis). This reference plane shift moves the diagonal elements to the location of an ideal admittance resonator. It is important to note, that this reference plane adjustment gives good results for a number of arrays as examined in [P1] and the shift of 23° can be used for correction of the microstrip arrays with the same feed structure. In [P1] only array patterns were shown. In Fig. 5 we see how the correction using scattering matrix done in [P1] affects the amplitude patterns of the array elements. It makes the element patterns smoother. As well it corrects the phase perturbations (not shown).

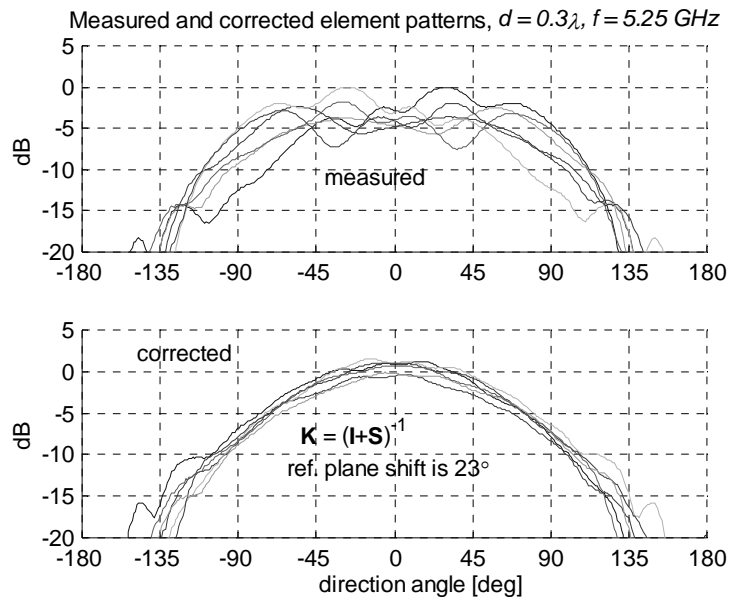


Figure 5. *The effect of the pattern correction on power patterns of the antenna elements using scattering matrix correction method with the reference plane shift done in [P1] in the case of the strongest coupling.*

3.4.3 Input circuit extraction

According to [P1] the same reference plane shift of 23° applied computationally in the reference plane adjustment can be done also with an inductance of about 0.7 nH. In the case of an input circuit, which in the simplest case can be the probe inductance, the pattern correction should be defined after the extraction of the (equivalent) input circuit [P4]. To extract the input circuit its impedance should be subtracted from the diagonals of the impedance matrix and its admittance from the diagonals of the admittance matrix [P4]. This is similar to working on the Smith diagram in the single-antenna case. The transformations between different matrices were presented in Equations (2.2.4) and (2.2.5).

When the input circuit is extracted and the wanted input, the vector of the array input voltages, is defined on the antenna side of the input circuit, the reference point for the correction should be moved again to the generator side of the input circuit. For microstrip antennas the wanted input at the antenna side is a voltage vector, because microstrip antenna elements can best be modeled as an admittance resonator [49], [50], [51], [52]. The voltages at both sides of an admittance component are the same whereas the currents on both sides of an impedance component are the same. Thus only transformations between currents and voltages, i.e. impedance or admittance matrices, are needed when the reference point is moved to the generator side of the input circuit.

The simplicity of the final resonances and smooth frequency dependency of mutual impedances and admittances are used as criteria to find the input circuit in [P4]. Corresponding scattering parameters are always sensitive to resonant behavior and not used. In Fig. 6 we see the uncorrected, measured mutual impedance z_{16} and mutual admittance y_{16} in one examined array. We see that the mutual admittance is more corrupted than the mutual impedance and it indicates that the impedances dominate in the input circuit.

Paper [P4] is basically the same as [53], but with a table of most suitable component values of the input circuit with one or two components. The components of the input circuit considered in [P4] are presented in Fig. 7 and the table of the values of the lumped elements presented in [P4] is in Table I. In the case of the input circuit in Fig. 7 with reference points a, b, and c the corresponding correction matrix, giving the needed input voltage waves or generator voltages is

$$\mathbf{K} = (\mathbf{I} + \mathbf{S}_a)^{-1} \mathbf{z}_b \mathbf{y}_c = (\mathbf{I} + \mathbf{S}_a)^{-1} (\mathbf{y}_a - \mathbf{y}_{ab}^{in})^{-1} [(\mathbf{y}_a - \mathbf{y}_{ab}^{in})^{-1} - \mathbf{z}_{bc}^{in}]^{-1}, \quad (3.4.3)$$

where diagonal matrices \mathbf{y}_{ab}^{in} and \mathbf{z}_{bc}^{in} contain on the diagonals the input circuit admittances and impedances, respectively. Extraction of input inductance L_2 moves the resonant circle (see Fig. 4) with counterclockwise rotation to the position of the ideal resonator and removes the peak in the mutual admittance, for example, the peak in $|y_{16}|$ in Fig. 6. Extraction of input capacitance C_2 rotates the resonant circle clockwise and removes an additional peak at about 3.8 GHz caused with the extraction of L_2 (not shown). The same additional peak at about 3.8 GHz appears also with the reference plane shift of 23° considered in Section 3.4.2. The parallel capacitance C_1 in the input circuit removes partially the peak in mutual impedance as, for example, the peak in $|z_{16}|$ in Fig. 6. In the case of an admittance resonator the effect of resonance (at about 5.2 GHz) is seen in mutual impedances, which explains that the peak in $|z_{16}|$ is not fully removed. In principle this method could be used to find the input circuit for any radiator merely based on measured scattering parameters. The order of admittances and impedances in input circuit is arbitrary, if their values are low compared with the antenna radiation admittance or impedance.

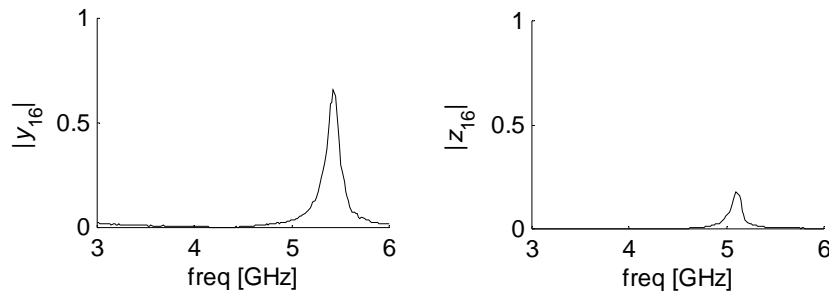


Figure 6. Mutual components $|y_{16}|$ and $|z_{16}|$ in the 6-element array with $d = 0.4\lambda$ before input circuit extraction.

Because the electrical length of the probe is not small, there is a distributed impedance and thus the lumped circuit is not exact when using only few elements [P4]. Also the complexity and the uncertainty problem of this method are noticed in [P4]. One problem with uncertainty is the size of the final more ideal resonant circle on the Smith diagram. Even in the case of two lumped circuit components the correction method with input circuit extraction is complicated and gave only a small increase in correction compared with the use of the reference plane shift [P4]. However, the input

circuit extraction helps to find better universal correction of element patterns, and also to understand better the electromagnetic behavior of the antenna under research.

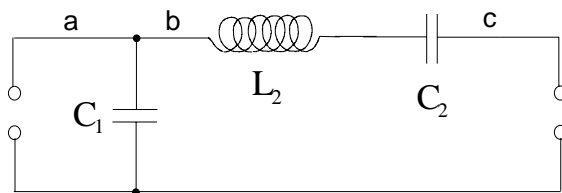


Figure 7. Components of the input circuit.

Table I Input circuit component values

	C_1	L_2	C_2
I	–	0.60 nH	–
II	–	0.79 nH	5 pF
III	0.5 pF	0.50 nH	–

Measurements were done in the frequency band from 3 GHz to 6 GHz. Increasing the upper limit would give more information on mutual components decreasing the uncertainty. Taking into account more electric model components in addition to the components related to the feed structure can complicate the situation further. For example, the end capacitance of the microstrip radiator can have an effect as well on the radiation, coupling and equivalent input circuits, and they can be included in the equivalent circuit of the antenna also with partially inductive character due to the distance from the reference point. The ground plate edge diffraction is obviously even more complicated to model with lumped elements and is thus not possible in practice.

According to [12] the good matching of the antenna element is a criterion for the simplicity of the use of the scattering matrix method. This is true for simple antennas and feeds, but even if antenna element is self-matched, then in the case of matching circuit or discontinuities the situation is complicated for the correction. In the case of complicated antenna feed structure this is a serious disadvantage of the scattering matrix method of correction.

3.4.4 Comparison of mutual coupling compensation with LSE method and based on scattering matrix

The comparison of mutual coupling compensation with the LSE method and with the scattering matrix method is done in [P1]. The array correction with the impedance matrix has been noted to improve the signal-to-noise-ratio in [44], [47] and the MIMO link capacity in [48]. In [P1] the effect of correction only on array patterns was studied. In the case of pronounced coupling the effect of correction using the scattering matrix presented in Fig. 5 is visually clear. However, as noted in [P1] this method of correction gives much poorer results than the LSE method even in this case. According to [P1] the difference in correction capability is due to the fact, that the scattering matrix method does not correct the edge diffraction. The role of the edge diffraction is greater in arrays with a large metallic ground plate. The advantage of the scattering matrix method is that it does not need measured element patterns, i.e. scattering matrix measurements are simpler to perform, as they do not require the use of a large anechoic chamber.

3.4.5 Correction over the frequency band

The effect of the correction over the frequency band is examined in [P1] and [P2]. The correction can be done for each frequency independently. This is not so convenient as to have a common correction matrix across the whole frequency band. The only correction method, which needs measurements over the frequency band is the scattering matrix method, if the needed reference plane shift or the input circuit to be extracted are not known, even when the correction itself is used at a single frequency. The frequency dependency of the scattering matrix is as easy to obtain as the

scattering matrix at the certain frequency, when the measurements are carried out with a vector analyzer. Over the frequency band only the capability of the LSE correction method was considered in the publications of this thesis.

In paper [P1] the correction matrix found for the center frequency defined using an isotropic desired array was tested over the whole frequency band. The correction for the array pattern with a forward beam became poorer for frequencies far from the center frequency, but was very satisfactory over the frequency band from 5.0 to 5.4 GHz, for arrays with spacings $d = 0.3\lambda$ and $d = 0.5\lambda$. In [P2] the correction using the virtual array method described in Section 3.3.2 was tested with a common correction matrix over the same frequency band for an array with virtual spacing equal to the real spacing of $d = 0.5\lambda$ and the results were good over the frequency band.

Comparing the results presented in [P1] and [P2] we can find that using the common correction matrix over the frequency band, the correction with the virtual array method with optimized desired pattern done in [P2], is better, than when the desired element pattern is fixed to isotropic, as in [P1]. In Fig. 8 we see the comparison of these two cases of correction as the correlation of the forward

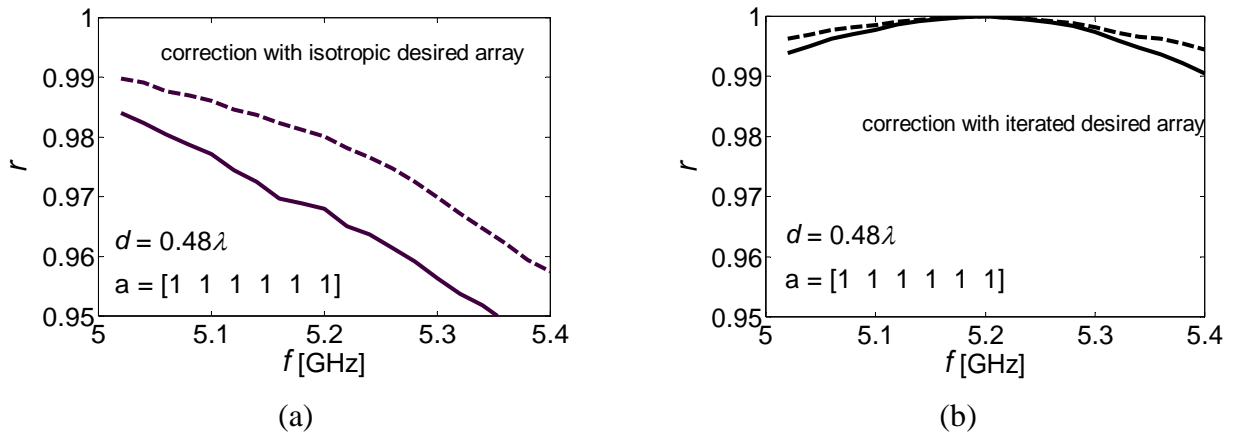


Figure 8. *Correction with predetermined desired element pattern and iterated desired element pattern over a frequency band. The input is $[1 \ 1 \ 1 \ 1 \ 1 \ 1]$. The generated array patterns are compared with the desired array pattern at the center frequency. The correction is evaluated by calculating the pattern correlation r of the uncorrected array pattern and corrected array pattern with the desired array pattern of the corresponding case. The uncorrected array is with solid line and the corrected array is with dashed line. The element spacing in this array is $d = 0.5\lambda$.*

beam with the wanted beam at the center frequency when the beam is with input $[1 \ 1 \ 1 \ 1 \ 1 \ 1]$. The effect of beam scan was examined in the case of an iterated wanted element pattern. During a 30° scan of the beam (not shown) the correlation with desired iterated patterns is more than 0.998 only over frequency band 5.15 GHz ... 5.25 GHz, but is still not less than 0.99 at the highest frequency of 5.4 GHz [P2]. For the case with iterated desired element patterns the uncorrected case for the forward beam correlates also well with the final desired array pattern and is a consequence of increased directivity and the average character of the corresponding desired element pattern, which is, unfortunately, not known previously. One important point is that the correlations of array patterns like in Fig. 8 are often very close to one. In [P1] even in the case when the uncorrected patterns are heavily corrupted the correlation is higher than 0.9.

3.4.6 Multiport model for mutual coupling

For a two-element array, impedance and admittance matrices can be presented with a two-port electrical circuit with elements, which can be calculated simply with impedance or admittance matrix values [P4]. For multielement arrays only the equivalent circuit based on admittance matrix components is possible [P4]. However, this fundamental limitation does not mean that the correction with the impedance matrix for arrays with current inputs could make no sense or that it could give worse results than the correction with the admittance matrix for arrays with voltage inputs. It means only, that we cannot find any multiport circuit with electric parameters describing exactly the impedance matrix of a multiport array.

3.4.7 The effect of mutual coupling compensation on the signal processing

The pattern correction with the scattering matrix (or mutual coupling compensation) provides a method to find the perfect feed, which is needed at the array input. According to [P4] this method removes the effect that internal reflections have on the element patterns. One might think, that it removes also mutual coupling, but this is not possible, because mutual coupling is a phenomenon caused by the proximity of the antenna elements. The mismatch remain and the correlation at least partly also after the pattern correction. Thus the array matching circuit is of interest [54], [55]. Also the stronger correlation between smooth and more identical corrected element patterns can be expected, when the elements are close to each other. However, the question whether the voltages or the currents are the less corrupted signals at the array input is important [P4], [P6]. From this point of view we can say, that mutual coupling compensation can in practice *remove the effect of additional internal mismatch* on the signal correlation within the antenna array caused by wrong input type.

4 Mutual coupling and element pattern correlation in antenna arrays

In this chapter mutual coupling and pattern correlation are considered in order to help the design of a compact antenna array or group. The connection between mutual coupling and pattern correlation is examined in papers [P5]-[P8]. In [P9] a compact multiantenna group with low coupling and low correlation between antenna elements was presented.

4.1 Experimental and simulated values of mutual coupling

The scattering matrix was measured for the different arrays [7], [11]. The absolute value of the scattering parameter S_{ij} between neighboring elements is in the examined arrays from -35 dB to -8 dB, when the self-matching, the diagonal element of the scattering matrix S_{jj} is from -35 dB to -15 dB at resonant frequencies from 5.14 GHz to 5.36 GHz. The resonant frequencies of the different arrays vary due to mutual coupling. Using Equation (2.2.8) the minimum, mean and maximum values of the relative reflected power are defined at the center frequency in [7] for the examined arrays. The square roots of the mean values of the relative reflected power calculated with the scattering matrix eigenvalues using Equation (2.2.7) are presented in [56] for different arrays. The mean relative reflected power is for all arrays lower than 10% with an exception of the array with the closest spacing of 0.3λ for which the mean value of the relative reflected power is 35% and the maximum value more than 90%. For the other arrays the maximum reflected powers are 4%-30% of input power at the resonant frequency. The coupling is lowest for the array with spacing of 0.7λ . The worst case with the highest eigenvalue of the relative reflected power matrix is somewhat exotic: the experiments with a beam scans in [7] and [11] show that during beam scans the worst cases do not occur. The maximum value for the power reflection from one port exceeds the mean of total reflected power for all the examined arrays [7].

4.2 Experimental and simulated values of pattern correlation

The pattern correlations are calculated in [7], [P5], [P9] based on the measured/simulated scattering parameters. A value less than 0.2 is typical for the examined arrays and groups. This means, that the corresponding estimate for envelope correlation, when a signal is arriving from many directions, is very low, $\rho_{env} = 0.2^2 = 0.04 \ll 0.7$, where $\rho_{env} \ll 0.7$ is a widely used criterion for the use of an antenna pair in diversity systems [13]. A stronger criterion $\rho_{env} \ll 0.5$ commonly used for a mobile terminal in mobile communications [13] is also well fulfilled.

4.3 Connection between mutual coupling and pattern correlation

In papers [P5]-[P8] the connection between mutual coupling and pattern correlation presented in equations (2.3.6) and (2.3.7) is used to estimate pattern correlation in an antenna pair and in antenna arrays. In recent years the interest has increased in the connection between mutual coupling and pattern correlation due to development of multiantenna systems [21]. The traditional way to correct the element patterns of a dipole antenna array is to find how to feed one element while the other elements are open circuited, without feed currents [44]. This can be done with all generator voltages or input wave voltages [P4]. The input voltage vector in an array depends only with a certain phase factor on the reference plane position and the input voltage waves are with a certain delay factor the

same as the generator voltages [P4], [P6]. The input voltages in the reference plane of the scattering matrix can be expressed with the corresponding input voltage waves

$$\vec{V} = (\mathbf{I} + \mathbf{S})\vec{V}^+. \quad (4.3.1)$$

When only one component of the voltage input vector \vec{V} is different from zero, we get the radiation of only that element, i.e. the radiation pattern of that element in the case of voltage feed. Because for a constant \vec{V} the corresponding voltage wave vector \vec{V}^+ depends on the reference plane of the scattering matrix, as we can see in Equation (4.3.1), we get that the element pattern for a certain voltage input is also dependent on the reference plane. Thus it is not surprising that the pattern correlation depends as well on the reference plane for voltage and current feeds [57], [P6]. For the scattering matrix presentation with voltage wave feeds the pattern correlation, presented in Equation (2.3.6) does not depend on the reference plane, because the complex conjugation in $\mathbf{S}^H \mathbf{S}$ removes the effect of a phase factor.

The dependency of pattern correlation on input type and reference plane is examined in the case of microstrip arrays in [P6]. In the impedance and admittance matrix presentations the dependency of pattern correlation on the reference plane has a clear minimum and the difference between minimum and maximum values is also significant. For arrays with strong coupling the minimum pattern correlation was less than the pattern correlation in the scattering matrix presentation. If the reference plane is adjusted to the minimum correlation we get the pattern correction to minimum pattern correlation. To correct the element patterns at the same time to regular ones and with low correlation, would be of interest [44]. The required reference plane shift varies for different arrays, and, unfortunately, it is not the same as required for the pattern shape correction, presented in Chapter 3.4.2.

In the other papers [P5], [P7], [P8], [P9] in this thesis the scattering matrix presentation with voltage wave inputs is used. It is the most natural presentation because antenna input voltage wave vectors are the same as the generator or load voltage vectors [P6].

4.3.1 Two-element array

In two-element lossless arrays the connection between mutual coupling and pattern correlation is given by Equation (2.3.7). An estimation of pattern correlation and mutual coupling is presented in [P5] for a symmetric antenna pair. In the case of antenna pair symmetry the pattern correlation is

$$r_{12} = \frac{2|S_{11}||S_{12}|\cos\varphi_{diff}}{1-|S_{11}|^2-|S_{12}|^2}, \quad (4.3.2)$$

where φ_{diff} is the difference of the phase angles of S_{11} and S_{12} . In [P5] the cosine is taken to be one and the dependency of envelope correlation on scattering parameter $|S_{12}|$ with fixed $|S_{11}|$ in the case of many paths is shown. In [P7] the optimum case, where the matching is the best for a given correlation (i.e. $S_{11} = \pm S_{12} = S_{22}$)[58], is presented. The connection between mutual coupling and pattern correlation allows us also to estimate coupling in a two-element array. For a lossless two-element array the minimum of the mean relative reflected power can be calculated when the patterns and the difference in gains are known [P7]. Asymmetry in the antenna pair can increase the

asymmetry in the relative reflected power matrix, which increases the minimum value of the mean relative reflected power.

4.3.2 Multi-element array

In multi-element arrays the connection between mutual coupling and pattern correlation is given by Equation (2.3.6). In [P5] the pattern correlation values are presented for arrays with different spacings calculated, on the other hand, using scattering matrix, and on the other hand measured element patterns with a two-dimensional horizontal plane approximation. The conclusion in [P5] is that these two correlations are so close to each other that the pattern correlation calculated with scattering parameters can approximate well the signal correlation also when signals are arriving only in the horizontal plane from many directions. A tendency to have true pattern correlations somewhat lower than those predicted with idealized patterns is detected in [7], [59] and [60]. It can be compared also with results showing that mutual coupling can decrease the signal correlation [29], [61], [62]. One explanation is, that the rippled structure of element patterns caused by mutual coupling leads to lower correlation than expected, i.e. when the element patterns remain unchanged [62], [63], [64].

4.3.3 Effect of element spacing in arrays

For small array design, there is an interest in the dependency of pattern correlation and mutual coupling on the element spacing. The pattern correlation is easy to calculate if the patterns are known. The pattern correlation is known as a function of element distance for different arrays and antenna pairs [7], [54], [55], [60], [65], [66], [67], [68], [69], [70], [71]. The connection between mutual coupling and pattern correlation in [13] allows obtain more information on pattern correlation, when simulating the dependency on element spacing. The effect of termination is presented in [29], [30], [54], [55], [57], [68], [72], [73]. The effect of input type and reference plane shift is examined in [P6] for arrays with different spacings. In [67] the dependency of pattern correlation on scattering parameters in Equation (2.3.7) is used for simulation of pattern correlation in a pair of monopoles for element spacings from 0 to 2λ . In [P7] the connection between pattern correlation and mutual coupling is used to estimate the dependency of the reflected power on the element spacing for different antenna pairs with different fixed element patterns, assuming that they are not changed when the elements are moved closer to each other. It is obvious that coupling can be low if the beams are directed away from each other. In [P7] is demonstrated that the coupling can be lowered also when beams are directed to the same direction, when the beams are different, as for example, with different widths.

4.3.4 Lower limits for correlation and coupling

The connection between mutual coupling and pattern correlation gives an opportunity to define the maximum level of pattern correlation for a lossless antenna pair when the absolute values of the scattering parameters are known. The minimum possible value of pattern correlation is always zero. As well the connection gives the opportunity to give a minimum level of mismatch, when antenna pattern correlation is known. In [P7] the limiting case is presented. For array with known element patterns (fixed pattern correlation) the limiting value helps to characterize the optimality of the array matching. Unfortunately, in the case of (strong) mutual coupling any matching network affects the patterns [P5], which should be taken into account, when matching circuits are used [54], [55], [57], [74], [75]. While only the mean of relative reflected power and the absolute values of the scattering

parameters are estimated from patterns in the general case in [P7], it is shown in the specific case of an array with small aperture elements that also the complex scattering parameters can be estimated [24].

4.3.5 Angle of arrival spread

When an array is used in a traditional scheme of diversity with a narrow angle of arrival the element spacing is several wavelengths [13], [76] and mutual coupling has no effect. The effect of the spread of the angle of arrival for signal correlation is presented in [77]. With increasing angular spread of the incoming signals their correlations become closer to the pattern correlation. In [P5] the allowed level of mutual coupling is simulated for a symmetric two-element array ($S_{11} = S_{22}$) with horizontal uniform patterns when the angular spread of the incoming signal is given. The criterion $\rho_{env} \leq 0.7$ is used. We can find, for example, that for 30° uniform distribution of spread of arrival angle the scattering parameter $|S_{12}|$ should be less than -12 dB if $|S_{11}| = -6$ dB. The best matching with the same level of correlation is obtained, however, for the case, when $S_{11} = S_{12}$ [P8], and to have a more optimal matching $|S_{11}|$ and $|S_{22}|$ could be better and $|S_{12}|$ worse. The effect of the antenna element directivity on the allowed mismatch level is obvious. In addition the allowed mismatch level depends also on the direction of arrival [P5]. Thus this method to find the allowed mismatch level should be used for each array and propagation environment separately, without generalizing.

4.3.6 Array with losses

When power dissipation in the antenna array is taken into account, Equation (2.3.4) needs to be used for power balance. When the relative radiated power matrix is multiplied left and right with a normalizing diagonal matrix the pattern correlation matrix diagonal elements of unity results. In the case with losses the pattern correlation is [P8]:

$$\mathbf{R}_{pat,L} = \mathbf{D}_L^{-1}(\mathbf{I} - \mathbf{S}_L^H \mathbf{S}_L - \mathbf{L}^H \mathbf{L}) \mathbf{D}_L^{-1} = \mathbf{D}_L^{-1} \mathbf{P}_{rad,L} \mathbf{D}_L^{-1} = \mathbf{D}_L^{-1} \mathbf{F}_L^H \mathbf{F}_L \mathbf{D}_L^{-1} = \mathbf{F}_{L,0}^H \mathbf{F}_{L,0}, \quad (4.3.3)$$

where subscript L denotes losses and the diagonal normalizing matrix \mathbf{D}_L contains the roots of the diagonals of the radiated power matrix $\mathbf{F}_L^H \mathbf{F}_L$. In two-dimensional case we can write for the square of pattern correlation amplitude

$$|r_{12}|^2 = \frac{\left| S_{11}^* S_{12} + S_{12}^* S_{22} + \{\mathbf{L}^H \mathbf{L}\}_{12} \right|^2}{\left((1 - |S_{11}| - |S_{22}| - \{\mathbf{L}^H \mathbf{L}\}_{11}) (1 - |S_{11}| - |S_{22}| - \{\mathbf{L}^H \mathbf{L}\}_{22}) \right)} = \frac{\left| S_{11}^* S_{12} + S_{12}^* S_{22} + \{\mathbf{L}^H \mathbf{L}\}_{12} \right|^2}{\eta_1 \eta_2}. \quad (4.3.4)$$

The sign of $\{\mathbf{L}^H \mathbf{L}\}_{12}$ is in discrepancy with that in [78]. Parameters η_1 and η_2 are the radiation efficiencies of antenna 1 and 2 [78], [79]. They are radiated powers related to the generator power, when the other antenna is with matched load. Their square roots are the components of the corresponding diagonal scaling matrix \mathbf{D}_L . They give possibility to write the equations more compactly. However, in the case of losses their values are difficult to obtain comparing with the scattering parameters. There is also a discrepancy with the definition of the radiation efficiency between [78], [79] and that in [80], where the efficiency is related to the coupled power. Mutual component of losses seems to be very difficult to obtain, if it can't be neglected as done in [78]. The

lack of exact information on losses leads to uncertainty problem in pattern correlation calculations [P8], [78], [79], [80].

One important difference between the cases with and without losses is that when there are losses in the array, the ideal matching with a scattering matrix with only zero components does not lead to absence of the pattern correlation [P8]. In [P8] is simulated a case with an antenna pair, when losses are present. The effect of the reflected power rate matrix and matrix of relative losses is similar to the pattern correlation matrix and if the mismatch and the losses are of equal strength, the maximum pattern correlation can be up to twice the pattern correlation in the case without losses. This is important for handheld devices, for which there are usually considerable near-field losses. An interesting possibility is noticed in [P8]: the off-diagonal elements both of the reflected power rate matrix and the matrix of losses can compensate each other resulting in zero cross-correlation between the elements. However, by the robust simulation for a two-element array, it is not the usual case [P8].

4.3.7 Mutual coupling and array performance in MIMO systems

Mutual coupling between antenna elements affect the signal correlation. For MIMO systems a decrease of link capacity due to mutual coupling is detected in [64], [69], [81]. On the other hand, some contrary observations are reported in [62], [82], [83] and [84]. According to [69] the total effect of decreased correlation caused by mutual coupling and increased mismatch is the lower MIMO capacity. In [82], [85] higher capacity caused by mutual coupling is noted for cases with close spacing, when compared to the case without mutual coupling. In [83] the increase of capacity is noted when the element spacing is reduced while increasing at the same time mutual coupling. The tendency to have lower pattern correlation than expected with decreasing element spacing is reported in [7], [85], but without taking into account the changes in element patterns. The different contradictory observations on the effect of mutual coupling can be explained with the character of the dependency between pattern correlation and mutual coupling. Pattern correlation can have different values with a fixed level of mismatch and be even zero with mutual coupling present [7], [P5]. Thus the increase of mutual coupling does not always mean higher pattern correlation, which in MIMO systems with several paths is about the same as the signal correlation. The decorrelation effect of increased pattern distortions (beam direction changes and ripples) can be one practical explanation in cases, when the correlation increase is less than expected, when the antenna element spacing in an array is decreased [7], [29], [64].

4.4 Antenna configuration for MIMO terminals

A new antenna group was developed for a compact multielement smart antenna to be mounted on a laptop type device. The goal was to have a wideband antenna group with low coupling and with two polarizations. The basic solution is a two-layer microstrip patch element structure with two ports for different polarizations (see Fig. 9). The two-layer structure has been used to widen the frequency band [86], [87], [88], [89], [90]. Two identical dual-polarized antennas were stacked in such a way that they radiate to opposite directions [P9]. They were planned to be mounted on different sides of a device corner and having in total four connectors only on two edges of the antenna structure. Further the element patterns were used in simulations with measured channel data to find the performance of the developed structure; it was compared with usual dipoles in MIMO channel simulations with a good performance in [P9] and [91], [92]. The compactness and inexpensiveness of the microstrip structure are advantages.

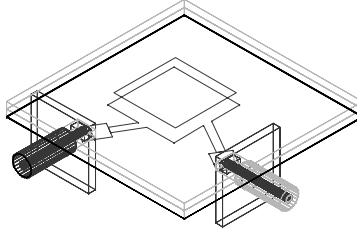


Figure 9. *Dual-polarized patch antenna examined in [P9].*

In the development process of the antenna group in [P9] the low pattern correlation was easier to realize than the low mutual coupling. Because the group is a prototype for laboratory experiments it has SMA connectors and thus skew microstrip lines were used at the antenna feeds to give place for the connectors at the opposite sides of the model of the laptop. The skewed feed lines decreased the orthogonality of the dual-polarized structure and it was compensated by optimizing the dimensions and locations of the patches. Widening the bandwidth of the dual-polarized antenna structure is more complicated than in the single-polarized case examined for example in [93]. The main problem is to keep the coupling low between the ports over the whole frequency band.

In the development process the goal was to minimize the mean reflected power over a wide frequency band. The mean reflected power is usually not characterized. Instead usually the scattering parameters $|S_{11}|$, $|S_{12}|$, and $|S_{22}|$ are characterized, as for example in [94]. In this thesis it is shown in [P9], that these parameters give an exact value for the mean reflected power but do not provide any information on the worst case, i.e. the maximum reflected power. In Fig. 10a) and Fig. 10b) we see the frequency behavior of the scattering parameters compared to that of the reflected power rate eigenvalues. It is very easy to see, that in this case the scattering parameter does not give a holistic view on the mismatch.

For multielement arrays the worst case is the eigenvector with the greatest eigenvalue of the reflected power rate matrix $\mathbf{S}^H \mathbf{S}$ at the array input. For multielement groups the largest eigenvalue of the reflected power rate matrix should be known to find the worst case. On the other hand, in the case of two-element arrays the largest value can not exceed the mean value by more than 3 dB, which can be used in practical validation of the two-antenna structures worst matching using Equation (2.2.11) with absolute values of the scattering parameters. In a critical case with matching difficulties, the eigenvalue analysis is important, as seen in Fig. 10. Because in the development process of a compact group a low value of the pattern correlation can be easily obtained, the emphasis should be in the enhancement of the matching.

In the development process of wide band antennas the resonant behavior on the Smith diagram is of interest [93]. Mutual coupling corrupts the resonant behavior of the scattering matrix eigenvalues. In [7] the scattering matrix eigenvalues presented on Smith chart were close to ideal resonances even in the case of strong coupling, when the reference plane was adjusted for pattern correction as described in Chapter 3. The reference plane shift might help to find more regular resonances, which could help in the development process of the compact antenna group. However, the feed structure in the case of the adaptive group is more complicated and a more complicated input circuit extracting work might be required to reach the ideal resonators.

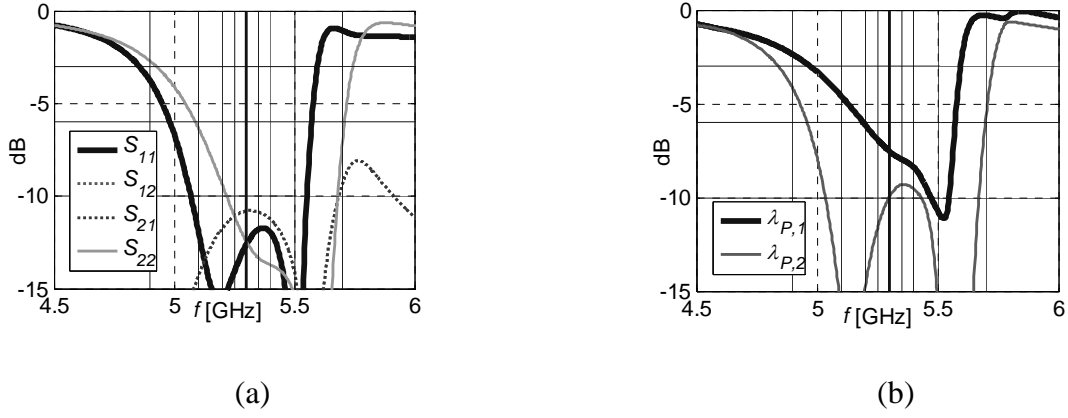


Figure 10. *Dual-polarized antenna resonance behavior (simulated parameters). In a) is the frequency behavior of the absolute values of scattering parameters and b) the behavior of the eigenvalues of the relative reflected power.*

In Fig. 11 we see the frequency behavior of the reflected power rate eigenvalues and pattern correlation for the measured dual-polarized antenna. Measurements are provided for one of the dual-polarized antennas in the stacked structure. In Fig. 11 we see a 600 MHz band with matching better than 5 dB and 1000 MHz band with correlation lower than 0.2. The low correlation is usually detected in simulations with scattering parameters [7], [P5], [P9], [95], [96]. The situation can be changed, when losses are taken into account [P8], [78]. The differences between simulated and measured results in Fig. 10b) and Fig. 11a) are due to the limitation of the computing power and practical limitations in the connecting cables to the antenna group, and finally, additional losses in the prototype. The coupling between the opposite directed dual-polarized antennas is very low, which means, that the radiation towards the edge direction of the laptop can be increased to improve the coverage of the whole 3D space. For low correlation a significant overlapping of beams is allowed [97]. This kind of compact multielement antenna structure is of interest for modern communications systems [94], [95]. The designed antenna system has not as good wideband matching as the one presented in [94] for a PDA device, but it is more compact. All four radiators are placed in a rectangular volume of $0.8 \times 0.8 \times 0.2$ wavelengths.

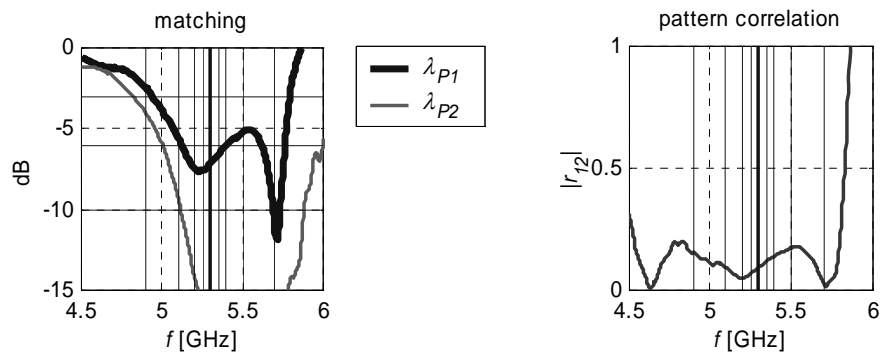


Figure 11. *Eigenvalues of reflected power rate matrix (left) and pattern correlation (right) for the realized antenna pair as a function of frequency. Calculations are performed with measured scattering parameters.*

5 Summary of publications

Paper [P1] presents the LSE method in linear pattern correction for a 6-element microstrip antenna array for arrays composed with the same radiating element and with different element orientations and spacings. The LSE method in pattern correction is compared with the correction performed by using the measured scattering matrix. For microstrip arrays the LSE method is better than the scattering matrix method. The reason is that the scattering matrix method is suitable for antennas with small aperture and does not take into account the substrate edge effect. The basic material for the publication is from [11]. However, more accurate calibration of the scattering matrix measurement was needed. Also the number of measured arrays was increased to seven. The element spacing in these arrays was from 0.3λ to 0.9λ . The center frequency was about 5.2 GHz. Due to mutual coupling it varied between different arrays and depended also on antenna element position. The effect of correction was that the array pattern nulls were located more exactly. The correction works best for beams in the forward direction. With a beam scan towards endfire the accuracy of the correction decreases. Compared with the results of pattern correction with single frequency the correction accuracy is lowered over a frequency band, when a common correction matrix is used over the whole frequency band.

Paper [P2] continues the research of linear pattern correction in real microstrip arrays using the same arrays as in [P1]. The linear pattern correction with a correction matrix used in [P1] requires desired complex-valued field patterns of the antenna elements. In [P2] another LSE method with iteration is presented. A desired element pattern, which gives good agreement with the corrected element pattern, is searched. For the desired array the element spacing is predetermined. The final corrected complex-valued element patterns (field patterns) are equal to each other when the reference point of each element is in its virtual, computational point of position. The element spacing in the corrected virtual array can alter and differ significantly from the real spacing. Thus it is called as virtual spacing. In the iteration the desired element pattern is varied, it is the mean of the corrected element patterns found in the previous iteration cycle. With this method the final element patterns are closer to each other than when using the method in [P1] and in the case of examined arrays they are also smooth and with a wide beam. The validity of the correction was studied also over a wider frequency band. With larger frequency bands the correction results become poorer when a common correction matrix over the whole band is used.

In [P3] the LSE method in pattern correction done in [P1] was developed further using pattern weighting. The basic LSE method gives a correction matrix, which can be used to find a complex-valued field pattern. Often the phase of the array pattern is arbitrary. The LSE method was used with iteration to generate an array amplitude/power pattern. When all the patterns (or rows in the pattern matrices) in the LSE method are multiplied component-wise with a weighting vector (usually called a cost function) the accuracy in different pattern directions changes. When a solution of relative error is needed, the inverse of the wanted array amplitude pattern should be used as the weighting vector. This robust weighting was presented in [P3]. The generation of box-type array patterns with wide nulls was examined and the array amplitude pattern generation with weighting was compared with the array amplitude pattern generation without weighting. The method with weighting was in general better for wide nulls from 0° to 60° and with the null depth of -20 dB to -40 dB for an array with element spacing of 0.5λ .

Paper [P4] deals with linear pattern correction using the scattering matrix. The scattering matrix method in pattern correction was presented in [P1], where it was compared with the LSE correction. The main advantage of pattern correction with scattering matrix is that it does not require measured element patterns. In [P1] the reference plane was shifted from the beginning of a coaxial probe towards the antenna to achieve better correction while in [P4] an equivalent input circuit consisting of up to three lumped elements was extracted from the array to achieve a more relevant scattering matrix for the correction. This correction improves slightly the scattering matrix method using reference plane shift done in [P1], but it is more complicated to perform.

In [P5] the connection between array mutual coupling and correlation between the array element patterns is established. The pattern correlation matrix of an array can be calculated if the scattering matrix is known and the array is lossless. For the examined arrays the pattern correlations were calculated using element patterns and also using measured scattering matrices. Results showed a good agreement between the two different ways to calculate the pattern correlation, even the patterns were defined only in the horizontal plane. For a two-element array a relation for scattering parameters $[S_{11}, S_{12}, S_{21}, S_{22}]$ and pattern correlation r_{12} was presented. Using the dependency between the scattering parameters and pattern correlation the level of acceptable coupling was estimated as a function of the angular spread of the incoming signal.

In papers [P6], [P7] and [P8] the connection between mutual coupling and pattern correlation is examined further. In [P6] the influence of the input signal type is studied. The input of the antenna group can be a vector of voltages, voltage waves or currents. These inputs are connected to each other with the scattering (or impedance) matrix. An important fact is that the antenna array element patterns are different for different input types when mutual coupling is present and if they are known in one of the three different presentations (scattering matrix, impedance matrix or admittance matrix presentation) they can be calculated in other presentations using the scattering matrix (or other coupling matrices). Because the element patterns are different in different presentations the pattern correlations are also different. In [P6] an example with a microstrip array is presented showing that the minimum pattern correlation can be reached with the reference plane shift in impedance and admittance presentations and it can be lower than the correlation in the scattering matrix presentation, which does not depend on the position of the reference plane. In paper [P7] the connection between mutual coupling and pattern correlation is presented in the most general formulation for lossless two-element arrays. The limiting case of lowest pattern correlation, when mutual coupling is given, is found, and also the lowest limit for mutual coupling is found, when the element patterns are given. This allows the antenna designer to estimate the level of pattern correlation or mutual coupling, when one of those is known. In paper [P8] the theory connecting mutual coupling and pattern correlation is generalized for antenna groups with losses and an example of a two-element array is presented. Losses complicate the estimations and in practice (the matrix of lost power is not trivial to obtain) the estimations can have only qualitative character.

In [P9] a dual-polarized antenna is used in MIMO channel analysis. The antenna was designed and manufactured so that two antennas of that kind could be stacked on different sides of a laptop type device. The center frequency was 5.3 GHz and a bandwidth of 400 MHz with 5 dB return loss was reached using a layered structure of patches. A band of 1000 MHz with a low pattern correlation of 0.2 was reached. When the signal is coming from many different directions the envelope correlations of the signals in the branches do not exceed $\rho_e = 0.2^2 = 0.04$, which is a very low value. The return loss of 5 dB was defined for the maximum possible reflected power (the worst case).

6 Conclusions

The effect of mutual coupling was examined for various antenna arrays with 6 elements and for an antenna group of 2 or 4 elements. For compensating the mutual coupling two basic methods for pattern correction were tested. The method using the measured scattering matrix was used in two modifications. In the first modification only one parameter, the reference plane of the scattering matrix was used. In the second case an equivalent input circuit of up to 3 lumped elements was extracted and the remaining scattering matrix was used for correction. The LSE method for pattern correction was tested using different modifications: The correction of element patterns when the desired element patterns were predetermined; finding a suitable element pattern and the corresponding correction matrix for a virtual array with arbitrary element spacing; and correction with weighting. In the last case of pattern correction the array patterns were generated directly from measured patterns without computing the corresponding correction matrix. The connection between mutual coupling and pattern correlation was examined with the manufactured antennas including the linear arrays and the dual-polarized antenna group, as well as with theoretical calculations. Basic relations were derived: For multielement antenna groups a matrix equation connecting the reflected power rate matrix and the pattern correlation matrix was presented. The reflected power rate matrix is a derivative of the scattering matrix. For a two-element array the equation was presented and analyzed with scattering parameters.

The LSE pattern correction method without predetermined element patterns was developed and examined for different arrays. In this method the element spacing in the ideal array is given and it can differ from the real spacing in the array. For a given element spacing an ideal array with equal element patterns is searched with iteration. Comparing with the basic LSE method with predetermined wanted element patterns this method gives element patterns that are very similar. The variation of virtual spacing gives a possibility for fine tuning of the pattern correction. Best results are detected when the virtual spacing is not far from the real spacing. The total length of the array can be extended to the dimensions of the ground plane, as was shown for an array with element spacing of 0.3λ . This property is an example on the possibility to use the array surroundings to extend the virtual array spacing to achieve better resolution for conformal arrays and arrays which are mounted on devices.

The pattern correction with the LSE method using weighting was tested for a 6-element microstrip array with element spacing of 0.5λ at 5.2 GHz. With weighting the expectation of the fitting accuracy is varied as a function of the direction angle. Correction with weighting was examined generating box-type amplitude/power patterns for the array. Thus an iteration of input coefficients was used with the pseudoinverse LSE method. The robust weighting was done using the inverse of the desired array amplitude pattern as a weighting vector. In this case the solution gives a LSE solution on the relative scale minimizing the relative error. The LSE method with robust weighting was compared with the basic LSE method without weighting by characterizing the error with a special error function for box-type desired array patterns. The main point of view was in wide null generation. The results showed that the method of robust weighting was better for the examined array when a wide null with a depth of -20 dB ... -40 dB and with a width $< 60^\circ$ is needed.

The basic pattern correction with the LSE method was compared to the method using the scattering matrix. The comparing criterion was LSE, for which the LSE method is definitely the best possible. The scattering matrix method gave only a small correction compared to the correction with the LSE method. When the more complicated method using a scattering matrix with input circuit extraction was used, then the results became only a little better. This shows that the LSE method is preferable

when pattern correction is needed. However, the scattering matrix method does not require measured or calculated element patterns, and is thus, due to its simplicity, potentially interesting. Also, it can give helpful model information for the antenna design. The main reason for the poor results using the scattering matrix method can be explained so that the scattering matrix method needs spatially constant aperture field which is realistic only for an electrically small antenna element. Additionally, it does not take into account the backplate edge diffraction.

Low element pattern correlation enables independent reception/transmission with the array branches. When the signals of interest are correlated in the array branches the received/transmitted information rate can be decreased. This is an actual problem in MIMO links which are predicted to be in wide use in the mobile communications in the near future. But also in the more traditional diversity use of the antenna groups the independence of the antenna reception/transmission is important. Mutual coupling can cause additional correlation between signals in array branches. However, the signals in the array branches are usually correlated also when there is no mutual coupling when a certain nearly plane wave is incoming. The pattern correlation is a pure array characteristic and gives a realistic estimate for the signal correlation, when signal arrive from many different directions. The calculated pattern correlation values are low in the examined cases. This shows that the problems with the signal correlation can arise mainly from the signal propagation environment, when the incoming signals are readily correlated, when impinging to antenna array.

To find out the pattern correlation level has a practical value also due to its connection to the array matching. When the array is lossless, the connection gives the opportunity to estimate the best possible matching with known pattern correlation and the highest value of pattern correlation with known matching. This connection can be used to estimate the matching when an antenna element for the array is planned and its pattern is known. When the array is lossy, in addition to the reflected power matrix also the matrix of lost power should be taken into account when the pattern correlation is calculated. Because the matrix of lost power is in practice difficult to find, the method to estimate pattern correlation should be used only in cases, when the mean of the lost power does not exceed the mean of the reflected power.

In antenna technology the antenna matching is a basic characteristic. In the multielement antenna groups mutual coupling causes that the matching is not trivial, because the antenna array matching cannot be achieved by matching each element independently to the system impedance. Pattern correlation always causes mutual coupling in the array. Thus, to avoid mutual coupling the element patterns should be changed. On the other hand, pattern and signal correlations can be zero with mutual coupling present. An antenna pair with a low pattern correlation and with moderate coupling over a wider frequency band was synthesized and manufactured. For the examined microstrip groups the matching seems to be a more complicated and challenging question than to obtain the low correlation. The use of the scattering matrix and relative reflected power matrix eigenvalues help to characterize the matching with more realistic way under the antenna group development process than the scattering parameters taken individually.

In this thesis non-idealities caused by mutual coupling in antenna groups were considered, namely the distortion of the patterns, mismatch and correlation between antenna elements. Different novel methods for estimation and compensation of these problems were developed and tested. A remaining challenge is to realize better matching and better space covering in the compact wideband multiantenna structures. The presented methods help in the design of small smart antennas for modern communications terminals.

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