Scatter Radio Receivers for Extended Range Environmental Sensing WSNs

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Abstract—**Backscatter communication, relying on the reflection principle, constitutes a promising-enabling technology for low-cost, large-scale, ubiquitous sensor networking. This work makes an overview of the state-of-the-art coherent and noncoherent scatter radio receivers that account for the peculiar signal model consisting of several microwave and communication parameters.**

I. INTRODUCTION

The need of ubiquitous environmental sensing has lead to the adoption of cost-effective, large-scale wireless sensor networks (WSNs). Such networks consists of several low-power and low-cost devices that sense environmental variables (such as soil humidity, soil moisture, temperature) and gathering the sensed information at a central unit. Existing commercial WSN equipment incorporates devices consisting of complex active radio frequency (RF) components that increase significantly the total monetary cost, as well as the overall energy consumption per sensor node.

Scatter radio adopts the reflection principle [1], which is achieved by generating a carrier wave that illuminates a set of RF tag/sensors. A RF tag terminates its antenna load according to the data to be transmitted. The incident signal is modulated and scattered back towards a software-defined radio (SDR) reader for processing. The above idea is depicted in Fig. 1.

This work proposes bistatic scatter radio technology with semi-passive tags and frequency shift-keying (FSK) modulation, ideal for the power limited regime [2]. The specific three-fold design mixture, not only reduces the energy cost per sensor tag, but also does not sacrifice in total communication range, achieving ranges on the order of a hundred of meters, as opposed to the conventional passive radio frequency identification (RFID) Gen 2 architecture that offers a limited range on the order of very few meters. A thorough overview of the state-of-the-art coherent and noncoherent scatter radio FSK receivers is subsequently conducted. The receivers are capable to achieve one hundred and fifty meters communication range.

II. BISTATIC SCATTER RADIO SIGNAL MODEL

The bistatic scatter radio architecture is employed, consisting of a carrier emitter, a RF tag, and a software-defined radio (SDR) reader (Fig. 1). Due to the relatively small bit-rate (on the order of few kilobits per second), along with the fact that the carrier emitter-to-SDR reader and tag-to-SDR reader links are on the order of a kilometer, i.e., small delay spread, frequency non-selective (flat) fading channel is assumed, where the baseband complex channel response for a duration of channel coherence time, T_{coh}, is given by $h_{\mathrm{k}}(t) = h_{\mathrm{k}} = a_{\mathrm{k}}\mathrm{e}^{-\mathrm{j}\phi_{\mathrm{k}}}$, $\mathrm{k} \in \{\mathrm{CR}, \mathrm{CT}, \mathrm{TR}\}$, where
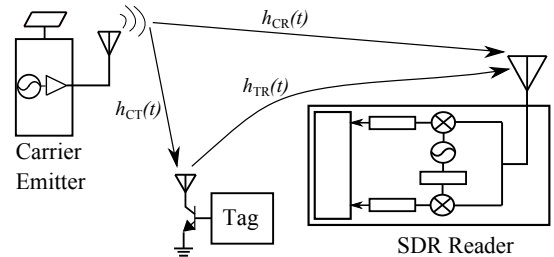

Fig. 1. Bistatic architecture system model: carrier emitter is displaced from SDR reader and RF tag modulates the incident RF signal from carrier emitter.

$a_{\mathrm{k}}, \in \mathbb{R}_+$, $\mathrm{k} \in \{\mathrm{CR}, \mathrm{CT}, \mathrm{TR}\}$, denote the channel attenuation parameters of the corresponding links and $\phi_{\mathrm{k}} \in [0, 2\pi)$, $\mathrm{k} \in \{\mathrm{CR}, \mathrm{CT}, \mathrm{TR}\}$, stand for the respective phases due to signal propagation delay, all independent of each other. Parameters $a_{\mathrm{CR}}, a_{\mathrm{CT}}, a_{\mathrm{TR}}$ are assumed Rician distributed due to strong line-of-site signals, commonly encountered in outdoors WSNs.

Carrier emitter transmits a continuous sinusoid wave at carrier frequency F_{car} that illuminates the tag. Tag modulates the received signal at passband using 50% duty cycle square waveform pulse of frequency F_i and random initial phase $\Phi_i \sim \mathcal{U}[0, 2\pi]$, $i \in \mathbb{B} \triangleq \{0, 1\}$ and backscatters it towards SDR reader. The received complex baseband signal at the SDR reader is given by the superposition of the carrier emitter sinusoid and the backscattered tag signal through channels h_{CR} and h_{TR}, respectively. The received signal also suffers from band-limited noise with power spectral density $N_0/2$ over SDR receiver bandwidth W_{SDR}. SDR reader applies carrier frequency offset (CFO) estimation and compensation using periodogram-based techniques, DC blocking, and synchronization.

For $|F_0 - F_1| \gg \frac{1}{T}$ and $F_i \gg \frac{1}{T}$, the noise-free, CFO-free, DC-blocked, and perfect synchronized baseband signal belongs to a four dimensional, time limited in $[0, T)$ signal space, whose orthonormal basis is denoted as $\mathcal{B} \triangleq \left\{ \frac{1}{\sqrt{T}}\mathrm{e}^{\pm\mathrm{j}2\pi F_i t}\Pi_T(t) \right\}_{i \in \mathbb{B}}$, where $\Pi_T(t)$ is the square waveform of duration T. The optimal demodulator projects received signal on basis $\mathcal{B}$ through correlators and the discrete baseband signal over bit duration T can be written as [3]

$$\mathbf{r} = \begin{bmatrix} r_0^+ \\ r_0^- \\ r_1^+ \\ r_1^- \end{bmatrix} = h\sqrt{\frac{\mathrm{E}}{2}} \begin{bmatrix} \mathrm{e}^{+\mathrm{j}\Phi_0} \\ \mathrm{e}^{-\mathrm{j}\Phi_0} \\ \mathrm{e}^{+\mathrm{j}\Phi_1} \\ \mathrm{e}^{-\mathrm{j}\Phi_1} \end{bmatrix} \odot \mathbf{s}_i + \begin{bmatrix} n_0^+ \\ n_0^- \\ n_1^+ \\ n_1^- \end{bmatrix}, \quad (1)$$

where $h = \alpha_{\mathrm{CT}}\alpha_{\mathrm{TR}}\mathrm{e}^{-\mathrm{j}\phi}$ incorporates the fading coefficients $h_{\mathrm{CT}}h_{\mathrm{TR}}$ as well as the phase difference of compound link

carrier emitter-to-tag-to-SDR reader. E is the average energy per bit in baseband, $\mathbf{s}_i = [(1-i)\ (1-i)\ i\ i]^\top$ is the four-dimensional symbol corresponding to bit $i \in \mathbb{B}$, and $\mathbf{n} = [n_0^+\ n_0^-\ n_1^+\ n_1^-]^\top \sim \mathcal{CN}\left(\mathbf{0}_4, \frac{N_0}{2}\mathbf{I}_4\right)$.

III. BISTATIC SCATTER RADIO FSK RECEIVERS

A. Uncoded Reception

1) Coherent Detector [3]: The coherent receiver estimates the compound channel $\mathbf{h}_c \triangleq h\sqrt{\frac{E}{2}}\left[\mathsf{e}^{+j\Phi_0}\ \mathsf{e}^{-j\Phi_0}\ \mathsf{e}^{+j\Phi_1}\ \mathsf{e}^{-j\Phi_1}\right]$ through the use of training symbols. After obtaining the least-squares estimate of compound channel $\mathbf{h}_c$, $\widehat{\mathbf{h}}_c$, the maximum-likelihood (ML) detector is applied through the rule [3]

$$\widehat{i}_{\mathrm{ML}} = \arg\max_{i\in\mathbb{B}} \Re\left\{\left(\widehat{\mathbf{h}}_c \odot \mathbf{s}_i\right)^H \mathbf{r}\right\}. \tag{2}$$

2) Noncoherent Detectors [4]: The first symbol-by-symbol noncoherent detector treats the parameter h as deterministic and parameters $\{\Phi_i\}_{i\in\mathbb{B}}$ as random; it is called hybrid composite hypothesis testing (HCHT) detector and is given by [5]

$$\arg\max_{i\in\mathbb{B}} \left\{\mathbb{E}_{\Phi_i}\left[\max_{h\in\mathbb{C}} \ln(\mathsf{f}(\mathbf{r}|i,h,\Phi_i))\right]\right\} \tag{3}$$

$$\Longleftrightarrow |r_0^+|^2 + |r_0^-|^2 \underset{i=1}{\overset{i=0}{\gtrless}} |r_1^+|^2 + |r_1^-|^2. \tag{4}$$

The second symbol-by-symbol detector is the generalized-likelihood ratio test (GLRT) detector, that treats all unknown parameters as deterministic and is expressed as [4]

$$\arg\max_{i\in\mathbb{B}} \left\{\max_{\Phi_i\in[0,2\pi]} \max_{h\in\mathbb{C}} \ln(\mathsf{f}(\mathbf{r}|i,h,\Phi_i))\right\} \tag{5}$$

$$\Longleftrightarrow |r_0^+| + |r_0^-| \underset{i=1}{\overset{i=0}{\gtrless}} |r_1^+| + |r_1^-|. \tag{6}$$

For a bit sequence of N bits satisfying $T_{\mathrm{coh}} \geq NT$, the received sequence a can be written as $\mathbf{r}_{1:N} = [\mathbf{r}_1\ \mathbf{r}_2\ \ldots\ \mathbf{r}_N] = h\left[\mathbf{x}_{i_1}(\Phi_{i_1})\ \mathbf{x}_{i_2}(\Phi_{i_2})\ \ldots\ \mathbf{x}_{i_N}(\Phi_{i_N})\right] + [\mathbf{n}_1\ \mathbf{n}_2\ \ldots\ \mathbf{n}_N]$, with $\mathbf{x}_{i_n}(\Phi_{i_n}) \triangleq \sqrt{\frac{E}{2}}\left[\mathsf{e}^{j\Phi_0}\ \mathsf{e}^{-j\Phi_0}\ \mathsf{e}^{j\Phi_1}\ \mathsf{e}^{-j\Phi_1}\right]^\top \odot \mathbf{s}_{i_n}$, with $i_n \in \mathbb{B}$, $n = 1, 2, \ldots, N$. To reduce the sequence error rate, a noncoherent sequence detector may be applied. For the above signal model, the GLRT sequence detector is expressed as

$$\arg\max_{\mathbf{i}\in\mathbb{B}^N} \left\{\max_{(\Phi_0,\Phi_1)\in[0,2\pi]^2} \max_{h\in\mathbb{C}} \ln(\mathsf{f}(\mathbf{r}_{1:N}|\mathbf{i},h,\Phi_0,\Phi_1))\right\}. \tag{7}$$

Work in [4] partitions the space of phases (Φ_0, Φ_1) in distinct M^2 points and solves the above problem with complexity $\mathcal{O}(N\log(N))$ using the same procedure with the work in [6].

B. Coded Reception

When channel coding is utilized, the transmitter encodes a sequence of K bits to a sequence of N coded bits, $\mathbf{c} = [c_1\ c_2\ \ldots, c_N]$ belonging to a code $\mathcal{C} \subset \mathbb{B}^N$.

1) Coherent Decoder [3]: After obtaining the least-squares estimate of compound channel $\mathbf{h}_c$, $\widehat{\mathbf{h}}_c$, the ML decoder can be expressed as [3]

$$\widehat{\mathbf{c}}_{\mathrm{ML}} = \arg\max_{\mathbf{c}\in\mathcal{C}} \sum_{n=1}^{N} \Re\left\{\left(\widehat{\mathbf{h}}_c \odot \mathbf{s}_{c_n}\right)^H \mathbf{r}_n\right\} \tag{8}$$

It can be shown that, if the above decoder is combined with interleaving technique, it can achieve diversity order $d_{\min}$, where $d_{\min}$ is the minimum distance of code $\mathcal{C}$ [3].
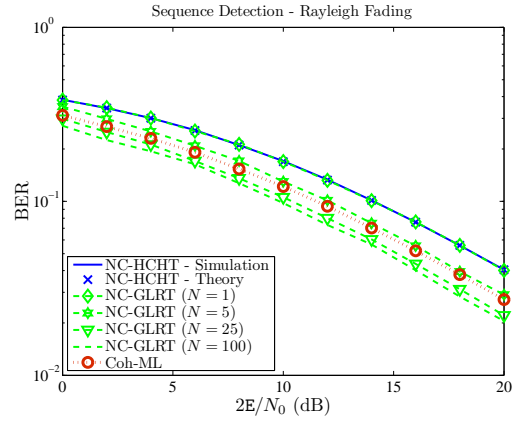

Fig. 2. BER vs average received SNR for uncoded schemes. Due to fixed energy budget per transmitted packet and the fact that coherent receiver employees extra 30 bits for channel estimation (i.e., it has less energy per bit), noncoherent GLRT sequence detector outperforms coherent one.

2) Nonoherent Decoder [4]: Suppose that we employ channel coding, noncoherent receiver, and the use of interleaving technique of interleaving depth D satisfying $DT \geq T_{\mathrm{coh}}$; accordingly with Eq. (3), HCHT decoder can be derived as [5]

$$\widehat{\mathbf{c}} = \arg\max_{\mathbf{c}\in\mathcal{C}} \sum_{n=1}^{N} \left(\|\mathbf{r}_n \odot \mathbf{s}_1\|^2 - (\|\mathbf{r}_n \odot \mathbf{s}_0\|^2\right) c_n. \tag{9}$$

For the case of $DT < T_{\mathrm{coh}}$, the above decoder is suboptimal, however, it is utilized for any value of D, due to its inherent simplicity.

IV. SIMULATION RESULTS

Simulations are conducted for bistatic scatter radio system model assuming backscatter FSK transmissions with bit rate $T = 1$msec, over quasi-static Rician flat fading channel with coherence time $T_{\mathrm{coh}} = 100$msec. In all simulation results we have assumed perfect synchronization, perfect CFO estimation/compensation, and fixed energy budget per transmitted packet for all schemes.

Fig. 2 illustrates the bit error rate (BER) performance of all uncoded schemes studied in Section III-A as function of average received signal-to-noise ratio, $\frac{2E}{N_0}$. We observe that GLRT sequence detector outperforms all schemes. It is also noted that coherent ML detector offers 3dB better BER performance compared to noncoherent symbol-by-symbol schemes.

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