Antenna mutual coupling and amplifier effects in transmission

Veli-Pekka Kutinlahti
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Abstract

The oncoming fifth generation (5G) telecommunication standard utilizes multi-antenna systems to implement multiple-input multiple-output and beam-steering capabilities in most wireless devices, including mobile devices. This shift in transceiver architecture will introduce each antenna element with its own feed control, along with an amplifier and phase shifter chain. The high integration level of these components prohibits the use of traditional ferrite circulators as isolators between the components, introducing the non-ideality of active reflections in the antenna elements to the amplifier outputs. The change in amplifier load impedance causes variation in amplifier output power, linearity, efficiency and can possibly even cause breakage of components in extreme cases. This development is parallel to the fact that 5G will use higher frequencies and wider bandwidth signals, driving the development of innovative design methods to achieve wide-band high-gain antennas with beam-steering capability.

The first part of the thesis describes optimizing different aspects of amplifier-antenna systems with mutual-coupling-induced mismatch. First, the equivalent isotropic radiated power (EIRP) of a 2x2 patch antenna array in an amplifier-antenna system is optimized through phase tuning. Phase tuning achieves a maximum 0.7-dB improvement in EIRP within the -3-dB beam steer range at 2.5 GHz, compared to progressive phase shift. Second, the 3rd order intermodulation is minimized with respect to the carrier by adjusting input feed power and phase in a two-tone excited 1x4 amplifier-antenna array, where the beam at each tone is independently steered. Optimization results in a 25-dB improvement in the signal-to-3rd-order-intermodulation ratio without decreasing far-field power density. However, this improvement comes at the cost of sacrificing beam integrity in terms of side-lobe level. Third, 3rd order intermodulation with respect to the carrier is minimized by antenna impedance matching using co-simulations of the amplifier and antenna.

The second part considers the optimization of realized gain in antenna arrays. First, an antenna array driven with element-specific amplifiers with varying output impedance is examined. Changes in amplifier gain may lead to altered output impedance and increased mismatch in the antenna interface, a phenomenon often neglected. An iterative method that accounts for the change in impedance is introduced, resulting in increased realized gain. Second, a cluster array concept is proposed to achieve high coverage gain over a wider band compared to a simple patch antenna array with similar elements. The cluster array utilizes patch elements with different resonant frequencies and high inter-element coupling to achieve wide-band matching with feeding weight tuning.

Keywords amplifier, antenna, beam-steering, load-pull, mutual coupling

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Tiivistelmä

Työn ensimmäinen osa käsittää erilaisten ominaisuuksien optimointia vahvistinantennijärjestelmissä, joissa esiintyy keskinäiskyttekkäen aiheuttama epäosivitusta. Ensimmäisenä ekivalenttinen isotrooppinen säteilyteho (EIRP) optimoidaan vahvistinantennijärjestelmässä vahviasäädöllä 2x2 mikroluiska-antenniryhmässä. Vahviasäädöllä saavutetaan korkeimmillaan 0.7 dB EIRP paranauksen puolestaan keilankääntöalueella 2.5 GHz taajuudella verrattaaessa progressiiviseen vaiheensiirtoon. Toisena kolmannen asteen intermodulaatio minimoidaan suhteessa cantotaajuuteen kaksitaajuisessa 1x4 vahvistinantenniryhmässä, jossa molempien taajuusosien keiloja käännetään itsenäisesti, säätämällä syötötehoja ja -vaiheita. Optimointi saavuttaa 25 dB paranuksen signaalin suhteessa kolmannen asteen intermodulaatioon ilman pääkeilojen tehohoiden alenemista. Keilojen eheys kuitenkin kärsii sivukellatajosten nousun muodossa. Kolmantena kolmannen asteen intermodulaatio suhteessa cantotaajuuteen minimoidaan antenniosoituksella käyttäen vahvistimen ja antennin yhteissimulaaatioita.


Avainsanat
antenni, keilankääntö, keskinäiskytentä, kuormanvento, vahvistin


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Preface

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Turku, March 16, 2024,

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List of Publications

This thesis consists of an overview and of the following publications which are referred to in the text by their Roman numerals.


Author’s Contribution

Publication I: “Amplifier-Antenna Array Optimization for EIRP by Phase Tuning”

The author is the main writer of the paper and conducted the research himself. This research was suggested by Prof. Viikari. This work was instructed by Dr. Lehtovuori and supervised by Prof. Viikari.

Publication II: “Analyzing and Optimizing the EIRP of a Phase-Tunable Amplifier-Antenna Array”

The author is the main writer of the paper and conducted the research himself. This work was instructed by Dr. Lehtovuori and supervised by Prof. Viikari.

Publication III: “Over-the-Air Suppression of Third-Order Intermodulation in a Two-Beam Steered Amplifier-Antenna Array”

The author is the main writer of the paper and conducted the research himself. The idea for the research was suggested by the author. This work was instructed by Dr. Lehtovuori and supervised by Prof. Viikari.

Publication IV: “Minimizing Intermodulation Distortion in a Transmitting Antenna Array with Matching”

The author is the main writer of the paper and conducted the research himself. The idea for the research was suggested by the author. This work was instructed by Dr. Lehtovuori and supervised by Prof. Viikari.
Author's Contribution

**Publication V: “Optimizing RF Efficiency of a Vector-Modulator-Driven Antenna Array”**

The author is the main writer of the paper and conducted the research himself. This research was suggested by Prof. Viikari. This work was instructed by Dr. Lehtovuori and supervised by Prof. Viikari.

**Publication VI: “Coupled Array of Diverse Elements for Wideband High Spherical Coverage”**

The idea of this work was proposed by Prof. Viikari. The author developed the mathematics and MATLAB code for analyzing the performance. Mr. Chen designed, simulated, and measured the antenna. Mr. Haarla contributed to the fabrication of the antenna prototype. Mr. Chen and Dr. Lehtovuori are responsible for the writing of the paper. This work was instructed by Dr. Lehtovuori and supervised by Prof. Viikari.
List of abbreviations and symbols

Abbreviations

3D  three dimensional
4G  fourth generation
5G  fifth generation
AiA  active integrated antenna
AM-AM  amplitude modulation to amplitude modulation
AM-PM  amplitude modulation to phase modulation
ARC  active reflection coefficient
CDF  cumulative distribution function
DC  direct current
DPA  Doherty power amplifier
DUT  device-under-test
EIRP  equivalent isotropically radiated power, effective isotropically radiated power
EM  electromagnetic
HPBW  half-power beam-width
IM3  3rd order intermodulation
IM3C  3rd order intermodulation to carrier ratio
IMD  intermodulation distortion
IoT  internet of things
IP3  3rd order intercept point
LNA  low noise amplifier
PA  power amplifier
List of abbreviations and symbols

**PAE**  power-added-efficiency
**PCB**  printed circuit board
**PHD**  polyharmonic distortion
**RF**  radio frequency
**RQ**  Rayleigh quotient
**SI3R**  signal-to-IM3-ratio
**SLL**  side-lobe level
**S-parameters**  scattering parameters
**TARC**  total active reflection coefficient
**TC**  threshold curve
**VNA**  vector network analyzer

**Symbols**

\( \Gamma_2 \)  vector of output reflection coefficients of amplifiers
\( \Gamma_{iv,\text{act}} \)  active voltage reflection coefficient at port \( i \)
\( \Gamma_{ip,\text{act}} \)  active reflection coefficient at port \( i \) for power wave
\( \Gamma_p \)  reflection coefficient for power waves
\( \Gamma_v \)  voltage reflection coefficient
\( \Delta_S \)  determinant of a two-port scattering parameter matrix
\( \eta \)  impedance of free space
\( \hat{\theta} \)  \( \theta \)-unit vector
\( \theta \)  \( \theta \)-angle
\( \lambda \)  free-space wavelength
\( \hat{\phi} \)  \( \phi \)-unit vector
\( \phi \)  \( \phi \)-angle
\( \varphi_i \)  phase shift of antenna element \( i \)

\( a \)  power wave into a component
\( a_{1,i} \)  power wave into the input port of \( i \)th amplifier
\( a_{2,i} \)  power wave into the output port of \( i \)th amplifier
\( \mathbf{a}_1 \)  a vector of power waves into the input ports of the amplifiers
**List of abbreviations and symbols**

- **a** subscripts:
  - \( a_2 \) a vector of power waves into the output ports of the amplifiers
  - \( a_i \) power wave into port \( i \) of a network
  - **a** vector of power waves into a multi-port network

- **B**
  - \( B \) amplifier model function
  - **B** set of amplifier model functions

- **b**
  - \( b \) power wave coming from a component
  - \( b_{1,i} \) power wave from the input port of \( i \)th amplifier
  - \( b_{2,i} \) power wave from the output port of \( i \)th amplifier

- **b** subscripts:
  - \( b_1 \) a vector of power wave from the input ports of the amplifiers
  - \( b_2 \) a vector of power wave from the output ports of the amplifiers
  - \( b_i \) power wave coming from port \( i \) of a network
  - **b** vector of power waves coming from a multi-port network

- **D**
  - \( D \) directivity of an antenna
  - \( D_{arr} \) directivity of an antenna array
  - \( D_{el} \) directivity of a single element in an array

- **E**
  - \( E \) total electric field
  - \( E_{\hat{\theta}} \) total electric field with \( \theta \)-polarization
  - \( E_{\hat{\phi}} \) total electric field with \( \phi \)-polarization
  - \( E_{0i} \) normalized embedded element pattern
  - \( E_{arr} \) total electric field of an array
  - \( E_i \) embedded element pattern

- **f** subscripts:
  - \( f_i \) frequency of tone \( i \)

- **G**
  - \( G \) antenna gain, power gain of an amplifier
  - \( G_A \) available power gain of an amplifier
  - \( G_T \) transducer power gain of an amplifier
  - \( G_{real} \) realized gain

- **I**
  - \( I \) total current

- **K**
  - \( K \) K-factor
List of abbreviations and symbols

\( N \) number of elements in an array
\( P_L \) power-to-the-load
\( P_{1\text{dB}} \) 1dB compression point
\( P_{\text{acc}} \) accepted power
\( P_{\text{av}} \) available power
\( P_{\text{avn}} \) available power from the network
\( P_{\text{avs}} \) available power from the source
\( P_{\text{DC}} \) direct current power
\( P_{\text{in}} \) power into an amplifier
\( P_{\text{rad}} \) radiated power
\( \vec{r}_i \) position vector of element \( i \) with respect to element 1
\( S_{ji} \) scattering parameter from port \( i \) to port \( j \)
\( S \) power density of a plane wave
\( \mathbf{S} \) scattering matrix
\( V^- \) reflected voltage wave
\( V_i^- \) reflected voltage wave at port \( i \)
\( V^+ \) incident voltage wave
\( V \) total voltage
\( Z_0 \) impedance of transmission line 0, characteristic impedance of a transmission line
\( Z_1 \) impedance of transmission line 1
\( Z_A \) input impedance of an antenna
\( Z_{iA,\text{act}} \) active antenna impedance of port \( i \)
\( Z_R \) reference impedance
\( Z_{\text{out}} \) output impedance
1. Introduction

1.1 Background

It could be said that, whereas the iPhone, the fourth-generation (4G) Long-Term-Evolution telecommunications standard, and the HTML5 brought the internet to everybody [1–3], the fifth-generation (5G) telecommunications standard will bring the internet to everything [4]. The Internet of Things (IoT) is envisioned to have most of our electronic devices connected to the internet and communicate with each other and with us. The IoT is said to have already happened, with the number of devices on the internet surpassing the number of people [5], but it has not visibly reached the common consumer yet. One major application promised by 5G, which is getting very prominent coverage, is the self-driving car. Requiring constant communication with other self-driving cars and high reliability, but offering complete change in personal transportation, it will be, upon realization, a groundbreaking change in our daily lives.

5G will offer higher data rates, lower latency, and increased mobility [6]. The gigabytes per second throughput promised by 5G is to be achieved with the utilization of higher communication frequencies. The currently allocated bands above 24GHz can support this, at the expense of higher transmission losses. Because of the increased losses, even mobile devices are required to be equipped with antennas capable of beamforming [7]. Designing wide-band beamforming-capable antennas in volume-restricted mobile devices is a difficult task [8], which requires modern computer-aided design software to perform three-dimensional (3D) electromagnetic (EM) simulations.

With the advent of new technology, the engineering required to achieve it will also undergo a drastic change. It can be said that engineering "tries to make everything as simple as possible, but no simpler," a quote often attributed to Albert Einstein but not definitively verified [9]. The current paradigm of designing modern wireless communication systems
is becoming more complex, and as such, simpler traditional solutions are no longer sufficient. As 5G requires multiple individually controlled antenna elements to perform beamforming while simultaneously requiring the system to use a small volume and have high efficiency, traditional solutions that aim to simplify system design need to be excluded. Antennas will exhibit high coupling because of tight spacing, which will affect their matching. Moreover, the matching will change because of beam-steering. Isolators, a common component at the amplifier-antenna interface allowing for almost completely independent design of amplifiers and antennas, are becoming too large to be incorporated into the systems [10]. With the lack of isolation and changing matching, the load-pull effect, not prevalent previously, will affect the performance of the whole system. This needs to be taken into account in modern systems to ensure linearity, efficiency, and output power.

Realizing 5G systems will require advances in multiple areas of radio frequency (RF) engineering. The transmitter is responsible for modulating, amplifying, and radiating the communication signal. With the stringent space and visual requirements of modern mobile devices, achieving the communication standard requires higher integration of different components than before. This integration will require taking into account non-idealities that have previously been absent.

1.2 Objectives of the thesis

The main objective of this thesis is to increase understanding of the effects of antenna coupling and load pull at the amplifier-antenna interface on transmitters. This includes feed optimization, modeling of load pull, and matching network design. The findings are universal, as the methods are based on theory, which can be used to describe the operation of antennas and amplifiers at any frequency range. The findings can be especially useful in the upcoming 5G applications, where closely spaced isolation-free transmitters are expected to be commonplace.

1.3 Main scientific merits

The main scientific merits of this thesis are as follows:

1. Showing that considering amplifier load-pull effects in an amplifier-antenna array allows for optimizing the equivalent isotropic radiated power (EIRP) of the system through feed control, especially when significant coupling is present. [I], [II]
2. Demonstrating that feed control can be utilized to decrease intermodulation distortion in the radiated fields of a system with two independently steered single-tone beams. [III]

3. Presenting an amplifier-antenna design method for minimizing intermodulation through matching, utilizing co-simulation of the amplifier-antenna system. [IV]

4. Developing a method to optimize the realized gain of an antenna array that is driven with amplifiers whose output impedances vary with the variable gain of the amplifiers. [V]

5. Developing an analysis method for antenna design to leverage mutual coupling for wide-band impedance matching, with demonstrated operation at mmWave frequencies. [VI]

1.4 Organization of the thesis

The summary is organized with antenna array, RF amplifier, and load-pull fundamentals described in Chapter 2. Chapter 3 holds load-pull accounted antenna array modeling and optimization methods with respect to feed control and matching networks. Chapter 4 contains methods to maximize realized gain over a large bandwidth with antenna array design and feed optimization in the presence of changing port impedance. Chapter 5 holds the brief summaries of the publications, and conclusions are drawn in Chapter 6.
2. Amplifiers and antennas in transmitters

A transmitter is a device responsible for mixing, amplifying, and radiating a processed input signal containing information towards a receiver at a distance. In modern systems, the transmitter can employ multiple antennas, working independently or together depending on the application, with each equipped with its own amplifier and feed control. Carefully selecting the feeding weight of each element is needed to utilize the benefits of the whole system.

This chapter describes the basics of antenna arrays and amplifiers necessary to understand the findings of the thesis. Additionally, a literature review of active antenna array research is presented.

2.1 Antenna arrays

An antenna array is, by definition, "an antenna comprised of a number of identical radiating elements in a regular arrangement and excited to obtain a prescribed radiation pattern" [11]. Within this definition, fits a wide variety of different antenna arrays that differ by element type, element count, and feed network, among other characteristics. This section describes the operation of antenna arrays as linear devices and the metrics used to characterize them.

2.1.1 Matching and S-parameters

Matching describes how power is transmitted between two microwave components. Matching, in this case the reflections coefficient [12], is generally presented in decibels, and typical matching levels are −3dB, −6dB, −10dB, and −20dB, which describe how much power is reflected back into the source. As microwave systems are usually unidirectional, matching generally refers to how the input of the next component is matched to the output of the previous component.

Transmission line theory tells us that the reflection $\Gamma_v$ of a voltage wave...
V+ at an interface of two different impedances Z₁ and Z₀ is
\[ \Gamma_v = \frac{V^-}{V^+} = \frac{Z_1 - Z_0}{Z_1 + Z_0}, \] (2.1)
where V− is the reflected wave, and Z₀ is the complex characteristic impedance of the transmission line. The reflection is analogous to an antenna input impedance, usually denoted with Zₐ, connected to a transmission line. With high-frequency signals, currents and voltages are difficult to measure [13], which is why rather than using voltage waves V+ and V− for forward and backward propagating waves, the used format is power waves a and b, respectively.

Power waves a and b are defined with the total voltage V and total current I peak amplitudes with respect to a complex reference impedance Zₐ as [14]–[16]
\[ a = \frac{V + Z_R I}{2\sqrt{\Re(Z_R)}}, \] (2.2)
\[ b = \frac{V - Z_R^* I}{2\sqrt{\Re(Z_R)}}, \] (2.3)
and they technically have a unit of [a] = \( \sqrt{\text{W}} \). The reference impedance can be selected freely but is quite universally Zₐ = 50Ω, which is the characteristic impedance of the coaxial cables used in vector network analyzer (VNA) measurements.

Waves a and b are not physical but a mathematical tool that satisfies the maximum power transfer theorem, which states that power to the load (P_L) is maximized with conjugate matching. Time-averaged power to the load is calculated from a and b with
\[ P_L = \frac{|a|^2}{2} - \frac{|b|^2}{2} = \frac{|a|^2}{2} (1 - |\Gamma_p|^2), \] (2.4)
where \( \Gamma_p \) is the reflection coefficient for power waves. \( \Gamma_p \) of antenna impedance Zₐ for power waves is
\[ \Gamma_p = \frac{b}{a} = \frac{Z_A - Z_R^*}{Z_A + Z_R}. \] (2.5)
From (2.5), it is evident that with Zₐ = Z_R*, the power reflection is minimized, and maximum power transfer is satisfied. Also, because of the Zₐ = Z_R* relation, matching typically means impedance matching, i.e., keeping the impedance at the interface the same for both components. Note that in the case of purely real Z_R [18], (2.1) and (2.5) are equal.

Equation (2.5) is useful when analyzing, for example, linear amplifiers when designing their input and output matching, as their matching typically requires complex impedances [17]. Note that the power reflection coefficient (2.5) cannot be plotted on a Smith chart in the case of a complex reference impedance Z_R [18].
With multiple antenna elements comprising an antenna array, single port matching is not sufficient to describe the whole antenna system. As signals couple from element to element, a scattering parameters (S-parameters) $S_{ji}$ are used [16], [19]. Arranged in a matrix called the S-matrix $S$, S-parameters describe the interaction of the incident wave $a_i$ into port $i$ with the output waves $b_j$ at all the ports in the array. Parameter $S_{ii}$ is the reflection coefficient of port (antenna element) $i$ and $S_{ji}$ is the transmission (coupling) coefficient from port $i$ to port $j$.

To calculate the reflected waves $b_i$ when multiple inputs of an array are excited simultaneously with input waves $a_i$, one simply calculates the linear equation

$$b = \begin{bmatrix} b_1 \\ \vdots \\ b_n \end{bmatrix} = \begin{bmatrix} S_{11} & \cdots & S_{1n} \\ \vdots & \ddots & \vdots \\ S_{n1} & \cdots & S_{nn} \end{bmatrix} \begin{bmatrix} a_1 \\ \vdots \\ a_n \end{bmatrix} = Sa. \quad (2.6)$$

The previously defined reflection coefficient (2.1) and (2.5) are perfectly usable in a situation with multiple ports excited together, influencing the resulting reflected waves $V_i^-$ and $b_i$, respectively. Extending (2.5) to include the effect of feeding weights $a^T = [a_1 \ldots a_n]$ on the reflection coefficient in port $i$ gives

$$\Gamma_{pi,\text{act}} = \sum_j S_{ij} a_j / a_i. \quad (2.7)$$

The term used when multiple ports are excited simultaneously is the active reflection coefficient (ARC) to distinguish it from a single-element case [11].

As was previously mentioned, matching means impedance matching. Thus, evaluating the behavior of an antenna array in terms of impedances, even in an active situation, is sometimes used. The term used in such a case is active impedance [11]. The formulas for reflection coefficients (2.1) and (2.5) can be rearranged to calculate the active impedance $Z_{i,\text{act}}$ of port $i$ when the array is excited as

$$Z_{i A, \text{act}} = Z_0 \frac{1 + \Gamma_{iv, \text{act}}}{1 - \Gamma_{iv, \text{act}}}, \quad (2.8)$$

$$Z_{i A, \text{act}} = Z_R + \Gamma_{ip, \text{act}} Z_R / 1 - \Gamma_{ip, \text{act}}. \quad (2.9)$$

Including the effect of feeding weights $a^T = [a_1 \ldots a_n]$ by replacing $\Gamma_{ip, \text{act}}$ in (2.9) with (2.7) gives

$$Z_{i A, \text{act}} = \frac{Z_R^* + Z_R \sum_j S_{ij} a_j / a_i}{1 - \sum_j S_{ij} a_j / a_i}, \quad (2.10)$$
from which it can be seen, that active impedance can vary tremendously by changing the feeding weights $\mathbf{a}$.

For an antenna array or other multiport antennas, it might be beneficial to utilize the concept of the total active reflection coefficient (TARC) \[20\]. TARC is the ratio of the total reflected power in all ports to the total excited power. In antenna arrays, the elements are usually excited with similar power, and in regular arrays, the elements are quite identical in electrical properties, justifying the use of TARC over individual ARCs. TARC is formulated in matrix form as

$$\text{TARC} = \sqrt{\frac{\mathbf{b}^H \mathbf{b}}{\mathbf{a}^H \mathbf{a}}} = \sqrt{\frac{\mathbf{a}^H \mathbf{S} \mathbf{a} \mathbf{S}^H \mathbf{a}}{\mathbf{a}^H \mathbf{a}}},$$

(2.11)

where $^H$ denotes the conjugate transpose or Hermitian of the matrix or vector.

With single-feed electrically small antennas in mobile phones, the problem of matching is the intrinsically small radiation resistance of the antennas \[21\]. For antenna arrays and multi-port antennas in general, the difficulty of matching arises partly from the coupling between the elements. As described by (2.5), (2.6), and (2.9), the active impedance of an element depends on how all the ports in the array are fed, assuming the coupling coefficients $S_{ji}$ are non-zero. In beam-steering, the feeding weights $\mathbf{a}$ of the array are changed, and thus, the impedance at each port varies during operation, making the matching of an array non-constant.

The problem of varying active impedance in beam-steered antenna arrays has led to increased research to lower the coupling between elements. Numerous methods exist for reducing coupling, including decoupling networks \[22\]–\[25\], neutralization lines \[26\]–\[28\], band-gap structures \[28\]–\[31\], and defected ground structures \[28\], \[32\]–\[35\]. Some of these methods require altering the 3D structure of the antenna, while others can be implemented after the antenna’s design by utilizing specialized feeding circuits.

### 2.1.2 Parameters of radiation pattern

The radiation pattern of an antenna describes how the antenna excites traveling EM waves in the far-field and how it couples to EM waves produced by other antennas. The radiation pattern is frequency-specific and can be described in a multitude of coordinate systems. In this thesis, the coordinate system used is the standard circular coordinate system, and this notion extends to all the publications as well. The broadside of antennas and antenna arrays systematically faces the z-axis, i.e., $\theta = 0^\circ$. Fig. 2.1 shows the coordinate system and the orientation of antenna arrays.

At the basis for further analysis of patterns is the total electric field $\mathbf{E}(\theta, \phi)$ at a constant distance $r$. The total electric field is the sum of two
orthogonal polarizations $\mathbf{E}_{\hat{\theta}}(\theta, \phi)$ and $\mathbf{E}_{\hat{\phi}}(\theta, \phi)$

$$\mathbf{E}(\theta, \phi) = \mathbf{E}_{\hat{\theta}}(\theta, \phi) + \mathbf{E}_{\hat{\phi}}(\theta, \phi),$$

(2.12)

where the used linear polarization vectors are $\hat{\theta}$ and $\hat{\phi}$. Usually, the electric field is not used directly to describe an antenna, but rather the power density pattern $S(\theta, \phi)$ is used. The total power density is calculated from orthogonal electric field components with

$$S(\theta, \phi) = \frac{|\mathbf{E}_{\hat{\theta}}(\theta, \phi)|^2}{2\eta} + \frac{|\mathbf{E}_{\hat{\phi}}(\theta, \phi)|^2}{2\eta}.$$

(2.13)

Antennas are usually designed to work well with a single polarization, and this polarization contributes the majority of the power density. In these situations, (2.13) can usually be utilized with only one polarization without too much error in the result.

The directivity pattern $D(\theta, \phi)$ of an antenna describes to what directions the antenna radiates the radiated power $P_{\text{rad}}$. Typically, an antenna is characterized by its maximum directivity $D$

$$D = \frac{S_{\text{max}}}{\int_{0}^{2\pi} \int_{0}^{\pi} S(\theta, \phi) \, d\Omega} = \frac{S_{\text{max}}}{P_{\text{rad}}/4\pi r^2} = \frac{S_{\text{max}} r^2}{P_{\text{rad}}/4\pi} = \frac{U_{\text{max}}}{P_{\text{rad}}/4\pi},$$

(2.14)

where $S_{\text{max}}$ is the maximum power density excited by the antenna in the far-field and $U_{\text{max}}$ is the maximum radiation intensity. The rightmost form is the standard definition of directivity, but the formulation used by utilizing $S_{\text{max}}$ is equally valid, as the distance $r$ is kept constant for all the steps of the calculation.

The directivity does not take into account losses in the dielectric or metals in the antenna or mismatch. The antenna gain $G$ accounts for losses in the antenna, and realized gain $G_{\text{real}}$ accounts for both loss and mismatch [11]. Both can be also presented as patterns with angular dependence. Antenna gain $G$ accounts for both dielectric and ohmic losses.
in the antenna structure by relating the power density in the far-field to the accepted power \( P_{\text{acc}} \) rather than the radiated power \( P_{\text{rad}} \)

\[
G = \frac{S_{\text{max}}}{P_{\text{acc}}/4\pi r^2}.
\]  

Realized antenna gain \( G_{\text{real}} \) finally accounts for the losses in the structure as well as from mismatch by using the available power \( P_{\text{av}} \)

\[
G_{\text{real}} = \frac{S_{\text{max}}}{P_{\text{av}}/4\pi r^2}.
\]  

The directivity and both gains can be defined separately for orthogonally polarized fields of the antenna. In this case, the term used is partial directivity, partial gain or partial realized gain [11].

Another useful parameter used with antennas, which relates to the whole system than just solely the antenna, is the equivalent isotropic radiated power (EIRP) [11]. It describes the total power radiated by an ideal isotropic antenna with far-field power density equal to the maximum far-field power density of the evaluated antenna. It accounts for the input power, increasing if the power fed to the antenna is increased. It is, by definition, calculated by multiplying antenna gain \( G \) with the accepted power \( P_{\text{acc}} \)

\[
\text{EIRP} = GP_{\text{acc}},
\]

but it can also be calculated straight from directivity \( D \) or realized gain \( G_{\text{real}} \) with the respective power in the denominators because of the similar formulation of the three parameters.

In addition to the previous parameters, which have an angular dependency, there are a multitude of parameters that describe individual aspects of the radiation pattern. Half-power beamwidth (HPBW) describes the angular span of the main beam, where the radiation intensity decreases at most by half of the maximum intensity [11]. A null is a minima in the radiation pattern; in an ideal case no radiation is directed towards a null. Side-lobes are beams located between nulls. Side-lobe level (SLL) is the ratio of side-lobe maxima to the maxima of the main beam. Fig. 2.2 describes these parameters visually. A back lobe is the side-lobe in the opposite direction to the broadside.

### 2.1.3 Radiation pattern of antenna array

An antenna array, synonymous with array antenna, is composed of multiple similar antenna elements in a regular arrangement fed simultaneously to form a combined field. The combination of fields from multiple antennas leads to a more directive far-field. The most relevant approximation for the maximum directivity of an antenna array \( D_{\text{arr}} \) is for an array where
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Figure 2.2. A single element antenna and an antenna array.

Element-to-element spacing is half of a free-space wavelength. The approximation for an array with $N$ elements is

$$D_{arr} = ND_{el},$$

(2.18)

where $D_{el}$ is the directivity of a single individual element. Equation (2.18) assumes no mutual coupling between the elements, i.e., the coupling coefficients $S_{ji}$ of the array are zero. Larger spacing up to approximately a full wavelength would increase the directivity. Spacing of half-wavelength or less is, however, more relevant for beam-steering capable systems, as this rejects grating lobes, i.e., side-lobes with equal magnitude to the main lobe [36].

In reality, (2.18) serves as an upper bound for an early stage of antenna design. When an antenna is placed close to another to form an array, the embedded element patterns $E_i(\theta, \phi)$ of both antennas change from the isolated patterns due to coupling [36]. The total electric field pattern of the array $E_{arr}(\theta, \phi)$ in the far-field is the superposition of the fields $E_i(\theta, \phi)$ of all elements 1 to $n$ [37]

$$E_{arr}(\theta, \phi) = \sum_{i=1}^{n} E_i(\theta, \phi),$$

(2.19)

from which the power density in the far-field can be calculated with (2.13).

Besides the increase in directivity, an antenna array with individually controlled feed phases and amplitudes allows for beam steering and beam shaping during operation. This is a cornerstone in modern telecommunications, as the maturing 5G systems will utilize high-gain beam-steerable antenna arrays to overcome losses in transmission caused by the high frequencies required by the increasing bandwidths. In beam-steering, the input waves $a_i$ into each element $i$ in an array are weighted to control the total electric field. The total electric field pattern $E_{arr}(\theta, \phi)$ of an antenna array in the far-field is calculated from the normalized embedded element...
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Patterns \( \mathbf{E}_{\Omega i}(\theta, \phi) \) weighted by the input waves \( a_i \):

\[
\mathbf{E}_{\text{arr}}(\theta, \phi) = \sum_{i=1}^{n} \mathbf{E}_{\Omega i}(\theta, \phi)a_i.
\]  

(2.20)

Normalized element patterns are attained by exciting the array elements one at a time with other ports matched to 50 \( \Omega \) with \(-3\) dBW time-averaged input power. \(-3\) dBW input power corresponds to a power wave amplitude of \(|a| = 1\). Normalized element patterns account for the effects of element pattern distortion because of nearby elements, as well as mismatch of the element itself and coupling to other elements.

A multitude of antenna array feeding schemes exist to control the total field \( \mathbf{E}_{\text{arr}}(\theta, \phi) \). The most common beam-steering method is the progressive phase shift [38], where the array geometry and the wanted beam direction define the phase shifts \( \varphi_i = \angle a_i \) of the input signals of all elements. Fundamentally, one element of the array is defined as the origin to calculate all the other phase shifts and has a phase shift value of \( \varphi_1 = 0^\circ \). Other phase shifts \( \varphi_i \) (in radians) are calculated for a planar array on the xy-plane with

\[
\varphi_i = -\frac{2\pi r_i}{\lambda} \cdot (\hat{x}\sin\theta_0\cos\phi_0 + \hat{y}\sin\theta_0\sin\phi_0),
\]  

(2.21)

where \( r_i \) is the position vector of the element \( i \) with respect to element 1, \( \lambda \) is the free-space wavelength of the radiation, \( \hat{x} \) and \( \hat{y} \) are the orthogonal unit vectors on the plane, and \( (\theta_0, \phi_0) \) is the direction of the desired main beam [36]. Equation (2.21) is precise only for an infinite array with identical element patterns, and results in a slight misalignment of the beam the further it is steered from the broadside direction.

For a linear array on the x-axis with the first element at the origin, (2.21) simplifies to

\[
\varphi_i = -kd(i - 1)\sin\theta_0,
\]  

(2.22)

where \( k \) is the wavenumber and \( d \) is the element-to-element spacing. The term \((i - 1)\) guarantees that the first element has a phase shift of 0°. Expressing the waves \( a_i \) in terms of amplitudes \( A_i \) and phase components \( \exp(-jkdi\sin\theta_0) \), (2.10) becomes

\[
Z_{ip,\text{act}} = \frac{Z_R^* + Z_R \sum_{h=1}^{n} S_{ih}A_h e^{-jkd(h-1)\sin\theta_0}}{1 - \sum_{h=1}^{n} S_{ih}A_h e^{-jkd(h-1)\sin\theta_0}},
\]  

(2.23)

from which the scan-angle dependence of the ARCs and active impedances is evident.

In addition to controlling the phase shift of the input signals, also input signal amplitude can be controlled. Amplitude control allows for a general
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increase or decrease the radiated power density, but when combined with phase control it is also used to control SLLs and possible nulls \[39\], \[40\]. With amplitude tapering, the most common type of amplitude control, the input power is decreased from center element towards the edge elements. Side-lobes can be effectively eliminated completely with the main beam in the broadside direction at the expense of the main beam width \[36\]. Typical amplitude tapering schemes include triangular, binomial and Dolph-Chebyshev.

The previously mentioned control methods usually ignore array non-idealities, such as embedded element distortion and coupling. In modern systems, the input signal weights can be optimized using simulated or measured embedded element patterns and an optimization routine with \ref{2.20} to find the best set of weights \(a\). This can be used to manipulate the pattern more freely, allowing, for example, the formation of multiple nulls in the desired directions simultaneously while controlling the main beam direction \[41\]–\[47\].

2.2 RF amplifiers

In RF applications, amplifiers are used extensively to boost signals to usable levels. Different types of amplifiers are utilized in different sections of systems to leverage their primary operation for most benefit. A low noise amplifier (LNA) is the first component in a receiver after the antenna to keep overall noise level as low as possible, and a power amplifier (PA) is the last component in a transmitter before the antenna to boost the signal level high enough to reach the end-user. For 5G hand held devices, the maximum total radiated power is set to 23dBm, which effectively represents the total power from the amplifiers \[48\]. Amplifier linearity, the ability to maintain signal integrity in amplification, is an important subject, as it affects efficiency as well as users on adjacent channels.

Besides differentiating amplifiers with respect to the type (LNA or PA),
amplifiers are also differentiated by their class, indicating whether they are linear or non-linear and whether they are differential or not. The class inherently describes the arrangement of transistor or transistors used to make up the amplifying structure. Linear amplifiers are driven far below compression and non-linear amplifiers are driven near or even in compression to boost efficiency. Whether an amplifier is differential or not describes the input and output structure. Choosing an application-specific amplifier requires knowledge of the benefits and trade-offs of all classes and structures to make the correct choice [49].

Figure 2.3 illustrates the basic operation of the amplifier as well as the relevant input and output wave notations for this thesis. This section provides a general description of amplifier characteristics without delving into the specific advantages of different amplifier types, thereby aiding the understanding of amplifier-antenna interface analysis in Chapter 3.

2.2.1 Amplifier characteristics

Stability is typically an inherent characteristic of an amplifier. Instability arises from the constructive amplification of noise in the active device, eventually leading to saturation and a constant output at the oscillating frequency. A common parameter used to describe stability is the K-factor $K$ [50], defined with S-parameters as [51]

$$K = \frac{1 - |S_{21}|^2 - |S_{22}|^2 + |\Delta S|^2}{2|S_{12}S_{21}|},$$

(2.24)

where $\Delta S$ is the determinant of the two-port S-parameter matrix of the amplifier. Instability can in simple terms be traced to the input and output mismatch, where output of the amplifier carries back to the input without sufficient dampening, cascading until saturation occurs.

Multistage amplifier design requires careful consideration of stability, especially since the output power of subsequent stages is considerably higher than that of the preceding stages. This situation can lead to the mismatch of the last stage cascading down the line and causing the previous stages to become unstable [52]. Instability is best addressed, especially in the case of passive loads, by ensuring sufficient matching levels and isolating waves traveling backward in the amplifier chain. K-factor cannot be used to assess possible internal instabilities of a circuit, just the instabilities visible from the input and output ports. In multistage amplifiers, the stability conditions need to be evaluated for each stage separately.

After stability, perhaps the most crucial parameter of an amplifier is its power gain, which describes the amplifier’s ability to amplify the input signal. As RF analysis mostly involves power waves, characterizing RF amplifiers with power gain—over voltage or current gain commonly used in electronics—is apparent. Generally, there are three different types of
power gain for any two-port device. First, the power gain $G$:

$$G = \frac{P_L}{P_{in}} = \frac{|b_2|^2 - |a_2|^2}{|a_1|^2 - |b_1|^2},$$  \hspace{1cm} (2.25)$$

where $P_L$ is the power delivered to the load (power-to-the-load) and $P_{in}$ is the power into the amplifier. Second, the available power gain $G_A$:

$$G_A = \frac{P_{avn}}{P_{avs}} = \frac{|b_2|^2}{|a_1|^2},$$  \hspace{1cm} (2.26)$$

where $P_{avn}$ and $P_{avs}$ are the available powers from the network and the source, respectively. Third, the transducer power gain $G_T$:

$$G_T = \frac{P_L}{P_{avs}} = \frac{|b_2|^2 - |a_2|^2}{|a_1|^2}. \hspace{1cm} (2.27)$$

These different gains are useful in the design of linear amplifiers in the presence of passive loads and relate to the reflections in presence of different types of matching [12].

Efficiency is the measure of the amplifier’s ability to convert direct current (DC) power to RF power. Multiple efficiencies exist, but typically only the power-added-efficiency (PAE) is considered. PAE relates the DC power $P_{DC}$ to the power added to the input RF signal

$$\text{PAE} = \frac{P_L - P_{in}}{P_{DC}}. \hspace{1cm} (2.28)$$

Low efficiency leads to heating of the active circuitry, which, in turn, affects the gain, matching and efficiency of the amplifier stage [49].

Amplifier non-linearities include single tone amplitude and phase distortion (AM-AM and AM-PM), harmonic distortion, intermodulation distortion (IMD), and others. AM-AM is another term of gain compression, where an increase in input power at some point starts to degrade the gain, finally leading to output saturation. AM-AM is usually characterized by 1-dB compression point $P_{1\text{dB}}$, after which the gain of the amplifier has decreased by more than 1 dB compared to the linear region. AM-PM describes the phase shift between a low input signal and a high input signal. Harmonic distortion is the production of harmonics of the input frequency. IMD is the mixing of two signals with different frequencies $f_1$ and $f_2$, which produces signals at new frequencies $nf_1 + mf_2$. Maybe the most important IMD metric is the level of 3rd order intermodulation products (IM3), which are the two closest intermodulation products ($2f_1 - f_2$ and $2f_2 - f_1$) to the main tones. IM3 level is typically characterized with 3rd-order-intercept point (IP3), which is a hypothetical point at which the main tone and IM3 tone powers are equal. The point is hypothetical because an amplifier output saturates well before IP3 is reached, but it is a commercially used metric to evaluate amplifier linearity. As modern communication systems utilize wide-band modulated signals, the use of IP3 is not descriptive enough.

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A more used method is to evaluate the adjacent channel power leakage (ACPL) also known as adjacent channel power ratio. ACPL is measured by feeding the amplifier with the modulated signal in question and either taking the average or maximum powers on the main channel and the adjacent channels [53].

2.2.2 Load-pull of RF amplifiers

Load-pull is a technique that originated from the design of oscillators. Oscillators utilize transistors, similar to amplifiers, but instead of amplification, they are designed to output a single frequency at constant power and phase. Load-pull refers to the change in the load impedance, which induces varying effects on the operation of the device being load-pulled. For oscillators designed to operate at a specific frequency, changing the load impedance causes the frequency to shift. Load-pull equipment was developed to simplify the design process by providing the ability to systematically change the load impedance to study the behavior of the oscillator. This later became a practice utilized in amplifier design as well [54].

As an amplifier is inherently a non-linear device, it is not always adequately described by S-parameters. The linear region of an amplifier is loosely described as the region where utilizing S-parameters is sufficiently accurate, but as the amplifier is driven towards compression, the accuracy when using S-parameters decreases. As amplifier efficiency is highest with high output power, as evident from (2.28), designing high-efficiency non-linear amplifiers typically requires load-pull measurements or simulations [54].

Load-pull measurement systems are only briefly discussed in this work. The main focus is on the results obtained from them, regardless of the measurement setup itself. Different load-pull measurement systems can be roughly divided into two categories: passive and active. Active systems are generally preferred over passive systems due to their speed, flexibility in measurements, and the ability to measure reflection coefficients larger than one. However, this comes at the cost of increased complexity in the measurement equipment [54].

In a passive system, the device-under-test (DUT) is connected to a passive load, which can be tuned to control the reflections. This tuning can be achieved, for example, with a length of a transmission line with a dielectric rod on top. The distance of the rod from and along the line can be adjusted, changing the impedance of the line and the phase of the reflections. This allows the load-pulled impedance to be moved on the Smith chart. In a passive system, the reflection coefficient cannot be made larger than unity, and it is largely affected by the losses in the transmission lines used in the measurements [54].

In an active system, the reflected wave $a_2$ is controlled with external...
amplifiers and signal generators much more freely. With an active system, the magnitude of the reflections can be made as large as the amplifiers in the measurement equipment allow. Passive and active load-pull measurements are equivalent, as the DUT does not know if the reflections are caused actively or passively. In an active system, the impedance set for the measurement is found iteratively, as change in $a_2$ leads to change in $b_2$ and thus changes the reflection coefficient $b_2/a_2$, i.e., the wanted impedance $Z_{\text{out}}$.[54]

Principally, load-pull presents a set of impedances $Z_{\text{out}}$ to the DUT output and measures the response. The main interest is in the $b_2$ spectrum, i.e., how the gain is affected and how the frequencies not included in the input signal (IMD and harmonics) are affected. Drawn DC power is also monitored to evaluate efficiency changes. The results acquired from load-pull are plotted on a Smith chart as constant contours showing decreasing performance around a maximum or minimum. Fig. 2.4 illustrates a representation of the plots. From the plots, a compromise fulfilling the design parameters is chosen for the matching. As maximum efficiency, gain and linearity do not coincide, selecting a correct design requires deep understanding of the application requirements.

### 2.3 Active integrated antennas

Current trend in transmitter systems is toward higher degrees of integration. The amplifier-antenna interface has been seen as an ample location for improvements in recent years by co-designing the impedance of the antenna element to match the amplifier without intermediate matching networks [55]–[59]. As the separate design of the components requires both to be independently matched to the same impedance value, typically 50Ω, integration can effectively eliminate the need for separate matching networks, which lowers losses. These types of active antennas are called active integrated antennas (AiA).
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The main driver in the integration is, as mentioned, the reduction of losses. This requires more on the part of the antenna design, as it needs to be possible to achieve the optimal impedance needed by the amplifier not just at the fundamental frequency but at the second harmonic as well \[59\] in order to achieve the peak PAE of the amplifier. The issue is that power amplifiers generally prefer low impedance values, whereas antennas have a higher input impedance \[60\]. This leads to a compromise between optimal PAE of the amplifier and the radiation efficiency of the antenna to maximize the combined efficiency.

Besides providing impedance matching, antennas can fulfill other requirements of amplifiers designs. The Doherty power amplifier (DPA) \[61\], an amplifier type working efficiently in back-off, uses two separate amplifiers with a feeding network in between. The feeding network can be realized by utilizing two separate antennas \[62\], \[63\] or with a single antenna with two feed points \[64\]. However, achieving optimal operation of DPA-based AiA requires non-symmetric coupling between the amplifier ports, which is a challenging design target \[65\].

In the design of AiA arrays, a significant challenge is the ARC. As the antenna impedance varies with beam-steering, the integrated amplifiers become mismatched, leading to a deterioration in performance. This necessitates co-design efforts for both the array and the individual elements \[66\], \[67\]. These co-simulations are time-consuming, as 3D simulations of the antenna and circuit simulations of the amplifiers are both typically computationally intensive, especially for large arrays. Other challenges in the AiA design are cost and size \[68\].

2.4 Research equipment and software

This section describes the equipment and software used to conduct the research in this thesis.

2.4.1 Simulation softwares and numerical tools

Antennas in this thesis were designed using CST Studio Suite \[69\]. CST is a 3D EM simulation environment that can calculate S-parameters, near fields, and far-fields of an antenna structure. It features an integrated circuit simulation environment, which can be used to design matching networks for 3D simulated structures. CST also offers a Visual Basic interface for interacting with other software, specifically MATLAB in the context of this thesis.

AWR Microwave Office is a circuit simulation software that can be utilized for load-pull simulations of non-linear devices \[70\]. AWR was used to design the amplifier used in Publications [I] and [II], covering the entire
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process from transistor biasing to load-pull simulations of the final design.

MATLAB is a numerical tool originally developed for matrix calculations [71]. MATLAB was used in almost all publications for calculating the plotted results from either measured or simulated parameters. As far-field electric field data is conveniently presented in matrix format, MATLAB was used to calculate the total far-fields of antenna arrays from embedded element patterns with the input weights. MATLAB also offers a variety of numerical optimization tools in the Global Optimization Toolbox, which were used to optimize non-analytical optimization problems arising from load-pull accounted systems. Eigenvalue problems were also solved with MATLAB.

2.4.2 Measurement equipment

Manufactured amplifier prototypes were measured with load-pull using HITECH’s MT2000 load pull system [72]. MT2000 is an active load-pull system that enables measurements with single-tone, two-tone and modulated input signals. Both source and load side impedances can be swept simultaneously. In Publications [III] and [IV], MT2000 was used to perform load-pull with two-tone signal, and the load impedances at the two-tones were swept independently, a feat not easily possible with a passive system.

S-parameters of antennas were measured with a two-port VNA. The antennas in all of the publications of this thesis have more than two ports, thus the complete S-matrices were composed from multiple two-port measurements with a VNA, and the results were combined with MATLAB. Element patterns, as well as the full system operation with amplifiers in Publication [II], were measured with a full-spherical near-field scanner, the MVG Starlab 6-GHz system, at Aalto University [73]. An anechoic chamber was used to measure the antenna in [VI] at Aalto University.
3. Amplifier load-pull accounted antenna arrays

This chapter discusses the use of load-pull accounted amplifier modelling to enhance the operation of amplifier-antenna systems. The general system schematic is shown in Fig. 3.1. To clarify, the term "amplifier-antenna" used in Publications [I]-[IV] means a system, where an amplifier is feeding an antenna. This can mean an integrated setup, such as AiA, or a more typical case where a coaxial is used to connect the components.

Section 3.1 reviews the problems that require accounting for load-pull in literature. In Section 3.2, the iterative method used in Publications [I]-[IV] is described collectively. The rest of the chapter presents the work in Publications [I]-[IV]. In Publications [I] and [II], beam-steered EIRP is optimized with phase tuning. In Publication [III], IM3 distortion is minimized with amplitude and phase tuning when two independent beams at different frequencies are steered. In Publication [IV], IM3 distortion is minimized by utilizing load-pull accounted system simulation in the matching network design process using progressive phase shift.
3.1 Dynamic mismatch in antenna systems

4G systems have utilized simple radiating systems, where a single antenna cover a 120°-sector, and three such antennas with no overlapping coverage can cover the whole area surrounding the base station [74]. In the transition to 5G, the systems will need more complexity to support multiple-input multiple output and beamforming capabilities, and this requires the usage of a large number of antenna elements, with the majority of them having their own driver circuitry [75]. The shift will introduce load-pull of the driving amplifiers through antenna mutual coupling into the system as a real challenge, because utilization of isolators at each amplifier-antenna element interface will be infeasible [76]. This makes it mandatory to account for load-pull introduced non-idealities in the design and operation of the amplifiers and antennas.

The effects of load mismatch for amplifiers are many. [77] and [78] list issues with change in efficiency, power, linearity, as well as stability. At extreme mismatch conditions, outright breakage of the amplifier is a possibility to some degree [79]. For base stations and mobile devices, changes in loading can occur due to operational reasons, while with mobile device the changes in environment (e.g., hand placement) can also effect the matching. Amplifier protection against extremely high mismatch needs to be very effective at registering the possible power incoming to the amplifier output and managing to activate without affecting the amplifier’s operation otherwise [80]–[82]. For normal operation, the antenna itself can be designed to have lower mutual coupling to decrease load variation, with examples mentioned in Section 2.1.1. Designs for making the amplifiers themselves tolerant to dynamic mismatch include tunable matching networks [83]–[85], power recycling [86], [87] and supply control [88]. Additionally, using amplifier types inherently resistant to load variation, such as a balanced amplifier with load modulation [89]–[91] and DPA [92]–[94], with some degree of modifications, are common.

Pre-distortion (PD) can be considered a separate category for mismatch compensation, as it involves manipulation of the amplifier input signal [95]. To achieve this, a PD algorithm requires a behavioral model of the amplifier for which the coefficients are determined by monitoring the amplifier signal to make corrections. To account for mismatch, the amplifier model needs to take into account the input signal from both the input and output side [96]–[98]. By combining the amplifier model with the antenna S-parameters, the reflected wave into the output $a_2$ can be predicted, and proper pre-distortion can be applied to the input wave $a_1$ to produce the wanted output $b_2$ [99].

The majority of load-pull effects in research mostly concern amplifiers, which is only natural, as the amplifier is the component that is ultimately affected by the impedance match. Radiation characteristics, however,
Amplifier load-pull accounted antenna arrays are also affected by the changes in input waves to the antenna elements [100]. While the majority of effects are seen in SLLs, mismatch affects non-linearity as well [101], which is not typically attributed to antennas that are assessed at a single-frequency only. Therefore, research in future amplifier-antenna systems should be threefold, simultaneously taking into account the joint effects of the amplifier, the used signal profile, and wide-band antenna operation.

3.2 System simulation with load-pull accounted iteration

As all publications in this chapter utilize a similar iterative algorithm in solving the operation of the amplifier-antenna system in Fig. 3.1, the algorithm is covered collectively here. The reader should note that the input and output wave notation for the entire system is similar to the amplifier notation in Fig. 2.3 rather than the antenna notation in Fig. 2.2.

The algorithm requires separately measured or simulated results from the antenna array (or a multi-port antenna in general) and from the used amplifier or amplifiers, in the case of using multiple models. A similar iterative method to solve amplifier-antenna system operation has been employed in [100]. However, the approach used here utilizes direct load-pull data in a linearly interpolated look-up table format rather than polyharmonic distortion models (PHD) [102]. Generally, the standard PHD does not allow for large impedance swings in the model and fails when the impedance deviates significantly from the measured conditions. Still, there exists a load-dependent PHD to address this issue [103].

As described in Chapter 2, the behavior of an amplifier is dependent on the reflection, or rather the $a_2$-wave, at its output. Because the amplifier outputs in Fig. 3.1 are connected through the antenna array, this causes load-pull effect on the amplifiers. When the $b_{2,i}$-waves of the amplifiers change due to the active impedances, this causes the reflected $a_{2,i}$-waves also to change, causing the load impedances to change again. For a stable system there needs to be a solution where the waves $a_{2,i}$ and $b_{2,i}$ no longer change and both the single amplifier behavior under load-pull and the antenna array S-parameter relation of $a_{2,i}$- and $b_{2,i}$-waves is satisfied.

The algorithm uses a non-linear multi-input function set $B$ to model the amplifier and linear S-parameter and far-field data to model the antenna array. A single function $B$ calculates the output $b_{2,i}$ at a single frequency from the given input waves $a_1$ and $a_2$. Depending on the capabilities and file format of the measurement system providing the data, reflection coefficients $\Gamma_2$ can be used in place of $a_2$.

The algorithm iteratively calculates the outputs $b_2$, affecting the $a_2$ and influencing the load-pull until it converges to a solution. The algorithm is illustrated in Fig. 3.2. Initially, the outputs are assumed matched, and
the initial set of $b_2$ is calculated. Subsequently, $a_2$ is calculated from (2.6) using the antenna S-parameters, and this information is used to compute the next iteration of $b_2$. This process continues until the change between iterations is small. After convergence, the radiation pattern $E_{\text{arr}}(\theta, \phi)$ can be calculated with the converged $b_2$ using (2.20).

![Flowchart of the iterative algorithm used in the studies.][III]

In the publications, the algorithm has been used with single-tone and two-tone simulations, where only the excited tones were causing load-pull. In addition to the main tones, also harmonics and intermodulation products can be included as the inputs. Including more frequencies affects the complexity of the amplifier model steeply. As the data is complex valued, each added frequency adds two input arguments (the amplitude and the phase). Multidimensional linear interpolation is included in MATLAB and was tested up to six-dimensional data in gridded format, while non-gridded data consumes too much memory to be feasible.

3.3 **Effective isotropic radiated power optimization by phase tuning**

Publications [I] and [II] investigated whether EIRP could be improved with phase tuning in a beam-steered amplifier-antenna system. It is a well-known fact that the active impedance of the antenna elements changes during beam-steering. These changes in impedances can cause the feeding amplifiers to outside the optimal impedance range. This results in greater
decrease in the far-field power density, effectively decreasing EIRP when referencing to feeding amplifier inputs, than the isolated analysis of the radiation pattern of the antenna array would predict.

The initial simulation study in [I] predicted up to 0.7 dB improvement in EIRP with phase tuning in a system consisting of a 2x2 patch antenna array and a common collector BJT class-A amplifier. Both the array and the amplifier were designed for the study and matched to 50Ω. Fig. 3.3(a) shows the normalized EIRP of the reference case of geometry-based progressive phase shift, and in 3.3(b) the improvement provided by the phase optimization is presented. The enhancements are located at the edges of the steer range, with the highest values outside the –3-dB EIRP steer range of the reference, marked with black line in 3.3(b). In the broadside direction and the diagonals, no improvement is observed.

The analysis of what causes the improvement is not conclusive, but an examination of the behavior of the ARCs provides some insights. In Fig. 3.4(a), the ARC of element 1 for the optimized and reference cases are plotted, when the main beam is steered in the E-plane. The normalized amplitude of $b_2$-wave behavior of the amplifier model with respect to the reflection coefficient of the output is superimposed as a background. In Fig. 3.4(b), a closeup of the area of the ARCs is shown. The dashed line corresponds to the ARCs of the optimized case when the EIRP improves. In the improved steer directions the ARCs have shifted to the left on the chart, corresponding to the direction where $|b_2|$ is maximized.

This can be verified in Fig. 3.5(a) and (b), where the $|b_2|$ of the optimized feeds of elements 1 and 3 are plotted, respectively. The results are normalized to the reference $|b_2|$ in the same steer directions. The regions where EIRP has improved in Fig. 3.3(b) around $(\theta, \phi) = (60^\circ, 0^\circ)$, $(\theta, \phi) = (60^\circ, 90^\circ)$, $(\theta, \phi) = (60^\circ, 180^\circ)$ and $(\theta, \phi) = (60^\circ, 270^\circ)$ are also visible in Fig. 3.5(a) or (b) as regions where $|b_2|$ by over 0.1 dB. Note that the plot for elements 2 and
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Figure 3.4. ARC of element 1 on a complex plane when the main beam is steered in the E-plane. \((R, \phi) = (|\Gamma_{\text{out}}|, \angle \Gamma_{\text{out}}(\degree))\). [I]

Figure 3.5. Normalized \(|b_2|\) in different beam steer directions of elements (a) 1, and (b) 2. \((R, \phi) = (\theta(\degree), \phi(\degree))\). [I]

4 are similar to 1 and 3, respectively, but mirrored along the center line because of symmetry. Since \(b_2\) is the wave that excites the far-field electric fields of the antenna elements, an increase in \(|b_2|\) can be expected to result in increased power density in the far-field.

The findings in simulations were confirmed with measurements in [II]. The amplifier model and a similar antenna array used in [I] were manufactured. The amplifier was designed to maximize \(P_L\) with matching, and it can be verified from Fig. 3.6(b), that it was achieved, with the maximum marked with the cross only slightly off the center. A similar behavior of \(|b_2|\) is present in 3.6(a) compared to the simulated results in 3.4(a), but the change in \(|b_2|\) over the Smith chart is more pronounced. The behavior of the measured system’s reference case was confirmed to match that of the simulated system’s reference case, as shown in 3.7. The simulation results in 3.7(b) were obtained using the results from separately measured
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**Figure 3.6.** Measured amplifier prototype results with respect to the reflection coefficient: (a) the normalized $|b_2|$, and (b) the normalized $P_L$ of the amplifier. Maximum $P_L$ is marked with a cross. [II]

**Figure 3.7.** Normalized EIRP envelope of the reference case: (a) the measured result, with white crosses marking the measured directions, and (b) the simulated result. [II]

amplifier and antenna array prototypes.

In Fig. 3.8 are the results for the measured optimized case. The improvement is over 0.5 dB in large areas at the edge of the $-3$ dB steer range of the reference EIRP. Similar to the simulated case in Fig. 3.3 (b), there is practically no improvement in the broadside direction.

The study of the load-pull accounted system simulation in [I] and [II] shows, that phase tuning can be beneficial to the EIRP of an amplifier-antenna system. The ability to perform the analysis with separately simulated or measured amplifier and antenna models can provide time savings compared to a system simulated or measured as a whole. In future 5G systems, where the variation of the active impedance will pose a real concern, this analysis can be invaluable when designing the trans-
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Figure 3.8. Measured EIRP improvement with optimized phases when compared to reference feeding. [II]

Figure 3.9. Concept for IM3 minimization with feed tuning. [III]

mitting systems to spot possible issues related to amplifier and antenna co-operation.

3.4 IM3 minimization with feed control

To decrease interference caused by IM3 products, Publication [III] studied the possibility to use feed weight tuning in an amplifier-antenna system with simulations. By tuning the feeding weights of the amplifiers, IM3 products could be directed away from the main beam directions. The concept is presented in Fig. 3.9. As a single system can simultaneously service multiple users while beam-steering, inter-channel modulation can occur, causing interference to users. Some specifications, such as [104], might not however allow out-of-band emissions to increase above a certain limit even if they are directed away from the main beam direction.

The tuned weights were calculated using the load-pull accounting model with optimization in MATLAB for fixed amplifier and antenna array models. The optimization was compared to a linear model and a non-linear
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The linear model uses progressive phase shift at equal and constant input power across the amplifiers. The non-linear reference uses a non-linear matched model of the amplifier without load-pull data to optimize the powers and phases but accumulates errors due to lacking the varying impedance effects. The results of the system behavior for all three cases were evaluated with the load-pull accounted system model. The optimization target was to maintain signal-to-IM3-ratio (SI3R) above 40 dB in the beam directions while maximizing the radiated power densities at the main tones.

The simulation study used a four-element linear patch antenna array with amplifiers connected to each antenna. All amplifiers in the system were fed a two-tone signal at the frequencies 2.495 GHz and 2.505 GHz. Input power into each amplifier is equally split between the tones, but amplifiers can have different total input power. The phase of both tones in all amplifiers of the system can be adjusted freely. The beams at different main tone frequencies were independently swept on the H-plane of the array from \((\theta, \phi) = (60^\circ, 0^\circ)\) to \((\theta, \phi) = (60^\circ, 18^\circ)\) in 15° steps.

The load-pull data used to model the amplifiers were acquired with measurements from a commercial Freescale MMG38151BT1 amplifier mounted on a test printed circuit board (PCB) with coaxial connectors. The amplifier is inherently matched to 50 Ω. The measurement swept input power, as well as the load impedance at both tones independently, covering majority of the Smith chart. The independent sweeping of impedances is necessary because different frequencies can have drastically different loading, which cannot be adequately modelled with equal-value impedance sweeps. The measurement with two-tones does not provide the phase of the main tones or IM3 products, and thus AM-PM distortion is excluded from modelling. The output phases at main tones are assumed to be the same as the input phases. The phases of IM3 products IM3lo and IM3hi, denoted as \(\phi^{\text{IM3lo/IM3hi}}\), are calculated from the main tone phases using the formula:

\[
\phi^{\text{IM3lo/IM3hi}} = 2\phi^{f_1/f_2} - \phi^{f_2/f_1},
\]

where \(\phi^{f_1/f_2}\) represents phase shift at \(f_1 = 2.495\) GHz or \(f_2 = 2.515\) GHz.

An example of the resulting beam pattern for both beams pointed towards broadside is shown in Fig. 3.10. The linear reference has all beams collimated in the same direction, resulting in a maximum SI3R of 24.6 dB. Ideally, this could be increased to 40 dB by decreasing the amplifier input powers by 8 dB, but this would decrease the main tone power density by the same amount in the ideal case. It would also lower efficiency considerably, as the study was done with the amplifiers near compression. The non-linear reference has achieved an SI3R of 38.0 dB while maintaining the power density at similar levels to the linear reference. IM3 power densities have increased, but they are directed away from broadside, and
the main tone beam shapes have distorted. The optimization done with load-pull accounted model has managed to achieve the 40 dB SI3R ratio with similarly behaving IM3 beams as the non-linear reference.

Fig. 3.11 shows the complete results for the three cases and both main tone frequencies. The linear reference has a valley on the diagonal corresponding to the cases where all beams (main tone and IM3) point towards the same direction as the phases for them are equal. There are also a couple of cases just outside the diagonal where the phases at the main tones are selected so that the IM3 beams are pointed away from main beam directions. On the other hand, the non-linear reference has generally achieved the target and failed only slightly in a handful of directions. The failure is attributed to the absence of load-pull modeling, which indicates that the error in IM3 power is measured in decibels. The used antenna has a coupling of less than $-20$ dB between elements, a typical requirement for antenna arrays. The load-pull accounted model has achieved the target everywhere, showing that feed control can effectively lower IM3 interference in the main beam directions.

Fig. 3.12 shows the far-field power densities for the cases. Both the non-linear reference and the load-pull accounted model have distorted the $-3$ dB beam-steer range when compared to the linear reference. This is the trade-off when decreasing interference with feed weight tuning. It should be kept in mind, however, that depending on the orientation of the beams, the linear reference would need to decrease main tone power densities by up to 8 dB to achieve the required SI3R. From this perspective, the goal of achieving SI3R target and simultaneously retaining similar power densities is a great achievement with the non-linear reference and the load-pull accounted optimization. Both the non-linear reference and optimization with load-pull have managed to even increase power densities in some directions, with load-pull optimization performing better. The additional possibility to back off the system to further decrease interference without losing transmission is a good option to have in case SI3R needs to
Figure 3.11. SI3R for the calculated beam-steer configuration for 2.495 GHz (left) and 2.505 GHz (right). Linear reference (top), non-linear reference (middle) and load-pull accounted optimization (bottom). The results are normalized to 40 dB SI3R optimization target. [III]
Figure 3.12. Far-field power densities for the calculated beam-steer configuration for 2.495 GHz (left) and 2.505 GHz (right). Linear reference (top), non-linear reference (middle) and load-pull accounted optimization (bottom). The results are normalized to maximum of the linear reference. [III]
be increased even further.

Fig. 3.13 and 3.14 show SLL and HPBW of the three cases. Generally the outcome is the same as with the power density. Optimizing for SI3R deteriorates the beam integrity. However, this study is considering only a few design targets. Additional targets for beam integrity would alleviate these issues, but it ultimately depends on the system and the designer to know what performance metrics are most important. PAE of the non-linear reference and load-pull accounted optimization compared to the reference PAE are plotted in Fig. 3.15. The PAE of the linear reference was quite constant at 8.5% and was thus not plotted. As the amplifiers are driven quite close to compression in all cases, the PAE does not vary much.

The study in [III] shows that feeding weight tuning can be used to decrease IM3 interference in beam-steering. Taking into account the effect of coupling in the behavior of an amplifier in the form of a change in active impedance is crucial in accurate prediction of the system output. Even in the case of lower than −20dB coupling from element to element, not accounting for load-pull can cause errors in evaluating the linearity of the system.
Figure 3.14. HPBW for the calculated beam-steer configuration for 2.495GHz (left) and 2.505GHz (right). The linear reference (top), the non-linear reference (middle) and the load-pull accounted optimization (bottom). [III]

Figure 3.15. PAE of the non-linear reference (left) and load-pull accounted optimization (right) when compared to the linear reference. [III]
3.5 IM3 minimization with antenna input matching

In Publication [IV], a matching network for a 1x4 linear patch array was designed to minimize IM3 interference. The idea for utilizing a matching network arose from the load-pull measurement results of the amplifier used in [III], which showcased a minimum IM3 region on the Smith chart in the upper-left quadrant. Fig. 3.16 shows the load-pull results, where the minimum of the maximum IM3-to-carrier ratio (IM3C) is shown for sweeping the impedance at 2.495GHz while keeping the impedance at 2.505GHz matched. In the study, the two main tones have equal input power to the amplifiers, and the beams are pointed towards the same steer direction during beam-steering, emulating a modulated signal beam of a 10MHz channel.

Three different matching networks were designed. The reference, later REF, was designed without system simulations by minimizing antenna ARCs with the two main tone beams pointed towards broadside with equal power to the elements. Two optimized networks, later OPT, were designed with one having the beam pointed similarly to broadside and the other with the beam pointed towards the beamsteer edge of 40° with progressive phase shift. The systems are driven similarly with progressive phase shift and have an identical amplifier and antenna model apart from the matching network. The matching network is the same for all antenna elements for a given case.
The optimized matching networks were designed by utilizing co-simulations, performing 3D EM and circuit simulations on CST, and calculations with the load-pull accounted system model in MATLAB. The EM structure was designed with EM-simulations in CST and was then kept fixed. The optimization of the matching network was then performed by calculating the S-parameters and far-field patterns for different matching networks with CST and assessing the system performance with MATLAB. The performance metric for the systems is the maximum IM3C in the radiated far-field. As an additional requirement, both main tone beams need to have the same power densities within 0.2 dB for the result to be valid.

The matching network topology and the final component values of the three cases are shown in Fig. 3.17. Fig. 3.18 shows the ARCs of the three cases when the beams are steered in the \( \theta = [-40^\circ, 40^\circ] \) (H-plane) steer range. Only ARCs of elements 1 and 2 are shown because of symmetry. With the beam pointed towards broadside, ARCs of REF are quite well matched in the center. As the beam is steered, the ARCs move into the lower-right quadrant, which is disadvantageous with regard to IM3C, as can be estimated from Fig. 3.18. ARCs of both OPT results are sturdily in the upper-left quadrant, the favorable region with regard to IM3C.

Fig. 3.19 shows the beam-steer envelopes for IM3C and the minimum far-field power density \( S \). To account for the ideal behavior of IM3 output power of an amplifier, the threshold curve TC is plotted with the definition of

\[
TC = IM3C_{REF} - 2(S_{REF} - S_{OPT}).
\]  

TC describes how the IM3C of REF would ideally behave if the input power to its amplifiers was adjusted so that \( S_{REF} \) matched \( S_{OPT} \). TC is included as backing-off an amplifier is a good way to improve linearity, which TC emulates. Fig. 3.19 (a) shows that OPT improves IM3C by 1.46 dB in the broadside direction when compared to REF directly. As the main tone power is decreased slightly, comparing OPT to TC gives an improvement of 1.32 dB. For both REF and OPT, IM3C degrades monotonically when the beam is steered, resulting in a difference of 1.24 dB in IM3Cs at the steer edge. In Fig. 3.19 (b), the result at the steer edge is better, with
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Figure 3.18. ARCs of the three studied cases with the amplifiers accounted. ARCs of elements 3 and 4 are excluded because of symmetry. [IV]

Figure 3.19. The beam steer envelopes of the three cases. (a) OPT designed with broadside beam, and (b) OPT designed with the beam pointed towards the steer edge $\theta = 40^\circ$. Power densities are normalized to REF maximum (broadside). [IV]

the difference between OPT and REF IM3Cs being 1.77 dB. Interestingly, the IM3C of OPT is quite constant and having a minimum at around 30° steer angle. In the broadside, the power density of OPT is decreased by 0.29 dB, resulting in a worse result when compared to TC, even though the absolute IM3C level is lower than when optimized with a broadside beam. IM3C, however, has improved everywhere with both optimized matching networks when compared to REF directly or when compared to TC.

The simple study in [IV] proposes a method to design matching networks for antenna arrays that can improve the system linearity. The method utilizes the load-pull model of an amplifier to evaluate the system performance with the active impedances of the antenna array affecting the
amplifiers. As modern systems have high standards for linearity and PD being responsible for achieving most of it, small contributions to linearity could alleviate the burden from PD.
4. Optimizing realized gain

This chapter presents a method in [V] to optimize the realized gain of an antenna array driven with variable gain amplifiers, whose output impedance depends on the gain, and a cluster array concept in [VI] which utilizes mutual coupling to achieve high coverage gain over a wide band. Section 4.1 describes Rayleigh Quotient (RQ), which is used in both publications.

4.1 Rayleigh quotient in optimizing antenna feeding

In recent years, the use of RQ [105] has become a useful tool in optimizing multi-port antennas with active feeding. For example, the cluster antenna method [106] utilizes RQ to increase impedance matching bandwidth using multiple antenna elements with different resonance frequencies. The method utilizes coupling between elements to cause passively mismatched elements to be matched in an active feeding scenario.

RQ is used in the min-max theorem [107], and has the form

\[
RQ = \frac{x^H M x}{x^H x},
\]

(4.1)

where \( x \) is an \( n \times 1 \) complex vector and \( M \) is an \( n \times n \) Hermitian matrix. The minimum value of RQ is the minimum eigenvalue \( \lambda_{\text{min}} \) of matrix \( M \), and it can be achieved with the corresponding eigenvector \( x_\lambda \). Conversely, RQ can be maximized by choosing the eigenvector corresponding to the maximum eigenvalue \( \lambda_{\text{max}} \), which is the maximum value of RQ.

As an antenna is a linear device, various parameters of antennas can be formulated as RQ from which the maximum or minimum results can be easily solved. RQ has been used in minimizing TARC [106], [108]–[110], reducing user effect [111]–[113], maximizing radiation and total efficiencies [114], [115], and maximizing realized gain [116], [117]. The work done in [V] and [VI] has employed RQ in improving the performance of their respective systems.
4.2 Accounting output impedance variation of a feeding amplifier

As modern beamforming systems require amplitude and phase control, different methods are researched to control both. As amplifiers are driven at maximum output power for high efficiency, the simple method of using attenuators to control the output wave amplitudes is not ideal. Changing the supply voltage of an amplifier changes its gain and can be used to control output power without backing off from the region of high efficiency [118]. Using supply voltage tuning in an amplifier to control the gain will change the output impedance of the amplifier, which will affect the matching to a constant 50-Ω load, such as an antenna. Publication [V] studied the effect of amplifier output impedance variation on the realized gain of an antenna array fed with an amplifier at each port. A novel method utilizing S-parameter re-normalization with eigenvector optimization was proposed.

The method to optimize the realized gain of an antenna array with \( n \) ports in the described situation utilizes S-parameters of the system \( S_{SYS} \) comprised by the transmitting array TX and the receiving antenna RX. \( S_{SYS} \) is defined as

\[
S_{SYS} = \begin{bmatrix} S_{TX} & S_{TR}^T \\ S_{TR} & S_{RX} \end{bmatrix}, \tag{4.2}
\]

where \( S_{TR} \) is the transmission vector from TX to RX and \( S_{TX} \) and \( S_{RX} \) are the typical S-matrices of TX and RX, respectively. When the TX and RX are in the far-fields of each other, \( S_{TR} \) effectively describes the far-field pattern of the TX in the direction of RX. \( S_{TR} \) includes different non-idealities related to the RX, but they do not affect the analysis related to the far-field of TX.

Fig. 4.1 (a) shows the general system diagram and (b) shows the diagram with the matrix partitions that comprise \( S_{SYS} \) as well as the used wave notation. The transmission efficiency \( \eta_{TR} \) is the ratio of power received at the RX port \( P_{rec} \) and the total available power at the antenna array \( P_{av} \) when \( a_{RX} = 0 \)

\[
\eta_{TR} = \frac{P_{rec}}{P_{av}} = \frac{b_{RX}^H b_{RX}}{a_{TX}^H a_{TX}} = \frac{a_{TX}^H S_{TR}^H S_{TR} a_{TX}}{a_{TX}^H a_{TX}}, \tag{4.3}
\]

where \( a_{TX} = [a_1, \ldots, a_n]^T \). Equation (4.3) can be maximized with the proper eigenvector. \( \eta_{TR} \) is directly related to the realized gain of TX, as RX is kept fixed.

The solution for maximizing realized gain with the eigenvector feeding alone is not, however, sufficient in the studied case. The S-parameter matrix is formed from measurements (or simulations) with 50-Ω impedances at the system ports. In the case of varying port impedances \( Z_{pi}, i = 1, \ldots, n \), the S-parameters with 50-Ω port impedances are not correct. The S-matrix
S can, however, be re-normalized to any set of port impedances by utilizing the Z-matrix $Z$ of the system \[ \tilde{S} = F(Z - Z_p)(Z + Z_p)^{-1}F^{-1}, \] where $\tilde{S}$ is the re-normalized S-matrix,

$$Z_p = \begin{bmatrix} Z_{p1} & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & Z_{pn} \end{bmatrix}$$

and

$$F = \begin{bmatrix} \frac{1}{2\sqrt{Z_{p1}}} & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & \frac{1}{2\sqrt{Z_{pn}}} \end{bmatrix}.$$

The Z-matrix can be solved from (4.4) using a measured S-matrix with known port impedances. From the re-normalized S-matrix of the system $\tilde{S}_{SYS}$, the re-normalized transmission vector $\tilde{S}_{TR}$ can be extracted for the situation with new port impedances.

By utilizing (4.3) and (4.4) iteratively, the method takes into account the variation of port impedances with respect to gain. In full, the method starts by calculating the initial eigenvector for maximizing realized gain with (4.3) from the initial S-matrix of the system. The port impedances of the amplifier are then adjusted to match the output impedances $Z_{pi}$ of the amplifiers, which vary because of the change in supply voltage being used to control amplifier gain. The S-matrix of the system is then re-normalized with (4.4) using the new impedances $Z_{pi}$. The eigenvector is then calculated again with (4.3), and re-normalization is then reapplied until the solution converges.

The method was tested with a simulated example case of a linear 4-element vertical patch antenna array as TX. TX elements were matched to...
50Ω at 3.1GHz. In the example, TX and RX were placed in the far-fields of each other. RX was then steered from the broadside direction of TX in 10° steps. The 50-Ω S-parameters were simulated in each steer direction. The situation is shown in Fig. 4.2(a), and in (b), the used amplifier output impedance behavior is shown. The output impedance is always purely real and has a minimum value of 70Ω at maximum output power (gain) and increases monotonically when power is lowered. In the study, the amplifiers are assumed to be connected directly to the antenna element ports, making the antenna port impedances equal to the amplifier output impedances.

Fig. 4.3 shows the simulated S-parameters of TX with 50Ω port impedances and the improvement of η_{TR} when compared to η_{TR} of progressive phase shift at maximum gain. The antenna elements are matched on a wider band than typical two-layer PCB patches because the patches are constructed of two single-layer PCBs with a 16mm air gap between them. Larger ground clearance is the reason for increased matching bandwidth. In Fig. 4.3
(b) it can be seen that the proposed method always improves realized gain over that of progressive phase shift in the steer range of \([0^\circ,40^\circ]\). Around the center frequency of 3.1GHz, there are a few areas of minor deterioration at extreme steer angles, but the method could be enhanced to select the progressive phase shift if the proposed method does not improve transmission. On average, the realized gain improves by 3.5dB in the studied case.

As novel feeding schemes are becoming more popular for multi-port antennas, the non-idealities of amplifiers need to be taken into account. Publication [V] proposes a method to take into account the variation of output impedance with gain control of an amplifier in feeding an antenna array for maximum gain. As the method takes into account the behavior of both components, it is more beneficial compared to the analysis of the amplifiers and the antenna separately.

### 4.3 Leveraging mutual coupling for high coverage realized gain

Publication [VI] proposes a flexible cluster array concept to improve the coverage gain for wide-band operation. The concept utilizes non-identical antenna elements and leverages mutual coupling to enhance the matching for a wider bandwidth. To address the challenges of 5G, wide-band impedance matching alone is not sufficient, as mobile devices are also required to be capable of beamforming at millimeter wave frequencies. Utilizing the limited volume for antennas in mobile devices with maximum efficiency is the key to meeting the stringent requirements. The concept achieves over 6 dBi at the 50%-tile cumulative distribution function (CDF) of maximum realized gain on the 24.5–29.5 GHz band.

The concept relies on utilizing eigenvector feeding to optimize the realized gain in beam-steering. To formulate RQ, the excited electric field by a single-fed radiating element \(E_i(\theta,\phi)\) is calculated from the normalized electric field \(E_{0i}(\theta,\phi)\) scaled with the complex incident wave of the element \(a_i\) by

\[
E_i(\theta,\phi) = E_{0i}(\theta,\phi)a_i. \tag{4.5}
\]

When all the element fields \(E_i(\theta,\phi)\) are combined into a vector \(\mathbf{E}(\theta,\phi)\), then the far-field power density \(S(\theta,\phi)\) can be expressed as

\[
S(\theta,\phi) = \frac{\mathbf{E}^H(\theta,\phi)\mathbf{E}(\theta,\phi)}{2\eta} = \frac{a^H(\theta,\phi)\mathbf{E}^H(\theta,\phi)\mathbf{E}(\theta,\phi)a(\theta,\phi)}{2\eta}. \tag{4.6}
\]

Finally, the formulation for realized gain \(G\) is

\[
G^r = \frac{4\pi r^2 S}{P_{\text{inc}}} = \frac{4\pi r^2 a^H E^H E a}{\eta a^H a}. \tag{4.7}
\]
where angle dependency has been dropped for brevity. Equation (4.7) can be maximized with the proper eigenvector.

To analyze the performance of 5G mobile devices, 3GPP has created a standard for spherical coverage EIRP [119]. The standard evaluates EIRP in contrast to the realized gain of an antenna, but the metric is extended here to be used as a way to analyze antennas. Coverage gain is the CDF of the realized gain over the steer angles at a single frequency. Mathematically, the coverage gain \( F(G_r) \) is described as

\[
F(G_r) = P(G_r(\theta, \phi) \leq G_r),
\]

i.e., how large a portion of the beam steer envelope of the realized gain \( G_r(\theta, \phi) \) is less than or equal to the value \( G_r \).

The concept was demonstrated with simulations with patch antennas, as they are simple elements with a low profile and easy to produce, making them ideal for antennas in mobile devices. The conventional four-element patch antenna array geometry used as a reference and the eight-element cluster patch antenna array are presented in Fig. 4.4 with the dimensions in Table 4.1. Both antennas were simulated with Rogers-4003 substrate (\( \varepsilon_r = 3.5, \tan \delta = 0.002 \)) of thickness 0.508 mm and had a probe feeding through the substrate for each element. The common element-to-element distance \( \lambda_1 \) varies from 0.45\( \lambda \) to 0.55\( \lambda \) with increasing frequency while the cluster array distance \( \lambda_2 \) varies from 0.14\( \lambda \) to 0.17\( \lambda \).

Fig. 4.5 (a) and (b) show the simulated reflection and transmission coefficients of the cluster array, respectively. The elements are designed to have different resonance frequencies covering the whole 24.5-29.5 GHz.
Table 4.1. Dimensions of the simulated structures. [VI]

<table>
<thead>
<tr>
<th>Parameters</th>
<th>$W_{\text{sub}}$</th>
<th>$L_{\text{sub}}$</th>
<th>$d_1$</th>
<th>$d_2$</th>
<th>$W$</th>
<th>$F_1$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Values (mm)</td>
<td>5.5</td>
<td>24</td>
<td>5.5</td>
<td>1.7</td>
<td>1</td>
<td>0.45</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Parameters</th>
<th>$F_2$</th>
<th>$L_1$</th>
<th>$L_2$</th>
<th>$L_3$</th>
<th>$L_4$</th>
<th>$L_5$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Values (mm)</td>
<td>0.3</td>
<td>3.1</td>
<td>2.85</td>
<td>2.9</td>
<td>2.65</td>
<td>3</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Parameters</th>
<th>$L_6$</th>
<th>$W_T$</th>
<th>$F_T$</th>
<th>$D$</th>
<th>$h$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Values (mm)</td>
<td>2.75</td>
<td>2.9</td>
<td>0.55</td>
<td>0.15</td>
<td>0.1</td>
</tr>
</tbody>
</table>

Figure 4.5. Results of simulated structures (a) reflection coefficients and TARC envelope of cluster array, (b) transmission coefficients of cluster array, and (c) S-parameters of the reference patch array. [VI]

band to allow tuning the active matching across the band. The simulated S-parameters of the reference array are presented in Fig. 4.5 (c). There is a significant difference in coupling, the cluster array reaching up to $-5$ dB coupling between elements, while the reference only up to $-20$ dB. Fig 4.5 (a) and (c) also show the calculated TARC envelopes of the respective simulated structures, which are achieved with eigenvector feeding.

Fig. 4.6 (a) shows the simulated realized gain CDFs of the reference and the cluster array. CDFs are calculated for the upper hemisphere only.
because of the directive nature of the patch antenna. Eigenvector feeding has been utilized for both, but as the reference cannot be actively matched on a wide band, it falls considerably short at 27.5, 28.5, and 29.5 GHz. At the other frequencies, the reference can keep TARC sufficiently low, as can be seen in Fig. 4.5. Both antennas, however, have relatively equal gains.
on the whole band, as illustrated in Fig. 4.6 (b). This tells us that the difference in realized gain arises from the active matching. The cluster array has 5 dB better realized gain at 50% CDF for 29.5 GHz in simulations.

The simulated antennas were manufactured, and the prototypes are shown in Fig. 4.7. The prototypes were manufactured with an additional layer on the backside and made wider and higher to facilitate feeding lines and connectors for measurements. The substrate of the additional layer is 0.1 mm high R-5785 ($\varepsilon_r = 3.55, \tan\delta = 0.004$). The added area on the PCBs is covered with metal, leaving the same area as in simulations around the antenna for radiation as is in the simulated structures.

The individual element patterns and array S-parameters were measured for the reference and the cluster array. The results were processed in MATLAB to produce the full array results, rather than measuring the full array. The CDF of the realized gain is plotted in Fig. 4.8, showing good agreement with simulated results in Fig. 4.6. The cluster array has over 6 dB coverage gain at the 50%-tile CDF at all frequencies. The reference array, on the other hand, has only 2 dB realized gain at the 50%-tile at 29.5 GHz.

Fig. 4.9 shows the realized gain of individual steered beams for the reference and the cluster array. The results are equal at the resonant frequency of the reference 24.5 GHz, but at 29.5 GHz, the cluster array has on average 4.5 dB higher realized gain. Table 4.2 contains the feeding weights for the measured cluster array prototype that maximize the realized gain in different steer directions.

Publication [VI] offers the cluster array concept as a solution to solve wide-band steering requirements of 5G. The concept was demonstrated with patch antennas at millimeter waves, but the concept offers high adaptability, which can be utilized with other antenna types as well. As tight spacing of mobile device antennas and the requirement for beam-steering will inevitably lead to increased coupling between elements, the
Table 4.2. Example optimized feeding weights for individual steered beams. [VI]

<table>
<thead>
<tr>
<th>port</th>
<th>$\theta = -60^\circ$</th>
<th>$\theta = -30^\circ$</th>
<th>$\theta = 0^\circ$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0.43$\angle$ -74.5</td>
<td>0.19$\angle$ 119.2</td>
<td>0.25$\angle$ 17.8</td>
</tr>
<tr>
<td>2</td>
<td>0.41$\angle$ -86.6</td>
<td>0.24$\angle$ 101.1</td>
<td>0.30$\angle$ 15.5</td>
</tr>
<tr>
<td>3</td>
<td>0.53$\angle$ 15.0</td>
<td>0.35$\angle$ 125.3</td>
<td>0.35$\angle$ -31.9</td>
</tr>
<tr>
<td>4</td>
<td>0.17$\angle$ -72.5</td>
<td>0.34$\angle$ 54.4</td>
<td>0.47$\angle$ -108.5</td>
</tr>
<tr>
<td>5</td>
<td>0.37$\angle$ -164.8</td>
<td>0.45$\angle$ -135.6</td>
<td>0.38$\angle$ -27.3</td>
</tr>
<tr>
<td>6</td>
<td>0.26$\angle$ 74.7</td>
<td>0.45$\angle$ 142.9</td>
<td>0.52$\angle$ -105.6</td>
</tr>
<tr>
<td>7</td>
<td>0.21$\angle$ 87.7</td>
<td>0.30$\angle$ 63.2</td>
<td>0.12$\angle$ 69.4</td>
</tr>
<tr>
<td>8</td>
<td>0.31$\angle$ 0</td>
<td>0.42$\angle$ 0</td>
<td>0.29$\angle$ 0</td>
</tr>
</tbody>
</table>

Figure 4.9. Individual steered beam with optimal weights for both arrays at (a) 24.5 GHz, and (b) 29.5 GHz. [VI]

The cluster array concept can be used to address this by leveraging the coupling for improved spherical coverage gain.
5. Summary of Publications

**Publication I: Amplifier-Antenna Array Optimization for EIRP by Phase Tuning**

In this work, the effects of load-pull on an amplifier are accounted to optimize the EIRP of a transmitting beam-steered antenna array by phase tuning. The effect of load-pull is accounted in the system with an iterative model utilizing single-tone load-pull simulation results of the amplifier at a single input level and antenna array S-parameters. The algorithm is implemented in MATLAB. The maximum attained EIRP improvement with phase tuning in the −3-dB EIRP steer range of the reference is 0.6dB.

**Publication II: Analyzing and Optimizing the EIRP of a Phase-Tunable Amplifier-Antenna Array**

In this work, an amplifier prototype and antenna array prototypes are measured separately, and the separate results are combined with co-simulations to predict the combined behavior, which is verified with measurements. The work confirms the system model devised previously, as well as the result of EIRP improvement with phase tuning when compared to progressive phase shift.

**Publication III: Over-the-Air Suppression of Third-Order Intermodulation in a Two-Beam Steered Amplifier-Antenna Array**

IM3 products are minimized in the beam-steer directions with feed control by accounting the effects of amplifier load-pull with co-simulations. Compared to a method that accounts for amplifier behavior only in the matched case, the load-pull accounted method provides a maximum improvement
Publication IV: Minimizing Intermodulation Distortion in a Transmitting Antenna Array with Matching

In this work, an antenna array matching network is tuned to minimize transmitted intermodulation distortion with co-simulations of the array and the amplifier model. Utilizing co-simulations in the design process allows the effects of load-pull of the amplifier to be accounted and decreases the IM3C by a maximum of 1.7 dB and a minimum of 1.24 dB over the reflection-minimizing matching reference in the studied cases.

Publication V: Optimizing RF Efficiency of a Vector-Modulator-Driven Antenna Array

This work presents a method to optimize transmission efficiency by controlling feeding weights when the transmitting array is driven by variable-gain amplifiers, which have an output impedance varying with the gain. The method utilizes the known output impedance behavior of the amplifiers in re-normalizing the S-parameter matrix of the system, which is used iteratively to find the optimal solution. In the case study with a 1x4 patch antenna array, accounting for the impedance mismatch improves the transmission efficiency by 0.9 dB in comparison to just accounting for the antenna behavior, and a 3.5-dB improvement when compared to progressive phase shift feeding.

Publication VI: Coupled Array of Diverse Elements for Wideband High Spherical Coverage

In this work, a concept of a wide-band beamforming-capable cluster array is demonstrated. By utilizing feed tuning and patch antenna elements with different resonant frequencies, the cluster array achieves superior realized gain when compared to an array with patch elements with identical resonance frequencies. The bandwidth of the cluster array consisting of eight elements is 24.5-29.5 GHz whereas the reference array with four elements has 24.5-29.5 GHz when spherical coverage gain is considered. The concept relies on high coupling between elements to allow for reconfigurable matching with feed tuning.
6. Conclusion

This doctoral thesis investigated the effects of mutual coupling and amplifier non-idealities in beam-steered antenna arrays. The primary scientific contributions of the work include the study of optimizing radiation in amplifier load-pull affected array systems and realized gain in antenna arrays with high inter-element coupling.

Publications [I]-[II] studied how transmitting amplifier-antenna systems can be optimized for EIRP when the feeding amplifiers are affected by load-pull of the active impedance of the antenna elements. By utilizing an iterative method with a new type of data set, the system behavior can be modelled accurately and used to calculate the optimal feeding phases for maximizing radiation in different beam steer directions. The studies show that system-level simulations are beneficial for overall performance and can be done with more cost-effective behavioral modelling with the new datatype rather than with costly circuit simulations.

Publications [III]-[IV] studied the effect of load-pull on IM3 behavior of beam-steered antenna arrays. A similar iterative method as in [I]-[II] was used with two-tone excitation. Publication [III] found that feed tuning could be beneficial in controlling IM3 interference by using feed tuning to radiate the interference in another direction compared to the intended signal direction. The main tone powers were not significantly affected, while the interference could be drastically decreased. Publication [IV] studied the use of system-level analysis when designing the matching network of the antenna in minimizing IM3 interference in the main beam direction and found the analysis to be an effective method.

Publication [V] devised a method to take into account the output impedance variation of a variable-gain amplifier. As modern transmitters will require the ability to control antenna feeding weights more freely and still achieve high efficiency, the amplifier non-idealities will be more prominent. The method utilized re-normalization of port impedances of the antenna system S-matrix in the system modeling and eigenanalysis in calculating the optimal feeding weights for maximizing transmission efficiency in a beam-steered array.
In [VI], the cluster concept offering wide-band matching by utilizing diverse antenna elements was extended to a beam-steered array. The demonstrated concept can provide simultaneously wide bandwidth at higher frequencies, beam-steering capability, and high gain required by 5G mobile device antennas. The example cluster array utilizes eight elements compared to four elements of the reference traditional array to extend spherical coverage gain bandwidth from 24.5-26.5 GHz to 24.5-29.5 GHz.

The work done in this thesis demonstrates the need to take into account the interaction of different components in the design of radiating systems, as well as the bandwidth potential of antenna arrays with high mutual coupling. As 5G communication systems mature, the analysis of the whole system simultaneously over individual components will surely be needed in the design processes. The research in this thesis is likely to contribute to the work in that field.

Future research should evaluate when a system will benefit from system analysis. The typical approach to designing large antenna arrays is to use unit cell simulations, which assume an infinite array, as it is sufficiently accurate and more computationally efficient compared to finite array simulations. A similar approach is probably true for modeling load-pull accounted systems, as the effect will average over large arrays. The size of arrays in mobile devices cannot, however, ever be as large as the size in base stations, which might ultimately determine the research in this thesis to be more usable in mobile device design.
References


References


References


As future mobile communication systems are to utilize beam-steering capable systems with high gain, the problem of the varying active impedance will most likely become uncircumventable with bulky isolators used traditionally. This will make the traditional system design, where each component is designed separately, obsolete, and require more involved co-design of separate components. This thesis investigates the problem of varying active impedance in antenna arrays to the feeding amplifiers in transmitters. The problem is approached by utilizing co-simulations where the amplifiers and antennas are simulated together.