Backscatter Radio Frequency Systems and Devices for Novel Wireless Sensing Applications

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Abstract

This thesis examines the use of backscatter radio frequency identification (RFID) in novel wireless sensing applications. Backscatter RFID in sensor networks relies on the radio communication between an RFID reader, acting as a control unit, and a multitude of passive or semi-passive RFID transponders (tags), acting as sensor nodes. All power for the transmission of the sensor data is drawn from the electromagnetic field radiated by the reader. Hence, their low-power consumption makes backscatter tags appropriate for sensing applications that require small, light-weight, and low-maintenance sensor nodes.

It is vital to ensure a reliable power transfer and wireless communication between the reader and the tags. Thus, the major design goal in this work is to realize backscatter RFID systems and devices which lead to high system performances.

Particular attention is paid to the design of a tag antenna at 864 MHz for the wheel unit (WU) of an advanced tire monitoring system (ATMS). One premise of the application is that the antenna is directly attached to a car tire. The tire’s material parameters vary strongly with type and vendor of the tire. Thus, a tag antenna is prototyped which can cope with a change in the tire environment. An analysis shows that the antenna prototype assures a stable power supply to the tag’s chip for different tire environments and leads to a good system performance.

In addition, an on-body backscatter RFID system is evaluated for remote health monitoring applications at 900 MHz and 2.45 GHz. The system performance is evaluated in a realistic indoor scenario through on-body channel measurements. An analysis of a state-of-the-art system example shows that the use of semi-passive chips leads to a reliable performance in the system’s forward link. A strategy to overcome limitations in the system’s backward link is to use a phase-modulated backscatter signal.

Finally, a backscatter RFID sensor tag is designed that monitors the varying curvature of an object at 5.8 GHz. The backscatter sensor includes a transducer prototype which changes its impedance as a function of bending and directly modulates the carrier signal sent from the RFID reader. The transducer prototype is optimized with respect to the sensor’s sensitivity to bend and with respect to the sensor tag’s modulation efficiency. It is found that the prototype qualifies for integration in the sensor tag and assures a good RFID system performance.
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1 Introduction

Since the concept of modulating backscatter for communication was proposed in 1948 [1], considerable research and development has been invested in the area of backscatter radio frequency (RF) systems and devices, or rather backscatter radio frequency identification (RFID)\(^1\) systems and devices [2, 3].

Backscatter RFID in the ultra high frequency (UHF) and microwave frequency ranges is a promising communication technology for wireless sensing applications [4, 5]. In particular, the operation in the unlicensed frequency bands around 900 MHz [3], 2.45 GHz [3], and 5.8 GHz [6], is attractive for many sensor systems because of relatively large communication distances, comparatively small devices, and potentially high data rates. Backscatter RFID in sensor networks relies on the radio communication between an RFID reader, acting as a control unit, and a multitude of passive or semi-passive RFID transponders (tags), acting as sensor nodes. The principle of communication for transmitting information from the tag to the reader relies on a modulated backscatter signal. All power for the transmission of the sensor data is drawn from the electromagnetic field radiated by the reader. Hence, their low-power consumption makes backscatter tags appropriate for sensing applications that require small, light-weight, and low-maintenance nodes.

In backscatter RFID systems, it is vital to ensure a reliable power transmission to the backscatter tags and to realize a robust wireless communication between the reader and tags. Thus, the proper design of backscatter RFID devices, which are included in sensor tags, and the investigation of backscatter radio channels are key study areas.

This thesis examines the use of backscatter RFID in sensing applications like advanced tire monitoring, remote health monitoring, and curvature monitoring. Special attention is paid to prototyping and evaluation of backscatter RFID systems and devices, particularly to the development of systems and devices that are reliable in adverse operating environments like a car tire and the human body.

\(^1\)In general, the terms “backscatter RF” and “backscatter RFID” are used interchangeably in this work.
Original Publications Related to This Chapter


1.1 Backscatter RFID System

A backscatter RFID system relies on the wireless communication between an RFID reader and a backscatter tag [3]. A schematic of the backscatter communication system with emphasis on the RFID tag is depicted in Fig. 1.1. The wireless communication link between the RFID reader and the backscatter tag can be divided into a forward link and a backward link. In the forward link, the RFID reader transmits RF power and data to the tag. In the backward link, whenever a reader command requires a tag’s response, the tag starts its data transfer using a modulated backscatter signal, i.e., the signal from the reader is reflected by the tag depending on the transmitted data. Typically, the reflected signal switches between two states and thus represent a logical ‘0’ and ‘1’.

1.1.1 Backscatter Transponder

A conventional RFID tag consists of an antenna and an application specific integrated circuit chip. Fig. 1.1 shows such a tag. The tag’s antenna and chip are characterized by their complex impedances: by the impedance of the antenna, \( Z_{\text{Ant}} = R_{\text{Ant}} + jX_{\text{Ant}} \) and by the absorbing and reflecting impedances of the chip, \( Z_{\text{Abs}} = R_{\text{Abs}} + jX_{\text{Abs}} \) and \( Z_{\text{Ref}} = R_{\text{Ref}} + jX_{\text{Ref}} \).

As stated earlier, the tag uses a modulated backscatter signal to communicate with the reader. Such a signal is realized by switching the chip’s input impedance between its absorbing impedance, \( Z_{\text{Abs}} \), and its reflecting impedance, \( Z_{\text{Ref}} \) (see Fig. 1.1). Hence, the signal from the reader is reflected at the chip’s input depending on the tag’s data.

The reflected power at the chip’s input is related to reflection coefficients in the absorbing mode, \( S_{\text{Abs}} \), and the reflecting mode, \( S_{\text{Ref}} \), which are defined by the antenna impedance and the respective chip impedance [7],

\[
S_{\text{Abs}} = \frac{Z_{\text{Abs}} - Z_{\text{Ant}}^*}{Z_{\text{Abs}} + Z_{\text{Ant}}} \quad \text{and} \quad S_{\text{Ref}} = \frac{Z_{\text{Ref}} - Z_{\text{Ant}}^*}{Z_{\text{Ref}} + Z_{\text{Ant}}} \tag{1.1}
\]
1.1 Backscatter RFID System

Figure 1.1: Backscatter RFID system with emphasis on the RFID tag: The backscatter tag consists of an antenna and a microchip which are characterized by their complex input impedances, $Z_{\text{Ant}}$, $Z_{\text{Abs}}$, and $Z_{\text{Ref}}$.

In the case of an ideally amplitude-modulated backscatter signal\(^2\), the reflection coefficient in the absorbing mode is zero, $S_{\text{Abs}} = 0$ [3]. This is true for an impedance match between the antenna and the chip, i.e., the antenna’s impedance is the complex conjugate of the chip’s absorbing impedance [7],

$$Z_{\text{Ant}} = Z_{\text{Abs}}^\ast.$$ \hfill (1.2)

As a consequence, the power available at the chip’s circuitry reaches its maximum. That is, the power transmission in the system’s forward link is maximized which corresponds to a power transmission coefficient of $\tau = 1$ or rather $\tau = 100\%$. This transmission coefficient is defined as [8, 9]

$$\tau = 1 - |S_{\text{Abs}}|^2 = \frac{4R_{\text{Abs}}R_{\text{Ant}}}{|Z_{\text{Abs}} + Z_{\text{Ant}}|^2}.$$ \hfill (1.3)

The magnitude of the reflection coefficient in the reflecting state is one, $|S_{\text{Ref}}| = 1$, in the case of an ideally amplitude-modulated backscatter signal [3]. Consequently, the power in the backward link of the system is maximized. The modulation efficiency, $\eta$, compares the backscatter signal-relevant power reflected at the chip’s

\(^2\)In this work, an amplitude-modulated backscatter signal is assumed where not otherwise stated.
input with the available power at the antenna’s output and is defined as [9, 10]
\[ \eta = \frac{2}{\pi^2} |S_{\text{Abs}} - S_{\text{Ref}}|^2 = \frac{4R_{\text{Ant}}^2}{\pi^2} \cdot |Z_{\text{Abs}} - Z_{\text{Ref}}|^2 \cdot |Z_{\text{Ref}} + Z_{\text{Ant}}|^2. \] (1.4)

From Eqn. 1.4, it follows that a modulation efficiency of \( \eta \approx 20\% \) can be achieved for an ideally amplitude-modulated backscatter signal, i.e., \( S_{\text{Abs}} = 0 \) and \( |S_{\text{Ref}}| = 1 \). A maximum modulation efficiency of \( \frac{8}{\pi^2} \approx 81\% \) can be realized for reflection coefficients which have magnitudes of 1 and contrary phases [10], i.e., for an ideally phase-modulated backscatter signal with \( S_{\text{Abs}} = 1 \) and \( S_{\text{Ref}} = -1 \) [3].

**Transponder Chip**

In general, an off-the-shelf RFID chip is a nonlinear load whose complex impedance in each state varies with the frequency and input power. The power dependence is determined by the details of the chip’s RF frontend and the power consumption of the specific chip, while the frequency dependence is mostly due to the chip’s parasitic and packaging effects [11].

To maximize the power transfer in the forward and backward links of a backscatter system, knowledge of the chip’s impedances is mandatory. As an example, Fig. 1.2 plots the measured impedance in the absorbing mode, \( Z_{\text{Abs}} \), of a passive UHF RFID chip at a frequency of \( f = 864 \text{ MHz} \). The knowledge of \( Z_{\text{Abs}} \) is especially important in the case of a passive RFID chip which does not have its own power supply and thus relies completely on the wireless power transfer from the reader. In Fig. 1.2, the absorbing impedance of the exemplary chip, \( Z_{\text{Abs}} = R_{\text{Abs}} + jX_{\text{Abs}} \), is plotted versus the chip’s input power, \( P_{\text{Tag}} \). It can be seen that \( Z_{\text{Abs}} \) is highly reactive. Additionally, the impedance changes drastically at high power levels. This is due to an internal power regulation in the chip, the absorbing impedance converges with increasing power to the reflecting impedance to protect the chip’s internal circuit [9].

A backscatter tag chip needs a certain minimum power to turn on its circuitry. This power level is known as the chip’s sensitivity, \( T_{\text{Chip}} \). Again, this threshold is defined by the details of the chip’s RF frontend. The chip’s sensitivity, \( T_{\text{Chip}} \), can be found by measuring the chip’s input impedances versus its input power [9]. Off-the-shelf passive RFID chips, which do not have their own power supply, have chip sensitivities of about \(-15 \text{ dBm} \), e.g., \( T_{\text{Chip}} = -17.8 \text{ dBm} \) [12], while semi-passive chips feature sensitivities down to \(-40 \text{ dBm} \) [13]. Semi-passive chips have their own power supply to activate the chip’s circuitry. Additionally, they harvest power and use power-saving backscattering for the communication with the reader [14].
1.1 Backscatter RFID System

Figure 1.2: Absorbing impedance of an exemplary chip, \( Z_{Abs} = R_{Abs} + jX_{Abs} \), versus the chip’s input power, \( P_{Tag} \), at 864 MHz

Transponder Antenna

Proper impedance matching between the tag’s antenna and the chip is very important for backscatter RFID systems. Typically, the input impedance of the chip cannot be chosen arbitrarily due to technological limits. Thus, the antenna’s impedance is optimized to match the chip’s impedance.

As stated earlier, knowledge of the chip’s absorbing impedance allows a tag antenna design which is optimized to maximize the forward link of the backscatter RFID system. This impedance matching is usually done at the chip’s sensitivity, \( T_{Chip} \) [15]. In the example presented above (see Fig. 1.2), the absorbing impedance is \( Z_{Abs} = (68 - j442) \Omega \). Fig. 1.3 plots the power transmission coefficient, \( \tau \), versus the antenna’s impedance, \( Z_{Ant} \) (see Eqn. 1.3). It can be seen that \( \tau \) is indeed maximized for an antenna impedance of \( Z_{Ant} = (68 + j442) \Omega \) as stated in Eqn. 1.2.

In addition, the antenna impedance can be optimized to maximize the signal that is reradiated towards the reader which is equivalent to the use of a phase-modulated backscatter signal. For this optimization, the knowledge of both chip impedances is mandatory [9],

\[
Z_{Ant} = \sqrt{\frac{R_{Abs} R_{Ref} (R_{Abs} + R_{Ref})^2 + (X_{Abs} - X_{Ref})^2}{(R_{Abs} + R_{Ref})^2}} - j \frac{R_{Ref} X_{Abs} + R_{Abs} X_{Ref}}{R_{Abs} + R_{Ref}} \tag{1.5}
\]
In Fig. 1.4, the modulation efficiency, $\eta$, is plotted versus the antenna’s impedance, $Z_{\text{Ant}}$, for an absorbing impedance of $Z_{\text{Abs}} = (68 - j442) \, \Omega$ and a reflecting impedance of $Z_{\text{Ref}} = (2 - j0.1) \, \Omega$ (see Eqn. 1.4). $\eta$ reaches its maximum for an antenna impedance of $Z_{\text{Ant}} = (75 + j13) \, \Omega$, while the power transmission coefficient is modest 10% (see Fig. 1.3). Thus, such an optimization is only advisable if the forward link can tolerate small power transmission coefficients. For an antenna impedance of $Z_{\text{Ant}} = (68 + j442) \, \Omega$, $\eta$ is 20% according to Eqn. 1.4 for $S_{\text{Abs}} = 0$ and $|S_{\text{Ref}}| = 1$.

The variation of the chip’s impedances as a result of the input power can drastically affect the matching between the tag’s antenna and chip. It is possible to have a situation where a significant variation of the chip’s impedances results in dead spots within the operational range of the tag [16]. Today’s backscatter RFID systems combat this effect by using for example transmit diversity at the RFID reader [17].

**Scattering Properties of Antennas**

In general, the power reflected from a receiving antenna for any load impedance can be written as the sum of two terms [18], the power scattered due to the antenna mode and the power scattered due to the structural mode of the antenna [9]. Consequently, it is incorrect to assume that the receiving antenna scatters as much as its absorbs under matched load conditions [19], i.e., $Z_{\text{Ant}} = Z_{\text{Abs}}^*$. In general, the reflected power may be larger, equal to, or smaller than the absorbed power [19].

The reflected power due to the antenna mode depends on the chip’s impedances, while the power due to the antenna’s structural mode is not influenced by an impedance change. Thus, the structural scattering component cannot be used for the data transmission in an RFID system and is not relevant for RFID applications [9]. This scattering component is similar to a reflected signal that is caused by an interacting object in the propagation path between the reader and tag and is furthermore treated as such (see Sec. 1.1.3).
Figure 1.3: Power transmission coefficient, $\tau$, versus the antenna’s impedance,
$Z_{\text{Ant}} = R_{\text{Ant}} + jX_{\text{Ant}}$, for $Z_{\text{Abs}} = (68 - j442) \Omega$: The optimized antenna impedances, $Z_{\text{Abs}} = (68 - j442) \Omega$ according to Eqn. 1.2 and $Z_{\text{Ant}} = (75 + j13) \Omega$ according to Eqn. 1.5, are highlighted.
Figure 1.4: Modulation efficiency, $\eta$, versus the antenna’s impedance, $Z_{\text{Ant}} = R_{\text{Ant}} + jX_{\text{Ant}}$, for $Z_{\text{Abs}} = (68 - j442) \Omega$ and $Z_{\text{Ref}} = (2 - j0.1) \Omega$: The optimized antenna impedances, $Z_{\text{Abs}} = (68 - j442) \Omega$ according to Eqn. 1.2 and $Z_{\text{Ant}} = (75 + j13) \Omega$ according to Eqn. 1.5, are highlighted.
1.1 Backscatter RFID System

1.1.2 Backscatter Radio Channel

In a backscatter RFID system, a bidirectional radio link is established between the reader and tag — the reader-tag-reader link — which can be subclassified into the forward link and backward link [6]. The link budget of the backscatter radio channel is outlined in Fig. 1.5.

In the forward link, the reader transmits RF power, $P_{\text{TX}, \text{Reader}}$, and data to the tag. The power absorbed by the tag’s chip, $P_{\text{Chip}}$, is defined by [9]

$$P_{\text{Chip}} = \tau P_{\text{Tag}} = \tau |S_{21}|^2 P_{\text{TX}, \text{Reader}},$$

(1.6)

where $P_{\text{Tag}}$ is the chip’s input power and $S_{21}$ is the channel transfer function of the forward link. $P_{\text{Chip}}$ should be higher than the chip’s sensitivity, $T_{\text{Chip}}$. If $P_{\text{Chip}}$ is smaller than $T_{\text{Chip}}$, the backscatter communication is limited in its forward link [16].

In the backward link, the tag responds to the reader by modulating the backscattered signal. The power of the tag’s signal at the receiver (RX) of the reader, $P_{\text{RX}, \text{Reader}}$, can be written as [9]

$$P_{\text{RX}, \text{Reader}} = |S_{12}|^2 \eta P_{\text{Chip}} = |S_{12}|^2 \eta |S_{21}|^2 P_{\text{TX}, \text{Reader}},$$

(1.7)

where $S_{12}$ is the channel transfer function of the backward link. $P_{\text{RX}, \text{Reader}}$ should be higher than the RX’s sensitivity, $T_{\text{RX}, \text{Reader}}$, which is defined as the minimum input power at the reader to assure a successful reception of the tag’s data. If $P_{\text{RX}, \text{Reader}}$ is smaller than $T_{\text{RX}, \text{Reader}}$, the communication system is limited in its backward link [16].

The channel transfer functions, $S_{21}$ and $S_{12}$, depend on the antenna characteristics of the reader and tag (e.g., antenna gain, polarization) and the properties of the propagation channel (e.g., path loss, fading) [20].
1 Introduction

1.1.3 Interrogator

Fig. 1.6 depicts the RF architecture of an RFID reader with a monostatic antenna configuration [11]. Such a configuration applies a single antenna for the transmit and receive paths which are — in this example — separated by a circulator. The transmitter (TX) produces the RF signal to power the tag and to send commands, while the RX takes care of amplifying and purifying the received signal [21]. A baseband processor generates the command sequences, demodulates the received signals, and controls the protocol.

The reader constantly transmits a continuous wave signal to the tag except during the times of interrogation when the carrier is modulated. After a certain idle period, the tag responds to the reader by backscattering the carrier [22]. During the entire process, the transmit signal leaks into the receive path. The amplitude of the leakage signal depends on the TX to RX decoupling concept [23]. Here, the amplitude of the leakage depends on the isolation of the circulator and the antenna matching. In addition, the leakage signal includes scattered signal components from static scatterers in the propagation environment as well as static reflections due to the tag antenna’s structural mode and antenna mode (see Sec. 1.1.1). In a practical system, the carrier leakage can be 65 dB to 90 dB stronger than the backscattered signal [23], making it necessary to estimate its extent and subtract it from the received signal, e.g., by the use of a carrier compensation unit [24].

Figure 1.5: Link budget of the backscatter radio channel
1.1 Backscatter RFID System

Baseband Signal Constellations at the Reader Receiver

At the RX part of the baseband processor, the received signal is downconverted to
the baseband. Fig. 1.7 shows an exemplary signal constellation in the baseband’s
inphase (I) and quadrature (Q) plane of the reader RX [23]. If the tag is absorbing
energy, the reader finds the tag’s absorb state in its I/Q plane, $S^{(A)}$, which is
essentially the carrier leakage, $L$ [23]. If the tag starts to transmit information to
the reader, the reader encounters the tag’s reflect state, $S^{(R)}$. The tag’s signal, $h$, adds
up with the carrier leakage, $L + h$ [23]. $h$ depends on the channel transfer functions,
$S_{21}$ and $S_{12}$, and on the modulation behavior of the tag which is determined by
$S_{\text{Abs}}$ and $S_{\text{Ref}}$ [23].

Digital reader RX architectures are capable of estimating the two states, $S^{(A)}$
and $S^{(R)}$, by employing a carrier cancelation and a channel estimation [23]. Such
RXs feature sophisticated detection algorithms at the reader and allow for example
the detection of multiple tags during a collision [23].
Figure 1.7: Exemplary baseband constellation at the RFID reader RX [23]
1.2 Scope of Work

This thesis examines the use of backscatter RFID systems in novel wireless sensing applications. Particular attention is paid to the tag antenna design of a wheel unit (WU) in an advanced tire monitoring system (ATMS), the performance of an on-body RFID system for a wireless body area network (WBAN), and the design of a transducer for a backscatter bend sensor to monitor the curvature of an object.

As stated earlier, it is vital to ensure a reliable power transfer and wireless communication between the reader and the tag. For example, if the power at the tag’s chip is smaller than the chip’s sensitivity, the backscatter communication system is limited in its forward link. If the power at the reader RX is smaller than the RX’s sensitivity, a limitation in the backward link occurs.

Thus, the major design goal in this work is to realize backscatter RFID systems and devices which lead to high system performances. Special care is taken to realize systems and devices which operate efficiently in a car tire and on the human body whereas in the case of the backscatter sensor, the focus is on low power consumption and an acceptable sensitivity to bend.

1.3 Outline and Related Work

The main contributions of this thesis comprised in Chap. 2, Chap. 3, and Chap. 4 are briefly summarized in this section.

1.3.1 Transponder Antenna for Car Tire Monitoring

Chap. 2 deals with the design of a tag antenna at 864 MHz for the WU of an ATMS based on backscatter RFID. One premise of the application is that the antenna is directly attached to a car tire. Thus, the antenna’s surroundings and proximity effects caused by the tire environment are explored. Measurements of the dielectric properties of the tire rubber show that the material parameters vary strongly with type and vendor of the tire. This fact leads to a detuning of the WU antenna which is rather difficult to predict. Consequently, a tag antenna is prototyped which can cope with a change in the tire environment. The prototype is power-matched to the tag’s chip over a wide range of frequencies. Further analyses show that the antenna prototype assures a stable power supply to the tag’s chip and leads to a good performance in an ATMS based on passive UHF RFID. In addition, another proximity effect due to the car tire — the distortion of the antenna’s radiation pattern — is beneficially exploited and leads to a further enhancement in the tag’s performance.
1 Introduction

Related Work
Previous research has focused on the design of WU antennas for classical tire pressure monitoring systems (TPMSs) without the use of backscatter RFID. For example, Brzeska et al. [25] studies and characterizes a WU with an integrated antenna mounted on the valve of a car tire at 434 MHz. Investigations of a small loop antenna operating at 315 MHz, which is mounted on the rim of a car tire, are presented by Tanoshita et al. [26], while Teranishi et al. [27] provides simulations of a helical WU antenna mounted on the rim of a truck tire.

Car tire monitoring based on a backscatter RFID system has received less attention in the literature. Lasser and Mecklenbräuker [28, 29] characterize the propagation channel of an ATMS at 868 MHz and 2.45 GHz. A detailed investigation of such a system in terms of its backscatter performance in the forward link using dual-antenna WU RFID tags is done by Lasser et al. [30]. Basat et al. [31] present passive RFID tag antennas at 915 MHz which are designed for inventory purposes of car tires. Peyerl [32] investigates in his diploma thesis tire structures and characterizes the material properties of tire rubbers at UHFs for the purpose of a proper tag antenna design.

1.3.2 Communication System for Remote Health Monitoring
Chap. 3 investigates an on-body backscatter RFID system for remote health monitoring applications in the UHF range at 900 MHz and 2.45 GHz. The investigation is done for two different on-body antenna types. Monopole antennas act as a best-case reference, while less efficient patch antennas are used to give insight into practical RFID system implementations. The antennas are designed by means of a body model to account for proximity effects that are caused by human tissue. The system performance is evaluated in a realistic indoor scenario through on-body channel measurements. The evaluation provides outage probabilities for the system’s forward and backward links. These probabilities help to identify limitations in the backscatter system and to evaluate strategies to overcome these barriers for the realization of a reliable on-body RFID system. An analysis of a state-of-the-art system example shows that the use of semi-passive chips leads to a reliable performance in the system’s forward link. A strategy to overcome limitations in the system’s backward link is to use a phase-modulated backscatter signal.

Related Work
Previous studies on UHF RFID based WBANs have focused on in-body and off-body communication systems. This classification of WBANs has been introduced by Hall and Hao [33].

For example, Occhiuzzi et al. [34] investigate backscatter sensor tags implanted in the human body and their communication to an external reader for vascular
monitoring applications. Sani et al. [35] compare passive and active RFID tags implanted in the human body for tracking and identification purposes, while Schmidt et al. [36] concentrate on the antenna design of implanted RFID tags.

RFID sensor tags which are mounted on the human body for the monitoring of sleep diseases are explored by Occhiuzzi et al. [37]. Backscatter identification of people based on on-body backscatter tags is the driving force behind the research of Polivka et al. [38]. Cotton et al. [39] explore the backscatter communication links between an active, wrist-worn tag and four readers in an indoor environment. On-body tag antenna designs for identification and tracking applications are presented by Kellomäki et al. [40], by Rajagopalan and Rahmat-Samii [41], and by Ziai and Batchelor [42].

So far, the investigation of backscatter communication systems on the human body has received less attention in the literature. A first feasibility study of on-body backscatter RFID systems is presented by Manzari et al. [43, 44] and is based on a backscatter measurement at 870 MHz. The RFID system consists of a short range reader composed of a patch antenna and five on-body tags consisting of custom-built wearable felt antennas.

1.3.3 Sensor for Curvature Monitoring

Chap. 4 provides a design for a backscatter RFID sensor tag that monitors the varying curvature of an object at 5.8 GHz. The backscatter sensor includes a transducer, which changes its impedance as a function of bending, in the tag antenna’s load and thus directly modulates the carrier signal sent from the RFID reader. In addition, the transducer, which acts as the chip’s reflecting impedance, assures a stable power supply to the chip’s circuitry and benefits the detection at the RFID reader. A microstrip line resonator is prototyped for integration in such a backscatter bend sensor. The resonator’s backscatter transducer efficiency is evaluated and optimized with respect to the sensor’s sensitivity to bend and with respect to the sensor tag’s modulation efficiency. It is found that the prototype qualifies for integration in the sensor tag and assures a good RFID system performance.

Related Work

Previous research on the integration of sensing abilities in backscatter RFID tags, without the use of additional RF circuitry, has focused on the use of antenna transducers, i.e., the tag’s antenna acts as the transducer. For example, Capdevila et al. [45] present the use of passive RFID tags for continuous monitoring of the water temperature and discrete monitoring of liquid levels. Bhattacharyya et al. [46] present tag antenna-based sensing of temperature thresholds, displacements, and liquid levels. Siden et al. [47] present another antenna transducer which is used to sense the moisture content in walls. Caizzone et al. [48] realize an antenna trans-
ducer by using a shape memory alloy as part of the tag’s antenna to detect a
temperature threshold violation. An approach to realize an antenna transducer de-
sign for gas detection using single-walled carbon nanotubes (SWCNTs) is presented
by Occhiuzzi et al. [49].

The chip transducer approach has received less attention in the literature. Tentzeris
and Nikolaou [50] present a chip transducer design for a chipless RFID tag which
uses a film of SWCNTs as the antenna load to detect toxic gas. Another chipless
RFID tag to sense gas is presented by Balachandran et al. [51, 52].
2 Transponder Antenna for Car Tire Monitoring

Car tire monitoring is an important safety feature in modern vehicles and strongly improves the reliability of tires and tire control systems [53, 54]. As a consequence, a lot of effort is put into the development of ATMSs which upgrade conventional TPMSs [55] by additionally acquiring sensor data like tire temperature, contact area, vertical load, slip angle, and street conditions.

An ATMS is composed of sensor nodes — WUs — mounted in each tire and a control unit — the on-board unit (OU) — located in the car body. Backscatter RFID in the UHF range around 900 MHz is a promising communication technology to power up the WUs, read out the sensor data, and support lifecycle management and identification of the tires.

In conventional TPMSs, the WU and its antenna is integrated into the valve or is mounted on the rim of the wheel. In ATMSs, the WU must directly contact the tire tread to enable the measurement of additional sensor data. Thus, one of the challenges in realizing backscatter UHF RFID in an ATMS is the design of efficient tag antennas which are directly attached to the tires. An antenna in the complex environment of a car tire will be influenced by this surroundings.

This chapter deals with the design of a tag antenna for a WU of an ATMS based on backscatter RFID at 864 MHz. Sec. 2.1 explores the antenna’s environment as well as the proximity effects caused by the tire. In Sec. 2.2, the tag antenna’s requirements are listed and an appropriate antenna prototype is designed. Sec. 2.3 analyzes the tag’s performance in terms of varying tire environments and the system’s forward link using a passive tag.

Original Publications Related to This Chapter


2 Car Tire Monitoring


2.1 Tire Environment

It is important to have a thorough knowledge of the tire’s structure and its material properties to evaluate the influence of the tire on the tag antenna. Thus, the structure of a modern radial car tire is investigated. Its construction is shown in Fig. 2.1. The tire is composed of multiple rubber layers as well as metal and fiber reinforcements. These components ensure the main functions of the tire: load carrying and transfer of acceleration, deceleration, and lateral forces [56].

From an RF point of view, the tire can be divided into two sections, the tire sidewalls and the tire tread [25]. The sidewalls are composed of different rubber layers, a carcass which is mainly composed of fiber reinforcements, and tire beads with steel cores to anchor the tire on a rim. In contrast, the tread includes different rubber layers, the carcass, and a steel belt. The steel belt is made of two parallel layers of steel wires, in each layer the wires draw an angle of plus and minus 25° (depending on the tire type) with respect to the direction of motion.
2.1 Tire Environment

2.1.1 Dielectric Properties of Tire Rubbers

The different rubber layers of the tire can be considered as lossy dielectrics. A dielectric material is characterized by a complex relative permittivity [58],

$$\tilde{\varepsilon}_r = \varepsilon_r - j\varepsilon''_r,$$  \hspace{1cm} (2.1)

at a certain frequency. In the dielectric material, the arrangement of constituent atoms and molecules, i.e., the orientation of their charges, is changed as a reaction to an external electric field. A measure of the amount of polarization that occurs for the applied field is the relative permittivity, $\varepsilon_r$. $\varepsilon''_r$ is a measure of the losses in the material due to the friction associated with the changing polarization and the drift of conduction charges. The loss tangent, $\tan(\delta)$, is defined as [58]

$$\tan(\delta) = \frac{\varepsilon''_r}{\varepsilon_r},$$  \hspace{1cm} (2.2)

and characterizes the relative loss in the dielectric, which is the ratio of the energy lost and the energy stored within one cycle of the RF field.

Open-Ended Coaxial Probe Technique

The dielectric properties of each rubber layer of the tire are measured by means of an open-ended coaxial probe technique. Here, an open-ended coaxial probe is...
pressed against the rubber under test (RUT). The field of the probe fringes into the dielectric material and is measured by means of a vector network analyzer (VNA). The measured reflection coefficient, $S_{11}$, at the end of the coaxial probe leads to the RUT’s material parameters [59].

The open-ended coaxial probe technique is a broadband testing method [60] and is non-destructive in the sense that it does not require a specimen to be shaped to fit a particular sample holder geometry. This is an advantage in the case of the car tire because the different rubber layers of the tire are difficult to separate.

To perform the method, the following preconditions must be fulfilled [61]:

- The surface of the material has to be flat. Thus, a piece of the tire is cut out and polished to obtain a flat surface (see Fig. 2.7).
- The material has to have semi-infinite thickness. This means that the rubber sample must be thick enough to appear infinite to the probe’s field. A simple practical approach to verify this requirement is to check if the measurement results are affected by the presence of metal at the material’s perimeter [61].
- Another precondition is that there must not be air gaps between the probe and the RUT. Thus, the probe is fixed in a micromanipulator (see Fig. 2.2) and carefully pressed against the RUT till a stable measurement result is realized.
- The RUT has to be non-magnetic. It is assumed that the magnetic response of the tire rubber is very weak and thus negligible.
- The RUT has to be homogenous and isotropic. The RUT is assumed to be linear, homogenous, and isotropic.

The measurement setup with the micromanipulator, the coaxial probe, and the tire specimen can be seen in Fig. 2.2. The reflection coefficient, $S_{11}$, is measured at room temperature with Rohde & Schwarz’s ZVA24 VNA versus a frequency range of 100 MHz - 5 GHz. The VNA is connected to the 50 Ω coaxial probe via a test cable and is calibrated by means of short, open, and matched load standards, while the electrical length of the coaxial probe is de-embedded by means of an electrical short [62].

There are two basic approaches to derive the complex relative permittivity, $\varepsilon_r$, from the measurement results [63]. The first approach uses an equivalent circuit of the probe’s fringing field [59]. The second approach calculates the probe’s field by numerically solving Maxwell’s equations [64]. Here, the latter one is used which leads to the benefit that no reference materials are needed [59]. In particular, $\varepsilon_r$ is obtained by solving the electromagnetic field using Ansys’ HFSS [65].

An HFSS model is designed which is practically equivalent to the measurement setup. The model is shown in Fig. 2.3. The coaxial probe is composed of an inner conductor with a radius of 0.65 mm, a Teflon substrate with a thickness of
2.1 Tire Environment

1.35 mm, and an outer conductor with a thickness of 0.35 mm. The RUT is modeled as a cylinder with a radius and a height of 10 mm. Simulations show that these dimensions are large enough to contain the fringing field in the dielectric. The material’s dielectric properties are defined by $\varepsilon_r$ and $\tan(\delta)$. A wave port is used to excite the structure. The blue arrow in Fig. 2.3 denotes the de-embedding of the port to the end of the coaxial probe. As with the measurement, the port impedance is 50 $\Omega$. In addition, an air box surrounds the whole simulation setup to avoid perfectly conducting boundaries at the tire material [66].

The effect of the RUT’s material parameters on the reflection coefficient is illustrated in Fig. 2.4 by means of simulations. The reflection coefficient, $S_{11}$, is plotted in a Smith chart versus frequency for different material parameters. The Smith chart relates the reflection coefficient, $S_{11}$, to an impedance, $Z = R + jX$, by [7]

$$S_{11} = \frac{Z - Z_0}{Z + Z_0},$$

where the characteristic impedance, $Z_0$, is in this case 50 $\Omega$. Starting at a frequency of $f = 100$ MHz, the fringing capacitance at the end of the coaxial probe is $C \approx 39$ fF for $\varepsilon_r = 1$ and $\tan(\delta) = 0$. This capacitance, $C$, corresponds to a reactance of $X = -1/(2\pi f C) = -40809 \Omega$ [67] which is close to the open circuit point in the Smith chart: $Z = \infty$, or rather $S_{11} = 1$. An increase of the relative permittivity — while disregarding losses in the dielectric material ($\tan(\delta) = 0$) — as well as an increase in frequency lead to an increase of the coaxial probe’s end reactance.
Because there are no losses in the dielectric, the reflection coefficient remains at the unit circle of the Smith chart (see Fig. 2.4, $\varepsilon_r = 5$, tan($\delta$) = 0). For tan($\delta$) > 0, losses increase with the frequency and can be identified by a deviation of $S_{11}$ from the unit circle (see Fig. 2.4, $\varepsilon_r = 15$, tan($\delta$) = 0.1).

In the next step, the simulated reflection coefficient, $S_{\text{Sim}}$, is fitted to be consistent with the measured reflection coefficient, $S_{\text{Meas}}$. During a sequence of simulation runs the dielectric properties, $\varepsilon_r$ and tan($\delta$), of the RUT are varied until the simulated and measured reflection coefficients at a specific frequency are equal within a certain tolerance. A comparison of simulation and measurement results, $S_{\text{Sim}}$ and $S_{\text{Meas}}$, of one exemplary RUT can be seen in Fig. 2.5. The reflection coefficients are fitted for a frequency of 866 MHz. The figure indicates that there are losses in the RUT which increase with frequency. The absolute error of simulated and measured reflection coefficients, $|S_{\text{Sim}} - S_{\text{Meas}}|$, of this exemplary RUT is plotted in Fig. 2.6. The figure shows that there is only a small deviation in reflection coefficients (< 0.04) versus frequency. This deviation is caused by a dispersive behavior of the tire material [68] which is not considered in the simulation model. As expected, the smallest error occurs at about 866 MHz.

Figure 2.3: HFSS design of the measurement setup: The enhanced illustration of the open-ended coaxial probe on the right gives an insight in the probe’s structure.
2.1 Tire Environment

Figure 2.4: Reflection coefficient, $S_{11}$, versus frequency (100 MHz – 5 GHz) for different material parameters displayed in a Smith chart.
Figure 2.5: Comparison of the simulated and measured reflection coefficients, $S_{\text{Sim}}$ and $S_{\text{Meas}}$, of one exemplary RUT versus frequency (100 MHz – 5 GHz) displayed in a Smith chart: The reflection coefficients are fitted for a frequency of 866 MHz.
Figure 2.6: Absolute error of the simulated and measured reflection coefficients, $|S_{\text{Sim}} - S_{\text{Meas}}|$, of one exemplary RUT versus frequency (100 MHz – 5 GHz): The reflection coefficients are fitted for a frequency of 866 MHz.
Measurement Results

Using the open-ended coaxial probe technique, the individual rubber layers of a standard tire (Continental Premium Contact 2 205/55 R16 91V) and a run-flat tire (Continental Premium Contact SSR 205/55 R16 91H) are characterized. The run-flat tire has reinforced sidewalls (see Fig. 2.8) which allow to continue driving uninflated for a limited distance and low speed [69].

The investigated tire specimens and its material parameters can be seen in Fig. 2.7 and Fig. 2.8. The single rubber layers are distinguished from each other by their varying gray scales and different grit sizes.

The relative permittivity, $\varepsilon_r$, and the loss tangent, $\tan(\delta)$, are evaluated for 866 MHz. It is found that the rubber materials used in the different layers of a car tire vary strongly. Some layers show very high values of permittivity and loss tangent. For the standard tire, dielectric permittivities between 1.5 and 6.1 and loss tangents between 0.003 and 0.17 are determined (see Fig. 2.7). For the run-flat tire, dielectric permittivities between 3.9 and 6.9 and loss tangents between 0.04 and 0.23 are found (see Fig. 2.8). Additionally, it is discovered that the dielectric properties vary strongly with the type and vendor of the tire. Thus, a wide range of material parameters has to be considered for the ATMS tag antenna design.

Figure 2.7: Dielectric material parameters, $\varepsilon_r$ and $\tan(\delta)$, of a standard car tire at 866 MHz (tire tread on the left, tire sidewall on the right)
2.1 Tire Environment

2.1.2 Proximity Effects

The tire — with its metal reinforcements and dielectric rubber layers — and the wheel rim are going to influence the performance of the WU antenna. To explore the proximity effects induced by the standard tire, the characteristics of a half-wave dipole directly attached to the tire rubber are investigated. The geometry of the lightweight, brass dipole is sketched in Fig. 2.9. It has a length of $l = 112$ mm, a width of $w = 10$ mm, and a feed gap length of $g = 2$ mm. The dipole is mounted on the thinnest part of the standard tire’s inner sidewall which is favorable with respect to the antenna’s performance (see Sec. 2.2.1).

![Figure 2.8: Dielectric material parameters, $\varepsilon_r$ and $\tan(\delta)$, of a run-flat car tire at 866 MHz (tire sidewall)](image)

![Figure 2.9: Top view of the planar dipole antenna with a length, $l$, width, $w$, and feed length, $g$: The antenna’s feed is denoted by two points.](image)
Proximity effects induced by tire might affect the dipole’s resonant frequency and thus its input impedance. This effect can be observed by investigating the input impedance of the dipole in free space and attached to the tire sidewall.

The dipole’s input impedance, $Z$, is measured by means of Rohde&Schwarz’s ZVA24 VNA. A differential input signal at the dipole’s symmetric input is applied to overcome errors due to common mode currents that will cause radiation and alter the antenna properties. Fig. 2.10 shows an exemplary measurement setup with a test antenna. The antenna is fed by two flexible 50 Ω coaxial cables. The cables are connected to the antenna via a small, custom-built feeding structure on FR-4 substrate with miniature coaxial connectors (U.FL series by Hirose Electric Co. Ltd.). The differential signal is generated by the VNA. A custom-built calibration kit [70] made of FR-4 substrate — including match, short, open, and through standards — is used to shift the reference plane of the measurement directly to the input of the antenna.

It is found that this differential measurement method works quite well for antennas on a substrate, while measurements of antennas on an air substrate are impaired by the feeding structure due to capacitive coupling between the antenna and the metal parts of the feed. Thus, the input impedance of the dipole attached to the tire rubber is obtained by measurements, while the impedance curve of the dipole in free space is obtained by simulations using HFSS.

In Fig. 2.11, the real and imaginary parts of the dipole’s input impedance, $Z = R + jX$, are plotted versus frequency, $f$. The figure shows the comparison of the dipole’s impedance in free space and attached to the inner tire sidewall. The resonances — half-wave and full-wave resonances — of the dipole can be found at the frequencies where the reactance, $X$, is zero [8]. It can be seen from the figure that there is a shift to lower resonant frequencies due to the tire.

This effect can be also seen in Fig. 2.12. The magnitude of the reflection coefficient at the dipole’s input, $|S_{11}|$, in decibel is plotted versus frequency, $f$. $|S_{11}|^2$ characterizes the dipole’s matching to the 50 Ω feed (see Eqn. 2.3, with $Z_0 = 50$ Ω) and is defined as the ratio of the power reflected to the power available at the antenna’s input. A frequency shift of about 460 MHz can be observed in the dipole’s matching due to the tire. In addition, the dipole’s bandwidth increases. Here, the bandwidth is defined as the frequency range with an antenna matching smaller than $-3$ dB. The increase is due to a higher power absorption in the lossy rubber material.

In general, the resonant frequency of the dipole when attached to the tire sidewall is strongly shifted towards lower frequencies. On the one hand, this impedance detuning is due to the dielectric rubber of the tire with $\varepsilon_r > 1$ [8]. On the other hand, the dipole’s impedance is affected by the metal reinforcements of the tire and
the rim in the reactive near field of the antenna.

Figure 2.10: Setup for the input impedance measurement of a symmetric test antenna: The calibration standards (match, short, through, and open), the antenna feed, and the measurement cables leading to the VNA are shown.
Figure 2.11: Dipole’s input impedance, $Z = R + jX$, versus frequency in free space and attached to the tire sidewall

Figure 2.12: Magnitude of the dipole’s reflection coefficient, $|S_{11}|$, in decibel versus frequency in free space and attached to the tire sidewall


Radiation Pattern and Efficiency

Other proximity effects might influence the antenna’s radiation pattern and radiation efficiency. For a detailed analysis of these effects, the radiation pattern and efficiency of the dipole are measured while it is attached to the tire sidewall.

The measurement setup is depicted in Fig. 2.13. The measurement method is outlined in [9]. For the investigation, a standard tire specimen is mounted on a rotation unit. Again, the dipole is attached to the thinnest part of the inner tire sidewall, parallel to the tire tread. A small battery-driven oscillator is connected to the dipole and provides a sinusoidal transmit signal at 864 MHz. The oscillator is equipped with a tunable matching network which allows to realize perfect power matching to the antenna. In addition, the radiation efficiency of the dipole can be determined. The dipole’s radiation is measured by a pick-up antenna which consists of two cross-polarized dipole antennas to measure the power for linear polarization in the $\vartheta$-direction and $\varphi$-direction of a spherical coordinate system.

The transmission distance between the dipole and the pick-up antenna is $d = 0.95\text{ m}$ sufficient to be in the far field. That is, $d$ is bigger than the Rayleigh distance [8], $r_R = 2D^2/\lambda = 0.92\text{ m}$, where $D = 0.4\text{ m}$ is the maximal lateral dimension of the pick-up antenna and $\lambda = 0.35\text{ m}$ is the wavelength at 864 MHz. A measurement computer controls the rotation unit and the rest of the setup and automatically samples a series of power values at the surface of a sphere with diameter, $d$. The measurements are taken at pre-defined angle increments of $\vartheta$ and $\varphi$. From these power values, the directional pattern and the radiation efficiency are derived.

The spherical coordinate system of the antenna is depicted in Fig. 2.14. Note that there is a slight parallel offset between the rotation axes of the tire specimen and the antenna’s coordinate system. The influence of this spatial offset (about 7 cm) has been calculated in terms of antenna gain and turned out to be negligible.

The observed radiation pattern is plotted in Fig. 2.15. The graph shows the square root of the measured antenna gain in $\vartheta$-polarization compared to an isotropic radiator, $G_{\text{ISO,}\vartheta}$, normalized to the maximum gain, $G_{\text{Max}}$,:

$$F_\vartheta(\vartheta, \varphi) = \sqrt{\frac{G_{\text{ISO,}\vartheta}(\vartheta, \varphi)}{G_{\text{Max}}}},$$

(2.4)

where $G_{\text{Max}}$ is defined as

$$G_{\text{Max}} = \max_{\vartheta, \varphi} (G_{\text{ISO,}\vartheta} + G_{\text{ISO,}\varphi})(\vartheta, \varphi).$$

(2.5)

The dipole attached to the tire shows a maximum gain of 4.21 dBi in the $\vartheta = -90^\circ$ and $\varphi = 280^\circ$ direction. It can be seen that most of the power is radiated outwards through the tire sidewall and that the power is maximal in the plane perpendicular to the dipole. In comparison, a half-wave dipole in free space has a gain of 2.15 dBi.
and is omnidirectional at \( \vartheta = 90^\circ \), i.e., in the azimuth plane [8]. The high gain of the dipole attached to the tire sidewall is due to a strong directional effect caused by the tire’s metal reinforcements. The dipole’s efficiency is about 81\% when attached to the tire rubber. In comparison, the half-wave dipole in free space has an efficiency of about 100\% [8]. The reduction in radiation efficiency is primarily due to dielectric losses in the rubber material with a loss tangent greater than zero (see Sec. 2.1.1).

Additionally, Fig. 2.15 shows the radiation pattern of the dipole mounted on the tire with a wheel rim. The rim is emulated by means of an aluminum foil. The radiation pattern shows that the directional effect due to the tire is further increased by the rim. The dipole then has a maximum gain of 4.51 dBi in the \( \vartheta = -90^\circ \) and \( \varphi = 280^\circ \) direction. The radiation efficiency is reduced to about 73\% which is a consequence of a detuning of the antenna by the rim.

Further measurements of the dipole attached to the inner tire sidewall verify that the directional effect occurs due to the metal reinforcements of the tire. Fig. 2.16 shows the dipole’s radiation pattern when it is mounted on a modified tire specimen. The tire bead which is farthest away from the dipole is removed. In comparison to the measurement of the original tire specimen, only a small change in the radiation pattern can be noticed. A big change in the radiation pattern can be observed when both beads are removed (see Fig. 2.17). As a consequence, the direction of the maximum gain is shifted from \( \vartheta = -90^\circ \) and \( \varphi = 280^\circ \) to \( \vartheta = -90^\circ \) and \( \varphi = 340^\circ \). Fig. 2.18 shows the radiation pattern of the dipole mounted on the tire sidewall without any influence of the tire’s metal reinforcements (without beads and without the steel belt). It can be seen that the pattern resemble the pattern of a dipole in free space. Variations from a perfect omnidirectional radiation characteristic occur due to the complex rubber material. Then, the maximum gain is \( G_{\text{Max}} = 0.68 \text{ dBi} \) at \( \vartheta = -90^\circ \) and \( \varphi = 80^\circ \).

In summary, it can be stated that the dipole’s radiation pattern and efficiency are significantly affected by the tire. The radiation pattern shows pronounced directivity and is favorably influenced by the tire’s metal reinforcements. This is also verified by HFSS simulations of the dipole attached to the sidewall of a complete tire model including the rim. In addition, the radiation efficiency of the dipole is decreased due to the lossy rubber material.
2.1 Tire Environment

Figure 2.13: Setup for the measurement of the dipole’s radiation pattern and efficiency: The tire specimen with the dipole is mounted on the rotation unit, while the pick-up antenna receives the power transmitted in the sphere with a diameter of $d = 0.95$ m. The dipole’s position is denoted by the white bar.

Figure 2.14: Spherical coordinate system of the dipole antenna attached to the sidewall of the tire specimen
Car Tire Monitoring

Figure 2.15: Comparison of radiation patterns, $F(\vartheta,\phi)$, of the dipole mounted on the inner tire sidewall with and without the rim at 864 MHz. The coordinates are defined in Fig. 2.14.
2.1 Tire Environment

Figure 2.16: Comparison of radiation patterns, $F_{\theta}(\theta, \varphi)$, of the dipole mounted on the inner tire sidewall with and without the bead which is farthest away from the dipole at 864 MHz. The coordinates are defined in Fig. 2.14.
with and without beads at 864 MHz. The coordinates are defined in Fig. 2.14.

Figure 2.17: Comparison of radiation patterns, \( P(\theta, \phi) \), of the dipole mounted on the inner tire sidewall with and without beads at 864 MHz. The coordinates are defined in Fig. 2.14.

The coordinates are defined in Fig. 2.14.
Figure 2.18: Radiation pattern, $F_{\theta}(\theta, \phi)$, of the dipole mounted on the inner tire sidewall without metal reinforcements at 864 MHz. The coordinates are defined in Fig. 2.14.
2.2 Antenna Prototype

Sec. 2.1.2 shows that an antenna is considerably influenced by a car tire. In particular, if the antenna is mounted directly on or in the tire rubber, the proximity effects are strong. Here, the design goal is to create an efficient WU antenna with best performance in the complex environment of the tire. To do this, the antenna designer must take into consideration the above mentioned proximity effects to realize an antenna which suits the environment.

2.2.1 Antenna Requirements

A UHF RFID tag antenna practical for the WU of an ATMS should fulfill the following requirements:

- The WU antenna should show a strong radiation in the direction to the OU antenna. In an ATMS, the OU ideally uses a single antenna to communicate with all four WUs which is placed on the bottom of the car in the middle of the baseplate (see Sec. 2.3.2).
- The antenna should show a high radiation efficiency. Thus, conduction losses due to a complex antenna structure should be prevented. Losses due to the dielectric tire rubber cannot be completely avoided. Although, the mounting of the antenna on the thinnest part of the tire sidewall is beneficiary because of smaller dielectric losses due to the thin sidewall rubber.
- Additionally, the antenna should be power matched to the WU’s microchip. The impedance matching of the tag antenna and chip strongly influences the communication performance between the OU and WU (see Eqn. 1.6). Since the input impedance of the chip cannot be chosen arbitrarily due to technological limits, the antenna has to be designed to match the chip’s impedance. Thus, the antenna impedance should be the complex conjugate of the chip’s absorbing impedance (see Eqn. 1.2). Here, the tag antenna is designed for an absorbing impedance of \( Z_{Abs} = (68 - j442) \, \Omega \) at 864 MHz (see Fig. 1.2).
- The polarization of the WU antenna should be ideally the same as the polarization of the OU antenna, which is — in the investigated ATMS (see Sec. 2.3.2) — vertically polarized with respect to the car’s baseplate. This requirement is not easily met because of the rotation of the tire.
- Mechanical requirements are that the antenna should be lightweight and low profile to avoid big unbalanced masses in the tire. In addition, the dimensions of the WU antenna should be around the size of the thinnest part of the tire sidewall to assure small dielectric losses.

An antenna based on a dipole structure fulfills most of the requirements listed above. First, if a dipole-based antenna is mounted directly on the thinnest part
of the inner tire sidewall, parallel to the tire tread, the radiation pattern shows a pronounced directivity in the direction of the OU antenna (see Sec. 2.1.2). Second, a dipole operates quite efficient in the tire environment (see Sec. 2.1.2). Third, a dipole-based antenna with a T-feed integrated in its layout [71] can match the high reactive input impedance of the WU chip. Forth, a dipole-based antenna made of a thin brass sheet is lightweight and low profile. However, the polarization of such an antenna which is mounted parallel to the tire tread changes with the tire’s rotation. This drawback is outweighed by the mass of met requirements. Though, the use of a circularly polarized OU antenna can possibly reduce this polarization losses.

2.2.2 T-Matched Dipoles

In the following, tag antenna prototypes are engineered based on a planar dipole with a T-matched feeding structure. The antenna’s geometry can be seen in Fig. 2.19. Different dimensions of the dipole’s length, $l$, width, $w$, and its feeding structure, $b$, $a$, $wis$, and $g$, lead to diverse input impedances.

The prototypes are designed in HFSS to be directly attached to the thinnest part of the inner tire sidewall with a thickness of 5 mm. An average relative permittivity of $\varepsilon_r = 6.2$ and a loss tangent of $\tan(\delta) = 0.1$ models the electromagnetic properties of the sidewall and are found to be appropriate to simulate the detuning due to the sidewall (see Fig. 2.23).

The picture of the first prototype can be seen in Fig. 2.20. Its dimensions are $l = 60$ mm, $w = 10$ mm, $b = 15$ mm, $a = 24$ mm, $wis = 5$ mm, and $g = 2$ mm. The antenna shows an optimized power transmission coefficient of about 91% at 864 MHz and a tag antenna bandwidth of approximately 100 MHz. Here, the bandwidth is defined as the frequency range with a power transmission coefficient greater or equal to $\tau \geq 60\%$.

For comparison, a second T-matched dipole is realized (see Fig. 2.21). Its dimensions are $l = 113$ mm, $w = 10$ mm, $b = 21$ mm, $a = 20$ mm, $wis = 5$ mm, and $g = 2$ mm which are optimized for a wide tag antenna bandwidth of approximately 200 MHz. The wide impedance bandwidth is realized by a feeding geometry which shows double tuned matching to the chip’s impedance [72]. However, an increase in bandwidth comes at the cost of the maximum achievable power transmission coefficient. $\tau$ is about 63% at 864 MHz.

Fig. 2.22 plots measurement results of the antenna’s input impedance, $Z_{\text{Ant}} = R_{\text{Ant}} + jX_{\text{Ant}}$, versus frequency, $f$, for both antenna prototypes. The impedances are measured by means of the differential method presented in Sec. 2.1.2. The simulated and measured power transmission coefficients of both antennas are plotted in Fig. 2.23. It can be seen that the simulation and measurement results fit quite well versus frequency. The chip’s absorbing impedance, $Z_{\text{Abs}} = (68 - j442)$ $\Omega$, is
assumed to be constant versus frequency.

Figure 2.19: Top view of the planar T-matched dipole with parameters \( l \), \( w \), \( b \), \( a \), \( wis \), and \( g \): The antenna’s feed is denoted by two points.

Figure 2.20: Prototype of the narrowband tag antenna mounted on the inner tire sidewall
2.2 Antenna Prototype

Figure 2.21: Prototype of the broadband tag antenna mounted on the inner tire sidewall

Figure 2.22: Measured input impedance, $Z_{\text{Ant}}$, versus frequency of the narrowband and broadband antennas directly attached to the tire sidewall
Figure 2.23: Simulated and measured power transmission coefficients, $\tau$, versus frequency of the narrowband and broadband antennas directly attached to the tire sidewall: The chip’s absorbing impedance, $Z_{\text{Abs}} = (68 - j 442) \, \Omega$, is assumed to be constant versus frequency.
2.3 Transponder Performance

As stated earlier, the matching of the RFID tag antenna and the chip strongly influences the wireless power transmission and the communication performance between the OU and WU. In particular, passive RFID systems suffer from low power transmission coefficients. In this section, the antenna prototypes are evaluated in terms of their performance in passive WU tags.

2.3.1 Change in Tire Environments

The investigation in Sec. 2.1.1 shows that the dielectric properties of tire rubbers strongly vary with the type and vendor of the tire. These properties lead to unpredictable detuning effects of the antenna’s resonance when it is attached to different types of tires and consequently influences the matching between the antenna and chip. Thus, the tag antenna prototypes are investigated in terms of their tolerance to varying proximity effects due to different rubber environments.

It is assumed that both tags with the narrowband and broadband antenna prototypes are attached to different types of tires which are represented by different values of $\varepsilon_r$ and tan($\delta$). In Fig. 2.24 and Fig. 2.25, simulation results of the tags’ transmission coefficients are plotted for different rubber environments at 864 MHz. It can be seen that detuning due to different environments leads in the majority of cases to a decrease in $\tau$. In particular, the narrowband antenna prototype is vulnerable to detuning effects. For example, the narrowband tag antenna reaches $\tau = 91\%$ for a rubber environment of $\varepsilon_r = 6.2$ and tan($\delta$) = 0.1, while a power transmission coefficient of merely 10% can be achieved when the antenna is attached to a tire environment of $\varepsilon_r = 3$ and tan($\delta$) = 0.1 (see Fig. 2.24). In comparison, the broadband antenna achieves nearly 50% in the latter environment (see Fig. 2.25) and is thus more suitable for the ATMS application.

In general, to achieve best transponder performance, the WU antenna which will be attached to various tires should be maximized in bandwidth to grant better robustness to detuning effects. Another benefit of the broadband antenna is its robustness to detuning due to mechanical deformation of the tire or due to different sidewall thicknesses.
Figure 2.24: Simulated power transmission coefficient, \( \tau \), versus relative permittivity, \( \varepsilon_r \), and loss tangent, \( \tan(\delta) \), of the tire rubber at 864 MHz for the narrowband antenna prototype: Exemplarily, the points, \( \varepsilon_r = 6.2 \) and \( \varepsilon_r = 3 \), at \( \tan(\delta) = 0.1 \) are highlighted.
Figure 2.25: Simulated power transmission coefficient, $\tau$, versus relative permittivity, $\varepsilon_r$, and loss tangent, $\tan(\delta)$, of the tire rubber at 864 MHz for the broadband antenna prototype: Exemplarily, the points, $\varepsilon_r = 6.2$ and $\varepsilon_r = 3$, at $\tan(\delta) = 0.1$ are highlighted.
2.3.2 Advanced Tire Monitoring System

Static radio channel measurements for the ATMS at 868 MHz are presented in [29]. The WU antenna is a dipole and is directly attached to various positions inside the tire sidewalls. The OU antenna is situated in the middle of a car’s baseplate at the bottom of the vehicle and radiates in an omnidirectional, vertically polarized fashion. The measurements validate that the mounting of the dipole on the inner tire sidewall — closest to the OU antenna and parallel to the tire tread — shows the best performance in terms of channel gain. Channel gains of $-30 \text{ dB}$ to $-64 \text{ dB}$ have been observed due to different tire positions.

In the following, the power transfer to the passive RFID tag, which is comprised of the broadband antenna prototype, is roughly estimated. A passive RFID chip has a typical chip sensitivity of $T_{\text{Chip}} = -15 \text{ dBm}$. If the power absorbed by the chip, $P_{\text{Chip}}$, is smaller than the chip’s sensitivity, the backscatter communication is limited in its forward link (see Sec. 1.1.2). Fig. 2.26 plots $P_{\text{Chip}}$ for a channel gain of $|S_{21}|^2 = -30 \text{ dB}$ using Eqn. 1.6 with $P_{\text{TX,Reader}} = 33 \text{ dBm}$. The power transmission coefficient, $\tau$, varies with the rubber environment defined by $\varepsilon_r$ and $\tan(\delta)$. It can be seen that there is no limitation in the forward link of the passive RFID system for a wide range of rubber environments.

In comparison, in the case of a maximum channel loss of 65 dB the system is definitely limited in its forward link. To overcome this limitation, an OU antenna with a directional radiation pattern or circular polarization can be used. In addition, various OU antennas can be placed at shorter distances to the WUs, e.g., one antenna for the communication with the tires in the front of the car and one antenna for the communication with the rear tires. Another possibility to enhance the system performance is to use dual-antenna techniques in the WU tags [30].

In summary, the broadband T-matched dipole proves feasible for the car tire monitoring application. The prototype assures a good tag performance in a passive RFID system especially in the case of a varying tire environment. However, limitation in the forward link can occur due to high channel losses.
Figure 2.26: Power absorbed by the tag chip, $P_{\text{Chip}}$, when using the broadband antenna prototype versus relative permittivity, $\varepsilon_r$, and loss tangent, $\tan(\delta)$, at 864 MHz: If $P_{\text{Chip}}$ is smaller than the chip’s sensitivity, $T_{\text{Chip}} = -15$ dBm, the backscatter communication is limited in its forward link.
2.4 Summary

This chapter deals with the design of a tag antenna at 864 MHz for the WU of an ATMS based on backscatter RFID. One premise of the application is that the antenna is directly attached to a car tire. Thus, the antenna’s surroundings and proximity effects caused by the tire environment are explored. It is found that the tire with its complex structure has a considerable influence on the antenna parameters like input impedance and gain. Measurements of the dielectric properties of the tire rubber show that its material parameters vary strongly with type and vendor of the tire. This fact leads to a detuning of the antenna’s resonant frequency which is rather difficult to predict. Additionally, the radiation pattern of a dipole antenna, which is attached to the thinnest part of the inner tire sidewall, is favorably influenced by the metal reinforcements of a standard car tire.

Consequently, a tag antenna prototype is realized which can cope with a change in the tire environment. The prototype — a T-matched dipole — is power-matched to the chip over a wide range of frequencies. Further analyses show that the antenna prototype assures a stable power supply to the tag’s chip and thus leads to a good performance in the ATMS using passive WU tags. In addition, the distortion of the antenna’s radiation pattern due to the tire is beneficially exploited and leads to a further enhancement in system performance.

In a next step, the tag antenna’s matching to the chip should be investigated for varying chip input powers which lead to a variation in the chip’s impedance. In general, the concept of a broadband tag antenna — which is exploited in this chapter — proves feasible for a multitude of applications where a change in the environment is foreseeable, e.g., when a tag is attached to the human body.
3 Communication System for Remote Health Monitoring

Remote health monitoring applications are often based on WBANs and support medical experts, patients, and humans with physiological data to improve the quality of the clinical environment. In addition, such applications favor therapy support at home and improve the well-being of people. Because of this, the attention of the industry and the scientific community is highly drawn to the promising field of WBANs [73, 74].

WBANs connect sensor nodes situated in clothes, on the body, or under the skin of a person through a wireless communication channel. Backscatter RFID at UHFs is a promising communication technology for WBANs. RFID sensor tags can be used to monitor the physiological parameters of persons (e.g., blood pressure, temperature, heartbeat, body motion). As an example, the backscatter bend sensor proposed in Chap. 4, which senses an object’s curvature, may be used to monitor the motion sequence of people for sports analysis or for human computer interaction devices when a patient is undergoing physical therapy.

This chapter investigates an on-body backscatter RFID system for remote health monitoring applications at 900 MHz and 2.45 GHz. Sec. 3.1 describes the arrangement of the on-body RFID reader and sensor tags as well as the design of their antennas. Sec. 3.2 deals with the on-body radio channel measurement, which is used to evaluate the system performance in the forward and backward links in Sec. 3.3.

Original Publications Related to This Chapter


3 Remote Health Monitoring


3.1 On-Body System

Fig. 3.1 shows the arrangement of the investigated RFID system situated on the body of an adult female. In particular, the positions of the RFID reader and tags are highlighted. The RFID reader is situated on the stomach of the female, while four RFID tags are placed at various positions on the female’s body: on the right chest, on the middle of the back, on the left side of the head, and on the right wrist. These links represent two trunk-to-trunk, a trunk-to-head, and a trunk-to-limb link following a classification of on-body links introduced in [33].

Figure 3.1: On-body RFID system: The sensor tag on the female’s back is represented by the orange circle.
3.1 On-Body System

3.1.1 Body Model

The antennas of the RFID reader and tags are influenced by the close proximity to the human body. Proximity effects, which are experienced by an antenna attached to the body, include a shift in the antenna’s resonant frequency, a distortion of its radiation pattern, and a reduction in its radiation efficiency [75]. The strength of this effects depends on the antenna type, the antenna’s placement on the body, and the antenna-to-body separation distance.

Previous investigations and measurements of human tissues have shown that the electromagnetic properties vary significantly with tissue type and frequency. These properties — the relative permittivity, $\varepsilon_r$, and the electric conductivity, $\sigma$ — are plotted versus frequency in Fig. 3.2 and Fig. 3.3 for muscle, skin, and fat tissues. The curves are based on a parametric model for the complex relative permittivity of human tissues [76]:

$$\tilde{\varepsilon}_r(\omega) = \varepsilon_\infty + \sum_{m=1}^{4} \frac{\Delta \varepsilon_m}{1 + (j\omega\tau_m)^{1-\alpha_m}} + \frac{\sigma_i}{j\omega\varepsilon_0},$$  \hspace{1cm} (3.1)

where $\omega = 2\pi f$ is the angular frequency, $\varepsilon_\infty$ is the material’s permittivity at terahertz frequencies, $\varepsilon_0 = 8.854\text{pF/m}$ is the free space permittivity, $\sigma_i$ is the ionic conductivity, and $\Delta \varepsilon_m$, $\tau_m$, and $\alpha_m$ are material parameters for each dispersion region. The relative permittivity, $\varepsilon_r(\omega)$, is the real part of the complex relative permittivity as denoted in Eqn. 2.1,

$$\varepsilon_r(\omega) = \Re\{\tilde{\varepsilon}_r(\omega)\}. \hspace{1cm} (3.2)$$

The electric conductivity, $\sigma(\omega)$, is defined by [58],

$$\sigma(\omega) = -\omega \varepsilon_0 \varepsilon''(\omega), \hspace{1cm} (3.3)$$

where $\varepsilon''(\omega) = \Im\{\tilde{\varepsilon}_r(\omega)\}$ is the imaginary part of the complex relative permittivity (see Eqn. 2.1). Tab. 3.1 lists the individual parameters for different body tissues which are needed to calculate $\tilde{\varepsilon}_r(\omega)$ (see Eqn. 3.1). The figures, Fig. 3.2 and Fig. 3.3, show that there is only a minor change in $\varepsilon_r$ at UHFs, while the losses increase with frequency.

Due to this adverse electromagnetic properties of human tissues, the performance of on-body antennas is rather difficult to predict. For an adequate antenna design and computation of the on-body channel, a human body model has been created in HFSS. The body model is composed of two-third muscle equivalent tissue which varies with frequency [58]. $\varepsilon_r$ and $\sigma$ of this tissue type are also plotted in Fig. 3.2 and Fig. 3.3. The body model is realized as a simple rectangular torso to efficiently solve the electromagnetic problem in a reasonable time. Fig. 3.4 shows the HFSS model for the computation of the stomach-chest link using monopole antennas at 900 MHz. The spacing of the monopoles is 30 cm which is approximately the length of the line of sight path on the female’s body.
Figure 3.2: Relative permittivity, $\varepsilon_r$, of muscle, skin, fat, and 2/3 muscle equivalent tissues versus frequency.

Figure 3.3: Electric conductivity, $\sigma$, of muscle, skin, fat, and 2/3 muscle equivalent tissues versus frequency.
### 3.1 On-Body System

<table>
<thead>
<tr>
<th>Tissue</th>
<th>$\varepsilon_\infty$</th>
<th>$\varepsilon_1$</th>
<th>$\varepsilon_2$</th>
<th>$\varepsilon_3$</th>
<th>$\varepsilon_4$</th>
<th>$\Delta\varepsilon_1$</th>
<th>$\Delta\varepsilon_2$</th>
<th>$\Delta\varepsilon_3$</th>
<th>$\Delta\varepsilon_4$</th>
<th>$\tau_1$ (ns)</th>
<th>$\tau_2$ (ns)</th>
<th>$\tau_3$ (ms)</th>
<th>$\tau_4$ (ms)</th>
<th>$\sigma_1$</th>
<th>$\sigma_2$</th>
<th>$\sigma_3$</th>
<th>$\sigma_4$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Muscle</td>
<td>4</td>
<td>50</td>
<td>7.234</td>
<td>0.1</td>
<td>7000</td>
<td>0.1</td>
<td>1.26</td>
<td>1.26</td>
<td>0.1</td>
<td>2.5e-7</td>
<td>2.5e-7</td>
<td>2.5e-7</td>
<td>0</td>
<td>0.2</td>
<td>0.2</td>
<td>0.2</td>
<td>0.2</td>
</tr>
<tr>
<td>Skin</td>
<td>3</td>
<td>32</td>
<td>7.234</td>
<td>0.1</td>
<td>7000</td>
<td>0.1</td>
<td>1.26</td>
<td>1.26</td>
<td>0.1</td>
<td>2.5e-7</td>
<td>2.5e-7</td>
<td>2.5e-7</td>
<td>0</td>
<td>0.2</td>
<td>0.2</td>
<td>0.2</td>
<td>0.2</td>
</tr>
<tr>
<td>Fat</td>
<td>2</td>
<td>5</td>
<td>7.958</td>
<td>0.2</td>
<td>35</td>
<td>0.2</td>
<td>14.915</td>
<td>15.915</td>
<td>0.2</td>
<td>0.002</td>
<td>0.002</td>
<td>0.002</td>
<td>0</td>
<td>0.01</td>
<td>0.01</td>
<td>0.01</td>
<td>0.01</td>
</tr>
</tbody>
</table>
Figure 3.4: HFSS model for the channel computation of the stomach-chest link: Monopole antennas operating at 900 MHz are attached to the human trunk. The length of the line of sight path is 30 cm.
3.1.2 Antennas

The primary requirement for on-body antennas is that the mutual influence between the antenna and the human body is low [77]. This decoupling can be achieved by metallic shields which are integrated in the antennas as groundplanes [37]. Suitable antenna types are for example monopoles or patch antennas, their groundplanes mounted parallel to the body surface.

Monopole Antennas

Practically, a monopole antenna is not suitable for WBAN applications because it is not low profile. However, monopoles show the best performance in on-body systems [75, 77]. A monopole shows an omnidirectional radiation pattern on the body, i.e., a maximum radiation along the body’s surface, and a vertical polarization. Such a radiation is favorable for on-body links where the main mechanism for propagation around the body is via a surface wave [78]. Thus, monopoles are used as a best-case reference in this work and help to define an upper bound for the performance of practical system implementations.

In the following, monopole antennas resonant at 900 MHz and 2.45 GHz are designed by means of the human body model and realized on FR-4 substrate with a thickness of 1.6 mm [75]. The design is done for an antenna-to-body separation distance of 5 mm. Pictures of the realized monopoles and their dimensions are depicted in Fig. 3.5.

The monopoles are matched to a 50 Ω feed and show an overall matching of some $-10$ dB (see Fig. 3.11 and Fig. 3.12) which is sufficient for real-world systems. The planar feed geometry is essential for wearable applications because probe-fed antennas cannot be mounted close to the body surface. Furthermore, the addition of the microstrip line provides an extra degree of freedom for the antenna matching [75].

The simulated radiation efficiency of the monopole operating at 900 MHz is 77% in comparison to the monopole in vacuum which shows an efficiency of 97%. The maximum antenna gain compared to an isotropic radiator is $G_{\text{Max}} = 1.3$ dBi on the body and 2.1 dBi in vacuum. The monopole at 2.45 GHz shows a radiation efficiency of 71% in comparison to an efficiency of 97% in vacuum. The maximum gain is 1.6 dBi on the body and 2.2 dBi in vacuum. The smaller efficiency at 2.45 GHz, although the groundplane is bigger than the radiator length in comparison to the 900 MHz monopole, is due to higher losses in the human tissue at higher frequencies (see Fig. 3.3).
3 Remote Health Monitoring

Patch Antennas

Patch antennas are low profile and show a broadside radiation, i.e., a maximum radiation away from the body, when they are excited at their fundamental mode [8]. This radiation characteristic is suitable for on-body links where the propagation path is a free space path or a shadowed free space path with diffraction around the body (e.g., the stomach-wrist link) [77]. In addition, there is the possibility to realize patch antennas which operate at a higher mode with a maximum radiation along the body’s surface [75]. This radiation pattern is suitable for on-body links where the propagation is predominantly due to a surface wave (e.g., the stomach-chest link). Hence, patch antennas are especially suitable for on-body applications. In this work, the less efficient patch antennas — in comparison to the monopole antennas — represent typical tag antennas and provide an insight in the performance of practical system implementations.

An initial set of patch antennas is designed and realized on FR-4 substrate with a thickness of 1.6 mm for the 900 MHz and 2.45 GHz regime. Again, the antennas are designed for an antenna-to-body separation of 5 mm. Photographs and the dimensions of the patch antennas can be seen in Fig. 3.6.

The antennas are fed by a microstrip line and matched to 50 Ω. The matching is about −10 dB in the operating frequency range (see Fig. 3.13 and Fig. 3.14).

The patch antennas operate at their fundamental mode and thus show a broad-
side radiation. The patch at 900 MHz shows a simulated radiation efficiency of 22\% and a maximum gain of $-1.2\, \text{dBi}$. In comparison, the patch antenna in vacuum shows an efficiency of 24\% and a gain of $-0.7\, \text{dBi}$. While the patch antenna for the 2.45\, \text{GHz} regime shows a radiation efficiency of 42\% on the body and of 47\% in vacuum. The maximum gain is 2.7\, \text{dBi} on the body and 3.7\, \text{dBi} in vacuum. The higher radiation efficiency in comparison to the patch antenna at 900\, \text{MHz} is due to the larger size of the groundplane relative to the patch size.

Figure 3.6: On-body patch antennas resonant at 900\, MHz (on the left) and 2.45\, GHz (on the right)
3.2 Radio Channel

In the following, the radio links of the on-body RFID system, which is composed of monopole or patch antennas, are investigated. The antennas operate as both reader and tag antennas. The investigation of the on-body system is done by means of channel measurements at 900 MHz and 2.45 GHz. The channel transfer functions of the forward link, $S_{21}$, and the backward link, $S_{12}$, between the reader antenna on the stomach and the four different tag antennas are captured by means of a VNA [20]. As stated earlier (see Sec. 1.1.2), these functions depend on the antenna characteristics of the reader and the tag and on the properties of the on-body propagation links.

3.2.1 Measurement Setup

The measurement setup is depicted in Fig. 3.7. The custom-built on-body antennas are connected via two coaxial cables to the VNA. The calibrated VNA measures the scattering parameters — reflection coefficients, $S_{11}$ and $S_{22}$, and transmission coefficients, $S_{12}$ and $S_{21}$ — at the antenna inputs and communicates via a local area network (LAN) with a personal computer (PC). The PC controls the measurement and defines parameters like measurement duration and rate.

With Rohde&Schwarz’s ZVA8 VNA, a good trade-off between the accuracy and the temporal resolution of the measurement is found. This trade-off manifests in a noise floor of about $-100$ dB for the transmission coefficients.

The coaxial cables are equipped with ferrite beads which are used to reduce sheath currents and to mitigate influences in the measurement due to the cables [79]. This is adequate for frequencies up to 1 GHz. Additionally, the positioning of the cables is arranged to minimize their influence. This is done by guiding the cables along the body. In this way, the near field of the on-body antenna is predominately influenced by the high complex permittivity of the human body and not by the measurement cables. The variations of the cable loss are less then 0.1 dB for movements similar to those performed during the measurement (see Tab. 3.2).

The scattering parameters are measured for 18 different stationary and moving body postures versus time [33] in an indoor scenario. The postures are listed in Tab. 3.2. Snapshots of these postures during the measurement of the stomach-head and stomach-back links are shown in Fig. 3.8. Each posture is held 20s. The repetition rate of the measurement is 5 Hz.
3.2 Radio Channel

Coaxial cables with ferrite beads
Vector network analyzer: measurement of scattering parameters \((S_{11}, S_{12}, S_{21}, S_{22})\)
LAN connection
PC

Figure 3.7: Measurement setup: The reflection and transmission properties of the on-body antennas are analyzed by means of a VNA.

Table 3.2: List of body postures performed during the channel measurement [33]

<table>
<thead>
<tr>
<th>Number</th>
<th>time</th>
<th>description</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0 – 20 s</td>
<td>standing, upright</td>
</tr>
<tr>
<td>2</td>
<td>20 – 40 s</td>
<td>standing, body turned left</td>
</tr>
<tr>
<td>3</td>
<td>40 – 60 s</td>
<td>standing, body turned right</td>
</tr>
<tr>
<td>4</td>
<td>60 – 80 s</td>
<td>standing, body leaning forward</td>
</tr>
<tr>
<td>5</td>
<td>80 – 100 s</td>
<td>standing, head leaning forward</td>
</tr>
<tr>
<td>6</td>
<td>100 – 120 s</td>
<td>standing, head turned left</td>
</tr>
<tr>
<td>7</td>
<td>120 – 140 s</td>
<td>standing, head turned right</td>
</tr>
<tr>
<td>8</td>
<td>140 – 160 s</td>
<td>standing, arms stretched out to sides</td>
</tr>
<tr>
<td>9</td>
<td>160 – 180 s</td>
<td>standing, arms above head</td>
</tr>
<tr>
<td>10</td>
<td>180 – 200 s</td>
<td>standing, arms reaching forward</td>
</tr>
<tr>
<td>11</td>
<td>200 – 220 s</td>
<td>standing, forearms forward</td>
</tr>
<tr>
<td>12</td>
<td>220 – 240 s</td>
<td>sitting, arms hanging body</td>
</tr>
<tr>
<td>13</td>
<td>240 – 260 s</td>
<td>sitting, hands in lap</td>
</tr>
<tr>
<td>14</td>
<td>260 – 280 s</td>
<td>standing, upright</td>
</tr>
<tr>
<td>15</td>
<td>280 – 300 s</td>
<td>standing, moving arms, head, and body randomly</td>
</tr>
<tr>
<td>16</td>
<td>300 – 320 s</td>
<td>sitting, moving arms, head, and body randomly</td>
</tr>
<tr>
<td>17</td>
<td>320 – 340 s</td>
<td>walking, back and forth</td>
</tr>
<tr>
<td>18</td>
<td>340 – 360 s</td>
<td>walking, moving arms, head, and body randomly</td>
</tr>
</tbody>
</table>
Figure 3.8: Snapshots of different body postures taken during several measurement runs: The snapshots show measurements of the stomach-head link (postures 1-11) and of the stomach-back link (postures 12-18).
3.2.2 Comparison of Simulations and Measurements

In this section, results obtained by the simulation are compared with the measurement. The comparison is done for the on-body system composed of monopole antennas operating at 900 MHz.

Fig. 3.9 and Fig. 3.10 compare the simulated and measured magnitudes of the scattering parameters in decibel versus a frequency range of 100 MHz – 3 GHz. The squared magnitudes of the reflection coefficients, $|S_{11}|^2$ and $|S_{22}|^2$, define the reader and tag antennas’ matching at their respective feed points. It can be seen that a good matching occurs at about 900 MHz and 2.6 GHz. The squared magnitudes of the channel transfer functions, $|S_{12}|^2$ and $|S_{21}|^2$, define the channel gain. $|S_{21}|^2$ is defined as the ratio of the power received at the tag antenna’s output (port 2 of the VNA) to the power available at the reader antenna’s input (port 1 of the VNA), while $|S_{12}|^2$ is defined as the ratio of the power received at the reader antenna’s output (port 1 of the VNA) to the power available at the tag antenna’s input (port 2 of the VNA). It can be seen that the channel gain is maximum at about 900 MHz and 2.6 GHz.

Fig. 3.9 shows a comparison of simulation and measurement for the stomach-back link, while Fig. 3.10 shows results of the stomach-chest link. The measurement curves picture a snapshot in time at $t = 15$ s in the standing, upright position. It can be seen that the simulation results are well confirmed by the measurements, although the simulation does not model body movements due to respiration, the indoor multipath environment, and the exact shape of the body. Consequently, the simplified human body model is an appropriate tool to design on-body antennas and roughly evaluate on-body links.
Figure 3.9: Simulated and measured magnitudes of scattering parameters ($|S_{11}|$, $|S_{12}|$, $|S_{21}|$, and $|S_{22}|$) in decibel versus frequency of the stomach-back link in a standing, upright position (measurement at $t = 15$ s)
Figure 3.10: Simulated and measured magnitudes of scattering parameters ($|S_{11}|$, $|S_{12}|$, $|S_{21}|$, and $|S_{22}|$) in decibel versus frequency of the stomach-chest link in a standing, upright position (measurement at $t = 15$ s)
3.2.3 Measurement Results

As mentioned above, the scattering parameters, $S_{11}$, $S_{22}$, $S_{12}$, and $S_{21}$, are measured at 900 MHz and 2.45 GHz in a realistic, multipath environment versus stationary and moving body postures (see Tab. 3.2).

Antenna Matching

Fig. 3.11 illustrates the matching of the reader’s monopole antenna, $|S_{11}|^2$, and the matching of the tag antennas, $|S_{22}|^2$, at 900 MHz versus stationary and moving body postures over time. Fig. 3.12 illustrates the matching of the monopole antennas at 2.45 GHz, while Fig. 3.13 and Fig. 3.14 depicts the matching of the patch antennas at 900 MHz and 2.45 GHz.

It can be seen that the antenna matching depends on the body posture, which influences the antenna-to-body separation distance, and on the antenna’s position on the body, which influences the reactive near field of the antenna due to varying effective permittivities. Thus, the tag antennas show a wider variation in matching in comparison to the reader antennas which are mounted on the female’s stomach. This behavior can be observed in Tab. 3.3 which lists the temporal mean of the antennas’ matching for all four on-body links. To account for varying proximity effects, the use of broadband antennas is advisable (see Chap. 2). In practice, the antennas’ matching is better than $-10$ dB which is deemed to be sufficient for real-world systems.

Table 3.3: Temporal mean of antennas’ matching for all four on-body links (stomach-back, stomach-chest, stomach-wrist, and stomach-head links)

<table>
<thead>
<tr>
<th>Reader antennas</th>
<th>back (dB)</th>
<th>chest (dB)</th>
<th>wrist (dB)</th>
<th>head (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Monopole @ 900 MHz</td>
<td>-26</td>
<td>-32</td>
<td>-26</td>
<td>-24</td>
</tr>
<tr>
<td>Monopole @ 2.45 GHz</td>
<td>-16</td>
<td>-19</td>
<td>-17</td>
<td>-17</td>
</tr>
<tr>
<td>Patch @ 900 MHz</td>
<td>-36</td>
<td>-36</td>
<td>-34</td>
<td>-33</td>
</tr>
<tr>
<td>Patch @ 2.45 GHz</td>
<td>-12</td>
<td>-12</td>
<td>-12</td>
<td>-12</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Tag antennas</th>
<th>back (dB)</th>
<th>chest (dB)</th>
<th>wrist (dB)</th>
<th>head (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Monopole @ 900 MHz</td>
<td>-23</td>
<td>-25</td>
<td>-17</td>
<td>-23</td>
</tr>
<tr>
<td>Monopole @ 2.45 GHz</td>
<td>-15</td>
<td>-17</td>
<td>-19</td>
<td>-17</td>
</tr>
<tr>
<td>Patch @ 900 MHz</td>
<td>-35</td>
<td>-32</td>
<td>-40</td>
<td>-33</td>
</tr>
<tr>
<td>Patch @ 2.45 GHz</td>
<td>-14</td>
<td>-13</td>
<td>-14</td>
<td>-14</td>
</tr>
</tbody>
</table>
Figure 3.11: Measured magnitudes of reflection coefficients of the reader and tags’ monopole antennas at 900 MHz, $|S_{11}|$ and $|S_{22}|$, in decibel versus stationary (0 – 280 s) and moving (280 – 360 s) body postures over time: Each posture is held 20 s.
Figure 3.12: Measured magnitudes of reflection coefficients of the reader and tags’ monopole antennas at 2.45 GHz, $|S_{11}|$ and $|S_{22}|$, in decibel versus stationary (0 – 280 s) and moving (280 – 360 s) body postures over time: Each posture is held 20 s.
Figure 3.13: Measured magnitudes of reflection coefficients of the reader and tags’ patch antennas at 900 MHz, $|S_{11}|$ and $|S_{22}|$, in decibel versus stationary (0 − 280 s) and moving (280 − 360 s) body postures over time: Each posture is held 20 s.
Figure 3.14: Measured magnitudes of reflection coefficients of the reader and tags’ patch antennas at 2.45 GHz, $|S_{11}|$ and $|S_{22}|$, in decibel versus stationary (0 – 280 s) and moving (280 – 360 s) body postures over time: Each posture is held 20 s.
3.2 Radio Channel

Channel Gain

Exemplarily, Fig. 3.15 plots the measured channel gain in the forward link, $|S_{21}|^2$, of the on-body system versus stationary and moving body postures. The reader and tag antennas are monopoles operating at 900 MHz. Another illustration of these measurement results is shown in Fig. 3.16. The figure depicts the cumulative distribution function (CDF) of the channel gain, $|S_{21}|^2$, for all four on-body links. The CDF describes the probability that the channel gain is less or equal to values plotted on the x-axis of Fig. 3.16 [20].

The figures show that on-body links with shorter path lengths (e.g., the stomach-chest link) have a higher channel gain in comparison to links with longer distances (e.g., the stomach-back link). In addition, it can be observed that, depending on the on-body link, the link geometry and thus the channel gain is influenced by the body movements. For example, the stomach-chest link shows a smaller variation in channel gain in comparison to the stomach-wrist link, which is more affected by movements. In general, trunk-to-trunk links are less influenced by the body’s movement in contrast to trunk-to-wrist and trunk-to-head links [33].

Any real propagation environment is symmetrical [15], i.e., the channel gains of the forward and backward links between the two communicating antennas are equal: $|S_{21}|^2 = |S_{12}|^2$. This is also found within these channel measurements. $|S_{21}|^2$ and $|S_{12}|^2$ are almost captured simultaneously during the measurement and are found to be virtually the same (see Fig. 3.9 and Fig. 3.10).
Figure 3.15: Measured magnitudes of the channel transfer function, $|S_{21}|$, in decibel for all four on-body links using the 900 MHz monopoles versus stationary (0 – 280 s) and moving (280 – 360 s) body postures over time: Each posture is held 20 s.

Figure 3.16: CDFs of the channel gain, $|S_{21}|^2$, for all four on-body links using the 900 MHz monopoles
3.3 System Performance

The evaluation of the performance of the on-body RFID system is done by means of the measured channel transfer functions, $S_{21}$ and $S_{12}$. As stated earlier, $|S_{21}|^2$ defines the channel gain in the forward link of the RFID system, while $|S_{12}|^2$ defines the channel gain in the backward link.

The CDF of $|S_{21}|^2$ directly relates to an outage probability in the system’s forward link [30], more precisely to the probability that the backscatter system operates at its limit. This probability is

$$P_F = P\{|S_{21}|^2 \leq F_{Th}\}. \quad (3.4)$$

The threshold, $F_{Th}$, is defined as the channel gain which is necessary to realize $P_{\text{Chip}} = T_{\text{Chip}}$, i.e., the power absorbed by the chip is equal to the chip’s sensitivity (see Eqn. 1.6):

$$F_{Th} = \frac{T_{\text{Chip}}}{\tau P_{\text{TX,Reader}}} \quad (3.5)$$

The CDF of the product of the channel gain in the forward link and backward link, $|S_{21}|^2|S_{12}|^2$, relates to an outage probability in the system’s backward link:

$$P_B = P\{|S_{21}|^2|S_{12}|^2 \leq B_{Th}\}. \quad (3.6)$$

The threshold, $B_{Th}$, is the total channel gain of the forward and backward links which is necessary to realize a RX power at the reader equal to the RX’s sensitivity, $P_{\text{RX,Reader}} = T_{\text{RX,Reader}}$ (see Eqn. 1.7):

$$B_{Th} = \frac{T_{\text{RX,Reader}}}{\eta P_{\text{TX,Reader}}} \quad (3.7)$$

With the probabilities defined in Eqn. 3.4 and Eqn. 3.6, the backscatter system performance can be analyzed. Such an analysis allows to explore each system parameter, i.e., $P_{\text{TX,Reader}}$, $\tau$, $T_{\text{Chip}}$, $\eta$, and $T_{\text{RX,Reader}}$, individually (see Eqn. 3.5 and Eqn. 3.7).

Usually, the outage probabilities which are allowed in an on-body system are governed by the application. In the case of a system which monitors life parameters of patients in clinical care, the outage probabilities should be close to zero. While systems used in sports analysis can deal with higher outage probabilities of, e.g., 10%.

3.3.1 Forward Link

Fig. 3.17, Fig. 3.18, Fig. 3.19, and Fig. 3.20 plot the outage probabilities in the forward link, $P_F$, versus the gain threshold, $F_{Th}$, for the respective links of the on-body
system using all four antenna configurations, i.e., monopoles or patch antennas operating at 900 MHz or 2.45 GHz.

As observed in Sec. 3.2.3, the curves show that links with longer path lengths show higher outage probabilities in comparison to links with shorter distances. In addition, depending on the on-body link, the link geometry and thus the channel gain is influenced by body movements. Thus, on-body links with higher mobility, i.e., trunk-to-limb links, experience a wider range of outage probabilities than trunk-to-trunk links with lower mobility. These phenomena can be observed for both antenna types in both frequency ranges.

As expected from theory, the outage probabilities at 900 MHz are lower than the probabilities at 2.45 GHz. This is due to an increased energy absorption in human tissues at higher frequencies (see Fig. 3.3). This behavior can be observed for both antenna types, although there are quite some differences in the patch antennas’ radiation efficiencies (see Sec. 3.1.2).

In addition, the probability curves show that the monopoles are indeed a best-case reference for on-body systems.

![Figure 3.17: Outage probability, $P_F$, versus gain threshold, $F_{Th}$, of the stomach-back forward link](image)

Figure 3.17: Outage probability, $P_F$, versus gain threshold, $F_{Th}$, of the stomach-back forward link
3.3 System Performance

Figure 3.18: Outage probability, $P_F$, versus gain threshold, $F_{Th}$, of the stomach-chest forward link

Figure 3.19: Outage probability, $P_F$, versus gain threshold, $F_{Th}$, of the stomach-wrist forward link
Figure 3.20: Outage probability, $P_F$, versus gain threshold, $F_{Th}$, of the stomach-head forward link
3.3 System Performance

**System Example**

Subsequently, the performance of an on-body RFID system is evaluated using state-of-the-art reader and tag chips.

A modern RFID system provides for example a gain threshold of $F_{\text{Th}} = -47$ dB (see Eqn. 3.5, with $T_{\text{Chip}} = -17.8$ dBm [12], $P_{\text{TX,Reader}} = 30$ dBm [80], and $\tau = 100\%$). In Tab. 3.4, the corresponding outage probabilities of each on-body link and for each antenna configuration are listed. The table shows that the investigated RFID system is mostly limited in its forward link.

There are different strategies to overcome this limitation. An increase in the TX power at the on-body reader is not an option because of the safety regulations and power constraints in on-body systems [81, 82]. A promising approach is to use semi-passive backscatter tags with chip sensitivities down to $-40$ dBm (see Sec. 1.1.1). Such a sensitivity corresponds to a gain threshold of $F_{\text{Th}} = -70$ dB. Tab. 3.5 shows that semi-passive tags lead to a good performance in the system’s forward links.

However, there is still a rather large limitation in the stomach-back link of the patch antenna systems. This constraint can be resolved by the use of more efficient patch antennas realized on a low-loss substrate or by the use of higher mode patches which benefit surface waves on the body. Another solution would be the use of a second RFID reader on the person’s back to reduce the distance of the propagation path.

| Table 3.4: Outage probabilities for a gain threshold of $F_{\text{Th}} = -47$ dB |
|-----------------|---|---|---|---|
|                 | Monopoles | Patch antennas |
| 900 MHz | back (%) | chest (%) | wrist (%) | head (%) |
| 30   | 0     | 6     | 21       |
| 2.45 GHz | 100 | 0     | 20       | 64       |
| 900 MHz | 100 | 40    | 46       | 86       |
| 2.45 GHz | 100 | 50    | 95       | 90       |
Table 3.5: Outage probabilities for a gain threshold of $F_{Th} = -70$ dB

<table>
<thead>
<tr>
<th>Antenna Type</th>
<th>Back (%)</th>
<th>Chest (%)</th>
<th>Wrist (%)</th>
<th>Head (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>900 MHz</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>2.45 GHz</td>
<td>2</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Antenna Type</th>
<th>Back (%)</th>
<th>Chest (%)</th>
<th>Wrist (%)</th>
<th>Head (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>900 MHz</td>
<td>10</td>
<td>0</td>
<td>2</td>
<td>2</td>
</tr>
<tr>
<td>2.45 GHz</td>
<td>44</td>
<td>0</td>
<td>4</td>
<td>1</td>
</tr>
</tbody>
</table>
3.3.2 Backward Link

Fig. 3.21, Fig. 3.22, Fig. 3.23, and Fig. 3.24 plot the outage probabilities of the backward links, $P_B$, versus the gain threshold, $B_{Th}$, for all antenna configurations. The same characteristic behavior of on-body links as described in Sec. 3.3.1 can be observed.

Figure 3.21: Outage probability, $P_B$, versus gain threshold, $B_{Th}$, of the stomach-backward link
Figure 3.22: Outage probability, $P_B$, versus gain threshold, $B_{Th}$, of the stomach-chest backward link

Figure 3.23: Outage probability, $P_B$, versus gain threshold, $B_{Th}$, of the stomach-wrist backward link
3.3 System Performance

Figure 3.24: Outage probability, $P_B$, versus gain threshold, $B_{Th}$, of the stomach-head backward link
A state-of-the-art RFID system provides for example a gain threshold of $B_{\text{Th}} = -118 \text{ dB}$ (see Eqn. 3.7, with $T_{\text{RX,Reader}} = -95 \text{ dBm}$ [80], $P_{\text{TX,Reader}} = 30 \text{ dBm}$, and $\eta = 20 \%$). In Tab. 3.6, the outage probabilities of each on-body link and for each antenna configuration are listed. Some limitations in the backward links of the system can be observed.

To overcome these, a phase-modulated backscatter signal could be used (see Sec. 1.1.1). This modulation scheme provides a maximum modulation efficiency of $\eta = 81 \%$ which corresponds to gain threshold of about $-124 \text{ dB}$. Tab. 3.7 lists the corresponding outage probabilities. It can be seen that the 6 dB difference in the threshold sufficiently improves the performance in the stomach-chest, stomach-wrist, and stomach-head links. If a phase-modulated backscatter signal is applied, there should be no substantial limitations in the system’s forward link because of a reduced power transmission coefficient (see Fig. 1.3). Other strategies to overcome backward link limitations are, e.g., the use of a second reader unit, the realization of more sophisticated antennas, or the use of a reader RX with a lower sensitivity.

Table 3.6: Outage probabilities for a gain threshold of $B_{\text{Th}} = -118 \text{ dB}$

<table>
<thead>
<tr>
<th>Monopoles</th>
<th>back (%)</th>
<th>chest (%)</th>
<th>wrist (%)</th>
<th>head (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>900 MHz</td>
<td>4</td>
<td>0</td>
<td>2</td>
<td>1</td>
</tr>
<tr>
<td>2.45 GHz</td>
<td>44</td>
<td>0</td>
<td>1</td>
<td>5</td>
</tr>
<tr>
<td>Patch antennas</td>
<td>back (%)</td>
<td>chest (%)</td>
<td>wrist (%)</td>
<td>head (%)</td>
</tr>
<tr>
<td>900 MHz</td>
<td>56</td>
<td>1</td>
<td>12</td>
<td>18</td>
</tr>
<tr>
<td>2.45 GHz</td>
<td>100</td>
<td>2</td>
<td>34</td>
<td>36</td>
</tr>
</tbody>
</table>

Table 3.7: Outage probabilities for a gain threshold of $B_{\text{Th}} = -124 \text{ dB}$

<table>
<thead>
<tr>
<th>Monopoles</th>
<th>back (%)</th>
<th>chest (%)</th>
<th>wrist (%)</th>
<th>head (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>900 MHz</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td>2.45 GHz</td>
<td>23</td>
<td>0</td>
<td>1</td>
<td>2</td>
</tr>
<tr>
<td>Patch antennas</td>
<td>back (%)</td>
<td>chest (%)</td>
<td>wrist (%)</td>
<td>head (%)</td>
</tr>
<tr>
<td>900 MHz</td>
<td>33</td>
<td>0</td>
<td>8</td>
<td>11</td>
</tr>
<tr>
<td>2.45 GHz</td>
<td>98</td>
<td>0</td>
<td>18</td>
<td>15</td>
</tr>
</tbody>
</table>
3.4 Summary

This chapter investigates an on-body backscatter RFID system for remote health monitoring applications at 900 MHz and 2.45 GHz. The investigation is done for two different on-body antenna types. Monopole antennas act as a best-case reference, while less efficient patch antennas are used to give insight into practical RFID system implementations. The antennas are designed by means of a human body model to account for proximity effects that are caused by human tissues. The body model reflects the behavior of the electromagnetic properties of human tissues and is found to be an appropriate tool for the design of on-body antennas.

The system’s performance is evaluated by means of on-body channel measurements in a realistic indoor scenario. In particular, the channel transfer functions of the system are examined versus different body postures and lead to outage probabilities of the system’s forward and backward links. These probabilities help to identify limitations in the backscatter system and to evaluate strategies to overcome these barriers for the realization of a reliable on-body RFID system. An analysis of a state-of-the-art system example shows that the use of semi-passive chips leads to a reliable performance in the system’s forward link. A strategy to overcome limitations in the system’s backward link is to use a phase-modulated backscatter signal.

It is worth pointing out that the presented analysis can be performed for any kind of backscatter RFID system. The analysis gives an initial overview of a backscatter system and ultimately allows to realize a robust power transmission and wireless communication.
4 Sensor for Curvature Monitoring

Curvature monitoring using low-cost and low-maintenance backscatter sensor tags is attractive for a multitude of different applications. For example, such backscatter bend sensors can be used in the entertainment industry where visitors in a theme park interact with robotic characters. Other possible applications of backscatter bend sensors are in structural health and remote health monitoring systems. In the latter, these sensor tags may be used to monitor the motion sequence of people for sports analysis or for human computer interaction devices when a patient is undergoing physical therapy.

Such a sensor tag can be realized by including a transducer in a backscatter RFID tag. The transducer changes its electrical impedance as a function of bending and thus directly modulates the backscatter signal’s amplitude and phase. Consequently, the RFID reader can wirelessly detect the sensor data without power consuming analog to digital converters (ADCs) or other conventional RF circuitry on the backscatter sensor.

This chapter provides a design for a wireless bend sensor using backscatter RFID at 5.8 GHz. Sec. 4.1 gives a detailed overview of the integration of a transducer in a backscatter RFID tag. In Sec. 4.2, a transducer prototype for the backscatter bend sensor is presented. An evaluation and optimization of the performance of the backscatter sensor is done in Sec. 4.3.

Original Publications Related to This Chapter


4 Curvature Monitoring

4.1 Sensing Approach

As described in Sec. 1.1, a backscatter RFID system relies on the wireless communication between an RFID reader and a tag. The reader transmits RF power and data to the tag, while the tag communicates by modulating the electromagnetic waves scattered at the input of its chip. The amplitude and phase of the scattered waves can be changed by creating an impedance mismatch between the impedance of the tag antenna and the tag chip.

A backscatter RFID sensor for curvature monitoring can be formed from an RFID tag by integrating a transducer that changes its impedance as a function of the curvature of an object to which the transducer is attached. As the transducer is bent its changing impedance causes an amplitude and phase change in the backscattered signal. This information can be wirelessly detected at the reader by observing the tag’s absorb state, $S^{(A)}$, and reflect state, $S^{(R)}$ (see Sec. 1.1.3). Assuming a perfect knowledge of the carrier leakage and the channel, these states are directly related to the tag’s reflection coefficients in the absorbing mode, $S_{\text{Abs}}$, and reflecting mode, $S_{\text{Ref}}$ [23]. In the following, it is assumed that the carrier signal and the channel are perfectly known.

There are two promising approaches to integrate a transducer in a backscatter RFID tag. One approach is to use the tag’s antenna as the transducer, denoted as an antenna transducer. The other approach is to integrate a specially-designed transducer structure in the antenna’s load. This is termed a chip transducer.

4.1.1 Antenna Transducer

In the case of the antenna transducer, the tag’s antenna is used to act as the transducing structure. Thus, the antenna’s impedance changes with the object’s curvature, $\Delta$. This relation is denoted as $Z_{\text{Ant}}(\Delta)$. A schematic of a backscatter sensor tag using the antenna as the transducer is sketched in Fig. 4.1, which is based on the tag’s block diagram shown in Fig. 1.1.

Due to the change in the antenna’s impedance with $\Delta$, there is also a change in the reflection coefficients in both the absorbing and reflecting modes, $S_{\text{Abs}}(\Delta)$ and $S_{\text{Ref}}(\Delta)$ (see Eqn. 1.1). Consequently, the chip’s power supply varies too, $\tau(\Delta)$ (see Eqn. 1.3). This is a disadvantage of the antenna transducer approach because a resultant power loss limits the backscatter system’s forward link.

Aside from that, the information about the object’s curvature lies within both reflection coefficients, $S_{\text{Abs}}(\Delta)$ and $S_{\text{Ref}}(\Delta)$, which necessitates a sophisticated detection at the reader.

The main advantage of the antenna transducer approach is that it allows the use of an off-the-shelf microchip. However, the absorbing and reflecting impedances of
the chip vary with the chip’s input power, $Z_{\text{Abs}}(P)$ and $Z_{\text{Ref}}(P)$ (see Sec. 1.1.1). Thus, the reflection coefficients change with the object’s curvature and the chip’s input power, $S_{\text{Abs}}(\Delta, P)$ and $S_{\text{Ref}}(\Delta, P)$, which makes an accurate detection of the bending state even more complex [83].

![Diagram of backscatter sensor tag using the antenna as the transducer](image)

**Figure 4.1:** Backscatter sensor tag using the antenna as the transducer, $Z_{\text{Ant}}(\Delta)$

### 4.1.2 Chip Transducer

Fig. 4.2 depicts a schematic of a backscatter sensor tag with a specially-designed transducer structure included in the chip’s reflecting impedance. A change of the chip’s reflecting impedance with the object’s curvature, $Z_{\text{Ref}}(\Delta)$, leads to a variation of the reflection coefficient in the reflecting mode, $S_{\text{Ref}}(\Delta)$. The reflection coefficient in the absorbing mode, $S_{\text{Abs}}$, stays constant with respect to $\Delta$ and leads to a stable power supply of the chip.

Thus, the chip transducer approach — where the transducer is included as the reflecting impedance of the chip, $Z_{\text{Ref}}(\Delta)$ — avoids the problem mentioned in Sec. 4.1.1. A transducer included in the chip’s absorbing impedance would again lead to an unstable power supply with respect to $\Delta$. Tab. 4.1 allows to compare the chip transducer approach with the antenna transducer approach.

In addition, the detection at the reader simplifies in comparison to the antenna transducer approach because the information about the object’s curvature is solely included in $S_{\text{Ref}}(\Delta)$.
4 Curvature Monitoring

In Fig. 4.2, the chip’s absorbing impedance is a function of the chip’s input power, $Z_{\text{Abs}}(P)$, which reflects the fact that the custom-built chip harvests power as it is the case in a conventional tag chip (see Fig. 1.2). Consequently, the reflection coefficient in the absorbing mode changes with the power too, $S_{\text{Abs}}(P)$. An other RF frontend architecture in the custom-built chip might lead to a different behavior of the sensor tag with $P$.

A disadvantage of the chip transducer solution is that it is necessary to design a custom-built chip including the transducer structure which is considerably bigger than an off-the-shelf chip used in the antenna transducer approach.

![Diagram of backscatter sensor tag with the transducer included in the tag’s reflecting impedance, $Z_{\text{Ref}}(\Delta)$](image)

**Figure 4.2:** Backscatter sensor tag with the transducer included in the tag’s reflecting impedance, $Z_{\text{Ref}}(\Delta)$

<table>
<thead>
<tr>
<th>Antenna transducer</th>
<th>chip transducer</th>
</tr>
</thead>
<tbody>
<tr>
<td>$Z_{\text{Ant}}(\Delta)$</td>
<td>$Z_{\text{Ant}} = \text{const.}$</td>
</tr>
<tr>
<td>$Z_{\text{Abs}}(P)$</td>
<td>$Z_{\text{Abs}}(P)$</td>
</tr>
<tr>
<td>$S_{\text{Abs}}(\Delta, P)$</td>
<td>$S_{\text{Abs}}(P)$</td>
</tr>
<tr>
<td>$Z_{\text{Ref}}(P)$</td>
<td>$Z_{\text{Ref}}(\Delta)$</td>
</tr>
<tr>
<td>$S_{\text{Ref}}(\Delta, P)$</td>
<td>$S_{\text{Ref}}(\Delta)$</td>
</tr>
</tbody>
</table>

Table 4.1: Comparison of the antenna and chip transducer approaches
4.1 Sensing Approach

**Ideal Chip Transducer**

The reflection coefficients in the absorbing and reflecting modes, $S_{\text{Abs}}$ and $S_{\text{Ref},1}$, of an exemplary backscatter RFID tag are depicted in Fig. 4.3. The antenna’s impedance is defined as $Z_{\text{Ant}} = (20 + j350)\Omega$, the chip’s impedances are assumed to be $Z_{\text{Abs}} = (20 - j350)\Omega$ and $Z_{\text{Ref},1} = (2 - j0.1)\Omega$. The figure shows the tag’s response and can be understood as a Smith chart representation with a characteristic impedance of $Z_{\text{Ant}} = (20 + j350)\Omega$. The reflection coefficient in the absorbing mode is zero, $S_{\text{Abs}} = 0$, because $Z_{\text{Ant}}$ is defined to be $Z_{\text{Abs}}^*$ (see Eqn. 1.1).

In addition, Fig. 4.3 pictures the response of an idealized backscatter sensor tag which includes an ideal chip transducer structure in its reflection impedance, $Z_{\text{Ref}}(\Delta)$. The diagram shows exemplarily four different reflecting states, $S_{\text{Ref},1}$, $S_{\text{Ref},2}$, $S_{\text{Ref},3}$, and $S_{\text{Ref},4}$, which are related to four different bending states or rather to four distinct reflecting impedances of the transducer structure: $Z_{\text{Ref},1} = (2 - j0.1)\Omega$, $Z_{\text{Ref},2} = -j332.5\Omega$, $Z_{\text{Ref},3} = -j351\Omega$, and $Z_{\text{Ref},4} = -j372\Omega$. The four different reflecting states show high phase differences of about $90^\circ$ and thus optimally exploit the available signal space. Such a wide separation between the different reflecting states assures a proper detection at the reader in the presence of noise and increases the sensitivity of the backscatter tag sensor to different bending states. Additionally, the different reflecting states feature a maximum modulation efficiency of $\eta = 20\%$ (see Sec. 1.1.1). For illustration purposes, contour lines of constant modulation efficiencies are plotted in Fig. 4.3.

An ideal chip transducer structure included in an RFID tag leads to high performances in the forward and backward links of a backscatter system and features a high sensitivity to variations in the object’s curvature.

Such an ideal transducer structure can be realized by obeying the following guidelines:

- High phase differences in the reflecting states can be achieved by reflecting impedances with $R_{\text{Ref}} \leq R_{\text{Ant}}$ and $X_{\text{Ref}} \approx -X_{\text{Ant}}$ while conjugate impedance matching to the antenna’s impedance, $Z_{\text{Ref}} = Z_{\text{Ant}}^*$, must be avoided. This behavior is demonstrated in Fig. 4.4, which plots the normalized phase of the reflection coefficient in the reflecting mode, $\varphi'_{\text{Ref}}$ (normalized to $180^\circ$), versus the real part, $R_{\text{Ref}}$, and imaginary part, $X_{\text{Ref}}$, of the reflecting impedance. Again, the antenna impedance is defined to be $Z_{\text{Ant}} = (20 + j350)\Omega$.

- High modulation efficiencies can be achieved by reflecting impedances with their real part close to zero or infinity, $R_{\text{Ref}} \approx 0$ or $R_{\text{Ref}} \approx \infty$, or impedances far away from the complex conjugate of the antenna’s impedance, $Z_{\text{Ant}}^*$. This is demonstrated in Fig. 4.5, which plots the normalized modulation efficiency, $\eta'$ (normalized to 0.2), versus $R_{\text{Ref}}$ and $X_{\text{Ref}}$ for $Z_{\text{Ant}} = (20 + j350)\Omega$.

In summary, an ideal chip transducer should feature impedances with $R_{\text{Ref}} \approx 0$ and $X_{\text{Ref}} \approx -X_{\text{Ant}}$. This is true for the example presented in Fig. 4.3.
4 Curvature Monitoring

Backscatter Transducer Efficiency

A performance parameter is introduced for an easy evaluation of a chip transducer structure in terms of its performance within a backscatter sensor tag and for a fast comparison of different transducer structures within a design process.

The backscatter transducer efficiency, $\alpha$, is defined as

$$\alpha = \sqrt{\frac{\eta_{\text{Min}} \cdot \Delta \varphi_{\text{Min}}}{0.2 \cdot \frac{360^\circ}{N}}}, \quad (4.1)$$

where $\eta_{\text{Min}}$ is the minimum modulation efficiency of all reflecting states, $\Delta \varphi_{\text{Min}}$ is the minimum phase difference of all reflecting states, and $N$ is the number of the individual reflecting states.

The idealized backscatter sensor tag as depicted in Fig. 4.3 reaches a backscatter transducer efficiency of $\alpha = 100\%$ with $\eta_{\text{Min}} = 0.2$, $\Delta \varphi_{\text{Min}} = 90^\circ$, and $N = 4$. 
4.1 Sensing Approach

Figure 4.3: Tag response showing the reflection coefficients in the absorbing mode, $S_{\text{Abs}}$, and the reflecting modes, $S_{\text{Ref,1}}$, $S_{\text{Ref,2}}$, $S_{\text{Ref,3}}$, and $S_{\text{Ref,4}}$ which correspond to $Z_{\text{Ant}} = (20 + j350) \Omega$, $Z_{\text{Abs}} = (20 - j350) \Omega$, $Z_{\text{Ref,1}} = (2 - j0.1) \Omega$, $Z_{\text{Ref,2}} = -j332.5 \Omega$, $Z_{\text{Ref,3}} = -j351 \Omega$, and $Z_{\text{Ref,4}} = -j372 \Omega$. In addition, contour lines of constant modulation efficiency are plotted.
Figure 4.4: Normalized phase of the reflection coefficient in the reflecting mode, \( \varphi'_{\text{Ref}} \) (normalized to 180°), versus the real part, \( R_{\text{Ref}} \), and imaginary part, \( X_{\text{Ref}} \), of the reflecting impedance: The antenna impedance is \( Z_{\text{Ant}} = (20 + j350) \, \Omega \).
Figure 4.5: Normalized modulation efficiency, $\eta'$ (normalized to 0.2), versus the real part, $R_{\text{Ref}}$, and imaginary part, $X_{\text{Ref}}$, of the reflecting impedance. The antenna impedance is $Z_{\text{Ant}} = (20 + j350) \, \Omega$. 
4 Curvature Monitoring

4.2 Transducer Prototype

In the following sections, a chip transducer structure for a backscatter bend sensor is realized and evaluated.

4.2.1 Microstrip Line Resonator

A resonant structure, e.g., a microstrip antenna or a transmission line resonator, shows a shift in its resonant frequency when it is bent along the dimension determining its resonance [84, 85, 86]. In addition, a resonator features impedances which are suitable for a sensor tag based on a chip transducer (see Sec. 4.1.2).

Thus, an open-circuited microstrip line resonator, which is bent in the direction of its transmission line, is explored to act as a chip transducer. The realized prototype can be seen in Fig. 4.6. The structure is made of single-sided, copper-clad flexible laminate, DuPont’s Pyralux FR9150 [87]. Its thickness is 0.127 mm and the material parameters have been provided by the manufacturer. The microstrip line is made of copper tape applied to the un-clad side of the laminate and has a width of 2 mm and a length of 55 mm which has been chosen to realize a high backscatter transducer efficiency at 5.8 GHz. The size of the groundplane is 30 mm × 80 mm.

Figure 4.6: The transducer prototype in the input impedance measurement setup.
4.2 Transducer Prototype

4.2.2 Input Impedance Measurement

The transducer’s input impedance is investigated for three exemplary bending states defined by their bending radius, $R_{\text{Bend}}$, i.e., the radius of the cylinder around which the transducer is bent. $R_{\text{Bend}} = \infty$ represents the planar case, while $R_{\text{Bend}} = 84.1\text{ mm}$ and $R_{\text{Bend}} = 21.1\text{ mm}$ represent two bending states with increasing curvature. Polyvinyl chloride (PVC) sheets and pipes are used to realize the different bending states (see Fig. 4.6). The transducer’s input impedance versus bending is measured by means of a VNA, Agilent’s E5071C. A sub-miniature-A (SMA) connector is used to feed the structure and the VNA’s calibration is adjusted accordingly. The relatively large dimensions of the prototype, which operates at 5.8 GHz, are chosen to allow the input impedance measurement using the VNA. It is found that the measurement setup is only suitable for longer resonator lengths because the behavior of the resonant frequencies with bending for short resonators is impaired by the rigidity of the attached SMA connector.

Fig. 4.7 shows the mean values of the structure’s input impedance, $Z = R + jX$, for a frequency range of $f = 5.0 - 5.8$ GHz and for all three different bending states, $R_{\text{Bend}} = \infty$, $R_{\text{Bend}} = 84.1\text{ mm}$, and $R_{\text{Bend}} = 21.1\text{ mm}$. The mean value is calculated from five different measurements in which the prototype is repeatedly attached to the PVC material and connected to the VNA. These repeatability measurements are done to verify the accuracy of the measurement. It can be seen from Fig. 4.7 that the structure’s resonant frequency shifts to lower frequencies when it is bent. This shift is also found in simulations with CST’s Microwave Studio [88]. Another prototype with the same dimensions is measured and, although the resonant frequencies differ because of manufacturing variances, it shows the same resonant frequency trend as the first.

The investigated transducer prototype shows the desired change in its input impedance with bending. The transducer, when integrated in a backscatter RFID tag, is operated in the resonator’s capacitive range of its third resonance at 5.8 GHz to realize a high backscatter transducer efficiency. Tab. 4.2 lists the realized impedance values at this frequency. In addition, operating the resonator at 5.8 GHz and not at its self-resonance reduces unintentional radiation from the structure as shown by simulations.

<table>
<thead>
<tr>
<th>$R_{\text{Bend}}$ (mm)</th>
<th>$Z$ (Ω)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\infty$</td>
<td>1.21 $- j21$</td>
</tr>
<tr>
<td>84.1</td>
<td>1.15 $- j20.3$</td>
</tr>
<tr>
<td>21.1</td>
<td>1.37 $- j18.1$</td>
</tr>
</tbody>
</table>
Figure 4.7: Measured transducer input impedance, $Z = R + jX$, versus frequency for three different bending states, $R_{\text{Bend}} = \infty$, $R_{\text{Bend}} = 84.1 \text{ mm}$, and $R_{\text{Bend}} = 21.1 \text{ mm}$
4.3 Sensor Performance

An initial design of a backscatter bend sensor including the proposed transducer prototype is shown in Fig. 4.8. The assembly includes an antenna, a microwave switch which needs a power supply, the transducer prototype, and a constant load defining the tag’s absorbing impedance, $Z_{\text{Abs}}$. The antenna should not bend with the object’s curvature or should be insensitive to bending. This insensitivity can be realized for example by using a broadband antenna structure as proposed in Chap. 2.

Fig. 4.9 shows the calculated response of the backscatter sensor tag at 5.8 GHz. The reflection coefficient in the absorbing mode, $S_{\text{Abs}}$, is zero, while the reflection coefficient in the reflecting mode, $S_{\text{Ref}}$, changes with the object’s curvature, $R_{\text{Bend}} = \infty$, $R_{\text{Bend}} = 84.1$ mm, and $R_{\text{Bend}} = 21.1$ mm. The phase differences between the reflecting states are $16.5^\circ$ and $60.2^\circ$ with respect to $R_{\text{Bend}} = \infty$. Additionally, modulation efficiencies of about 8% are achieved (see Tab. 4.3). This good tag response is realized by an optimized antenna impedance of $Z_{\text{Ant}} = (5.5 + j20.5)$ $\Omega$ and features a backscatter transducer efficiency of about 23% (see Eqn. 4.1, with $\eta_{\text{Min}} = 0.08$, $\Delta \varphi_{\text{Min}} = 16.5^\circ$, and $N = 3$). The optimization is done by searching for the maximum backscatter transducer efficiency, $\alpha$, with respect to the antenna’s impedance, $Z_{\text{Ant}} = R_{\text{Ant}} + jX_{\text{Ant}}$, at $f = 5.8$ GHz. The choice of the antenna impedance range is guided by the formulations introduced in Sec. 4.1.2 for an ideal chip transducer design. By means of this optimization, it is possible to realize a good backscatter transducer efficiency, although the prototype features only small changes in its input impedance with respect to the object’s curvature.

Furthermore, it is found that resonant structures with strong and narrowband resonances feature highest backscatter transducer efficiencies. Fig. 4.10 and Fig. 4.11 plot the measured impedance curves of the transducer prototype, $Z = R + jX$, for all three different bending states versus frequency (compare with Fig. 4.7). In addition, Fig. 4.10 shows the corresponding contour lines of $\varphi_{\text{SRef}}'$ and Fig. 4.11 plots the corresponding contour lines of $\eta'$ for $Z_{\text{Ant}} = (5.5 + j20.5)$ $\Omega$. A stronger resonance

<table>
<thead>
<tr>
<th>$R_{\text{Bend}}$ (mm)</th>
<th>$Z$ ($\Omega$)</th>
<th>$\Delta \varphi$ ($^\circ$)</th>
<th>$\eta$ (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\infty$</td>
<td>1.21 – j21</td>
<td>–</td>
<td>8</td>
</tr>
<tr>
<td>84.1</td>
<td>1.15 – j20.3</td>
<td>16.5</td>
<td>9</td>
</tr>
<tr>
<td>21.1</td>
<td>1.37 – j18.1</td>
<td>60.2</td>
<td>9</td>
</tr>
</tbody>
</table>

Table 4.3: Transducer impedance, $Z$, phase difference, $\Delta \varphi$, with respect to $R_{\text{Bend}} = \infty$, and modulation efficiency, $\eta$, of the individual reflecting states at 5.8 GHz.
of the prototype would result in larger circles of the impedance curves and would thus feature impedances closer to $R_{\text{Ref}} \approx 0$ and $X_{\text{Ref}} \approx -X_{\text{Ant}}$ which lead to higher modulation efficiencies and phase differences of the reflecting states (see Sec. 4.1.2). In the case of the microstrip line resonator, a stronger resonance can be achieved by the use of a thinner microstrip line, by the use of a low-loss substrate, and by the use of the first resonance which is typically the strongest.

Figure 4.8: Initial design of a backscatter bend sensor including an antenna, a microwave switch, the transducer prototype, and a constant load, $Z_{\text{Abs}}$. 
Figure 4.9: Tag response of the backscatter bend sensor for an optimized antenna impedance of $Z_{\text{Ant}} = (5.5 + j20.5) \Omega$ at 5.8 GHz
Figure 4.10: Impedances of the transducer prototype, $Z = R + jX$, for all three different bending states versus frequency ($5.0 - 5.8$ GHz): In addition, the normalized phase of the reflection coefficient in the reflecting mode, $\varphi'_{\text{Ref}}$ (normalized to 180°), is plotted versus the reflecting impedance, $Z_{\text{Ref}} = R_{\text{Ref}} + jX_{\text{Ref}}$, for $Z_{\text{Ant}} = (5.5 + j20.5) \Omega$. 
Figure 4.11: Impedances of the transducer prototype, $Z = R + jX$, for all three different bending states versus frequency ($5.0 - 5.8$ GHz): In addition, the normalized modulation efficiency, $\eta'$ (normalized to 0.2), is plotted versus the reflecting impedance, $Z_{\text{Ref}} = R_{\text{Ref}} + jX_{\text{Ref}}$, for $Z_{\text{Ant}} = (5.5 + j20.5) \Omega$. 

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4 Curvature Monitoring

4.4 Summary

This chapter provides a design for a backscatter RFID sensor tag that monitors the varying curvature of an object at 5.8 GHz. The backscatter sensor includes a transducer in the tag antenna’s load. In particular, the transducer, which acts as the chip’s reflecting impedance, assures a stable power supply to the chip’s circuitry and benefits the detection at the RFID reader. The transducer changes its impedance as a function of bending and directly modulates the carrier signal sent by the reader. Thus, there is no need to include a power consuming ADC or other RF transmitting hardware in the sensor tag.

A microstrip line resonator is prototyped for integration in such a backscatter bend sensor. It is found that the resonator shows the desired change in input impedance with bending. Furthermore, the resonator’s backscatter transducer efficiency is evaluated and optimized with respect to the sensor’s sensitivity to bend and with respect to the sensor tag’s modulation efficiency. It is found that the prototype qualifies for integration in the sensor tag and assures a good RFID system performance.

In order to validate the work, backscatter measurements should be carried out. In addition, further studies are needed to assess the influence of the chip’s input power on the sensor tag’s response and to evaluate the influence of an imperfect carrier cancelation and channel estimation on the detection of the individual bending states at the reader. Finally, it should be noted that with the introduction of the backscatter transducer efficiency, it is possible to easily compare different transducer prototypes and to realize high sensor performances.
5 Summary and Conclusions

Backscatter RFID in the UHF and microwave frequency ranges is a promising communication technology for wireless sensing applications. In particular, the operation in the unlicensed frequency bands around 900 MHz, 2.45 GHz, and 5.8 GHz, is attractive for many sensor systems because of relatively large communication distances, comparatively small devices, and potentially high data rates. Backscatter RFID in sensor networks relies on the radio communication between an RFID reader, acting as a control unit, and a multitude of passive or semi-passive RFID tags, acting as sensor nodes. The principle of communication for transmitting information from the tag to the reader relies on a modulated backscatter signal. All power for the transmission of the sensor data is drawn from the electromagnetic field radiated by the reader. Hence, their low-power consumption makes backscatter tags appropriate for sensing applications that require small, light-weight, and low-maintenance sensor nodes.

In backscatter RFID systems, it is vital to ensure a reliable power transmission to the backscatter tags and to realize a robust wireless communication between the reader and tags. For example, if the power at the tag’s chip is smaller than the chip’s sensitivity, the backscatter communication system is limited in its forward link. If the power at the reader RX is smaller than the RX’s sensitivity, a limitation in the backward link occurs. Thus, the proper design of backscatter RFID devices, which are included in sensor tags, and the investigation of backscatter radio channels are key study areas.

This thesis examines the use of backscatter RFID in sensing applications like advanced tire monitoring, remote health monitoring, and curvature monitoring. Particular attention is paid to the tag antenna design of a WU in an ATMS, the performance of an on-body RFID system for a WBAN, and the design of a transducer for a backscatter bend sensor to monitor the curvature of an object. The major design goal in this work is to realize backscatter RFID systems and devices which lead to high system performances. Special care is taken to realize systems and devices which operate efficiently in a car tire and on the human body whereas in the case of the backscatter sensor, the focus is on low power consumption and an acceptable sensitivity to bend.
Chap. 2: Transponder Antenna for Car Tire Monitoring

This chapter deals with the design of a tag antenna at 864 MHz for the WU of an ATMS based on backscatter RFID. One premise of the application is that the antenna is directly attached to a car tire. Thus, the antenna’s surroundings and proximity effects caused by the tire environment are explored.

It is found that the tire with its complex structure has a considerable influence on the antenna parameters like input impedance and gain. Measurements of the dielectric properties of the tire rubber show that its material parameters vary strongly with type and vendor of the tire. This fact leads to a detuning of the antenna’s resonant frequency which is rather difficult to predict. Additionally, the radiation pattern of a dipole antenna, which is attached to the thinnest part of the inner tire sidewall, is favorably influenced by the metal reinforcements of a standard car tire.

Consequently, a tag antenna prototype is realized which can cope with a change in the tire environment. The prototype — a T-matched dipole — is power-matched to the chip over a wide range of frequencies. Further analyses show that the antenna prototype assures a stable power supply to the tag’s chip and thus leads to a good performance in the ATMS using passive WU tags. In addition, the distortion of the antenna’s radiation pattern due to the tire is beneficially exploited and leads to a further enhancement in system performance.

In a next step, the tag antenna’s matching to the chip should be investigated for varying chip input powers which lead to a variation in the chip’s impedance. In general, the concept of a broadband tag antenna — which is exploited in this chapter — proves feasible for a multitude of applications where a change in the environment is foreseeable, e.g., when a tag is attached to the human body.

Chap. 3: Communication System for Remote Health Monitoring

This chapter investigates an on-body backscatter RFID system for remote health monitoring applications at 900 MHz and 2.45 GHz. The investigation is done for two different on-body antenna types. Monopole antennas act as a best-case reference, while less efficient patch antennas are used to give insight into practical RFID system implementations. The antennas are designed by means of a human body model to account for proximity effects that are caused by human tissues. The body model reflects the behavior of the electromagnetic properties of human tissues and is found to be an appropriate tool for the design of on-body antennas.

The system’s performance is evaluated by means of on-body channel measurements in a realistic indoor scenario. In particular, the channel transfer functions of the system are examined versus different body postures and lead to outage probabilities of the system’s forward and backward links. These probabilities help to identify limitations in the backscatter system and to evaluate strategies to overcome these barriers for the realization of a reliable on-body RFID system. An analysis of a state-of-the-art system example shows that the use of semi-passive chips leads
to a reliable performance in the system’s forward link. A strategy to overcome limitations in the system’s backward link is to use a phase-modulated backscatter signal.

It is worth pointing out that the presented analysis can be performed for any kind of backscatter RFID system. The analysis gives an initial overview of a backscatter system and ultimately allows to realize a robust power transmission and wireless communication.

Chap. 4: Sensor for Curvature Monitoring

This chapter provides a design for a backscatter RFID sensor tag that monitors the varying curvature of an object at 5.8 GHz. The backscatter sensor includes a transducer in the tag antenna’s load. In particular, the transducer, which acts as the chip’s reflecting impedance, assures a stable power supply to the chip’s circuitry and benefits the detection at the RFID reader. The transducer changes its impedance as a function of bending and directly modulates the carrier signal sent by the reader. Thus, there is no need to include a power consuming ADC or other RF transmitting hardware in the sensor tag.

A microstrip line resonator is prototyped for integration in such a backscatter bend sensor. It is found that the resonator shows the desired change in input impedance with bending. Furthermore, the resonator’s backscatter transducer efficiency is evaluated and optimized with respect to the sensor’s sensitivity to bend and with respect to the sensor tag’s modulation efficiency. It is found that the prototype qualifies for integration in the sensor tag and assures a good RFID system performance.

In order to validate the work, backscatter measurements should be carried out. In addition, further studies are needed to assess the influence of the chip’s input power on the sensor tag’s response and to evaluate the influence of an imperfect carrier cancelation and channel estimation on the detection of the individual bending states at the reader. Finally, it should be noted that with the introduction of the backscatter transducer efficiency, it is possible to easily compare different transducer prototypes and to realize high sensor performances.
<table>
<thead>
<tr>
<th>Acronym</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>ADC</td>
<td>analog to digital converter.</td>
</tr>
<tr>
<td>ATMS</td>
<td>advanced tire monitoring system.</td>
</tr>
<tr>
<td>CDF</td>
<td>cumulative distribution function.</td>
</tr>
<tr>
<td>I</td>
<td>inphase.</td>
</tr>
<tr>
<td>LAN</td>
<td>local area network.</td>
</tr>
<tr>
<td>OU</td>
<td>on-board unit.</td>
</tr>
<tr>
<td>PC</td>
<td>personal computer.</td>
</tr>
<tr>
<td>PVC</td>
<td>polyvinyl chloride.</td>
</tr>
<tr>
<td>Q</td>
<td>quadrature.</td>
</tr>
<tr>
<td>RF</td>
<td>radio frequency.</td>
</tr>
<tr>
<td>RFID</td>
<td>radio frequency identification.</td>
</tr>
<tr>
<td>RUT</td>
<td>rubber under test.</td>
</tr>
<tr>
<td>RX</td>
<td>receiver.</td>
</tr>
<tr>
<td>SMA</td>
<td>sub-miniature-A.</td>
</tr>
<tr>
<td>SWCNT</td>
<td>single-walled carbon nanotube.</td>
</tr>
<tr>
<td>tag</td>
<td>transponder.</td>
</tr>
<tr>
<td>TPMS</td>
<td>tire pressure monitoring system.</td>
</tr>
<tr>
<td>TX</td>
<td>transmitter.</td>
</tr>
<tr>
<td>UHF</td>
<td>ultra high frequency.</td>
</tr>
<tr>
<td>VNA</td>
<td>vector network analyzer.</td>
</tr>
<tr>
<td>WBAN</td>
<td>wireless body area network.</td>
</tr>
<tr>
<td>WU</td>
<td>wheel unit.</td>
</tr>
</tbody>
</table>
List of Symbols

List of Latin symbols

\( a \)  
geometry parameter of the T-matched dipole.

\( b \)  
geometry parameter of the T-matched dipole.

\( B_{\text{Th}} \)  
channel gain threshold of the combined forward and backward links.

\( C \)  
capacitance.

\( d \)  
transmission distance.

\( D \)  
maximal lateral dimension of the antenna.

\( f \)  
frequency.

\( F_{\text{Th}} \)  
channel gain threshold of the forward link.

\( F_\vartheta \)  
radiation pattern of the \( \vartheta \)-polarization field.

\( g \)  
feed gap length.

\( G_{\text{ISO},\vartheta} \)  
antenna gain in \( \vartheta \)-polarization compared to the isotropic radiator.

\( G_{\text{ISO},\varphi} \)  
antenna gain in \( \varphi \)-polarization compared to the isotropic radiator.

\( G_{\text{Max}} \)  
maximum antenna gain.

\( h \)  
transponder’s signal in the baseband.

\( l \)  
dipole length.

\( L \)  
carrier leakage in the baseband.

\( m \)  
index of dispersion regions.

\( N \)  
number of bending states.

\( P \)  
power.

\( P_B \)  
outage probability of the backward link.

\( P_{\text{Chip}} \)  
power absorbed by the chip.
\( P_F \) outage probability of the forward link.
\( P_{\text{Tag}} \) chip input power.
\( P_{\text{TX,Reader}} \) reader transmitter power.
\( P_{\text{RX,Reader}} \) reader receiver power.

\( R \) real part of the input impedance, resistance.
\( R_{\text{Abs}} \) real part of the chip’s absorbing impedance.
\( R_{\text{Ant}} \) real part of the antenna impedance.
\( R_{\text{Bend}} \) bending radius.
\( r_R \) Rayleigh distance.
\( R_{\text{Ref}} \) real part of the chip’s reflecting impedance.

\( S^{(A)} \) transponder’s absorb state in the baseband.
\( S_{\text{Abs}} \) reflection coefficient in the absorbing mode.
\( S_{\text{Meas}} \) measured reflection coefficient.
\( S^{(R)} \) transponder’s reflect state in the baseband.
\( S_{\text{Ref}} \) reflection coefficient in the reflecting mode.
\( S_{\text{Sim}} \) simulated reflection coefficient.
\( S_{11} \) reflection coefficient.
\( S_{12} \) channel transfer function of the backward link.
\( S_{21} \) channel transfer function of the forward link.
\( S_{22} \) reflection coefficient.

\( t \) time.
\( T_{\text{Chip}} \) chip sensitivity.
\( T_{\text{RX,Reader}} \) reader receiver sensitivity.

\( w \) dipole width.
\( w_{\text{is}} \) geometry parameter of the T-matched dipole.

\( x \) \( x \)-coordinate.
\( X \) imaginary part of the input impedance, reactance.
\( X_{\text{Abs}} \) imaginary part of the chip’s absorbing impedance.
\( X_{\text{Ant}} \) imaginary part of the antenna impedance.
\( X_{\text{Ref}} \) imaginary part of the chip’s reflecting impedance.

\( Z \) input impedance.
\( Z_{\text{Abs}} \) chip’s absorbing impedance.
\( Z_{\text{Ant}} \) antenna impedance.
\( Z_{\text{Ref}} \) chip’s reflecting impedance.
\( Z_0 \) characteristic impedance.
List of Greek symbols

\( \alpha \) backscatter transducer efficiency.
\( \alpha_m \) material parameter of the dispersion region \( m \).

\( \tan(\delta) \) loss tangent.
\( \Delta \) displacement.
\( \Delta \varepsilon_m \) material parameter of the dispersion region \( m \).
\( \Delta \varphi \) phase difference.
\( \Delta \varphi_{\text{Min}} \) minimum phase difference.
\( \Delta \varphi'_{\text{Min}} \) normalized minimum phase difference.

\( \tilde{\varepsilon}_r \) complex relative permittivity.
\( \varepsilon_r \) real part of the complex relative permittivity.
\( \varepsilon''_r \) imaginary part of the complex relative permittivity.
\( \varepsilon_0 \) free space permittivity.
\( \varepsilon_\infty \) material permittivity at terahertz frequency.

\( \eta \) modulation efficiency.
\( \eta' \) normalized modulation efficiency.
\( \eta_{\text{Min}} \) minimum modulation efficiency.
\( \eta'_{\text{Min}} \) normalized minimum modulation efficiency.

\( \vartheta \) \( \vartheta \)-coordinate.

\( \lambda \) wavelength.

\( \sigma \) electric conductivity.
\( \sigma_i \) ionic conductivity.

\( \tau \) power transmission coefficient.
\( \tau_m \) material parameter of the dispersion region \( m \).

\( \varphi \) \( \varphi \)-coordinate.
\( \varphi_{\text{SRef}} \) phase of the reflection coefficient in the reflecting mode.
\( \varphi'_{\text{SRef}} \) normalized phase of the reflection coefficient in the reflecting mode.

\( \omega \) angular frequency.
Bibliography


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