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SIGNAL DESCRIPTIONS AND FORMULATIONS FOR LONG RANGE UHF RFID READERS

Z. G. Fan, S. Qiao, J. T. Huangfu, and L. X. Ran †

Department of Information and Electronic Engineering
Zhejiang University
Hangzhou 310027, China

Abstract—In this paper, signal descriptions and formulations for the radio frequency (RF) front-end of a passive backscatter radio frequency identification (RFID) reader working at ultra high frequencies (UHF) are discussed in detail, and a set of design considerations aiming to improve the read range are outlined. The reader's architecture is proposed and the design details of its RF frond-end are presented. The read range is formulated through calculating the time-averaging power absorbed by the tag and the signal-noise-ratio (SNR) of the demodulation, and accordingly RFID systems can be classified into tag-determining and reader-determining ones. It is concluded that the gain of the reader antenna, the phase noise of the local oscillation (LO) and the receive-transmit isolation coefficient dominate the demodulation output noise of the reader, and consequently the reader-determining maximum operational distance. A prototype reader working at the frequency of 915 MHz was built with off-the-shelf components and was evaluated with a commercial tag in an indoor environment. The measured results show that this RFID system is of tag-determining and has a read range of 8.4 meters, which are in good agreement with the calculated results.

1. INTRODUCTION

Radio frequency identification (RFID) that just began its explosive development in the last decade is actually with a long history of more than half a century [1]. Passive backscatter ultra high frequency (UHF) RFID readers are interrogators for tags, or transponders,

† The second author is also with Zhejiang University City College, Hangzhou 310027, China. The 3rd and 4th authors are also with The Electromagnetics Academy at Zhejiang University, Zhejiang University, Hangzhou 310027, China.

based on the operational principles of wireless power transmission [2] and backscatter communication [3–5]. The reader transmits electromagnetic waves into air and the tag draws operational energy from the radiation fields. The electromagnetic waves are partially backscattered and modulated by the tag, from which the reader retrieves the tag's information.

An important characterization of the performance of an RFID system is "read range", which is defined as the maximum distance between a reader and a tag where the radiation field from the reader is strong enough to power up the tag and the backscatter signal from the tag is strong enough to be detected correctly by the reader. In quite a few RFID applications such as vehicle toll collection, retail item management, animal tracking and building access control [6–9], a reader that can provide long read range is usually preferred or required. In addition to long range applications, other versatile designs are also required, for example, the power consumption is also emphasized in portable cases. However, the design of an RFID system with optimized performance is not easy, and due to the complex RF environments and the very weak backscatter signals, the design for the RF front-ends involved in both tags and readers are especially difficult. Hence, a complete understanding of the architectures and the signal descriptions is important for the optimal performances. For tags, considerable researches have been conducted in various aspects [10–14]. For readers, a compact antenna of dual polarizations is investigated in [15]. However, to our knowledge, no paper on the reader has been published that provides detailed, systematic signal descriptions and formulations or design considerations, particularly in the aspect of the radio frequency (RF) frond end design of the reader, although in some tag papers, valuable discussions on the reader design are also incidentally presented, such as in [13].

In the current paper, signal descriptions and formulations are presented in details for a basic passive backscatter UHF RFID system consisting of a tag, a reader and some obstacles in operational circumstance. It is concluded that a non-coherent receiver needs to be adopted for the reader design. We propose the reader's architecture and present the design details of its RF frond-end section. The read range of the reader is formulated through calculating the time-averaging absorption power of the tag and the signal-noise-ratio (SNR) of the demodulation output signal of the reader and accordingly RFID systems are classified into two kinds: one is of tag-determining and the other is of reader-determining. It is observed that the gain of the reader antenna, the phase noise of the reader's local oscillation (LO) and the receive-transmit isolation coefficient mainly determine the demodulation output noise of the reader and consequently the reader-determining maximum operational distance. Then, a set of design considerations for the RF Front-end of the reader are summarized to help to achieve long read range.

We also demonstrate a design example of RFID system operating at the frequency of 915 MHz. The reader is built with off-the-shelf components and a commercial tag is chosen to compose the whole system. The calculated tag-determining maximum operational distance takes a value of 8.3 m and the calculated reader-determining maximum operational distance takes a value of 11.8 m, which shows that this is a tag-determining RFID system. Through conducting two comparative experiments, we confirm the above conclusion and also get an estimated value of 8.2 m for the tag-determining maximum operational distance. In an open indoor environment, we measured the SNR of the reader's demodulation output signal for several different operational distances and got a maximum operational distance of 8.4 m, where the SNR of the demodulation output signal well meets the user-specified SNR requirement. Thus, this RFID system indeed is of tag-determining and accordingly the read range takes a value of 8.4 meter, which shows good agreement with the calculated result.

2. SIGNAL DESCRIPTION

Fig. 1 illustrates a basic passive backscatter UHF RFID system which consists of a reader, a tag and some obstacles in operational circumstance. The tag is composed of a microchip and an antenna.

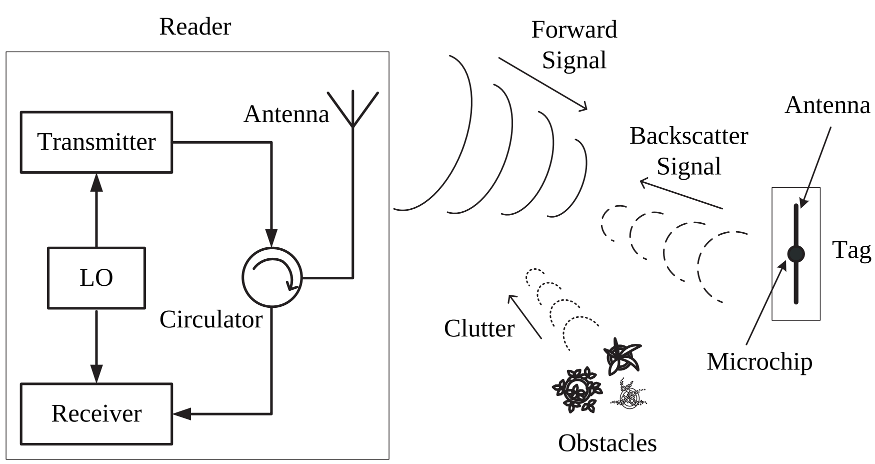

Figure 1. Basic passive backscatter UHF RFID system.

The microchip stores the data information for the object attached by the tag. There is no internal energy source for the microchip operation in the tag. The tag draws energy from the forward electromagnetic wave signal transmitted by the reader. Part of the electromagnetic wave signal is backscattered/reflected back to the reader by the tag. The backscatter signal is also modulated by the tag through varying the microchip's input impedance in terms of the stored data information to realize backscatter communication. Generally, the tag toggles the microchip's input impedance between two discrete states, which produces amplitude shift keying (ASK) or phase shift keying (PSK) impedance modulation [9]. The reader is composed of a local oscillator (LO), a transmitter, a receiver and a receiver-transmit duplex antenna plus a circulator as duplexer. The circulator is a non-reciprocal three-port device, where the signals travel from the transmitter port to the antenna port or from the antenna port to the receiver port. The LO provides two sine-waveform signals of the same frequency, of which one is for the transmitter and the other is for the receiver. The transmitter amplifies the LO signal to achieve high power level. The amplified signal feeds into the reader antenna via the circulator and then radiates into air. Simultaneously, the reader antenna also picks up the backscatter signals from the tag and the clutter from the obstacles. These received signal then enter the receiver via the circulator. The receiver utilizes the LO signal to demodulate the backscatter signal to retrieve the tag's data information.

It is also practicable to adopt two antennas for the reader design, i.e., one receiving antenna and one transmitting antenna. Then, the isolation between the receiver and the transmitter is achieved due to the antennas' directivity and the circulator is not needed any more. However, there is no difference in nature in terms of operational principles of these two reader designs. Thus, in order to save space, the current paper will only focus on the investigations into the former, whose analyze methods are also applicable to the later completely.

As mentioned above, the transmitter and receiver operate at one time and one frequency and the circulator acts as the receive-transmit duplexer. However, the circulator can not isolate the transmitter from the receiver totally due to the inherent leakage between its ports. In practice, due to inevitable impedance mismatching between the circulator and the reader antenna, the reflection also gives rise to leakage signal into the receiver, which deteriorates the circulator's isolation performance further. Here, we define a power transmission coefficient α and a power leakage coefficient β to characterize the insertion loss and the isolation between any two ports of the circulator, respectively. Let P_T denote the signal power feeding into the reader antenna by the transmitter and then the power of the signal leaking into the receiver via the circulator is given by βP_T. Let P_B denote the power of the backscatter signal received by the reader antenna and then the power of the signal entering the receiver via the circulator is given by αP_B. Then, we define a quantity ξ that is equal to α/β and call ξ the receive-transmit isolation coefficient of the reader.

It is assumed here that the reader antenna is of polarization matching with the tag antenna. Let r denote the operational distance between the tag and the reader operating in free space. Then, using the Friis electromagnetic wave propagation equation, the power P_R of the signal received by the tag antenna is given by

$$P_R = \left(\frac{c}{2\omega r}\right)^2 P_T G_T G_R, \tag{1}$$

where c is the speed of light in free space, ω is the signal angular frequency, G_R is the gain of the reader antenna and G_T is the gain of the tag antenna. One portion of P_R is absorbed by the tag to provide operational energy for the microchip circuits and the other portion of P_R is reflected back to the tag antenna by the microchip and then reradiated into air by the tag antenna. The tag varies the amount of P_R through modulating the microchip's input impedance in terms of its stored data to realize backscatter communication. As mentioned above, there are two kinds of impedance modulation methods, i.e., ASK and PSK. In the case of ASK tag, the power reflection coefficient typically takes a value of "0"/"Γ" for a data bit of "0"/"1", respectively; in the case of PSK tag, it takes the same value of Γ for both data bit of "0" and "1". Here, we assume that data bits of "0" and "1" have the same transmission probability. Let P_A denote the time-averaging absorption power of the tag. Then, in the case of ASK tag, we can get

$$P_A = (1 - \Gamma/2)P_R; \tag{2a}$$

in the case of PSK tag, we can get

$$P_A = (1 - \Gamma)P_R. \tag{2b}$$

As can be seen in Fig. 1, the electromagnetic wave signals picked up by the reader antenna consist of two components: one is the backscatter signal from the tag and the other is the clutter that is the sum of the backscatter signals from the obstacles in operational circumstance. The backscatter signal from the tag is indeed composed of the modulated backscatter signal due to the tag microchip's impedance modulation as mentioned above and the unmodulated structural backscatter signal due to the induced current on the surface of the tag antenna [16]. Both the backscatter signal and the clutter enter the receiver via the circulator. In addition, the leakage signal between the transmitter and receiver ports of the circulator also enters the receiver. These input signals can be summed up into two components: one is the unmodulated signal $X_U(t)$ that is the sum of the clutter from the obstacles, the structural backscatter signal from the tag antenna and the receiver-transmit leakage signal, which all act as harmful interference to the receiver; the other one is the modulated backscatter signal $X_M(t)$ that conveys the tag's data information. For a long range reader, $X_U(t)$ is usually much stronger than $X_M(t)$, which makes it hardly possible for the receiver to achieve a LO signal $X_S(t)$ being synchronized with $X_M(t)$. In other words, $X_S(t)$ recovered from $X_M(t)$ indeed would mainly tracks the phase of $X_U(t)$ not that of $X_M(t)$. Thus, non-coherent demodulation scheme needs to be adopted for the reader's receiver design.

In practice, the receive-transmit leakage signal generally dominates among the unmodulated signal $X_U(t)$. Thus, we can neglect other components and proximately express $X_U(t)$ as

$$X_U(t) = A_U \sin(\omega t + \theta_U), \tag{3a}$$

where θ_U denotes the signal phase and A_U denotes the signal amplitude that is given by

$$A_U = \sqrt{2\beta R_0 P_T}, \tag{3b}$$

where R_0 is the input resistance of the receiver. The modulated backscatter signal $X_M(t)$ can be expressed as

$$X_M(t) = A_M[S(t)] \sin(\omega t + \theta_{M0} + \theta_M[S(t)]), \tag{4}$$

where $S(t)$ denotes the tag's binary data sequence of "0"/"1", θ_{M0} denotes the unmodulated part of the signal phase, $\theta_M[S(t)]$ denotes the part of the signal phase modulated by $S(t)$ and $A_M[S(t)]$ denotes the signal amplitude modulated by $S(t)$. In the case of ASK modulation, $A_M[S(t)]$ takes a value of "0"/"A_M" while the tag transmits a data bit of "0"/"1", respectively, and $\theta_M[S(t)]$ takes a constant value of θ_M; in the case of PSK modulation, $\theta_M[S(t)]$ takes a value of $-\theta_M/\theta_M$ while the tag transmits a data bit of "0"/"1", respectively, and $A_M[S(t)]$ takes a constant value of "A_M". Using the Friis electromagnetic wave propagation equation again and (1), A_M can be given by

$$A_M = \left(\frac{c}{2\omega r}\right)^2 \sqrt{2\Gamma\alpha R_0 P_T} G_T G_R. \tag{5}$$

The LO signal $X_S(t)$ for the receiver can be expressed as

$$X_S(t) = A_S \sin(\omega t + \theta_S), \tag{6}$$

where θ_S denotes the signal phase and A_S denotes the signal amplitude. In practice, the LO providing $X_U(t)$, $X_M(t)$ and $X_S(t)$ always has phase and magnitude noise. Thus, these signals' phases, i.e., θ_U, θ_{M0} and θ_S, all should be of random variable depending on time, i.e., stochastic process, to reflect the phase noise from the LO. However, considering that the amplitude noise is usually much lower than the phase noise, in order to simplify analysis in the later section, here we neglect the former's effect and still utilize constant quantities to denote these signals' amplitude.

3. READER DESIGN

3.1. Reader Architecture

Based on the analysis in Section 2, we design a passive UHF RFID reader, whose block diagram is illustrated in Fig. 2. We divide it into two sections, i.e., the RF frond-end section and the digital baseband section. The RF Front-end is composed of a receiver, a transmitter, a LO, a circulator and an antenna. The receiver consists of a RF band-pass filter, a power splitter, a $\pi/2$-phase-shifting power splitter, two mixers and two baseband band-pass filters. The RF band-pass filter is for rejecting the interference outside the operational frequency band. The power splitter, the power splitter and two mixers compose a dual-channel (I channel and Q channel) non-coherent quadrature

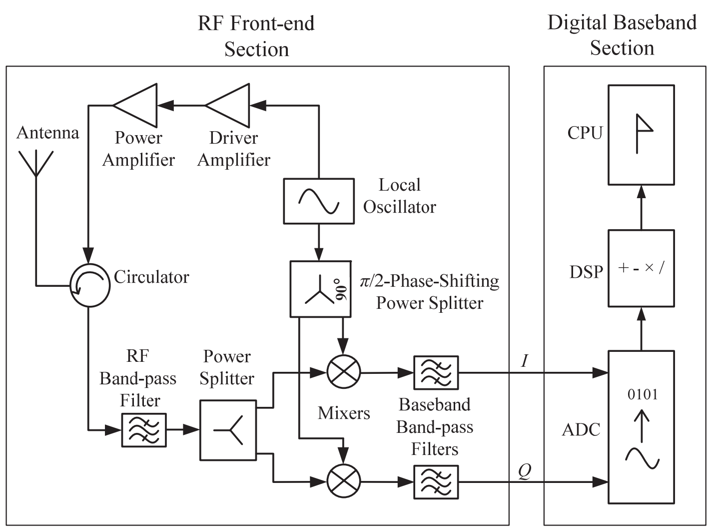

Figure 2. Reader architecture.

demodulator, from which that one of better SNR is chosen to retrieve the tag's data information. The purpose of baseband band-pass filters is to remove unuseful or even harmful DC, low frequency and high frequency noise signals. The transmitter consists of a driver and a power amplifier to achieve high power level signal. The circulator acts as the receive-transmit duplexer. The antenna radiates the power signal from the transmitter into air and simultaneously picks up the backscatter signals from the tag and the obstacles in operational circumstance.

The digital baseband section is generally composed of analog-digital-converter (ADC), digital signal processor (DSP), central processing unit (CPU), etc. Its functions mainly include signal sampling, channel choosing, data judging, data decoding, etc. This aspect is beyond the scope of the current paper and will not be discussed further here.

3.2. Read Rang Formula

The read range of the reader is determined by two conditions: a) the tag could draw enough energy for its microchip's normal operation from the incident electromagnetic wave; b) the modulated backscatter signal from the tag should be strong enough to make the reader's demodulation output signal meet a user-specified SNR requirement.

Firstly, we need to calculate the time-averaging absorption power P_A of the tag. It is obvious that increasing the transmitting power P_T of the reader would directly improve P_A. However, in practice, P_T is generally constrained by regional regulations to be no larger than a specified power value P_{EIRP} (efficient isotropic radiated power). For example, the P_{EIRP} is equal to 4 watt within 915 MHz ISM band (902–928 MHz) in the United States. Here, let P_T take the maximum allowed value, i.e., $P_T = P_{EIRP}/G_R$. Then, in the case of ASK tag, using (1) and (2a), we get

$$P_A = (1 - \Gamma/2) P_{EIRP} G_T \left(\frac{c}{2\omega r}\right)^2; \tag{7a}$$

in the case of PSK tag, using (1) and (2b), we get

$$P_A = (1 - \Gamma) P_{EIRP} G_T \left(\frac{c}{2\omega r}\right)^2. \tag{7b}$$

In order to power up or activate the tag, P_A must be no less than the tag microchip's input threshold power level P_{TH}, i.e., $P_A \geq P_{TH}$, which gives a upper limit of the operational distance and is called the tag-side condition of determining the read range. Here, we define a quantity r_{TAG} to denote the tag-determining maximum operational distance. Then, for the ASK tag, using (7a), we get

$$r_{TAG} = \frac{c}{2\omega}\sqrt{\frac{(1-\Gamma/2)P_{EIRP}G_T}{P_{TH}}}; \tag{8a}$$

for the PSK tag, using (7b), we get

$$r_{TAG} = \frac{c}{2\omega}\sqrt{\frac{(1-\Gamma)P_{EIRP}G_T}{P_{TH}}}. \tag{8b}$$

Secondly, we need to calculate the SNR of the demodulation output signal of the reader. Considering that the baseband band-pass filter utilized in the reader's receiver have sharp frequency selectivity, in order to simplify the later analysis, we characterize it approximately with an ideal rectangular transfer function and denote its low-end frequency and high-end cutoff frequency as f_L and f_H, respectively. Then, referring to Fig. 2 and using (3a), (4) and (6), we can express the demodulation output signals of two channels of the receiver as

$$\begin{aligned} X_I(t) &= k_D A_S A_U \cos(\theta_U - \theta_S) + k_D A_S A_M[S(t)] \\ &\quad \cos[\theta_{M0} - \theta_S + \theta_M[S(t)]] + n_0(t) && \text{(9a)} \\ X_Q(t) &= k_D A_S A_U \sin(\theta_U - \theta_S) + k_D A_S A_M[S(t)] \\ &\quad \sin[\theta_{M0} - \theta_S + \theta_M[S(t)]] + n_0(t), && \text{(9b)} \end{aligned}$$

respectively, where θ_U, θ_{M0} and θ_S are all of stochastic process to reflect the phase noise of the LO, $n_0(t)$ denotes the internal thermal noise of the receiver and k_D denotes the transfer coefficient of the receiver that takes into consideration the total loss or gain of the RF band-pass filters, the power splitter, the mixers and the baseband band-pass filters. In practice, $X_U(t)$ is usually orders of magnitude larger than both $X_M(t)$ and $n_0(t)$, which means that the SNR of the demodulation output signal is mainly determined by the phase noise of $X_U(t)$ and consequently we can neglect the noise contributions of $n_0(t)$ and $(\theta_{M0} - \theta_S)$ terms in (9a) and (9b). In addition, since $X_U(t)$ and $X_S(t)$ originate from the same LO, we can assume that their phase noise can be characterized using the same stochastic process except for a time delay. Then, we can rewrite (9a), (9b) as

$$\begin{aligned} X_I(t) &\approx k_D A_S A_U \cos[\omega\Delta t + \theta_P(t+\Delta t) - \theta_P(t)] \\ &\quad + k_D A_S A_M[S(t)] \cos[\theta_{M0S} + \theta_M[S(t)]] && \text{(10a)} \\ X_Q(t) &\approx k_D A_S A_U \sin[\omega\Delta t + \theta_P(t+\Delta t) - \theta_P(t)] \\ &\quad + k_D A_S A_M[S(t)] \sin[\theta_{M0S} + \theta_M[S(t)]], && \text{(10b)} \end{aligned}$$

respectively, where $\theta_P(t)$ is a stochastic process characterizing the phase noise of the LO, Δt denotes the time delay between $X_U(t)$ and $X_S(t)$, and θ_{M0S} denotes the phase difference between $X_M(t)$ and $X_S(t)$. Then, the SNRs of $X_I(t)$ and $X_Q(t)$ can be given in frequency domain by

$$SNR_I = \frac{\int_{f_L}^{f_H} \boldsymbol{S}\{A_M[S(t)]\cos[\theta_{M0S} + \theta_M[S(t)]]\}df}{\int_{f_L}^{f_H} \boldsymbol{S}\{A_U\cos[\omega\Delta t + \theta_P(t+\Delta t) - \theta_P(t)]\}df} \quad (11a)$$

$$SNR_Q = \frac{\int_{f_L}^{f_H} \boldsymbol{S}\{A_M[S(t)]\sin[\theta_{M0S} + \theta_M[S(t)]]\}df}{\int_{f_L}^{f_H} \boldsymbol{S}\{A_U\sin[\omega\Delta t + \theta_P(t+\Delta t) - \theta_P(t)]\}df}, \quad (11b)$$

respectively, where $\boldsymbol{S}\{\ \}$ denotes the operator of calculating the power density spectrum (PDS) of a stochastic process, i.e., the Fourier transform of the auto-correlation function of a stochastic process. It is obvious that the SNRs of both $X_I(t)$ and $X_Q(t)$ strongly depend on the unknown θ_{M0S} and Δt. In other words, there is an inherent uncertainty in terms of the SNR performance of a nor-correlate demodulation receiver. Thus, two channels could have different SNRs and the digital baseband section would choose that one of higher SNR to retrieve the tag's data information. In the worst case where θ_{M0S} takes a value of $\pi/4$, $\omega\Delta t$ takes values of $(2n+1)\pi/2$ in (11a) or $n\pi$ in (11b), where n is of integer number, and the stochastic processes $\theta_P(t)$ and $\theta_P(t+\Delta t)$ are assumed to be non-correlated to each other, considering that $\theta_P(t)$ and $\theta_P(t+\Delta t)$ are mostly taking values near to zero for a practical LO having low phase noise, the minimal achievable SNR can be approximately expressed as

$$SNR_{MIN} \approx \frac{\int_{f_L}^{f_H} \boldsymbol{S}\{A_M[S(t)]\cos[\pi/4 + \theta_M[S(t)]]\}df}{2\int_{f_L}^{f_H} \boldsymbol{S}\{A_U\theta_P(t)\}df}, \quad (12)$$

which gives a lower boundary of the SNR value of the demodulation outputs. Here, we neglect the correlation effect between $\theta_P(t)$ and $\theta_P(t+\Delta t)$, which indeed would dramatically reduce the demodulation output noise [17]. In order to characterize the correlation effect, we define a phase noise improvement factor ψ, which can be achieved experimentally as shown in the later section. Then, we rewrite (12) as

$$SNR_{MIN} = \frac{\displaystyle\int_{f_L}^{f_H} \boldsymbol{S}\{A_M[S(t)]\cos[\pi/4+\theta_M[S(t)]]\}df}{\displaystyle\psi\int_{f_L}^{f_H} \boldsymbol{S}\{A_U\theta_P(t)\}df} \tag{13}$$

Now, substituting (3b) and (5) into (13), in the case of ASK tag, we get

$$SNR_{MIN} = \frac{\Gamma\xi G_T^2 G_R^2}{2\psi}\left(\frac{c}{2\omega r}\right)^4 \frac{\displaystyle\int_{f_L}^{f_H} \boldsymbol{S}\{S(t)\}df}{\displaystyle\int_{f_L}^{f_H} \boldsymbol{S}\{\theta_P(t)\}df}; \tag{14a}$$

in the case of PSK tag, we get

$$SNR_{MIN} = \frac{\Gamma\xi G_T^2 G_R^2}{\psi}\left(\frac{c}{2\omega r}\right)^4 \frac{\displaystyle\int_{f_L}^{f_H} \boldsymbol{S}\{\cos[\pi/4+\theta_M[S(t)]]\}df}{\displaystyle\int_{f_L}^{f_H} \boldsymbol{S}\{\theta_P(t)\}df}. \tag{14b}$$

In (14a) and (14b), the integral items in the denominators denote the single-sideband in-band phase noise power of the LO (a dimensionless quantity being normalized to the carrier power), expressed as P_{PN}, which can be readily obtained by means of numerical integral calculation on the known LO phase noise data or direct measurement with a spectrum analyzer (SPA) device; the integral items in the numerators denote the single-sideband in-band signal power of the tag's binary data sequence (also being treated as a dimensionless quantity in fact), expressed as P_{DATA}, which can be derived according to specific data coding schemes such as unipolar/bipolar coding, return-to-zero/non-return-to-zero (RZ/NRZ) coding, Manchester coding, Miller coding and FM0 (bi-phase-space) coding [9, 18]. Then, the read range can be calculated through requiring the SNR_{MIN} to be no lower than a user-specified SNR_{USER}, i.e., $SNR_{MIN} \geq SNR_{USER}$, which is called the reader-side condition of determining the read range. Here, we define a quantity r_{READER} to denote this reader-determining maximum operational distance. Then, for the ASK tag, using (14a), we get

$$r_{READER} = \frac{c}{2\omega}\left(\frac{\Gamma\xi G_T^2 G_R^2}{2\psi SNR_{USER}} \frac{\displaystyle\int_{f_L}^{f_H} \boldsymbol{S}\{S(t)\}df}{\displaystyle\int_{f_L}^{f_H} \boldsymbol{S}\{\theta_P(t)\}df}\right)^{1/4}; \tag{15a}$$

for the PSK tag, using (14b), we get

$$r_{READER} = \frac{c}{2\omega} \left(\frac{\Gamma \xi G_T^2 G_R^2}{\psi SNR_{USER}} \frac{\int_{f_L}^{f_H} \boldsymbol{S}\{\cos[\pi/4 + \theta_M[S(t)]]\} df}{\int_{f_L}^{f_H} \boldsymbol{S}\{\theta_P(t)\} df} \right)^{1/4}; \tag{15b}$$

Based on (14a), (14b), (15a) and (15b), for a fixed P_{EIRP} at an operation frequency of ω and a tag of the specific P_{TH}, G_R, Γ and coding/modulation schemes, in order to realize a reader of long read range, we summarize several key design considerations for the RF Front-end of the reader as follows:

1) A large reader antenna gain G_R is preferred for achieving a long range reader. A reader antenna of high gain could efficiently improve the SNRs of the demodulation output signals of the reader and consequently the read rang. Large G_R is also helpful to dense-multiple-reader applications, which would dramatically reduce the interference between multiple readers. In addition, it can be noticed that simply increasing the transmitting power P_T to a higher level, even without considering the limit of P_{EIRP}, would not contribute to improving the receiver's SNR performance any further, which is more or less beyond one's intuition.
2) A large receive-transmit isolation coefficient ξ is recommended for realizing a long range reader. The receive-transmit isolation performance of the reader, which is mainly limited by the circulator acting as the duplexer, has quite important effect on the SNR of the demodulation output signal of the reader. In order to improve this specification, we should choose a circulator having both low inherent leakage and low insertion loss. In addition, since the reflection signal from the reader antenna also enters the receiver and consequently deteriorates the circulator's isolation performance, the impedance matching between the reader antenna and the circulator should be tuned as elaborately as possible to improve ξ further.
3) A LO of low phase noise is strongly required for increasing the read range of a reader. It is the phase noise of the LO, not the internal thermal noise of the receiver, that has become the most significant noise source of the demodulation output signal of the reader, which is quite distinguished from traditional RF transceivers. Incidentally, considering that the phase noise is concentrated within a narrow band surrounding the carrier frequency and consequently the receiver noise is mainly

of low frequency component, the coding and modulation schemes adopted in a tag should be capable of suppressing DC and low frequency component of its data signal as more as possible. That would facilitate utilizing band-pass filters in the receiver to reject the harmful noise and simultaneously preserve the useful data signal as more as possible, which consequently could relax the requirement of the phase noise performance of the LO.

4) There could be two kinds of RFID systems according to the relations between the read range r_{READER} and r_{TAG}. One is called tag-determining RFID system, where the tag-side condition of $P_A \geq P_{TH}$ would not be met earlier than the reader-side condition of $SNR_{MIN} \geq SNR_{USER}$ while one separates the tag and the reader at a longer distance gradually, i.e., $r_{READER} > r_{TAG}$; the other is called reader-determining RFID system, which is just on the contrary. A good reader design should belong to the former in order to provide more compatibility and robustness for varieties of tags. It can be noticed that an RFID system could reach a compromise between r_{READER} and r_{TAG} through optimizing the power reflection coefficient Γ of the tag.

4. DESIGN EXAMPLE AND RESULTS

Based on the discussions in Section 3, we built a prototype reader with off-the-shelf components and set up a basic RFID system together with a commercial tag. The system operates at the frequency of 915 MHz. The tag has a half-wavelength dipole antenna that has a gain G_T of 2.15 dBi. Its input threshold power level P_{TH} is -15 dBm. The tag's data rate is 160 kbps. The unipolar FM0 coding is utilized for the tag's data sequence. The ASK modulation method is utilized for backscatter communication and the power reflection coefficient Γ takes a value of 1. The allowed maximum P_{EIRP} for the reader is 36 dBm according to the regulations of the United States. The reader has a vertical-linear-polarization panel antenna that has a gain G_R of 13 dBi and then the transmitting power P_T is constrained to 23 dBm. The receive-transmit isolation coefficient ξ of the reader is 20 dB, which is mainly limited by the circulator's performance. The baseband band-pass filter approximately has an ideal rectangular transfer function, whose low-end cutoff frequency f_L is 10 kHz and high-end cutoff frequency f_H is 320 kHz.

Utilizing (8a) in the current example, we can figure out that the tag-determining maximum operational distance r_{TAG} takes a value of 8.3 m. However, it could not be an easy task to evaluate r_{TAG} directly through measuring the time-averaging absorption power P_A

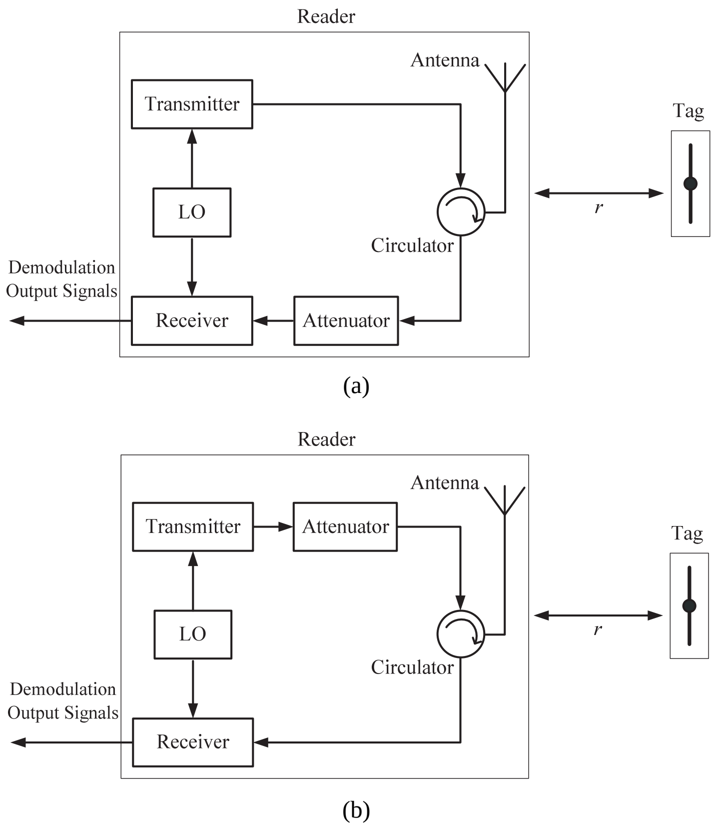

Figure 3. Comparative operational distance experiments. (a) Receiver-attenuation case. (b) Transmitter-attenuation case.

of the tag. Fortunately, here it is enough for us only to judge whether the system is of tag-determining or reader-determining. For this purpose, we conducted two comparative operational distance experiments where receiver-attenuation and transmitter-attenuation were utilized as illustrated in Fig. 3(a) and Fig. 3(b), respectively. We observed that the reader could not detect the tag's data signal earlier in the case of transmitter-attenuation than in the case of receiver-attenuation while the operational distance r being increased. Thus, we can conclude that this is a tag-determining RFID system and accordingly the calculated read range is given by the calculated r_{TAG}, i.e., 8.3 m. In addition, in the experiment of transmitter-attenuation, we got a maximum operational distance of 4.1 m while the attenuation takes a value of 6 dB. Then, based on (7a), we can get an estimated value of 8.2 m for r_{TAG}, which is close to both the calculated r_{TAG} above and the measured read range in the later experiment.

Before using (15a) to calculate the reader-determining maximum operational distance r_{READER}, we need to know the single-sideband in-band signal power P_{DATA} of the tag's binary data sequence and the band-limited phase noise power P_{PN} of the LO. Firstly, we manage to figure out the PDS of P_{DATA}. Since each tag has a unique data sequence of finite-length (64 bits for the current tag), one tag's PDS is generally different from another one's more or less. Here, instead of directly calculating the PDS of a specific tag, we evaluate it in a statistical-averaging-sense through constructing a random-generated data sequence of infinite-length. We assume that this sequence is a wide-sense stationary ergodic stochastic process. Then, we can achieve a good approximate result of its PDS through applying discrete fast Fourier transform (DFFT) to its any segment of finite-length (1024 bits here). Then, we numerically integrate its PSD within the receiver's pass-band of [10 kHz, 320 kHz], which gives that P_{DATA} takes a value of -6.8 dB. Secondly, we utilize a SPA directly to measure the normalized phase noise power of the LO within the receiver's bandwidth, which gives that P_{PN} takes a value of -39.9 dB. In addition, we need to know the phase noise improvement factor ψ. In order to evaluate ψ, we set up two comparative demodulation output noise experiments as illustrated in Fig. 4(a) and Fig. 4(b), respectively. The phase noise of the signal generator is much lower than the LO's and they are uncorrelated to each other. In addition, considering that a SPA device generally does not support accurate measurement of very-low-frequency signals, we utilize a digital oscilloscope to measure the power of demodulation output noise in terms of root mean squared (RMS) value. Then, through evaluating the ratio of the noise power measured in Fig. 4(b) to that measured in Fig. 4(a), we get that ψ takes a value of -40.9 dB. Now, substituting the user-specified SNR_{USER} of 12 dB into (15a), we get that the reader-determining maximum operational distance r_{READER} takes a value of 11.8 m, which shows further that this is a tag-determining RFID system.

We measured this RFID system in an open indoor environment. Here, in order to characterize the reader's performance in a practical operational scenario where there is lots of clutter from obstacles, we did not choose to conduct the measurements in an anechoic chamber. Fig. 5 shows the measured SNRs of the reader's demodulation output signals for several different r values of 2.0 m, 3.0 m, 4.0 m, 5.0 m, 6.0 m, 7.0 m and 8.4 m together with the calculated SNR_{MIN} using (14a). We get a maximum operational distance of 8.4 m, where the SNR of the demodulation output signal is 17.5 dB, which is well beyond the

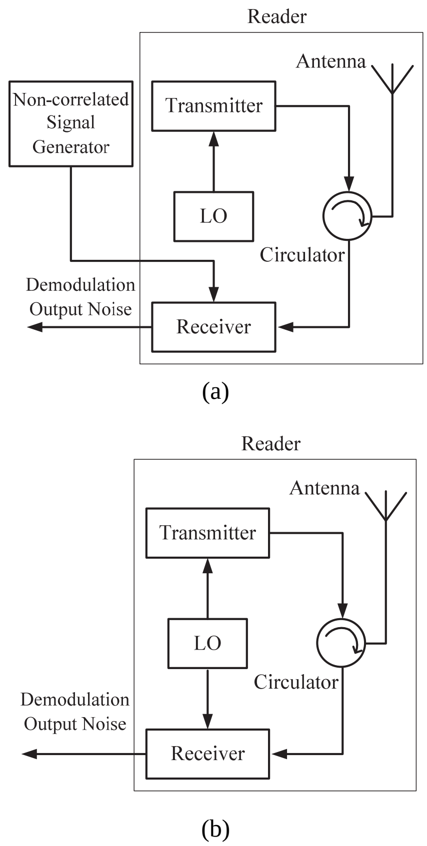

Figure 4. Comparative demodulation output noise experiments. (a) Non-correlation case. (b) Correlation case.

user-specified SNR_{USER} of 12 dB. Thus, this RFID system indeed is of tag-determining and accordingly the read range takes a value of 8.4 m.

In Fig. 5, we can see some offsets between the measured SNR and the calculated SNR_{MIN}. In addition, the calculated r_{TAG} is a little less than the measured read range. There could mainly be two reasons for the offsets. One is the indoor multi-path effects of electromagnetic wave propagation [19–21], which reduces the accuracy of the free space Friis electromagnetic wave propagation equation that has been utilized in (1) and (5). The other one is the inherent SNR uncertainty of the nor-correlate demodulation scheme as discussed in Section 3. However, these calculated results are still satisfactory enough to provide an insight into designing long read range readers.

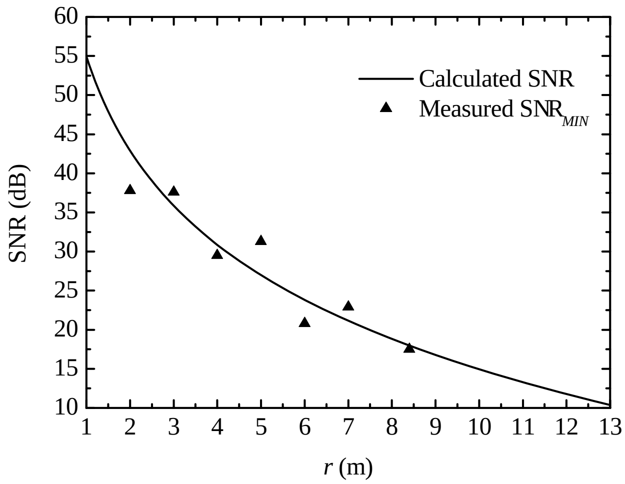

Figure 5. Calculated SNR_{MIN} and measured SNR of the reader's demodulation output signal for several different r values of 2.0 m, 3.0 m, 4.0 m, 5.0 m, 6.0 m, 7.0 m and 8.4 m.

5. CONCLUSION

In this paper, we outline a set of design considerations for the RF Front-end of passive backscatter UHF RFID readers in order to achieve long read range. We discuss signal descriptions for a basic passive backscatter UHF RFID system, propose the reader's architecture and present the design details of its RF frond-end section. We formulate the read range through calculating the time-averaging absorption power of the tag and the SNR of the demodulation output signal of the reader, which accordingly classifies RFID systems into tag-determining ones and reader-determining ones. It is observed that the gain of the reader antenna, the phase noise of the reader's local oscillation (LO) and the receive-transmit isolation coefficient mainly determine the demodulation output noise of the reader and consequently the reader-determining maximum operational distance. We also built a prototype reader operating at the frequency of 915 MHz with off-the-shelf components and evaluated it together with a commercial tag in an open indoor environment. The measurement shows that this RFID system is of tag-determining and has a read range of 8.4 m, which is in good agreement with the measured result and validates the design considerations presented in this paper.

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