Scalar Method for Reading of Chipless RFID Tags Based on Limited Ground Plane Backed Dipole Resonator Array

Jan Kracek, Member, IEEE, Milan Svanda, Member, IEEE, and Karel Hoffmann, Senior Member, IEEE

Abstract—In this paper, we propose a novel frequency-domain reading method for the chipless radio frequency identification (RFID) tags. It is based on a single scalar measurement of transmission coefficient magnitude between two antennas by the presence of a tag. The method enables to reliably read the chipless RFID tags based on a dipole resonator array, which is backed by a limited ground plane, in a real environment outside the anechoic chamber. Moreover, it suits simple, yet not necessarily well impedance matched, reader antennas represented by an open end of the rectangular waveguide. The method was verified experimentally in the frequency band of 7–11 GHz.

Index Terms—Chipless radio frequency identification (RFID), frequency-domain reading method, scalar measurement, sensor, radar cross section (RCS), tag.

I. INTRODUCTION

T

HE use of radio frequency identification (RFID) systems has spread into numerous application areas, such as acceleration of logistic storing procedures, production process monitoring in industry, supply chain management, and tracking and identification of animals or people in access and monitoring systems [1]. The identification information can be stored in various types of tags. Most applications operate in the UHF band and require tag antennas tolerant to metallic and dielectric platforms, e.g., metallic boxes, foodstuffs, and animal or human bodies [2]–[7]. Conventional RFID systems operating in the RF and microwave frequency bands use passive tags with semiconductor chips that enable N-bit information to be stored.

The said tags with semiconductor chips suffer from high production costs, which limit their application in situations when a high number of tags is required. For comparison, the conventional tags are about hundred times more expensive than optical barcodes [8]. A promising solution for tag cost reduction is represented by a chipless RFID concept [9], which includes different possibilities of storing identification information without chip employment.

In the case of chipless RFID, the identification information is stored in the electromagnetic properties of a tag structure within the given frequency band. The information is read through the tag scattering response, which is dependent on its structure. The response is invoked by reradiation of a part of the incident electromagnetic wave that is sent to the tag. The same structure can be used for reradiation of the response as well as for encoding the information into the response [12], or the tag can include two separate structures for these purposes [9].

Reading methods constitute an important part of chipless RFID systems development. They can be divided into two main groups: the methods based on the time-domain reading and those based on the frequency-domain reading. Note that the division is done based on the display of the identification information in the time or frequency domain and it is not done based on the means used for its acquisition.

In the case of the time-domain reading, bits of identification information are encoded in the scattering response consisting of a set of pulses with a different time spacing between them [10], [11]. The response is invoked by an electromagnetic wave pulse irradiating the tag.

As far as the frequency-domain reading is concerned, bits of identification information are encoded in the frequency spectrum of the scattering response. The complete response is invoked usually by a continuous electromagnetic wave with a swept frequency within the given range irradiating the tag. It can also be invoked by an electromagnetic wave pulse containing the frequency spectrum within a given range. The spectrum of the response typically contains a set of peaks or drops. Tag modifications, such as adding or removal of a resonator structure, result in the presence or absence of spectrum extremes corresponding to the resonances of the structures, which can be used to encode a logical 1 or 0 of bits [12]–[15].

On the basis of the frequency-domain reading, relatively simple and potentially fully printable structures of chipless tags can be designed. The scattering response of these tags is invoked by the electromagnetic wave radiated by a transmitting antenna and detected by a receiving antenna of the RFID system. It is read as the reflection coefficient at the transmitting/receiving antenna or as the transmission coefficient between the transmitting antenna and the receiving antenna.
ways for the acquisition of the response, whose overview is mentioned further, were published. However, each of them embodies some disadvantages.

The tag scattering response can be read with the help of a radar cross section (RCS) of the tag which can be measured by two approaches: the bistatic (different transmitting and receiving antennas [12], [16], [17]) and monostatic (common transmitting and receiving antenna [15], [18], [19]). The method in the bistatic approach is usually based on the vector measurement of a set of transmission coefficients $S_{21\text{tag}}$, $S_{21\text{ref}}$, and $S_{21\text{iso}}$ between the transmitting and receiving antennas [12]. $S_{21\text{tag}}$ refers to the transmission coefficient by the tag presence, $S_{21\text{ref}}$ represents the transmission coefficient by the presence of a reference object with a known RCS $\sigma_{\text{ref}}$, and $S_{21\text{iso}}$ stands for the transmission coefficient without any presence of the tag or reference object. The RCS $\sigma_{\text{tag}}$ of the tag is then expressed by the relation

$$\sigma_{\text{tag}} = \left| \frac{S_{21\text{tag}} - S_{21\text{iso}}}{S_{21\text{ref}} - S_{21\text{iso}}} \right|^2 \sigma_{\text{ref}}. \quad (1)$$

A rectangular metal plate with dimensions $a$ and $b$ can be used as the object with the known RCS $\sigma_{\text{ref}}$ given by the formula [20]

$$\sigma_{\text{ref}} = 4\pi \frac{a^2 b^2}{\lambda^2} \quad (2)$$

where $\lambda$ is the wavelength. For the monostatic case, the transmission coefficients $S_{11\text{tag}}$, $S_{11\text{ref}}$, and $S_{11\text{iso}}$ correspond to reflection coefficients $S_{11\text{tag}}$, $S_{11\text{ref}}$, and $S_{11\text{iso}}$ of the common transmitting/receiving antenna and the relation (1) then becomes [18]

$$\sigma_{\text{tag}} = \left| \frac{S_{11\text{tag}} - S_{11\text{iso}}}{S_{11\text{ref}} - S_{11\text{iso}}} \right|^2 \sigma_{\text{ref}}. \quad (3)$$

Costa et al. [19] employed the fact that the relation (1) can be modified and the tag scattering response can be evaluated based on the relation

$$\Delta_{21} = \left| S_{21\text{tag}} - S_{21\text{iso}} \right|^2 = \left| \frac{S_{21\text{ref}} - S_{21\text{iso}}}{\sigma_{\text{ref}}} \right|^2 \sigma_{\text{tag}} = \nu_{21} \sigma_{\text{tag}} \quad (4)$$

where

$$\nu_{21} = \left| \frac{S_{21\text{ref}} - S_{21\text{iso}}}{\sigma_{\text{ref}}} \right|^2 \approx \text{const.} \quad (5)$$

It means that the quantity $\nu_{21}$ is considered to be slow varying with frequency, i.e., approximately constant, related to the quantity $\Delta_{21}$. Thus, it is just a scaling constant without information about the tag scattering response. The quantity $\Delta_{21}$ can be then used for the evaluation of the response instead of the RCS $\sigma_{\text{tag}}$. Similarly, the relation (3) can be modified to

$$\Delta_{11} = \left| S_{11\text{tag}} - S_{11\text{iso}} \right|^2 = \left| \frac{S_{11\text{ref}} - S_{11\text{iso}}}{\sigma_{\text{ref}}} \right|^2 \sigma_{\text{tag}} = \nu_{11} \sigma_{\text{tag}} \quad (6)$$

where

$$\nu_{11} = \left| \frac{S_{11\text{ref}} - S_{11\text{iso}}}{\sigma_{\text{ref}}} \right|^2 \approx \text{const.} \quad (7)$$

The main disadvantage of the respective methods consists in the need for a set of three or two precise vector measurements of the coefficients $S_{11\text{tag}}$, $S_{11\text{ref}}$, $S_{11\text{iso}}$, $S_{21\text{tag}}$, $S_{21\text{ref}}$, and $S_{21\text{iso}}$. These measurements are then used for the evaluation of the tag scattering response according to (1) and (3) or (4) and (6) since the coefficients $S_{21\text{tag}}$, $S_{21\text{iso}}$ or $S_{11\text{tag}}$, $S_{11\text{iso}}$ are comparable. It is caused by the properties of the transmitting and receiving antennas arrangement and by a low RCS of some tags. It means that the crosstalk between the transmitting and receiving antennas represents the main contribution to the coefficients $S_{21\text{tag}}$ and $S_{21\text{iso}}$. Similarly, the self-reflection of the common transmitting/receiving antenna represents the main contribution to the coefficients $S_{11\text{tag}}$ and $S_{11\text{iso}}$. Thus, the information about the tag constitutes a minor contribution to the coefficients $S_{21\text{tag}}$ or $S_{11\text{tag}}$ and has to be gained through (1) and (4) or (3) and (6). On the contrary, these methods enable to read tags with relatively low RCS (even $-60$ dBsm), yet provided that the measurements are performed in an anechoic chamber with a very stable background.

Further simplification of evaluation of the tag scattering response can be achieved provided $\left| S_{21\text{tag}} \right| \gg \left| S_{21\text{iso}} \right|$ or $\left| S_{11\text{tag}} \right| \gg \left| S_{11\text{iso}} \right|$ and the relation (4) or (6) then becomes

$$\Delta_{21} \approx \left| S_{21\text{tag}} \right|^2 \quad (8)$$

$$\Delta_{11} \approx \left| S_{11\text{tag}} \right|^2. \quad (9)$$

It means that only one scalar measurement of the magnitude $\left| S_{21\text{tag}} \right|$ or $\left| S_{11\text{tag}} \right|$ is necessary. This approach to a chipless tag design and reading can be found in [8], [9], and [21]. Here, the tag contains two cross-polarized UWB antennas, a microstrip transmission line connected between them and a set of various resonators, which are coupled to the microstrip transmission line. The microstrip transmission line with the set of resonators encodes information into the tag scattering response whereas one UWB antenna intercepts a part of the incident electromagnetic wave and the other UWB antenna radiates it as the response that is read using the magnitude of the transmission coefficient $S_{21\text{tag}}$ between cross-polarized transmitting and receiving horns. Main benefits of this solution consist in low crosstalk due to the cross-polarized transmitting and receiving antennas and a relatively high power level of the response reached because of the integrated UWB antennas, enabling to read the tag information from the magnitude $\left| S_{21\text{tag}} \right|$ only. Nevertheless, the UWB antennas cause a significant increase in tag size, which is the fundamental drawback of this solution. Moreover, the tag’s edge has to be situated right opposite the transmitting and receiving antennas, which is very inconvenient for the identification of some objects.

This paper proposes a novel implementation of a frequency-domain reading method for chipless tags or sensors that are based on a single scalar measurement of the magnitude of the transmission coefficient $S_{21\text{tag}}$ between two simple electrically small, yet not necessarily well impedance matched, transmitting and receiving antennas. Thus, the idea of evaluation of the tag scattering response based on (8) is also adopted. The suitability of using (8) over (9) for the implementation of the reading method is discussed in detail in connection with a
The transmission coefficient of the transmitting/receiving antenna or the magnitude of the reflection coefficient at a measurement of the magnitude of the reflection coefficient \( S_{11\text{tag}} \) of the transmitting/receiving antenna or the magnitude of the transmission coefficient \( S_{21\text{tag}} \) between the transmitting and receiving antennas by the presence of the tag, i.e., only one scalar measurement is required. The second goal is the employment of simple, compact, low-cost transmitting and receiving antennas. The third goal is compactness of a tag, which leads to a tag whose information encoding structure is also employed for radiation of the scattering response.

### A. Measurement Approach

To meet the first goal, an analysis of the measurement approaches is done. In principle, the monostatic or bistatic measurement approaches are possible and they correspond to a measurement of the reflection coefficient \( S_{11\text{tag}} \) or the transmission coefficient \( S_{21\text{tag}} \), respectively (see Fig. 1).

The signal flow paths in the monostatic and bistatic measurement approaches are described further. A signal \((\text{In.})\) carried by a continuous electromagnetic wave, whose frequency is swept in the frequency band of interest, enters into the input of the transmitting antenna \(1\). In real conditions, a part of this signal \((\text{Refl.})\) is reflected back due to mismatching of the antenna \(1\). The other part is radiated from the antenna \(1\) as a transmitted signal \((\text{Tr.})\) to a chipless tag and invokes its scattering response. A scattered signal \((\text{Rec.} 1)\) enters the transmitting/receiving antenna \(1\) in the monostatic measurement approach. In the case of the bistatic measurement approach, a scattered signal \((\text{Rec.} 2)\) enters the receiving antenna \(2\). The antenna \(2\) also receives a signal \((\text{Cross.})\) due to a crosstalk between both antennas. The reflected \((\text{Refl.})\) and crosstalk \((\text{Cross.})\) signals are unwanted parasitic components of output signals \((\text{Out.} 1 \text{ and Out.} 2)\) of the antennas \(1\) and \(2\) that are very unpleasant, namely, in scalar measurements. It is useful to know their levels compared to the signals \((\text{Rec.} 1 \text{ or Rec.} 2)\) containing the tag scattering response.

Hereafter, a quantitative comparison of both measurement approaches is performed taking into account the presented implementation of the reading method. Let us consider identical antennas \(1\) and \(2\) represented by \(R100\) open-ended rectangular waveguide stubs (see Section II-C). The antennas \(1\) and \(2\) receive powers \(P_{R1}\) and \(P_{R2}\) corresponding to the signals \((\text{Rec.} 1 \text{ and Rec.} 2)\) containing the tag scattering response. The problem can be simplified assuming

\[
P_R = P_{R1} = P_{R2}
\]

and the equal gain \(G\) of both antennas, i.e., the directional dependence of the tag scattering response and the gain of both antennas is neglected. Then, the power \(P_R\) can be related to the power \(P_T\) corresponding to the signal \((\text{Tr.})\) transmitted by the antenna \(1\) with the help of the well-known radar equation [22]

\[
P_R = \frac{\sigma_{\text{tag}} \lambda^2 G^2}{(4\pi)^3 R^4} P_T
\]

where \(R\) is a distance between the antennas and the tag with the RCS equal to \(\sigma_{\text{tag}}\). Note that this equation is valid for the far-field region of the antenna. Let us suppose the tag formed only by a single half-wavelength dipole resonator. Its RCS \(\sigma_{\text{tag}}\) can be determined using [22]

\[
\sigma_{\text{tag}} = 0.86\lambda^2.
\]

It means that \(\sigma_{\text{tag}} = -31.1 \text{ dBSm}\) in the case of \(\lambda = 30 \text{ mm}\), which corresponds to reasonable dimensions of the tag. Similar value \(\sigma_{\text{tag}} = -31.5 \text{ dBSm}\) is also determined by a simulation using the MoM full-wave simulator Zeland IE3D (IE3D), which is suitable for fast modeling of 2.5-D sandwich planar structures. The gain \(G = 5.8 \text{ dB}\) of the selected antennas (see Section II-C) and their reflection coefficient \(|S_{11A}| = -10 \text{ dB}\) are determined with the help of the finite integral full-wave simulator CST Microwave Studio (CST), which is applicable for the modeling of 3-D general structures. The distance \(R = 50 \text{ mm}\) is chosen corresponding well to the far-field region of the antennas. Under these conditions, (11) results in

\[
P_R = 1.05\sigma_{\text{tag}} P_T.
\]
Both measurement approaches shown in Fig. 1 can be represented by the well-known signal flow graph suggested, e.g., in [23] and widely used in correction methods for vector network analyzers (VNAs) (see Fig. 2). The graph visualizes in the formalism of S-parameters signal flow paths between the reference planes 1, 1’, 2, and 2’, which are also marked in Fig. 1.

The antennas 1 and 2 are characterized by S-parameters $S_{11A}$, $S_{12A}$, $S_{21A}$, $S_{22A}$, i.e., every antenna is considered as a two-port between its connector and the free space (FS). The connectors of the antennas 1 and 2 correspond to the reference planes 1 and 2. The reference planes 1’ or 2’ are fictitious boundaries between the antenna 1 or 2 and the FS. For the antenna 2, the S-parameters $S_{12A}$ and $S_{22A}$ are displayed for simplicity only since the signal flows only from the reference plane 2’ to the reference plane 2. The crosstalk between both antennas related to the reference planes 1 and 2 is modeled by the S-parameter $S_{12iso}$ in accordance with the formalism established in Section I, i.e., the transmission between the antennas 1 and 2 when the tag is not present. The path between the antennas 1 and 2 including the tag is also considered as a two-port related to the reference planes 1’ and 2’ and it is characterized by S-parameters $S_{11D}$, $S_{12D}$, $S_{21D}$, and $S_{22D}$ corresponding to S-parameters of the device under test in common VNA measurements.

$S_{11A}$ represents the reflection at the connector of the antenna 1 (reference plane 1) when the tag is not present, i.e., it is the reflection coefficient $S_{11iso}$ in the sense of the formalism of Section I. $S_{21A}$ is the transmission between the connector of the antenna 1 (reference plane 1) and the FS (reference plane 1’). $S_{12A}$ is transmission between the FS (reference plane 1’) and the connector of the antenna 1 or 2 (reference plane 1 or 2), respectively, and it is equal to $S_{21A}$ due to reciprocity. $S_{22A}$ is the reflection of the antenna 1 or 2 at the reference plane 1’ or 2’, respectively.

$S_{11D}$ includes transmission from the antenna 1 (reference plane 1’) to the tag, scattering due to the tag and transmission from the tag to the antenna 1 (reference plane 1’). $S_{21D}$ includes transmission from the antenna 1 (reference plane 1’) to the tag, scattering due to the tag and transmission from the tag to the antenna 2 (reference plane 2’). $S_{12D}$ includes transmission from the antenna 2 (reference plane 2’) to the tag, scattering due to the tag and transmission from the tag to the antenna 1 (reference plane 1’). $S_{22D}$ includes transmission from the antenna 2 (reference plane 2’) to the tag, scattering due to the tag and transmission from the tag to the antenna 2 (reference plane 2’).

Magnitudes $|S_{11D}|$, $|S_{12D}|$, $|S_{21D}|$, and $|S_{22D}|$ are related to the power balance expressed by (10) and (13) by

$$|S_{11D}| = |S_{12D}| = |S_{21D}| = |S_{22D}| = \sqrt{1.05\sigma_{tag}}. \quad (14)$$

Using the signal flow graph in Fig. 2, the measured S-parameters $S_{11tag}$ and $S_{21tag}$ corresponding to the monostatic and bistatic measurement approaches can be then expressed as

$$S_{11tag} = S_{11iso} + S_{12A}S_{21A}S_{11D}(1 - S_{22A}S_{22D}) + S_{12A}S_{21A}S_{22A}S_{12D}S_{21D}$$

$$S_{21tag} = S_{21iso} + S_{12A}S_{21A}S_{21D}$$

Let us consider for the antennas 1 and 2 represented by the R100 open-ended waveguide stubs (see Section II-C)

$$|S_{11A}| = |S_{22A}| = -10 \text{ dB} = 0.32$$

and corresponding

$$|S_{12A}| = |S_{21A}| = \sqrt{1 - |S_{11iso} + S_{21iso}|^2} \\ \approx \sqrt{1 - |S_{11iso}|^2} \\ = -0.45 \text{ dB} = 0.95$$

considering $|S_{11A}| \gg |S_{21iso}|$. Supposing the previously mentioned RCS $\sigma_{tag}$ of the tag

$$\sigma_{tag} = -31.1 \text{ dBsm} = 7.7 \cdot 10^{-4} \text{ m}^2$$

it yields through (14)

$$|S_{11D}| = |S_{12D}| = |S_{21D}| = |S_{22D}| = -31 \text{ dB} = 0.028. \quad (21)$$

Then, (17) results in

$$D \approx 1. \quad (22)$$

Neglecting further terms with insignificant magnitudes, (15) and (16) can be simplified using (18), (19), (21), (22), and (28) to

$$S_{11tag} = S_{11iso} + S_{12A}S_{21A}S_{11D} \\ = S_{11iso} + |S_{12A}|^2|S_{11D}|e^{j \arg (S_{12A}S_{11D})} \\ = 0.32e^{j \arg (S_{11iso})} + 0.025e^{j \arg (S_{12A}S_{11D})} \quad (23)$$

$$S_{21tag} = S_{21iso} + S_{12A}S_{21A}S_{21D} \\ = S_{21iso} + |S_{12A}|^2|S_{21D}|e^{j \arg (S_{12A}S_{21D})} \\ = 0.010e^{j \arg (S_{21iso})} + 0.025e^{j \arg (S_{12A}S_{21D})}. \quad (24)$$

In the special case, when the tag is not present, there is only the FS and no signal (Rec. 1 or Rec. 2) is received by the antenna 1 or 2. Then, it holds true

$$|S_{11D}| = |S_{12D}| = |S_{21D}| = |S_{22D}| = 0$$

$$S_{11tag} = S_{11iso}$$

$$S_{21tag} = S_{21iso}. \quad (27)$$
Fig. 3. Simulation of transmission between transmitting and receiving antennas in setup according to Fig. 8 by the presence of different objects in transmission path.

For the monostatic measurement approach, the unwanted term \( S_{11iso} \) in (23) due to mismatching of the antenna 1 has about ten times greater magnitude than the term of interest \( S_{12A}S_{21A}S_{11D} \) corresponding to the tag scattering response. Therefore, the response is very difficult to distinguish on the background of the disturbing reflection, namely, in the scalar measurement.

The situation is much better in the bistatic measurement approach. The magnitude of the unwanted term \( S_{21iso} \) in (24) is estimated using the simulations in CST as (see Fig. 3)

\[
|S_{21iso}| = -40 \text{ dB} = 0.010. \tag{28}
\]

It means, contrary to the monostatic measurement approach, in this bistatic measurement approach, the unwanted crosstalk term \( S_{21iso} \) has several times smaller magnitude compared to the term of interest \( S_{12A}S_{21A}S_{21D} \) corresponding to the tag scattering response. This dynamic makes possible to distinguish the response using the scalar measurement and, thus, a suitable reading method for the tag can be based on this approach. This theoretical conclusion is also confirmed by simulations in CST (see Fig. 3). The \( |S_{21tag}| \) trace by the presence of the dipole in the FS is clearly distinguishable above the frequency 9 GHz from the special case \( |S_{21tag}| = |S_{21iso}| \) trace of the FS corresponding to the crosstalk.

**B. Interferometric Improvement**

The maximum around the frequency 9.5 GHz of the scattering response of the tag represented by the dipole resonator displayed in Fig. 3 is very flat in frequency due to the low \( Q \) factor of the resonator. This makes impossible to create a tag consisting of more resonators with different resonant frequencies in the R100 rectangular waveguide frequency band to encode a bit word with more bits. Therefore, it is essential to increase the frequency selectivity of the individual resonators.

Initially, the solution was sought experimentally. It led to the design of the tag described in Section III. It was discovered that, if a limited ground plane is placed in a proper distance behind the dipole resonator, a significant frequency selective drop in the magnitude \( |S_{21tag}| \) measured in accordance with Figs. 1, 2, and 8 can be observed. Then, the problem was analyzed using simulations. A tag consisting of one dipole resonator backed by the ground plane is considered in the arrangement according to Fig. 9(a) and (b). However, a single resonator in a middle position is taken into account only. A dipole resonator strip is supported by the Rogers RO4350 substrate with the relative permittivity \( \varepsilon_r = 3.66 \), loss tangent \( \tan \delta = 0.003 \), and thickness of 0.1 mm. A foam layer with the relative permittivity \( \varepsilon_r = 1.3 \), loss tangent \( \tan \delta = 0.02 \), and thickness of 1 mm is inserted between the ground plane and substrate. The magnitude \( |S_{21tag}| \) obtained through simulations in CST is shown in Fig. 3 and, for completeness, the case of the ground plane without the dipole resonator is also depicted. The influence of the ground plane dimensions on the RCS \( \sigma_{tag} \) simulated in IE3D is displayed in Fig. 4. It can be observed that an optimally spaced ground plane with the dimensions 30 × 30 mm² backing the dipole resonator creates about 15-dB deep selective drop at the frequency 10 GHz. Such drop is able to form a single bit of the bit word which requires a reasonable frequency bandwidth.

Some physical background and explanation of the drop presence in the RCS \( \sigma_{tag} \) should be given. It can be explained using an interferometric approach. The incident electromagnetic wave induces a current in the dipole resonator, which results in a backward wave scattered from the resonator. The incident wave also reflects from the ground plane forming the second backward scattered wave interfering with the first backward wave. If these waves are out-of-phase, a destructive interference occurs, which results in the suppression of the RCS \( \sigma_{tag} \) and also the corresponding magnitude \( |S_{21tag}| \) around the interference frequency (see Figs. 3 and 4).

This concept is related to a structural mode scattered signal corresponding to the ground plane and an antenna mode scattered signal corresponding to the dipole resonator, which is considered in [24]. The phenomenon is also partially analyzed in [25] using MATLAB where superposition of the two individual RCS is studied. It is also supposed in [26] that the intensity of the electric field \( E_R \) of the total scattered wave can be written as

\[
E_R = E_G + E_D \tag{29}
\]

where \( E_G \) is the intensity of the electric field of the wave scattered by the ground plane and \( E_D \) is the intensity of the electric field of the wave scattered by the dipole resonator. If these intensities are out-of-phase and both have similar magnitudes,
i.e., $E_G \approx -E_D$, a significant drop in the RCS $\sigma_{\text{tag}}$ occurs. Note that the tag structure described in Section III-A consisting of the dipole resonator array backed by the ground plane is advantageously optimized with respect to this condition with the help of the parametric analysis in the frequency domain (see Fig. 4). Therefore, it is not necessary to separate the signals corresponding to the structural and antenna modes in the time domain as in [24]. On the other hand, the parametric approach at a single frequency is not able to explain the shapes of traces of the simulation results displayed in a given frequency band in Fig. 4. This is why a better explanation based on an equivalent circuit of the tag is suggested.

The RCS $\sigma_{\text{tag}}$ is defined as

$$\sigma_{\text{tag}} = \lim_{R \to \infty} \left( \frac{4 \pi R^2 W_R(r = R)}{W_T(r = 0)} \right)$$  \hspace{1cm} (30)

where $W_R$ and $W_T$ are the respective active power densities of scattered and incident waves [22], [27]. The distances $r$ and $R$ are measured from the tag [see Figs. 1 and 8(a)]. In the definition (30), the density $W_R$ is considered in the distance $r = R \to \infty$ where it corresponds to the scattered far field. In reality, it is a distance of the receiving antenna in which the scattered field can be considered to have the character of the far field. The RCS $\sigma_{\text{tag}}$ is defined in such a way that it expresses the objectively existing and measurable density $W_R$ at the distance $R$ produced by the tag scattering response. It is based on an equivalent scenario supposed that this density $W_R$ is produced by an isotropically scattered wave of the total power $P_T$ from the place of the tag where $r = 0$. On the other hand, the power $P_R$ also corresponds to the density $W_T$ of the incident quasi-planar wave intercepted in the place of the tag by an area equal to the RCS $\sigma_{\text{tag}}$. The condition $r = R \to \infty$ results from experimental demands on measurement in the far field. However, in theory, the relation (30) remains valid for an arbitrary value of the distance $R$.

Formally, the relation (30) can be divided (normalized) by the RCS $\sigma_{\text{norm}} = 1$ m$^2$ and written as

$$\frac{\sigma_{\text{tag}}}{\sigma_{\text{norm}}} = \lim_{R \to \infty} \left( \frac{4 \pi R^2 W_R(r = R)}{\sigma_{\text{norm}} W_T(r = 0)} \right) = \frac{P_R}{P_T} = |S_{11\text{cir}}|^2$$  \hspace{1cm} (31)

where $P_T$ corresponds to the power intercepted by an area equal to 1 m$^2$ due to the density $W_T(r = 0)$ and the ratio $P_R/P_T$ can be considered as the squared magnitude of a reflection coefficient $S_{11\text{cir}}$ characterizing the tag in the FS. It is numerically equal to the RCS $\sigma_{\text{tag}}$ and it is not dependent on the distance $R$.

An equivalent circuit of the tag placed in the FS is suggested based on the coefficient $S_{11\text{cir}}$ defined by (31) (see Fig. 5). The tag, characterized by its impedance $Z_{\text{tag}}$, is connected at the end of a line with the FS characteristic impedance $Z_0 = 120 \pi$. Power waves $P_T$ and $P_R$ are represented by well-known normalized voltage waves $a$ and $b$ used for the definition of S-parameters. They are related by the formulas

$$P_T = \frac{1}{2} |a|^2 \quad P_R = \frac{1}{2} |b|^2 \quad |S_{11\text{cir}}| = \frac{|b|}{|a|}.$$  \hspace{1cm} (32)

As mentioned, the magnitude $|S_{11\text{cir}}|$ in (31) is not dependent on the distance $R$. Similarly, in this equivalent circuit, the magnitude $|S_{11\text{cir}}|$ is not dependent on the length of the line. Therefore, it is possible to analyze the properties of the tag consisting of a single dipole resonator backed by a limited ground plane in the reference plane of the tag corresponding to the plane of the dipole resonator.

The structure of the impedance $Z_{\text{tag}}$ corresponding to the coefficient $S_{11\text{cir}}$ is shown in Fig. 5. The dipole resonator is characterized by a parallel resonant circuit (capacitance $C_P$, inductance $L_P$, and resistance $R_P$). It is connected in series with (29) with a short-ended line of the characteristic impedance $Z$ and a resistor $R_R$ with the value given by

$$R_R = Z_0 - Z.$$  \hspace{1cm} (33)

Both the line and resistor model the limited ground plane. The electrical length $L_L$ of the line corresponds to the distance between the resonator and ground plane. In the particular case, which takes into consideration the results from Fig. 4, it accounts for the length $L_L = 14.2$ at 10 GHz. The basic characteristics of the equivalent circuit are mentioned in Table I.

The loaded Q factor and resonant frequency of the parallel resonant circuit can be properly set by the values $C_P$, $L_P$, and $R_P$. Therefore, the equivalent circuit makes it possible to include variations in the $Q$ factor of the resonator due to different dimensions of the ground plane. In addition, the circuit enables to involve a mirror dipole resonator due to the presence of the ground plane. Nonetheless, its influence is negligible and it is neglected for simplicity. For the same reason, neither the circuit takes into account the frequency dependence of the RCS $\sigma_{\text{tag}}$.

The proper values of the individual components of the equivalent circuit were determined experimentally in AWR Microwave Office to fit the results in Fig. 4. Fig. 6 displays the magnitude $|S_{11\text{cir}}|$ for different value settings summarized
Fig. 6. Simulation of reflection coefficient of equivalent circuit of tag for different settings of values of components given in Table II.

TABLE II
VALUES OF COMPONENTS OF EQUIVALENT CIRCUIT

<table>
<thead>
<tr>
<th>Ground plane size</th>
<th>Z (Ω)</th>
<th>C_E (pF)</th>
<th>L_E (nH)</th>
<th>R_E (Ω)</th>
<th>Q (-)</th>
</tr>
</thead>
<tbody>
<tr>
<td>No ground plane</td>
<td>0</td>
<td>6.2</td>
<td>41</td>
<td>18.4</td>
<td>7.1</td>
</tr>
<tr>
<td>7 × 15 mm²</td>
<td>25</td>
<td>15</td>
<td>16.9</td>
<td>44</td>
<td>41.6</td>
</tr>
<tr>
<td>15 × 15 mm²</td>
<td>34.9</td>
<td>15</td>
<td>16.9</td>
<td>47.2</td>
<td>44.6</td>
</tr>
<tr>
<td>30 × 30 mm²</td>
<td>66</td>
<td>15</td>
<td>16.9</td>
<td>58</td>
<td>53.8</td>
</tr>
<tr>
<td>40 × 40 mm²</td>
<td>113.1</td>
<td>25</td>
<td>10.1</td>
<td>41.6</td>
<td>65</td>
</tr>
<tr>
<td>50 × 50 mm²</td>
<td>169.6</td>
<td>41.6</td>
<td>6.1</td>
<td>30.4</td>
<td>79</td>
</tr>
</tbody>
</table>

Fig. 7. Simulation of reflection coefficient of equivalent circuit of tag for different settings of values of components given in Table II. Very good agreement with the results in Fig. 4 can be observed.

An explanation of behavior shown in Figs. 4 and 6 can be provided if the Smith chart is used for S-parameter $S_{11cir}$ display (see Fig. 7). It can be seen that the deep selective drops in the magnitude $|S_{11cir}|$ in Fig. 6 occur when the part of the resonant loop near the resonant frequency of the parallel resonant circuit passes close to the center of the chart. In principle, if the loop crosses the center that corresponds to zero reflection, a very deep drop in the magnitude $|S_{11cir}|$ can occur. The phase shift of the loops with respect to 180° results from the distance between the dipole resonator and the ground plane. It leads to the nonsymmetric shapes of the traces in Figs. 4 and 6 around the resonant frequency.

C. Implementation of Transmitting and Receiving Antennas

It is explained in Section II-A, the crosstalk between the transmitting antenna 1 and receiving antenna 2 without the presence of the tag has to be acceptably low and the RCS $\sigma_{tag}$ of the tag relatively high in order to clearly distinguish the tag scattering response in the magnitude of the transmission coefficient $S_{21tag}$.

Therefore, given the second goal, the open end of the rectangular waveguide represents a suitable candidate for the transmitting and receiving antennas. In addition, this type of antenna shows a favorably close distance of the far-field region, which decreases the influence of the surroundings and read tag on antenna characteristics.

The transmitting and receiving antennas are implemented as a pair of R100 open-ended rectangular waveguide stubs (shorted on the other end) with SMA connectors in the arrangement according to Fig. 8(a). The antennas in a complete setup of the reading method are shown in Fig. 8(b). The length of the stubs is 33 mm and the distance between their narrow walls equals 18 mm. This antenna arrangement ensures the
acceptably low crosstalk corresponding to the special case $|S_{21tag}| = |S_{21iso}| = -40$ dB without the presence of the tag. It has also very compact total size of $60 \times 12 \times 34$ mm$^3$ (without SMA connectors and adapters). The transmitting and receiving antennas are backed by an absorber that improves the performance for the greater distance $R$; however, even without the absorber, the tag scattering response is readable. The absorber prevents an additional path of propagation of the transmitted signal with multiple reflections between the tag and background of the antennas before it reaches the receiving antenna.

The distance $r_{far}$ of the far-field region of the antennas is evaluated according to the formula [22]

$$r_{far} = \frac{2D^2}{\lambda} = \frac{2D^2 f}{3 \cdot 10^8}$$

(34)

where $D = 25$ mm is the dimension of the diagonal of the R100 rectangular waveguide cross section and $f$ is the frequency. In the considered frequency band $f \in [7$ GHz, 11 GHz $]$ for the evaluation of the tag scattering response, it implies from (34) for the distance $r_{far} \in [29$ mm, 46 mm $]$. The analysis of the measurement approach in Section II-A takes into account the distance $R = 50$ mm $> r_{far}$, whereas the experimental verification in Section IV is performed for the distance $R = 20$ mm $< r_{far}$. The operation at the latter distance is also acceptable since the evaluation of the response is based on the drops in the response spectrum, and the exact character of the electromagnetic field is less important.

### III. TAGS

#### A. Topology

To meet the third goal, the chipless tag whose information encoding structure consists of a dipole resonator array backed by a limited ground plane is chosen (see [28]).

The dipole resonators are tuned to a set of different frequencies. The tag scattering response at each resonant frequency is used for encoding of a single bit of information. The presence or absence of a single resonator, which corresponds to the existence or nonexistence of a drop in the magnitude $|S_{21tag}|$ at the frequency of the given bit, encodes the logical 1 or 0, respectively.

The method is verified by means of tags operating in the frequency band from 6.8 to 10.5 GHz, encoding information of a 16-bit word (see Fig. 9). The tags encoding 16-bit words 1111111111111111, 1101111111110111, and 1010101010101010 are used. The 16-bit word 1111111111111111 is considered as a reference [see Fig. 9(b)]. Two manners of zero bit encoding are applied and compared. The first one implements the absence of the single dipole resonator, which encodes the logical zero of the given bit by removing or interrupting corresponding dipole resonators in (c) and (d) or (e) and (f), respectively. Dimensions are stated in millimeters.

#### B. Simulation

The tags were simulated in IE3D. The scattering response of the tag is read in the simulation with the help of its RCS $\sigma_{tag}$, which is, in this case, almost directly proportional to $|S_{21tag}|^2$. This claim holds true since the crosstalk between the chosen transmitting and receiving antennas is acceptably low and the RCS $\sigma_{tag}$ of the chosen tags is relatively high. Thus, the main contribution to $|S_{21tag}|^2$ by the presence of the tag is caused dominantly by its scattering response. Furthermore, the RCS $\sigma_{tag}$ represents the intrinsic property of the tag and it is not necessary to simulate the transmitting and receiving antennas. Due to that, the simulation is speeded up.

Every graph in Figs. 10 and 11 contains three RCS $\sigma_{tag}$ traces corresponding to different 16-bit words encoded by
Fig. 10. Scattering response of tags simulated by means of RCS: tag with 16-bit word 1111111111111111 in comparison to tags with 16-bit words 110111111110111 and 1010101010101010; dipole resonators are in the longitudinal position with respect to incident wave polarization; zero bit resonators are (a) removed and (b) interrupted.

Fig. 11. Scattering response of tags simulated by means of RCS: tag with 16-bit word 1111111111111111 in comparison to tags with 16-bit words 1101111111110111 and 1010101010101010; dipole resonators are in 30° rotated position with respect to incident wave polarization; zero bit resonators are (a) removed and (b) interrupted.

the tags displayed in Fig. 9. The black solid trace represents the tag with the reference 16-bit word 1111111111111111 from Fig. 9(b), whereas the red dashed trace and blue dashed-and-doted trace symbolize the tags with the 16-bit words that comprise two or more zero bits according to configurations in Fig. 9(c)–(f). Figs. 10 and 11 depict the scattering response of the tags simulated by means of the RCS $\sigma_{\text{tag}}$ in the longitudinal position of the dipole resonators with respect to the incident wave polarization and in the 30° rotated position, respectively. The longitudinal position is optimal for excitation of the dipole resonators.

For the case of the tag with the 16-bit word 1111111111111111, the RCS $\sigma_{\text{tag}}$ contains the drops at all frequencies corresponding to all 16 bits as it is expected according to the system of information encoding described in Section III-A. For the tags with two or more zero bits, the drops in the RCS $\sigma_{\text{tag}}$ at the frequencies corresponding to the zero bits are apparently missing and the drops corresponding to the one bits remain almost unchanged with respect to the reference tag. In the case of the encoding, the zero bits by removing the dipole resonators [see Fig. 9(c) and (d)] the frequency stability of the RCS $\sigma_{\text{tag}}$ corresponding to the one bits is very satisfying with respect to the reference 16-bit word. Nevertheless, this feature can be additionally improved by the encoding the zero bits by interruption of the dipole resonators according to Fig. 9(e) and (f).

IV. EXPERIMENTAL VERIFICATION

The proposed concept was verified through an experiment. The scattering response of the tags was determined by the measurement of the magnitude $|S_{21}^{\text{tag}}|$ between the transmitting and receiving antennas by the presence of the tag, as it is described in Section II-A and depicted in Figs. 1, 2, and 8. The Agilent PNA E8364A VNA in the frequency band from 7 to 11 GHz was used. The calibration was done at the reference planes of the SMA connectors of the antennas. Although the vector data were obtained, only the magnitude $|S_{21}^{\text{tag}}|$ was employed for the reading of the tag scattering response.

The crosstalk between the transmitting and receiving antennas depicted in Fig. 8, which is measured through the special case $|S_{21}^{\text{tag}}| = |S_{21}^{\text{iso}}|$ without the presence of the tag, is lower than $-40$ dB in the whole considered frequency band (see Fig. 12). These antennas are not well matched and, thus, inappropriate for, e.g., the method employing the monostatic approach based on the measurement of the antenna reflection [14]. However, their reflection coefficient, which is measured through the special case $|S_{11}^{\text{tag}}| = |S_{11}^{\text{iso}}|$ without the presence of the tag, is around $-10$ dB (see Fig. 12) and
Fig. 13. Photograph of tested tags with 16-bit words constituted by dipole resonator array backed by limited ground plane.

Fig. 14. Scattering response of tags measured through transmission coefficient between transmitting and receiving antennas by the presence of tag: tag with 16-bit word 1111111111111111 in comparison to tags with 16-bit words 1101111111110111 and 1010101010101010; dipole resonators are in the longitudinal position with respect to incident wave polarization; zero bit resonators are (a) removed and (b) interrupted.

Fig. 15. Scattering response of tags measured through transmission coefficient between transmitting and receiving antennas by the presence of tag: tag with 16-bit word 1111111111111111 in comparison to tags with 16-bit words 1101111111110111 and 1010101010101010; dipole resonators are in 30° rotated position with respect to incident wave polarization; zero bit resonators are (a) removed and (b) interrupted.

it influences the quality of the reading of the tag scattering response through the magnitude $|S_{21\text{tag}}|$ negligibly.

The chipless tags considered for the verification of the proposed reading method and described in Section III-A were manufactured (see Fig. 13). In the experimental setup, they were placed in the distance $R = 20$ mm above the apertures of the transmitting and receiving antennas in the longitudinal position of the dipole resonators with respect to the polarization of the radiated wave as well as in the 30° rotated position from this position [see Fig. 8(a)]. The tags operate in the frequency band from 7.1 to 10.5 GHz.

The measured magnitude $|S_{21\text{tag}}|$ between the transmitting and receiving antennas by the presence of the tag in the longitudinal position and in the 30° rotated position is shown in Figs. 14 and 15, respectively. In both positions, the tag scattering response read through the magnitude $|S_{21\text{tag}}|$ with the significant drops encoding the one bits is confirmed. The frequency and amplitude stability of the response discussed in Section III-B for both manners of encoding of the zero bits was confirmed as well. Small frequency shifts of the drops and maxims that can be observed in Figs. 14 and 15 in comparison to the simulated RCS $\sigma_{\text{tag}}$ in Figs. 10 and 11 are caused by manufacturing tolerances and employment of the foam layer. The simulated RCS $\sigma_{\text{tag}}$ shows more stable level of the drops and maxims than the measured magnitude $|S_{21\text{tag}}|$. It is caused by the inclusion of the reflection and gain of the transmitting and receiving antennas, which are frequency variable, into the magnitude $|S_{21\text{tag}}|$, whereas the RCS $\sigma_{\text{tag}}$ is an intrinsic property of the tag, which does not include information about other parts of the RFID system and shows for the given tag mentioned frequency stable behavior.

The quantity $|S_{21\text{tag}} - S_{21\text{iso}}|$ measured through the transmission coefficients $S_{21\text{tag}}$ and $S_{21\text{iso}}$ between the transmitting and receiving antennas by the presence and without the presence of the tag in the longitudinal position and in the 30° rotated position is displayed in Figs. 16 and 17, respectively. Comparison with the magnitude $|S_{21\text{tag}}|$ in Figs. 14 and 15 shows a good agreement. Thus, it demonstrates that the primary assumption $|S_{21\text{tag}}| \gg |S_{21\text{iso}}|$ for the implementation of the presented reading method holds true.

It is apparent that the presented reading method based on the measurement of the magnitude $|S_{21\text{tag}}|$ provides the ability in distinguishing of the zero and one bits as the methods based
Fig. 16. Scattering response of tags measured through the difference in transmission coefficients between transmitting and receiving antennas by the presence and without the presence of tag: tag with 16-bit word 1111111111111111 in comparison to tags with 16-bit words 1101111111011110 and 1010101010101010; dipole resonators are in the longitudinal position with respect to incident wave polarization; zero bit resonators are (a) removed and (b) interrupted.

on the measurement of the quantity $|S_{21\text{tag}} - S_{21\text{iso}}|$ or the RCS $\sigma_{\text{tag}}$ since the detection of the presence or absence of a drop of the measured quantity at a given frequency is for all methods necessary only. However, the presented method requires advantageously only one scalar measurement.

The drops in the magnitude $|S_{21\text{tag}}|$ are well distinguishable for the distance $R$ between the tag and the apertures of the transmitting and receiving antennas from 20 to 80 mm. The distance $R$ which is acceptable for proper function of the proposed reading method is restricted by the following conditions. The lower bound is a distance where the tag is far enough from the transmitting antenna to be properly irradiated by the wave transmitted by this antenna. The complete dipole resonator array and backing ground plane of the tag have to be irradiated to produce a proper scattering response for which the tag is designed. However, when the tag is too close to the transmitting antenna, only a part of the tag can be irradiated. In addition, the properties of the transmitting and receiving antennas can be degraded when the tag is too close. The upper bound is a distance when the signal (Rec. 2) produced by the tag scattering response at the receiving antenna is comparable to the signal (Cross.) of the crosstalk between the transmitting and receiving antennas (see Fig. 1).

The immunity of the reading method to the presence of perturbing elements in the vicinity of the tag was tested using a metal sheet (see Fig. 18). Even for the close proximity of the sheet to the edge of the tag, the tag scattering response is still readable.

Fig. 17. Scattering response of tags measured through difference in transmission coefficients between transmitting and receiving antennas by presence and without presence of tag: tag with 16-bit word 1111111111111111 in comparison to tags with 16-bit words 1101111111011110 and 1010101010101010; dipole resonators are in 30° rotated position with respect to incident wave polarization; zero bit resonators are (a) removed and (b) interrupted.

Fig. 18. Setup for the testing of immunity of reading method to the presence of perturbing elements in vicinity of tag.

V. CONCLUSION

We carried out the analysis of frequency-domain reading methods for chipless tags based on monostatic and bistatic measurement approaches. An original explanation of properties of a tag consisting of a dipole resonator array backed by a limited ground plane was provided using an equivalent circuit.

The novel implementation of the frequency-domain reading method for chipless tags based only on a single scalar measurement of the magnitude of the transmission coefficient between two antennas by the presence of the tag was proposed and experimentally verified in the frequency band from 7 to 11 GHz.
The applied antennas are low-cost, simple, compact, and formed by the R100 open-ended rectangular waveguide stubs. The distance of the tag from the antennas as well as the tag partial rotation with respect to the optimal mutual polarization of the tag and antennas are not critical. The drops and maxims of the measured frequency dependencies of the magnitude of the transmission coefficient, which encode the tag information, are clearly distinguishable and do not need any sophisticated method to be revealed.

Thanks to the above-mentioned features, this scalar reading method in conjunction with the tags based on the dipole resonator array backed by the limited ground plane seems to be suitable for low-cost chipless RFID systems working in a real environment outside the anechoic chamber.

REFERENCES


Jan Kracek (M’09) was born in Slany, Czech Republic, in 1984. He received the Ing. degree in telecommunications and radio engineering and the Ph.D. degree in radioelectronics from Czech Technical University in Prague, Czech Republic, in 2008 and 2016, respectively.

He is currently a Researcher with the Department of Electromagnetic Field, Faculty of Electrical Engineering, Czech Technical University in Prague. His current research interests include theory of electromagnetic fields, wireless power transfer, and antennas.

Milan Svanda was born in Prague, Czech Republic, in 1982. He received the Ing. and Ph.D. degrees in radioelectronics from Czech Technical University in Prague, Prague, in 2007 and 2011, respectively.

He is currently a Researcher with the Department of Electromagnetic Field, Faculty of Electrical Engineering, Czech Technical University in Prague. His current research interests include antennas operating in the close vicinity of the human body, low-profile and wearable antennas, UHF radio frequency identification (RFID), and tag antennas and sensors for special applications.

Karel Hoffmann (M’93–SM’99) was born in Prague, Czech Republic, in 1951. He received the Ing. degree (Hons.) and the Ph.D. degree from Czech Technical University in Prague, Prague, in 1974 and 1981, respectively.

From 1994 to 2002, he was an Associate Professor with the Faculty of Electrical Engineering, Czech Technical University in Prague, where he has been a Full Professor since 2002. His current research interests include the design of active and passive microwave integrated circuits, precise microwave measurements, and modeling of microwave components.