Ambient Backscatterers using FM Broadcasting for Low Cost and Low Power Wireless Applications

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Abstract—Nowadays, the explosive growth of Internet-of-Things-related applications has required the design of low-cost and low-power wireless sensors. Although backscatter radio communication is a mature technology used in radio frequency identification (RFID) applications, ambient backscattering is a novel approach taking advantage of ambient signals to simplify wireless system topologies to just a sensor node and a receiver circuit eliminating the need for a dedicated carrier source. This paper introduces a novel wireless tag and receiver system that utilizes broadcast frequency modulated (FM) signals for backscatter communication. The proposed proof-of-concept tag comprises of an ultra-low-power microcontroller (MCU) and a radio frequency front-end for wireless communication. The MCU can accumulate data from multiple sensors through an analog-to-digital converter, while it transmits the information back to the receiver through the front-end by means of backscattering. The front-end uses On-Off keying modulation and FM0 encoding on ambient FM station signals. The receiver consists of a commercial low-cost software defined radio which downconverts the received signal to baseband and decodes it using a suitable signal processing algorithm. A theoretical analysis of the error rate performance of the system is provided and compared to bit-error-rate measurements on a fixed transmitter-tag-receiver laboratory setup with good agreement. The prototype tag was also tested in a real-time indoor laboratory deployment. Operation over a 5 m tag-reader distance was demonstrated by backscattering information at 2.5 Kbps featuring an energy per packet of 36.9 μJ.

Index Terms—Ambient backscattering, backscatter communication, FM modulation, inkjet printing, internet-of-things (IoT), radio frequency (RF) identification (RFID) sensors, software-defined radio (SDR).

I. Introduction

RECENTLY, Internet-of-Things (IoT) has become the trend for networking every day objects so as to automate and make easier our daily lives. The most important challenge

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for IoT applications, is the minimization of the cost and energy dissipation of the sensors. Keeping the massive number of energy-constrained IoT sensors active with low cost designs is a key issue. Commercial radio modules used in IoT devices typically use power-hungry radio frequency (RF) chains including oscillators, mixers and digital-to-analog converters (DACs) resulting in significant limitations of the battery life. One particularly promising approach to alleviate these issues is backscatter communication [1] that allows IoT sensor nodes-tags to transmit data by reflecting and modulating an incident RF wave [2]. Communication using backscatter principles has been widely deployed in the application of radio frequency identification (RFID) for passive tags. The RF front-end part of the tags consists of only one RF transistor or switch. In this case the tags are battery-free and can operate using only RF power transmitted from a RFID reader resulting in communication ranges up to several meters [3], [4].

Ambient backscattering is an idea based on the bistatic backscatter philosophy and could constitute a very promising novel approach for extremely low power and low cost communication systems [5]. Utilizing ambient signals for backscattering, the communication scheme is simplified since it requires only a receiver eliminating the need for a continuous wave (CW) emitter. For example, ambient backscattering devices, such as RFID tags, can communicate with a reader by backscattering ambient RF signals that are available from multiple sources, such as mobile communications, television [5], and FM-AM radio [6] and WiFi [7] that are typically widely available in urban areas indoors and outdoors during day and night. In [5] two battery-free tags communicate via ambient backscatter TV signals. In [7], a WiFi backscatter deployment was designed to connect battery-free devices with off-the-shelf WiFi devices. Also a full-duplex ambient communication system was introduced in [8], where a WiFi access point (AP) can cooperate with backscatter IoT sensors with high data throughput. The use of ambient RF signals as the only source of both the CW carrier and the tag power is an extremely energy-efficient communication technique compared to the general backscattering technique.

In [9] preliminary results for a wireless sensor node prototype for agricultural monitoring were presented. The sensor node measures the temperature difference between the leaf and the atmosphere in order to estimate the water stress of a plant [10]. The tag modulates and reflects a fraction of the ambient frequency modulated (FM) station signals back to the reader as it is shown in Fig. 1.
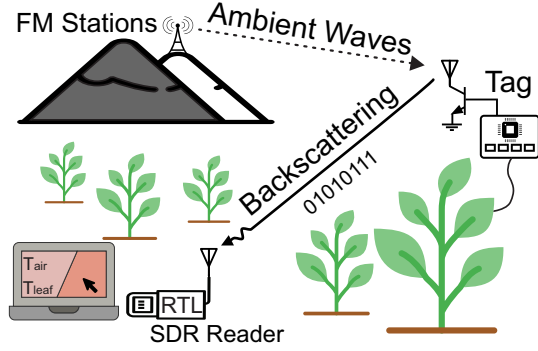
Fig. 1. Deployment of ambient backscattering in smart agriculture applications. Backscatter communication is achieved using ambient frequency modulated (FM) signals. The differential temperature (T_{leaf}- T_{air}) is measured by the tag-sensor and is transmitted back to a SDR receiver.
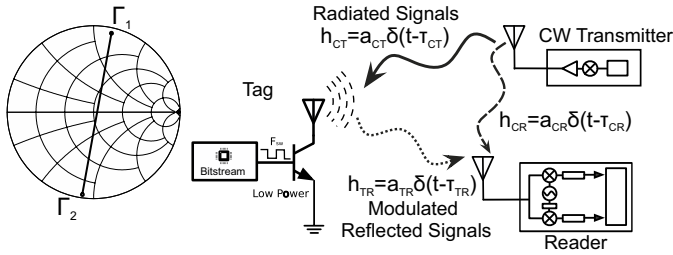

Fig. 2. Left: Two-state antenna S_{11} parameters on a Smith chart. Right: Bistatic backscatter principle. The emitter transmits a carrier signal and the tag reflects a small amount of the approaching signal back to the reader. The tag modulates the backscattered signal by changing the load connected to its antenna terminals resulting in a Γ_i change between two values (states).

This paper is an extensive presentation of the novel ambient FM backscatter monitoring system [9] with low complexity and low power. We propose an improved version of this system for generic environmental monitoring applications by designing an improved receiver algorithm. In addition to the receiver implementation, we provide additional details about the tag circuitry, a theoretical tag-receiver framework for the operation of the ambient backscatter system and a series of packet error rate (PER) and bit error rate (BER) measurements in a proof-of-concept indoor environment. The operation of the system prototype was demonstrated in the lab using an existing FM transmitter broadcasting 34 km away from the tag. Operation over a 5 m tag-to-reader distance was achieved by backscattering sensor data at 0.5, 1 and 2.5 Kbps bit rates.

Our work is different from [5], which first proposed ambient backscattering, in that it used ambient digitally modulated television signals (DTV) whereas the system proposed in this paper uses analog FM signals. Also, a moderately expensive software defined USRP-N210 radio ($\sim 1 - 5$ KUSD) used in [5] to receive and decode the signals whereas in our work a low cost Realtek (RTL) SDR (22 USD) was used. Recently, [6] also proposed ambient backscattering using FM signals but only for two state frequency-shift-keying (2-FSK) modulated signals. In our work we used on-off keying (OOK) modulation with FM0 encoding. In addition, an arbitrary waveform generator was used in [6] to generate the ambient FM signals contrary to signals from existing broadcast FM stations in this paper. Therefore, this paper takes into account all the

signal characteristics of an FM radio broadcasting and serves as the proof-of-concept for practical ambient backscatter deployments. The findings reported are equally useful for indoors and outdoors, where FM broadcasting signals are pervasive.

The structure of the paper is as follows: Section II provides the principles of ambient backscatter communication. Section III describes the design and implementation of the sensor node-tag parts. Section IV provides the theory and performance analysis of the FM ambient backscatter system. Section V discusses the hardware and software part of the low cost receiver. Section VI presents proof-of-concept experimental results, including an indoor demonstration and range measurements. Finally, section VII includes concluding remarks.

II. FM AMBIENT BACKSCATTERING

A. Backscatter Principles

A general bistatic backscatter system consists of three devices: a backscatter node (i.e. a tag), a reader and a CW emitter. The tag receives a CW carrier signal with frequency F_c and scatters a fraction of it back to the reader as shown in Fig. 2 (left). It superimposes the sensor information on top of the carrier by appropriately changing the load connected to its antenna terminals according to [11]:

$$\Gamma = \frac{Z_{\text{L}} - Z_{\text{a}}^*}{Z_{\text{L}} + Z_{\text{a}}}, \tag{1}$$

with Z_{L} and Z_{a} denoting the load and the tag antenna impedance. For binary modulation, the reflected signal is modulated by switching the load between two discrete values (Z_1 and Z_2) effectively resulting in two reflection coefficient values, (Γ_1 and Γ_2) over time. The 180 degrees difference between the two load values (Fig. 2, right) is necessary for maximization of backscatter performance. The reader captures the reflected signal at a frequency $f_c + \Delta F$ and an additional phase ϕ and then filters out the high frequency components. ΔF is the carrier frequency offset (CFO) between the emitter and the reader. According to [12] the received signal can be expressed in the following complex baseband form:

$$y_r(t) = n(t) + \frac{A_c}{2}e^{-j2\pi\Delta Ft}\big[\alpha_{\text{CR}}e^{-j\phi_{\text{CR}}}+ \\ s\alpha_{\text{CT}}\alpha_{\text{TR}}e^{-j\phi_{\text{CTR}}}\Gamma(t - \tau_{\text{TR}})\big], \tag{2}$$

where A_c is the carrier amplitude, $\alpha_{\text{CR}}, \alpha_{\text{CT}}, \alpha_{\text{TR}} \in \mathbb{R}$ and $\phi_{\text{CR}}, \phi_{\text{CTR}} \in [0, 2\pi)$. Moreover τ_{TR} is the time delay constant of the tag-reader channel. Term s is related to the tag scattering efficiency and tag antenna gain at a given direction. The term $\alpha_{\text{CR}}e^{j\phi_{\text{CR}}}$ defines the component which depends on the emitter-to-reader channel (h_{CR} in Fig. 2). The tag signal is a direct function of Γ over time and the term $\alpha_{\text{CT}}\alpha_{\text{TR}}e^{j\phi_{\text{CTR}}}$ scales and rotates the modulated part of the tag signal. This term depents on the transmiter-to-tag and tag-to-reader channel parameters (h_{CT} and h_{TR} in Fig. 2). Finally, $n(t)$ is the complex thermal Gaussian noise at the receiver.

B. FM Broadcasting Operation

The FM broadcasting technology was first utilized in 1940 radio-audio transmissions and nowadays FM radio broadcasts
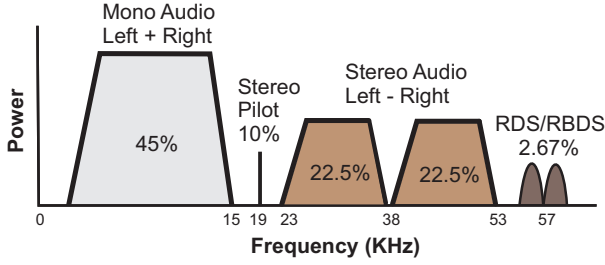
Fig. 3. Baseband Spectrum of a generic modern-day FM audio station. The signal contains Left (L) and right (R) channel information (L+R) for monophonic and stereo reception.

take place between radio frequencies of 88 MHz to 108 MHz with a channel bandwidth of 200 kHz. Each FM station uses frequency modulation in order to transmit the audio signals and the information signals varying the frequency of a carrier wave accordingly. A typical FM output signal is given by the following equation [13]:

$$x_{\text{FM}}(t) = A_{\text{c}} \cos \left[2\pi f_{\text{c}} t + 2\pi K_{\text{VCO}} \Delta f \int_0^t m(x) \mathrm{d}x \right] \quad (3)$$

where $m(x)$ is the baseband message signal, and Δf is the frequency deviation which is equal to the maximum frequency shift from f_{c} while K_{VCO} is the gain of the transmitter's voltage-controlled oscillator (VCO). Generally, it is not straightforward to analyze the properties of $x_{\text{FM}}(t)$ due to its non-linear dependence to the $m(x)$. The baseband message signal of a typical FM station as shown in Fig. 3 can be expressed as:

$$m(t) = A_0 \big[S_{\text{L}}(t) + S_{\text{R}}(t) \big] + A_1 \cos(2\pi f_1 t) + \\ A_0 \big[S_{\text{L}}(t) - S_{\text{R}}(t) \big] \cos(2\pi f_2 t) + A_2 \text{RDS}(t) \cos(2\pi f_3 t) \quad (4)$$

with $f_1 = 19$ KHz, $f_2 = 38$ KHz, $f_3 = 57$ KHz. The S_{L} and S_{R} define the time domain signals from the "stereo left" and "stereo right" channels, respectively, while $\text{RDS}(t)$ is the time domain signal of the Radio Data System (RDS) and Radio Broadcast Data System (RBDS). The gain factors A_0, A_1, and A_2 are used to appropriately scale the amplitude of S_{L} and S_{R} waveforms. As it can be easily observed in Fig. 3, the $0 - 15$ kHz part of the message signal consists of the left and right channel information [(Left)+(Right)] for monophonic sound. Stereophonic sound is the result of the amplitude modulation of the [(Left)-(Right)] message onto a suppressed 38 kHz subcarrier in the $23 - 53$ kHz region of spectrum. Furthermore, there is a 19 kHz pilot tone to enable receivers to recognize and decode the two stereo channels. Modern FM radio signals also include a 57 kHz subcarrier that carries RDS and RBDS data.

C. Ambient FM backscatter

In the case of typical ambient FM backscatter systems, incident "CW carrier" to the tag antenna is the signal in (3). The SDR receiver receives the superposition of this signal and the backscattered tag signal. Following the same procedure described in [12], but using a FM modulated carrier instead of
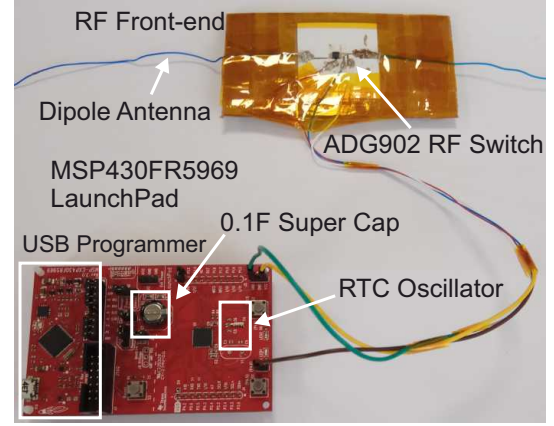

Fig. 4. The proposed tag prototype consists of a MSP430 development board [14] connected with a RF front-end board. The RF front-end consists of the ADG902 RF switch and was fabricated using inkjet printing technology on a paper substrate. A MCU digital output pin was connected with the control signal of the RF switch. The operation power of RF front-end was supplied by the MCU development board and the hole system was supplied by an embedded super capacitor for duty cycle operation.
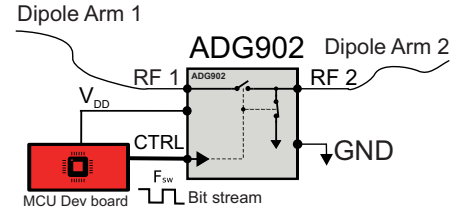

Fig. 5. Schematic of the RF switch utilized for the load modulation and of the dipole antenna arms.

a CW signal one may obtain the following complex baseband signal at the receiver:

$$y_{\text{amb}}(t) = n(t) + \frac{A_c}{2} e^{-j2\pi\Delta F t} \big[\alpha_{\text{CR}} e^{-j\phi_{\text{CR}}} e^{-jM(t-\tau_{\text{CR}})} \\ + s\alpha_{\text{CT}}\alpha_{\text{TR}} e^{-j\phi_{\text{CTR}}} e^{-jM(t-\tau_{\text{TR}})} \Gamma(t-\tau_{\text{TR}}) \big] \quad (5)$$

and

$$M(t) = 2\pi K_{\text{VCO}} \Delta f \int_0^t m(x) \mathrm{d}x. \quad (6)$$

The received signal y_{amb} contains the desired information Γ but also the carrier, FM modulation and frequency offset. The magnitude square of the received complex waveform will be formulated below in order to eliminate the frequency offset. If the desired magnitude square is formed, a component proportional to the desired information will be generated along with DC and other interference terms. We show theoretically and experimentally in sections IV and VI respectively that it is possible to successfully decode the signal provided there is a sufficiently high signal-to-noise-ratio (SNR).

III. TAG DESIGN

A. Tag

The main DIGITAL part of proposed tag is based on a 16-bit microcontroller (MCU) development board MSP-EXP430FR5969 [14] (Fig. 4). The development board is powered from a 0.1 F supercapacitor. The tag also includes

a real-time clock (RTC) to wake up the MCU from the "sleep" operation mode, where the current consumption of the board is $0.02\ \mu$A. The MCU generates 50% duty cycle pulses that control the RF switch, thus generating an OOK modulated backscattered signal. The OOK modulation is described in more detail in the subsection III-B. The MCU was programmed at 1 MHz clock speed using the internal local oscillator. The current consumption at 1 MHz was $126\ \mu$A at 2.3 V ($290\ \mu$W).

The MCU has a 16 channel, 12 bit analog-to-digital converter (ADC) which was used to read analog output signals from sensors. In this work the tag is programmed to read four analog inputs and the voltage level of the super capacitor. These analog inputs can be used to provide information from out to four sensors. This work focuses on the telecommunication aspect of the system and specific sensing examples form part of future work. When a tag wants to communicate with the reader, it sends a packet that contains the information of only one sensor each time. In [9] only two ADC inputs for two high precision, analog temperature sensors were used.

The backscatter communication of the tag is achieved with a separate RF front-end board. It consists of a 1.5 m wire dipole antenna in order to resonate within the FM band (95 MHz) and a single-pole, single-throw (SPST) RF switch ADG902 by Analog Devices. The circuit schematic of the front-end is provided in Fig. 5, while the fabricated prototype is shown in Fig. 4. The switch element varies the antenna load between two impedance values and is selected due to its low insertion loss (~ 0.5 dB @ 100 MHz) and high off isolation (~ 57.5 dB @ 100 MHz). The RF switch is a CMOS reflect-mode (i.e. not terminated) switch with high off-port VSWR and consumes less than $1\ \mu$A at 2.75 V [15]. It is driven by a digital output of the MCU as shown in Fig. 5. The power consumption of the RF switch follows the equation $\frac{1}{2}C_{RF}V_{DD}^2 F_{\text{sw}}$ which is the CMOS dynamic consumption [16]. The F_{sw} is the control switching frequency and C_{RF} the dynamic power dissipation capacitance at RF path when it is ON. For $F_{\text{sw}} = 2.5$ KHz, which equals to our maximun bit rate 2.5 Kbps, $V_{DD} = 3.3$ V, and $C_{RF} = 1.2$ pF (@ 1 MHz) the power consumption is estimated at 16.3 nA. As the data rate increases (switching speed) the DC consumption increases. The front-end printed circuit board (PCB) was fabricated using inkjet printing technology on a paper substrate. The characteristics of the the substrate was: $\epsilon_r = 2.9$, $\tan\delta = 0.045$ and substrate height $210\ \mu$m. The traces were printed with conductive silver nanoparticle (SNP) ink and conductive epoxy deposition was used in order to attach the switch to the substrate.

In order to minimize the average power consumption, a duty cycle operation was programmed where the tag was active only for a desired minimum period of time. The duty cycle operation was set using the RTC and the sleep mode of the MCU. A future challenge for the tag is to employ RF harvesters and solar cell s for powering as it is shown for example in [17], [18].

B. Telecommunication Protocol

The tag uses amplitude-shift keying (ASK) modulation to transmit its data via backscattering. More specifically, by

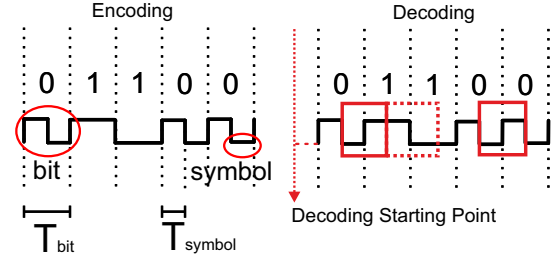
Fig. 6. Left: In FM0 encoding, the boundaries of the bits must always be different. Two sequential "on" or "off" correspond to the bit "1". Right: FM0 decoding technique, after shifting by T_{symbol}, receiver has to detect only two possible pulse shapes (line square or dash line square).
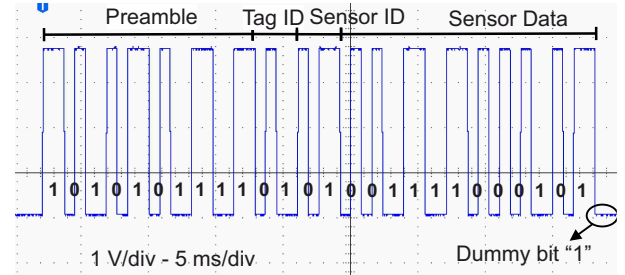

Fig. 7. Example of the oscilloscope-measured transmitted rectangular pulses (MCU output). The packet ("bit stream") consists of the Preamble, Tag ID, Sensor ID and Sensor Data bits and an extra dummy bit "1" at the end.

changing the RF switch states between "on" and "off" and backscattering the ambient FM broadband signals, a binary ASK modulated signal of OOK type can be created described by (5). Using OOK modulation, the information-containing received tag signal of (5) can be expressed as [12]:

$$\Gamma(t - \tau_{\text{TR}}) = \sum_{n=0}^{N-1} x_n \Pi[t - nT_{\text{symbol}} - \tau_{\text{TR}}], \qquad (7)$$

where $x_n \in \{-1, 1\}$ are the N transmitted symbols and $\Pi(t)$ is the pulse (symbol) with duration T_{symbol}. In addition to the OOK modulation, the low-power consuming FM0 technique is utilized to encode the sensor data. For binary OOK, x_n would be the bits and for FM0-coded OOK, x_n are the binary symbols. In FM0 encoding there is an inversion of the phase at every bit boundary (at the beginning and at the end of every bit), and additionally bit "0" has an additional phase inversion in the middle (Fig. 6, left). Each bit includes two symbols, as shown in Fig. 6. The duration of a bit and of a symbol are denoted as T_{bit} and T_{symbol} respectively. The data bit rate is $1/T_{\text{bit}}$ bits per second (bps). The FM0 encoding always ends with a dummy "1" bit in order to detect easily the end of the bitstream. In the case that the received backscatter waveform finishes with a "LOW", it would be indistinguishable from receiving the reader's CW only (i.e. no packet transmission).

The tag is programmed to send the data in packets to the reader and the reader tries to receive and decode them. The length of each packet is fixed. Fig. 7 shows a typical packet format. The packet has the length of 26 bits and begins with the preamble bits. After that follow the "Tag ID" bits, the "Sensor ID" bits and finally the "Sensor Data" bits. The preamble is useful for bit-level synchronization at the receiver

and was fixed to be `1010101111` (10 bits) in our proof-of-concept tests. The "Tag ID" (2 bits) is utilized in the case of simultaneous multiple tag utilization. As mentioned before, the tag can support up to four sensors, and therefore the "Sensor ID" (2 bits) is used to identify the sensor the data is coming from.

IV. RECEIVER THEORY

In this work, ambient backscatter modulation based on On-Off Keying (OOK) modulation with FM0 encoding is used, as in conventional passive RFID tags [19]. An example of FM0 encoding is shown in Fig. 6, left. As a result, four possible waveforms are transmitted corresponding to a two-dimensional bi-orthogonal constellation. However, if one observes the FM0 signal shifted by one symbol, only two possible waveforms exist, which are the ones of bit "0". These two waveforms correspond to a one-dimensional antipodal constellation which is easier to study and decode [20]. The detected bits from the half-bit time shifted signal correspond to the originally transmitted bits after differential encoding. Therefore, one can proceed to decode the FM0 signal in two steps, first detecting the time-shifted bits and then using a differential decoder to recover the originally transmitted bits. In this section, we first derive the error probability P_s of the time shifted signal. Once P_s is obtained, the error probability of the originally transmitted bit stream P_e is given by [20], [21]:

$$P_e = 2P_s(1 - P_s). \qquad (8)$$

As it is shown in [20], in addition to the simplification of the detection process, the fact that an antipodal constellation is used leads to a SNR improvement of approximately 3 dB in comparison to the standard detection method based on the bi-orthogonal constellation.

In order to derive P_s, one may proceed following references [12] and [22]. In [12] a thorough analysis of traditional backscattering in a bi-static configuration is presented using a CW carrier signal. OOK modulation was assumed but without considering FM0 encoding. In [22], the analysis of the error probability of ambient backscatter systems was presented considering randomly modulated signals. In addition the special case of phase-shift-keying (PSK) modulation is treated in Appendix B, which is similar in analysis to FM signals used in this work, in that the carrier amplitude is constant. However, [22] also does not use FM0 encoding. In this work, we proceed by following the formulation of [22] but treat the case of FM0 encoding taking into account [20] as described in the previous paragraph. The received signal complex envelope was given in (5) and repeated here for convenience in a more compact form:

$$y(t) = Ae^{-jD}\big(\alpha_1(t)e^{-jK_1} + \alpha_2(t)b(t)e^{-jK_2}\big) + n(t). \quad (9)$$

The term D includes the frequency and phase offset, K_1 is the delayed modulation signal arriving directly from transmitter to the receiver and K_2 delayed modulation signal arriving through the tag. $b(t)$ is the information signal and $n(t)$ is additive zero mean complex white Gaussian noise added at the receiver $n(t) \sim \mathbb{N}(0, N_w)$. Following [22] we assume that

$K_1 = K_2$ due to the fact that the two paths are approximately equal. In addition, any thermal noise generated in the tag is ignored as very low [22] value. The obtained equation is:

$$y(t) \approx Ae^{-jD}e^{jK}h(t) + n(t). \qquad (10)$$

Where $h(t) = a_1(t) + a_2(t)b(t)$ is the complex valued signal containing the information from the tag and the channel effects. In order to eliminate the frequency and phase offset in the receiver we form the magnitude square of the envelope:

$$Z(t) = A^2|h(t)|^2 + |n(t)|^2 + 2\Re\{Ae^{-jD}e^{jK}h(t)n^*(t)\}$$
$$= A^2|h(t)|^2 + w(t). \qquad (11)$$

Following the Appendix B of [22], and invoking the Central Limit Theorem (CLT), $w(t)$ is a real Gaussian process with mean and variance given by:

$$w(t) \sim \mathcal{N}\big(N_w, N_w^2 + 2A^2N_w|h(t)|^2\big). \qquad (12)$$

The N_w is the noise power at the receiver. One should note that before the decoding process the receiver applies a low pass filter consisting of an averaging operation of approximately 1000 samples, which further supports the reasoning of invoking the CLT. The receiver applies a synchronization algorithm to derive the beginning of the information signal which is described in more detail in Section V. In order to facilitate the synchronization process a dc offset removal was applied to $Z(t)$. Due to the fact that the dc offset removal does not affect the detection process it will not be considered in this section. Once synchronization is achieved a time shifted version of the received bits $Z(t)$ is considered and detection based on an antipodal constellation is applied. Specifically, the received signal $Z(t)$ is correlated with pulse:

$$q(t) = \begin{cases} +1, & \text{if } 0 < t \le \frac{T_{\text{bit}}}{2} \\ -1, & \text{if } \frac{T_{\text{bit}}}{2} < t \le T_{\text{bit}} \end{cases} \qquad (13)$$

giving:

$$Z(t) = X + V$$
$$= \int_0^{T_{\text{bit}}} A^2|h(t)|^2 q(t)\mathrm{d}t + \int_0^{T_{\text{bit}}} w(t)q(t)\mathrm{d}t. \qquad (14)$$

Due to binary modulation $|h(t)|^2$ takes one of two values $|h_H|^2$ or $|h_L|^2$. It is straightforward to show that V is a real gaussian process with mean and variance given by

$$V \sim \mathcal{N}\big(0, 2T_{\text{symbol}}N_w^2 + T_{\text{symbol}}A^2N_w\big(|h_H|^2 + |h_L|^2\big)\big). \qquad (15)$$

Similarly:

$$X_\pm = \pm\frac{T_{\text{bit}}}{2}A^2\big(|h_H|^2 - |h_L|^2\big), \qquad (16)$$

with the sign depending on whether $q(t)$ or $-q(t)$ was transmitted. Assuming equal probability of transmission of the two possible symbols, one derives:

$$P_s = P\{Z < 0|+\} = Q\left(\frac{X_+}{\sigma_V}\right), \qquad (17)$$

where $Q(x)$ is the tail probability of the normal distribution function [20], [21]. $P(Z < 0|+)$ denotes the probability that

$Z < 0$ when $q(t)$ was transmitted. Therefore, the originally transmitted bit error probability is:

$$P_\text{e} = 2Q\left(\frac{X_+}{\sigma_\text{V}}\right)\left(1 - Q\left(\frac{X_+}{\sigma_\text{V}}\right)\right). \qquad (18)$$

It should be noted that in order to compute P_e one needs information of the signal at the two different states $|h_\text{H}|^2$ and $|h_\text{L}|^2$ but also of the noise power N_w, something which was also highlighted in [22], Appendix B.

A method to compute P_e is outlined in order to compare the theoretical analysis with bit-error rate measurements. Consider a given setup of transmitter, tag and receiver, and perform the following three power measurements, i.e. on signal $Z(t)$. While transmitting a modulated signal, set the tag to a fixed state and measure the received power, to obtain:

$$P_\text{yH} = A^2|h_\text{H}|^2 + N_\text{w} \qquad (19)$$

and

$$P_\text{yL} = A^2|h_\text{L}|^2 + N_\text{w}. \qquad (20)$$

Then turn the transmitter off and measure the noise power N_w. The most significant noise contribution is due to the receiver electronics and thus the state of the tag during the noise measurement is not important. Using the three measurements one has

$$X_\pm = \pm\frac{LT_\text{s}}{2}A^2\left(P_\text{yH} - P_\text{yL}\right) \qquad (21)$$

and

$$\sigma_v^2 = LT_\text{s}N_\text{w}\left(P_\text{yH} + P_\text{yL}\right). \qquad (22)$$

where T_s is the sampling period. In our implementation $L = 10$ samples per bit were used. Using the P_yH, P_yL and N_w measurements, one can apply (21) and (22) in (18) to compute the theoretical BER for a given transmitter power level.

V. RECEIVER IMPLEMENTATION

A. Software-Defined Radio (SDR)

In this work, the low cost (22 USD) RTL SDR (Nooelec NESDR SMArt) was used as receiver. It is an improved version of RTL SDR dongle that was used in [9], [23] and it is based on the same RTL2832U Demodulator/USB interface IC and R820T2 tuner. The new version provides a better oscillator, temperature stability and antenna improvements compared to the old one. It comes with an ultra-low phase noise 0.5 PPM temperature compensated crystal oscillator (Phase noise @100 kHz: -152 dBc/Hz). The dongle was redesigned with a RF-suitable voltage regulator with under 10 μVRMS of noise for lower power consumption. Power consumption has been reduced by an average of 10 mA according to manufacturer [24]. A custom heatsink is affixed to the primary PCB for temperature improvement and it comes with a low-loss RG58 feed cable and SMA antenna connector for better signal reception. In general the RTL SDR has a tuning frequency range from 24 MHz to 1850 MHz and it can support sampling rates up to 2.8 MS/s. The SDR downconverts the received RF signal to baseband and sends in-phase (I) and quadrature (Q) samples to the PC through the USB interface, while it is connected to an improved telescopic monopole antenna in order to receive the FM signals.
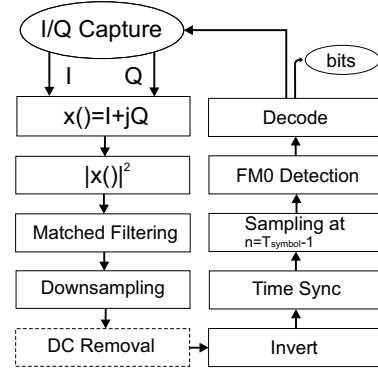

Fig. 8. Flow chart of the real-time receiver algorithm.

B. Receiver algorithm

The received signal of Eq. 5 contains the useful bits in FM0 encoding (rectangular pulses in Eq. 7). A real-time receiver and digital signal processing was implemented in order to read the backscattered information sent from the tag. The steps of the algorithm are briefly shown in Fig. 8 and the software that was used was Matlab and GNU radio framework. The GNU radio provides the I and Q samples to Matlab through a FIFO file and the samples are interleaved for further processing. The received digitized signal after sampling with a sampling period T_s, can be written as:

$$y_\text{r}[k] = y_\text{amb}(kT_\text{s} + \tau_{TR}) = x_\text{r}[k] + n[k] = I[k] + jQ[k], \quad (23)$$

with $n[k] = n(kT_\text{s})$ and $n[k] \sim \mathcal{N}(0, \sigma_n^2)$. The term $x_\text{r}[k]$ is the signal without noise that consists of a DC component, a modulated component and the ambient FM signal utilized for the backscattering. The algorithm collects and process the data in a window with duration: $3 \times$ packet duration.

The first step of the signal processing algorithm is the CFO correction. In our case, CFO is the frequency difference between the FM transmitter and SDR reader and, if not properly removed it causes a performance loss at the receiver. In order to eliminate this term without using an a-priori CFO estimation and correction algorithm, the absolute value $|y_\text{r}[t]|^2$ was taken, which is an established CFO compensation technique in digital communication textbooks, such as [25].

A matched filter was then applied to the samples in order to filter out noise and interference terms and maximize the SNR, consisting of a square pulse with duration T_symbol. Fig. 9 (Top) depicts the received packet of Fig. 7 after absolute square operation. The same packet after matched filtering is shown in Fig. 9 (Bottom). Matched filtering was followed by downsampling by a factor of 10 in order to reduce the computational cost of the subsequent operations without compromising the detection quality.

The DC offset of the received window was estimated by averaging some samples when the tag is not transmitting data (average at the start or at the end of the window). The DC offset was removed by subtracting the above estimate from all the values within the receive window. The outcome of this step can be an upright or an inverted waveform. In the case shown in Fig. 9 (bottom), an inverted waveform will result
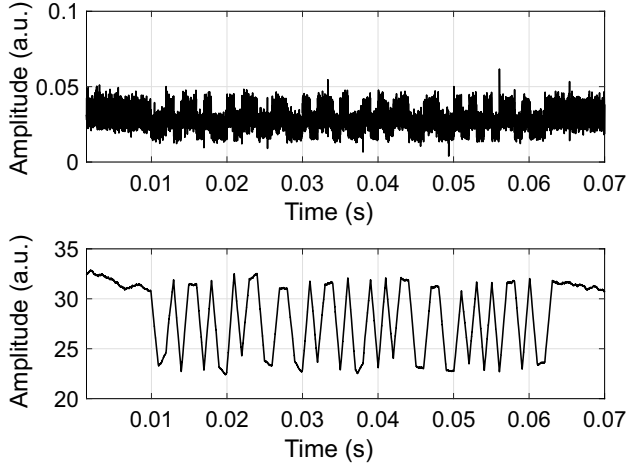
Fig. 9. Received signal including a data packet. Top: Squared absolute value signal. Bottom: Received signal after matched filtering for a symbol period, $T_{\text{symbol}} = 1$ ms. The packet is flipped due to the channel characteristics.

Fig. 10. Scotland FM radio outdoor deployment. The BBC 95.8 MHz station in "Radio 2" band was selected for measurements. The FM transmitter was 34.5 Km away from the measurement's setup and its transmission power was 250 kW.

after the DC offset removal. Upright or inverted waveforms may result due to the channel propagation characteristics. If an inverted waveform is detected after the DC offset removal, it is flipped so that only upright waveforms y_{fl} are forwarded to the synchronization block.

The received signal must be symbol-synchronized in order to determine when the packet starts. In order to find the starting sample of the packet, cross-correlation with the known preamble symbol sequence (1101001011010011011) was used. The similarity of the waveform y_{fl} and the preamble sequence p was evaluated as a function of the time-lag according to:

$$C[n] = \sum_{t=1}^{\infty} p[t] y_{\text{fl}}[t+n], \ n \in [0, N_s/2] \quad (24)$$

with N_s the number of received packet samples. The starting point of the packet is defined as:

$$I_{\text{start}}[n] = \arg \max_n C. \quad (25)$$

which corresponds to the position of the peak of the cross-correlation between the known sequence p and the received waveform.

In FM0-encoded signals, the received bits can be determined by comparing two neighbouring symbols. In order to begin decoding, $y_{\text{fl}}[t]$ is shifted to sample $I_{\text{start}} + P - T_{\text{symbol}}$, where P is the length of the preamble. Two possible orthogonal pulse waveforms can be received, as shown in [20] and used in [26]. The two waveforms are indicated in Fig. 6, (right) with a solid line square and a dash line square. With this observation the algorithm has to easily decode two adjacent received symbols in order to detect a whole bit. This method gives a gain of 3 dB compared to maximum likelihood symbol-by-symbol detection [27]. The two orthogonal waveforms can be expressed as:

$$D_1[k] = \begin{cases} +1, & \text{if } 0 < k \le \frac{M}{2} \\ -1, & \text{if } \frac{M}{2} < k \le M \end{cases} \quad (26)$$
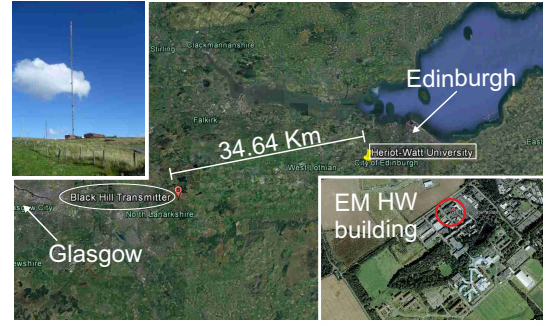
and $D_2[k] = -D_1[k]$ with M the oversampling factor T_{symbol}/T_s. The shifted signal is correlated with $D_1[k]$ and $D_2[k]$ and it is possible to determine which bit has been sent according to [28]:

$$S_{\text{k}} = \begin{cases} 1, & \text{if } \sum_{i=1}^{N_s} y_{\text{sh}}[i] D_1[i] > \sum_{i=1}^{N_s} y_{\text{sh}}[i] D_2[i] \\ 0, & \text{elsewhere} \end{cases} \quad (27)$$

with $y_{\text{sh}}[t]$ is the shifted version of waveform $y_{\text{fl}}[t]$. The results from the above calculation were stored in a vector L and the estimated bit $a_{\text{k}+1}$ that was sent is determined by:

$$a_{\text{k}+1} = \begin{cases} 0, & \text{if } L_{\text{k}} = L_{\text{k}+1} \\ 1, & \text{elsewhere.} \end{cases} \quad (28)$$

It is noticed that the first waveform derived by this decoding procedure is from the last preamble symbol (decoding starting point in Fig. 6, right). The following waveforms will be either D_1 or D_2. This means that if the first waveform is D_1 and the second is D_2 and vice versa, the bit "1" was sent, otherwise the bit "0" was transmitted.

VI. EXPERIMENTAL RESULTS

In this work we tried to produce a systematic set of measurements and compare them with the theoretical result of (18). For the systematic characterization in a controlled environment the system was demonstrated first in the anechoic chamber of the Heriot-Watt Microwaves and Antennas Laboratory. The tag was placed in a far-field anechoic chamber together with an analog signal generator used as a transmitter (TX). The SDR receiver (RX) was also placed at the edge of the anechoic chamber (Fig. 11). The tag, TX and RX are in fixed locations with fixed distances tag-TX 1.5 m and tag-RX 1.5 m. The TX and RX use commercial passive FM antennas with gain 2.5 dBi while the tag antenna is a wire dipole. The anechoic chamber was not specified to work at FM frequencies but it was used to minimize multipath and external interference.

The analog signal generator produces an FM modulated signal with a carrier centered at 98.5 MHz and frequency deviation of 75 KHz. The carrier frequency was selected so as to utilize a frequency band without any interference from any external stations. We used a sinusoidal signal with a frequency of 15 KHz to modulate the TX carrier. The 15 KHz is equal
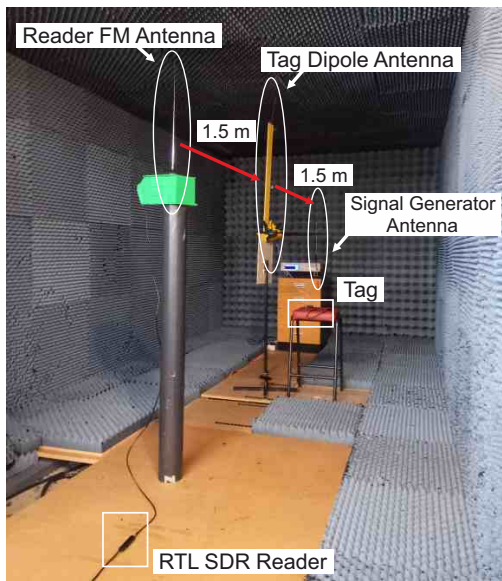
Fig. 11. Anechoic chamber experimental setup. The receiver antenna was placed at 1.5 m away from the tag and the tag was placed at 1.5 m away from the signal generator.
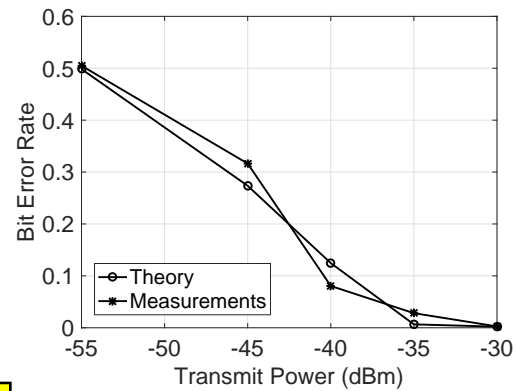

Fig. 12. Measured and theory calculated Bit Error Rate (BER) versus the signal generator transmit power for 0.5 Kbps.
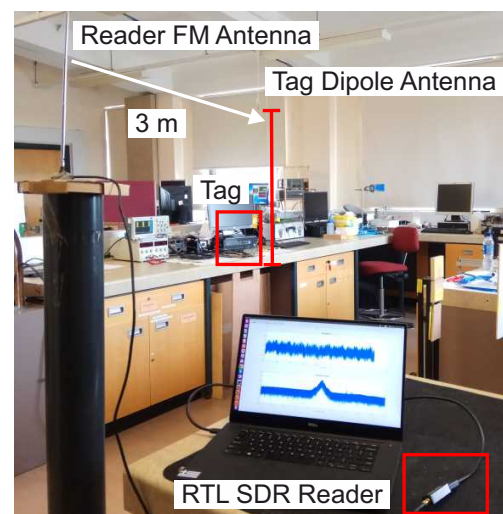

Fig. 13. Indoor experimental setup. The tag with the FM dipole antenna was set in a vertical position and the receiver was tuned at the most powerful FM station. For communication measurements, the receiver was placed at a maximum of 5 m away from the tag with the receiver antenna on top of a beam.

to the end of mono audio (left and right) signal frequency of a standard stereo FM signal transmission (Fig. 3 and that FM stations typically use a 75 KHz deviation [13].

The tag was programmed to send packets with fixed information bits for bit rate: 500 bps. An oscilloscope measurement of the packet transmitted at 500 bps is presented in Fig. 7. The data information was the 12-bit binary representation of 965 mV: 001111000101.

The receiver has a bandwidth of 1 MHz around the carrier frequency. The noise power P_w at the receiver was computed over the 1 MHz bandwidth while TX was off. Then, for a given transmit power at the TX, the received power at RX was recorded while the tag was set to a fixed load state A or B, resulting in P_{yH} or P_{yL}. The measured data consist of downconverted time domain values, which were converted in the frequency domain by taking a fast fourier transform (FFT) and the total power was computed by taking the sum of the squared magnitude values of the FFT operation. It is noted that P_{yH} and P_{yL} correspond to the total signal plus noise power measurement. Two sets of P_{yH} and P_{yL} measurements were collected for a varying transmit power from -55 dBm to -25 dBm. In order to compute an estimate of average power values, for each transmit power, 200 sets of data were collected and an average power value was computed.

The BER was measured for each value of the transmitted power while the tag rate was backscattering a fixed package with bit rate of 500 bps. In addition, the BER was recorded for each transmit power level. The resulting BER vs TX transmit power curves are shown in Fig. 12 along with the theoretical BER results (P_e). To calculate the analytical BER, the measured values of P_{yH} and P_{yL} and P_w were used with (18). One can see a good agreement between simulation and measurement. BER measurements were performed for transmitted power levels up to -30 dBm where the BER

approached 10^{-3}. Due our system memory limitations it was not possible to setup longer measurements containing a sufficient number of data to ensure a good confidence level of BER measurement. For example for a typical confidence level of 0.95 the required number of bits to test without any errors is 2.99573×10^7 in comparison to our case that we had only 9616 transmitted bits.

The proposed system was also tested indoors in the Heriot-Watt Microwaves and Antennas Laboratory, selecting the most powerful FM station as the ambient RF source to use in backscattering. Thus, the receiver was tuned to BBC 95.8 MHz station with 1 MS/s sampling rate. The station is located 34.5 Km away at the "Black Hill" location between the town of Edinburgh and Glasgow as depicted in Fig. 10. The transmission power of the station is 250 kW. The power of the FM station carrier signal was measured in the vicinity of the tag antenna in the lab at -51 dBm. The reader was placed close to the tag at different reader-to-tag distances with a maximum range of 5 m (Fig. 13). The antenna of the reader
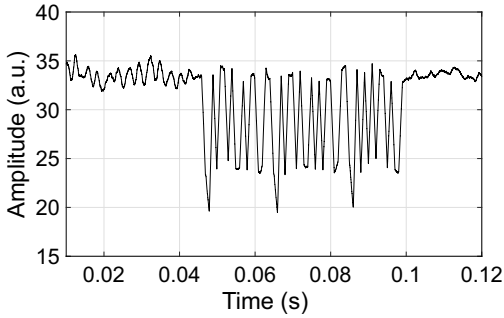
Fig. 14. Corrected received packet after matched filtering at $T_{\text{symbol}} = 1$ ms (500 bps) featuring a smaller channel fluctuation. igh frequency noise components can be observed.
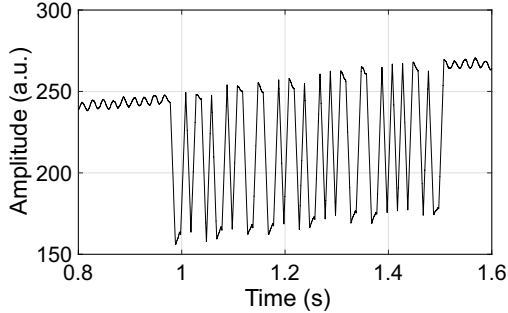

Fig. 15. Corrected received packet after matched filtering for $T_{\text{symbol}} = 10$ ms (50 bps) including the channel fluctuation effects. A better filtering quality is observed.
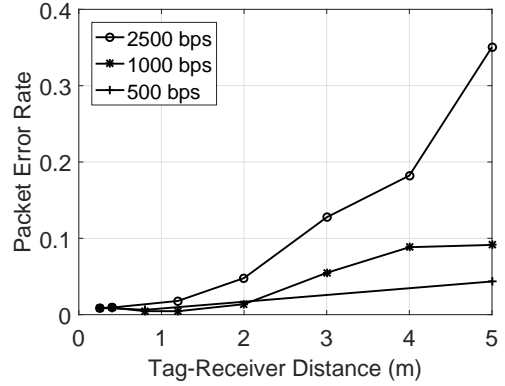

Fig. 16. Measured Packet Error Rate (PER) versus the tag-receiver distance for 0.5, 1 and 2.5 KBps.

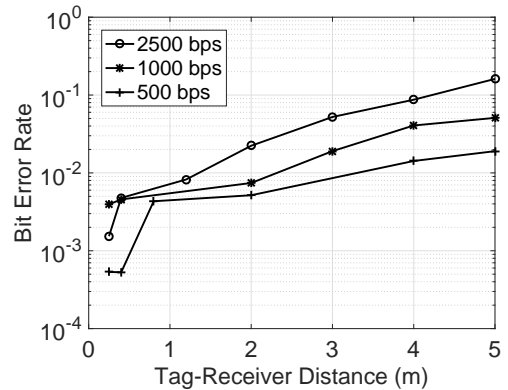Fig. 17. Measured Bit Error Rate (BER) versus the tag-receiver distance for 0.5, 1 and 2.5 Kbps.

was placed on top of a plastic stick with height 1.5 m for better reception.

The tag was programmed to send packets with the fixed information bits (same as above) for the following different bit rates: 50, 100, 500, 1000, 1250 and 2000 bps. The received packets for 500 bps and 50 bps after the matched filtering step are illustrated in Fig. 14 and Fig. 15 respectively. One can see that the packets are inverted due to the channel conditions i.e. random, unknown channel phase. It is clear that there is trade-off between bit rate and efficient filtering. In case that a high bit rate is employed (Fig. 14), there is less channel fluctuation, and the matched filtering operation is not able to remove the high-frequency components of the ambient FM signal, due to the wider bandwidth of the matched filter. In the case of low bit rate transmission (Fig. 15) the filtering operation is more effective, corresponding to a higher SNR, but a channel fluctuation effect is visible. When channel fluctuation is present it is more difficult to decode the packet due to the fast varying signal level.

In order to validate the effectiveness of our digital backscatter communication system, numerous range measurements were performed indoors with the setup described above. Figures 16 and 17 display the BER and PER performance as a function of the tag-to-reader distance for the three different data rates. The minimum PER and BER value at 5 m was measured to be 0.043 and 0.0019 respectively. As the tag-to-receiver distance decreases, the reader can decode successfully more the bit packets. It is also seen that for a given distance

value, reducing the bit rate improves the PER and BER performance. However, transmitting packets at lower bit rates result in increased transmission time and energy per packet while the MCU and the front-end staying in "on" state for longer time. There is a direct and inversely proportional relationship between the bit rate and the energy that a tag consumes sending a packet as shown in Table I, where the energy per packet for six bit rates is presented. The table also provides the tag power consumption for each bit rate. A higher power consumption of the MCU electronics is observed when operating at a higher bit rate. In order to compile the measurements shown in Table I, the tag was programmed to wake up every 3 sec, transmit a packet and go to sleep mode, while being powered from the supercapacitor.

The power consumption of the tag can be reduced by the following modifications which are the object of future work. First, it is possible to use a more energy efficient MCU such as PIC16LF1459 (25 μA/MHz at 1.8 V) [29] . Similarly, one can select sensing elements with minimum power dissipation or even employ some passive sensing technique such as for example [30], [31]. Second, the RF front-end can be modified to use instead of an off-the-shelf switch, a single transistor based switch such as the ones in [32], [33] with pJ/bit energy consumption. Finally, a customized CMOS based IC may provide an even further reduction of dissipated power, as sug-

TABLE I
Tag Power Characteristics.

Bit Rate (bps)	Power (mW)	Energy/Packet (μJ)
2000	2.838	36.9
1250	2.087	43.4
1000	1.785	46.4
500	1.283	66.73
100	0.751	195.45
50	0.677	352.15

gested in [6]. In addition to reducing the circuit consumption, battery-less operation can be achieved by exploring energy harvesting techniques. There are several studies related to the availability of ambient RF energy [34]–[37] as well as demonstrations of sensors powered by harvesting ambient RF energy from TV [18], WiFi [38] or even microwave oven signals [39], which could be used for smart house-targeted sensors. In addition, multiple technology of energy harvesters such as solar and electromagnetic energy harvesters can be employed in order to combine the different forms of ambient energy availability [17], [40].

Finally, the potential interfering effects of ambient backscatter systems on the performance of the ambient systems it utilizes should be considered. In U.S.A., according to the Federal Communications Commission (FCC), it is illegal to broadcast unlicensed signals on FM band (88 MHz to 108 MHz) [41]. However, devices that communicate with backscatter signals (e.g. RFID tags) have not been reviewed or tested by FCC. The reason is that the RF front ends of backscatter tags are not active components (have no amplifiers) and they only modulate the reflections of the incoming signals. Consequently the power of the reflected signals is at very low levels. The ambient backscatter operation such as our developed system belongs to the category of RFID tags so it is legal under current rules. However, the reflected signals of existing FM signals could be synchronized with FM transmissions to interfere the commercial FM receivers. Experimentally it was observed that the transmissions did not affect typical FM receivers, due to the low power level of ambient backscatter signals, and the different type of modulation (ASK). A FM-receiver equipped smartphone was tested in the worst-case interference scenario, where random data was continuously backscattered. The received audio quality was practically unchanged when the backscatter tag was placed right at the vicinity of the smartphone.

VII. CONCLUSION

In this work, we present a novel FM backscatter tag and receiver system. The tag communicates with a low cost SDR reader by backscattering the ambient FM signals. Data acquisition from sensors with low power operation and communication ranges up to 5 m has been demonstrated experimentally. The communication was implemented with OOK modulation over the modulated carrier of the most powerful FM station. This concept can be the next novel way for low power and low cost long range communication.

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