Active Microwave Tagging Systems

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# Contents

Contents ............................... i
Abstract ................................ 5
Zusammenfassung ......................... 7

1. Introduction ......................... 9
   1.1 Motivation ....................... 10
   1.2 Project Description .............. 10
   1.3 Organization ..................... 10

2. Classification of Tagging Systems ... 13
   2.1 Radio Frequency Ranges and Licensing Regulations 13
   2.2 Inductively Coupled RFID Systems .................. 15
      2.2.1 Operation Principle of Inductively Coupled RFID Systems 15
   2.3 Surface Acoustic Wave Transponder Systems ....... 27
   2.4 Microwave Tagging Systems ................. 27
      2.4.1 Remotely Powered Transponders ............. 27
      2.4.2 Battery Powered Transponders using Passive Backscatter Modulation 30
      2.4.3 Battery Powered Transponders using Active Backscatter Modulation 34
   2.5 Summary ................................ 35

3. Active Backscatter Tagging System ... 37
   3.1 Basic Operation Principle .......... 37
   3.2 Downlink Communication ............. 38
      3.2.1 Circular-Polarization-Shift-Keying (CPSK) Modulation 39
      3.2.2 Signal Representation and Analysis of CPSK Modulation 41
   3.3 Uplink Communication ................. 43
      3.3.1 Theory of Modulated Scatterers .......... 44
      3.3.2 System Topologies for Backscatter Modulation 49
      3.3.2.1 Single Sideband Modulator on the Transponder 51
      3.3.2.2 In-phase/Quadrature-phase Demodulator at the Reader 53
3.3.2.3 Heterodyne Receiver at the Reader 55
3.3.3 System Topology of the Active Read/Write Tagging System Demonstrator operated in the 2.4 GHz ISM Band 58
3.4 Summary 60

4. System Setup 61

4.1 RF Detector 61
4.1.1 Detector Circuit Design, Detector and Diode Equivalent Circuits 61
4.1.2 Voltage Sensitivity 67
4.1.3 Tangential Signal Sensitivity 70
4.1.4 Diode Measurements 74
4.1.5 Matching 76
4.1.6 Design of the Tag Demodulator 80
4.1.7 Bit Error Rate Measurement 82
4.1.8 Voltage Doubler 85

4.2 Antennas 87
4.2.1 Aperture-Coupled Patch Antennas for Tagging Applications 88
4.2.2 Definition of Antenna Parameters and Antenna Requirements for the Active Tagging System 89
4.2.3 Analysis of a Linearly-polarized Aperture-Coupled Patch Antenna 94
4.2.4 Circularly Polarized Antennas with Switchable Polarization Sense 96
4.2.5 Antenna Parameter Design Rules 98
4.2.6 Measured Radiation Characteristics 101
4.2.7 Tag Antenna 105
4.2.8 Reader Antenna 110
4.2.9 Phased Array Antenna for the Reader 111

4.3 Active Modulator 115
4.3.1 Transmission Distance in the Downlink and Uplink Communication 115
4.3.2 Active Modulator consisting of Hybrid Circuits 118
4.3.3 Active Modulator as MMIC 123

4.4 RF Synthesizer 127
4.5 Signal Processing in the Downlink/Uplink Communication 133
<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>4.5.1 Downlink Communication</td>
<td>133</td>
</tr>
<tr>
<td>4.5.2 Uplink Communication</td>
<td>135</td>
</tr>
<tr>
<td>4.6 Summary</td>
<td>141</td>
</tr>
<tr>
<td>5. System Measurements</td>
<td>143</td>
</tr>
<tr>
<td>6. Conclusions and Outlook</td>
<td>149</td>
</tr>
<tr>
<td>7. Appendix</td>
<td>151</td>
</tr>
<tr>
<td>7.1 Anticollision Procedures</td>
<td>151</td>
</tr>
<tr>
<td>7.2 Multiple Access Procedure for the Active Tagging System</td>
<td>152</td>
</tr>
</tbody>
</table>

References 155

Abbreviation and Symbols 159

Acknowledgements 161

Curriculum Vitae 163

Publications 165
Abstract

Microwave tags are small low cost microwave transmitter/receivers suited to transmit identification or data to a reader over a small distance of 0.5 to 10 m. Typical applications are: contactless identification of persons, containers, parcels, railway cargo carriages. Mainly three different tagging system types are distinguished: (1) remotely powered fully passive tags, (2) battery powered tags applying a passive reflex modulator and (3) active tags where active RF circuits are used on the tag to increase transmission distance. The development and investigation of an active tagging system is the topic of this thesis. In particular, the objectives of this work include the development of lowest power read/write tags for the 2.4 GHz ISM band and the increase of transmission distance at a maximum RF power of 10 mW EIRP.

In a first part, an overview on different RFID and tagging systems is presented. Particular interest is given to the investigation of battery powered transponders applying active backscatter modulation. This modulation scheme is applied in the experimental setup of an active tagging system which will be discussed in the following. A further part of this work is dedicated to the discussion of the circuit topologies and modulation schemes in the downlink and uplink communication. In order to reduce demodulation complexity and DC power consumption on the tag, circular-polarization-shift-keying modulation is used for the downlink communication where data or commands are transmitted from the reader to the tag. In the uplink communication active backscatter modulation is applied for reading data from the transponder. The transponder’s response signal is typically double sideband modulated and thus a simple homodyne receiver at the reader would cause signal cancellation of the demodulated baseband signal. Circuit topologies like a single sideband modulator on the transponder or an I/Q and a heterodyne receiver at the reader are discussed which eliminate this signal cancellation phenomenon. In the experimental system setup a heterodyne receiver topology is used that is additionally optimized for the demodulation of low data rate signals. The baseband modulation scheme in the uplink communication is BFSK modulation.

In the next part of the work, the key circuits and components of the experimental system setup including RF detector circuits, aperture-coupled microstrip patch antennas, the active modulator of the transponder and the RF synthesizer used in the reader are presented. Zero bias Schottky detector diodes are applied for the RF detector circuits. A voltage sensitivity between 20 to 25 mV/μW and a tangential signal sensitivity of about -60 dBm at a video bandwidth smaller than 10 kHz are typical values that could be achieved by the used detector diodes. Aperture-coupled microstrip patch antennas are investigated because they allow to separate the feed layer from the radiating patch, resulting in a high front-to-back ratio. In particular, the multi-layered structure is optimum for developing small and compact transponders. Circularly polarized aperture-coupled antennas with switchable polarization sense are presented which show a polarization isolation exceeding 15 dB across the entire 2.4 GHz ISM band. For the active modulator a simple circuit topology consisting of a balanced upconverter preceded by an RF amplifier is chosen. A conversion gain of about 6 dB is feasible at a DC power consumption of 10 mW. Off-the-
shelf components as well as MMICs are used to built the active modulator. The RF synthesizer of the reader has to show frequency hopping capabilities. A circuit topology for the RF synthesizer is investigated which consists of a phase-locked-loop circuit whose reference signal is generated by a direct digital synthesizer performing frequency hops.

The last part of the work is dedicated to measurements with the experimental active tagging system. At an RF power of 10 mW EIRP a transmission distance of about 4.2 m is achieved in the downlink communication. The transmission distance in the uplink communication is slightly smaller. A communication link scenario in a multiple-reader multiple-tag environment is also investigated with the experimental system setup. However, the tagging system is not capable to perform a true multiple access communication since no anticollision procedures are implemented yet.
Zusammenfassung


Im nachfolgenden Teil der Arbeit werden ausgewählte Schaltungen und Komponenten des Prototyp-Objektidentifikationssystems vorgestellt. Es sind dies HF-Detektorschaltungen, aperturgekoppelte Mikrostreifenantennen, der aktive Modulator des Datenträgers sowie der HF-Synthesizer der Ablese-Einrichtung. Für die HF-Detektorschaltungen werden Zero-bias Schottky-Detektordioden verwendet. Eine Spannungsempfindlichkeit zwischen 20 und 25 mV/µW sowie eine tangentielle Signalempfindlichkeit um -60 dBm bei einer Videobandbreite kleiner 10 kHz sind typische Werte, welche sich mit der verwende-
ten Detektordiode erzielen lassen. Die Verwendung von apertur gekoppelten Mikrostrei-
fenantennen ermöglicht, das Zuführungssubstrat (engl. Feed Layer Substrate) vom Sub-
strat der eigentlichen abstrahlenden Antenne (engl. Patch Layer Substrate) zu trennen.
Dadurch ergibt sich ein guter Wert für die Unterdrückung der Abstrahlung gegen die
Antennenhinterseite. Insbesondere vereinfacht der mehrschichtige Antennenaufbau die
Realisierung von kleinen und kompakten Datenträgern. Es werden zirkularpolarisierte
apertur gekoppelte Mikrostreifenantennen vorgestellt, bei denen die Polarisationsrichtung
umgedreht werden kann, um Drehrichtungsmodulation auszuführen. Diese Antennen
weisen über die gesamte Bandbreite des ISM Bands eine Unterdrückung der jeweils uner-
wünschten Polarisation von über 15 dB auf. Für den aktiven Modulator wurde eine Schalt-
tungstopologie bestehend aus einem balancierten Mischer mit vorgeschaltetem HF-
Verstärker verwendet. Die realisierte Schaltung weist einen Konversionsgewinn von 6 dB
bei einem DC Leistungsverbrauch von 10 mW auf. Beim Aufbau des aktiven Modulators
werden sowohl kommerzielle Bauteile als auch integrierte Mikrowellenschaltungen (engl.
MMIC) eingesetzt. Der HF-Synthesizer der Ablese-Einrichtung erlaubt, Frequenzhüpf-
Modulation auszuführen. Es wurde eine Schaltungstopologie untersucht, welche aus
einem Phasenregelkreis (engl. Phase-Locked-Loop) besteht, dessen Referenzsignal von
einem digitalen Frequenzsynthesizer (engl. Direct Digital Synthesizer) in der Frequenz
moduliert wird.

Der letzte Teil der Arbeit befasst sich mit Messungen des gesamten Prototyp-Objektiden-
tifikationssystems. Bei einer abgestrahlten Leistung von 10 mW EIRP lässt sich eine
Übertragungsdistanz von 4,2 m in der Abwärtsstrecke erzielen. Die Übertragungsdistanz
in der Aufwärtsstrecke ist etwas kleiner. Weiter wurde mit dem Prototyp-Objektidentifi-
fikationssystem ein Übertragungssystem bestehend aus mehreren Ablese-Einrichtungen
und Datenträgern aufgebaut und ausgetestet. Das realisierte System ermöglicht jedoch
nicht, mehrere Datenträger gleichzeitig auszulesen, da noch kein Übertragungsverfahren
verwendet wird, welches ein Mehrfachzugriffsverfahren (engl. Anticollision Procedure)
unterstützt.

Zusammenfassung
1. Introduction

In recent years wireless identification systems have become very popular in the fields of factory automation, purchasing and distribution logistics, personnel and security management and material flow systems where objects need to be identified by means of attached transponders or tags.

There are mainly two types of identification systems: (1) inductively coupled radio frequency identification (RFID) systems operated below 30 MHz and (2) microwave tagging systems using backscatter modulation for transmitting identification or data over a small distance to a reader station. Fig. 1.1 shows a typical application example of a microwave tagging system used for an automatic identification procedure in the car manufacturing industry.

![Fig. 1.1: Application example of an automatic identification procedure: car manufacturing using a microwave tagging system. (Reproduced by permission of TagMaster AB, Sweden)](image)

The inductively coupled RFID systems typically provide slower data transfer and must work at closer distances (frequently less than 0.5 meters) than microwave tagging systems. Many applications requiring a maintenance-free operation of the transponder (e.g. ski tickets) use inductively coupled RFID systems because the transponders are remotely powered by the carrier signal of the reader.

Microwave tagging systems are well suited for the identification of large objects like containers, pallets, trucks etc., especially if the object speed is high and the environment is dirty. In the 2.4 GHz industrial-scientific-medical (ISM) band, tagging systems using battery powered transponders show typically a transmission distance up to 4 meters at 25 mW effective isotropic radiated power (EIRP). Modern tags have such a low power consumption that a battery life of about 10 years can be expected when passive backscatter modulation is applied.
1.1 Motivation

Future microwave tagging systems will provide new functions and services. In particular new applications will require tags with larger transmission distance, higher data rates and lower acceptable error rates. To increase the transmission distance, and to be able to use more sophisticated modulation schemes, a decrease in battery life has however to be accepted.

The focus of the research for this thesis is therefore on the investigation of an active microwave tagging system where the transponder either applies active backscatter modulation or shows a complete transceiver. In order to keep battery life high and tag size and cost low, the microwave circuitry (active modulator on the tag) must be very efficient and able to operate from a low voltage.

1.2 Project Description

The research for this thesis was performed under the project ACTMIT (Active Microwave Tagging Systems for the ISM Bands 2.4 GHz, 5.8 GHz and 24 GHz) which was funded by the Swiss Priority Program in Micro & Nano System Technology (MINAST). The project was in cooperation with the academic partners HSR University of Applied Sciences, Rapperswil and Communication Technology Laboratory (IKT), at ETH, Zurich, and with the industrial partners Baumer Ident GmbH, Weinheim (Germany), and Baumer Electric AG, Frauenfeld. The objectives of the project can be summarized as follows:

- Lowest power read/write tags primarily for the 2.4 GHz ISM band.
- Increased operating distance due to the active modulator on the tag of up to 10 meters with an RF power of maximum 10 mW EIRP.
- Evaluation of coding schemes suitable for multi-tag multi-reader operation.

1.3 Organization

After a short overview on the radio frequency ranges and the licensing regulations for tagging systems, Chapter 2 is focused on the classification of the different tagging system types. The presented identification systems include: (1) inductively coupled RFID systems, (2) tagging systems using surface acoustic wave (SAW) transponders and (3) microwave tagging systems. Moreover, microwave tagging systems are subdivided into remotely powered systems and battery powered systems using passive or active backscatter modulation.

In chapter 3 system issues of an active backscatter tagging system are discussed. After a short introduction into the basic operation principle of a backscatter system the optimum modulation schemes and system topologies for the down- and uplink communication are analyzed. In the discussion of the downlink communication, circular-polarization-shift-keying modulation is proposed as optimum modulation scheme meeting the requirements of a low power and low complexity transponder. The investigations of the uplink commu-
nication are focused on the discussion of the optimum system topology for an active backscatter system. Finally, the complete system topology is presented as it is implemented in a demonstrator set-up of the tagging system using active backscatter modulation in the 2.4 GHz ISM band.

Chapter 4 is dedicated to the discussion of the tagging system setup. The key components and circuits of the system setup include

- RF detector circuit
- Aperture-coupled microstrip patch antennas
- Active modulator
- RF synthesizer

For each component design rules, simulation results and measured data are shown. Moreover, the signal processing applied in the downlink and uplink communication is also presented in this chapter.

Measurement results obtained for the complete tagging system are shown in Chapter 5. This chapter is also illustrated by photographs of the experimental active tagging system including readers, transponders and the application software running on the host PC.

In the Appendix, some commonly used anticollision procedures are presented. A multiple access procedure is also discussed which is suitable for the experimental active tagging system.
2. Classification of Tagging Systems

In this chapter an overview on wireless identification systems is presented. Tagging systems may be classified by the frequency range and the operation principle of the transponder. For frequencies below 30 MHz, the interaction between transponder and reader is based on inductive coupling. In the frequency range above a few 100 MHz, microwave identification systems are applied which use electromagnetic backscatter coupling. Several types of transponders can be distinguished: read-only or read/write tags, remotely powered or battery powered tags, LC-tags, backscatter modulated tags or surface acoustic wave tags.

In order to classify the various identification systems according to their operation frequencies, this chapter summarizes in the first part the frequency range allocations and the radio licensing regulations for RFID and tagging systems. In the second part of the chapter, three types of tagging systems are presented that represent the most common applied systems for wireless identification of objects. These examples include an inductive RFID system, an identification system using surface acoustic wave transponders and a microwave tagging system applying backscatter modulation.

It should be noted that the following technical terms are used as synonyms in the next chapters:

- reader, interrogator
- tag, transponder, data carrier

2.1 Radio Frequency Ranges and Licensing Regulations

Most applications require that wireless identification systems are operated in license free frequency bands. These frequencies are classified worldwide as ISM frequency ranges (Industrial-Scientific-Medical). International licensing regulations\(^1\) give a limit for the maximum permissible electrical field strength, the maximum permissible transmitter output power or the permissible spurious emissions. For instance, in the ETSI regulation EN 300220 the effective isotropic radiated power (EIRP) for a reader operated in the 2.4 GHz ISM band is limited to 500 mW. The most important frequency ranges for RFID systems are listed in the following section. More informations about regulation issues are specified in reference [1].

- frequency range 9 - 135 kHz

This frequency range does not belong to the ISM frequency bands. It is heavily used by long wavelength radio services such as aeronautical and marine navigational services (e.g. OMEGA), time signal services or military radio services. Frequencies of this band show a high penetration depth in non-metallic materials. Hence, injectable transponders can be built e.g. for animal identification. A low power consumption (low clock frequency of the

\(^1\) ETSI standards: EN 300330, EN 300220, EN 3000440 (Europe) or FCC part 15 (USA)
control unit) and a miniaturized transponder format (transponder antenna consisting of a ferrite coil) are some further advantages of identification systems operated in this frequency range (e.g. TIRIS® system from Texas Instruments).

- **frequency range 13.56 MHz**

A large number of inductive RFID systems are operated in this worldwide ISM frequency band (e.g. MIFARE® from Philips/Mikron). Many RFID chips show an on-chip parallel capacitor for the resonance matching of the transponder coil. Clock frequency and data transmission (typically 106 kbit/s) can be chosen sufficiently high for performing cryptological functions in the microprocessor of the transponder. In addition to RFID systems, other ISM applications include remote control systems, demonstration radio equipment and pagers. This frequency band is located in the middle of the short wavelength band and permits, due to its propagation characteristics, transcontinental connections.

- **frequency range 27.125 MHz**

This frequency range is used only for special applications as it is not an ISM frequency band. An example of an application is the EURO-Bahse system of the European train control system (ETCS) where the power transmission frequency is located in this frequency range. Similar to the 13.56 MHz ISM band, this frequency range shows a large bandwidth and thus allows to transmit data at high data rates (typically 424 kbit/s). Other applications operated in this ISM frequency band include CB radio (26.565-27.405 MHz) and high frequency welding equipment which may cause interference problems with inductive RFID systems operated in industrial environments.

- **frequency range 433.920 MHz**

Mainly backscatter tagging systems are operated in this ISM frequency range which is located approximately in the middle of the amateur radio frequency range (430.0-440.0 MHz). It belongs to the UHF frequency range where the wave propagation is approximately optical. Obstacles or buildings lead to a strong damping or reflecting of incident electromagnetic waves. This frequency range is used by various radio equipments like handheld transmitters for vehicle central lockings, cordless headphones or telemetry transmitters (e.g. wireless thermometer).

- **frequency range 915.0 MHz**

In Europe, this frequency range is not recognized for license free applications. However, in Australia and the USA the frequency ranges 888-889 MHz and 902-928 MHz may be used for RFID applications. Especially, tolling systems are operated in this frequency band.

- **frequency range 2.45 GHz**

Microwave backscatter tagging systems are operated in this S-band ISM frequency range. It overlaps partially with frequency bands used by the amateur radio and radiolocation services. The propagation characteristics of the electromagnetic waves are quasi-optical. For the data transmission between reader and transponder a line-of-sight (LOS) path is necessary as buildings and other obstacles damp electromagnetic waves very strongly at trans-
mission. Due to the strong free space loss (40 dB at 1 m distance from the transmitter), the wireless transfer of sufficient power is difficult so most transponders have a battery. A wide range of other applications is operated in this frequency band. Major interference problems for tagging systems may be caused by the increasing number of PC local area networks for the wireless networking of PCs (e.g. Bluetooth™).

- **frequency range 5.8 GHz**

As well as at the 2.45 GHz ISM band, this ISM frequency range is partly shared with amateur radio and radiolocation services. This band is allocated for future use. It is not yet achievable cost effectively with current techniques for conventional transponders. Targeted applications concern particularly electronic toll collection (ETC) systems. A potential application is also the intelligent transport system (ITS), in which information is exchanged between roadside units and on-board transponders [3].

- **frequency range 24.125 GHz**

Apart of the ISM applications, this frequency range is occupied by amateur radio and radiolocation services plus satellite links for digital broadcasting. ISM applications include police radar for speed measurements, movement sensors, automatic door openers. Due to the strong free space loss, transponders with an on-board RF source need to be used for obtaining large transmission distance (> 4 m) [5]. Up to now there is no commercially available identification system operating in this frequency range.

Table 1.1 summarizes the ISM frequency ranges that are relevant for tagging or remote sensing applications.

<table>
<thead>
<tr>
<th>ISM frequency ranges</th>
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<tr>
<td>6.765 - 6.795 MHz</td>
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<tr>
<td>13.553 - 13.567 MHz</td>
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<tr>
<td>26.957 - 27.283 MHz</td>
</tr>
<tr>
<td>40.660 - 40.700 MHz</td>
</tr>
<tr>
<td>433.050 - 434.790 MHz</td>
</tr>
<tr>
<td>2.400 - 2.4835 GHz</td>
</tr>
<tr>
<td>5.725 - 5.875 GHz</td>
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<tr>
<td>24.00 - 24.25 GHz</td>
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### 2.2 Inductively Coupled RFID Systems

Identification systems using inductive coupling between transponder and reader are operated below 135 kHz or in the 6.78 MHz, 13.56 MHz and 27.125 MHz frequency ranges. The coupling of two electric circuits via the magnetic field is the physical principle upon which inductively coupled RFID systems are based. Inductive coupling is therefore only possible in the near-field of the transmitter antenna where the electromagnetic field has
not been separated from the antenna yet. For loop antennas which are mostly used by RFID systems operated in these frequency ranges, the transition between the near-field region (Fresnel region) and the far-field region (Fraunhofer region) is approximately given by [6]

\[ r_1 \approx \frac{\lambda_0}{2\pi} \]  

E.g. for an RFID system operated in the 13.56 MHz ISM frequency range (wavelength \( \lambda_0 \approx 22.1 \text{ m} \)), the far field region where no inductive coupling is possible begins at a distance \( r_1 \) of 3.5 m from the reader antenna.

Inductively coupled transponders are operated passively as the operation energy is provided by the reader. The reader fulfills two tasks: (1) transmitting data and commands to the transponder or receiving data from the transponder and (2) generating RF transmission power to activate the transponder and supply it with power. The data transmission from the reader to the transponder is termed downlink communication. The term uplink communication is used for the data transfer from the transponder to the reader. Energy transfer for the power supply of the transponder is performed either continuously during the uplink and downlink communication (duplex mode) or just during the downlink communication (sequential mode). In order to ensure a continuous power supply to the transponder even during data transfer, special coding can be applied (e.g. modified Miller coding or pulse-pause coding (PPC) [1]). For instance in PPC a logical 1 is represented by a pause duration \( t \) before the next pulse while a logical 0 is represented by a pause duration \( 2t \) before the
next pulse. Due to the very short pulse durations ($t_{\text{pulse}} << T_{\text{bit}}$) a continuous power supply is possible.

For the illustration of the operation principle, the block diagram of a reader and a transponder is shown in Fig. 2.1 and Fig. 2.2. The block diagram of the reader is split into two functional blocks: the control unit and the radio frequency (RF) front end, comprising a transmitter and a receiver. The transmitter part of the RF front end consists of an oscillator, a modulator and a power amplifier which is connected to the loop antenna. Binary data from the control unit (Tx data) is fed to the mixer to be upconverted. Depending on the modulator type, amplitude-shift-keying (ASK) or phase-shift-keying (PSK) modulation is performed on the oscillator signal. For frequency-shift-keying (FSK) modulation, the baseband signal has to be fed directly into the frequency synthesizer. Before emitting the modulated signal by the loop antenna, the power amplifier amplifies it to the required power level.

The receiver part of the reader consists of a steep-edged bandpass filter, a low noise amplifier and a demodulator whose binary output signal (Rx data) is fed to the control unit. The steep edge bandpass filter in the receiver has the task of filtering out the response signal from the transponder (transponder operated in sequential mode) and additionally blocking the strong carrier signal from the transmitter part of the reader (transponder operated in duplex mode). After amplifying the received response signal by a low-noise amplifier (LNA), it is further demodulated and led to the control unit. The reader’s control unit performs the functions of signal coding and decoding and of handling the communication with the application software where commands or data can be entered.
The transponder signal is either modulated as double-sideband (DSB) signal on the fundamental frequency (load modulation of the main carrier) or on a sub harmonic frequency (subharmonic modulation) of the reader’s oscillator frequency. In addition, RFID systems based on load modulation may use a subcarrier for the uplink communication. The subcarrier frequency is obtained by binary division of the operating frequency. A typical subcarrier frequency in a 13.56 MHz RFID system is 212 kHz (13.56 MHz / 64 ≈ 212 kHz). Systems using load modulation with subcarrier are suited for full duplex operation where the downlink and uplink communication is performed at the same time.

The RF front end depicted in Fig. 2.2 is typical for a read/write inductively coupled transponder. Generally, transponders can be subdivided into read-only and writable (read/write) transponders. Read-only transponders show a unique identification number that is incorporated into the transponder during the chip manufacturing. The communication with the reader is unidirectional as the transponder transmits continuously its identification (ID) code when entering the interrogation range of the reader. This type of transponder represents the low end, low cost segment of the range of RFID data carriers. On the other hand, writable transponders show a ROM where permanent data such as the ID number is stored and an EEPROM or FRAM for the storage of temporary data received from the reader. A microcontroller handles the EEPROM control and read/write control.

The depicted inductively coupled transponder belongs to the type of passive data carriers which do not have their own power supply (battery). The operation power of the control unit devices (microcontroller and memory chips) is derived from the RF carrier of the reader that is rectified and supplied to the control unit as a regulated supply voltage. The antenna of the transponder consists of a transponder coil and a capacitor. It is connected in parallel with the transponder coil to form a parallel resonant circuit with a resonance frequency that corresponds to the carrier frequency of the reader. The resonant frequency of the parallel resonance circuit (see Fig. 2.1 and Fig. 2.2) can be calculated by

\[
fr = \frac{1}{2\pi \sqrt{L_2 \cdot C_2}} = \frac{1}{2\pi \sqrt{L_1 \cdot C_1}}.
\]

The magnetic alternating field of the reader induces at resonance frequency a voltage in the transponder coil that is used to provide the power supply. A bridge rectifier with smoothing capacitor may be used to generate the required supply voltage. Frequently, a parallel voltage regulator at the rectifier output is used to prevent the supply voltage from being subject to an uncontrolled increase when the transponder approaches the reader antenna. As shown in Fig. 2.2 the clock signal for the digital circuitry on the transponder is also derived from the reader signal by a clock recovery circuit. For the demodulation of the received data from the reader, a video detector is commonly used as frequently ASK modulation is applied for the downlink communication.

In the uplink communication, the binary data from the transponder is transmitted using ohmic load modulation or capacitive load modulation. In capacitive load modulation an additional capacitor (parallel to the capacitor at the loop coil) is switched on and off in time with the data stream or in time with a modulated sub carrier. This causes an amplitude and phase modulation. In the case of ohmic load modulation (as shown in Fig. 2.2), a
parallel resistor is switched on and off according to the binary data of the transponder. Ohmic load modulation generates amplitude modulation.

### 2.2.1 Operation Principle of Inductively Coupled RFID Systems

In the following, the physical operation principle of inductively coupled RFID systems is discussed in greater detail by investigating the power transmission in the downlink communication and the data transmission in the uplink communication [1]. Fig. 2.3 shows the equivalent circuit diagram of the magnetically coupled transponder in the vicinity of the reader. For the power transmission, the reader generates a time variant current \( i_1 \) in the antenna coil represented by the inductivity \( L_1 \). Due to the inductively coupling a change in the magnetic flux \( \Phi_M \) (or the interlinked flux \( \Phi_{IM} \), respectively) of the transponder coil (inductivity \( L_2 \)) is generated which induces a voltage \( U_{ind} \) that is given by

\[
U_{ind} = \oint E_i \cdot ds = \frac{d\Phi_M(t)}{dt} = -N \cdot \frac{d\Phi_M(i_1)}{dt}. \tag{2.3}
\]

The contour integral of the electric field strength \( E_i \) is equal to the voltage appearing at the open-circuited terminals of the transponder coil. The reader coil and the transponder coil are coupled by their mutual inductance \( M \) which is defined as

\[
M = \frac{N_1 \cdot \Phi_M}{i_1}. \tag{2.4}
\]

---

**Fig. 2.3:** Equivalent circuit diagram of the transponder (on the right) and the reader (on the left) at power transmission.
The mutual inductance is proportional to the number of windings $N_j$ of the transponder coil and to the mutual magnetic flux $\Phi_M$ divided by the current $i_1$ flowing through the reader coil. Only a small fraction of the magnetic flux of the reader’s conductor loop is coupled to the conductor loop of the transponder. This mutual flux is determined by

$$\Phi_M = \int_{A_2} B \cdot dA_2 = \int_{A_2} \mu_0 H \cdot dA_2 , \tag{2.5}$$

where $A_2$ denotes the cross-section of the transponder’s conductor loop. $B$ and $H$ ($\mu_0 = 4\pi \cdot 10^{-7}$ Vs/Am) are the magnetic inductance and the magnetic field strength, respectively, occurring in the cross-section $A_2$.

Frequently, rectangular transponder coils are used at the reader. For a rectangular conductor loop with edge length $a-b$ and $N_j$ windings, the magnetic field strength in a distance $x$ from the center of the antenna is given by [1]

$$H = \frac{N_j \cdot i_1 \cdot ab}{4\pi \sqrt{\left(\frac{a}{2}\right)^2 + \left(\frac{b}{2}\right)^2 + x^2}} \left(\frac{1}{\left(\frac{a}{2}\right)^2 + x^2} + \frac{1}{\left(\frac{b}{2}\right)^2 + x^2}\right) . \tag{2.6}$$

Fig. 2.4 shows an illustration of the path of the magnetic flux lines around a rectangular conductor loop as it may be used for a reader antenna.

![Fig. 2.4: Lines of the magnetic flux around a rectangular coil representing a typical reader antenna.](image)
The inductive coupling between reader and transponder can also be described by the coupling coefficient $k$. The following relationship applies

$$ k = \frac{M}{\sqrt{L_1 \cdot L_2}} \quad \text{(2.7)} $$

where $L_1$ and $L_2$ denote the inductances of the reader and transponder coil. For a rectangular conductor coil with edge length $a \cdot b$ and a wire diameter $D$ being very small compared to the diameter $q = \sqrt{a^2 + b^2}$ of the loop, the inductance is given by [6]

$$ L = 4a \cdot \left( \ln\left(\frac{5ab}{D \cdot (a + q)}\right) + \frac{b}{a} \cdot \ln\left(\frac{5ab}{D \cdot (b + q)}\right) + 2 \cdot \left(1 - \frac{q - b}{a}\right) \right) \quad \text{(2.8)} $$

This formula yields the inductance in $nH$ when the dimensions of the edge lengths are given in $cm$. An analytical calculation of the coupling coefficient is only possible for very simple antenna configurations. In practice, inductively coupled transponder systems are operated with coupling coefficients lower than 15% [1].

In the following, the voltage $U_2$ (see Fig. 2.3) appearing at the terminals of the transponder coil is calculated. This voltage is used by the rectifier for providing the transponder’s power supply. Based on this expression, the minimum interrogation field strength $H_{\min}$ (and thus the maximum transmission distance) is derived. Finally, considerations about the load modulation scheme are shown which are also related to the formula for $U_2$.

Assuming that the reader generates a sinusoidal alternating magnetic field in the transponder coil, the voltage $U_2$ is determined by

$$ U_2 = U_{\text{ind}} - j\omega L_2 \cdot i_2 - R_2 \cdot i_2 \quad \text{(2.9)} $$

where $R_2$ is the resistance of the conductor loop with inductance $L_2$. The voltage $U_{\text{ind}}$ that is generated by the mutual inductance $M$ and the current $i_2$ can be expressed as

$$ U_{\text{ind}} = j\omega M \cdot i_1 \quad \text{(2.10)} $$

$$ i_2 = \frac{U_2}{R_L + \frac{1}{j\omega C_p}} \quad \text{(2.11)} $$

Substituting these relations into Eq. (2.9) yields

$$ U_2 = \frac{j\omega M \cdot i_1}{1 + \frac{R_2}{R_L} - \omega^2 L_2 C_p + j\omega \left(\frac{R_2 C_p + \frac{L_2}{R_L}}{R_L}\right)} \quad \text{(2.12)} $$

Moreover, Eq. (2.7) can be solved for the mutual inductance and substituted into Eq. (2.12) to obtain...
The curve of $U_2$ versus frequency shows a steep step-up at resonance frequency which can be measured by the Q-factor. For maximizing the energy transfer, a high Q-factor is desirable. However, for the data transmission during the load modulation, a minimum bandwidth is necessary. If the Q-factor is too large, the modulation sidebands will be damped so much that the transmission range becomes very small. Therefore, a trade-off between maximum power transmission and maximum data rate has to be made when developing inductively coupled RFID systems. The Q-factor is defined by

$$ Q = \frac{f_r}{\Delta f_{-3dB}}. \quad (2.14) $$

The bandwidth $f_{-3dB}$ can be calculated by using the relationship $Re\{Z_{2j}\} = Im\{Z_{2j}\}$ where $Z_{2j}$ is obtained from Eq. (2.13) after dividing by $i_1$. Substituting the resulting $f_{-3dB}$ into Eq. (2.14) yields

$$ Q \approx \frac{1}{L_2} \frac{C_p}{L_2 + \frac{1}{R_L \cdot C_p}}. \quad (2.15) $$

For a fixed resonance frequency, the Q-factor is only dependent on the load resistance $R_L$ and the coil resistance $R_2$. When optimizing the power transmission range, it should be noted that for every pair of parameters $(R_2, R_L)$, there is an inductance value $L_2$ at which the Q-factor is at a maximum. By setting the first derivative of Eq. (2.15) with respect to $L_2$ equal to zero and evaluating this expression at the capacitance value $C_p$ (calculated at resonance frequency $f_r = 1/(2\pi L_1 C_p)^{1/2}$), the optimum inductance is determined by

$$ L_{2, opt} = \frac{R_2 \cdot R_L}{(2\pi f_r)^2 \cdot L_1}. \quad (2.16) $$

The magnetic field strength $H$ at which the supply voltage of the transponder is just high enough for the operation of the transponder is denoted as minimum interrogation field strength $H_{min}$. It is derived from Eq. (2.12) where the induced voltage is replaced by

$$ U_{ind} = j\omega M \cdot i_1 = j\omega N\mu_0 H_{min} A. \quad (2.17) $$

Substituting Eq. (2.17) into Eq. (2.12) and solving for $H_{min}$ yields

$$ H_{min} = \frac{U_2 \cdot \left(1 - \omega^2 L_2 C_p + \frac{R_2^2}{R_L^2} \right)^2 + \omega^2 \left(\frac{L_2}{R_L} + R_2 C_p \right)^2}{\omega \mu_0 AN}. \quad (2.18) $$
The minimum interrogation field strength depends on the number of windings and the coil cross-section. For an example transponder ($L_2 = 3.5 \, \mu H$, $C_p = 39.4 \, pF$, $N = 4$, $A = 5cm \times 8cm$, $R_2 = 5\Omega$, $R_L = 1.5 \, k\Omega$) operated in the 13.56 MHz ISM band at 5V (assumption: $U_2 \approx$ supply voltage) $H_{min}$ is 0.629 A/m. This corresponds to a transmission distance of 0.12 meters when $H_{min}$ is substituted in Eq. (2.6) and the dimensions of the reader antenna ($i_j = 0.5 \, A$) are assumed to be equal to that of the transponder.

Fig. 2.5 illustrates the influence of the Q-factor on the modulation sidebands. A transponder optimized for efficient energy transfer shows a high Q-factor. In order to prevent that the modulation sidebands are affected too much by the edge of the voltage step-up, the data rate of the load modulation has to be small. The degradation of the modulation products becomes even worse if load modulation with subcarrier is used. Load modulation with subcarrier is therefore mainly used for close coupling RFID systems where the Q-factor may be lower due to the small transmission distance. As component tolerances may strongly detune data carriers with high Q-factor, some transponder chips show on-chip capacitors that are trimmed during manufacturing.

In order to conclude this discussion of the physical operation principles of inductively coupled RFID systems, the transmission of the load modulated data to the reader is discussed in the following. Based on the equivalent circuit diagram shown in Fig. 2.3, the voltage $U_0$ at resonance frequency (the reactances of $L_1$ and $C_1$ cancel each other) can be expressed as

$$U_0 = R_1 \cdot i_1 - j\omega M \cdot i_2.$$  \hspace{1cm} (2.19)

For the transponder current $i_2$ the relation applies

$$i_2 = \frac{U_{ind}}{R_2 + j\omega L_2 + Z(R_L || C_p)} = \frac{j\omega M \cdot i_1}{R_2 + j\omega L_2 + \frac{R_L}{1 + j\omega R_L C_p}}.$$  \hspace{1cm} (2.20)

Substituting Eq. (2.20) into Eq. (2.19) yields the total impedance

$$Z_0 = \frac{U_0}{i_1} = R_1 + \frac{\omega^2 k^2 L_1 L_2}{R_2 + j\omega L_2 + \frac{R_L}{1 + j\omega R_L C_p}} = R + Z_T,$$  \hspace{1cm} (2.21)

where Eq. (2.7) has been solved for the mutual inductance $M$. The second term $Z_T$ in Eq. (2.21) is equal to the transformed transponder impedance. It is important for the demodulation of the load modulated data from the transponder. As mentioned above there is ohmic or capacitive load modulation. For instance, ohmic load modulation is performed by switching a resistance $R_{mod}$ parallel to $R_L$. If $R_{mod}$ is chosen equal to $R_L$ - which corresponds to 50% ASK modulation - and the component values are chosen as mentioned above ($R_L = 1 \, k\Omega$, $L_1 = 2 \, \mu H$, $L_2 = 3.5 \, \mu H$, $R_2 = 5 \, \Omega$, $C_p = 39.4 \, pF$, $k = 3\%$), the transformed transponder impedance takes the two states $Z_T(R_L/R_{mod}) = 0.5 \angle -15^\circ \, \Omega$ and
load modulation with subcarrier (f = 847 kHz)

Fig. 2.6: (1) Induced voltage $U_2$ at the transponder coil as function of the load resistance ($R_L = 1 \, k\Omega$, $Q = 3$, $R_L = 1.5 \, k\Omega$, $Q = 4.5$, $R_L = 2 \, k\Omega$, $Q = 6$); (2) Influence of the Q-factor on the damping of the modulation sidebands (transponder and reader parameters: $L_2 = 3.5 \, \mu H$, $C_p = 39.4 \, pF$, $R_2 = 5 \, \Omega$, $k = 3\%$, $L_1 = 2 \, \mu H$, $I_1 = 0.5 \, A$).

$Z_\tau(R_L) = 0.29 \angle -30^\circ \, \Omega$. As shown in Fig. 2.1 the resulting voltage $U_0 = (R_f + Z_\tau) \cdot i_f$ is demodulated after bandpass filtering by a simple video detector.

The section above showed a short overview on the physical operating principles and the main components of a typical inductively coupled RFID system. Most of the commercial applications and manufacturers of RFID systems are in the field of inductively coupled systems. In the following two types of tagging systems are presented which are suited for more specific applications where for instance a longer transmission range, higher data rates or heat resistant tags are demanded.

### 2.3 Surface Acoustic Wave Transponder Systems

Surface acoustic wave (SAW) transponder systems apply *time division access* for the data transmission in the uplink. SAW transponders utilize the speed of sound of a surface acoustic wave propagating on a piezoelectric substrate. The acoustic storage time has to be longer than the decay time of the environmental electromagnetic wave echoes. SAW transponders respond therefore *delayed* to the downlink.

Fig. 2.6 shows the layout of a surface acoustic wave transponder. On the surface of a plain polished piezoelectric substrate, a finger-shaped electrode structure is arranged. This
metallic structure is called *interdigital transducer*. Due to the piezoelectricity, an electrical signal at the interdigital transducer excites an acoustic wave that propagates on the surface. Conversely, surface acoustic waves that are generated in the SAW device can be converted into an electric field. For transponder applications, lithium niobate (LiNbO₃) or lithium tantalate (LiTaO₃) are used as substrate materials.

The conversion between electrical signals and acoustic waves is performed by an interdigital converter. When a high frequency scanning pulse (e.g. in the 2.4 GHz ISM band) transmitted by the reader is supplied from the dipole transponder antenna to the interdigital converter, the electromagnetic waves are converted into much slower acoustic surface waves which propagate through the substrate in longitudinal direction. The propagation velocity of the SAW on commonly used substrates lies between 3000 and 4000 m/s. A small part of the surface acoustic waves is reflected by the edges of the reflective strips. These reflectors are made of aluminium. The reflections at the rising and descending edges occur with opposite signs. If the width of the electrode is \( \lambda/4 \), then the reflected waves are superpositioned in-phase [6]. The reflected parts of the surface acoustic waves propagate back to the interdigital converter where they are converted into an RF pulse sequence which is emitted by the dipole antenna. This pulse sequence is received and demodulated by the reader. According to the number of reflectors on the transponder, the downconverted RF signal contains pulses whose delay times between the individual pulses correspond to the spatial distances between the reflector strips on the substrate. The *arrangement of the reflectors* can be chosen in a way that it represents a *binary sequence* which is equal to the identification code of the transponder. The layout of the reflectors and thus the identification code is determined at the fabrication of the SAW device. Therefore, SAW transponders belong to the category of read-only tags.

Fig. 2.6: Layout of a surface acoustic wave transponder.
Fig. 2.7: Block diagram of a reader using the FM CW radar principle for the demodulation of the transponder's response signals. The interference frequencies $f_1$ and $f_2$ are caused by the SAW reflectors 1 and 2, respectively.

Fig. 2.8: Radiolocation of a SAW transponder with time domain division between the transmitted impulse and response signal. The reader receives the response from the transponder after the decay time of the environmental echoes $T_f$.

Classification of Tagging Systems
Fig. 2.7 shows the radio request of an SAW transponder. The reader transmits a strong RF pulse. Due to multipath propagation and the resulting reflections on the metal surfaces of the environment, the reader receives at first the environmental echoes. Only after the decay of the environmental echoes the pulsed response signal of the surface acoustic wave transponder arrives at the reader antenna. This delayed response in the uplink is called time division access. The signal power of the response is dependent on the code length. As the identification code sequence corresponds to the number of reflectors, a long code word results in a strong attenuation of the reflected surface acoustic waves on the transponder. The transmission range of a SAW system depends mainly on the transmission power of the reader. For a system operated in the 2.45 ISM band a transmission distance of 6 m can be expected at a transmission power of 500 mW.

A commonly used method for interrogating transponders with time division access is the principle of FM-CW radar. The FM-CW radar scheme uses the time delay to frequency shift conversion and is utilized in the reader to measure the delay times of the pulses in the transponder’s response signal. A block diagram of an FM-CW reader is depicted in Fig. 2.8. As an example, it is illustrated how the delay time caused by two reflectors is converted into a frequency shift ($f_1 - f_2$). An alternative topology to the FM-CW reader is the pulse radar where the impulse response of the transponder is evaluated in the reader.

### 2.4 Microwave Tagging Systems

There are three types of microwave transponders: (1) remotely powered tags, (2) backscatter modulated tags and (3) tags which have their own RF source. Backscatter modulated tags can further be split into transponders using passive or active backscatter modulation. The operating principles of these systems will be discussed in the following three chapters.

#### 2.4.1 Remotely Powered Transponders

Remotely powered transponders are designed to be operated in the near-field region of the reader antenna. This is the case because remotely powered transponders need to take their power supply from the transmitted RF carrier of the reader. In the near-field region the power transmission from the reader to the tag is maximum. This can be shown under the following assumptions:

- The power density $S_N$ in the near-field is constant.
- The cross-section of the near-field is equal to the effective receiving area $A_{e, rd}$ of the reader antenna.

The constant power density is given by

$$S_N = \frac{P_{tx} \cdot \eta_{rd}}{A_{e, rd}}, \quad (2.22)$$
where \( P_{tx} \): power fed to the input of the reader antenna

\( \eta_{rd} \): efficiency of the reader antenna

The effective receiving area \( A_{e, tg} \) of the tag antenna is determined by

\[
A_{e, tg} = \frac{P_{rx}}{\eta_{tg} \cdot S_N}.
\]  

(2.23)

where \( P_{rx} \): power appearing at the terminal of the tag antenna

\( \eta_{tg} \): efficiency of the terminal of the tag antenna

Solving Eq. (2.23) for the power density \( S_N \) and substituting into Eq. (2.22) yields the power transfer function \( T_N \)

\[
T_N = \frac{P_{rx}}{P_{tx}} = \frac{A_{e, tg}}{A_{e, rd}} \cdot \eta_{tg} \cdot \eta_{rd}.
\]  

(2.24)

Eq. (2.24) states that the power transfer from the reader to the tag is independent from the distance between reader and tag as long as the tag is in the near-field region of the reader antenna.

The power transfer function \( T_F \) for the far-field region is easily deduced from the transmission formula [37]

\[
P_{rx} = P_{tx} \cdot G_{tg} \cdot G_{rd} \cdot \left( \frac{\lambda}{4\pi \cdot d} \right)^2,
\]  

(2.25)

where \( G_{tg} \): gain of the tag antenna (receiving antenna)

\( G_{rd} \): gain of the reader antenna (transmitting antenna)

\( d \): transmission distance

\( \lambda \): free-space wavelength

Substituting the following relationship for the antenna gain of the reader (and for the antenna gain of the tag as well) into Eq. (2.25)

\[
G_{rd} = D_{rd} \cdot \eta_{rd} = A_{e, rd} \cdot \frac{4\pi}{\lambda^2} \cdot \eta_{rd}
\]  

(2.26)

yields the formula for the power transfer function \( T_F \) in the far-field region
Fig. 2.9: Block diagram of a remotely powered microwave transponder.

\[
T_F = \frac{P_{tx}}{P_{rx}} = G_{tg} \cdot G_{rd} \cdot \left(\frac{\lambda}{4\pi \cdot d}\right)^2 = A_{e,rd} \cdot A_{e,tg} \cdot \eta_{rd} \cdot \eta_{tg} \cdot \left(\frac{1}{\lambda \cdot d}\right)^2.
\]

\(T_F\) is inverse proportional to the power of two of the transmission distance. The directivity 
\(D_{rd}\) used in Eq. (2.26) is defined as power density in the main beam to the average power 
density, \(D_{rd} = S_{rd}/S_i\) where \(S_i\) denotes the power density if the antenna radiates isotropically. The boundary between near- and far-field can be determined by equating Eq. (2.24) and Eq. (2.27) and solving for the distance \(d_{N,max}\)

\[
d_{N,max} = \frac{A_e,rd}{\lambda} \approx \frac{\Omega^2}{\lambda},
\]

where \(\Omega\) denotes the diameter of the reader antenna. The distance \(d_{N,max}\) is the maximum transmission distance for remotely powered systems. As \(d_{N,max}\) is approximately equal to the diameter of the reader antenna, remotely powered microwave tagging systems show a very limited operating range of just a few decimeters for practical applications. Two examples of such short-range systems are the identification of rental garments and mats in a laundry and the animal identification where miniaturized fully passive transponders are implanted under the skin of the animals.

The block diagram of a remotely powered microwave transponder is shown in Fig. 2.9. Its RF front end consists of an antenna, an RF filter and an RF rectifier. In a remotely powered transponder, the RF rectifier is a key component because it has to fulfill three tasks. First it rectifies the RF carrier in order to get a DC power supply for the operation of the control and signal processing unit on the transponder. Frequently, the RF carrier from the
reader is AM modulated with the data to be transmitted to the tag. A second task of the RF
rectifier is therefore to demodulate the superimposed amplitude modulation of the RF car-
rier. In this mode the RF detector is operated as envelope detector. For transmitting data
from the tag to the reader, a backscatter modulator can be used which consists of a pin
diode or a capacitance diode. In order to loose as little RF power as possible for the RF
rectifier, the backscatter modulator has to be loosely coupled to the antenna. Another
approach is to combine the backscatter modulator with the RF rectifier. In this case the RF
rectifier fulfills its third task by backscatter modulating the RF carrier with the data to be
transmitted to the reader. The backscatter modulation of the RF rectifier is performed by
applying the low frequency data signal from the signal processing unit to the DC port of
the rectifier. This data signal acts as pumping signal of the nonlinearity of the rectifier
diode. Thus, the transponder's response signal is up-converted to the RF carrier frequency
and emitted by the antenna to the reader. As the mixing process generates higher harmon¬
ics, a bandpass filter is required between the antenna and the RF rectifier. The filter may
also be used to increase the efficiency (conversion efficiency $P_{dc}/P_{RF}$) of the RF rectifier
when the higher harmonics are rejected to the RF rectifier with the appropriate phase.

### 2.4.2 Battery Powered Transponders using Passive Backscatter Modulation

In contrast to remotely powered tags, transponders applying backscatter modulation can
be operated in the far-field of the reader antenna. The power supply is delivered by a bat-
tery on the tag. A sensitive wake-up detector on the tag switches on the battery power
when a dedicated wake-up signal from the reader is detected. The wake-up detector may
perform either DC or AC detection. If DC detection is applied, a valid signal is detected
when the DC voltage of the rectified RF carrier is beyond a threshold voltage. In most of
the cases the frequency selectivity of the transponder's RF front end is not sufficient to
prevent that false detections may occur due to other users in the license free ISM band. In
order to circumvent this problem, it is convenient to use AC detection where the RF car-
rier contains a additional subcarrier which is only used for the wake-up detection. This
subcarrier can be detected by a bandpass filter following the RF detector. Detection errors
may be reduced due to the frequency selectivity of the bandpass filter.

Since transponders using passive backscatter modulation may be operated in the far field
of the reader antenna, the transmission distance is much higher than it can be achieved
with a remotely powered system. At a remotely powered tagging system the maximum
transmission distance is achieved when the DC voltage supply from the rectified RF car-
rier is lower than a minimal threshold voltage. On the other hand, at passive backscatter
modulation systems the transmission distance is limited by the sensitivity of the reader.
This difference in the operation between remotely powered and battery powered tagging
systems is illustrated in the link budget diagram depicted in Fig. 2.10. The curve labeled
with $A,A'$ represents the power level diagram of a remotely powered tagging system. Con-
sidering curve $A$, it should be noted that the received power from the reader of 10 dBm
remains constant within the zone of the near-field which ranges up to the normalized dis-

dance
In the illustration a directivity ratio $D_{rd}/D_{tg}$ between reader and tag antenna of 11 dB has been assumed. This yields a normalized near-field range $d/(\lambda D_{rd})$ of approximately 1. In the far-field region the power level decays with the power of two of the transmission distance. Therefore the minimum required power level $P_{min, tg}$ for the operation of the remotely powered tag is reached shortly after the beginning of the far field zone. The transmission distance is thus limited by the power level $P_{min, tg}$. As the backscatter modulator on the remotely powered tag is fully passive, it shows conversion loss. The curve $A'$ corresponds to the power level diagram of the up link communication when the transponder transmits backscatter modulated data to the reader. In this example, the backscatter modulated signal from the tag arrives at the reader at a power level of -40 dBm. Assuming a sensitivity of the reader of -90 dBm, the demodulation of the transponder’s response signal is no problem in a remotely powered tagging system.

However, this is different for a battery powered tagging system which applies passive backscatter modulation. The power level diagram of such a tagging system is illustrated by the curves $B, B'$. Since the transponder is operated in the far field of the reader antenna, it is not mandatory to use a high-gain antenna at the reader. In the example, it is assumed that the ratio of the directivities $D_{rd}/D_{tg}$ is 0 dB. Hence, the region of the near-field is correspondingly smaller than it is for curve $A$. The passive backscatter modulator on the tag
reflects the modulated RF carrier from the reader again with conversion loss. Due to the larger free space loss of the channel, the backscatter modulated signal arrives at the reader with a power of -85 dBm which is very close to the minimum sensitivity of the reader. In a

---

**Fig. 2.11**: Typical block diagram of a microwave tagging system using passive backscatter modulation.
battery powered tagging system using passive backscatter modulation the transmission distance is limited thus by the sensitivity of the reader. The sensitivity of the transponder’s RF detector (zero bias Schottky diodes) is about -70 dBm for a video bandwidth of 10 kHz. However, the minimum required power level is reached prior at the reader than at the tag. Different relations apply for active backscatter modulated systems (curves C, C') which are discussed in the next chapter.

A typical block diagram of a battery powered tagging system is depicted in Fig. 2.11. In this system, the reader transmits at the beginning of the communication a modulated RF carrier which is received by the tag. The Schottky diode detector demodulates the signal and a wake up detector tuned to a dedicated wake up signal enables the control and signal processing unit (state machine) on the tag from the standby mode. After this initial wake up phase the reader starts to transmit data. Frequently, delay coded (Manchester coded) amplitude-shift-keying (ASK) modulation is used. This means that the microwave signal is switched on or off in the middle of each data bit. After the demodulation of this pulsed RF signal by the Schottky diode detector, it is led to a Manchester decoder whose binary output signal is fed to the state-machine for further signal processing. A cyclic redundancy check (CRC) calculates the CRC checksum of the received binary sequence in order to recognize transmission errors.

For reading data from the tag, the reader stops sending modulated data and illuminates the tag with a continuous wave (CW) or an unmodulated signal. The tag’s FSK encoder and switch driver switch the load of the tag antenna between two impedance states, causing the radar cross-section of the tag to be changed. As a result, the weak signal reflected from the tag is modulated. This signal is then detected by a homodyne detector at the reader. If the backscatter modulated signal is a double-side band signal, an I/Q demodulator has to be used for the homodyne detection in order to prevent signal cancellation caused by the non-coherent demodulation (see Chapter 3.3.2).

From the point of view of the reader, the backscatter modulation performed at the tag can be considered as a modulation of the radar cross-section $\sigma$ of the tag antenna. The transponder reflects a power $P_{\text{back}}$ that is proportional to the incident power density $S_{\text{in}}$

$$P_{\text{back}} = \sigma \cdot S_{\text{in}}.$$  \hspace{1cm} (2.30)

The maximum radar cross-section is equal to the effective area $A_{e,\text{tg}}$ of the tag antenna

$$\sigma_{\text{max}} = A_{e,\text{tg}} = D_{\text{tg}} \cdot \frac{\lambda^2}{4\pi}.$$ \hspace{1cm} (2.31)

Battery powered tagging systems using passive backscatter modulation are mainly operated in the 2.4 GHz ISM band. This type of system is the most widespread identification system among the commercial available microwave tagging systems. Some typical examples are: (1) Ident-M System from Pepperl & Fuchs with a transmission range of 5 meters at an uplink data rate of 76.8 kbit/s; (2) Confident SI255 from TagMaster with a transmission distance of 4 meters at an uplink data rate of 16 kbit/s and (3) OIS-P RF/ID System
from Baumer Ident with a transmission distance of 6 meters at an uplink data rate of 10.8 kbyte/s.

### 2.4.3 Battery Powered Transponders using Active Backscatter Modulation

Tagging systems with an active backscatter modulator on the transponder may be used to increase the transmission distance without the need of an increased effective isotropic radiated power. This is illustrated by curve $C, C'$ in Fig. 2.10. It has been shown above that the conversion loss of the passive backscatter modulator causes a limitation in the transmission distance. The larger the conversion loss of the passive backscatter modulator the sooner the minimum required power level is reached. If an active backscatter modulator is used the modulated RF carrier is reflected with conversion gain. As illustrated by curve $C'$ this conversion gain is directly converted into an increase of transmission distance.

Active backscatter modulation shows the drawback of an increased power consumption of the transponder. A trade-off between increased power consumption and gain of transmission distance has therefore to be made. Whether the transmission distance is limited by the sensitivity of the reader or that one of the tag is now dependent on the conversion gain of the backscatter modulator. Due to the increased transmission distance, an active backscatter modulation system must be able to be operated in a multi-reader multi-tag environment where several tags are within the range of the reader antenna and narrow spaced readers may interfere with each other. Investigations related to tagging systems using active backscatter modulation are presented in detail in the next chapter.

For the sake of completeness, wireless identification systems are discussed in this section that show an on-board RF source on the transponder. The power consumption of such a transponder is larger than that of a tag applying active backscatter modulation. Transponders with an on-board RF source are utilized only for special applications (e.g. electronic traffic control systems or vehicle tolling applications) where the power consumption is of minor importance and is not limiting the lifetime of a transponder. Transponders having their own transceiver show the possibility of transmitting data to the reader without awaiting a readout command from the reader. Hence, their operation principle is different from a conventionally defined transponder system. The major problem at the operation of a transponder with an on-board RF source is the frequency stability of the oscillator. Typically, the clock rate of the digital signal processing unit is chosen low in order to save DC power. Thus, the data rate is low as well and it might be in the order of magnitude of the FM jitter bandwidth of the RF oscillator on the tag. A frequency stabilization circuit (phase-locked loop) is therefore required on the transponder when homodyne or heterodyne detection is used at the reader. When the RF frequency is not sufficiently stable, only a square-law detector diode could be used for the demodulation at the reader which shows however a much lower receiver sensitivity. Another approach to circumvent this problem is proposed in reference [3] for a short-range communication system. In this system, a linearly polarized CW carrier from the reader is received by the transponder antenna and used as locking signal of an on-board RF oscillator. The output signal of this oscillator is then radiated with orthogonal polarization sense to the reader. The proposed system is to be used for the dedicated short-range communications (DSRC) in the 5.8 GHz ISM band.
2.5 **Summary**

In this chapter, the most widespread tagging systems were presented which include inductively coupled RFID systems or remotely powered microwave tagging systems applying passive backscatter modulation for short-range applications and microwave tagging systems using SAW transponders or battery powered tags for long-range applications. It was shown by means of a power level diagram analysis that an active backscatter modulator may be used to increase the transmission distance. Compared to a complete transceiver the increase of transmission distance is smaller. However, for most applications this is less important than the power consumption and the circuit complexity on the tag which is also larger for a complete transceiver.
3. Active Backscatter Tagging System

This chapter is focused on the investigation of a tagging system that applies active backscatter modulation. Active tagging systems may be used to increase the transmission distance without increasing the effective isotropic radiated power at the reader. This allows to use the identification system for long range applications (max. 20 m) such as e.g. identification of containers. However, the active components on the transponder decrease the battery life. Therefore, a trade-off between gain of transmission distance and loss of battery life has to be done. An increased transmission distance may also have the consequence that a larger number of transponders are within the range of the reader antenna. Hence, the difficulty of operating a tagging system in a multi-reader multi-tag environment becomes of prime importance. This affects the choice of the modulation and coding schemes used in the uplink and downlink communication. System aspects including link budget analysis, modulation schemes and transmitter/receiver topologies for the reader and the tag are therefore the topic of this chapter.

3.1 Basic Operation Principle

Fig. 3.1 illustrates the data transmission between reader and tag in the downlink and uplink communication of an active tagging system. It is assumed that two readers and three tags are operated in the same environment. The transponders tag1 and tag2 are within the transmission distance of the interrogator labeled with reader1. The transmission range of reader2 includes the transponders tag2 and tag3. Each transponder has a unique identification number ID. During the downlink communication, the reader transmits data or commands to the tag. Prior to the transmission of these data, the identification number of the tag is transmitted. As shown in the example, the transponders tag1 and tag2 receive the data from the interrogator reader1 but only tag2 is enabled from the stand by mode in order to store the transmitted data into its memory.

As long as a transponder is not within the beam of two or more reader antennas, the downlink communication is not affected by multiple access problems. However, when two readers are transmitting at the same time, a reliable data transmission is no more possible since the frequency selectivity of the RF detectors used as demodulator on the transponders is not sufficient. For instance, when the interrogator reader2 in the example is transmitting data to tag3 at the same time as the downlink communication between reader1 and tag2 occurs, the latter data transmission will fail because tag2 would not be able to demodulate the signal of reader1 due to the superimposed signal of reader2. This data collision has to be detected by a transmission protocol where the tags must send a data acknowledgement message to the reader as soon as the downlink communication is finished. When no tag response has occurred after a predefined time, the downlink communication has to be repeated.

In the uplink communication, data is read from the transponder. Normally, a tagging system which applies active backscatter modulation performs the uplink communication in two steps. First, a 'read' command has to be transmitted via downlink communication in...
order to instruct the transponder that the reader transmits next a CW carrier for the backscatter modulation. The second step is the backscatter modulation itself where the transponder’s data is modulated on the RF carrier emitted by the reader. The ‘read’ command may also include the request of transmitting the identification number. This case is illustrated in the lower part of Fig. 3.1 where reader2 transmits a CW carrier and the transponders tag2 and tag3 respond with the backscatter modulated identification number. In contrast to the receiver on the transponder, the demodulator and decoder topology of the reader can be chosen such that the superposition of the response signals may not cause a transmission error. In chapter 7.1 and 7.2, modulation and coding schemes are discussed which are well suited for the described multiple access communication in the uplink.

3.2 Downlink Communication

Data or commands are transmitted from the reader to the tag during the downlink communication. In this chapter the optimal choice of the modulation scheme for the downlink communication is discussed. It is shown that circular-polarization-shift-keying (CPSK)
modulation is a modulation scheme that optimally meets the requirements of the downlink communication for an active tagging system. In this work, the application of this modulation scheme for a tagging system is reported for the first time.

3.2.1 Circular-Polarization-Shift-Keying (CPSK) Modulation

The specifications of an ideal transponder ask for small dimensions and a low power consumption. Hence, the following requirements for the transponder’s demodulator used for the downlink communication are demanded

- Low complexity of the circuitry
- Low power consumption
- High sensitivity of the receiver

Particularly, the first and second requirement disagree with the third requirement. As shown in Fig. 2.10, the sensitivity of the receiver on the transponder is less important than the sensitivity of the reader’s receiver. Therefore, crystal video receivers consisting of zero bias Schottky detector diodes are frequently used in a transponder applying passive backscatter modulation. In an active backscatter system, the sensitivity of the receiver on the tag becomes more and more important due to the increased transmission distance. However, this would not justify using a much more sensitive superheterodyne receiver topology on the transponder since the power-saving requirement is still rated higher than the maximization of the transmission distance.

Another important issue of a microwave tagging system is the alignment of the reader and the tag antenna. If the emitted waves from the reader are linearly polarized and the linearly polarized transponder antenna is aligned perpendicular to the E-field vector of the RF carrier, the received signal on the transponder is minimized and the signal might be too weak to be demodulated. No alignment is necessary when the reader transmits circularly polarized waves. Circularly polarized antennas are also preferable for the transponder, if the bandwidth and size specifications make it possible. Compared to a linearly polarized antenna, 3 dB more signal power is gained by using a circularly polarized tag antenna.

It can be summarized from the above discussion that a low complexity receiver (e.g. a crystal video receiver) has to be used on the transponder and the modulated data should be transmitted by a circularly polarized RF carrier. Therefore, amplitude-shift-keying (ASK) modulation is frequently used for the downlink communication due to its low demodulator complexity (video signal of the detector needs just to be sampled by the analog/digital converter of the tag’s control unit). Frequency-shift-keying (FSK) and phase-shift-keying (PSK) modulation are used as well but less frequently. Analog modulation schemes are commonly not applied as the data to be transmitted are binary.

Starting from the requirements concerning a low demodulator complexity and a high sensitivity, a modulation scheme has been invented in this work that is novel for microwave tagging systems. This modulation scheme is called circular-polarization-shift-keying (CPSK) modulation. As shown in Fig. 3.2, it is based on the principle that a logical 1 is transmitted as left-hand circularly polarized (LHCP) waves and a logical 0 is transmitted...
Fig. 3.2: Circular-polarization-shift-keying (CPSK) modulation. A logical 1 is transmitted as left-hand circularly polarized (LHCP) waves and a logical 0 is transmitted as right-hand circularly polarized (RHCP) waves.

as right-hand circularly polarized (RHCP) waves. CPSK can also be considered as on/off-keying modulation of LHCP and RHCP waves.

Fig. 3.3 shows the block diagram of the downlink communication where circular-polarization-shift-keying modulation is used. The modulator as well as primarily the demodulator show a very low complexity of the circuitry. In terms of the demodulator complexity, circular-polarization-shift-keying modulation is superior to most of the conventional modulation schemes.

The circular-polarization-shift-keying modulator consists mainly of a 50 Ω terminated single-pole-double-throw (SPDT) switch. As shown in Fig. 3.3, the terminated SPDT switch connects the RF source to the ports of the polarizer used for emitting left-hand and right-hand circularly polarized waves. The binary data signal $s_{data}$ is modulating the SPDT switch for performing the circular-polarization-shift-keying modulation. The combination of a polarizer (3dB/90° hybrid) and a dual-polarized antenna corresponds to a circularly polarized antenna with switchable polarization sense.

On the tag side, the demodulator consists of a circularly polarized antenna that is able to separate LHCP and RHCP waves. The power of the separated waves is rectified by two identical zero bias Schottky detector diodes which are connected to the two output terminals of the polarizer belonging to the dual polarized antenna. The output signal of the comparator corresponds to the demodulated baseband signal $s_{data}$ and needs only to be sampled by the control unit of the tag for further signal processing (decoding). Due to the comparison of two instantaneous power levels, a higher sensitivity can be achieved than
Fig. 3.3: Block diagram of the downlink communication using circular-polarization-shift-keying modulation. The modulator (reader) consists of a 50 Ω terminated single-pole-double-throw (SPDT) switch and a circularly polarized antenna with switchable polarization sense. The demodulator (tag) consists of a circularly polarized antenna for the reception of LHCP and RHCP waves, two zero bias Schottky detector diodes (crystal video receiver) and a low-power comparator.

by the conventional demodulation of an on/off-keying modulated signal where the output signal of the detector is compared with a fixed threshold voltage.

The presented demodulator for circular-polarization-shift-keying modulated signals shows also the advantage that ideally linearly polarized jamming signals do not affect the demodulation. The power of a linearly polarized jamming signal would be split equally to the two identical RF detector receivers by an ideal circularly polarized antenna. Since the instantaneous power levels are compared for the demodulation, the jamming signal power will be cancelled.

3.2.2 Signal Representation and Analysis of CPSK Modulation

The direction of the E-field vector determines the polarization of an electromagnetic wave. A circularly polarized wave can be decomposed into two orthogonal fields $E_x$ and $E_y$. The circularly polarized radiation fields propagating in z-direction can then be written as [6]

$$E_{RHCP}(z, t) = E \cdot (e_x + je_y) \cdot \exp(j(\omega \cdot t - k \cdot z))$$  \hspace{1cm} (3.1)

$$H_{RHCP}(z, t) = H \cdot (-je_x + e_y) \cdot \exp(j(\omega \cdot t - k \cdot z))$$  \hspace{1cm} (3.2)

$$E_{LHCP}(z, t) = E \cdot (e_x - je_y) \cdot \exp(j(\omega \cdot t - k \cdot z))$$  \hspace{1cm} (3.3)

$$H_{LHCP}(z, t) = H \cdot (je_x + e_y) \cdot \exp(j(\omega \cdot t - k \cdot z))$$  \hspace{1cm} (3.4)

where the subscripts RHCP and LHCP represent the right-hand and left-hand circular polarization sense. The underlined expressions denote complex values and the bold figures represent vectors. The wave number $k$ is defined as
A circularly polarized wave is defined as left-hand circularly polarized if the E-field vector is turning in counterclockwise direction in respect of the propagation direction. Vice versa, the E-field vector of a right-hand circularly polarized wave is turning in clockwise direction.

For the transmission of a bit sequence s(t) where the logical values 0 and 1 are represented by the values -1 and +1, the circular-polarization-shift-keying modulation can be expressed as

\[
E_{RHCP, LHCP} = E \cdot \{e_x + j \cdot s(t) \cdot e_y\} \exp(j \cdot (\omega \cdot t - k \cdot z))
\]

and

\[
H_{RHCP, LHCP} = H \cdot \{-j \cdot s(t) \cdot e_x + e_y\} \exp(j \cdot (\omega \cdot t - k \cdot z)) .
\]

In the analysis of digitally modulated signals, the similarity between any pair of signal waveforms, say \(x_m(t)\) and \(x_k(t)\), is measured by the normalized cross-correlation coefficient

\[
\rho_{m, k} = \frac{1}{\sqrt{\varepsilon_m \cdot \varepsilon_k}} \int_0^T x_m(t) \cdot x_k(t) dt ,
\]

where the signal energy is expressed as

\[
\varepsilon_m = \int_0^T x_m^2(t) dt
\]

and \(T\) denotes the symbol duration. A similar analysis can be performed for the circular-polarization-shift-keying modulated waveforms. By substituting the signal waveforms of Eq. 3.8 by the signal waveforms given by Eq. 3.1 and Eq. 3.3, the normalized cross-correlation function can be obtained

\[
\rho_{RHCP, LHCP} = \frac{1}{\sqrt{\varepsilon_{RHCP} \cdot \varepsilon_{LHCP}}} \int_0^T E_{RHCP}(t) \cdot E^*_{LHCP}(t) dt = 0 ,
\]

where the asterix denotes the conjugate complex and the normalization is given by the energy density functions which can be written as the inner product between the electric displacement \(D\) and the electric field vector \(E\)

\[
\varepsilon_{RHCP} = \frac{1}{2} \cdot D_{RHCP} \cdot E^*_{RHCP} ,
\]
The cross-correlation function in Eq. 3.10 is zero and thus, the signal waveforms of circular-polarization-shift-keying modulation are orthogonal. Circular-polarization-shift-keying is therefore comparable to BFSK modulation. As shown in reference [24] the bit-error-rate for non-coherent binary orthogonal signalling is given by

$$P_b = \frac{1}{2} \exp \left( -\frac{\gamma_b}{2} \right) ,$$

(3.13)

where $\gamma_b$ denotes the signal-to-noise ratio (SNR) per bit.

### 3.3 Uplink Communication

In the uplink communication, the identification code or data from the transponder is read by the reader. As shown in chapter 3.1, the uplink communication is always initiated by the reader by transmitting first a command or a request signal via downlink communication.

This chapter is focused on the investigation of backscatter modulation schemes used for the uplink communication. Backscatter modulation is mainly applied in passive tagging systems where no RF source can be used on the transponder due to power saving requirements. In an active tagging system the increase of transmission distance has to be weighted against a decrease of battery life and thus the performance of an active backscatter modulator and an RF source on the transponder has to be compared carefully in terms of power consumption, circuit complexity and increase of transmission distance. In this work, system topologies using active backscatter modulation are investigated because of the following disadvantages when using an RF source on the transponder:

- Typically, transponders are operated at low data rates. This requires a high frequency stability of the RF source on the transponder as otherwise the transponder signal can not be demodulated by homodyne or heterodyne detection.

- RF oscillators and PLL circuits for this kind of low power applications are not commercially available. Even if the RF source is designed as MMIC (microwave monolithic integrated circuits), its power consumption is still higher than that of an active backscatter modulator. On the other hand, the transmission distance of an active backscatter modulator is shorter. However, for many applications the increase of transmission distance is rated less than the power consumption of the transponder.

In the first part of this chapter, the theory of modulated scatterers is presented. Thereafter, it is shown that homodyne detection of double sideband (DSB) backscatter modulated signals may cause an undesired phase sensitivity of the demodulated signal. The second part is focused therefore on the investigation of system topologies avoiding this undesired phase sensitivity. Each system topology is also discussed in terms of its advantages and disadvantages when implemented in an active tagging system.
3.3.1 Theory of Modulated Scatterers

There is a variety of modulated scatterers (electrical, mechanical, electro-optical, electromechanical modulation) for different applications. The theory presented in this chapter is however limited to backscatter modulation where a nonlinearity (diode, FET) is electrically modulated in a non-resonant scatterer [15].

A good parameter describing the effect of backscattering is given by the scattering cross-section \( \sigma \). It is defined as the area for which the incident field contains sufficient power to produce the same field as is scattered by the scatterer when radiating omnidirectionally. Mathematically, this definition can be expressed by

\[
S_r \cdot 4\pi r^2 = S_i \cdot \sigma, \tag{3.14}
\]

or equivalently,

\[
\sigma = 4\pi r^2 \cdot \frac{S_r}{S_i}, \tag{3.15}
\]

where \( S_r \) denotes the scattered power density at the receiver (reader), and \( S_i \) is the incident power density at the scatterer (tag). The reader and the tag are separated by the distance \( r \).

In order to determine the scattering cross-section \( \sigma \), the illustration of a backscatter system depicted in Fig. 3.4 is used. The backscatter system consisting of one reader and one transponder can be represented by a 3-port system. The first and third port belong to the transmitting and receiving part of the reader and the second port represents the backscatter modulator. This network has the following matrix representation

\[
\begin{bmatrix}
U_1 \\
U_2 \\
U_3
\end{bmatrix} =
\begin{bmatrix}
Z_{11} & Z_{12} & Z_{13} \\
Z_{21} & Z_{22} & Z_{23} \\
Z_{31} & Z_{32} & Z_{33}
\end{bmatrix}
\begin{bmatrix}
I_1 \\
I_2 \\
I_3
\end{bmatrix}, \tag{3.16}
\]

Each matrix element can be computed in terms of the incident electric field \( E_i \) (due to the source \( i \)) at the terminal \( j \) with current distribution \( J_j \)

\[
Z_{ij} = \frac{1}{I_i J_j} \iiint_V E_i \cdot J_j dV = \frac{U_{ij}}{I_j}, \tag{3.17}
\]

where \( U_{ij} \) is the input voltage at terminal \( i \) when driven by a current \( I_j \) with all other terminals open-circuited. The latter relation is a consequence of the reciprocity principle.
The interesting part of Eq. 3.16 for the investigation of backscattering is the determination of the open-circuited voltage \( U_3 \) appearing at the receiving antenna. If \( U_2 = -Z_L I_2 \) is required at the load terminal of the scatterer, and \( I_3 \) is set to zero because the open-circuit voltage at the receiving antenna is desired, the matrix representation of Eq. 3.16 reduces to

\[
\begin{bmatrix}
0 \\
U_3
\end{bmatrix} = \begin{bmatrix}
Z_{21} & Z_{22} + Z_L \\
Z_{31} & Z_{32}
\end{bmatrix} \begin{bmatrix}
I_1 \\
I_2
\end{bmatrix}.
\]  

(3.18)

Eq. 3.18 can now be solved for \( U_3 \) by eliminating \( I_2 \)

\[
U_3 = \left( Z_{31} - \frac{Z_{21} Z_{32}}{Z_{22} + Z_L} \right) I_1.
\]  

(3.19)

\( Z_{31} \) represents the direct signal from the transmitter to the receiver of the reader. This term is not dependent on the backscatter modulator and can be ignored for the analysis of the backscatter modulation.
The interesting part of Eq. 3.19 is the denominator \(Z_{22} + Z_t\) of the second term. \(Z_{22} = R_0 + jX_{22}\) is the self-impedance of the scatterer and \(Z_L = R_L + jX_L\) denotes its load impedance. The radiation resistance \(R_0\) is difficult to vary. Therefore, mainly the resistive part of \(Z_L\) is used for the backscatter modulation.

If short-circuited electric dipoles are used as receiving and transmitting antennas at the reader and the tag, an analytical expression can be derived for the backscattering cross-section [15]. All antennas have the same length \(l\). The electric field in the far-field region of a \(z\)-directed Hertzian dipole of moment \(\mathbf{i} \cdot I\) is

\[
E = e_z \cdot \frac{Z_F \cdot I \cdot l}{2 \lambda r} \cdot e^{-jkr},
\]

where

\(Z_F = 120 \pi \Omega\) : free-space wave impedance

\(I\): current at antenna terminal

\(r\): distance from dipole

\(k = \frac{2 \pi}{\lambda}\) : wave number and \(\lambda\) : wave length

The incident power density \(S_i\) at the scatterer can now be determined using the magnitude of the incident electric field strength \(E_i\) given by Eq. 3.20

\[
S_i = \frac{|E_i|^2}{Z_F} = Z_F \cdot \left(\frac{l}{2 \lambda r}\right)^2 \cdot I_1^2,
\]

where \(I_1\) denotes the current flowing in the transmitting antenna of the reader. The reflected power density \(S_r\) from the scatterer appearing at the receiving antenna of the reader can be expressed by means of the voltage \(U_3\) given by Eq. 3.19

\[
S_r = \frac{|E_r|^2}{Z_F} = \frac{1}{Z_F} \cdot \left(\frac{U_3}{I}\right)^2 = \frac{1}{Z_F} \cdot \left(\frac{Z_{21} \cdot Z_{32}}{Z_{22} + Z_L'}\right)^2 \cdot \left(\frac{I_1}{I}\right)^2.
\]

The relation \(|E_i| = \frac{U_3}{I}\) holds because the length of the electrical dipole is assumed to be much smaller than the wave length \(l \ll \lambda\). Moreover, the term \(Z_{3f}\) has been neglected in Eq. 3.19. Substituting the incident and reflected power densities of Eq. 3.21 and 3.22 into Eq. 3.18 yields
\[
\sigma = 4\pi r^4 \left( \frac{2\lambda}{Z_F \cdot l^2} \right)^2 \cdot \left| \frac{Z_{21} \cdot Z_{32}^2}{Z_{22} + Z_L^2} \right| . \quad (3.23)
\]

Eq. 3.23 can also be expressed in terms of the scatterer’s antenna gain \( G(\theta_r, \phi_r) \) directed towards the reader’s antenna at angles \( (\theta_r, \phi_r) \) and the scatterer’s effective receiving area \( A(\theta, \phi) \) which is dependent on the angles \( (\theta, \phi) \) of the incident waves from the reader. The antenna gain \( G(\theta_r, \phi_r) \) is defined as

\[
G(\theta_r, \phi_r) = \frac{\text{effective power density in } (\theta_r, \phi_r) \text{ direction}}{\text{average power density assuming zero losses}} \quad (3.24)
\]

The average power density emitted by the backscatterer is determined by the radiation resistance \( R_0 = \text{Re}(Z_{22}) \) and the current \( I_2 \) appearing at the terminal of the scatterer’s antenna. Furthermore, the electric field strength at the reader caused by the backscatterer can be written as

\[
E = \frac{U_{32}}{I} = \frac{Z_{32} \cdot I_2}{l} . \quad (3.25)
\]

Thus, the following relation applies

\[
G(\theta_r, \phi_r) = \frac{\left| E_r \right|^2}{Z_F} = \frac{\left| I_2 \right|^2 \cdot R_0}{4\pi r^2} = 4\pi r^2 \left( \frac{Z_{32}}{l} \right)^2 \frac{Z_{22}}{Z_F \cdot R_0} . \quad (3.26)
\]

The effective receiving area \( A(\theta, \phi) \) is defined as

\[
A(\theta, \phi) = \frac{\text{power delivered to a matched load}}{\text{power density of incident wave}} . \quad (3.27)
\]

This expression can be evaluated by first determining the current \( I_2 \) appearing at the antenna terminals of the backscatterer

\[
I_2 = \frac{\frac{1}{2} U_{21}}{R_0} = \frac{I_1 \cdot Z_{21}}{2 \cdot R_0} . \quad (3.28)
\]

The expression for the incident electric field strength \( E_i \) given by Eq. 3.20 is now used together with Eq. 3.28 to obtain the following relation:
Eq. 3.26 and 3.29 can be solved for $Z_{32}$ and $Z_{2j}$, respectively, and substituted into Eq. 3.22. This yields the final expression for the backscattering cross-section

$$
\sigma = 4 \cdot A(\theta_\rho, \phi_\rho) \cdot G(\theta_\rho, \phi_\rho) \cdot \frac{R_0^2}{|Z_{22} + Z_L|^2}.
$$

(3.30)

For coincident transmitting and receiving antennas at the reader, Eq. 3.30 is simplified to

$$
\frac{\sigma}{\lambda^2} = \frac{1}{\pi} G^2(\theta, \phi) \cdot \frac{R_0^2}{|Z_{22} + Z_L|^2},
$$

(3.31)

where the relation $A(\theta_\rho, \phi_\rho) = G(\theta_\rho, \phi_\rho) \frac{\lambda^2}{4\pi}$ has been applied. Numerical results are obtained from Eq. 3.31 by inserting the directivity of 1.5 and the radiation resistance of

$$
R_0 = \frac{h^2}{\lambda^2} \ 790 \ \Omega \quad \text{of a Hertzian dipole antenna.}
$$

Varying the resistive part of the load impedance is the most widely used means for backscatter modulation. Typically, the junction impedance of an RF diode is electrically changed between the two states $R_L, high$ and $R_L, low$. This corresponds also to a change of the backscattering cross-section from $\sigma_{max}$ to $\sigma_{min}$ and vice versa. For the Hertzian dipole antenna, this modulation of the backscattering cross-section can be easily determined using Eq. 3.31. The analytical determination of the backscattering cross-section for other antenna types is very difficult. Microwave tagging systems show frequently patch antennas at the reader and the tag. For instance for patch antennas, the determination of the backscattering cross-section has to be done by using a full-wave Green’s function/Galerkin approach [19].

The backscatter modulation can also be characterized by a modulation index $m$ defined by the relation

$$
m = \frac{\max - \min}{\max + \min}
$$

(3.32)
The terms \( max \) and \( min \) refer to the maximum and minimum backscattering cross-sections \( \sigma_{max} \) and \( \sigma_{min} \). As it can be shown by Eq. 3.31, \( \sigma_{max} \) is inversely proportional to \( z_{L, min} \) and vice versa \( \sigma_{min} \) is inverse proportional to \( z_{L, max}^2 \). Thus, the modulation index for the backscatter modulation of the Hertzian dipole is given by

\[
m = \frac{1}{\sqrt{\sigma_{min}}} - \frac{1}{\sqrt{\sigma_{max}}} = \frac{\sqrt{\sigma_{min}}}{\sqrt{\sigma_{max}} - 1} \frac{1}{\sqrt{\sigma_{min}}} + \frac{1}{\sqrt{\sigma_{max}}} + 1
\] (3.33)

The variation of the load impedance by a data signal on the transponder results in an amplitude modulation index \( m \) according to Eq. 3.33.

### 3.3.2 System Topologies for Backscatter Modulation

Low system complexity, low cost and low power consumption of the transponder are the main requirements for the design of a tagging system. Therefore, backscatter modulation performed by modulating the junction impedance of a semiconductor connected to the antenna terminal is frequently applied. This kind of load modulation generates a backscatter modulated double sideband (DSB) signal. The carrier frequency of the transmitted and received signal at the reader are the same when using backscatter modulation. Thus, it would be convenient to use homodyne detection for the demodulation of the backscatter modulated signal. Homodyne detection of a DSB modulated signal causes however an undesired phase sensitivity of the demodulated signal when a DSB downconverter is applied. This undesired effect can be shown by means of the system topology depicted in Fig. 3.5. First, the reader emits an unmodulated RF carrier given by

\[
x_c(t) = \text{Re}\{A_c \cdot e^{j\omega t}\}.
\] (3.34)

At the tag the incident waves are backscatter modulated by the load modulating data signal

\[
s(t) = \sum_n I_n \cdot g(t - nT) \cdot \cos(\omega_m t).
\] (3.35)

where \( I_n \) is a symbol of the information sequence, \( g(\cdot) \) is the pulse shape function and \( \omega_m \) denotes the subcarrier frequency. For transponders where the load modulation is performed by toggling a pin of the on-board microcontroller, the pulse shape \( g(\cdot) \) is rectangular with \( I_n = \{0, 1\} \) or \( I_n = \{\pm 1\} \) and \( \omega_m = 0 \). The backscattered signal shows a round-trip phase delay of

\[
2\phi = 2kr = \frac{4\pi}{\lambda} r.
\] (3.36)
when it is received by the reader after passing two times the transmission distance $r$. Thus, the backscatter modulated signal can be written as

$$x_{bs}(t) = \text{Re}\{m'(s(t)) \cdot e^{j(\omega t + 2\phi)}\} \approx \text{Re}\{A_{bs} \cdot m \cdot s(t) \cdot e^{j(\omega t + 2\phi)}\},$$  \hspace{1cm} (3.37)

which is a double sideband suppressed carrier amplitude modulated signal (DSBSC-AM). The modulation index $m'(s(t))$ denotes the modulation index of the backscattering cross-section given by Eq. 3.32 which is dependent on the load modulating signal $s(t)$. As explained above, an analytical expression of this modulation index is difficult to obtain for commonly used antenna types such as e.g. slot, patch or printed dipole antennas. The following considerations are focused on system aspects where the knowledge of the analytical expression for $m'(.)$ is of minor importance. Therefore, $m'(.)$ is replaced by the modulation index $m$ of the modulating data signal and thus, the second part of Eq. 3.37 is only an approximation of the backscattered signal.

Using a balanced downconverter with conversion factor $K$ for the homodyne detection of the backscatter modulated signal yields

$$y(t) = K \cdot m \cdot \cos(2\phi) \cdot s(t).$$  \hspace{1cm} (3.38)

The factor $\cos(2\phi)$ is due to the phase difference between the reference signal used for the downconversion and the incident backscatter modulated data signal. It is dependent on the transmission distance and may lead to a cancellation of the demodulated signal if the transmission distance is
where $n$ is an integer. Of course, this phase sensitive term is undesired for the data transmission. In the next part of this chapter, three system topologies are therefore presented that avoid this phase sensitivity of the demodulated signal. Each system topology is also discussed in terms of its advantages and disadvantages for the implementation in an active tagging system.

### 3.3.2.1 Single Sideband Modulator on the Transponder

An approach to circumvent the undesired phase sensitivity of the demodulated signal is to use a single sideband (SSB) modulator on the tag instead of a DSB modulator as shown in the previous section. A block diagram of a passive SSB modulator on the tag is depicted in Fig. 3.6. Before explaining the operation principle of this system topology, it should be demonstrated which effects occur when demodulating the transponder signal by a DSB homodyne detector. The single sideband suppressed carrier signal of the transponder shows the following form

$$x_{bs}(t) = \Re\{A_{bs} \cdot m \cdot [s(t) \pm j \cdot \hat{s}(t)] \cdot e^{j(\omega t + 2\phi)}\},$$

(3.40)

where $\hat{s}(t)$ denotes the Hilbert transform of the modulating data signal $s(t)$ and $\pm$ designates the lower and upper sidebands. The Hilbert transform is defined by the convolution between the filter impulse response $h(t) = 1/\pi t$ and the signal $s(t)$

$$\hat{s}(t) = h(t) \cdot s(t) = \frac{1}{\pi t} \cdot s(t) = \frac{1}{\pi} \int_{-\infty}^{\infty} \frac{s(\tau)}{t-\tau} d\tau.$$

(3.41)

Applying conventional DSB homodyne detection to the signal $x_{bs}(t)$ given in Eq. 3.40, yields the demodulated signal

$$y(t) = K \cdot m \cdot [s(t) \cdot \cos(2\phi) \mp \hat{s}(t) \cdot \sin(2\phi)],$$

(3.42)

which shows a phase-shift distortion if the round-trip phase delay $2\phi$ is not equal to an integer multiple of $2\pi$. The problem of the phase-shift distortion disappears when the modulating signal is a single frequency. For instance, for $s(t) = \cos(\omega_m \cdot t)$ the Hilbert transform is $\hat{s}(t) = -\sin(\omega_m \cdot t)$ and Eq. 3.42 yields

$$y(t) = K \cdot m \cdot \cos(\omega_m \cdot t \mp 2\phi).$$

(3.43)

The passive SSB modulator depicted in Fig. 3.6 consists of a power splitter connected to two mixer diodes whose terminal impedance is modulated by two bit sequences generated in the transponder’s control unit. One of the two bit sequences is delayed by a quarter bit length which corresponds to the required 90° phase-shift of the modulating waveform.
(normally, IF or AF signal) in a conventional SSB modulator [21]. The required $90^\circ$ phase-shift of the RF signal is achieved by an additional $\lambda/8$ transmission line - which is passed two times - in one of the two signal paths. The type of power splitter ($0^\circ$, $90^\circ$ or $180^\circ$) may be arbitrarily chosen depending only on the sideband and carrier suppression. The operation principle of the SSB backscatter modulator using a $3\text{dB}/180^\circ$ hybrid will be illustrated by a short example. Assuming that the incident unmodulated RF carrier has the form

$$x_1(t) = A_c \cdot \cos(\omega \cdot t) , \quad (3.44)$$

then the power divided ($3\text{dB}/180^\circ$ hybrid) signals are

$$x_2(t) = \frac{A_c}{\sqrt{2}} \cdot \cos(\omega \cdot t) \quad \text{and} \quad x_3(t) = \frac{A_c}{\sqrt{2}} \cos(\omega \cdot t) . \quad (3.45)$$

Furthermore, it is assumed that the modulating signal is a single frequency, and hence the signals modulating the terminal impedance of the diodes are

$$x_4(t) = -\sin(\omega_m \cdot t) \quad \text{and} \quad x_5(t) = \cos(\omega_m \cdot t) . \quad (3.46)$$
The backscattered signal consists of the sum of the two modulation products $x_2(t) \cdot x_4(t)$ and $x_3(t) \cdot x_5(t)$. Thus, it can be expressed as

$$x_{bs}(t) = K \cdot \frac{A_c}{\sqrt{2}} \cdot \cos\left(\omega \cdot t + \frac{\pi}{8}\right) \cdot \cos\left(\omega_m \cdot t + \frac{\pi}{4}\right) - \cos(\omega \cdot t) \cdot \cos(\omega_m \cdot t), \quad (3.47)$$

where $K$ denotes the conversion constant of the diode mixers and the phase delay of the additional transmission line has also been considered appropriately. Eq. 3.47 can now be simplified to

$$x_{bs}(t) = K \cdot \frac{A_c}{\sqrt{2}} \cdot \sin((\omega - \omega_m)t), \quad (3.48)$$

which represents the lower sideband of a single sideband suppressed carrier (SSBSC) signal. In practice, a sideband suppression of about 20 dB can be expected.

For a passive tagging system the presented SSB modulator is well suited as it shows low circuit complexity for the reader and for the tag as well. The presented topology shows similarities to a reflection-type phase-shifter when a quadrature hybrid is used as power splitter. In order to convert such a passive SSB modulator into an active SSB modulator, an RF amplifier could be used at the decoupled port of the hybrid. The amplified backscatter modulated signal has to be decoupled from the incident unmodulated waves as well as from the downlink signal. Moreover, the downlink signal and the input signal of the backscatter modulator have also to be decoupled. Ideally, all these signals need to be orthogonal which is however difficult to achieve. In order to assure a sufficient separation of these signals either a second antenna or additional switches are required. In short, the conversion of the described passive SSB modulator into an active modulator shows drawbacks in terms of the tag size and the circuit complexity. Therefore, system topologies are preferred which allow to use a conventional DSB upconverter for the active backscatter modulation on the transponder. In the following sections system topologies are presented where the active modulator generates DSB backscatter modulated signals.

### 3.3.2.2 In-phase/Quadrature-phase Demodulator at the Reader

The undesired phase sensitivity of the demodulated signal can be eliminated by using in-phase/quadrature-phase (I/Q) demodulation in the reader. Fig. 3.7 shows the block diagram of an I/Q demodulator which is similar to an image-rejection mixer where the combining network (3dB/90° hybrid) for the baseband signal has been omitted. When the double sideband modulated signal of the transponder as given by Eq. 3.37 is demodulated by the I/Q demodulator, the signals of the $I$- and $Q$-path have the following form

$$y_I(t) = K \cdot m \cdot s(t) \cdot \cos(2\phi), \quad (3.49)$$

$$y_Q(t) = K \cdot m \cdot s(t) \cdot \sin(2\phi). \quad (3.50)$$
The signals of the I- and Q-path are still phase sensitive. However, combining the signal of the I-path with the 90° phase-shifted signal of the Q-path would eliminate the undesired effect. This can easily be shown when the modulating signal \( s(t) = \cos(\omega_m t) \). Then

\[
y(t) = K \cdot m \cdot \left( \cos(\omega_m t) \cdot \cos(2\phi) + \cos(\omega_m t + \frac{\pi}{4}) \cdot \sin(2\phi) \right),
\]

or, equivalently

\[
y(t) = K \cdot m \cdot \cos(\omega_m t + 2\phi).
\]

The difficulty of this system topology is the implementation of the broadband 90° phase-shifter. For a rectangular pulse shape of the modulating signal (e.g. ASK modulation) on the transponder - like the modulating signal generated by a microcontroller - the phase-shifter for the baseband signal has to be operated over a frequency range of several octaves (compare Fourier series of a rectangular pulse sequence). Such a large bandwidth is not feasible with a conventional analog phase-shifter in this low frequency range. In
order to reduce the bandwidth requirements of the baseband combining network, a modula-
tion scheme must be used where just the fundamental frequency of the modulating bit
sequence is used for the demodulation. E.g. binary FSK modulation is a modulation
scheme which is ideally suited for solving this problem since BFSK modulation can easily
be performed by the microcontroller on the transponder (switching between two rectangu-
lar waveforms). Thus, the block diagram of the reader shown in Fig. 3.7 has been imple-
mented in a system demonstrator set-up. Instead of combining the signals \( y_f(t) \) and \( y_Q(t) \) as
described above by a 3dB/90° hybrid in order to get rid of the phase sensitivity the follow-
ing baseband signal processing has been applied. First, the signals of the I and Q-path are
analog/digital converted and fed to two digital signal processor (DSP) demodulators that
perform the Fourier transform of the sampled signal. A logic unit compares the magnitude
of the BFSK modulated signal components (only the two fundamental frequencies) of the
I and Q-path. As these magnitudes are phase-sensitive as well the logic unit is built in a
way that it evaluates automatically the largest of these two baseband signals. The maxi-
mum deviation from the signal magnitude obtained by using a quadrature hybrid is 30%
(\( \approx 1-\cos(\pi/8) \)).

It has turned out in the demonstrator set-up that an I/Q receiver in the reader is not a pref-
erable system topology for backscatter systems. Apart of the rather complex baseband sig-
nal processing the main problem of this topology is the cross-talk power from the emitted
unmodulated RF carrier into the I/Q receiver. For two separated transmit/receive antennas
an isolation of about 40 dB can be achieved and for a single transmit/receive antenna
where orthogonal polarization senses are applied for the decoupling an isolation of 20 dB
is a typical value. In order to increase the receiver sensitivity a low noise amplifier (LNA)
might be used in front of the I/Q demodulator. However, this results also in the amplifica-
tion of the cross-talk signal which shows the following negative impacts. If the cross-talk
signal becomes too strong it acts as second local oscillator signal. Since its phase is turned
by the RF hybrid in the same way as the phase of the backscattered data signal, the demod-
ulated signals are no more in phase-quadrature. However, if the amplification is chosen
not as high as it may act as pumping signal of the downconverters this effect is negligible.
A much worse impact of the cross-talk signal is that it causes a large DC signal that is
superimposed with the weak data signal. This DC part has to be eliminated as otherwise
the dynamic range of the signal processing part is strongly limited or self-biasing effects
of the mixers occur. The lowpass filters of Fig. 3.7 have therefore replaced in the demon-
strator system by active bandpass filters in order to suppress the undesired DC signal.
Despite of the effort in eliminating the described cross-talk problem a transmission dis-
tance of just 0.5 m has been achieved at 10 mW EIRP with this demonstrator set-up.
Hence, in the next section the third system topology is presented which appeared to be
suited best for active tagging systems as (1) the backscatter modulated signal may be a
DSB signal and (2) the difficulty in the baseband signal processing as shown at the I/Q
demodulator is eliminated.

3.3.2.3 Heterodyne Receiver at the Reader

It can easily be shown that the undesired phase sensitivity of the demodulated signal does
not occur when heterodyne detection is used at the reader. The block diagram of a hetero-
dyne receiver is depicted in Fig. 3.8a. When the incident signal at the reader is assumed to have the form as given by Eq. 3.37, the intermediate frequency of the first downconverter stage becomes

\[ y_{IF1}(t) = K \cdot m \cdot s(t) \cdot \cos((\omega_{LO1} - \omega) t + 2\phi) = K \cdot m \cdot s(t) \cdot \cos(\omega_{IF1} t + 2\phi) \] (3.53)

Due to the non-zero IF, the round-trip phase delay does not cause a phase dependence of the amplitude in the downconverted signal. If the IF is chosen appropriately the undesired sideband can be easily filtered out and the signal will be sampled by the A/D converter for further signal processing.

Generally, tagging systems are operated with relatively low data rates (typically lower than 100 kbit/s) in order to save battery power at the transponder. In the demonstrator system data rates even lower than 10 kbit/s have been used due to restrictions of the micro-

---

**Fig. 3.8:** a) Heterodyne receiver topology using two RF synthesizers; b) Heterodyne receiver optimized for the demodulation of low data rate signals.
controller speed. However, for low data rates of the backscatter modulated signal the FM jitter bandwidth (rms phase jitter) of the RF synthesizers in the reader is in the order of magnitude of the data signal bandwidth. Thus, the demodulation of the IF signal is no more possible.

The system topology used for the final system demonstrator is therefore optimized for the demodulation of low data rate signals. Its block diagram is depicted in Fig. 3.8b. The basic idea of this optimized heterodyne topology is to derive the local oscillator signal of the first downconverter stage directly from RF synthesizer in the transmit path. Thus, the FM jitter of the local oscillator signal and the backscatter modulated signal are identical and the IF signal which corresponds to the difference of these two signals is not affected by the FM jitter of the RF signals. This approach can be achieved by replacing the undesired RF synthesizer of the first downconverter stage by a single sideband upconverter where the output signal of the RF synthesizers in the transmit path is used as RF signal of the SSB upconverter. The modulating frequency of the SSB upconverter is derived from a direct digital synthesizer (DDS) which shows a much better FM jitter performance than the RF synthesizer. In order to reduce the leakage of the transmit signal into the receiver, the SSB upconverter uses double balanced mixers suppressing the RF carrier. In practice, a carrier suppression of typically 20 dB can be achieved. Since the output signal of the SSB upconverter is amplified for being used as local oscillator signal, the leakage power is increased as well which leads to a strong DC signal in the IF signal of the first downconverter stage. The IF frequency has now to be chosen such that the undesired frequency components can be filtered out easily. As the intermediate frequency of the first downconverter stage is too high to be sampled directly, a second downconverter stage is used. The reference oscillator of the DDS used as local oscillator signal for the second downconverter stage is derived from the reference oscillator of the DDS used at the SSB upconverter for the first downconverter stage. This approach helps to reduce the impact of the FM jitter caused by the direct digital synthesizers on the data signal. After filtering the desired sideband, the IF signal of the second downconverter stage is sampled and the further demodulation is performed in the DSP demodulator (Fourier transform of the sampled data) as described above.

Another approach for getting rid of the FM jitter problem at the first downconverter stage would have been to use the same reference signal for the two RF synthesizers depicted in Fig. 3.8a. However, this approach would not have been successful because the lock-in frequency range of the PLL has to be chosen higher than the admissible maximum frequency deviation. As it will be explained in chapter [4.5.2] the BFSK modulation scheme in the uplink uses two FSK tones at $f_1=15$ kHz and $f_2=21$ kHz which give a limit for the loop bandwidth. Due to the demodulation method by determining the Fourier transform of the sampled AF frequency a maximum admissible frequency deviation of $(f_2-f_1)/2=3$ kHz is however required.
3.3.3 System Topology of the Active Read/Write Tagging System Demonstrator operated in the 2.4 GHz ISM Band

In this chapter the final system topology of the demonstrator system is presented. The reader block diagram of the active read/write tagging system is depicted in Fig. 3.9. It can be subdivided into four parts: (1) the host computer for entering or reading data (ID number, commands, etc.) and the control unit (μC) at the reader; (2) the modulator for the downlink communication performing circular-polarization-shift-keying modulation; (3) the heterodyne receiver used in the uplink communication and (4) the RF synthesizer providing the RF carrier generation for the down- and uplink communication. In addition, the RF synthesizer shows frequency hopping capabilities in order to reduce the interference between different readers in a multi-reader environment. The application of frequency hopping methods for the uplink link communication has been motivated by the system demonstrator built at the project partner 'wireless lab' (Hochschule Rapperswil).

The block diagram of the tag is depicted in Fig. 3.10. It consists of three parts: (a) the receiver for the circular-polarization-shift-keying modulated downlink signals; (b) the active modulator for the backscatter modulation in the uplink communication and (c) the control unit (μC) performing demodulation, modulation and data storing tasks. The control unit modulates the IF port of the balanced mixer in the active modulator by an $m$-ary FSK signal with rectangular pulse shape. In order to prevent instabilities of the active

![Fig. 3.9: The reader block diagram as used in the demonstrator tagging system.](image-url)
modulator, the backscattered waves are emitted with orthogonal polarization sense related to the incident waves. By means of two absorptive switches either the receive path (RF detectors) or the transmit path (active modulator) is connected to the antenna which is left-hand and right-hand circularly polarized. The switching between the two paths occurs according to the transmission protocol (readout command from the reader). There is no
dedicated wake up detector circuit on the tag. The wake up procedure is performed by the transmission of a preamble bit sequence in the downlink communication before initiating the data transmission.

Based on the presented block diagrams a system demonstrator of the active tagging system has been built. In Fig. 3.11 the complete system is depicted which consists of a reader connected to a host computer and several active tags. In the following chapter the implementation of the system topology is presented. This includes the design and the characterization of (1) the RF detectors, (2) the antennas with switchable polarization sense, (3) the active modulator, (4) the RF synthesizer and (5) the modulator and demodulator firmware in the reader and the tag.

3.4 Summary

In this chapter, the modulation schemes and system topologies used for the down- and uplink communication were discussed. Circular-polarization-shift-keying modulation was proposed as novel modulation scheme for the downlink communication. It shows orthogonal signalling and thus, its bit error rate performance is equal to that of e.g. noncoherent binary FSK. However, its circuit complexity is much lower than that of a BFSK demodulator.

Active backscatter modulation is used in the uplink communication as the implementation of a complete transceiver on the tag would decrease the battery life too much. It was shown that non-coherent detection of backscatter modulated signals may cause signal cancellation of the demodulated signals due to the well-known phase sensitivity of double-sideband modulated signals. However, a conventional DSB modulator is preferred on the tag in order to reduce the circuit complexity and power consumption. Therefore, a heterodyne receiver was presented as optimum demodulator topology in the reader. An optimized heterodyne topology was shown that is well suited for the demodulation of low data rate signals when the FM jitter of the RF synthesizer is in the order of magnitude of the data rate of the backscattered signal.
4. **System Setup**

In this chapter, the implementation of the system topology shown in Fig. 3.10 (reader) and Fig. 3.11 (tag) is discussed. The key components and circuits of the system setup include:

- RF detector circuit
- Aperture-coupled microstrip patch antennas
- Active modulator
- RF synthesizer

The RF detector circuit is used on the tag for the demodulation of the circular-polarization-shift-keying modulated signals in the downlink communication. In order to increase the transmission distance in the uplink communication, the active modulator on the tag performs backscatter modulation with conversion gain. The performance of the tag antenna is important for the RF detector as well as for the active modulator. In the receive mode, the antenna has to separate LHCP and RHCP waves as good as possible for achieving a high receiver sensitivity of the RF detector. In the transmit mode, the tag antenna has the task of isolating the incident and reflected waves and thus, it helps to assure a stable operation of the active modulator. A high isolation between LHCP and RHCP waves is also required for the reader antenna. At the reader another key component is the RF synthesizer which must be able to perform fast frequency hops.

The following chapters are dedicated to the discussion of the above components and circuits. In the last chapter, the signal processing in the downlink and uplink communication is presented.

4.1 **RF Detector**

Due to severe cost, size and DC power constraints simple crystal video receivers are used in the demodulator of the tag. Schottky-barrier diodes are frequently employed for microwave detector applications. These diodes are majority carrier devices and therefore they do not suffer from charge-storage effects which limit the switching speed of pn-junctions. Especially, zero bias Schottky diode detectors are well suited for tagging system applications as no power supply must be applied. This chapter presents the design of the circular-polarization-shift-keying demodulator which uses low barrier zero bias Schottky detector diodes.

4.1.1 **Detector Circuit Design, Detector and Diode Equivalent Circuits**

Some common types of diode detector circuits are shown in Fig. 4.1. The most common type is the single diode detector circuit because it is low cost and simple. The voltage doubler shows a higher voltage sensitivity and a lower input impedance which simplifies the input impedance matching network. Externally-biased detector circuits are also shown in Fig. 4.1. However, these circuits are not suited for tagging applications. They are typically
formed on n-type silicon with medium and high barrier height. Due to their barrier height, the saturation current is low (in the range of nanoamperes) and they must have an external bias to be operated at small signals. On the other hand zero bias diodes are formed on p-type silicon and they show high values of the saturation current (in the range of 1 to 10 μA).

Diode detectors are essentially low sensitivity receivers which perform direct rectification of the RF signal by means of their non-linear $I$-$V$-characteristic. Generally, detectors can be classified into two distinct types: (1) square-law detectors and (2) linear or peak detectors. In the square-law detector region (or small-signal region), the output voltage of the detector is proportional to the input power of the diode, that is, the output voltage is proportional to the square of the input voltage. In the linear detection region (or large-signal region), the output voltage is proportional to the square root of the input power. The detector diode functions as a switch where it conducts over a portion of the input cycle. Since the output current follows the peaks of the input signal waveform, there is a linear relationship between the output current and input voltage. The detection law for both regions follows the formula

$$V_{out} = K \cdot (\sqrt{P_{in}})^x,$$

where $x$ is two in the small-signal region. At higher power levels in the linear region the diode impedance changes with power and $x$ can even be lower than one. Fig. 4.2 shows a

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**Fig. 4.1: Common types of diode circuits.**

<table>
<thead>
<tr>
<th>Single diode circuits</th>
<th>DC bias</th>
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</thead>
<tbody>
<tr>
<td>Voltage doubler circuits</td>
<td>DC bias</td>
</tr>
<tr>
<td>Low barrier zero bias diodes</td>
<td>Conventional detector diodes</td>
</tr>
</tbody>
</table>
Fig. 4.2: Measured detector output voltage versus input power at 2.44 GHz of a low barrier zero bias Schottky detector diode (HP HSMS-2855) at a video bandwidth of 2MHz.

typical detector transfer characteristic ($V_{out}$ vs. $P_{in}$) which was determined by a low barrier zero bias Schottky detector diode (Hewlett Packard HSMS-2855). For tagging applications, the detector on the transponder is operated almost exclusively in the square-law region as the input signal power at the transponder is mostly below -20 dBm. The dynamic range of the square-law region can be defined as the difference between the 1 dB compression point (1 dB deviation from the ideal square-law response) and the input power corresponding to the tangential signal sensitivity (TSS). Fig. 4.3 illustrates the definition of the TSS. The tangential signal sensitivity is defined as the input power where the negative noise peaks superimposed to the detected voltage $V_{det}$ are equal to the noise peaks when no signal is present. For zero bias diodes, the TSS is dependent on the video bandwidth and the RF frequency due to the frequency dependent voltage sensitivity $\beta_v$.

Fig. 4.4a shows the schematic of a single diode detector and its representation by different equivalent circuits (Fig. 4.4b and Fig. 4.4c). In the schematic, the shunt inductance $L$ provides a current return path for the diode. It is chosen to be large compared to the diode’s impedance at the RF frequency (open-circuited at $f_{RF}$). The bypass capacitance $C$ is chosen such that it provides a short-circuit for the RF signal at the output terminal of the detector. This capacitance has to be sufficiently large that its reactance is small compared to the diode impedance but small enough to avoid that the video circuit is loaded too much by its reactance ($RC$-time constant of the load impedance). Either lumped elements or distributed elements can be used for the shunt inductance and the bypass capacitance in the
detector design. $R_L$ denotes the load impedance which is normally chosen large (e.g. 100 kΩ).

The cross-section of a Schottky barrier diode chip and the corresponding diode equivalent circuit, which was substituted into the RF and video equivalent circuits, are shown in Fig. 4.4d and Fig. 4.4e. In the diode equivalent circuit consisting of five elements, the package parasitic inductance $L_p$ and the package parasitic capacitance $C_p$ are also included. The parasitic series resistance $R_s$ of the diode consists of the sum of the bondwire and lead frame resistance, the resistance of the bulk layer of silicon and other parasitic resistances [29]. RF energy which is coupled into $R_s$ is lost as heat and does not contribute to the rectified output of the diode. This series resistance reduces also the diode sensitivity by reducing the voltage across the diode junction.

The thickness of the epitaxial layer and the diameter of the Schottky contact control the parasitic junction capacitance $C_j$ of the diode. $C_j$ and $R_s$ are also dependent on the barrier height. For a biased Schottky diode, the junction capacitance obeys the equation

$$C_j = \frac{C_{j0}}{\left(1 - \frac{V}{\phi_{bi}}\right)^2}, \quad (4.2)$$
where $V$ is the applied voltage and $\phi_{bi}$ denotes the built-in potential. $C_{j0}$ is the junction capacitance at zero bias voltage. The exponent $1/2$ in the denominator comes from the assumption that the doping density is constant throughout the semiconductor.

$R_j$ is the junction resistance (sometimes also referred as video resistance $R_v$) of the diode where the RF power is converted into the video output voltage. At higher frequencies the junction capacitance $C_j$ shorts out the junction resistance $R_j$ and shunts the RF energy to the series resistance $R_s$. The junction resistance $R_j$ is the inverse of the average slope of the I-V characteristic at the operating point.
Fig. 4.5: Measured I-V characteristic and transconductance of the zero bias Schottky detector diode HP HSMS 2855.

\[ R_j = \left( \frac{di}{dV} \bigg|_{V_j} \right)^{-1} \]  \hspace{1cm} (4.3)

where the I-V characteristic is given by

\[ i = I_s \cdot \left( \exp \left( \frac{1}{n} \Lambda v \right) - 1 \right) \]  \hspace{1cm} (4.4)

The term \( \Lambda = \frac{q}{kT} \) denotes the reciprocal of the thermal voltage with \( q \) being the electron charge (1.602x10^{-19} Coulomb), \( k \) is the Boltzmann constant (1.38x10^{-23} Joules/°K), \( T \) denotes the physical temperature in Kelvins, and \( n \) is the diode ideality factor. The saturation current \( I_s \) at temperature \( T \) is given by [39]

\[ I_s(T) = I_{s0}(T_0) \cdot \left( \frac{T}{T_0} \right)^{2n} \exp \left[ -\frac{q\psi_{ms}}{k}\left(\frac{1}{T} - \frac{1}{T_0}\right) \right] \]  \hspace{1cm} (4.5)

where \( T_0 \) is 293° K (room temperature), \( I_{s0} \) is the saturation current measured at room temperature and \( \psi_{ms} \) is the metal-semiconductor Schottky barrier height. The applied zero
bias Schottky detector diodes of the series Hewlett Packard HSMS 285x (x = #0, #1, #2, #5) show the diode parameters \( I_{s0} = 3 \mu A, n = 1.2 \) and \( \psi_{ms} = 0.35 eV \).

Performing the derivative of Eq. 4.4 with respect to the voltage \( v \) across the junction as indicated by Eq. 4.3 yields the following expression for the junction resistance

\[
R_j = \left( \frac{\Lambda}{n} \cdot I_s e^n \bigg| I_0 \right)^{-1} = \frac{n k T}{q \cdot (I_{s0} + I_0)}.
\]  

(4.6)

The junction resistance \( R_j \) is a function of the saturation current \( I_{s0} \) which is constant at a given temperature and the current \( I_0 \) which depends on the rectification of the RF signal, the RF input power and the RF input impedance match.

### 4.1.2 Voltage Sensitivity

For tagging applications the voltage sensitivity \( \beta_v \) and the tangential signal sensitivity TSS are the most important features of the detector diodes. In this section the voltage sensitivity will be derived.

By expressing the I-V relationship (Eq. 4.4) of the diode into a Taylor’s expansion and applying an RF carrier of the form \( v_d \cos(\omega t) \) (where \( v_d \) is the peak amplitude) to the detector diode’s input terminal it can be shown that the rectified current mainly stems from the second order term in the Taylor’s expansion [38]

\[
\Delta i = \frac{v_j^2}{2} \cdot \frac{1}{\frac{d^2 i}{dv^2}} \bigg|_{I_0},
\]  

(4.7)

where \( v_j \) denotes the peak of the ac voltage across the diode junction. The second order derivation of the diode current \( i \) with respect to the voltage \( v \) across the junction can be derived from Eq. 4.6

\[
\frac{d^2 i}{dv^2} \bigg|_{I_0} = \left( \frac{\Lambda}{n} \right)^2 \cdot (I_{s0} + I_0) = \frac{\Lambda}{n} \cdot \frac{1}{R_j}.
\]  

(4.8)

Substituting Eq. 4.8 into Eq. 4.7 and adding the effect of the parasitic series resistance \( R_s \), which reduces the rectified current by the factor \( R_j/(R_s + R_j) \) [40], yields

\[
\Delta i = \frac{v_j^2}{4} \cdot \frac{\Lambda}{n \cdot R_j} \cdot \frac{R_j}{R_s + R_j}.
\]  

(4.9)
In the next step the power absorbed in the diode is determined. The small signal diode equivalent circuit including the package parasitics as shown in Fig. 4.4 is used for calculating the diode’s admittance

\[ Y_d = \frac{1 - \omega^2 R_s R_j C_p C_j + j\omega (R_s C_p + R_j C_j + R_j C_j)}{R_j + R_s - \omega^2 L_p (R_s C_p + R_j C_j + R_j C_j) + j\omega (L_p - \omega^2 R_s R_j L_p C_p C_j + R_j R_s C_j)}. \] (4.10)

The absorbed power in the diode can be expressed by

\[ P = \frac{V_d^2}{2} \cdot \text{Re}[Y_d]. \] (4.11)

The voltage \( v_j \) across the junction is determined by the voltage divider

\[ v_j = \frac{R_j}{1 + j\omega R_j C_j} \cdot |Y_d| \cdot v_d. \] (4.12)

Substituting for \( v_d \) from Eq. 4.12 into Eq. 4.11 yields

Fig. 4.6: Calculated voltage sensitivity \( \beta_v \) versus junction capacitance \( C_j \) and junction resistance \( R_j \) based on the parameters of the Schottky diode HP HSMS 2855 [29]: \( n=1.055, R_s=25 \ \Omega, \ C_\theta=0.08 \ \mu\text{F}, \ L_p=2 \ \text{nH, } R_L=100 \ \text{k}\Omega \) and \( \tau=0.316 \) (\( S_1=-10 \ \text{dB at } f=2.44 \ \text{GHz} \)).
where

\[ N_1 = (1 - \omega^2 R_s R_j C_p C_j) \cdot \left[ R_j + R_s - \omega^2 L_p (R_s C_p + R_j C_p + R_j C_j) \right], \quad (4.13a) \]

\[ N_2 = \omega^2 \left( L_p - \omega^2 R_s R_j L_p C_p C_j + R_j R_s C_j \right) \cdot (R_s C_p + R_j C_p + R_j C_j). \quad (4.13b) \]

The ratio between the detected current (Eq. 4.9) and the power absorbed in the diode (Eq. 4.13) is defined as detector current sensitivity

\[ \beta_i = \frac{\Delta i}{P} \quad (4.14a) \]

\[ = \frac{\Delta}{2} \cdot \frac{R_j^2}{n} \cdot \frac{1 - \omega^2 R_s R_j C_p C_j}{(R_j + R_s) \cdot (1 + (\omega R_j C_j)^2) \cdot (N_1 + N_2)}. \quad (4.14b) \]

The voltage sensitivity \( \beta_v \) is defined as the product of the current sensitivity \( \beta_i \) times the junction resistance \( R_j \). Since the detector diode may be considered as a video voltage source of impedance \( R_v \) feeding a load resistance \( R_L \), the voltage sensitivity is reduced by the ratio of \( R_L \) to \( R_v + R_L \). The video resistance is equal to the sum of the junction resistance \( R_j \) and the parasitic series resistance \( R_s \) (the reactive elements of the equivalent circuit do not affect the reduction of \( \beta_v \)).

So far the analysis has assumed that all incident power is absorbed by the diode. In particular for a zero bias diode, the design of a broadband matching network is not possible (no bias is applied for lowering the junction resistance) and therefore reflection losses have also to be considered in the calculation of the voltage sensitivity. The voltage sensitivity will be reduced by a factor \( (1 - |r|^2) \) where \( r \) denotes the reflection coefficient of the diode. The final expression for the calculation of the voltage sensitivity is thus given by

\[ \beta_v = \beta_i \cdot R_j \cdot \frac{R_L}{R_j + R_L} \cdot (1 - |r|^2). \quad (4.15) \]

The voltage sensitivity given by Eq. 4.15 has been calculated for the zero bias Schottky detector diode HP HSMS 2855 which is used in the tag demodulator. Fig. 4.6 shows the calculated voltage sensitivity versus the junction resistance \( R_j \) and the junction capacitance \( C_j \) which are dependent on the operation point. The voltage sensitivity is maximum for a certain \( R_{j,\text{opt}} \) which belongs to an operation point with the total current \( I_{T,\text{opt}} = I_{0,\text{opt}} + I_{\text{s,0}} \). For currents greater than \( I_{T,\text{opt}} \) or, correspondingly, for a junction resistance \( R_j \) lower than \( R_{j,\text{opt}} \), \( \beta_v \) drops due to the reduced voltage across the diode junction.
the current is less than \( I_{\text{opt}} \), the sensitivity is reduced due to the \( R_L/(R_f+R_s+R_L) \) voltage divider because \( R_f \) gets large relative to \( R_L \). As shown in Fig. 4.6, the maximum voltage sensitivity of 38 mV/\( \mu \)W at \( R_{j,\text{opt}}=12 \) k\( \Omega \) is close to \( \beta_v \) of the zero biased diode obtained at \( R_{j0}=9 \) k\( \Omega \) and \( C_{j0}=0.8 \) pF.

In order to compare the calculated voltage sensitivity with measurements, Fig. 4.7 shows the measured \( \beta_v \) at an input power of -30 dBm across the 2.4 GHz ISM band. It can be seen that the maximum voltage sensitivity is about 26 mV/\( \mu \)W which is slightly less than the maximum \( \beta_v,\text{max} \) of 30 mV/\( \mu \)W specified in the diode’s data sheet [29] that can be achieved with a narrow band impedance match. As shown by the characteristic of \( \beta_v \) in Fig. 4.7, the measured detector is designed to cover the whole ISM band (about 4% relative bandwidth) and that is why the impedance match is not as good as to achieve \( \beta_{v,\text{max}} \).

### 4.1.3 Tangential Signal Sensitivity

There are three noise sources that can be distinguished in the detector diode: (1) shot noise, (2) thermal noise and (3) flicker noise which varies as the inverse of frequency (1/\( f \) noise). These noise sources limit the minimum detected signal strength.

The mean-square shot noise current generated in the junction resistance of the Schottky barrier diode within the bandwidth \( B \) is given by [41]

\[
\overline{I^2_s} = 2e \cdot I_0 \cdot B + 4e \cdot I_{s0} \cdot B ,
\]

(4.16)
which can also be transformed into a noise voltage

\[ \overline{v_s^2} = 2e \cdot B \cdot R_s^2 (I_0 + 2I_{s0}) . \]  

(4.17)

The thermal noise due to the parasitic series resistance \( R_s \) can be expressed as \[42\]

\[ \overline{v_{th}^2} = 4k \cdot T \cdot B \cdot R_s . \]  

(4.18)

The junction of metal, silicon and passivation around the rim of the Schottky contact (see Fig. 4.4d) is the main source of the flicker noise \[43\]. This noise can severely reduce the sensitivity of the Schottky diode detector if the video frequency is below the noise corner of the flicker noise (the corner frequency \( f_c \) is defined as the frequency where the flicker noise equals the shot noise generated in the junction resistance). The flicker noise can be approximated by \[38\]

\[ \overline{v_{1/f}^2} = 2n \cdot k \cdot T \cdot R_j \cdot f_c \ln \left( 1 + \frac{B}{f_c} \right) , \]  

(4.19)

where \( f_c \) denotes the lower frequency limit of the video bandwidth \( B \). p-type silicon Schottky diodes used for detector applications show much lower flicker noise than n-type Schottky diodes. In particular, zero bias Schottky diodes have very low values of flicker noise. Therefore, in the following derivation of the tangential signal sensitivity the contribution of the flicker noise is neglected. The total noise generated in the diode is then

\[ \overline{v_n^2} = 2e \cdot B \cdot R_j^2 (I_0 + 2I_{s0}) + 4k \cdot T \cdot B \cdot R_s . \]  

(4.20)

In order to measure the detector output voltage \( \beta_v P_{in} \), a post amplifier with gain \( G \) is used as shown in the setup depicted in Fig. 4.8. The output voltage of the amplifier consists of a signal term

\[ v_s = \beta_v \cdot G \cdot P_{in} \]  

(4.21)

and a noise term

\[ v_s = v_s + v_{n,a} \]

Fig. 4.8. Measurement setup for the determination of the detector’s noise performance.
\[ \overline{v_{n,a}^2} = (2e \cdot B \cdot R_j^2(I_0 + 2I_{s0}) + 4k \cdot T \cdot B \cdot (R_s + R_a)) \cdot G^2, \]  

(4.22)

where the thermal noise generated in the amplifier has been taken into account by the equivalent noise resistance \( R_a \) of the amplifier. The ratio between Eq. 4.21 and the square root of Eq. 4.22 is defined as signal-to-noise ratio

\[ S = \frac{\beta_v \cdot P_{in}}{\sqrt{2e \cdot B \cdot R_j^2(I_0 + 2I_{s0}) + 4k \cdot T \cdot B \cdot (R_s + R_a)}}. \]  

(4.23)

As mentioned above, the tangential signal sensitivity is measured by an oscilloscope (see Fig. 4.3) where the TSS is defined as the input power when the noise peaks (denoted below by \( v_{peak} \) which is equal to \( \sqrt{2}v_n \)) without applied signal are about half the detected voltage with applied RF signal. Thus, the condition for the tangential signal sensitivity can be expressed as

\[ v_{peak} = v_{det} - v_{peak} \]  

(4.24)

or equivalently

\[ v_{det} = 2v_{peak} = 2\sqrt{2}v_n. \]  

(4.25)

Since \( v_{det} = \beta_v P_{in} \) the tangential signal sensitivity can be written as

\[ TSS = \frac{2\sqrt{2} \cdot v_n}{\beta_v} = \frac{2.82 \cdot \sqrt{2e \cdot B \cdot R_j^2(I_0 + 2I_{s0}) + 4k \cdot T \cdot B \cdot R_s}}{\beta_v}. \]  

(4.26)

Fig. 4.9 shows a comparison between the calculated and measured tangential signal sensitivity. The calculations are based on Eq. 4.26 where the values given in Table 4.2 have been used.

**TABLE 4.2**

<table>
<thead>
<tr>
<th>( B )</th>
<th>[ \text{variable from } 1\text{kHz to } 2\text{MHz} ]</th>
</tr>
</thead>
<tbody>
<tr>
<td>( R_j )</td>
<td>9 kΩ</td>
</tr>
<tr>
<td>( I_{s0} )</td>
<td>3 μA</td>
</tr>
<tr>
<td>( I_0 )</td>
<td>0</td>
</tr>
<tr>
<td>( k )</td>
<td>1.38·10⁻²³ Ws/K</td>
</tr>
<tr>
<td>( T )</td>
<td>293 K</td>
</tr>
<tr>
<td>( R_s )</td>
<td>25 Ω</td>
</tr>
<tr>
<td>( e )</td>
<td>1.602·10⁻¹⁹ As</td>
</tr>
<tr>
<td>( \beta_v )</td>
<td>30 mV/µW</td>
</tr>
</tbody>
</table>
The setup for the measurement of the tangential signal sensitivity is depicted in Fig. 4.10. The generation of the on/off-keying modulated RF signal is performed by upconverting a unipolar (‘0’ and ‘+A’) square wave signal. The DC offset (‘+A/2’) in the pulsed video signal causes the RF carrier to be switched on and off by the upconverter (a polar (‘-A’ and ‘+A’) signal would only cause a binary phase-shift-keying modulation of the RF carrier).

**Fig. 4.9:** Calculated and measured tangential signal sensitivity of a zero bias low barrier Schottky diode (HSMS 2855) at $f = 2.44$ GHz.

**Fig. 4.10:** Setup for the measurement of the tangential signal sensitivity (TSS).
A small part of the pulsed RF carrier is branched off by a 10 dB coupler and is measured by a spectrum analyzer. After passing a video amplifier, the detector output signal is displayed on an oscilloscope. The video amplifier shows at its input an adjustable lowpass filter.

As can be seen in Fig. 4.9, the lowest tangential signal sensitivity that could be measured with the described setup is slightly less than -60 dBm. For an input power below -60 dBm or, equivalently, for a video bandwidth smaller than 100 kHz, measurements of the TSS are no more possible as the feed through of the synchronization signal from the square wave generator to the oscilloscope gets noticeable and prevents accurate measurements. Above a video bandwidth of 100 kHz, the deviation between measured and calculated tangential signal sensitivity is admissible when considering that there is some scope in the determination of the TSS on the oscilloscope.

The determination of the tangential signal sensitivity by means of an oscilloscope where the bottom level of the pulse coincides with the top level of the noise on either side of the pulse is to a certain degree a subjective measurement. An alternative measurement for the determination of the TSS will be described below.

An equivalent expression for the tangential signal sensitivity is the signal-to-noise ratio at the detector output terminal. A generally accepted signal-to-noise ratio which corresponds well with the TSS condition is 8 dB [44]. This SNR corresponds to a voltage ratio of 2.5 at the output. Since the detector is operated in the square law region, the corresponding SNR at the input is 4 dB. The measurement of the tangential signal sensitivity is then performed by using a root-mean-square (rms) voltmeter instead of displaying the pulse signal (audio signal of the detector) on an oscilloscope. First, the noise level is observed on the rms voltmeter when the RF signal is switched off. Then, the RF signal is applied and adjusted such that the increase in the rms voltmeter reading is 8 dB or more which means that the peak signal voltage is 2.5 times the rms noise voltage $v_n$. For an AC coupled voltmeter, the square wave signal is symmetrical with amplitude $1.25v_n$. Thus the total voltage on the rms voltmeter is

$$v_t = \sqrt{v_n^2 + (1.25v_n)^2}.$$  \hspace{1cm} (4.27)

### 4.1.4 Diode Measurements

In order to determine the diode parameters and the elements of the small signal equivalent circuit depicted in Fig. 4.4e, DC and RF measurements must be performed. DC measurements are suited for the determination of the ideality factor $n$, current parameter $I_{s0}$, and junction resistance $R_j$ which can be found from the $I-V$ characteristic. At low currents the change in voltage per decade of current is constant as only little voltage drop occurs across the series resistance. Thus, some simple modifications of Eq. 4.4 yield
Fig. 4.11: $S_{21}$ of the diode HSMS 2855 mounted in shunt to a transmission line. Maximum loss of 11 dB occurs at a resonance frequency $f_0=7.38$ GHz. The frequencies corresponding to a 3 dB reduced loss are $f_1=8.15$ GHz and $f_2=6.5$ GHz.

\[
\begin{align*}
n &= \Lambda \cdot \log(e) \cdot \Delta v = \frac{\Delta v}{0.05783}, \quad (4.28) \\
I_{s0} &= I(v) \cdot \exp\left(\frac{-\Lambda \cdot v}{n}\right), \quad (4.29) \\
R_j &= \frac{n}{\Lambda} \cdot \frac{1}{I_{s0} + I_0}. \quad (4.30)
\end{align*}
\]

where $\Delta v$ is the change in junction voltage per decade of current. Knowing $n$, the saturation current $I_{s0}$ is determined by

where $I(v)$ is the diode current at a voltage drop $v$ across the diode and the additive -1.0 in Eq. 4.4 has been ignored. According to Eq. 4.6 the junction resistance is given by

In Eq. 4.30 $I_0$ denotes the total external current which is dependent on the diode's operation point.

RF measurements may be performed to determine the series resistance $R_s$, junction capacitance $C_{j0}$, and parasitic series inductance $L_s$. These measurements are based on the determination of the transmission scattering parameter in a measurement setup where the diode is mounted in shunt to a transmission line of impedance $Z_0$ (measurement setup proposed by DeLoach [45]). By means of the scattering parameter $S_{21}$ the maximum loss $L$ can be
found. The frequencies \( f_1 \) and \( f_2 \) corresponding to a loss less than 3 dB of the maximum loss are also needed for the following equations of \( R_s \), \( C_{j0} \) and \( L_p \):

\[
L = 10^{\frac{L[\text{dB}]}{20}}
\]  

(4.31)

\[
R_s = \frac{Z_0}{2} \cdot \frac{1}{\sqrt{L - 1}}
\]  

(4.32)

\[
C_{j0} = \frac{1}{\pi \cdot Z_0 \cdot f_1 \cdot f_2} \left( \sqrt{L - 1} \right) \sqrt{1 - \frac{2}{L}}
\]  

(4.33)

\[
L_p = \frac{1}{4 \cdot \pi^2 \cdot f_1 \cdot f_2 \cdot C_{j0}}
\]  

(4.34)

Substituting the parameters \( L, f_1 \) and \( f_2 \) given by Fig. 4.11 into Eq. 4.31 - 4.34 yields \( R_s = 28 \, \Omega \), \( C_{j0} = 1.15 \, pF \), \( L_p = 4 \, nH \).

### 4.1.5 Matching

The most difficult part of the detector circuit design is the input impedance matching network. When maximum sensitivity is required over a narrow band of frequencies, a reactive matching network is optimum. Such networks can be realized in either lumped or distributed elements, depending upon frequency, size constraints and cost limitations.

![Matching Network Diagram](image)

**Fig. 4.12:** 2.4 GHz matching network for the detector circuit using the diode HP HSMS 2855 on the substrate Duroid 6010 (\( \varepsilon_r = 10.2 \), \( h = 25 \, \text{mil} (= 0.635 \, \text{mm}) \) and \( \tan(\delta) = 0.0035 \) @ 10 GHz).
A typical block diagram of a detector circuit is shown in Fig. 4.4a. It consists of a separate impedance network and DC current return path at the input of the detector diode. For a distributed matching network, it is possible to include the DC current return path into the matching network by means of a short-circuited shunt reactance. However, vias used as short-circuits show a parasitic series inductance which might be difficult to include into the matching network design. Moreover, the termination of a short-circuited transmission line is not as well defined as that of an open-circuited transmission line. High sensitivity detectors require narrow band matching networks and thus tolerances in the matching network elements must be small. Therefore, in the present design a separation of the DC return path and the matching network was preferred.

For tagging applications very low loss matching networks are demanded. Instead of matching the diode impedance to 50 Ω, it might be matched to the conjugate complex of the antenna impedance. Since the diode impedance shows typically a high susceptance, the impedance of the antenna should be in the upper half of the Smith Chart close to the conjugate complex of the diode’s impedance so that no impedance transformer is necessary between diode and antenna. As shown in Fig. 3.11, there is an absorptive switch at the antenna terminals in order to select the operation mode (RF detector or active modulator) for the downlink and uplink communication. Due to this absorptive switch, the terminal impedance of the antenna has to be matched to 50 Ω and thus the detector diodes must also be matched to 50 Ω.

![Graph](image_url)

**Fig. 4.13:** Reflection scattering parameter $S_{11}$ from 2.3 GHz to 2.6 GHz at different points of the detector circuit as marked in Fig. 4.12. Curves (A) and (C) are based on measured data; Curve (B) is simulated.
Frequently, the DC current return path is located between matching network and detector diode. However, in the present detector design the DC current return path is located in front of the matching network. The reason is that the impact of the DC current return path on the input impedance caused by the parasitics of the via is less as the diode impedance is much larger than the characteristic impedance.

Fig. 4.12 shows a typical schematic of a matching network consisting of distributed elements for the zero bias Schottky detector diode HSMS 2855. Distributed elements have also been used to build the capacitance after the diode for providing a short-circuit to the RF signal and the inductance of the DC current return path. These open-circuited and short-circuited shunt transmission lines show both an electrical length of about λ/4. The actual matching network consists of a series transmission line and an open-circuited stub. Fig. 4.13 shows the impacts of these matching elements. The diode is mounted on a 25 mil (0.635mm) thick substrate with a relative permittivity of ε_r = 10.2 and a dissipation factor of 0.0035 specified at 10 GHz. Curve (A) on the Smith chart indicates the impedance of the unmatched diode from 2.3 GHz to 2.6 GHz (ISM band reaches from 2.4 GHz to 2.4835 GHz). It can be seen that the input impedance of the diode is highly capacitive. In order to match the diode input impedance to 50 Ω, first a series transmission line rotates the diode impedance to the inductive upper part of the Smith chart. This series transmission line corresponds to a series inductance in a lumped element matching network. The

![Fig. 4.14: Comparison between calculated and measured S11 of the zero bias Schottky detector circuit shown in Fig. 4.12.](image-url)

\[ S_{11} [dB] \]
resulting impedance locus is labeled with (B). An open-circuited stub line corresponding to a shunt capacitance is now used to bring the impedance locus to the 50 Ω point in the Smith chart. The length of the stub line is adjusted such that midband match is traded for bandwidth. As shown by the resulting curve (C), the impedance locus in the Smith chart has been stretched due to the effect of dispersion from the matching elements. Since the input impedance of the unmatched diode is at the edge of the Smith chart, large reactances must be used in the matching network to obtain a 50 Ω impedance match. Hence, the effect of dispersion is getting strong and the bandwidth of the detector is reduced significantly. In order to get the diode impedance farther away from the Smith chart edge and thus making the impedance matching easier either a shunt resistance could be added at the diodes input terminal or the video resistance could be reduced by biasing the diode. However, both effects would lower the sensitivity of the detector.

Fig. 4.14 shows the comparison between measured and simulated $S_{11}$ of the detector circuit depicted in Fig. 4.12. This detector circuit was also used for the measurements of the
voltage sensitivity and the tangential signal sensitivity shown in Fig. 4.7 and Fig. 4.9. It is obvious by comparing the voltage sensitivity plot with the magnitude of $S_{ij}$ that there are some similarities. Indeed, the calculation of the voltage sensitivity $\beta_v$ given by Eq. 4.15 states that $\beta_v$ is multiplied by the term $(1 - |r|^2)$ where $r$ is the input reflection factor which corresponds to the magnitude of $S_{jj}$. The impedance matching of about 4 dB for the measured detector circuit is relatively poor. This can also be verified by the maximum measured voltage sensitivity (see Fig. 4.7) of 26 mV/μW which is less than $\beta_{v, max} = 30$ mV/μW obtained for a narrow band match. This less than optimum performance is caused by the trade-off between midband match and bandwidth. As already mentioned in Chapter 3.3.3, the reader may use frequency hopping methods in order to reduce the interference between different readers in a multi-reader environment or to circumvent channel fading effects. Therefore, the detectors have been designed to show a relatively flat characteristic of the voltage sensitivity over the 2.4 GHz ISM band (4% relative bandwidth) and thus midband match was traded for bandwidth as shown by curve (C) in Fig. 4.13. Moreover, a certain loss in the performance of $\beta_v$ and $\beta_{S}$ may be accepted as the transmission distance of an active tagging system is mainly limited by the sensitivity of the reader in the uplink communication and not by the sensitivity of the transponder in the downlink communication (see also Fig. 2.10).

The layout of the detector circuit presented in this chapter is depicted in Fig. 4.15b. The dimensions (indicated by the electrical length and the impedance) of the matching network, the DC current return path and the open-circuited stub line at the output terminal of the detector are given in Fig. 4.12. Fig. 4.15a shows the layout of a detector circuit which shows a slightly different matching network (tapered series transmission line) and lumped elements for the DC current return path (short-circuited inductivity) and the RF signal short (short-circuited capacitance). The layout of a detector circuit using a voltage doubler is depicted in Fig. 4.15c. A DC current return path is not necessary as this would short out the shunt diode of the voltage doubler.

Compared to the basic detector topology shown in Fig. 4.4a the detector circuits in Fig. 4.15 include no low pass filter at the output of the diode. The lowpass filter needed for limiting the noise bandwidth is inherently given by the comparator transfer characteristic on the transponder control unit PCB.

4.1.6 Design of the Tag Demodulator

A photography of the tag RF front end is shown in Fig. 4.16. The RF front end electronics (RF detector circuits, active modulator) is located on the feed layer of the tag’s aperture-coupled patch antenna. The nuts of the plastic screws holding the different layers together can be seen distinctly on the photography (some complete tags are shown on the photography of Fig. 3.12). In order to keep the size of the distributed elements in the RF electronics small, a high permittivity substrate has been used ($\varepsilon_r = 10.2$, thickness $h = 25$ mil). The size of the depicted RF front end is about 70 mm x 100 mm with a thickness of at maximum 8 mm. A second equally sized FR-4 printed circuit board including the battery and
the digital signal processing and control electronics is mounted by four bolts to the RF front end (the drilling holes for the spikes are visible at the corners of the RF front end).

The RF front end shown on the photography corresponds to the block diagram depicted in Fig. 3.11. It can be seen that the terminals of the antenna’s quadrature hybrid are connected to two absorptive switches (M/A-COM SW338) which are controlled by the tag microcontroller. In the downlink communication mode, the terminals of the quadrature hybrid, which correspond to the received left-hand and right-hand circularly polarized waves, are switched to the input of two identical RF detector circuits. For the presented RF front end, the RF detector shown in Fig. 4.15b has been used. The output of the detector circuits labeled with $v_{LHCP}$ and $v_{RHCP}$ are fed to a comparator circuit on the control unit PCB mounted on the top of the RF front end.

In order to determine the performance of the tag demodulator with respect to the transmission distance and the bit error rate, a separate setup of the downlink communication circuits (of the tag and the reader as well) has been built. This setup was used for the bit error rate measurements presented in the next chapter.
4.1.7 Bit Error Rate Measurement

In this section, the bit error rate performance is measured in the downlink communication. Fig. 4.17 shows the measurement setup. The modulator consists of an absorptive switch which is driven by the transmitter of the bit error rate test set (Wandel & Goltermann PF 4). It performs circular-polarization-shift-keying modulation by switching the RF signal either to the LHCP or RHCP antenna port at a data rate of 44 kbit/s. The conversion loss of the modulator is about 14 dB. At the demodulator the antenna separates the LHCP and RHCP waves and feeds the signals to two identical RF detectors consisting of zero bias Schottky detector diodes (HP HSMS 2855). The output of the detectors are compared by a comparator (OpAmp, National Semiconductor LF357N) whose binary ('0', '+A') output signal is fed to the receiver of the bit error rate test set. For a fixed transmission distance \(d=0.5\ m\) the transmitted power is varied in order to obtain different signal-to-noise ratios at the demodulator and thus different bit error rates can be measured.

Generally, bit error rate curves are plotted versus the signal-to-noise ratio \(\gamma_b\). However, the available measurement equipment does not allow to determine \(\gamma_b\) of the demodulator on the transponder. Therefore, a direct comparison between the theoretical BER performance given by Eq. 3.13 and the measured SNR is not possible. In the measurement setup the SNR at the demodulator is varied by means of the transmitted power \(P_{tx}\) at the modulator. Therefore, in the following Eq. 3.13 is modified such that the theoretical BER curve is related to the transmitted power in order to establish a connection to the measurement. Thereby, the SNR of the detector diodes given by Eq. 4.23 is used instead of the nonavailable SNR of the demodulator. It should be noted that this approach ignores any nonideali-

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**Fig. 4.17: Setup for measuring the bit error rate performance of the circular-polarization-shift-keying modulation used in the downlink communication.**
ties of the demodulator and assumes a perfect envelope detection of orthogonal signals (see Chapter 4.2.2 for the consideration of the antenna’s nonideality in the BER determination).

Substituting the signal-to-noise ratio given by Eq. 4.23 into Eq. 3.13 yields

\[
P_b = \frac{1}{2} \exp \left( -\frac{\gamma_b}{2} \right) = \frac{1}{2} \exp \left( -\frac{\beta_y \cdot P_{in}}{2 \sqrt{2} e \cdot B \cdot R_j^2 (I_0 + 2 I_{s0}) + 4k \cdot T \cdot B \cdot (R_x + R_a)} \right). \tag{4.35}
\]

Since \(P_{in}\) denotes the received power at the output terminals of the antenna it must be expressed in terms of the transmitted power \(P_{tx}\). The relationship between \(P_{in}\) and \(P_{tx}\) is given by Eq. 2.25 which can be substituted into Eq. 4.35

\[
P_b = \frac{1}{2} \exp \left( -\frac{\beta_y \cdot 1 mW \cdot 10^{\left( P_{tx} + G_{tx} + G_{rx} + 20 \log \left( \frac{\lambda}{4 \pi d} \right) \right) / 10}}{2 \sqrt{2} e \cdot B \cdot R_j^2 (I_0 + 2 I_{s0}) + 4k \cdot T \cdot B \cdot (R_x + R_a)} \right), \tag{4.36}
\]

\[\begin{align*}
\text{SNR [dB]} & \quad \text{(-26.3 dBm)} & \quad \text{(-21.3 dBm)} & \quad \text{(-16.3 dBm)} & \quad \text{(-11.3 dBm)} \\
P_b & \quad (P_{tx}) & \quad (P_{tx}) & \quad (P_{tx}) & \quad (P_{tx}) \\
\rho = 0.8 & \quad \rho = 0.6 & \quad \rho = 0.4 & \quad \rho = 0.2 & \quad \rho = 0 \\
\end{align*}\]

Fig. 4.18: Probability of bit error for noncoherent detection of circular-polarization-shift-keying modulated signals. Solid line: BER given by Eq. 3.13; Dashed lines: BER given by Eq. 4.45. On the horizontal axis the transmitted power \(P_{tx}\) is also indicated for the case that the BER curves are related to \(P_{tx}\) by means of Eq. 2.25 using the parameters given in Table 4.3.
where the values for $G_{rd}$ and $G_{tg}$ must be given in dB and $P_{tx}$ must also be expressed in dBm. Table 4.3 summarizes some of the parameters of the measurement setup which can be used in Eq. 4.36.

**TABLE 4.3**

<table>
<thead>
<tr>
<th>$\beta_v$</th>
<th>30 mV/μW</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P_{tx}$ (including conversion loss)</td>
<td>variable from -30 dBm to -16 dBm</td>
</tr>
<tr>
<td>$G_{rd}$ and $G_{tg}$</td>
<td>4.1 dB</td>
</tr>
<tr>
<td>$\lambda$ @ f = 2.44 GHz</td>
<td>0.123 m</td>
</tr>
<tr>
<td>$d$</td>
<td>0.5 m</td>
</tr>
<tr>
<td>$e$</td>
<td>$1.609 \times 10^{-19}$ As</td>
</tr>
<tr>
<td>$B$</td>
<td>44 kHz</td>
</tr>
<tr>
<td>$I_0$</td>
<td>0</td>
</tr>
<tr>
<td>$I_{s0}$</td>
<td>3 μA</td>
</tr>
<tr>
<td>$k$</td>
<td>$1.38 \times 10^{-23}$ Ws/K</td>
</tr>
<tr>
<td>$T$</td>
<td>293 K</td>
</tr>
<tr>
<td>$R_s$</td>
<td>25 Ω</td>
</tr>
<tr>
<td>$R_j$</td>
<td>9 kΩ</td>
</tr>
<tr>
<td>$R_d$</td>
<td>0</td>
</tr>
</tbody>
</table>

Fig. 4.18 shows the calculated BER curves for noncoherent detection of circular-polarization-shift-keying modulated signals. On the horizontal axis the transmitted power $P_{tx}$ is also indicated which is obtained by applying Eq. 2.25 on the SNR values as described above.

The measured BER values are listed in Table 4.4. For values of $P_{tx}$ larger than -10 dBm the BER test set was no more able to measure any bit errors. As mentioned above it should be noted that the calculated BER curve versus $P_{tx}$ in Fig. 4.18 (solid line) is based on the SNR of the diode and not on the SNR of the demodulator. Therefore, the calculated BER curve predicts a too optimistic BER performance. A comparison between measured and calculated BER results that the disagreement is in fact rather high.

**TABLE 4.4**

<table>
<thead>
<tr>
<th>$P_{tx}$ [dBm]</th>
<th>-15.8</th>
<th>-14.3</th>
<th>-12.8</th>
<th>-11.3</th>
</tr>
</thead>
<tbody>
<tr>
<td>measured BER</td>
<td>$1.04 \times 10^{-1}$</td>
<td>$6.25 \times 10^{-2}$</td>
<td>$4.27 \times 10^{-2}$</td>
<td>$2.37 \times 10^{-2}$</td>
</tr>
</tbody>
</table>
4.1.8 Voltage Doubler

Diode detectors may be combined in various ways to produce higher output voltages than would be produced by a single diode. For instance, two zero bias Schottky detector diodes may be combined to form a voltage doubler detector as it is shown in Fig. 4.1. The voltage doubler is a combination of a shunt diode and a detector. It works as follows: (1) The sine wave of the RF signal, which is symmetrical about zero volts, is raised by the shunt diode so that the minimum voltage is zero. (2) The input to the series detector diode is the input sine wave plus a DC component equal to the peak voltage. Thus, the detected voltage is the peak-to-peak voltage of the RF signal and therefore the peak amplitude detected by a single diode is doubled.

![Graph](image)

*Fig. 4.19: Voltage sensitivity $\beta_n$ versus RF input power $P_{in}$ of a single diode detector (HP HSMS 2855) and a voltage doubler detector (HP HSMS 2852). HSMS 2855: measured at a mid band match of $S_{11} = -7$ dB at 2.43 GHz; HSMS 2852: measured at a midband match of $S_{11} = -12$ dB at 2.41 GHz.*

Compared to a single diode detector, the voltage doubler shows the following advantages:

- For the RF signal, the two diodes of the voltage doubler are in shunt and thus the input impedance is halved. Matching networks for detectors using a voltage doubler are therefore easier to design. Smaller reactances are required and the matching networks becomes less dispersive. As a result the bandwidth is increased or, equivalently, less midband match must be traded for bandwidth.
Since the voltage doubler detector detects the peak-to-peak voltage of the incoming signal, a higher voltage sensitivity can be obtained than using a single diode detector. It can be shown by means of Eq. 4.26 that the voltage doubler detector shows also better values for the tangential signal sensitivity $TSS$ as the junction resistance $R_j$ is lowered and the voltage sensitivity $\beta_v$ is increased compared to a single diode detector.

Fig. 4.19 shows a plot of the voltage sensitivity versus RF input power of a single detector diode (HP HSMS 2855) and a detector circuit using a voltage doubler diode (HP HSMS 2852). Both circuits consist of low barrier Schottky detector circuits in a SOT-23/SOT-143 package. The layout of the voltage doubler detector is depicted in Fig. 4.15c. The single diode detector used for the measurements shows a midband match of $S_{11} = -7$ dB at a frequency of 2.43 GHz. As mentioned above, the voltage doubler detector is easier to match to 50 Ω. Indeed, $\beta_v$ could be measured at a midband frequency of 2.41 GHz where an insertion loss of -12 dB is obtained. In Fig. 4.19, the different operation regions of the detectors can be distinguished very well. For an input power below -30 dBm, the detectors are operated in the quadratic region since the voltage sensitivity is constant. In the range of $P_{in}$ from -30 dBm to -10 dBm, the voltage sensitivity is decaying linearly and the detectors are operated in the linear region. Above an input power of -10 dBm, the voltage sensitivity begins to flatten and thus the diodes are in the saturation region.

In the quadratic region where the performance of the detector is most important for the operation of the transponder, the voltage doubler shows a much better performance than the single diode detector. Additionally, the layout of a detector using a voltage doubler is smaller since the lengths of the distributed elements in the matching network are shorter and most important, no separate DC current return is necessary. Therefore, in an active tagging system voltage doubler detectors may be preferred to increase transmission distance as long as the transmission distance is limited by the sensitivity of the transponder at the downlink communication and not by the sensitivity of the reader in the uplink communication as it is illustrated by Fig. 2.10 in Chapter 2.4.2.
4.2 Antennas

Due to the applied circular-polarization-shift-keying modulation scheme in the downlink communication and the use of an active modulator in the uplink communication scheme, the antenna design for the reader and the tag must be performed very carefully since the communication between reader and tag is strongly affected by the antenna performance (e.g. polarization isolation). The tag and reader antennas have to fulfill partly different requirements which include the following items:

Requirements for the tag antenna:

- Since the antenna is the largest component on the transponder (in particular at 2.4 GHz), it must be combined with the RF front end electronics such that the transponder size gets as small as possible.

- The antenna plays an important role in the demodulation of the circular-polarization-shift-keying modulated signals because it separates the left-hand and right-hand circularly polarized signals and feeds them to the RF detectors (see Chapter 4.1.6). Thus, a high isolation between LHCP and RHCP waves is demanded. As shown below, this corresponds also to a low axial ratio which is a measure for the performance of the antenna to transmit or receive circularly polarized waves. A low axial ratio is desired as well in order that the reader and tag antenna need not to be aligned. Therefore, at most one antenna in a tagging system (mostly the reader antenna) should be circularly polarized.

- In the uplink communication, an active modulator consisting of a preamplifier succeeded by a mixer is used (the mixer's IF port is driven by the data signal from the microcontroller). In order to prevent an instable operation of the active modulator due to its conversion gain, the received and transmitted signals show orthogonal polarization sense. As will be shown below, not only a high polarization isolation is required within the ISM band but also for a frequency range outside of the band. This is due to the frequency response of the amplifier which typically shows gain outside the 4% bandwidth of the ISM band. An ideal tag antenna should have an isolation bandwidth much larger than the bandwidth of the ISM band.

Requirements for the reader antenna:

- Similar to the tag antenna, the antenna for the reader should also be circularly polarized with switchable polarization sense. The gain and impedance bandwidth requirements are the same as those for the tag antenna. But in contrast to the tag antenna, the requirements for the polarization isolation outside the ISM band are more relaxed.

- In order to increase the receiver sensitivity of the reader in the uplink communication, a high gain antenna may be preferred. This could be achieved by using an antenna array. However, a high gain antenna shows a reduced beam width and thus less tags can be scanned by the beam of the reader antenna. Depending on the application, this disadvantage might not be very bothering. Since the available space is not a limiting factor at the reader, the antenna array at the reader could be extended with beam steering capabilities in order to steer the antenna beam when interrogating the transponders.
It has turned out that the requirements listed above are optimally fulfilled by using aperture-coupled patch antennas. Therefore, in this chapter, first the setup of aperture-coupled patch antennas is presented and the advantages of using them for tagging applications are discussed. After summarizing the main antenna specifications for the active tagging system, a design example of a linearly polarized aperture-coupled antenna is shown. The antennas were designed by using 2.5 D EM solvers\(^1\) which are based on the method of moments. Afterwards the design of some circularly polarized aperture coupled patch antennas with switchable polarization sense for the transponder is presented together with the parameter design rules for these antennas. At the end of the chapter, an antenna array for the reader with beam steering capabilities is shown which is based on a transponder antenna discussed above.

### 4.2.1 Aperture-Coupled Patch Antennas for Tagging Applications

Fig. 4.20 shows the structure of an aperture-coupled patch antenna with switchable polarization sense. It consists of three layers, namely the patch, slot and feed layer. Aperture-coupled patch antennas are especially suited to meet the requirements of a transponder antenna because the multilayer structure allows to optimize the radiation characteristics (patch layer substrate) and the size of the feed network (feed layer substrate) indepen-

---

1. Momentum, HP Series IV, Ensemble, Ansoft and PCAAD, D.M. Pozar Antenna Design Associates

---

Fig. 4.20: Structure of an aperture-coupled patch antenna with switchable polarization sense consisting of patch, slot and feed layer.
dently. The radiating patch is located on a substrate with relative permittivity \( \varepsilon_{rf} \), and the feed network is etched onto the bottom side of a substrate with relative permittivity \( \varepsilon_{r2} \). These substrates are separated by a common ground plane that features an electrically small slot aperture for efficient coupling of power to the patch.

The slot aperture shown in Fig. 4.20 comprises two orthogonal slots that are fed by two open-circuited stub lines of a 3 dB/90° hybrid. This branch-line coupler produces fields of equal amplitude and 90° out of phase at its center frequency. Each input terminal of the hybrid provides an opposite circular polarization sense. The open-circuited stub length of the microstrip line extending beyond the slot aperture can be used for impedance matching and bandwidth enhancement. This stub, together with the aperture length, controls the input impedance over a wide range of values. Therefore, the lack of an additional matching network and the ability to choose a thin feed layer substrate help to minimize the size of the feed network. In addition, a thick patch layer substrate with low relative permittivity can be used to increase the impedance bandwidth.

Aperture-coupled patch antennas feature also a high front-to-back ratio due to the intermediate ground plane. This characteristic produces good shielding of the RF front end electronics (RF detectors, Rx/Tx switches and active modulator) and the signal processing electronics at the rear side of the tag antenna. Although aperture-coupled patch antennas show increased manufacturing costs compared to classical patch antennas, their advantages in terms of radiation performance and antenna dimensions makes it a viable alternative for microwave tag antennas.

4.2.2 Definition of Antenna Parameters and Antenna Requirements for the Active Tagging System

This chapter presents the definition of the main antenna parameters used in the design and discusses the antenna requirements for the active tagging system.

The input impedance characteristic of a tuned antenna behaves as a simple tuned circuit where the relative bandwidth \( B \) (related to a change of 3 dB from the magnitude of the impedance at resonance frequency) is about \( 1/Q \) with \( Q \) being the total quality factor. For microstrip antennas the bandwidth is determined by a permissible value of standing wave ratio (SWR), typically 1:2, and is calculated from

\[
B = \frac{\text{SWR} - 1}{Q \cdot \sqrt{\text{SWR}}} \tag{4.37}
\]

The relative bandwidth \( B \) can be increased by reducing \( Q \). This can be obtained for an aperture-coupled patch antenna by using a thicker and/or lower-permittivity substrate for the patch dielectric [54]. Since the beamwidths, sidelobe level, antenna gain and polarization properties vary with frequency, the bandwidth can also be defined by these parameters.
For the presented active microwave tagging system, the polarization properties of the antenna are most important for the system performance. The axial ratio and the cross-polarization isolation are two parameters which describe the polarization ellipticity of the antenna. As shown by Eq. 3.1 and Eq. 3.3, circularly polarized waves may be decomposed into two orthogonal fields $E_x$ and $E_y$. Then, the left-hand and right-hand circularly polarized waves can be written as [48]:

$$|E_{RHC}| = \frac{|E_x + jE_y|}{\sqrt{2}}$$  \hspace{1cm} (4.38)

$$|E_{LHC}| = \frac{|E_x - jE_y|}{\sqrt{2}}$$  \hspace{1cm} (4.39)

and the axial ratio is given by

$$AR = \frac{|E_{RHC}| + |E_{LHC}|}{|E_{RHC}| - |E_{LHC}|} = \frac{a}{b}.$$  \hspace{1cm} (4.40)

where $a$ and $b$ refer to the major and minor axis of the polarization ellipse depicted in Fig. 4.21. The ellipticity or axial ratio is frequently expressed in dB by

$$AR[dB] = 20 \cdot \log \left( \frac{a}{b} \right).$$  \hspace{1cm} (4.41)

The circularity can also be determined as the ratio of power in the desired sense to that in the opposite sense. This cancellation ratio is termed cross-polarization (CP) isolation

$$CP[dB] = 10 \cdot \log \left( \frac{\text{power in desired polarization sense}}{\text{power in opposite polarization sense}} \right).$$  \hspace{1cm} (4.42)
There is a relationship between the cross-polarization isolation and the axial ratio. When $CP$ and $AR$ are expressed in linear values, this relationship is given by

$$CP = \frac{AR + 1}{AR - 1}. \quad (4.43)$$

A plot of Eq. 4.43 is shown in Fig. 4.22 where the ellipticity and the corresponding cross-polarization isolation are expressed in dB.

In practice, there are mainly two different patch antenna configurations for generating circularly polarized waves, namely (1) singly-fed circularly polarized antennas and (2) dual-fed circularly polarized antennas. The latter antenna type is shown in Fig. 4.20 and Fig. 4.23d. It consists of a dual-linearly-polarized patch antenna whose antenna terminals are fed with equal amplitude and 90° out of phase using an external polarizer. For the polarizer, either a branch-line coupler or a power splitter with additional 90° delay line can be used. The branch-line coupler polarizer allows additionally to switch the sense of circular polarization.

Singly-fed circularly polarized antennas are mainly used for applications where the antenna must be small as no external polarizer is needed. The operational principle of such antennas is based on the generation of two orthogonal modes by the effect of a perturbation segment such as a slot or other truncated segment. Fig. 4.23a shows two typical examples of singly-fed circularly polarized antennas. In Fig. 4.23b and Fig. 4.23c, a typical
amplitude and phase diagram is depicted. The radiated fields excited by the orthogonal modes are circularly polarized. The polarization sense is dependent on the placement of the orthogonal segment relative to the feedline. As shown in the amplitude and phase diagram, modes 1 and 2 are excited in equal amplitude and 90° out of phase only in a small frequency range around the centre frequency. Singly-fed circularly polarized antennas are therefore very narrow band. Nevertheless, this antenna type is frequently used as transponder antenna due to its small dimensions.

Another interesting antenna parameter for tagging applications is the beamwidth. Frequently, the half-power beamwidths in the H and E planes are used for describing the characteristic of the major lobe of the radiation pattern. The half power beamwidth is equal to the angular width between directions where the gain decreases by 3 dB, or the radiated fields reduces to $1/\sqrt{2}$ of the maximal value. As a rule of thumb, the beamwidth in radians can be estimated by [55]

$$\theta_B \approx \frac{\lambda_D}{D}.$$  (4.44)
where \( D \) denotes the antenna diameter. The beamwidth of a microstrip antenna can be increased by selecting a substrate having a higher relative permittivity as this will decrease the antenna dimensions. As the beamwidth increases, the antenna gain and consequently the directivity decreases.

<table>
<thead>
<tr>
<th>Polarization</th>
<th>LHCP and RHCP</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cross-Polarization Isolation (Axial Ratio)</td>
<td>&gt; 20 dB (&lt; 1.8 dB)</td>
</tr>
<tr>
<td>Antenna Gain</td>
<td>&gt; 4 dBi</td>
</tr>
<tr>
<td>Impedance Bandwidth (VSWR=2:1)</td>
<td>&gt; 4%</td>
</tr>
</tbody>
</table>

Based on the above considerations, the antenna requirements listed in Table 4.5 can be specified. Due to the circular-polarization-shift-keying modulation scheme in the downlink communication and the required cross-polarization isolation in the uplink communication, the transponder antenna must be left-hand and right-hand circularly polarized. A cross-polarization isolation better than 20 dB (which corresponds to an axial ratio lower than 1.8 dB) is required. In the uplink communication, this specification allows to employ an RF amplifier in the active modulator with a gain of up to 20 dB. This value is a reasonable trade-off between conversion gain of the active modulator and power consumption.

The effects of the cross-polarization isolation on the downlink communication may be investigated by means of the probability of error for envelope detection of correlated binary signals. The error probability for the noncoherent detection of binary, equal-energy correlated signals is given by [24]

\[
P_h = Q_1(a, b) - \frac{1}{2} e^{-\frac{a^2 + b^2}{2}} I_0(ab),
\]

where

\[
a = \sqrt{\frac{\gamma_b}{2}} (1 - \sqrt{1 - |\rho|^2})
\]

\[
b = \sqrt{\frac{\gamma_b}{2}} (1 + \sqrt{1 - |\rho|^2}).
\]

\( Q_1(a, b) \) is the \( Q \) function defined as

\[
Q_1(a, b) = e^{-\frac{a^2 + b^2}{2}} \sum_{k=0}^{\infty} \left(\frac{a}{b}\right)^k I_k(ab) \quad b > a > 0
\]

and \( I_0(ab) \) is the modified Bessel function of order zero. The cross-correlation coefficient \( \rho \) is defined in Eq. 3.8. For orthogonal signals \( \rho \) is equal to zero and Eq. 4.45 reduces to Eq. 3.13 (cf. Fig. 4.18). The cross-polarization isolation of the transponder antenna is a...
measure for the separation of LHCP and RHCP waves. Thus, $\rho$ is a function of the cross-polarization isolation. This qualitative analysis shows that $\rho$ increases as the cross-polarization isolation decreases and hence the BER performance is decreased as well. However, an analytical expression for $\rho$ is difficult to derive since the nonideality of the circular-polarization-shift-keying modulator (axial ratio of the reader antenna) and the alignment between reader and transponder antenna must also be included in the analysis.

In Table 4.5, the antenna gain is specified to be larger than 4 dBi. This value has been chosen as a compromise between high directivity and wide beam width. Another specification is the impedance bandwidth which should be larger than 4% for a VSWR of 2 ($S_{11}=10dB$).

### 4.2.3 Analysis of a Linearly-polarized Aperture-Coupled Patch Antenna

In this chapter, some design considerations for a linearly-polarized aperture-coupled patch antenna will be discussed. Analytical expressions have been developed for the analysis of the coupling mechanism in a linearly-polarized aperture-coupled patch antenna [47], [48], [49]. Results of this analysis will be useful for the design of the circularly-polarized transponder and reader antennas.

Fig. 4.24 shows the geometry of a linearly-polarized aperture-coupled patch antenna and its coordinate system with the origin in the center of the slot. The slot aperture is located at $x=0$ in the middle under the patch. A simple theory of the aperture-coupling mechanism can be given when considering the patch to be resonating in its dominant $TM_{100}$ mode. The fields inside the cavity formed by the patch and the magnetic walls (magnetic current proportional to the tangential electric field) around the periphery of the patch are given by [49]

$$E_z(x) = \frac{k_0^2}{j \omega \varepsilon_0} \cdot \cos \left( \frac{\pi}{l_p} \cdot x \right)$$

$$H_y(x) = \frac{\pi}{l_p} \cdot \sin \left( \frac{\pi}{l_p} \cdot x \right),$$

where $l_p$ is the resonant dimension of the antenna (along the x-axis in Fig. 4.24), and $k_0=2\pi/\lambda_0$. Since the field components are not dependent on the y coordinate, the slot aperture may be shifted under patch in y-direction with little or no effect on the coupling to the $TM_{100}$ mode.

A displacement of the slot aperture in x-direction has been investigated in [49]. For an infinitely long microstrip feedline, it is shown that the electric and magnetic coupling coefficients $C_p$ and $C_M$, respectively, between the cavity fields and the feedline obey the following relationship

$$C_p \sim \sin \left( \frac{\pi}{l_p} \cdot x \right)$$

94 Antennas
The electric coupling is maximum at the ends of the patch ($x=-l_p/2$ or $x=l_p/2$), and minimum at the centre of the patch ($x=0$). The magnetic coupling has the opposite positions for its maxima and minima. Thus, by moving the aperture along the x-axis, the dominant coupling mechanism can be shifted from a pure electric to a pure magnetic coupling effect. Moreover, in reference [49] is shown that the maximum magnetic coupling $C_{M,\text{max}}$ is many times greater than the maximum electric coupling $C_{p,\text{max}}$. Hence, the magnetic coupling mechanism is the preferred one, and the slot aperture should be located at $x=0$ in the centre of the patch to maximize this effect. However, when circular polarization is
required, the slot aperture must be shifted along the resonant dimension of the patch in order to allocate two feed points.

The equivalent circuit of the linearly-polarized antenna is shown in Fig. 4.25. The aperture-coupled patch antenna can be tuned with an open-circuited stub line which is approximately λ/4 long. If the stub length is \( L_s \), the transformation of the equivalent series resistance is given by

\[
Z_{in} = Z - jZ_c \cdot \text{ctg}(\beta_m \cdot L_s),
\]

where \( Z \) denotes the equivalent series impedance of the slot and patch aperture, \( Z_c \) and \( \beta_m \) are the characteristic impedance and the propagation constant of the microstrip line.

4.2.4 Circularly Polarized Antennas with Switchable Polarization Sense

In this chapter, the design of four different circularly polarized, aperture-coupled patch antennas with switchable polarization sense are presented [46]. Fig. 4.26 and Fig. 4.27 show a top view and a detailed view on the characteristics of the antennas’ patch, slot and feedlayer. The antennas’ main differences are in the feed network, slot aperture shape and radiating patch. Three antennas are dual polarized and use a 3 dB hybrid as polarizer in the feed network. The slot aperture of antenna type A consists of two orthogonal slots that are placed at some offset distance below the edges of the patch. Two pairs of orthogonal slots are used for antenna type B. A bent λ/4 microstrip line is employed in the feed network of each pair to provide the required 90° phase difference between the orthogonal slots. Equal to a polarizer using a branch-line coupler, the polarization sense of antenna type B can be changed by switching between the two antenna ports. Antennas C and D show annular aperture shapes. The annular square slot aperture of antenna type C is fed at two adjacent corners by the open-circuited lines of a branch-line coupler. Antenna type D is similar to antenna type C except that the radiating patch and slot aperture are circular. The stub lines feed the slot aperture as well in orthogonal directions.

The four aperture-coupled patch antennas employ a 125 mil (3.18 mm) thick substrate (POLYFLON POLYGUIDE) with a relative permittivity of \( \varepsilon_r = 2.33 \) and a dissipation factor of 0.0011 specified at 3 GHz for the patch layer and a 20 mil (0.508mm) thick substrate (R/T Duroid 5880) with the parameters \( \varepsilon_r = 2.22 \) and \( \tan(\delta) = 0.0009 \) at 10 GHz for...
the feed layer. A planar electromagnetic (EM) solver\footnote{Momentum, HP Series IV} using the method of moments was employed for simulations. In order to produce a good understanding of the mode of action of the relevant antenna design parameters, the focus is mainly on antenna type A. This antenna type can be considered as dual-polarized antenna that is fed by the decoupled output ports of a branch-line coupler. Since the design of a 3 dB hybrid is well known, only the design steps of the dual-polarized antenna are illustrated.

The dual-polarized antenna consists of two linear polarized elements, one of which is shown in Fig. 4.28 along with its corresponding lumped-element equivalent circuit. Starting at the left end of the equivalent circuit, one approaches the slot aperture with the characteristic impedance of the feed network $Z_0$. The slot is modeled by a series inductance.

\begin{figure}[h]
\centering
\includegraphics[width=0.8\textwidth]{antennas.png}
\caption{Top view of the different antennas' patch, slot and feed layers.}
\end{figure}
**Fig 4.27: The four circularly polarized aperture-coupled antennas' characteristics.**

$L_{\text{slot}}$ since the coupling between the slot and the feed layer mainly occurs by the magnetic field component of the quasi-TEM mode propagating on the microstrip feed line. The radiating patch is modeled by a parallel resonant circuit where the resistance models the radiation losses. The impedance of the radiating patch shows no reactive part at resonance frequency. However, the series inductance of the slot contributes a reactive part to the input impedance that must be compensated by the capacitance of the open-circuited stub line in the feed network.

Based on the presented lumped element equivalent circuit, three design parameters are used. First, the length of the open-circuited stub line can be varied to match the slot aperture to the impedance of the microstrip feed line. The size of the slot aperture and the offset distance from the slot aperture to the patch edge are two parameters that determine the amount of coupling.

### 4.2.5 Antenna Parameter Design Rules

Initial values for the dimensions of the radiating patch can be determined using equations of the modal expansion model or the open-circuited radial resonator model [16]. Calculations for the patch layer substrate at the center frequency of the ISM band at 2.44 GHz produce a value of 38 mm and 18.5 mm for the antenna’s patch length and radius, respectively. A feed line width of 1.56 mm is calculated to connect the dual-polarized antenna to the 50 Ω microstrip lines of the branch-line coupler. Fig. 4.28a shows the initial antenna dimensions.
In Fig. 4.29a, the behavior of the antenna’s input impedance with changes in stub length of the feed network is depicted. As the stub length increases, the impedance locus rotates in a clockwise direction along a constant resistance circle at a fixed frequency. The input impedance curves on the Smith chart refer to different values of the stub length, which differ up to ±15 percent from the nominal value of the equivalent circuit. The open-circuited stub length can be adjusted to obtain the desired reactance. For a 50 Ω impedance match, the stub length is adjusted until the input impedance at the design frequency is purely real.

While the stub length rotates the impedance locus, the aperture length controls the amount of coupling, as shown in Fig. 4.29b. By increasing the aperture length, the input impedance curve moves toward the right in the direction of resistance values higher than the characteristic impedance. The aperture length can be adjusted to obtain the desired resistive part of the impedance (for example, 50 Ω). Curves of different values of aperture length are depicted on the Smith chart.

A similar behavior of the input impedance occurs for slot width changes. For a narrow slot, the patch is under coupled and the resonance resistance is less than the characteristic impedance of the feed line. An increase of the slot aperture also causes a reduction in the
Fig. 4.29: a) Input impedance characteristics with changes in stub length; b) input impedance characteristics with variations of slot aperture length (Scattering parameter plot from $f_{\text{min}}=2$ GHz to $f_{\text{max}}=3$ GHz).

front-to-back ratio and thus reduces the patch efficiency. A small slot width of 0.6 mm shows most effective coupling for the configuration. The slot aperture is operated below its resonance frequency.

The previous parameter study was focused on one of the linear-polarized elements. However, in the design of the dual-polarized antenna, it is important to obtain a high isolation between the two feed lines. If the requirement of high isolation is not met, the coupler connection to the two feed lines results in a poor antenna axial ratio. The limited bandwidth of the coupler causes an additional deterioration of the isolation. Therefore, investigations of appropriate values for the offset distance of the slot aperture from the patch edge also must include the isolation between the orthogonal slots. As the isolation becomes higher with longer distance between the nearest corners of the slot apertures, a trade-off between aperture length and isolation must be made while increasing the offset distance of the slots. In
the present case, these parameters are adjusted to obtain an isolation greater than 15 dB within the entire 2.4 GHz ISM band. For transmission of circular-polarization-shift-keying modulated data, a high isolation is of prime importance as the ellipticity characteristic follows the isolation characteristic.

**TABLE 4.6**

<table>
<thead>
<tr>
<th>Antenna type</th>
<th>A</th>
<th>B</th>
<th>C</th>
<th>D</th>
</tr>
</thead>
<tbody>
<tr>
<td>Slot width [mm]</td>
<td>0.6</td>
<td>0.6</td>
<td>0.6</td>
<td>0.6</td>
</tr>
<tr>
<td>Slot length [mm]</td>
<td>26</td>
<td>18</td>
<td>e = 14</td>
<td>r = 9</td>
</tr>
<tr>
<td>Offset of slot from patch midpoint [mm]</td>
<td>14.7</td>
<td>14.7</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Stub length [mm]</td>
<td>14.0</td>
<td>22.3</td>
<td>11.0</td>
<td>11.2</td>
</tr>
<tr>
<td>Feed line width [mm]</td>
<td>1.56</td>
<td>1.56</td>
<td>1.56</td>
<td>1.56</td>
</tr>
<tr>
<td>Patch width [mm]</td>
<td>37</td>
<td>37</td>
<td>35</td>
<td>r = 18.7</td>
</tr>
</tbody>
</table>

\(a \ e = \text{slot length}, \ r = \text{radius}\)

Although the presented design parameters are used mainly for the design of a type A antenna, some results for the design of the other antenna types also can be derived. The final dimensions of the presented antenna types are listed in Table 4.6. Antenna types C and D with annular slot apertures show the same slot width and radial patch dimensions as antenna type A. Therefore, only the placement of the circular and square slot aperture relative to the patch edges as well as the length of the open-circuited stub of the microstrip feed line must be subsequently determined. For an optimal coupling of the configuration, it has been determined that the distance from the center of the patch to the annular slot is slightly less than half the radius of the patch for antenna type D or approximately a quarter of the patch edge for antenna type C. The reactive part of the input impedance owing to the inductive slot aperture impedance can be tuned out as shown previously.

Antenna type B requires a different feed network. Two bent \(\frac{\lambda}{4}\) lines are used as polarizers instead of a branch-line coupler. The \(\frac{\lambda}{4}\) lines feed the orthogonal slot apertures 90° out of phase. Again, the open-circuited stub serves to match the first slot aperture. However, the reactance of the second slot must be compensated by a matching network that consists of a simple radial tuning stub.

### 4.2.6 Measured Radiation Characteristics

Small-signal scattering parameters have been measured within the frequency range of operation. A comparison of the measured S-parameters with the EM simulations of the antennas (including the polarizers) showed deviations in the resonance frequency of less than 1.5 percent. This discrepancy may be caused by different substrate losses. For simulations, only extrapolated values of the dissipation factor could be used. All antennas show an impedance bandwidth larger than four percent and thus, their impedance bandwidth exceeds the requirement for the relative bandwidth of the ISM band. The bandwidth specifications could be fulfilled because a thick substrate with low relative permittivity has been used for the patch layer.
Far-field measurements in an anechoic chamber have been performed to determine the antenna gain and radiation patterns. Fig. 4.30 shows the measured radiation patterns of antenna type A. Helix antennas transmitting LHCP and RHCP waves were used as reference antennas to determine an antenna gain of 7 dBi. There are only small differences in gain among the four antenna types. The same measurement setup has been used to determine a front-to-back ratio of approximately 16 dB.

Measurements of the polarization ellipticity were performed in the near and far field. In the far field, an automated electro-mechanical positioner was used to measure the axial ratio and polarizer isolation. The antenna under test (AUT) is mounted on the positioner such that it rotates around its boresight axis. The received power levels at the left-hand and right-hand ports of the AUT are recorded while transmitting LHCP and RHCP waves from a purely circular reference source. If the reference polarization is specified by LHCP
waves, power levels measured at the left-hand port are denoted as copolarized and power levels measured at the right-hand port are denoted as cross-polarized.

**TABLE 4.7**

<table>
<thead>
<tr>
<th>Antenna type</th>
<th>A</th>
<th>B</th>
<th>C</th>
<th>D</th>
</tr>
</thead>
<tbody>
<tr>
<td>Impedance bandwidth [%] (SWR =2)</td>
<td>8</td>
<td>7</td>
<td>6</td>
<td>6</td>
</tr>
<tr>
<td>Gain [dB]</td>
<td>7.1</td>
<td>6.8</td>
<td>7.4</td>
<td>6.7</td>
</tr>
<tr>
<td>Axial ratio [dB]</td>
<td>&lt;1.0</td>
<td>&lt;2.5</td>
<td>&lt;2.0</td>
<td>&lt;2.5</td>
</tr>
<tr>
<td>Cross-polarization isolation [dB]</td>
<td>&gt;13</td>
<td>&gt;8</td>
<td>&gt;12</td>
<td>&gt;10</td>
</tr>
</tbody>
</table>

*a. measured at 2.42 to 2.47 GHz*

Fig. 4.31 shows the resulting plot of the measured power levels at the two antenna ports of antenna type A. The axial ratio is determined as maximal aberration of the copolarized curve from a perfect circular curve. The polarization isolation is determined by the difference between the co- and cross-polarized curves. Although these measurements may differ from methods commonly used to determine the circular polarization (for example, the spinning dipole method [50], [51]), they are well suited for measurements of the circular polarization relative to a well characterized reference antenna such as a helix antenna.

Near-field measurements have been performed by means of an automated E-field scanner [52], [53] for visualizing the circular polarization. The automated E-field scanner measures the magnitude of two orthogonal E-field components and displays them as the minor and major axes of an ellipse. Fig. 4.32 shows the visualization of the circular polarization of antenna type A. The automated E-field scanner helps to visualize misalignments.
between patch, slot and feed layer, which produce an asymmetric distribution of the measured polarization ellipses across the radiating patch.

Based on the far-field measurement of the axial ratio and cross-polarization isolation, the values listed in Table 4.7 were determined. Within the measured bandwidth all antennas showed an axial ratio better than 3 dB. The best value of the axial ratio and cross-polarization was obtained by antenna A. As antenna types C and D use only one annular slot instead of two separated orthogonal slots, the isolation between their feed lines is reduced, which results also in a decrease in cross-polarization isolation. In addition, visualizations of the circular polarization with the automated E-field scanner showed that these antenna types are much more sensitive to misalignments between the different layers than in the case of antenna type A, possibly causing an additional deterioration of the axial ratio.

In order to minimize the size of the feed network, the $\lambda/4$ lines of antenna type B are connected directly between orthogonal slots instead of feeding the slots in an offset line configuration by using a power divider. Proper amplitude balance between the orthogonal slots is thus difficult to achieve, which leads to a degraded axial ratio compared to antenna type A. Due to the $\lambda/4$ lines, the ellipticity bandwidth is also slightly smaller compared to antenna types using a branch-line coupler as polarizer.

---

Fig. 4.32: Visualization (magnitude of orthogonal E-field vectors) of circular polarization for antenna type A measured with an automated E-field scanner.
4.2.7 Tag Antenna

Antenna type A shown in Fig. 4.26 was used for the setup of the tag antenna. In contrast to the antenna types presented in Chapter 4.2.4, a high permittivity substrate (Duroid 6010 or Rogers 3010, $\varepsilon_r = 10.2$, $h = 0.635 \text{ mm} = 25 \text{ mil}$) was employed for the feed layer substrate in order to reduce the size of the distributed elements for the RF front end electronics. The patch layer substrate (POLYFLON POLYGUIDE, $\varepsilon_r = 2.32$, $h = 3.18 \text{ mm} = 125 \text{ mil}$) is the same as the one used for the antennas presented in the previous chapters.

Fig. 4.33 shows a photography of the patch, slot and feed layer of the transponder’s aperture-coupled patch antenna. The slot and feed layer are etched on the same substrate. The total dimensions of the transponder including the signal processing PCB and the battery are $120 \text{ mm} \times 76 \text{ mm} \times 25 \text{ mm}$. The patch and feed substrate are hold together with plastic screws ($M3$-size).

![Fig. 4.33: Photography of the three layers of the transponder's aperture-coupled patch antenna.](image)

In this chapter, the simulation and measurement results of the tag antenna will be discussed. In Table 4.8 the antenna dimensions according to Fig. 4.28a are listed. The ground plane size of the test antennas whose measurements are shown below is $76 \text{ mm} \times 80 \text{ mm}$. For the presented antenna simulations, the software tools Momentum, HP Series IV and Ensemble, Ansoft were used. Both EM field simulators are 2.5 D solvers based on the Method of Moments (MoM).

<table>
<thead>
<tr>
<th>$w_{\text{patch}}$</th>
<th>$w_{\text{slot}}$</th>
<th>$l_{\text{slot}}$</th>
<th>$l_{\text{offset}}$</th>
<th>$w_{\text{feed}}$</th>
<th>$l_{\text{stub}}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>36.6 mm</td>
<td>1.3 mm</td>
<td>18 mm</td>
<td>6.3 mm</td>
<td>0.574 mm</td>
<td>9 mm</td>
</tr>
</tbody>
</table>

TABLE 4.8
The software package HP Series IV was used for simulating small-signal scattering parameters at the antenna ports in order to determine the impedance matching and the port coupling. The S-parameter data of the antenna was imported into the circuit simulator of HP Series IV for performing system simulations at the design of the active modulator (see Chapter 4.3.2).

The simulations of the near and far fields were performed by Ensemble because it allows to determine more antenna parameters than the employed version of Momentum. Namely, 2 D and 3 D plots of the co-polarization and cross-polarization radiation patterns for linear and circular polarizations as well as the simulation of the antenna gain and the axial ratio are feasible by the Ensemble’s EM postprocessing simulator.

Fig. 4.34 shows some simulation results obtained by Ensemble. In Fig. 4.34a and Fig. 4.34b, the antenna gain for right-hand and left-hand circular polarization at f=2.44 GHz is depicted. Both gain characteristics are plotted versus θ at φ = 0 where θ is the angle in the
Fig. 4.35: a) simulated and measured antenna gain; b) simulated cross-polarization isolation (difference between $G_{RHCP}$ and $G_{LHCP}$) and front-to-back ratio versus $\theta$ at $\varphi = \theta$ (simulation results from Ensemble).

The simulated antenna gain shows a maximum value of 6.6 dBi at $f=2.44$ GHz in the middle of the ISM band. The difference between the co-polarization and the cross-polarization in bore-sight direction is a measure for the suppression of the opposite polarization sense. This cross-polarization isolation was simulated from $f=2.3$ GHz to $f=2.6$ GHz and is plotted in Fig. 4.35b. As an alternative, the axial ratio could be simulated as shown in Fig. 4.34c and by the aid of the relationship depicted in Fig. 4.22 the cross-polarization isolation could be determined. By the same simulations, the front-to-back ratio was determined which is depicted as well in Fig. 4.35b. Compared to the peak value of the simulated gain

x-z plane and $\varphi$ is the angle in the y-z plane of the coordinate system shown in Fig. 4.24.
characteristic, the maximum values for the simulated cross-polarization isolation and front-to-back ratio are shifted towards the upper edge of the 2.4 GHz ISM band. The peak values for the simulated cross-polarization isolation and front-to-back ratio are 29 dB and 38 dB, respectively. The simulation results were verified in an anechoic chamber by means of two identical transponder antennas separated by 2.8 m and used as transmitting
Fig. 4.37: a) measured characteristic of the antenna gain and the cross-polarization isolation (measurements performed in an anechoic chamber with a distance of 2.8 m between the transmitting and receiving antenna); b) cross-polarization isolation: difference between the curves of the previous plot; c) measured and simulated transmission scattering parameters between the two antenna ports (simulations performed by Momentum, HP Series IV); d) antenna meshing in the simulation (top view).

and receiving antennas. Some measurement results are depicted in Fig. 4.35a and Fig. 4.37. As shown in Fig. 4.35a and Fig. 4.37c a very good agreement between measured and simulated gain characteristic (simulated by Ensemble) and $S_{21}$ characteristic (simulated by Momentum) was obtained. Measurement results showed also that the peak value of the cross-polarization of the transponder is located exactly in middle of the ISM band. The measured cross-polarization isolation characteristic is depicted in Fig. 4.37b. Across the 2.4 GHz ISM band, it exceeds 15 dB and it even outperforms the simulated CP isolation by a maximum value of 37 dB at $f=2.45$ GHz.

In Fig. 4.34d the simulated radiation pattern of the $E_\theta$ and $E_\phi$ fields are depicted. A direct verification of these radiated fields was not possible by the available measurement equipment in the anechoic chamber. However, the radiation patterns of the total radiated fields were measured by means of an automated electromechanical positioner. The diagrams of...
the radiation patterns measured at five frequencies are depicted in Fig. 4.36. In each diagram the co-polarized and cross-polarized radiation pattern is shown. Again, the cross-polarization isolation and the front-to-back ratio can be determined. In particular, the radiation pattern belonging to f=2.45 GHz illustrates the very good polarization and gain performance of the transponder antenna.

4.2.8 Reader Antenna

For the reader, an aperture-coupled patch antenna designed by the project partner¹ at the ETH Zurich was used. The top view of the reader antenna is shown in Fig. 4.38. This antenna is employed in the prototype tagging system (see Fig. 3.12). Due to its symmetrical geometry, it shows very good values for the cross-polarization isolation. This reader antenna is optimized for a broadband operation (CP isolation better than 20 dB across the whole ISM band). A summary of the antenna’s electrical specifications is published in [60].

¹ The reader antenna was designed by Matthias Fries, ETH Zürich, Institut für Feldtheorie und Höchstfrequenztechnik, Fachgruppe Elektromagnetische Felder.
4.2.9 Phased Array Antenna for the Reader

A phased array antenna enables the reader to steer its antenna beam in order to scan for transponders being within its transmission range. Due to the polarization requirements

a)

![Diagram of a phased array antenna](image)

Fig. 4.39: a) Layout of the phased array antenna (dimensions: 155 mm x 135 mm); b) Measurement setup in the anechoic chamber; c) Rear side of the antenna array with the control unit PCB (μC and D/A converters).
(circular polarization with switchable polarization sense), the feed network of a phased array antenna becomes rather complex and the gain of antenna directivity might easily be compensated by the additional losses in the transmission paths. Therefore, only a two-element array is feasible with reasonable effort. The antenna beam of a two-element array can be controlled only in one dimension. For a two-element array the pattern multiplication theorem is given by [18]

\[ E_f(\theta, \phi) = E(\theta, \phi) \cdot AF = E(\theta, \phi) \cdot 2 \cos \left( \frac{1}{2} (k \cdot d \cdot \cos \theta + \beta) \right) \, , \]  

\[ (4.54) \]

---

---

**Fig. 4.40:** a) \( S_{21} \) at \( f=2.44 \text{ GHz} \) for a tuning voltage range from 0V to 10V; b) magnitude and phase of \( S_{21} \) for the same tuning range and RF frequency; c) equivalent circuit of the reflection type phase shifter (the cascade of two loads consisting of a \( \lambda/4 \) transmission line with short-circuited varactor diode enhances the operation region of the phaseshifter).
where $E(\theta, \phi)$ is the element pattern and $AF$ is the two-element array factor for a reference point assumed in the middle between the two antenna elements. $\theta$ denotes the direction of the main beam, $k$ is the wave number defined by Eq. 3.5 and $d$ is the inter-element spacing. The array factor is maximum for $k \cdot d \cdot \cos \theta + \beta = 0$ and thus the direction of the main beam is given by [16]

$$
\theta = \arccos \left( \frac{\beta}{k \cdot d} \right) = \arccos \left( \frac{\beta \cdot \lambda_0}{2\pi \cdot d} \right). \tag{4.55}
$$

$\beta$ is the element phase factor which if varied will change $\theta$ and hence cause the beam to scan. The spacing between the elements determines the location of grating lobes and the maximum scan angle. The element spacing is chosen to prevent grating lobes. An element spacing greater than $\lambda/2$ in a phased array gives rise to grating lobes of amplitudes equal to that of the main beam. The element spacing can be determined by [16]

$$
d = \frac{\lambda_0}{1 + \sin \theta_{\text{max}}}, \tag{4.56}
$$

where $\theta_{\text{max}}$ is the maximum scan angle from the array normal. Grating lobes occur when $u = k \cdot d \cdot \cos \theta + \beta$ becomes equal to $u = (\pm 2\pi), \pm 4\pi$ etc.
The layout of the phased array is shown in Fig. 4.39a. The dimensions of the dual polarized antennas (patch and two pairs of orthogonal slots) are identical with that of the circularly polarized antenna given in Table 4.7. Each slot aperture is fed by a series phase shifter. The control voltages for the reflection-type phase shifters [61] are generated by a control unit PCB which is mounted on the antenna’s rear side in a similar fashion like the setup of the transponders. A photography of the antenna array’s rear side is shown in Fig. 4.39c. The schematic circuit of the reflection-type phase shifter is depicted in Fig. 4.40c. It consists of a quadrature hybrid with two identical loads. Each load is composed of two cascaded λ/4-lines with short-circuited varactor diodes (Siemens BB833). When designing a reflection-type phase shifter a trade off between ripple of insertion loss and range of phase adjustment has to be done. A cascaded load shows the advantage compared to a single diode load of increasing the phase adjustment range without increasing the insertion loss ripple. In Fig. 4.40c, the cascaded load is divided by a dashed stripline. Each section of the load contributes a phase rotation of about 180°. This operation principle is also indicated in the Smith chart of Fig. 4.40a by a straight line dividing the Smith chart into two halves where each half is associated to the phase rotation of one section. Fig. 4.40b shows the magnitude and the phase of the reflection-type phase shifter. The output port of the phase shifter is connected to the open-circuited stub line feeding the slot aperture of the aperture-coupled antenna.

In Fig. 4.41 the co- and cross-polarized radiation patterns for the transmission of LHCP waves are depicted. The patterns show different element phase factors β. For small values of β (<π/2) the polarization isolation which is approximately 25 dB does not change. The microcontroller on the attached control unit PCB can be programmed to change the phase shifter’s control voltages such that circular polarization-shift-keying modulation is performed while steering the antenna beam. Two phased array antennas (Tx and Rx) could be substituted for the fixed beam reader antenna (however, the phased array’s control unit is not compatible with the reader’s control unit and hence a full operation of the reader with the phased array antenna is not possible yet).
4.3 Active Modulator

The active modulator on the transponder used in the uplink communication increases the transmission distance due to its conversion gain. As described in Chapter 2.4.3 and illustrated in Fig. 2.10, the ratio between conversion gain and increased DC power consumption of the transponder must be chosen optimally for the corresponding tagging application and duty cycle (ratio between switched on and switched off state of the active modulator) in the uplink. As a compromise between increase of transmission distance and increase of power consumption, active backscatter modulation has been chosen as an alternative to a complete transceiver circuit on the transponder. As shown in Fig. 3.11, the active modulator consists of an RF preamplifier succeeded by an upconverter. In this chapter, the design of two types of active modulator will be discussed, namely (1) an active modulator built with off-the-shelf components and (2) a modulator optimized for low DC power consumption where an MMIC amplifier is used.

4.3.1 Transmission Distance in the Downlink and Uplink Communication

In order to demonstrate the effect of the active modulator on the transmission distance, some general considerations about the transmission distance in the downlink and uplink communication will be made prior to the presentation of the active modulator design.

\[ d \text{ [m]} \]

\[ G_{tg} \text{ [dBi]} \]

\[ P_{tg,rx,min} \text{ [dBm]} \]

*Fig. 4.42: Transmission distance \( d \) in the downlink communication versus the minimum admissible power \( P_{tg,rx,min} \) of the transponder's receiver and the transponder's antenna gain \( G_{tg} \).*

1. The transponder battery life is specified at a duty cycle of e.g. 1% in the *Objekt-Identifikations-Systeme Datenbuch* from Baumer Ident GmbH.
Based on Eq. 2.25, the power received at the transponder in the downlink communication is given by

\[ P_{tg, rx} = P_{rd, tx} \cdot G_{rd} \cdot G_{tg} \cdot \left( \frac{\lambda}{4\pi d} \right)^2 = EIRP \cdot G_{tg} \cdot \left( \frac{\lambda}{4\pi d} \right)^2. \] (4.57)

The term \( P_{rd, tx} \cdot G_{rd} \) is called effective isotropic radiated power (EIRP). As mentioned in Chapter 2.1, the maximum EIRP in the 2.4 GHz ISM band is 500 mW. However, a decrease of the maximum EIRP might be expected for future licensing regulations due to the emerging wireless LAN applications in the 2.4 GHz ISM band (e.g. bluetooth or HomeRF). Therefore, the EIRP in the present tagging system is limited to 10 mW. From Eq. 4.57, the maximum transmission distance in the downlink communication can be calculated by

\[ d = \frac{\lambda}{4\pi} \cdot \sqrt{\frac{EIRP \cdot G_{tg}}{P_{tg, rx}_{\text{min}}}}. \] (4.58)

For instance, for a transponder antenna gain \( G_{tg} \) of 4 dBi and a minimum admissible power of -50 dBm at the transponder’s demodulator (which corresponds to the TSS of the detector diodes plus additional losses occurring in the RF front end), the maximum transmission distance is about 15 m. In Fig. 4.42, the transmission distance in the downlink communication based on Eq. 4.58 is depicted.
In order to calculate the maximum transmission distance in the uplink communication, the reflected power from the reader can be written as

\[ P_{tg, tx} = C_{gain} \cdot P_{tg, rx} = C_{gain} \cdot \frac{EIRP \cdot G_{tg} \cdot (\frac{\lambda}{4\pi d})^2}{G_{rd}} \],

(4.59)

where \( C_{gain} \) denotes the conversion gain of the active modulator. The conversion gain is defined as ratio of the power transmitted at the frequency \( f = f_{carrier} + f_{data} \) and the power of the received RF signal at \( f = f_{carrier} \). Substituting Eq. 4.59 for the transmit power into Eq. 2.25 yields

\[ P_{rd, rx} = C_{gain} \cdot EIRP \cdot G_{tg}^2 \cdot G_{rd} \cdot \left(\frac{\lambda}{4\pi d}\right)^4, \]

(4.60)

which can be solved for the transmission distance

\[ d = \frac{\lambda}{4\pi} \cdot \left(\frac{C_{gain} \cdot EIRP \cdot G_{tg}^2 \cdot G_{rd}}{P_{rd, rx, min}}\right)^{\frac{1}{4}}. \]

(4.61)

If for instance a minimum required receiving power \( P_{rd, rx, min} \) of -70 dBm at the reader, a reader antenna gain of 4 dBi (EIRP is still limited to 10 mW) and a conversion gain \( C_{gain} \) of +5 dB is assumed, the maximum transmission distance in the uplink communication is 2.5 m. It can be seen by Eq. 4.61 that the main parameters suited to be adjusted for increasing the transmission distance are the conversion \( C_{gain} \) and the receiver sensitivity of the

**Fig. 4.44: Setup for testing the active modulator on the transponder.**
reader (represented by $P_{rd,rx,min}$). The gain of the reader and transponder antenna are less suited for optimizing the transmission distance as an increase of the antenna gain reduces the beam width and thus the adjusting of the reader and transponder antenna gets more difficult. Fig. 4.43 shows a plot of the transmission distance in the uplink communication versus $C_{gain}$ and $P_{rd,rx}$. The calculations are based on Eq. 4.61.

4.3.2 Active Modulator consisting of Hybrid Circuits

In the previous chapter, it has been shown that the transmission distance is dependent on the fourth root of the active modulator's conversion gain ($\sqrt[4]{C_{gain}}$). Therefore, the design goal of the active modulator is to achieve a high conversion gain at lowest possible DC power consumption. These parameters are optimized best by designing the active modulator as MMIC. However, for practical considerations (mainly manufacturing issues) transponders have been built that use off-the-shelf components for the active modulator.

The active modulator consists of an RF preamplifier succeed by an upconverter where the low frequency port is toggled by the data signal from the microcontroller. Since the expected RF power at the transponder is below -20 dBm (which corresponds to a trans-

![Antenna port isolation](image)

![Preamplifier gain](image)

![$S_{21}$ of mixer diodes](image)

Fig. 4.45: Measured transmission scattering parameters of a) preamplifier (M/A-COM MAAM22010), b) mutually decoupled antenna ports and c) mixer diodes (Mini-Circuits ADE18W) in the conducting state.
mission distance larger than 0.5 m at 10 mW EIRP), the intermodulation performance (1 dB compression point and third order intercept point) may be still good enough even when small transistors are used in the amplifier due to power economy reasons. For the hybrid active modulator, a low noise amplifier for the 2.4 GHz ISM band (M/A-COM MAAM22010) and a doubly balanced (diode ring) mixer (Mini-Circuits ADE18W) have been chosen. The amplifier shows a gain of 12 dB to 16 dB at a power consumption of 9 mW to 30 mW (there are two operation modes) and the output 1 dB compression point and the third order input intercept point are 3 dBm and 1 dBm, respectively. The conversion loss of the mixer is 5.5 dB. Thus, an overall conversion gain of the active modulator between 6.5 dB and 10.5 dB is feasible.

At the selection of components for the active modulator, it has turned out that the transfer characteristic of the RF amplifiers (where it shows a gain larger than 0 dB) is typically much larger than the cross-polarization isolation bandwidth of the antenna. Fig. 4.45 shows the transfer characteristic of the preamplifier and the characteristic of the antenna port decoupling. Even if the losses of the passive mixer and the switches are taken into account, it is obvious that the decoupling characteristic of the antenna is too small to compensate for the gain of the active modulator. In particular, outside the ISM band, where the decoupling performance of the polarizer is getting worse rapidly, the active modulator

![Diagram showing the effect of bandpass filters on S21](image)

**Fig. 4.46:** Effect on $S_{21}$ of the bandpass filters in the loop (antenna, switches, preamplifier and mixer) of the active modulator.
becomes instable. Therefore, in the following, the design of the active modulator will be discussed which takes measures to assure a stable operation.

An improvement in the decoupling of the input and output signal of the active modulator outside the ISM band can be achieved by introducing an additional bandpass filter in the loop consisting of the antenna polarizer, the RF amplifier and the upconverter. This configuration is depicted in Fig. 4.44. As the frequency selectivity of the antenna prevents that frequency components excited outside the ISM band may be transmitted, an additional 10 dB coupler has also been used for monitoring the stability of the loop by means of a spectrum analyzer. An alternative approach to the bandpass filter would be using a compensated branch-line coupler [57]. However, port isolation will be traded for bandwidth which might limit the admissible gain of the RF amplifier at midband.

Fig. 4.46 shows the effect of the additional bandpass filter in the loop of the active modulator. The depicted transmission scattering parameters have been obtained by opening the loop between the 10 dB coupler and the antenna polarizer. A 1\textsuperscript{st} order, a 2\textsuperscript{nd} order and a 3\textsuperscript{rd} order bandpass filter have been designed and used in the loop of the active modulator. The reflection and transmission scattering parameters of the different filters are depicted in Fig. 4.47. As shown in Fig 4.46, the open-loop transfer characteristic shows a gain

\begin{figure}[h]
\centering
\includegraphics[width=\textwidth]{figure4.47}
\caption{Reflection and transmission scattering parameters of 1\textsuperscript{st}, 2\textsuperscript{nd} and 3\textsuperscript{rd} order bandpass filters built on Duroid 6010, z=10.2, h=0.635 mm.}
\end{figure}
larger than 0 dB outside the ISM band. By introducing a bandpass filter, the transfer characteristic is affected such that the open-loop gain drops below 0 dB outside the ISM band. In the passband of the filter, the transfer characteristic of the open loop is only affected by the insertion loss (~1-2 dB) of the bandpass filter and the notch filter like transfer characteristic of the polarizer remains unchanged. For this specific antenna and preamplifier selection, a 1st order bandpass filter does not reduce the gain of the amplifier sufficiently. There is still an open-loop gain at about 2.55 GHz. Indeed, it can be shown by monitoring the branched off signal of the 10 dB coupler that there is an RF signal at 2.53 GHz due to the instability of the active modulator. Fig. 4.49 shows one part of the frequency spectrum where this oscillation occurs.

Another method used for the stability analysis is to determine the $k$-factor of the open-loop configuration of the active modulator. However, it must be taken into account that the open-loop configuration contains more than one loop (feedback path of the RF amplifier and cross-talk path of the dual-polarized antenna connected to the polarizer) and thus, the analysis based on the $k$-factor must be considered carefully. The $k$-factor is defined by
The $k$-factor must be larger than one for a conditional stable operation. It has been shown by means of the experimental setup depicted in Fig. 4.44 that the determination of the $k$-factor gives helpful predictions of the stability behavior of the active modulator. Fig. 4.48 depicts the $k$-factor of the open-loop configuration for the three different bandpass filter
types. Again, it is indicated by the $k$-factor that an instable operation will occur by using only a 1st order bandpass filter.

Fig. 4.50 shows the implementation of the active modulator on the RF front end of the transponder. It consists of a preamplifier (M/A-COM MAAM22010), an upconverter (MiniCircuits ADE18W, doubly balanced diode ring-quad), a bandpass filter 2nd order and two switches (M/A-COM SW338) which are connected to the mutually decoupled antenna ports. The power supply pin of the RF amplifier and the IF port of the upconverter are directly connected to the transponder’s microcontroller. In particular, the power supply of the preamplifier is switched on only during the transmitting period in the uplink communication. For this kind of active modulator consisting of off-the-shelf components, the bandpass filter is the space limiting component. Therefore, at the MMIC design of the active modulator, which will be discussed in the next subsection, one design goal is to eliminate the use of a bandpass filter. This goal could be achieved by designing a matching network for the MMIC RF amplifier which shows already the required frequency selectivity and thus, no external bandpass filter will be needed.

### 4.3.3 Active Modulator as MMIC

In order to reduce the size of the RF front end, some components may be integrated as MMIC. For the design of the integrated circuits, a 0.6 μm E/D ion-implanted MESFET
Prior to starting the MMIC design for the transponder's RF front end, it should be determined which components are suitable for the integration with the available process. An integration is only reasonable, if either the transponder's size is considerably reduced or electrical specifications can be obtained that exceed those of off-the-shelf components. Especially, the last issue prevents an integration of the RF detector circuits and the process (Triquint TQTRx) with a transit frequency of 18 GHz and a maximum frequency of oscillation above 50 GHz was available.

**Fig. 4.52: Comparison between simulated and measured scattering parameters of the MMIC cascode amplifier.**
antenna polarizer with the available MESFET process. The Schottky low-barrier diodes are particularly suited for detector applications due to their low barrier height and thus, MESFET diodes of the TQTRx process used for detector circuits will show a poorer performance (lower saturation current due to higher barrier height) than the optimized zero bias Schottky detector diodes. The integration of the antenna polarizer would reduce the size of the transponder’s RF front end significantly. However, an integrated branch-line coupler built with lumped elements would show a smaller bandwidth than the distributed
elements $90^\circ/3\text{dB}$ hybrid and the insertion loss will be higher as well due to the multi-turn spiral inductors in the MMIC design [56].

The disadvantage of the off-the-shelf absorptive switches is that negative control voltages are required. A large space of the transponder's control unit PCB is occupied by the circuitry generating these control signals. Therefore, a considerable reduction of the transponder's size could be achieved by designing absorptive switches with integrated TTL drivers where the control signals of the transponders microprocessor could directly be used to drive the switches. Again, the TQTRx process is not suited to design such a MMIC as the power consumption would be too high. The OpAmps (Maxim MAX409) used for the generation of the negative control voltages show a current consumption less than 10 $\mu$A which is even lower than the leakage current of a single MESFET.

As already mentioned above, the design of MMICs for the transponder's RF front end is limited to the integration of the active modulator. Due to its low conversion loss (~5.5 dB), the available diode ring quad mixer is not replaced by an MMIC and only the RF amplifier is designed as MMIC. Therefore, the goal of the MMIC design is first to obtain a transfer characteristic of the RF amplifier such that no bandpass filter is required and second to obtain a lower current consumption compared to the off-the-shelf RF amplifier. The latter requirement can be achieved by selecting smaller FETs for the amplifier stage of the active modulator.

A cascode amplifier consisting of two E-FETs with a gate width of 100 $\mu$m was chosen as circuit topology for the MMIC design. The circuit schematic and the corresponding chip photography are shown in Fig. 4.51 and Fig. 4.53. Two L-networks are used for the input and output matching. A comparison between simulated and measured scattering parameters is depicted in Fig. 4.52. The amplifier achieves a gain of 10 dB at a power consumption of about 7 mW (2.2 mA at 3V). In Fig. 4.54 the characteristic of the noise figure and the gain of the MMIC amplifier is depicted. Within the 2.4 GHz ISM band, a noise figure of 2.7 dB is obtained. Compared to the off-the-shelf RF amplifier the gain is 3 dB less but the current consumption is reduced by more than a factor of two. Due to the reduced gain and the more narrow band impedance match, no additional bandpass filter is needed to assure a stable operation of the active modulator outside the ISM band.
4.4 RF Synthesizer

As mentioned in Chapter 3.3.3, the RF synthesizer shows frequency hopping capabilities\(^1\). There are two reasons for using frequency hopping (FH) methods in the tagging system. First, tagging systems used for applications in the factory automation are frequently operated in rooms with metallic walls enforcing the effect of multipath propagation. The radio channel may be frequency-selective in such environments when the reciprocal of the multipath spread is smaller than the bandwidth of the transmitted signal. Fig. 4.55 shows a typical impulse response of a time-invariant channel. The depicted measurements have been performed in a metallic room (paintshop section of the assembly line) at the car manufacturer Audi\(^2\) in Neckarsulm, Germany. The channel displays deep frequency notches which may severely degrade the transmission reliability of a tagging system operated at a fixed carrier frequency. In order to reduce the vulnerability to channel fading effects, frequency hopping methods may be applied where the RF carrier is hopped according to a

![Figure 4.55: Measured S\(_{21}\)-parameters and power delay profile in a typical environment (Audi Werkshalle, Neckarsulm, Germany) for the operation of tagging systems (room dimensions: 6.5m x 9m x 2.1m, measurement bandwidth: 250 MHz, rms delay spread: 190 ns).](image)

---

1. The application of frequency hopping methods was motivated by the VAMPIR system demonstrator built at the wireless lab (Hochschule Rapperswil) [25].
2. The channel measurements at Audi, Neckarsulm, has been made possible by Baumer Ident GmbH, Weinheim, Germany.
A second reason for using FH methods is caused by the reduction of the mutual interference when two closely spaced readers are interrogating two transponders at the same time. If both readers are operated at the same carrier frequency, the strong cross-talk signal would lead to saturation effects in the downconverter stages. By using frequency hopping methods, the impact of the undesired cross-talk carrier is typically eliminated by the IF filter of the first downconverter due to the different hopping patterns of the two readers.

In this work, it has not been investigated which kind of hopping method (fixed hopping pattern or adaptive carrier frequency allocation) is optimal to achieve a maximum transmission reliability. However, frequency synthesizers were developed and implemented in the tagging system demonstrator that show the desired frequency hopping capabilities. The design of the frequency synthesizer is focused on two objectives: (1) high hop rates should be feasible in order to operate the system in a fast FH mode and (2) the complexity of the frequency synthesizer has to be low even if the frequency hops have to cover the whole ISM band.

To cope with these requirements a phased-lock-loop (PLL) operated as frequency multiplier is chosen for the circuit topology of the frequency synthesizer. The schematic of the RF synthesizer is depicted in Fig. 4.56. For operating the PLL in an FH mode either the prescaler values of the feedback and reference signal could be reprogrammed or the refe-
rence signal could be hopped with fixed prescaler values. The variant with the fixed prescaler values shows less programming effort and thus, it was preferred. The hopping reference signal is generated by a direct digital synthesizer (DDS) whose lowpass filtered output signal is frequency hopped within the bandwidth from 8.0 MHz to 8.3 MHz. The microcontroller (PIC 16F84) attached to the DDS (Analog Devices AD9830) can now be programmed to execute an arbitrary hopping pattern. The output frequency of the synthesizer (National Semiconductor LMX2326) will be changed according to the frequency hops of the reference frequency

\[ f_{out} + \Delta f_{out} = \frac{N}{R} \cdot (f_{ref} + \Delta f_{ref}) \],

where \( N/R = 300 \) and the minimum \( \Delta f_{ref} \) is 10 mHz (AD9830 with 50 MHz clock). The prescaler values\(^1\) (\( R = 8 \) and \( N = 2400 \)) are chosen such that the comparison frequency \( f_{comp} \) in the phase detector is 1 MHz. The choice of the comparison frequency is a compromise between the minimum programmable frequency increment and the phase detector noise floor degradation which is given by [58]

\[ PN = PN_{floor} + 20 \cdot \log(N) + 10 \cdot \log(f_{comp}) \],

where \( PN \) denotes the phase noise in dBC/Hz and \( PN_{floor} \) is the phase noise floor at the phase detector normalized to 1 Hz (\( PN_{floor} = -211 \) dBc/Hz for the LMX2326). Lowering the reference frequency shows also the disadvantage that the bandwidth of the loop filter has to be reduced and thus the frequency acquisition time will be enlarged (as a rule of thumb, the loop bandwidth has to be smaller than \( 1/10^{th} \) of the comparison frequency in order to sufficiently suppress spurs generated from the PD at \( f_{comp} \)).

The charge pump signal of the phase detector is led to a 3rd order loop filter that produces the tuning voltage for the voltage controlled oscillator (VCO). At the output of the VCO, a coupler consisting of three microstrip asymmetric coupled lines is used. This coupler is necessary because the applied VCO is very sensitive to changes of the load impedance as it shows no buffer amplifier at the output stage. By means of the coupler, changes of the

![Image of the PLL block diagram and 3rd order loop filter](image)

**Fig. 4.57: a) Basic loop block diagram; b) 3rd order loop filter**

---

1. The PLL design was performed by the programs ‘EasyPLL Design’ and ‘PLL Loop Filter Design’ available from National Semiconductor.
RF synthesizer’s load impedance, which would also affect the VCO output impedance, are reduced. One conductor of the coupler is used for the feedback path, the middle conductor where the VCO output signal is applied is connected to a fixed 50 Ω load and the third conductor is fed to the input of a power amplifier (PA) stage whose bandpass filtered output signal is led to the modulator in the reader. The coupling between the outer conductors to the middle conductor is 10 dB.

### TABLE 4.9

<table>
<thead>
<tr>
<th>phase transfer function (closed-loop transfer function)</th>
<th>( \frac{\theta_o(s)}{\theta_i(s)} = H(s) = \frac{A \cdot K_o \cdot K_d \cdot F(s)}{s + A \cdot K_o \cdot K_d \cdot F(s)} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>phase error transfer function</td>
<td>( \frac{\theta_e(s)}{\theta_i(s)} = 1 - H(s) = \frac{s}{s + A \cdot K_o \cdot K_d \cdot F(s)} )</td>
</tr>
<tr>
<td>open-loop transfer function</td>
<td>( G(s) = \frac{A \cdot K_o \cdot K_d \cdot F(s)}{s} )</td>
</tr>
<tr>
<td>phase error</td>
<td>( \theta_e(s) = \theta_o(s) - \theta_i(s) )</td>
</tr>
</tbody>
</table>
| loop parameters | \( K_d \): PD gain factor (LMX2326: 6.28 mA/2π rad)  
(\( K_o \): VCO gain (JTOS-3000: 120-160 MHz/V)  
(\( A \): loop filter gain  
(\( F(s) \): loop filter transfer function (3rd order)  
(\( \omega_n \): natural frequency (~20 kHz)  
(\( \xi \): damping factor (~1/\sqrt{2}) |

The order of the loop filter determines the tracking performance of the PLL. Due to the integrator in the VCO, the order of the PLL is equal to the loop filter order plus one. In general, PLLs having a 2nd or 3rd order loop filter are used. A 2nd order loop filter is capable of tracking a step change of the input phase of magnitude \( \Delta \theta \) and as well a step change of the input frequency of magnitude \( \Delta \omega \). However, if the input frequency is linearly changing with the time at a rate of \( \Delta \dot{\omega} \), a dynamic tracking error of \( \Delta \dot{\omega} / \omega_n^2 \) occurs where \( \omega_n^2 \) is the natural frequency of the loop filter. A 3rd order loop filter exhibits the improvement in tracking performance over a 2nd order loop filter by eliminating this steady-state tracking error. The analysis of the steady-state error can be performed by applying the final value theorem of the Laplace transforms to the phase error transfer function obtained by the linear PLL model. Using the phase error transfer function given in Table 4.9, the steady-state phase error is determined by

\[
\lim_{t \to \infty} \theta_e(t) = \lim_{s \to 0} \frac{s}{s + A \cdot K_o \cdot K_d \cdot F(s)} \cdot \theta_i(s), \tag{4.65}
\]

where for instance \( \theta_i(s) = \Delta \theta / s \) has to be substituted into Eq. 4.65 for a step change of the input phase.
In contrast to a 2nd order loop filter which is unconditionally stable, a 3rd order loop filter tends to be instable for low values of the loop gain. The stability behavior of PLLs can be analyzed by means of the Bode diagram (the phase margin must be larger than zero for unity gain of the PLL’s closed-loop transfer function) or the root locus plots (the roots of the PLL’s open-loop transfer function may not enter the right half-plane).

**Table 4.10**

<table>
<thead>
<tr>
<th>PLL parameters</th>
<th>Loop filter elements</th>
<th>Simulated values</th>
</tr>
</thead>
<tbody>
<tr>
<td>$K_d=6.28 \text{ mA/\text{rad}}$</td>
<td>$C_1=3.3 \text{ nF}$</td>
<td>$\omega_n=2\pi \cdot 19 \text{ kHz}$</td>
</tr>
<tr>
<td>$K_o=140 \text{ MHz/V}$</td>
<td>$C_2=22 \text{ nF}$</td>
<td>$\xi=0.76$</td>
</tr>
<tr>
<td>$f_{\text{comp}}=1 \text{ MHz}$</td>
<td>$C_3=330 \text{ pF}$</td>
<td>$T_L=85 \mu\text{s for } \Delta f/f_{\text{loop}}=0.01%$</td>
</tr>
<tr>
<td>$R=8$</td>
<td>$R_1=560 \Omega$</td>
<td>$\text{PN}=-83 \text{ dBc/Hz @ } f_{\text{offset}}=150 \text{ Hz}$</td>
</tr>
<tr>
<td>$N=2400$</td>
<td>$R_2=1.5 \text{ k\Omega}$</td>
<td>$\text{rms phase error} = 1.6^\circ$</td>
</tr>
<tr>
<td>$f_{\text{out}}=2.4 \text{ GHz}$</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table 4.10 shows the values of the loop filter elements depicted in Fig. 4.57b which were determined using the program ‘EasyPLL Design’ from National Semiconductor. The calculations of the natural frequency $\omega_n$ [rad/s] and the damping factor $\xi$ are based on the following equations [59]:

$$\omega_n = \sqrt{\frac{K_d \cdot K_o}{N \cdot (C_1 + C_2 + C_3)}}$$  \hspace{1cm} (4.66)

**Fig. 4.58:** Example of a hopping pattern with 300 hop/s (measurements were performed with the modulation domain analyzer HP 53310A; horizontal axis: time, vertical axis: frequency).
As described above the RF synthesizers will be used in the reader to generate frequency hops of the carrier frequency. For this application, it will be interesting how fast the PLL can change from one frequency to another within a specified frequency tolerance (e.g. 1% of the frequency step). In [59], this lock time is given as

$$T_L = -\ln\left(\frac{\sqrt{1 - \xi^2} \cdot f_{\text{tol}}}{\xi \cdot \omega_n} \cdot \frac{\Delta f}{\Delta f} \right),$$

where $f_{\text{tol}}$ specifies the settling tolerance and $\Delta f$ denotes the frequency jump.

Fig. 4.58 shows a typical hopping pattern of the RF synthesizer in the reader which was measured with a modulation domain analyzer. The maximum hopping rate could not be determined because it was limited to 2 khop/s due to the relatively low clock rate of the microcontroller programming the frequency hops of the DDS which is used as reference signal source.

The measured phase noise of the frequency synthesizer is depicted in Fig. 4.59. At an offset frequency of 10 kHz, the frequency synthesizer shows a phase noise of -80 dBc/Hz.
4.5 Signal Processing in the Downlink/Uplink Communication

Data to be transmitted from the reader to the tag is modulated by circular-polarization-shift-keying modulation. This data transmission occurs at the downlink communication where the transmitted data will be stored in the transponder’s memory or at the initializing part of the uplink communication when the read-out command is transmitted.

For the part of the uplink communication where data is transmitted from the transponder to the reader, binary frequency-shift keying modulation with noncoherent detection in the reader is applied. In this chapter, the signal processing for both modulation schemes in the transponder’s and reader’s modulator and demodulator is considered in detail.

![Fig. 4.60: Program structure for the communication protocols in the reader, tag and host PC.](image)

4.5.1 Downlink Communication

An equivalent block diagram of the tagging system during the downlink communication is depicted in Fig. 3.3. The structure of the programs used for the communication protocols in the down- and uplink communication is shown in Fig. 4.60. Prior to the data transmission, the tags have to be initialized by the air interface. That is, the identification number of the tag is transmitted via downlink communication to the transponder where it is stored into the transponder’s memory. Compared to transponders having a read-only ID, this initialization method shows advantages in testing the transponders in an experimental environment. As described in Chapter 3.1, each time the transponder is addressed by the reader, the ID number is transmitted prior to data or commands. For the three different downlink transmissions, namely (1) initializing the transponder, (2) transmitting data to the transponder and (3) transmitting the read out command, the setup of the transmitted bit stream is the same. It consists of a header used for the synchronization in the transponder’s demodulator succeeded by eight data bits. In contrast to commercially available tagging systems (see Chapter 2.4.2), no checksum for recognizing transmission errors is added to...
the payload data. The communication protocol and the data coding is kept as simple as possible in order to use the available 2 kbyte FLASH memory in the transponder’s microcontroller (PIC16F84).

The synchronization sequence consists of the transmission of a logical zero (sync 1 in Fig. 4.61) which is about one byte long, succeeded by the bit sequence 101 (sync 2 in Fig. 4.61). The first synchronization sequence serves for initiating the beginning of the data transmission (the detected pulse at the transponder must be within a specified time interval for being accepted as starting sequence for valid data) and the second synchronization sequence is used to determine the optimum sampling point of time for the payload data. Depending on the alignment between reader and tag, the duration of a logical one is extended at the expense of the duration of a logical zero or vice versa. This behavior is due to the nonideal axial ratio of the transponder’s antenna which causes that the receiver sensitivities for the LHCP and RHCP waves are not equal. For instance, if the antennas are aligned in such a way that LHCP waves are better received, the detector output voltage of the corresponding detector is higher. At the transition from a logical one to a logical zero (transition from LHCP to RHCP waves), the point of time for the transition of the logical signals at the comparator’s output port will change due to the impact of the detectors time constant (both detectors show the same RC-constant but different peak voltages). The higher the data rate the more distinctive this effect will be.

The data rate in the downlink communication is relatively low because low power OpAmps are used for the comparator circuit. Typically, such OpAmps show a gain band-

![Diagram](image.png)

Fig. 4.61: Example: Transmission of ASCII-code ‘a’ at the downlink communication (transmission distance is about 1.5 m). The comparator output voltage $v_{out, comparator}$ is equal to the difference between the detector output voltages $V_{LHCP}$ and $V_{RHCP}$.
width of a few tens of kHz (e.g. 40 kHz for MAX406 with 1.2 μA quiescent current consumption). As shown in Fig. 4.61, the data rate of 1 kbit/s in the downlink communication is even for tagging systems very low. If the application of the tagging system requires a higher speed, the data rate could be enhanced on the expense of a increased current consumption in the comparator circuit (faster comparator circuit).

### 4.5.2 Uplink Communication

The block diagram of the tagging system as used for the uplink communication is depicted in Fig. 4.62. Binary frequency-shift-keying modulation is applied for reading data from the transponder via the backscatter modulation principle. The demodulation of the BFSK modulated data in the reader is performed by a discrete Fourier transform. This type of demodulator can also be considered as a noncoherent correlation-receiver where the determination of the spectral components could be represented by a matched filter for each BFSK tone. The discrete Fourier transform (DFT) is defined as

\[ X(k \cdot \frac{f_s}{N}) = \sum_{n=0}^{N-1} x \left( n \cdot \frac{T}{N} \right) e^{-j \frac{2\pi}{N} k f_s n T}, \]  

(4.69)

where \( f_s = N/T \) is the sampling frequency, \( k \) denotes the \( k \)-th spectral component and \( N \) is the number of frequency components obtained by the discrete Fourier transform. If the number of points in the DFT may be expressed by a power of two, the DFT can be replaced by the fast Fourier transform which shows a reduced computational complexity.
sampled values in the time domain

spectral components in the frequency domain

\[ X(t=0) \rightarrow Y(f=0) \]
\[ X(t=1) \rightarrow Y(f=4)=X(f=-3) \]
\[ X(t=2) \rightarrow Y(f=2) \]
\[ X(t=3) \rightarrow Y(f=6)=X(f=-1) \]
\[ X(t=4) \rightarrow Y(f=1) \]
\[ X(t=5) \rightarrow Y(f=5)-Y(f=-2) \]
\[ X(t=6) \rightarrow Y(f=3) \]
\[ X(t=7) \rightarrow Y(f=7)=Y(f=0) \]

**Fig. 4.63: Signal flow diagram of the implemented 8-point fast Fourier transform with decimation in frequency.**

of only \( N \cdot \log_2(N) \). Fig. 4.63 illustrates the signal flow diagram of the implemented 8-point fast Fourier transform with decimation in frequency (DIF). For the implementation of the FFT-demodulator, a MOTOROLA DSP 56002 with a 16 bit stereo codec (CRYSTAL CS4215) is used. The A/D converter is operated at a sampling frequency of \( f_s = 24 \) kHz.

The BFSK modulation on the transponder is completely performed by the transponder’s microcontroller which toggles the IF port of the upconverter according to the BFSK tones. In order to reduce the requirements of the frequency selectivity at the bandpass filter after the second downconverter stage (suppression of the spectral components at DC and the local oscillator frequency of the second downconverter stage), the frequency of the BFSK tones is chosen as high as possible. The fastest and most accurate method for generating the BFSK tones by the \( \mu C \) is to program the FSK modulator in PIC Assembler instead of using the more convenient PIC Basic. By programming the FSK modulator in PIC Assembler, a maximum toggle frequency of about 25 kHz is feasible with a 10 MHz microcontroller clock rate. Therefore, the frequencies of the BFSK tones are chosen to be \( f_1 = 21 \) kHz for a logical one and \( f_0 = 15 \) kHz for a logical zero (due to the FFT-demodulator, the BFSK tones must be an integer multiple of \( f_s/N \) which is equal to 3 kHz for an 8-point
FFT). The pulse shape of the modulating signal is rectangular \((0,+A=5V)\) and can be expressed by the Fourier series

\[
r(t) = \frac{A_0}{2} \cdot \left[ 1 + 2 \sum_{n=1}^{\infty} \frac{\sin(n\pi/2)}{n\pi/2} \cdot \cos(n \cdot 2\pi \cdot f_{0/1} \cdot t) \right],
\]

where \(f_{0/1}\) represents the two FSK frequencies.

Considering the sampling rate, it can be seen that the BFSK signal will be undersampled since the Nyquist frequency is at \(f_s/2=12\,\text{kHz}\). However, the discrete Fourier transform assumes a periodic continuation of the time domain samples and thus, a frequency shift of the time domain signal \(x(nT)\) by an integer multiple of the Nyquist frequency \(q \cdot f_s/2\) yields the same spectral information since

\[
x(nT) = \sin(2\pi \cdot f_{0/1} \cdot nT) = \sin \left( 2\pi \cdot \left( f_{0/1} + q \cdot \frac{f_s}{2} \right) \cdot nT \right).
\]

The frequency separation between the two FSK tones is chosen such that the first order harmonics of the fundamental of the FSK tones do not produce a cross-talk signal at the evaluated spectral components at the output of the FFT demodulator. Fig. 4.64 illustrates the mapping of the frequency components of the sampled signal to the FFT demodulator output. The carrier frequency of the second downconverter stage is at \(f_{LO2} = f_{IF1} + f_{IF2} = 8\).
Fig. 4.65: Backscattered BFSK modulated data in the uplink communication: a) BFSK signal generated by the transponder’s microcontroller; b) last part of the transmitted sequence displaying the bit sequence 0101 (payload data is subdivided into nibbles); c) transition between a logical one and a logical zero; d) received signal at the input of the decision device in the demodulator of the reader (see Fig. 4.62).

MHz+36 kHz where \( f_{I_2} = 36 \) kHz is equal to three times the Nyquist frequency of the sampler. Thus, the FSK tone for a logical one is at \( f_{I_2} \), \( f_1 = 15 \) kHz which will be mapped to the spectral component \( f_{k=3} \) at the output of the FFT-demodulator. Correspondingly, the FSK tone belonging to a logical zero is at \( f_{I_2} \), \( f_0 = 21 \) kHz and will be mapped to the spectral component \( f_{k=1} \). Four different spectral components are available at the output of the FFT demodulator but only the two spectral components at \( f_{k=1} \) and \( f_{k=3} \) are used for the subsequent signal processing. Since these components are complex valued, the square of their absolute value is calculated as shown in the block diagram of Fig. 4.62. A decision device compares the two values and decides either in favor of \( f_{k=1} \) (logical one) or \( f_{k=3} \) (logical zero). The output signal of the decision device is sampled at a rate of \( f_T = 3 \) kHz \((=f_3/N)\) and thus, the bit rate in the uplink communication is 3 kbit/s which is again rather low compared with commercially available tagging systems. This reduced data speed is mainly caused due to the not appropriate modulator on the transponder and the low sampling frequency of the A/D converter in the reader.

Fig. 4.65 shows an example of the BFSK modulated data generated in the transponder (Fig. 4.65a to Fig. 4.65c) and demodulated in the reader (4.65d). Similar to the downlink communication the transmitted bit sequence consists of a header for the data clock syn-
Fig. 4.66: Frequency spectra at points a) to e) of Fig. 4.62: a) carrier signal at the LHCP antenna port; b) output signal of the SSB/SC upconverter used as local oscillator signal at the first downconverter stage; c) crosstalk signal at the input of the first downconverter stage; d) modulated signal at the input of the second downconverter stage; e) modulated signal at the input of the A/D converter; e') frequency spectrum at the input of the demodulator without BFSK tones.

chronization succeeded by the payload data. The payload data is subdivided into nibbles which are transmitted separately. This subdivision is necessary in order to maintain syn-
chronization. Since the FSK tones are generated by the microcontroller software which
also handles at the same time the data encoding and controlling functions of the RF front end electronics, it is not possible to generate exactly the required 21 kHz \((=\frac{7f_c}{2})\) and 15 kHz \((=\frac{5f_c}{2})\) tones - not even if the time critical parts of the modulator are programmed in PIC Assembler instead of PIC Basic (shorter instruction cycles). Thus, the demodulator in the reader will lose synchronization when longer bit sequences are transmitted and therefore, each nibble is transmitted individually so that the demodulator will be synchronized again.

So far the baseband signal processing in the uplink communication was discussed. Now the downconversion of the received signal to the baseband will be presented. The frequency spectra measured at the labeled points in Fig. 4.62 are depicted in Fig. 4.66. In Fig. 4.66a the frequency spectrum at the input port of the antenna is shown. The power of the carrier frequency is about +2 dBm (cable losses of approximately 2 dB are not accounted for in the frequency spectrum plots). Assuming an antenna gain of about 6 dBi, the effective isotropic radiated power is 8 dBm or approximately 6 mW. As indicated in the marker positions of the spectrum analyzer, the local oscillator signal of the first downconverter stage is suppressed by 51 dB. Remember that this local oscillator signal (at \(f=f_{carrier}-8 MHz\)) is derived from the transmit signal by a frequency shift operation in order to get rid of the frequency stability problem between two RF synthesizers (see Chapter 3.3.2.3).

The frequency spectrum of the local oscillator signal at the first downconverter stage is depicted in Fig. 4.66b. It is obtained by upconverting a filtered 8 MHz signal of a DDS to the transmitted RF frequency by means of an SSB/SC mixer. The power of the desired signal component is +5 dBm which is appropriate to drive the downconverter stage. A single sideband suppression of about 15 dB is achieved together with a carrier suppression of 23 dB.

Fig. 4.66c shows the feed-through signal from the transmit path to the path where the backscattered signal is received. The depicted frequency spectrum is measured with the LHCP and RHCP antenna ports terminated by 50 \(\Omega\) (broadband, ideal impedance match). Thus, the power level difference between the carrier signal of Fig. 4.66a and Fig. 4.66c represents the best case cross-talk suppression. Taking into account the gain of the LNA between the antenna port and the input port of the first downconverter stage, an isolation between the transmit and receive path of about 25 dB would be obtained with an ideal antenna which shows no coupling between its ports. However, the port isolation of the applied reader antenna is 20 dB and the cross-talk level is increased correspondingly. It can be seen from Fig. 4.66c that the gain of the LNA was adjusted such that the power level difference between the local oscillator signal (see Fig. 4.66b) and the RF signal at the first downconverter stage will be still larger than 10 dB which assures a proper operation of the mixer.

In Fig. 4.66d the frequency spectrum at the first intermediate frequency is depicted. It consists of the first intermediate frequency \(f_{IF1}=f_{carrier}-f_{LO1}=2.42\ GHz-(2.42\ GHz-8\ MHz)\), the two FSK-tones at \(f_0=f_{IF1}+f_0=8\ MHz+15\ kHz\) and \(f_1=f_{IF1}+f_1=8\ MHz+21\ kHz\) and the feed-through signal from the second local oscillator at \(f_{LO2}=f_{IF1}+36\ kHz\). The frequency spectrum of the second downconverter stage is shown in Fig. 4.66e. As described
above, the signal for a logical zero is located at $f_{IF2} - f_0 = 21$ kHz and a logical one is represented by the frequency $f_{IF2} - f_1 = 15$ kHz where $f_{IF2} = 36$ kHz. Fig. 4.66e shows the frequency spectrum of the second downconverter stage without the BFSK signals. An increase of the noise floor is visible within the bandwidth of the active bandpass filters. It has turned out in the prototype reader that the minimum admissible power of the FSK-tones is -60 dBm. This minimum required signal level corresponds to a receiver sensitivity of -70 dBm (power level of input RF signal).

Since the transmission distance is limited by the receiver sensitivity in the reader, it must be increased mainly by the following two measures. First, the cross-talk signal could be reduced by a higher port isolation of the Tx/Rx switch (see Fig. 3.10). Thereby, a higher gain of the LNA at the input of the first downconverter stage will be feasible. Second, the resolution (16 bit) of the A/D converter is not fully exploited in order to prevent numerically problems with strong FSK-tones when the tag is close to the reader. An implementation of the FFT-demodulator that deals better with the dynamic range requirements would also improve the receiver sensitivity.

### 4.6 Summary

The main components and the signal processing of the tagging system setup were presented in this chapter. For the demodulator on the transponder, RF detector circuits showing a voltage sensitivity of 25 mV/μW (Fig. 4.7) and a tangential signal sensitivity of about -65 dBm (Fig. 4.9) were designed. In order to build small and compact tags, apertured-coupled patch antennas were developed for the tag. An antenna gain of 6 dBi (Fig. 4.35a) and a cross-polarization isolation better than 15 dB (Fig. 4.37b) was obtained. By neglecting any losses and nonidealities (e.g. offset voltage of the comparator) in the transponder’s demodulator, a transmission distance larger than 10 m (Fig. 4.42) could be expected at 10 mW EIRP transmitted by the reader.

In order to increase the transmission distance, transponders using an active modulator for the uplink communication were built. With off-the-shelf components, a conversion gain of 6 dB at 10 mW DC power consumption is feasible. An MMIC design of the RF amplifier used in the active modulator has been performed which shows a reduced power consumption (Fig. 4.51). A complete reader using a heterodyne demodulator optimized for the demodulation of low data rate signals has been developed and tested. The receiver sensitivity is -70 dBm. For the reader, an RF synthesizer was built which may perform frequency hopping modulation of the RF carrier (Fig. 4.58).
5. System Measurements

This chapter is dedicated to the presentation of measurement results obtained by the complete prototype system consisting of two readers and five transponders. A photography of the measurement setup is shown in Fig. 5.1. The readers are located opposite of each other and the tags are positioned such that two tags (Tag1, Tag5) are accessible by Reader1 and four tags (Tag2, Tag3, Tag4, Tag5) are within the antenna beam of Reader2. This configuration corresponds to the communication link scenario depicted Fig. 3.1 where also one tag (here Tag5) is accessible by two readers. The largest distance between tag and reader is about 3 m. As described in Chapter 3.1 and 4.5, each tag is assigned an identification number prior to be used in the system setup. The application software for allocating the

![Measurement setup for the active tagging system consisting of two readers and five transponders.](image)

Fig. 5.1: Measurement setup for the active tagging system consisting of two readers and five transponders.

![Measurement setup for the determination of the transmission distance and the maximum azimuth angle θ.](image)

Fig. 5.2: Measurement setup for the determination of the transmission distance and the maximum azimuth angle θ.
identification numbers is shown in Fig. 5.3a. In this experimental setup at most four ID numbers may be assigned by the wireless interface. In a first step, this tagging system network is operated by means of the application software depicted in Fig. 5.3b (Visual Basic window) which runs on both host PCs. This program allows to enter data (e.g. one character) and to transmit the entered character to the transponder’s memory. For reading data from the transponder the corresponding button has to be clicked and the read data is displayed in the text box. Again, the program allows to communicate with a maximum of four transponders. It is not possible to communicate with more than one transponder at the same time. By means of this experimental system setup, the statements presented in Chapter 3.1 could be verified. In the downlink communication, the transmission of data to Tag5 (transponder which is accessible by both readers) is only successful for the case that not both readers are transmitting at the same time.
In the uplink communication, it could be demonstrated that the transponders may be read selectively. A command for reading data of several transponders at the same time was not implemented in the application software as the data coding is not appropriate for such a multiple access operation.

In a further measurement setup, the maximum transmission distance and the maximum azimuth angle in the downlink and uplink communication were determined. A sketch of the setup is depicted in Fig. 5.2. The maximum transmission distance is determined by moving the transponder on a straight line away from the reader. For measuring the maximum azimuth angle, the transponder is moved at a distance of 2 m perpendicular to the former movements. The tanglement between reader and transponder remains unchanged. Fig. 5.6 shows the maximum azimuth angle versus carrier frequency for the downlink and uplink communication. The measured characteristics are roughly similar to the antenna gain characteristic. The characteristic measured for the downlink communication shows a less distinctive frequency dependency. Since the transponder is moved perpendicular and not radial to the boresight direction of the reader antenna, the transmission distance changes. Thus, the depicted azimuth angle characteristics may change when the shortest distance between reader and tag is different from 2 m (the distance of 2 m is assumed to be typical for a work tracking application).
Fig. 5.6: Measured azimuth angle $\phi$ defined by Fig. 5.2 in the down- and uplink communication versus RF carrier frequency.

Fig. 5.7 shows the maximum transmission distance versus carrier frequency for the downlink and uplink communication. Despite of the active modulator on the transponder, the transmission distance in the downlink communication is larger than that in the uplink communication. Moreover, it is striking that the maximum transmission distance in the
downlink communication is not at the center frequency of the ISM band although the antenna gain characteristic is centered in the ISM band (see Fig. 4.37). This displacement may be caused by an unsymmetry in the impedance matching of the RF detector diodes. As discussed in Chapter 4.1.5, midband match has been traded for bandwidth and thus, the impedance matching is not optimal (see Fig. 4.14). Therefore, a slight displacement of the impedance matching characteristic of one diode may cause that the carrier frequency, where the maximum transmission distance is obtained, deviates from the center frequency. The worse the impedance match the more distinct this effect will be.

The measured transmission distance in the uplink communication shows a characteristic which is strongly frequency dependent. This strong frequency dependency may be caused by the cascaded transfer characteristics of the quadrature hybrid of the antenna and the bandpass filter in the active modulator. The transmission distance characteristic is roughly similar to the $S_{21}$ characteristic depicted in Fig. 4.43. The maximum measured transmission distance in the uplink communication for a compact tag as shown in Fig. 5.5a, 5.5b and 5.5e is about 3.5 m. This transmission distance is obtained at 6 mW EIRP (see Chapter 4.5.2).

Detailed photographs of the transponders are shown in Fig. 5.5. One transponder is composed of discrete components. This transponder uses the same antenna as the readers what allows to eliminate the bandpass filter (more broadband port isolation). Thus, the performance of this transponder is better than that of the compact ones (transmission distance in the uplink larger than 4 m). Fig. 5.5d shows the control unit PCB, the RF front end substrate and the interconnections between both circuit boards. In Fig. 5.4, the top and bottom side of a reader is depicted. In order to run the application software shown in Fig. 5.2, the reader is connected by an RS-232 interface to the host PC.
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6. Conclusions and Outlook

In this work, an active tagging system operated in the 2.4 GHz ISM was developed and investigated. In a first part, an overview on microwave tagging systems was presented which can be divided into the following three types: Remotely powered fully passive tags (type 1) have to be operated in the near-field region of the interrogator antenna. Therefore, they are used only for short transmission distances. A second type of tagging system (type 2) uses battery-powered tags, which possess a lifetime of several years, as the power supply is only used for the low-frequency signal-processing unit consuming a few microwatts. Due to power economy, no RF generator is available on such a tag; hence, a passive reflex modulator is used for the communication from the tag to the interrogator. The loss of the passive reflex modulator converts directly into a decrease of transmission distance. A third type of tagging system (type 3), which was implemented in the experimental set-up, uses an active modulator on the tag to increase transmission distance. Since the active RF modulator consumes much more battery power than the baseband signal-processing circuits, the decrease in battery lifetime has to be weighted against an increase of transmission distance. In order to reduce the demodulation complexity and thus to lower the DC power consumption on the tag, circular-polarization-shift-keying modulation is used in the downlink communication. For the uplink communication, several system topologies were investigated. Since the backscatter modulated signal is double sideband modulated, a cancellation of the baseband signal occurs when homodyne detection is applied. To prevent this signal cancellation phenomenon either single sideband backscatter modulation should be used at the transponder or the DSB backscatter modulated signal from the transponder should be demodulated at the reader using an I/Q receiver topology or a heterodyne detector. In the experimental system setup a heterodyne receiver topology was implemented which was optimized for the demodulation of low data rate signals. The receiver sensitivity achieved is about -70 dBm.

In the following, the key components and circuits of the experimental tagging system were discussed which include RF detector circuits for the transponder, aperture-coupled microstrip patch antennas for the reader and the transponder, the active modulator for increasing the transmission distance and the RF synthesizer in the reader. For the RF detector zero bias low barrier Schottky detector diodes were used. It was shown that the receiver sensitivity is lower than the optimum value since midband match has to be traded for bandwidth. Aperture-coupled patch antennas were investigated because they allow to separate the feed layer from the radiating patch, resulting in a high front-to-back ratio. In particular, the multi-layered structure is optimal for developing small and compact transponders where the RF front end electronics is located at the rear side of the antenna. Compact circular polarized antennas with switchable polarization sense were developed which show a polarization isolation exceeding 15 dB across the whole 2.4 GHz ISM band. For the active modulator a simple circuit topology consisting of an RF amplifier succeeded by a balanced upconverter was applied. A conversion gain of 6 dB is feasible with a reasonable DC power consumption of typically 10 mW. When using off-the-shelf components for the preamplifier, an additional bandpass filter is necessary to assure a stable operation of the active modulator outside the ISM band. MMICs were developed that eliminate the use of a
bandpass filter due to their optimized transfer characteristic. The design of the RF synthesizer was focused on the development of a fast frequency hopping synthesizer which has to show a preferably low circuit complexity. A circuit topology was chosen where the reference signal of a phased-lock-loop circuit is frequency hopped by a direct digital synthesizer. With the experimental system, measurements were performed to determine the maximum transmission distance and to investigate the operation in a multiple reader multiple tag environment. The transmission distance at 10 mW RF power (EIRP) is about 3.6 m in the uplink and 4.2 m in the downlink communication.

The transmission distance is still limited by the uplink communication. Despite of the active modulator the experimental active tagging system shows a behavior like a tagging system using a fully passive reflex modulator. Therefore, in further work, the transmission distance in the uplink must be considerably increased by improving the receiver sensitivity of the reader. Much sensitivity is lost by the non-optimal active filter circuits in the first and second downconverter stages (see noise floor enhancement at these stages in Fig. 3.78). Moreover, additional transmission distance could be gained in the downlink communication by more identical RF detector circuits on the transponder. Another issue to be addressed in further work is the multiple access capability of the reader in the uplink communication. Some anticollision and multiuser detection procedures are presented in the appendix of this work. However, none of these procedures was implemented in the experimental tagging system yet.
7. Appendix

When the reader sends out a command, this is processed by all transponders within its transmission range. Simultaneous data transmission by several transponders will lead to mutual interference and therefore to data loss. Data loss caused by multiple access to a transmission channel is known as collision. Due to competitive reasons, tagging system manufacturers do not publish the anticollision procedures that they use. For this reason, little information is available on this subject in the technical literature.

In this appendix, some commonly used anticollision procedures are first presented. In the subsequent part, a multiple access procedure is discussed which could be implemented in the experimental active tagging system.

7.1 Anticollision Procedures

The anticollision procedures commonly used in tagging systems can be classified into three groups [1]:

- **Spatial domain anticollision procedures**: The reader uses a tightly bundled directional antenna which scans the area around the reader and selects only a small number of transponders until the desired transponder is found.

- **Frequency domain anticollision procedures**: In this type of procedure, spread spectrum signalling is used for the data transmission from the transponder to the reader.

- **Time domain anticollision procedures**: Mainly two types of anticollision procedures can be distinguished. In the anticollision procedure which is controlled by the transponder the transponder’s response signals are randomly delayed and thus it is highly likely that two transponders will send their identification code at different times and no data collision will occur. However, these transponder driven procedures are very slow and inflexible. In the more flexible reader driven anticollision procedures a binary search procedure is applied. An illustration of this algorithm is given in Fig. 7.1. The transmit-

![Figure 7.1](image-url)
ted data from the transponders (e.g. their identification code) are Manchester coded (logical one: signal transition from low to high at $T/2$, logical zero: signal transition from high to low at $T/2$). First the superposition of all response signals is analyzed. Due to the signal transitions in the middle of the bits, a bit error can be detected if the signal level do not change for the duration of an entire bit. Once a bit error was detected, the reader may selectively eliminate the undesired response signals by limiting the number of responding transponders. A good strategy to limit the number of responding transponders is to allow only transponders to reply which show a logical zero at the MSB error bit. Thus at least one half of the transponders will no longer respond. This procedure is repeated until the desired transponder is detected. Another time domain anticollision procedure similar to that described above is discussed in reference [62].

### 7.2 Multiple Access Procedure for the Active Tagging System

The multiple access procedure discussed in this chapter is based on the references [63], [64]. It uses multilevel frequency-shift keying (MFSK) modulation and thus, it could be implemented with reasonable effort in the experimental active tagging system which already applies BFSK modulation in the uplink communication.

#### 7.2.1 Coding of the Finite Field (Galois Field)

- $GF(Q)$: field of $Q$ elements (e.g. $Q=2^3=p^3$, elements are expressed as three-digit binary number (k-tuplets): $1=001, ..., 7=111$).

- Addition in $GF(Q)$ is defined as mod $p$ addition of the element components (e.g. $5+4=101+100=001=1$).

- Multiplication in $GF(Q)$: first the k-tuplets are transformed into polynomials in $z$ of degree $k-1$ where the coefficients correspond to the digits in the tuplet (e.g. $5=101$ corresponds to $z+1$, $3=011$ corresponds to $z+1$, etc.). The multiplication rule in $GF(p^k)$ is defined as polynomial multiplication modulo an irreducible polynomial and mod $p$ for the coefficients (e.g. $5*3=(z+1)(z+1)=z^2+z+1=(z+1)+z^2+1=(z+1)+z^2+2z+2$ mod $2=z^2=100=4$).

**Fig. 7.2:** a) Time-frequency matrix; b) Tag ID number 4 represented by the time-frequency matrix and encoded by the address code $R_0$; c) Definition of the algebraic operations in the finite field $GF(Q)$. 
Fig. 7.3: An example of the multiuser detection scheme: The decoded matrices are obtained by subtracting the address code vectors from the received matrix. Then the majority rows of the decoded matrices are determined which yield the symbol matrices. Afterwards the address code vectors are added to the symbol matrices to obtain the cancelling matrices from whom the candidate matrices are derived by performing logical OR operations on the matrix entries. The decision is made in favor of the symbol matrix which shows the most number of coincident entries with the received time-frequency matrix.

The present multiuser detection scheme is based on the representation of the transponder’s ID number by a time-frequency matrix. An example of a time-frequency matrix is shown in Fig. 7.2a. On the horizontal axis the time slots are displayed and the vertical axis is divided by the MFSK tones. For illustration the tag with the ID number 4 is represented by the time-frequency matrix of Fig. 7.2b. Thus, in the uncoded state this tag would send the MFSK tone $f_4$ during the entire symbol duration $T = L \cdot T_c$. In order to apply error-correcting coding, the time-frequency matrix is encoded by an address code. The address code generation is based on the algebraic operations of a finite element field (Galois field) which are listed in Fig. 7.2c. The address codes used here are generated by the following vector

$$R_m = (\alpha_m, \alpha_m \beta, \alpha_m \beta^2, \ldots, \alpha_m \beta^{L-1}) \ ,$$

(7.72)
where \( \alpha_m \) is an element of \( GF(Q) \) assigned to the \( m \)-th user and \( \beta \) is a fixed primitive element of \( GF(Q) \). For instance, in \( GF(Q) \) with the parameters \( k=3, L=5 \) and \( \beta=2 \) the following address codes are obtained: \( R_0 = (1,2,4,3,6) \), \( R_1 = (2,4,3,6,7) \), \( R_2 = (3,6,7,5,1) \), \( R_3 = (4,3,6,7,5) \), \( R_4 = (5,1,2,4,3) \), \( R_5 = (6,7,5,1,2) \), \( R_6 = (7,6,1,2,4) \). In the example of Fig. 7.2b, the encoding of the tag ID \([4,4,4,4,4]\) is performed by the address code \( R_0 = [1,2,4,3,6] \) whereas a modulo-\( Q \) addition (\( Q=2^3 \)) is executed at each digit of both tuplets: \(([4,4,4,4,4] + [1,2,4,3,6] = [5,6,1,7,3])\). The coded ID number is then transmitted to the reader during the uplink communication. Clearly, the transponder begins the transmission with the MFSK tone \( f_5 \) in the first time slot and finishes the transmission with the MFSK tone \( f_3 \) in the last time slot. Fig. 7.3 shows an example of the multiuser detection scheme whereas the superposition of all transponder signals received by the reader is represented by the received time-frequency matrix. In a first step the decoding is performed by subtracting all address code vectors from the received time-frequency matrix. It is assumed for this example that there are three address vectors. The number of address code vectors used for decoding depends on how many responding tags are expected. A decoding error occurs when there are several rows having the largest number of entries, \( L \), in the time-frequency matrix. These rows are called majority rows. In the example an error occurred at decoding with address code \( R_0 \) since the decoded time-frequency matrix shows two majority rows. In the next step, the decoded time frequency matrices are duplicated by considering only the majority rows. This procedure yields the symbol matrices. Now the address code vectors are added again to the symbol matrices in order to get the canceling matrices. The canceling matrices are different from each other only for address code vectors where a decoding error occurred (in this example there are different canceling matrices at \( R_0 \)). All corresponding entries of the canceling matrices are combined by logical OR operations. The resulting matrices are termed candidate matrices. The final decision for the transmitted tag ID number is done by determining the number of coincident entries between the candidate matrices and the received matrix. The decision is made in favour of the candidate matrix having the most number of coincident entries.
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# Abbreviation and Symbols

<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Description</th>
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<tbody>
<tr>
<td>A/D</td>
<td>analog-to-digital (converter)</td>
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<tr>
<td>AGC</td>
<td>adaptive gain control</td>
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<td>AR</td>
<td>axial ratio</td>
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<tr>
<td>ASK</td>
<td>amplitude-shift-keying (modulation)</td>
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<tr>
<td>$\beta_v$</td>
<td>voltage sensitivity</td>
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<tr>
<td>BER</td>
<td>bit error rate</td>
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<tr>
<td>BFSK</td>
<td>binary frequency-shift-keying (modulation)</td>
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<tr>
<td>BP</td>
<td>bandpass (filter)</td>
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<tr>
<td>CPSK</td>
<td>circular-polarization-shift-keying (modulation)</td>
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<tr>
<td>CW</td>
<td>continuous wave</td>
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<tr>
<td>D/A</td>
<td>digital-to-analog (converter)</td>
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<tr>
<td>DDS</td>
<td>direct digital synthesizer</td>
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<td>DFT</td>
<td>discrete fourier transform</td>
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<td>DSB</td>
<td>double sideband</td>
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<tr>
<td>DSP</td>
<td>digital signal processor</td>
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<tr>
<td>DUT</td>
<td>device under test</td>
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<tr>
<td>EIRP</td>
<td>effective isotropic radiated power</td>
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<tr>
<td>EM</td>
<td>electromagnetic (solver)</td>
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<tr>
<td>$\varepsilon_r$</td>
<td>relative permittivity</td>
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<tr>
<td>FFT</td>
<td>fast fourier transform</td>
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<tr>
<td>FH</td>
<td>frequency hopping</td>
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<tr>
<td>FM</td>
<td>frequency modulation</td>
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<tr>
<td>FSK</td>
<td>frequency-shift-keying (modulation)</td>
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<tr>
<td>GaAs</td>
<td>Gallium Arsenide</td>
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<tr>
<td>ID</td>
<td>identification (number, code)</td>
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<tr>
<td>IF</td>
<td>intermediate frequency</td>
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<tr>
<td>ISM</td>
<td>industrial-scientific-medical</td>
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<td>Abbreviation</td>
<td>Description</td>
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<td>--------------</td>
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<tr>
<td>LHCP</td>
<td>left-hand circularly polarized</td>
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<td>LO</td>
<td>local oscillator</td>
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<tr>
<td>LP</td>
<td>lowpass (filter)</td>
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<tr>
<td>μC</td>
<td>microcontroller</td>
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<tr>
<td>MESFET</td>
<td>metal semiconductor field effect transistor</td>
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<tr>
<td>MMIC</td>
<td>monolithic integrated microwave circuits</td>
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<tr>
<td>NWA</td>
<td>network analyzer</td>
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<tr>
<td>PA</td>
<td>power amplifier</td>
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<tr>
<td>PCB</td>
<td>printed circuit board</td>
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<tr>
<td>PLL</td>
<td>phase-locked-loop</td>
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<td>PN</td>
<td>pseudo noise (sequence)</td>
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<td>RF</td>
<td>radio frequency</td>
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<td>radio frequency identification</td>
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<td>RHCP</td>
<td>right-hand circularly polarized</td>
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<td>rms</td>
<td>root-mean-square</td>
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<td>r/w</td>
<td>read/write (tag)</td>
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<tr>
<td>Rx</td>
<td>receive (antenna)</td>
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<tr>
<td>SAW</td>
<td>surface acoustic wave</td>
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<tr>
<td>S-parameter</td>
<td>scattering parameters</td>
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<tr>
<td>SPDT</td>
<td>single pole double through (switch)</td>
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<td>SSB</td>
<td>single sideband</td>
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<td>SWR</td>
<td>standing wave ratio</td>
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<tr>
<td>TSS</td>
<td>tangential signal sensitivity</td>
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<td>Tx</td>
<td>transmit (antenna)</td>
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<tr>
<td>VCO</td>
<td>voltage controlled oscillator</td>
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<tr>
<td>VSWR</td>
<td>voltage standing wave ratio</td>
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Curriculum Vitae

I was born on November 14 1972 in Bretzwil, Switzerland. From 1979-1992, I attended public schools in Bretzwil, Reigoldswil and Liestal and received a Matura type C in 1992.

In 1992 I started my studies at the Swiss Federal Institute of Technology (ETH) in Zürich where I received a Dipl. El. Ing. degree in 1996. In April 1996 I joined the Microwave Electronics Laboratory (IFH) at ETH. I participated in a research project investigating active tagging systems for the 2.4 GHz, 5.8 GHz and 24 GHz ISM bands. During this project, I had the opportunity to build an experimental setup of an active tagging system operated in the 2.4 GHz ISM band. Parts of the system could be patented by means of the financial support of the industrial project partners Baumer Electric AG and Baumer Ident GmbH. This system setup is the basis of this dissertation.

In 1999 I started post-diploma studies (Nachdiplomstudium Informationstechnik) at ETH.
Publications

The listed publications are related to the project ‘Active Microwave Tagging Systems for the ISM Bands 2.4 GHz, 5.8 GHz and 24 GHz (ACTM1T5.07)’ from the Swiss Priority Program in Micro & Nano System Technology (MINAST):

Journal Papers:


Conference Papers:


Patents


MINAST Proceedings


[16] MINAST Final Convention Proceedings, Bern, November 30, 1999, pp. 94-95


Technical Documentation

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