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An Information-Theoretic Approach to the Chipless RFID Tag Identification

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ABSTRACT In this paper, we focus on the chipless radio-frequency identification (RFID), where the tag information bits are encoded by the peak/notch pattern appeared in the frequency spectrum of the radar cross section (RCS) of the tag. In particular, we restrict our attention to a simple yet prevalent ''binary'' coding method, where a bit 0 or bit 1 is encoded by the absence or presence of the peak/notch, respectively. We provide an information-theoretic framework for the tag identification based on such a binary coding method. Our aim is to accommodate more bits in the limited bandwidth without degrading the identification performance. To this end, we first formulate the detection of each bit as a binary asymmetric channel (where a signal processing approach is integrated into each interrogation to enhance the underlying channel quality). Moreover, it is proposed to perform multiple interrogations with majority rule-based detection (in correspondence to the signal processing approach in each interrogation). Furthermore, we introduce some error-detecting codes to further improve the performance of tag identification. For instance, motivated by the asymmetric property of the channel model, we propose to apply the constant weight code and the Berger–Freiman code (as two representatives of non-systematic and systematic codes, respectively) to the problem to be addressed in this paper. In addition, an investigation is also conducted into the cyclic redundancy check (CRC) codes (as a representative of those codes that are not dedicated to the binary asymmetric channel but could be potentially competitive for error detection). The system's performance is analyzed through the key parameters, namely the successful transmission rate, the false identification rate (i.e., the probability of undetected errors), and the expected number of retransmissions/interrogations. The effectiveness of the proposed methods is demonstrated by the numerical results.

INDEX TERMS Chipless tags, binary asymmetric channel, error detecting codes, RFID, majority-rule interrogation, performance analysis, constant weight code, Berger–Freiman code, cyclic redundancy check (CRC) code.

I. INTRODUCTION

Radio frequency identification (RFID) is a technique that utilizes radio frequency (RF) waves to detect and identify information from tags attached to objects. The system comprises of two main components: a transponder/tag that contains a sequence of electronic codes used for the identification of an object, and an interrogator/reader that collects information from the tag. The RFID technology has the advantage of automatic detection and non-line of sight operation. Therefore, it is believed to have the potential to substitute the widely used barcode technology in the near future. The main barrier for RFID to replace barcode technique is the high per-tag price (about 10 cents) of RFID tags compared to barcode (less than 0.1 cents); while, chipless RFID, which, as its name suggests, does not contain any digital chip and the price of which can be made down to one cent if manufactured with printing technology [1], [2], could be a promising substitute of barcode technique [23].

Research on chipless RFID tags can be broadly classified into two main categories: time-domain reflectometry-based (TDR-based) chipless RFID and frequency-domain

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spectral-signature-based (FDSS-based) RFID. In TDR-based tags, the RFID reader transmits an ultra-wideband (UWB) RF interrogation pulse and listens to the reflections or echoes coming back from the tag [3]–[8]. By varying the structural properties of the tag, the time of arrival of these echoes can be controlled, providing a method for passive data storage in the tag. In FDSS-based tags [9]–[12], the frequency spectrum of the interrogation signal sent by the RFID reader is transformed by the tag to represent data bits. Generally speaking, an FDSS-based chipless tag consists of several resonators etched on a common grounded dielectric substrate. These resonators generate a series of notches at particular frequencies in the spectrum of the tag's frequency response. These notches can be used to encode the tag ID. For instance, a simple yet prevalent tag-ID coding method is to use a ''binary'' coding, i.e., encoding a bit 0 or 1 by the absence or presence of the notch, respectively. In this paper, we focus our discussion on this ''binary'' coding method.

In the current single-layer (2-D) design techniques, typically, each notch in an FDSS-based chipless tag occupies a bandwidth 100$\sim$200 MHz [13], [14]. In a multilayer (3-D) design structure, by folding several cascaded commensurate transmission-line sections into a package, it is possible to obtain several bits in a bandwidth of 100 MHz, as shown in a recent paper [24]. In order to accommodate reasonably high number of bits into a chipless tag, the tag should be operated in UWB frequency range. In addition, the transmit power of the reader is extremely low due to the spectral mask for UWB signals as specified by relevant authorities [15], [16]. As a direct consequence, weak notches in the spectrum of the backscattered signal might be difficult to be detected or even disappear in a noisy environment. Then the information bits corresponding to these weak notches will be erroneously identified, and so will be the tag ID. Therefore, an effective detection algorithm is desirable (with a high successful identification rate). Or, if there are errors, they should be detected (i.e., with a low false identification rate) and a retransmission can be automatically triggered. We note that differently from the problems in communications, to perform multiple times of interrogations is almost cost-free for the reader (within acceptable delay in reading time). In addition, the bandwidth for the FDSS-based chipless tag is very precious, thus it is also desirable to accommodate more bits within a certain bandwidth.

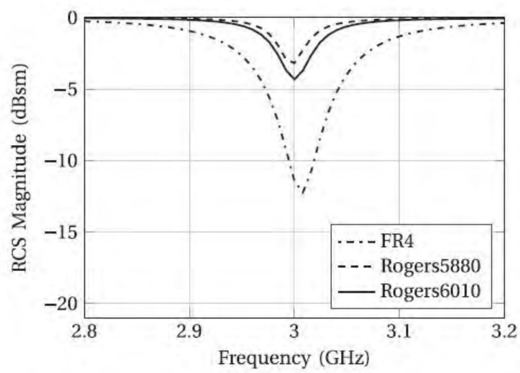
In this paper, we address the problem of RFID tag identification from an information-theoretic perspective. For simplicity, we restrict our attention to the binary coding for the tag IDs. Our first attempt is to treat the detection of each bit as a binary asymmetric channel (BAC), and the first insight gained is that false identification is unavoidable if the tag IDs are uncoded. In order to improve the performance of tag identification, we propose three countermeasures. The first is to take a signal processing approach (e.g., using a matched filter) before the bit detection takes place. The idea is to maximize the signal-to-noise ratio (SNR) of the underlying channel in the presence of the additive Gaussian noise. The second is to perform multiple times of interrogations and then decide each bit of the tag based on a majority rule. The idea is to increase the successful identification rate of each bit by further improving the underneath channel quality. Unlike the signal processing approach that assumes a time-invariant channel state, this approach does not need this assumption and thus could be more robust and implementable in practice. The last but not least, we introduce the error detecting code to further diminish the false identification rate. That is, if some bits in the tag are erroneously detected, most of the error patterns can be detected and a retransmission will be automatically triggered. Combining the signal processing approach in each interrogation, multiple interrogations & majority rule based detection, and the error detecting codes, the expected number of retransmissions/interrogations (i.e., waiting time for a successful identification) can be made acceptably small. All these make it possible to accommodate more data within a fixed bandwidth range, although the channel quality corresponding to each bit detection might be degraded due to the increased number of bits. Note that applying some kinds of error detecting/correcting codes to traditional communication systems is a de facto standard, but the coding performance is often analyzed under the assumption that the underlying channel is symmetric. The study of error detecting/correcting codes in a BAC scenario is far more challenging and as a consequence only performance bounds of the coded systems are available in literature [17]. In this paper, we will consider two classical codes for BAC, namely the constant weight code and Berger-Freiman code, as representatives of non-systematic and systematic codes; and one class of codes well-known for error detection, i.e., the cyclic redundancy check (CRC) codes, as a representative of those codes that are not dedicated to BAC but could be potentially competitive. Precise analytical and numerical results are provided to show the effectiveness of applying aforementioned codes to the chipless tag coding problem, which possesses a typical BAC model.

The rest of the paper is organized as follows. In Section II, we address the problem formulation of the chipless RFID tag detection, in which we integrate a signal processing approach. In Section III, we propose multiple interrogations with majority rule based detection and its impact is analyzed from an information theoretic point of view. A preliminary performance analysis is conducted in Section IV for the uncoded RFID tags. In Section V, we introduce the error detecting codes to reduce the false identification rate. Numerical results are presented in Section VI to illustrate the effect of the proposed signal processing approach and the majority rule based detection, and the performance of the error detecting coding schemes. Finally, Section VII concludes the paper.

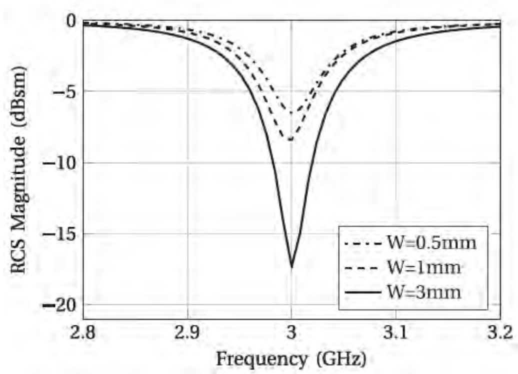
II. CHIPLESS TAG DETECTION

In this section, we will present the channel model for chipless tag identification problem. An explication for this issue can be found in [22]. For the reason of completeness and readers' convenience, the channel model will be briefly discussed in the following.

From an engineering point of view, the effectiveness of detecting a bit in an FDSS-based chipless tag can be characterized by the depth of the notch, which is determined by the RCS of the corresponding resonator. The deeper the notch in the RCS is, the more effectively the tag can represent the desired bit and more easily the reader can detect the bit. Fig. 1 gives a typical illustration for the shape of the RCS of a co-polarized resonator, where we can see that the depth and bandwidth of the notch change with the material of the substrate and the size (e.g. the length and width) of the slot line.



(a) RCS changes with substrate material



(b) RCS changes with the size of resonator slot line

FIGURE 1. RCS changes with material of the substrate and size of the resonator slot line. In (a), FR 4, Rogers 5880 and Rogers 6010 are three different dielectric materials for the substrate; in (b), *W* is the width of the dipole antenna. [From [18]. Reproduced with permission of ©IEEE 2017.].

We remark that in discussing tag ID coding, one can equivalently use a peak, instead of a notch, in the frequency spectrum of the resonator to encode a bit. Mathematically, it is more convenient to use a peak. Therefore, in the rest of the paper, we will adopt the peak coding way, and G_{peak} and B, represent the strength and bandwidth of the peak, respectively.

A. CHANNEL MODEL FOR CHIPLESS TAG DETECTION

Suppose that the peak frequencies in the resonators of a chipless tag are used to encode the tag ID. More specifically, the presence and absence of a peak in a specific frequency (window) are interpreted as "1" and "0", respectively in the corresponding bit.

To detect the tag, the reader transmits a frequency-sweeping signal or a UWB signal. Then the spectrum of the received signal is analyzed in order to decide whether the tag exists and what the coded bits are if the tag exists. In practical systems, the tag can be assumed to exist. Suppose that f_i is the resonating frequency of the i-th resonator and its calibrated receive signal can be expressed as

$$y_i = h_i \cdot x_i + n_i,$$

where $x_i = 0, 1$ with 1 representing that the i-th resonator is present while 0 not; n_i is an additive white Gaussian noise, the power spectral density of which is

$$N_0 = k_{\text{B}} T_0,$$

where $k_{\text{B}} = 1.3806 \times 10^{-23}$ J/K is Boltzmann constant and $T_0 = 300$ K is reference temperature. The power of the noise n_i added to the calibrated receive signal corresponding to the i-th resonator is

$$\sigma_{n_i}^2 = N_0 \, B_i = k_{\text{B}} T_0 B_i.$$

And, $h_i > 0$ stands for the channel coefficient. Suppose that the signal undergoes a free-space propagation. Then the forward and backward channel for the reader-to-tag link can be modeled with

$$|H_{\text{F},i}|^2 = |H_{\text{B},i}|^2 = \left(\frac{c}{4\pi f_i d}\right)^2, \tag{1}$$

where c is the speed of light, d the distance between the reader and tag, f_i the resonant frequency, and $H_{\text{F},i}$, $H_{\text{B},i}$ the transfer functions of the forward and backward channel, respectively. In this ideal scenario, we have

$$|h_i|^2 = |H_{\text{F},i}|^2 |H_{\text{B},i}|^2 \, G_{\text{peak}} P_{\text{rd}}, \tag{2}$$

where P_{rd} is the reader's transmit power. If we use $\sigma_{\text{RCS}}(f)$ to denote the RCS of the chipless tag at frequency f, then G_{peak} in equation (2) should be replaced with $\sigma_{\text{RCS}}(f_i)$. Since it is assumed that $\sigma_{\text{RCS}}(f_i)$ is of the same value for different notches (i.e., $\sigma_{\text{RCS}}(f_i) = G_{\text{peak}}$ for different resonant frequencies f_i), we directly use G_{peak} in equation (2) to simplify notations. For simplicity, we assume that h_i is a constant.

If the reader makes its decision after several rounds (say κ_{s}) of sending and recording the backscattered signals for each round, then a signal processing approach can be taken to enhance the SNR of the overall channel. (That is, in this paper, an *interrogation* involves of κ_{s} rounds of sending and recording received signals. Detection of bits can be done after each interrogation). Since both h_i and x_i are constants in one interrogation, i.e., κ_{s} times of sending and receiving signals, the received signals vary only due to the noise, i.e.,

$$y_{i,j} = h_i \cdot x_i + n_{i,j}, \quad \text{where } j = 1, \ldots, \kappa_{\text{s}},$$

where $n_{i,j}$, for $j = 1, \ldots, \kappa_s$ are i.i.d. white Gaussian noise with variance $\sigma_{n_i}^2$. Simply summing up all the received signals and considering $y_i' = \frac{1}{\kappa_s}\sum_{j=1}^{\kappa_s} y_{i,j}$, we have

$$y_i' = h_i \cdot x_i + n_i',$$

where $n_i' = \frac{1}{\kappa_s}\sum_{j=1}^{\kappa_s} n_{i,j}$ and n_i' is a white Gaussian noise with variance $\sigma_{n_i}^2/\kappa_s$.

The reader makes the following decision according to a given threshold θ_{th}:

$$\hat{x}_i = \begin{cases} 1, & \text{when } |y_i'| \geq \theta_{\text{th}}, \\ 0, & \text{when } |y_i'| < \theta_{\text{th}}. \end{cases}$$

An error occurs if $\hat{x}_i \neq x_i$. Clearly, we have two kinds of errors: where $x_i = 0$ while $\hat{x}_i = 1$ and $x_i = 1$ while $\hat{x}_i = 0$. The probability of these errors is given as follows:

$$\begin{aligned} \epsilon_0 :&= \Pr\{\hat{x}_i = 1 | x_i = 0\} = \Pr\left\{|n_i'| \geq \theta_{\text{th}}\right\} \\ &= \Pr\left\{\left|\frac{n_i'}{\sigma_{n_i}/\sqrt{\kappa_s}}\right| \geq \frac{\theta_{\text{th}}}{\sigma_{n_i}/\sqrt{\kappa_s}}\right\} \\ &= 2Q\left(\frac{\theta_{\text{th}}}{\sigma_{n_i}/\sqrt{\kappa_s}}\right), \end{aligned} \tag{3}$$

$$\begin{aligned} \epsilon_1 :&= \Pr\{\hat{x}_i = 0 | x_i = 1\} = \Pr\left\{\left|h_i + n_i'\right| < \theta_{\text{th}}\right\} \\ &= \Pr\left\{-\frac{h_i + \theta_{\text{th}}}{\sigma_{n_i}/\sqrt{\kappa_s}} < \frac{n_i'}{\sigma_{n_i}/\sqrt{\kappa_s}} < -\frac{h_i - \theta_{\text{th}}}{\sigma_{n_i}/\sqrt{\kappa_s}}\right\} \\ &= Q\left(\frac{h_i - \theta_{\text{th}}}{\sigma_{n_i}/\sqrt{\kappa_s}}\right) - Q\left(\frac{h_i + \theta_{\text{th}}}{\sigma_{n_i}/\sqrt{\kappa_s}}\right), \end{aligned} \tag{4}$$

where $Q(\cdot)$ stands for the Q-function.

So far, the channel for detecting a single bit of a chipless tag can be regarded as a BAC, which has a probability ϵ_0 that the bit symbol 0 will be read as 1 and a (possibly different) probability ϵ_1 that the bit symbol 1 will be read as 0. We denote it as BAC(ϵ_0, ϵ_1). See Fig. 2 (a).

One may argue that the channel can be made symmetric if the value of parameter θ_{th} is properly tuned so that $\epsilon_0 = \epsilon_1$. Even though this is theoretically possible, it is very difficult in practice to do this kind of tuning since ϵ_1 depends not only on the reader's transmit power, which can be fixed for some specified reading scenarios during a relatively long reading period, but also on the reading distance, which often varies constantly, while ϵ_0 does not. Therefore we often have that $\epsilon_0 \neq \epsilon_1$.

Note that the capacity of the BAC(ϵ_0, ϵ_1) is [19]

$$\begin{aligned} C_{\text{BAC}(\epsilon_0,\epsilon_1)} = \frac{\epsilon_0}{1-\epsilon_0-\epsilon_1} h_b(\epsilon_1) - \frac{1-\epsilon_1}{1-\epsilon_0-\epsilon_1} h_b(\epsilon_0) \\ + \log_2\left(1 + 2^{\frac{h_b(\epsilon_0)-h_b(\epsilon_1)}{1-\epsilon_0-\epsilon_1}}\right), \end{aligned} \tag{5}$$

where $h_b(\cdot)$ is the binary entropy function with $h_b(\epsilon) = -\epsilon\log_2(\epsilon) - (1-\epsilon)\log_2(1-\epsilon)$. In particular, if $\epsilon_0 = \epsilon_1$, the capacity reduces to the capacity of the binary symmetric channel (BSC) with crossover probability ϵ_0, which is $1 - h_b(\epsilon_0)$.

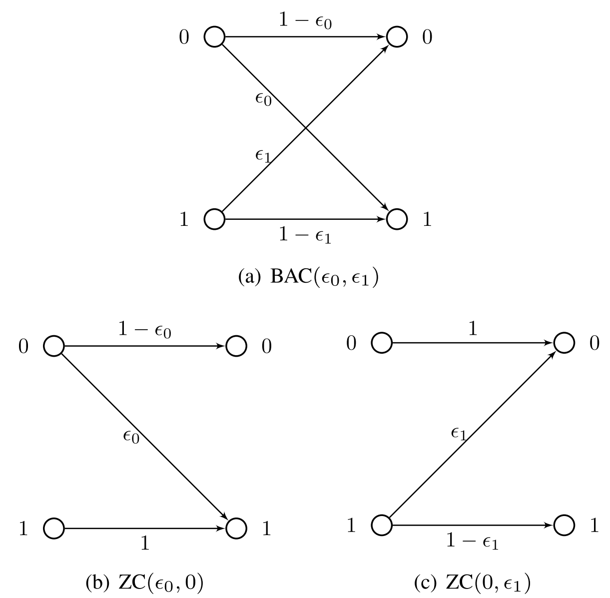


FIGURE 2. Channel model for detecting a single bit of a chipless tag.

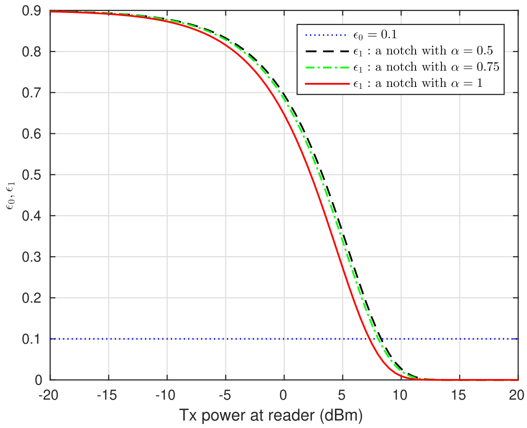
From equations (1) and (2) we see that h_i is different for a different frequency f_i. This will make the analysis below unnecessarily complicated. To further simplify notations, we assume that the effect of different frequencies on channel propagation loss is calibrated so that we can use the center frequency of the concerned frequency band to calculate h_i.

B. SOME OBSERVATIONS

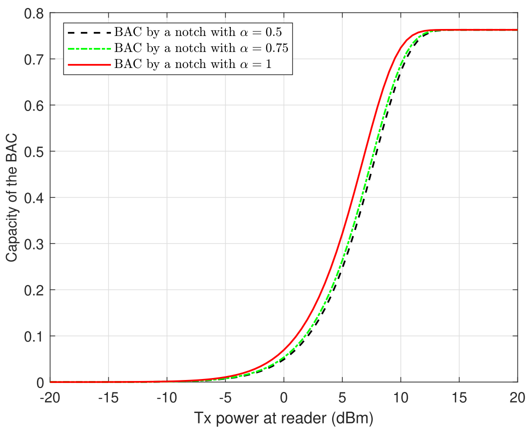
From Fig. 1, it can be seen that a deeper notch occupies a wider bandwidth. Therefore, if we want to use a deeper notch to encode a bit for the tag ID (for a more effective detection), fewer bits can be accommodated in the available frequency band. On the other hand, if a shallower notch is used, we can insert more bits to encode the tag ID. Note that by taking different choices of the notches (i.e., different choices of B and G_{peak}), the corresponding binary asymmetric channel varies.

To have a close look, we consider the following example with fixed $d = 1$ m and $f = 7.25$ GHz. At the starting point, we take $B = 112$ MHz and $G_{\text{peak}} = 8$ dB. To ease analysis, we assume that there is a simple relationship between the changes of parameters B (in MHz) and G_{peak} (in dB): if the notch bandwidth B is changed to $B_\alpha = \alpha B$ MHz, then the notch peak gain G_{peak} is changed to $G_{\text{peak},\alpha} = \alpha G_{\text{peak}}$ dB accordingly, where $0 < \alpha \leq 1$. Note that each notch may correspond to a different chipless tag design as indicated in Fig. 1.

For a fair and easy comparison, we fix that $\epsilon_0 = 0.1$ (the choice for this value of ϵ_0 is quite practical in chipless RFID systems) by taking $\theta_{\text{th}} = 1.645\sigma_{n_\alpha}$ for each design (in this case, the threshold θ_{th} needs to be tuned corresponding to different choices of α), where $\sigma_{n_\alpha}^2 = k_B T_0 B_\alpha/\kappa_s$, and κ_s is the number of the rounds (of sending and recording received signals in one interrogation) taken in the signal processing



(a) ϵ_1 w.r.t. P_{rd} for fixed $\epsilon_0 = 0.1$ and $\kappa_{\mathrm{s}} = 1$.



(b) Capacity of BAC defined by αB and αG_{peak}

FIGURE 3. BACs corresponding to notches defined by αB and αG_{peak}, for fixed $\kappa_{\mathrm{s}} = 1$.

approach. For a fixed κ_{s}, ϵ_1 becomes a function of the reader's transmit power P_{rd} (see (2) and (4)) as can be observed in Fig. 3 (a) (where $\kappa_{\mathrm{s}} = 1$).

Remarkably, a shallower notch may correspond to a worse channel (due to the higher ϵ_1, that results in a smaller channel capacity as can be observed in Fig. 3 (b)) at the same transmit power. Or, alternatively, more transmission power might be required to maintain the same channel quality as the one by using a deeper notch. Therefore, given the available frequency band, if using shallower notches, one can use more bits to code the tag ID; however, errors in each bit may occur more frequently. The general question of great interest is: how to increase the information bits of the tag IDs without degrading the identification performance of the chipless tags?

Another interesting observation (in the case of fixed ϵ_0 and κ_{s}) is that, as the reader's transmit power P_{rd} increases, the probability ϵ_1 will converge to 0. That is, in the high transmit power regime, the BAC channel model is reduced to a Z-channel (ZC), denoted by ZC(ϵ_0, 0) as shown in Fig. 2 (b). (Note that a ZC(ϵ_0, 0) is in fact a BAC(ϵ_0, 0); and a ZC(0, ϵ_1), as shown in Fig. 2 (c), is in fact a BAC(0, ϵ_1).)

III. MULTIPLE INTERROGATIONS & MAJORITY RULE BASED DETECTION

In the previous section, a signal processing approach (that involves κ_{s} rounds of sending and recording the received signals in one interrogation) is introduced; and a BAC is formulated for the bit detection that takes place after each interrogation. However, it is interesting to investigate how multiple interrogations can help the reader improve its successful identification rate quantitatively.

If the reader makes its decision after issuing several rounds (say $2\kappa_0 + 1$) of interrogation commands and recording the detected bits for each round of interrogation, then a majority rule can be applied to help identify each bit. That is, the bit 0 will be decoded as 1 after $2\kappa_0 + 1$ rounds if it is read as 0 in less than $\kappa_0 + 1$ rounds. So the probability ϵ_0' that a 0 will be read as 1 at each bit becomes

$$\epsilon_0' = \sum_{\kappa=0}^{\kappa_0} \binom{2\kappa_0 + 1}{\kappa} (1 - \epsilon_0)^{\kappa} \epsilon_0^{2\kappa_0+1-\kappa}. \tag{6}$$

Note that ϵ_0', as an increasing function of ϵ_0, equals to ϵ_0 at $\epsilon_0 = 0, 1/2$ and 1. Interestingly, for $0 \le \epsilon_0 \le 1/2$, we have $\epsilon_0' \le \epsilon_0$; while for $1/2 \le \epsilon_0 \le 1$, we have $\epsilon_0' \ge \epsilon_0$. That is, applying majority rule may result in a polarization effect in the error probability of $0 \to 1$.

Similarly, the probability ϵ_1' that a 1 will be read as 0 at each bit becomes

$$\epsilon_1' = \sum_{\kappa=0}^{\kappa_0} \binom{2\kappa_0 + 1}{\kappa} (1 - \epsilon_1)^{\kappa} \epsilon_1^{2\kappa_0+1-\kappa}. \tag{7}$$

If applying majority rule after $2\kappa_0 + 1$ rounds of interrogations, then the channel for detecting a single bit, can be regarded as a BAC(ϵ_0', ϵ_1'), as shown in Fig. 4. Remarkably, the channel quality could be polarized according to the values of ϵ_0 and ϵ_1. See a numerical illustration on such a polarization effect in Section VI-C.

In this paper, we say that one *transmission* involves of $2\kappa_0 + 1$ rounds of interrogations. Therefore, BAC(ϵ_0', ϵ_1') is formulated for the overall channel of bit detection after one transmission (that involves of $2\kappa_0 + 1$ rounds of interrogations; and each interrogation κ_{s} rounds of sending and recording received signals for the signal processing approach).

Recall that for a pre-specified $\kappa_{\mathrm{s}} = 1$ and $\epsilon_0 = 0.1$, we have $\epsilon_1 \to 0$ if the reader transmit power is quite high (as one can observe from Fig. 3 (a)). In this case, the channel to detect each bit can be regarded as a ZC(ϵ_0, 0) in the form of Fig. 2 (b). Besides, for those BAC(ϵ_0, ϵ_1) with $0 < \epsilon_0 \ll \epsilon_1 < 0.5$ (which can be obtained by taking appropriate choices of κ_{s} and θ_{th}), after $2\kappa_0+1$ rounds, we can obtain a BAC(ϵ_0', ϵ_1') with $\epsilon_0' \to 0$, then the channel can be regarded as a ZC(0, ϵ_1') in the form of Fig. 2 (c). Without loss of the generality, we restrict our attention to ZC(0, ϵ_1'). Nevertheless, our discussion, especially coding on ZC(0, ϵ_1') can be applied to ZC(ϵ_0', 0) (with minor modifications).

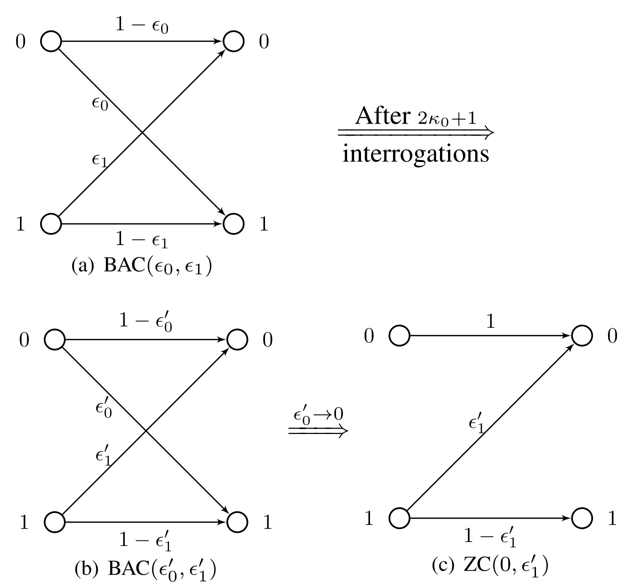


FIGURE 4. Channel model for each bit after $2\kappa_0 + 1$ interrogations.

IV. UNCODED CHIPLESS TAG ID

Suppose that there are l bits in a chipless tag and assume that there is no cross-interference between bits. The channel model for the detection of each bit is a BAC$(\epsilon_0', \epsilon_1')$ as shown in Fig. 4 (b) (for some κ_s and κ_0), where ϵ_0', ϵ_1' are as defined in (6) and (7), respectively.

In the uncoded case, l bits can be used to represent 2^l tag IDs (via a one-to-one mapping). Denote the set of binary sequences of length l to be $\mathcal{C}_{\text{uncoded}(l)}$. Then, the rate of $\mathcal{C}_{\text{uncoded}(l)}$ is

$$R_{\mathcal{C}_{\text{uncoded}(l)}} = 1. \tag{8}$$

For a l-bit tag ID, say $\mathbf{x} \in \mathcal{C}_{\text{uncoded}(l)}$, which has a Hamming weight $\mathrm{w}(\mathbf{x})$, (i.e., $\mathbf{x}$ has $\mathrm{w}(\mathbf{x})$ 1's and $l - \mathrm{w}(\mathbf{x})$ 0's), the probability of successful identification in one transmission is $(1-\epsilon_0')^{l-\mathrm{w}(\mathbf{x})}(1-\epsilon_1')^{\mathrm{w}(\mathbf{x})}$. Suppose that these 2^l IDs have a uniform distribution. Then the average probability of successful identification in one transmission is

$$\begin{aligned} p^{\text{suc}}_{\mathcal{C}_{\text{uncoded}(l)}} &= \sum_{\mathbf{x}\in\mathcal{C}_{\text{uncoded}(l)}} \frac{1}{2^l} \cdot (1-\epsilon_0')^{l-\mathrm{w}(\mathbf{x})}(1-\epsilon_1')^{\mathrm{w}(\mathbf{x})} \\ &= \frac{(1-\epsilon_0')^l}{2^l} \sum_{w=0}^{l} \binom{l}{w} \left(\frac{1-\epsilon_1'}{1-\epsilon_0'}\right)^w \\ &= \left(1 - \frac{\epsilon_0' + \epsilon_1'}{2}\right)^l. \end{aligned}$$

However, if there are errors occurred in the detection, then the tag ID will be wrongly identified (i.e., the errors are undetectable). Thus, we have the average probability of undetectable errors or the false identification rate in one transmission:

$$p^{\text{u.d.e.}}_{\mathcal{C}_{\text{uncoded}(l)}} = 1 - p^{\text{suc}}_{\mathcal{C}_{\text{uncoded}(l)}} = 1 - \left(1 - \frac{\epsilon_0' + \epsilon_1'}{2}\right)^l.$$

Clearly in the uncoded case, we have $p^{\text{suc}}_{\mathcal{C}_{\text{uncoded}(l)}} + p^{\text{u.d.e.}}_{\mathcal{C}_{\text{uncoded}(l)}} = 1$. And the false identification is unavoidable, since whenever there are errors occurred (either in the form of $0 \to 1$ or $1 \to 0$) in the detection, the tag ID will be falsely identified.

V. ERROR DETECTING CODES

The purpose of using the error detecting codes (over the uncoded case) is to detect a transmission error so that a retransmission can be automatically triggered. Suppose that a code $\mathcal{C}$ with weight distribution $\{w_i | 0 \le i \le l\}$ is used to encode the tag IDs. Thus, each tag ID is a codeword from $\mathcal{C}$. Clearly, $\mathcal{C}$ can be used to encode $\sum_{i=0}^{l} w_i$ tag IDs. Therefore, the rate is

$$R_{\mathcal{C}} = \frac{\log_2 \sum_{i=0}^{l} w_i}{l}. \tag{9}$$

For each tag $\mathbf{c} \in \mathcal{C}$, in general, we can define the following events for the transmission over BAC$(\epsilon_0', \epsilon_1')$:

- $\mathcal{E}_0$: the tag ID $\mathbf{c}$ is correctly transmitted. Thus we have
$$p^{\text{suc}}_{\mathbf{c}} = \Pr\{\mathcal{E}_0\} = (1-\epsilon_0')^{l-\mathrm{w}(\mathbf{c})}(1-\epsilon_1')^{\mathrm{w}(\mathbf{c})}. \tag{10}$$
where $\mathrm{w}(\mathbf{c})$ denotes the Hamming weight of $\mathbf{c} \in \mathcal{C}$.
- $\mathcal{E}_{\text{u.d.e.}}$: there are errors occurring but resulting into another codeword from $\mathcal{C}$ (therefore the errors will be undetectable). We denote
$$p^{\text{u.d.e.}}_{\mathbf{c}} = \Pr\{\mathcal{E}_{\text{u.d.e.}}\}. \tag{11}$$
- $\mathcal{E}_{\text{d.e.}}$: there are errors occurring and resulting into a sequence that is not in $\mathcal{C}$ (therefore the errors are detectable).
$$p^{\text{d.e.}}_{\mathbf{c}} = \Pr\{\mathcal{E}_{\text{d.e.}}\} = 1 - p^{\text{u.d.e.}}_{\mathbf{c}} - p^{\text{suc}}_{\mathbf{c}}. \tag{12}$$

Then, the probability that the first $\kappa - 1$ reads fail (i.e., error detected) while the κ-th read succeeds is $(p^{\text{d.e.}}_{\mathbf{c}})^{\kappa-1}(1-p^{\text{d.e.}}_{\mathbf{c}})$. As a result, we have *the expected number of retransmissions for an identification of tag ID* $\mathbf{c}$ as

$$\begin{aligned} T^{\text{id,t}}_{\mathbf{c}} &= \sum_{\kappa=1}^{\infty} \kappa \cdot (p^{\text{d.e.}}_{\mathbf{c}})^{\kappa-1}(1-p^{\text{d.e.}}_{\mathbf{c}}) \\ &= \frac{1}{1-p^{\text{d.e.}}_{\mathbf{c}}} = \frac{1}{p^{\text{u.d.e.}}_{\mathbf{c}} + p^{\text{suc}}_{\mathbf{c}}}. \end{aligned} \tag{13}$$

Since each transmission involves of $2\kappa_0 + 1$ rounds of interrogations, *the expected number of interrogations for a successful identification of tag ID* $\mathbf{c}$ is

$$T^{\text{id,i}}_{\mathbf{c}} = (2\kappa_0 + 1) \cdot T^{\text{id,t}}_{\mathbf{c}}. \tag{14}$$

Assume that the tag IDs have a uniform distribution. Then *the average probability of successful tag-ID identification* and *the average probability of undetectable errors* are

$$\begin{aligned} p^{\text{suc}}_{\mathcal{C}} &= \frac{1}{|\mathcal{C}|} \sum_{\mathbf{c}\in\mathcal{C}} p^{\text{suc}}_{\mathbf{c}} = \frac{1}{|\mathcal{C}|} \sum_{i=0}^{l} w_i (1-\epsilon_0')^{l-i}(1-\epsilon_1')^i \\ &= \mathrm{W}_{\mathcal{C}}\left(1-\epsilon_0', 1-\epsilon_1'\right); \end{aligned} \tag{15}$$

$$p_{\mathcal{C}}^{\text{u.d.e}} = \frac{1}{|\mathcal{C}|}\sum_{\mathbf{c}\in\mathcal{C}} p_{\mathbf{c}}^{\text{u.d.e}}, \tag{16}$$

where $\mathrm{W}_{\mathcal{C}}(x, y)$ is the average (homogeneous) weight enumerator of the code, and $\mathrm{W}_{\mathcal{C}}(1-\epsilon_0', 1-\epsilon_1')$ is its value at $x = 1-\epsilon_0'$ and $y = 1-\epsilon_1'$.

The average expected numbers of retransmissions and interrogations (noting that each transmission involves $2\kappa_0+1$ rounds of interrogations) are

$$T_{\mathcal{C}}^{\text{id,t}} = \frac{1}{|\mathcal{C}|}\sum_{\mathbf{c}\in\mathcal{C}} T_{\mathbf{c}}^{\text{id,t}} = \frac{1}{|\mathcal{C}|}\sum_{\mathbf{c}\in\mathcal{C}} \frac{1}{p_{\mathbf{c}}^{\text{u.d.e.}} + p_{\mathbf{c}}^{\text{suc}}}; \tag{17}$$

$$T_{\mathcal{C}}^{\text{id,i}} = (2\kappa_0+1)\cdot T_{\mathcal{C}}^{\text{id,t}}. \tag{18}$$

A. CONSTANT WEIGHT CODES

A simple idea to encode the chipless tag ID is to use a constant-weight code $\mathcal{C}_{\text{CW}(l,w)}$, where $w \le l/2$. That is, each tag ID has a constant Hamming weight w. Such a code can be used to encode $\binom{l}{w}$ tag IDs. Therefore, the rate is

$$R_{\mathcal{C}_{\text{CW}(l,w)}} = \frac{\log\binom{l}{w}}{l}. \tag{19}$$

Note that for a fixed l, $R_{\mathcal{C}_{\text{CW}(l,w)}}$ is maximized at $w = \lfloor \frac{l}{2} \rfloor$. In particular, we have $\lim_{l\to\infty} R_{\mathcal{C}_{\text{CW}(l,w)}} = h(\frac{w}{l})$. Noticing the fact that $h(\frac{1}{2}) = 1$, we can see that, with the choice of $w = \lfloor \frac{l}{2} \rfloor$, $R_{\mathcal{C}_{\text{CW}(l,w)}}$ approaches to 1 as l increases to infinity.

For each codeword $\mathbf{c} \in \mathcal{C}_{\text{CW}(l,w)}$, we have $\mathrm{w}(\mathbf{c}) = w$ and thus

$$p_{\mathbf{c}}^{\text{suc}} = (1-\epsilon_0')^{l-w}(1-\epsilon_1')^{w}. \tag{20}$$

Moreover, the undetectable errors for any codeword $\mathbf{c} \in \mathcal{C}_{\text{CW}(l,w)}$ are those occurring in both forms of $0 \to 1$ and $1 \to 0$ but the weight does not change. That is,

$$p_{\mathbf{c}}^{\text{u.d.e.}} = \sum_{i=1}^{\min\{w,l-w\}} \binom{w}{i}\binom{l-w}{i}(\epsilon_0'\cdot\epsilon_1')^{i}(1-\epsilon_0')^{l-w-i}(1-\epsilon_1')^{w-i}. \tag{21}$$

As a result, we have the expected number of retransmissions for an identification of $\mathbf{c}$ as

$$T_{\mathbf{c}}^{\text{id,t}} = \frac{1}{p_{\mathbf{c}}^{\text{u.d.e.}} + p_{\mathbf{c}}^{\text{suc}}}. \tag{22}$$

Note that for $\mathcal{C}_{\text{CW}(l,w)}$, both $p_{\mathbf{c}}^{\text{suc}}$ and $p_{\mathbf{c}}^{\text{u.d.e.}}$ depend only on the Hamming weight w, and so does $T_{\mathbf{c}}^{\text{id,t}}$. Since w is the same for all the codewords in $\mathcal{C}_{\text{CW}(l,w)}$, we have

$$p_{\mathcal{C}_{\text{CW}(l,w)}}^{\text{suc}} = (1-\epsilon_0')^{l-w}(1-\epsilon_1')^{w};$$

$$p_{\mathcal{C}_{\text{CW}(l,w)}}^{\text{u.d.e}} = p_{\mathbf{c}}^{\text{u.d.e.}} \text{ as defined in (21)};$$

$T_{\mathcal{C}_{\text{CW}(l,w)}}^{\text{id,t}}$ is given in (*), as shown at the bottom of this page;

$$T_{\mathcal{C}_{\text{CW}(l,w)}}^{\text{id,i}} = (2\kappa_0+1)\cdot T_{\mathcal{C}_{\text{CW}(l,w)}}^{\text{id,t}}.$$

Especially over $\mathrm{ZC}(0,\epsilon_1')$, only errors in the form of $1 \to 0$ are possible. Then, all errors will be detected if $\mathcal{C}_{\text{CW}(l,w)}$ is used (since if there are some ones are read as zeros, it will result in a sequence with a decreased weight), and a retransmission will be issued automatically. In this case, we have $\epsilon_0' = 0$ and $p_{\mathbf{c}}^{\text{u.d.e.}} = 0$ for each $\mathbf{c} \in \mathcal{C}_{\text{CW}(l,w)}$. The average probability of successful transmission, the average probability of undetectable errors and the expected number of transmissions/interrogations needed to make a successful identification become

$$p_{\mathcal{C}_{\text{CW}(l,w)}}^{\text{suc}} = (1-\epsilon_1')^{w};$$

$$p_{\mathcal{C}_{\text{CW}(l,w)}}^{\text{u.d.e}} = 0;$$

$$T_{\mathcal{C}_{\text{CW}(l,w)}}^{\text{id,t}} = \frac{1}{(1-\epsilon_1')^{w}};$$

$$T_{\mathcal{C}_{\text{CW}(l,w)}}^{\text{id,i}} = \frac{2\kappa_0+1}{(1-\epsilon_1')^{w}}.$$

That is, for the $\mathrm{ZC}(0,\epsilon_1')$, the constant-weight code $\mathcal{C}_{\text{CW}(l,w)}$ permits perfect error detection. The major disadvantage is that it is nonseparable (or nonsystematic), where by separability we mean that the bits containing the information to be transmitted and the bits provided for error detection are distinct. For $\mathcal{C}_{\text{CW}(l,w)}$, it is impossible to separate the information bits from the redundant bits.

B. BERGER-FREIMAN CODES

Differently from the constant-weight code $\mathcal{C}_{\text{CW}(l,w)}$, the Berger-Freiman code [20], [21] is a systematic code (i.e., the codeword contains information bits and check bits, with the information bits and the check bits being separated) that permits perfect error detection over a $\mathrm{ZC}(0,\epsilon_1')$ channel.

The Berger-Freiman code $\mathcal{C}_{\text{BF}(l,k)}$ can be formed as follows. Given k information bits, count the number of the zeros in the k information bits and take its binary representation as the check bits. So the number of the check bits is $r = \lceil \log_2(k+1) \rceil$ and the length of the codeword is $l = k + \lceil \log_2(k+1) \rceil$, where $\lceil a \rceil$ stands for the smallest integer that is larger or equal to a. Clearly, we have

$$R_{\mathcal{C}_{\text{BF}(l,k)}} = \frac{k}{l} = \frac{k}{k + \lceil \log_2(k+1) \rceil}. \tag{23}$$

Differently from the constant weight codes, the weight distribution of the Berger-Freiman code $\{w_i | 0 \le i \le l\}$ is not obvious. Nevertheless, for each codeword $\mathbf{c} \in \mathcal{C}_{\text{BF}(l,k)}$,

$$T_{\mathcal{C}_{\text{CW}(l,w)}}^{\text{id,t}} = \frac{1}{(1-\epsilon_0')^{l-w}(1-\epsilon_1')^{w} + \sum_{i=1}^{\min\{w,l-w\}} \binom{w}{i}\binom{l-w}{i}(\epsilon_0'\cdot\epsilon_1')^{i}(1-\epsilon_0')^{l-w-i}(1-\epsilon_1')^{w-i}}. \tag{*}$$

its Hamming weight $\mathrm{w}(\mathbf{c})$ can be calculated as follows. Suppose that there are k_0 zeros in the k information bits (i.e., $(k-k_0)$ ones). Note that in total there are $\binom{k}{k_0}$ such codewords. Then,

- if $k_0 = 0$, there are 0 ones in the check bits and the total Hamming weight is $\mathrm{w}(\mathbf{c}) = k$;
- if $k_0 \neq 0$, suppose $k_0 = \sum_{j=1}^{s} 2^{i_j}$, where $0 \le i_1 < i_2 < \ldots < i_s \le r-1$. Then there are s ones in the check bits, and the total Hamming weight $\mathrm{w}(\mathbf{c}) = k - k_0 + s$, where $1 \le s \le r$. (Let $\mathcal{T} = \{0, 1, \ldots, r-1\}$. Denote $\mathcal{S} = \{i_1, i_2, \ldots, i_s\} \subseteq \mathcal{T}$ with $|\mathcal{S}| = s$, where $|\mathcal{S}|$ is the cardinality of the set $\mathcal{S}$. Then $k_0 = \sum_{i\in\mathcal{S}} 2^i$, and the case of $k_0 = 0$ becomes the special case $\mathcal{S} = \emptyset$ if we use $\emptyset$ to denote the empty set and define $\sum_{i\in\emptyset} 2^i = 0$ and $|\emptyset| = 0$. In particular, corresponding to each choice of $\mathcal{S} \subseteq \mathcal{T}$ with $k_0 = \sum_{i\in\mathcal{S}} 2^i \le k$, there are $\binom{k}{k_0}$ codewords in $\mathcal{C}_{\mathrm{BF}(l,k)}$.)

Recall that for each codeword $\mathbf{c}$ of Hamming weight $\mathrm{w}(\mathbf{c})$, the probability of a successful transmission is $(1-\epsilon_0')^{l-\mathrm{w}(\mathbf{c})}(1-\epsilon_1')^{\mathrm{w}(\mathbf{c})}$. Thus, we have

$$p^{\mathrm{suc}}_{\mathcal{C}_{\mathrm{BF}(l,k)}} = \frac{1}{2^k} \sum_{\mathbf{c}\in\mathcal{C}_{\mathrm{BF}(l,k)}} (1-\epsilon_0')^{l-\mathrm{w}(\mathbf{c})}(1-\epsilon_1')^{\mathrm{w}(\mathbf{c})}. \quad (24)$$

For simplicity, we consider the transmission over a $\mathrm{ZC}(0, \epsilon_1')$ channel (i.e., $\epsilon_0' = 0$). In a $\mathrm{ZC}(0, \epsilon_1')$ channel, since only errors in the form of $1 \to 0$ are possible, so the number of zeros in k information bits can be only increased; while the value of the binary representation in the check bits can be only decreased. So any errors (appeared in either information bits or check bits, or both) will lead to an inequality between the numbers of zeros counted in the k information bits and represented by the check bits, respectively. As a consequence, we have $p^{\mathrm{u.d.e.}}_{\mathbf{c}} = 0$ for each $\mathbf{c} \in \mathcal{C}_{\mathrm{BF}(n,k)}$, i.e., all the errors will be detected. Thus, we have for a codeword $\mathbf{c} \in \mathcal{C}_{\mathrm{BF}(l,k)}$ over a $\mathrm{ZC}(0, \epsilon_1')$:

$$\begin{aligned} p^{\mathrm{suc}}_{\mathbf{c}} &= (1-\epsilon_1')^{\mathrm{w}(\mathbf{c})}; \\ p^{\mathrm{u.d.e}}_{\mathbf{c}} &= 0; \\ T^{\mathrm{id,t}}_{\mathbf{c}} &= \frac{1}{p^{\mathrm{u.d.e.}}_{\mathbf{c}} + p^{\mathrm{suc}}_{\mathbf{c}}} = \frac{1}{(1-\epsilon_1')^{\mathrm{w}(\mathbf{c})}}; \\ T^{\mathrm{id,i}}_{\mathbf{c}} &= \frac{2\kappa_0+1}{(1-\epsilon_1')^{\mathrm{w}(\mathbf{c})}}. \end{aligned}$$

Suppose that all the tag IDs have a uniform distribution. We obtain

$$\begin{aligned} p^{\mathrm{suc}}_{\mathcal{C}_{\mathrm{BF}(l,k)}} &= \frac{1}{2^k} \sum_{\mathbf{c}\in\mathcal{C}} (1-\epsilon_1')^{\mathrm{w}(\mathbf{c})} \\ &= \frac{1}{2^k} \sum_{\substack{\mathcal{S}\subseteq\mathcal{T} \\ k_0=\sum_{i\in\mathcal{S}} 2^i \le k}} \binom{k}{k_0} (1-\epsilon_1')^{k-k_0+|\mathcal{S}|}; \\ p^{\mathrm{u.d.e}}_{\mathcal{C}_{\mathrm{BF}(l,k)}} &= 0; \\ T^{\mathrm{id,t}}_{\mathcal{C}_{\mathrm{BF}(l,k)}} &= \frac{1}{2^k} \sum_{\mathbf{c}\in\mathcal{C}} \frac{1}{(1-\epsilon_1')^{\mathrm{w}(\mathbf{c})}} \\ &= \frac{1}{2^k} \sum_{\substack{\mathcal{S}\subseteq\mathcal{T} \\ k_0=\sum_{i\in\mathcal{S}} 2^i \le k}} \binom{k}{k_0} \frac{1}{(1-\epsilon_1')^{k-k_0+|\mathcal{S}|}}; \\ T^{\mathrm{id,i}}_{\mathcal{C}_{\mathrm{BF}(l,k)}} &= \frac{2\kappa_0+1}{2^k} \sum_{\substack{\mathcal{S}\subseteq\mathcal{T} \\ k_0=\sum_{i\in\mathcal{S}} 2^i \le k}} \binom{k}{k_0} \frac{1}{(1-\epsilon_1')^{k-k_0+|\mathcal{S}|}}, \end{aligned}$$

where $\mathcal{T} = \{0, 1, \ldots, l-k-1\}$.

C. CRC CODES

It is worth mentioning that our analysis is not limited to the constant weight codes and Berger-Freiman codes, although they are more dedicated to the BACs (for instance, both permit perfect error detection over a $\mathrm{ZC}(0, \epsilon_1')$ channel).

Another interesting error detection code family, which we would like to add to our discussion, is the CRC codes [25]. CS codes are widely used in digital networks and storage devices because of their easy implementation and their burst-error detection capability, the properties of which are due to the structure of the codes as shortened cyclic codes. Note that CRC codes are not dedicated to the BACs. Nevertheless, we wonder if it can still serve as an appropriate choice in coding chipless RFID tags.

A CRC code $\mathcal{C}_{\mathrm{CRC}(l,k)}$ of length l with k information bits can be specified by a generator polynomial $g(x) \in \mathbb{F}_2[x]$ of degree $l-k$. Given k information bits $[m_0 \cdots m_{k-1}]$, where $m_j \in \mathbb{F}_2$ for $0 \le j \le k-1$, one first forms the information polynomial $M(x)$ as

$$M(x) = \sum_{j=0}^{k-1} m_j x^j.$$

To determine the $l-k$ CRC check bits, i.e., $[c_0 \cdots c_{l-k-1}]$, where $c_j \in \mathbb{F}_2$ for $0 \le j \le l-k-1$, one calculates the corresponding polynomial form $C(x) = \sum_{j=0}^{l-k-1} c_j x^j$ by the following polynomial division over $\mathbb{F}_2$:

$$C(x) \equiv x^{l-k} M(x) \mod g(x). \quad (25)$$

In this way, for k information bits $[m_0 \cdots m_{k-1}]$, one forms the codeword of length l as $[m_0 \cdots m_{k-1}\ c_0 \cdots c_{l-k-1}]$. Clearly a CRC code $\mathcal{C}_{\mathrm{CRC}(l,k)}$ is systematic, which is similar to the Berger-Freiman codes; and also linear (i.e., a linear combination of codewords is still a codeword), unlike the constant weight codes and the Berger-Freiman codes.

Unfortunately, there is no general formula for the weight distribution of CRC codes. Even for CRC codes with the same l, k parameters, the weight distributions can be totally different due to the different choices of the generator polynomials $g(x)$. Besides, the calculation of $p^{\mathrm{u.d.e}}_{\mathcal{C}_{\mathrm{CRC}(l,k)}}$ is even more challenging, especially when we consider a BAC scenario. Both are open problems to be further studied. A generic brute-force approach for these two problems has a computation complexity of $\mathcal{O}(2^k)$ and $\mathcal{O}(2^{2k})$, respectively.

For simplicity, we consider the transmission of a CRC code $\mathcal{C}_{\mathrm{CRC}(l,k)}$ over a $\mathrm{ZC}(0, \epsilon_1')$ channel, where only errors in the form of $1 \rightarrow 0$ are possible. In this scenario, both constant weight codes and the Berger-Freiman codes permit perfect error detection. Unfortunately, this does not apply to the CRC codes, which can be seen from the fact that every nonzero codeword could become the zero codeword with a small but positive probability. More specifically, an undetectable error to a codeword $\mathbf{c} \in \mathcal{C}_{\mathrm{CRC}(l,k)}$ occurs only if $\mathbf{c} \rightarrow \mathbf{c}'$, for some $\mathbf{c}' \in \mathcal{C}_{\mathrm{CRC}(l,k)}$ and $\mathbf{c}' \neq \mathbf{c}$. That is, some positions with bit 1 in $\mathbf{c}$ change into 0 in $\mathbf{c}'$. In this case, we say that $\mathbf{c}$ covers $\mathbf{c}'$ and denote $\mathbf{c}' \preccurlyeq \mathbf{c}$. Clearly, in this undetectable error event $\mathbf{c} \rightarrow \mathbf{c}'$, there are

- $\mathrm{w}(\mathbf{c} - \mathbf{c}')$ positions with error in the form of $1 \rightarrow 0$;
- 0 positions with error in the form of $0 \rightarrow 1$;
- $\mathrm{w}(\mathbf{c}')$ positions in the form of $1 \rightarrow 1$ (note that $\mathrm{w}(\mathbf{c}') + \mathrm{w}(\mathbf{c} - \mathbf{c}') = \mathrm{w}(\mathbf{c})$ since $\mathbf{c}' \preccurlyeq \mathbf{c}$);
- $l - \mathrm{w}(\mathbf{c})$ positions in the form of $0 \rightarrow 0$.

Therefore, for a codeword $\mathbf{c} \in \mathcal{C}_{\mathrm{CRC}(l,k)}$ transmitted over $\mathrm{ZC}(0, \epsilon_1')$, we have

$$p_{\mathbf{c}}^{\mathrm{u.d.e}} = \sum_{\mathbf{c}' \preccurlyeq \mathbf{c}, \mathbf{c}' \in \mathcal{C}} (\epsilon_1')^{\mathrm{w}(\mathbf{c}-\mathbf{c}')}(1 - \epsilon_1')^{\mathrm{w}(\mathbf{c}')}.$$

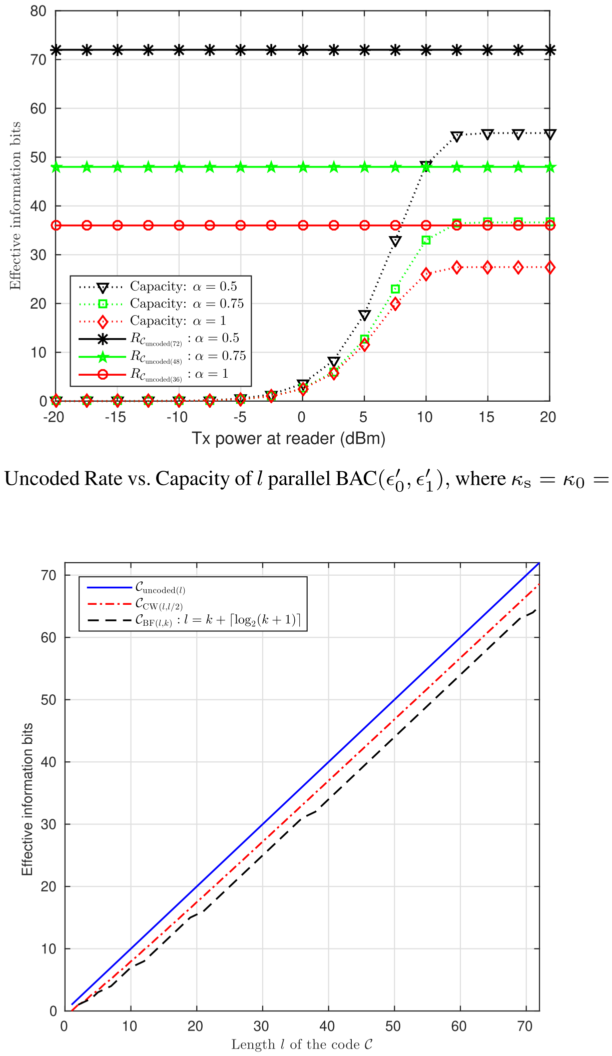
Recall that for $\mathbf{c} \in \mathcal{C}_{\mathrm{CRC}(l,k)}$ over $\mathrm{ZC}(0, \epsilon_1')$, we have $p_{\mathbf{c}}^{\mathrm{suc}} = (1 - \epsilon_1')^{\mathrm{w}(\mathbf{c})}$. If the weight distribution of the code $\mathcal{C}_{\mathrm{CRC}(l,k)}$ is known, then one can accordingly calculate $p_{\mathcal{C}_{\mathrm{CRC}(l,k)}}^{\mathrm{suc}}$ as defined in (15). Furthermore, $p_{\mathcal{C}_{\mathrm{CRC}(l,k)}}^{\mathrm{u.d.e}}$, $T_{\mathcal{C}_{\mathrm{CRC}(l,k)}}^{\mathrm{id,t}}$ and $T_{\mathcal{C}_{\mathrm{CRC}(l,k)}}^{\mathrm{id,i}}$ could be also calculated according to (16), (17) and (18), respectively.

VI. NUMERICAL RESULTS

Consider a deep notch defined by taking $B = 112$ MHz and $G_{\mathrm{peak}} = 8$ dB. At the same time, we consider two shallow notch alternatives by choosing $\alpha = 0.75$ and $\alpha = 0.5$, respectively (see the assumption made in Section II-B). The BACs defined by each notch choices are shown in Fig. 3. Clearly, for a fixed bandwidth $l_0 B$, the total bits (corresponding to the total bandwidth $l_0 B$) that can be used to encode the tag ID are $l = l_0 B/(\alpha B)$, which are l_0, $4l_0/3$ and $2l_0$, respectively for three different notch choices corresponding to $\alpha = 1$, $\alpha = 0.75$ and $\alpha = 0.5$. Here we take $l_0 = 36$. Then the code length l that can be used for the fixed bandwidth $l_0 * B$ with different notch choices is 36, 48 and 72, respectively.

A. CAPACITY OF L PARALLEL BACS(ϵ_0, ϵ_1)

Fig. 5(a) depicts the capacity of the l parallel BACs (corresponding to l notches/peaks defined by αB and αG_{peak}, with $\kappa_{\mathrm{s}} = 1$ and $\kappa_0 = 0$) versus the uncoded rates. As one can see, the uncoded rates for all possible α are beyond the capacities of the corresponding BACs. Recall that by definition, the capacity gives the fundamental limit of a reliable transmission with the probability of successful decoding approaching to 1. Clearly, for the uncoded case, false identification is unavoidable.

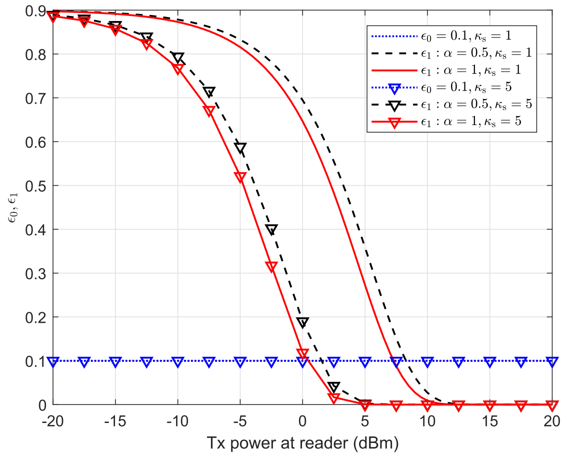


(a) Uncoded Rate vs. Capacity of l parallel $\mathrm{BAC}(\epsilon_0', \epsilon_1')$, where $\kappa_{\mathrm{s}} = \kappa_0 = 1$

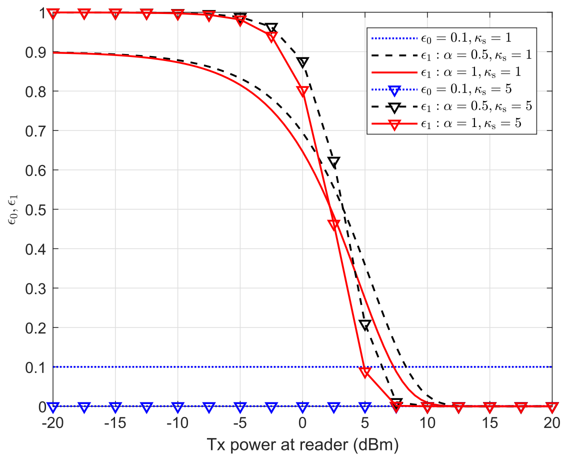
(b) Rate comparisons among $\mathcal{C}_{\mathrm{uncoded}}, \mathcal{C}_{\mathrm{CW}}$ and $\mathcal{C}_{\mathrm{BF}}$

FIGURE 5. Comparisons on effective information bits: Uncoded and coded cases.

To mitigate this problem, we propose several countermeasures. The first is to take the signal processing approach (this is characterized by the parameter κ_{s}) before the bit detection (for each interrogation); the second is to employ several rounds of interrogations and then apply the majority rule to make decision on each bit. Moreover, error detecting codes are also introduced to further reduce the false identification rate. As one can see from Fig. 5(b), if the two error-detecting codes, i.e., the constant-weight codes and Berger-Freiman codes, are used, the number of the effective information bits is only slightly less than that of the uncoded case; and the information rates of the two error-detecting codes approach to 1 when the code length becomes large.



(a) BACs for fixed $\epsilon_0 = 0.1$



(b) BACs for fixed θ_{th}

FIGURE 6. **BACs corresponding to different notch choices with varying κ_s and $\kappa_0 = 0$.**

B. EFFECT OF THE SIGNAL PROCESSING APPROACH

In this paper, a signal processing approach, characterized by the parameter κ_s, is suggested to be used in the bit detection in each interrogation. The idea is to enhance the SNR of the underlying channel and thus to improve the overall channel quality. The effect of taking such an approach is illustrated in Fig. 6. In particular, in Fig. 6, we consider the following choices at the detector:

- *The threshold θ_{th} is tunable*. In this case, for an easy comparison, we consider the case that θ_{th} is tuned such that the same ϵ_0 is maintained. That is, only ϵ_1 will vary corresponding to different signal processing parameter κ_s. The obtained BAC(ϵ_0, ϵ_1) is shown in Fig. 6 (a).
- *The threshold θ_{th} is not tunable*. In this case, both ϵ_0, ϵ_1 will vary corresponding to different signal processing parameter κ_s. The obtained BAC(ϵ_0, ϵ_1) is shown in Fig. 6 (b).

As one can see from Fig. 6 (a), if θ_{th} is tunable, then for $\kappa_s = 5$, a clear improvement on the underlying channel quality can be observed by the decreasing ϵ_1, which is the false detection rate of $1 \to 0$, whilst maintaining the same $\epsilon_0 = 0.1$ which is the false detection rate of $0 \to 1$. That is, ϵ_1 takes all the benefits from taking the signal processing approach.

However, if θ_{th} is not tunable, then as one can see from Fig. 6 (b), as κ_s increases, a smaller ϵ_0 (than the initial value 0.1) will be obtained; while a larger ϵ_1 will be obtained at the low transmit power regime and a smaller ϵ_1 will be obtained at the high transmit power regime. In other words, ϵ_0 will benefit from taking the signal processing approach; however, whether ϵ_1 will benefit or not, depends on the transmit power level (ϵ_1 will benefit only at the high transmit power regime).

In Fig. 8, we provide the corresponding capacities of the obtained BACs. As one can see from Fig. 8, it is advantageous to tune θ_{th} (over the fixed θ_{th} choice) at the low transmit power regime, since it leads to a BAC with a larger capacity. However, at the high transmit power region, it is actually superior to use the fixed threshold θ_{th} (such that ϵ_0 can benefit from the signal processing approach as well). In general, the signal processing approach together with an optimal choice of θ_{th} will improve the underlying channel quality from the perspective of channel capacity.

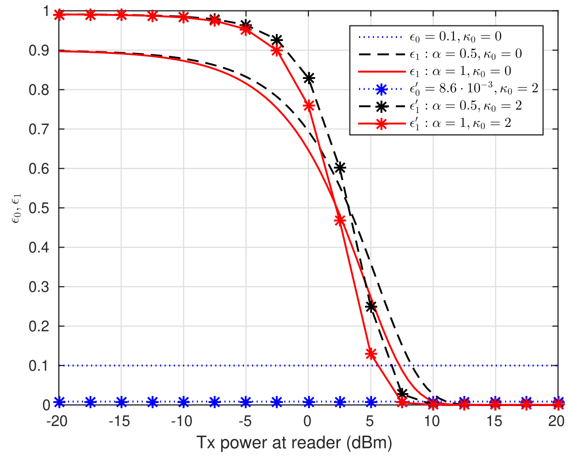
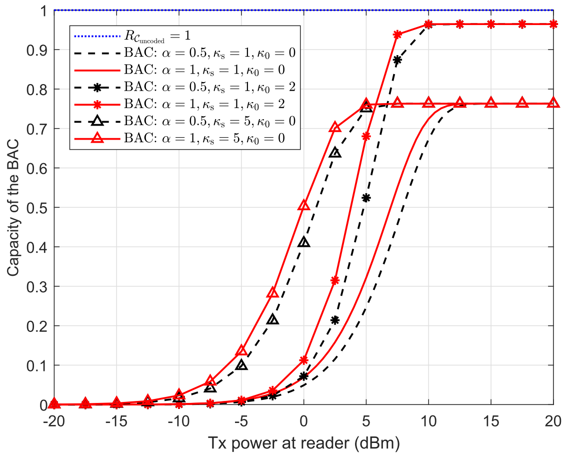


FIGURE 7. **BACs corresponding to different notch choices with $\kappa_s = 1$ and varying κ_0.**

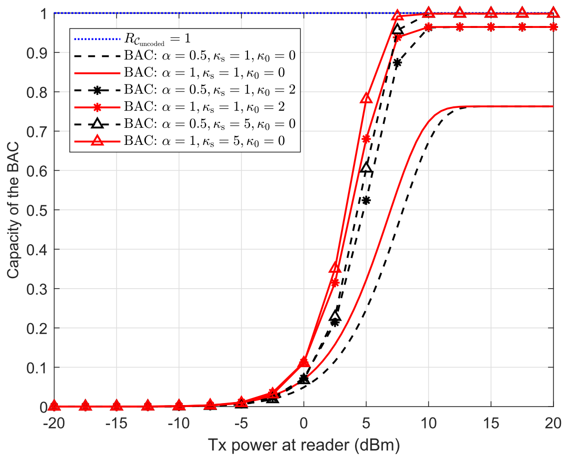
C. EFFECT OF MULTIPLE INTERROGATIONS & MAJORITY RULE BASED DETECTION

In Fig. 7, we illustrate the effect of applying $2\kappa_0+1$ rounds of interrogations and then using the majority rule to further help to identify each bit. The overall BAC (for each bit after $2\kappa_0+1$ rounds of interrogations) is a BAC(ϵ_0', ϵ_1') where (ϵ_0', ϵ_1') are as defined in (6) and (7), respectively. Here, in order to better understand the effect of the majority rule based detection, we take $\kappa_s = 1$.

An interesting observation from Fig. 7 is that majority rule based detection has a polarization effect. In particular, we obtain a smaller ϵ_1' for $\epsilon_1 < 0.5$; but a larger ϵ_1' for $\epsilon_1 > 0.5$. This polarization effect is due to the nature of the majority vote. Thus, such a strategy is recommended only for



(a) Tunable θ_{th} for varying κ_s (and fixed $\epsilon_0 = 0.1$)



(b) Fixed θ_{th} for varying κ_s

FIGURE 8. **Capacity of BACs corresponding to different notch choices, κ_s and κ_0.**

those BAC(ϵ_0, ϵ_1) with both $\epsilon_0, \epsilon_1 < 0.5$. This observation also implies the advantage of the signal processing approach especially at the low transmit power regime (e.g.: with tunable θ_{th}) in improving the underlying channel quality. In the following, we will focus on the BACs with $\epsilon_0, \epsilon_1 < 0.5$, by assuming that this could be obtained by taking appropriate choices of κ_s and θ_{th}.

1) COMPARISON TO THE SIGNAL PROCESSING APPROACH

Recall that the signal processing approach is to detect the bits based on a kind of soft decision; while the majority rule based detection is a kind of hard decision. Intuitively, for a fixed value of $\kappa_s \cdot (2\kappa_0 + 1)$ (i.e., the total number of rounds of sending the same signal), taking the former approach only will perform better than taking the later approach only. This can be confirmed in Fig. 8 from the capacities of the respectively obtained BACs. Note that for an easy and fair comparison, we take κ_s and κ_0 such that $\kappa_s \cdot (2\kappa_0 + 1) = 5$. Moreover, we observe that

- in Fig. 8(a), at the low transmit power, e.g., when P_{rd} is less than about 5 dBm, the signal processing approach (with *tunable* θ_{th}) is superior to the multiple interrogations with majority rule based detection, due to the larger capacity of the former BAC;
- in Fig. 8(b), at the high transmit power, e.g., when P_{rd} is larger than about 0 dBm, the signal processing approach (with *fixed* θ_{th}) is superior to the multiple interrogations with majority rule based detection, due to the larger capacity for its BAC.

In other words, the signal processing approach with an optimal θ_{th} performs always better than the multiple interrogations with majority rule based detection (from the capacity perspective).

However, the signal processing approach assumes that the channel state is time-invariant which is usually unrealistic in practice. Furthermore, without the perfect knowledge of the channel state information, it is difficult to find the optimal threshold θ_{th} in order to yield the best possible performance. Therefore, it is easier to implement the signal processing approach with a *fixed* θ_{th} than with an optimally *tuned* θ_{th}. Interestingly, we notice that the signal processing approach with a *fixed* θ_{th} and the approach of multiple interrogations with majority rule based detection have a very similar effect on the overall BAC, as one can see from Fig. 6(b), Fig. 7, and Fig. 8(b). Therefore, for simplicity, in the sequel, we simply take $\kappa_s = 1$ and vary κ_0 to further look into the effect of the multiple interrogations with majority rule based detection. Similar performance can be expected by employing the signal processing approach with a *fixed* θ_{th}; and better performance can be expected if combing both strategies in an optimal manner.

2) EFFECT ON THE PROBABILITY OF SUCCESSFUL IDENTIFICATION

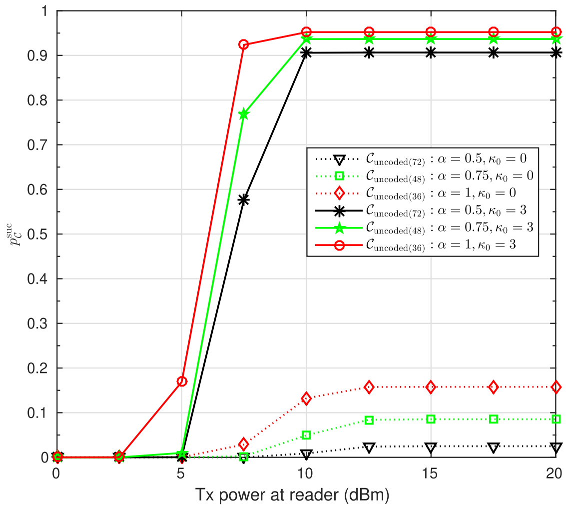
In Fig. 9, we plot the probability of successful identification (i.e., $p_{\mathcal{C}}^{\text{suc}}$) and the probability of undetectable errors (i.e., $p_{\mathcal{C}}^{\text{u.d.e}}$) for the uncoded tag IDs after $2\kappa_0 + 1$ rounds of interrogations. Taking $\kappa_0 = 0$ and $\kappa_0 = 3$, respectively, we see a significant improvement on the performance (i.e., significantly increased $p_{\mathcal{C}}^{\text{suc}}$ and decreased $p_{\mathcal{C}}^{\text{u.d.e}}$). Note that for the uncoded case, $p_{\mathcal{C}}^{\text{suc}} + p_{\mathcal{C}}^{\text{u.d.e}} = 1$ and $p_{\mathcal{C}}^{\text{u.d.e}}$ indicates the average probability of false identification.

Moreover, we note that if taking $\kappa_s = 1$ and $\kappa_0 = 3$, then we have $\epsilon_0' = 2.7 \cdot 10^{-3}$ and $\epsilon_1' \gg \epsilon_0'$ in the transmit power regime 0–7 dBm, where the BAC(ϵ_0', ϵ_1') could be approximated by a ZC($0, \epsilon_1'$).

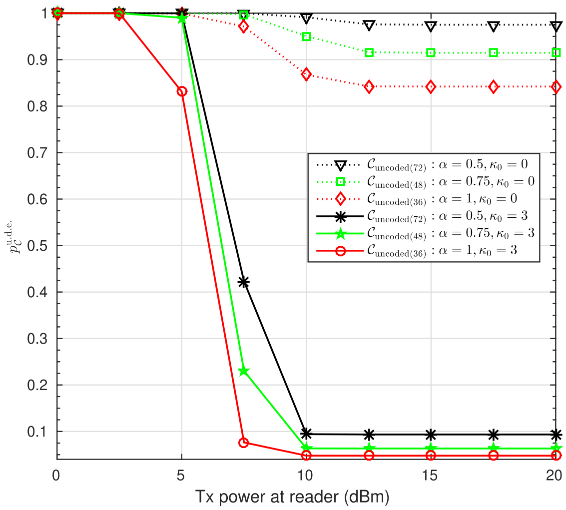
D. EFFECT OF CONSTANT WEIGHT CODES AND BERGER-FREIMAN CODES

1) ON THE PROBABILITY OF SUCCESSFUL TRANSMISSION

In Fig. 10, the (average) probability of successful transmission is plotted for the uncoded case, the constant weight codes $\mathcal{C}_{\text{CW}(l,\lfloor l/2 \rfloor)}$ and the Berger-Freiman codes $\mathcal{C}_{\text{BF}(l,k)}$. As one can observe from Fig. 10, the employment of $\mathcal{C}_{\text{CW}}(l, \lfloor l/2 \rfloor)$ and $\mathcal{C}_{\text{BF}(l,k)}$ does not bring much gain in the (average) probability of successful transmission (although $\mathcal{C}_{\text{BF}(l,k)}$ provides a probability of successful transmission very close to the one



(a) $p^{\mathrm{suc}}_{\mathcal{C}_{\mathrm{uncoded}(l)}}$ w.r.t. P_{rd}



(b) $p^{\mathrm{u.d.e.}}_{\mathcal{C}_{\mathrm{uncoded}(l)}}$ w.r.t. P_{rd}

FIGURE 9. $p^{\mathrm{suc}}_{\mathcal{C}_{\mathrm{uncoded}(l)}}$ **and** $p^{\mathrm{u.d.e.}}_{\mathcal{C}_{\mathrm{uncoded}(l)}}$ **after** $2\kappa_0 + 1$ **rounds of interrogations, where** $\kappa_s = 1$**.**

for the uncoded case). Nevertheless, the advantages of using the error detecting code are to decrease the probability of undetectable errors (i.e., to reduce the false identification rate of the tag ID) and to automatically issue a retransmission (until the tag ID is identified).

Especially, for a shallower notch defined by a smaller α, although more bits can be used to encode the tag IDs, this comes with a sacrifice on the performance of the successful identification rate in one transmission. Nevertheless, such a price diminishes as the transmission power increases and/or retransmissions are allowed (in case of a transmission error being detected).

2) ON THE PROBABILITY OF UNDETECTABLE ERRORS

In Fig. 11, the probability of undetectable errors is plotted for the uncoded case and the case where a constant-

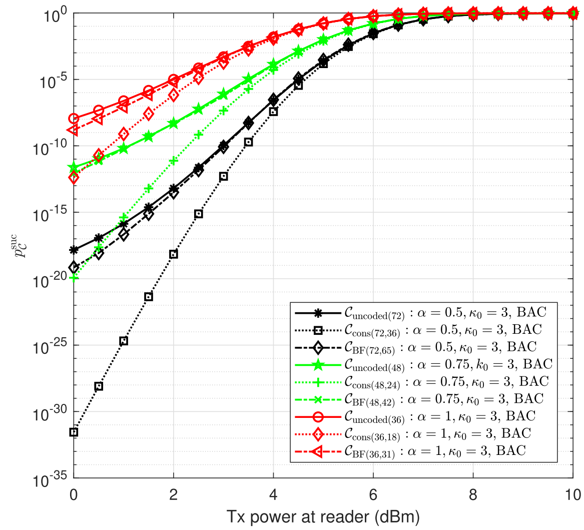


FIGURE 10. **Probability of successful transmission, where** $\kappa_s = 1$**.**

weight code $\mathcal{C}_{\mathrm{CW}(l,\lfloor l/2 \rfloor)}$ is employed. As one can observe from Fig. 11, the utilization of $\mathcal{C}_{\mathrm{CW}(l,\lfloor l/2 \rfloor)}$ can significantly decrease the probability of undetectable errors compared to the uncoded case by at least one order of magnitude (for 5 dB $\leq P_{\mathrm{rd}} \leq$ 20 dB).

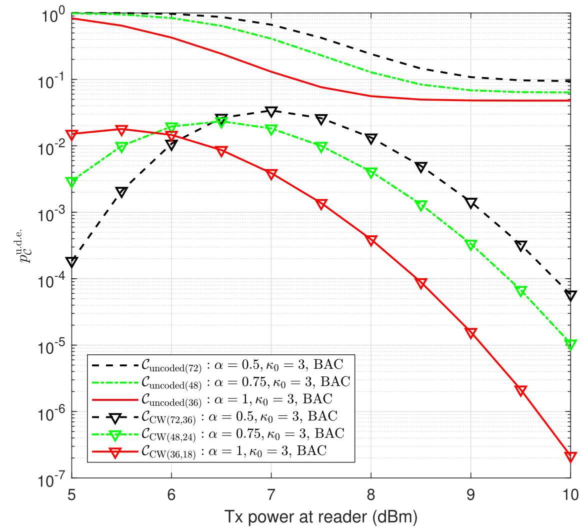
3) ON THE EXPECTED NUMBER OF RETRANSMISSIONS OR INTERROGATIONS

In Fig. 12 (a) and Fig. 13 (a), we plot the (average) expected numbers of retransmissions and interrogations for the constant weight codes $\mathcal{C}_{\mathrm{CW}(l,\lfloor l/2 \rfloor)}$. Both scenarios of $\mathrm{BAC}(\epsilon'_0, \epsilon'_1)$ and its approximation $\mathrm{ZC}(0, \epsilon'_1)$ are considered. As can be seen, the results for $\mathrm{BAC}(\epsilon'_0, \epsilon'_1)$ and the corresponding results for $\mathrm{ZC}(0, \epsilon'_1)$ are almost indistinguishable. This confirms that $\mathrm{ZC}(0, \epsilon'_1)$ can be a good approximation of $\mathrm{BAC}(\epsilon'_0, \epsilon'_1)$ as $\epsilon'_1 \gg \epsilon_0$.

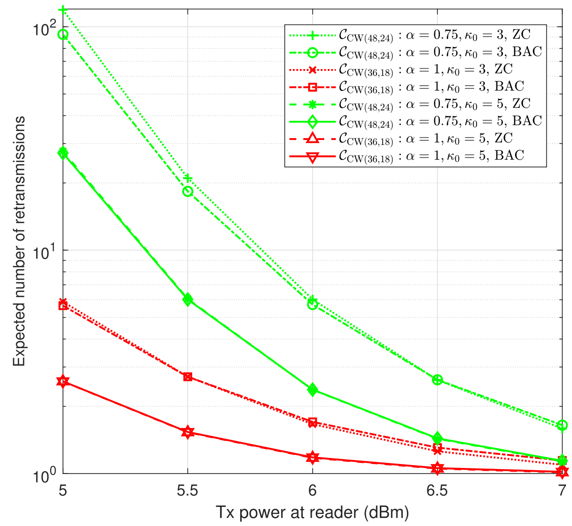
In Fig. 12 (b) and Fig. 13 (b), we use $\mathrm{ZC}(0, \epsilon'_1)$ to approximate $\mathrm{BAC}(\epsilon'_0, \epsilon'_1)$ and plot the (average) expected numbers of retransmissions and interrogations for the cases where either the constant code $\mathcal{C}_{\mathrm{CW}(l,\lfloor l/2 \rfloor)}$ or the Berger-Freiman codes $\mathcal{C}_{\mathrm{BF}(l,k)}$ are used.

Interestingly, as one can see from Fig. 12, the (average) expected number of retransmissions can be significantly decreased (for 5 dB $\leq P_{\mathrm{rd}} \leq$ 7 dB) as we increase κ_0 or increase α (i.e., decrease the code length). However, it is not the case for the (average) expected number of interrogations, as one can see from Fig. 13. The underlying reason is that, at the high transmit power regime, most probably no retransmission is needed and hence the expected number of interrogations approaches to $2\kappa_0 + 1$.

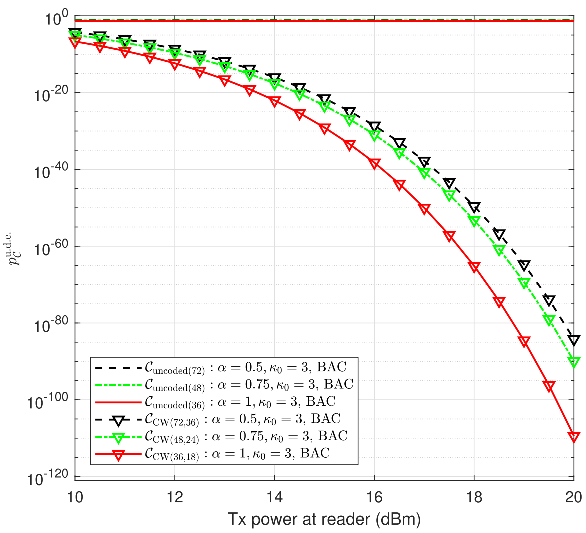
Moreover, we compare the performance on the expected numbers of retransmissions or interrogations if different error detecting codes are employed. As one observe from Fig. 12 (b) and Fig. 13 (b), at $\alpha = 0.75$ (i.e., $l = 48$) and $\kappa_0 = 3$, Berger-Freiman code shows its superiority to the constant code $\mathcal{C}_{\mathrm{CW}(l,\lfloor l/2 \rfloor)}$ due to the much less expected numbers of retransmissions and interrogations. However, their



(a) 5 dBm $\leq P_{\text{rd}} \leq$10 dBm

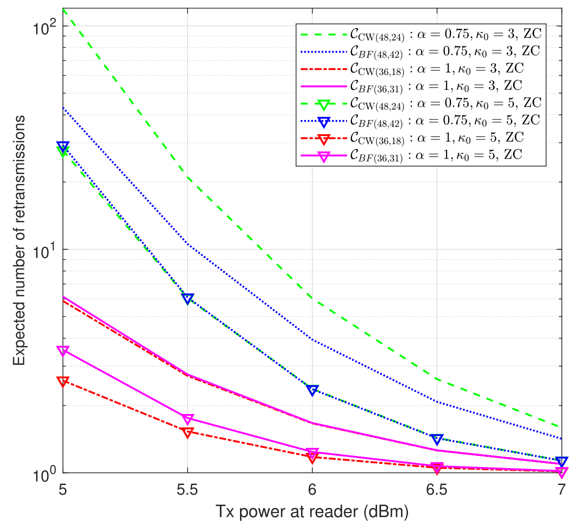


(a) $\mathcal{C}_{\text{CW}(l,\lfloor l/2 \rfloor)}$



(b) 10 dBm $\leq P_{\text{rd}} \leq$ 20 dBm

FIGURE 11. **Probability of undetectable errors: $\mathcal{C}_{\text{uncoded}(l)}$ vs. $\mathcal{C}_{\text{CW}(l,\lfloor l/2 \rfloor)}$, where $\kappa_s = 1$.**



(b) $\mathcal{C}_{\text{CW}(l,\lfloor l/2 \rfloor)}$ vs. $\mathcal{C}_{\text{BF}(l,k)}$

FIGURE 12. **Expected number of retransmissions for different κ_0, where $\kappa_s = 1$.**

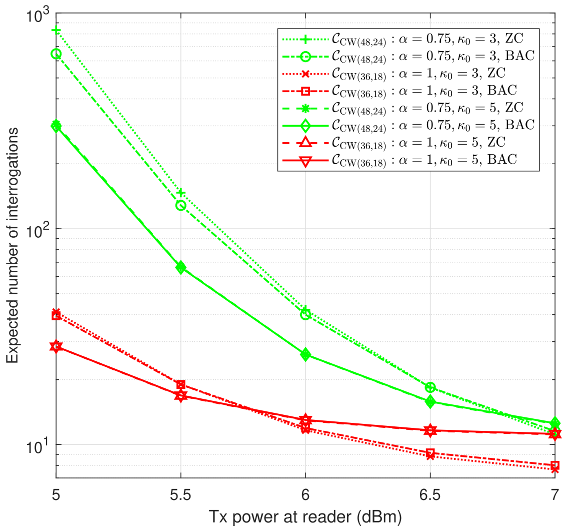
performance is quite close for $\alpha = 1$ (i.e., a shorter code length $l = 36$) and the scenario when more rounds of interrogations are taken (for instance $\kappa_0 = 5$).

E. EFFECT OF CRC CODES

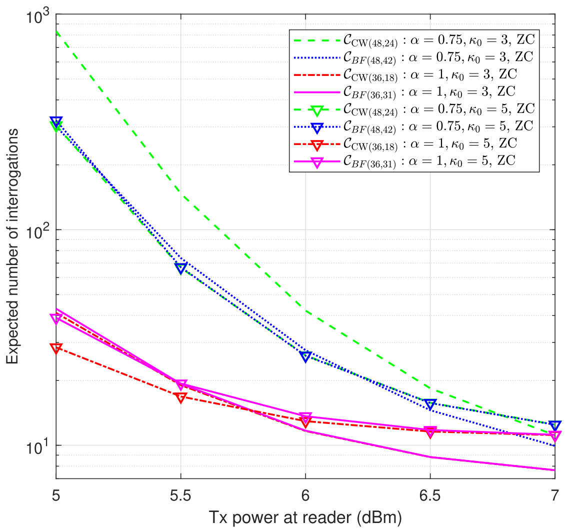
In this subsection, we bring CRC codes (as a representative of those codes that are not dedicated to the BAC but can be potentially competitive for error detection) into discussion. The goal is to shed light on their error detection performance (especially on $p_{\mathcal{C}}^{\text{suc}}$ and $p_{\mathcal{C}}^{\text{u.d.e}}$) over a BAC (which has not been extensively studied in the literature), by a comparison to the aforementioned constant codes and Berger-Freiman codes with the same code length (and the same number of check digits if possible).

Here we use codes of relatively short lengths (in comparison to those in the previous subsections), in order to obtain an accurate comparison within a reasonable computation complexity. The CRC codes with 5 check digits, for instance, $\mathcal{C}_{\text{CRC}(36,31)}, \mathcal{C}_{\text{CRC}(19,14)}$, and $\mathcal{C}_{\text{CRC}(17,12)}$, are constructed by the generator polynomial $g(x) = x^5 + x^3 + 1$; and the CRC codes with 4 check digits, for instance, $\mathcal{C}_{\text{CRC}(19,15)}, \mathcal{C}_{\text{CRC}(17,13)}, \mathcal{C}_{\text{CRC}(16,12)}, \mathcal{C}_{\text{CRC}(15,11)}$, are constructed by the generator polynomial $g(x) = x^4 + x + 1$. Note that both polynomials belong to the best general-purpose CRC polynomials according to [26].

In Fig. 14, we plot the average probability of successful identification (i.e., $p_{\mathcal{C}}^{\text{suc}}$) for the tag IDs coded with CRC codes $\mathcal{C}_{\text{CRC}(l,k)}$, constant codes $\mathcal{C}_{\text{CW}(l,\lfloor l/2 \rfloor)}$ and Berger-Freiman

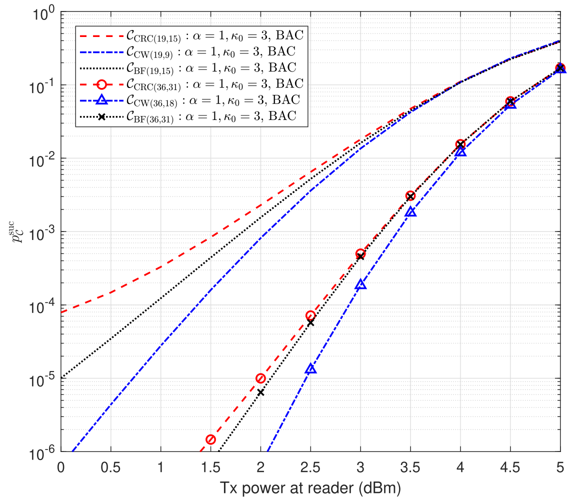


(a) $\mathcal{C}_{\mathrm{CW}(l,\lfloor l/2 \rfloor)}$

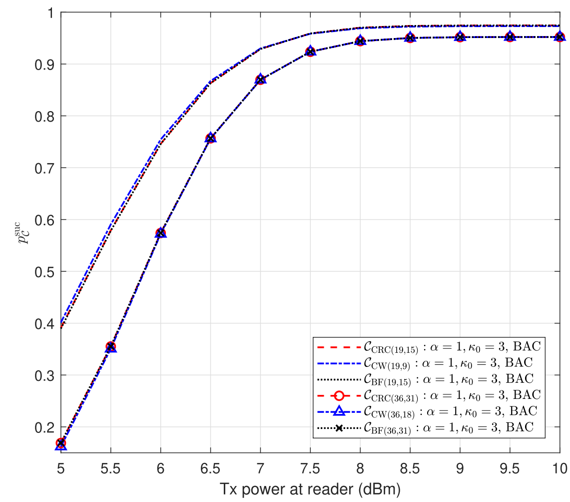


(b) $\mathcal{C}_{\mathrm{CW}(l,\lfloor l/2 \rfloor)}$ vs. $\mathcal{C}_{\mathrm{BF}(l,k)}$

FIGURE 13. **Expected number of interrogations for different κ_0, where $\kappa_S = 1$.**



(a) 0 dBm $\leq P_{\mathrm{rd}} \leq$5 dBm



(b) 5 dBm $\leq P_{\mathrm{rd}} \leq$10 dBm

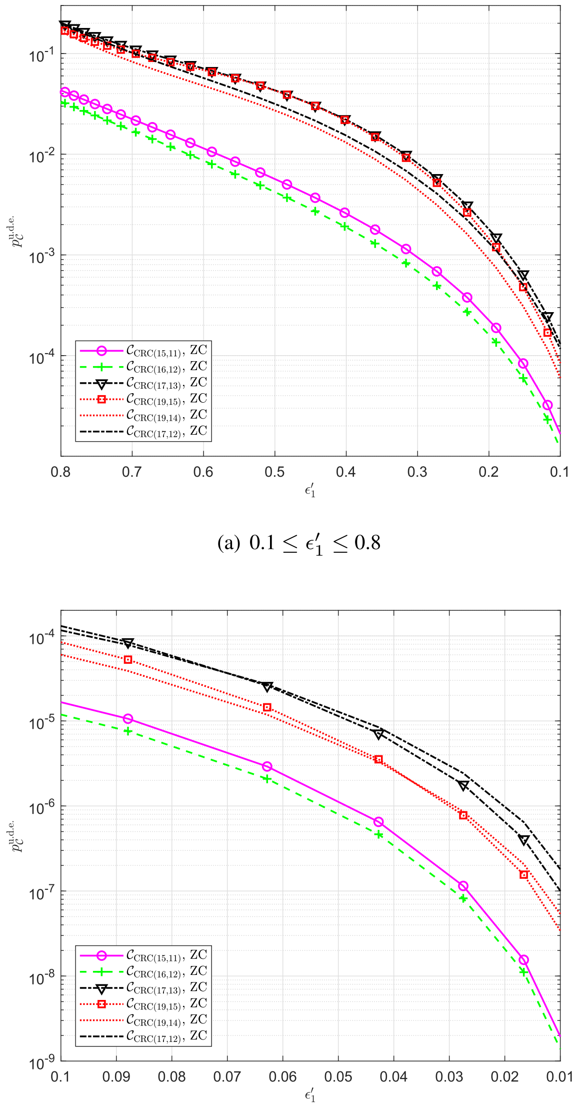
FIGURE 14. **Comparison on $p_{\mathcal{C}}^{\mathrm{suc}}$, where $\kappa_S = 1$.**

codes $\mathcal{C}_{\mathrm{BF}(l,k)}$, respectively. Under the same channel condition, we see that

- in the region of 0 dBm $\leq P_{\mathrm{rd}} \leq$ 5 dBm, the CRC code $\mathcal{C}_{\mathrm{CRC}(l,k)}$ has better performance on $p_{\mathcal{C}}^{\mathrm{suc}}$ than the corresponding constant weight code $\mathcal{C}_{\mathrm{CW}(l,\lfloor l/2 \rfloor)}$ and Berger-Freiman code $\mathcal{C}_{\mathrm{BF}(l,k)}$ with the same code length l at the low power region; however, its advantage diminishes as P_{rd} increases. In fact, for $p_{\mathcal{C}}^{\mathrm{suc}} > 0.1$, the constant code $\mathcal{C}_{\mathrm{CW}(l,\lfloor l/2 \rfloor)}$ tends to overtake in the short code length (e.g.: $l = 19$); whilst the Berger-Freiman code $\mathcal{C}_{\mathrm{BF}(l,k)}$ tends to have performance close to the CRC codes $\mathcal{C}_{\mathrm{CRC}(l,k)}$ (for both $l = 19, 36$).
- in the region of 5 dBm $\leq P_{\mathrm{rd}} \leq$ 10 dBm, the constant weight code $\mathcal{C}_{\mathrm{CW}(l,\lfloor l/2 \rfloor)}$ has slightly better performance on $p_{\mathcal{C}}^{\mathrm{suc}}$ among these three codes with a short code length (e.g.: $l = 19$); whilst CRC codes and Berger-Freiman codes have very similar performance on $p_{\mathcal{C}}^{\mathrm{suc}}$, regardless of the code length. In fact, the code performance on $p_{\mathcal{C}}^{\mathrm{suc}}$ converges as P_{rd} increases. In general, the coded tag IDs with more bits may suffer from some decrease on the $p_{\mathcal{C}}^{\mathrm{suc}}$.

So for a $p_{\mathcal{C}}^{\mathrm{suc}}$ in an acceptable range, the use of CRC code $\mathcal{C}_{\mathrm{CRC}(l,k)}$ does not bring any significant advantage on $p_{\mathcal{C}}^{\mathrm{suc}}$ over the constant codes $\mathcal{C}_{\mathrm{CW}(l,\lfloor l/2 \rfloor)}$ and the Berger-Freiman codes $\mathcal{C}_{\mathrm{BF}(l,k)}$.

In Fig. 15, we plot the average probability of undetectable errors (i.e., $p_{\mathcal{C}}^{\mathrm{u.d.e}}$) for the case where a CRC code $\mathcal{C}_{\mathrm{CRC}(l,k)}$ is employed. For simplicity we consider the code transmission over a ZC$(0, \epsilon_1')$. Remarkably, over a ZC$(0, \epsilon_1')$, both the constant codes $\mathcal{C}_{\mathrm{CW}(l,\lfloor l/2 \rfloor)}$ and the Berger-Freiman codes $\mathcal{C}_{\mathrm{BF}(l,k)}$ permit perfect error detection, i.e., $p_{\mathcal{C}_{\mathrm{CW}(l,\lfloor l/2 \rfloor)}}^{\mathrm{u.d.e}} = p_{\mathcal{C}_{\mathrm{BF}(l,k)}}^{\mathrm{u.d.e}} = 0$; whilst for $\mathcal{C}_{\mathrm{CRC}(l,k)}$, we have $p_{\mathcal{C}_{\mathrm{CRC}(l,k)}}^{\mathrm{u.d.e}} > 0$.



(a) $0.1 \leq \epsilon_1' \leq 0.8$

(b) $0.01 \leq \epsilon_1' \leq 0.1$

FIGURE 15. Comparison on $p_{\mathcal{C}}^{\text{suc}}$ over a ZC$(0, \epsilon_1')$.

To get more insight into the behavior of CRC codes over a ZC$(0, \epsilon_1')$, we consider CRC codes with varying lengths (i.e., l) and varying numbers of check digits (i.e., $l-k$). As one can observe from Fig. 15:

- if we fix the code length l, then the use of a CRC code $\mathcal{C}_{\text{CRC}(l,k)}$ with more check bits (i.e., $l-k$) does not necessarily lead to better performance, i.e., a decreasing on $p_{\mathcal{C}}^{\text{u.d.e}}$. For instance, we see that in Fig. 15 (a) as $\epsilon_1' > 0.1$, $p_{\mathcal{C}_{\text{CRC}(19,14)}}^{\text{u.d.e}} < p_{\mathcal{C}_{\text{CRC}(19,15)}}^{\text{u.d.e}}$ but a reverse behavior happens as ϵ_1' decreases as one can see in Fig. 15 (b). A similar observation can be made for the code pair $\mathcal{C}_{\text{CRC}(17,12)}$ and $\mathcal{C}_{\text{CRC}(17,13)}$;
- if we fix the number of check bits $l-k$, (i.e.,considering CRC codes $\mathcal{C}_{\text{CRC}(l,k)}$ generated by the same generator polynomial but with varying information bits k), then a CRC code with a smaller k may not necessarily lead to a decreasing on $p_{\mathcal{C}}^{\text{u.d.e}}$. For instance, $p_{\mathcal{C}_{\text{CRC}(15,11)}}^{\text{u.d.e}} < p_{\mathcal{C}_{\text{CRC}(17,13)}}^{\text{u.d.e}}$ but $p_{\mathcal{C}_{\text{CRC}(15,11)}}^{\text{u.d.e}} > p_{\mathcal{C}_{\text{CRC}(16,12)}}^{\text{u.d.e}}$, in all the plotted region of $0.01 \leq \epsilon_1' \leq 0.8$ in Fig. 15.

So the use of constant codes $\mathcal{C}_{\text{CW}(l,\lfloor l/2 \rfloor)}$ and Berger-Freiman codes $\mathcal{C}_{\text{BF}(l,k)}$ have a clear advantage over CRC codes $\mathcal{C}_{\text{CRC}(l,k)}$ on $p_{\mathcal{C}}^{\text{u.d.e}}$, especially over a ZC$(0, \epsilon_1')$.

VII. CONCLUDING REMARKS

In this paper, we have studied the problem of RFID tag identification from an information-theoretic perspective. Focusing on the so-called binary coding for the tag IDs, we aim to accommodate more bits in the limited bandwidth range without degrading the RFID tag identification performance within acceptable small number of retransmissions/interrogations. To this end, the detection of each bit was formulated as a binary asymmetric channel. We note that such a formulation was firstly introduced in [22]. In this paper, we have integrated the signal processing approach into this BAC model so as to enhance the SNR of the underlying channel and thus improve the overall channel quality. In addition, the signal processing approach is compared with the multiple interrogations with majority rule based detection. The signal processing approach is a kind of soft-decision technique, while the multiple interrogations with majority rule based detection is a kind of hard-decision technique. Not surprisingly and as shown by numerical results, the former yields better performance than the latter. However, the latter is more robust and easily implementable in practice since it does not need the assumption of a time-invariant channel as required by the former.

Note that the idea of accommodating more bits in the limited bandwidth range may lead to a worse underlying channel for each bit detection. Therefore, errors may occur more frequently in each transmission of the RFID tag. An effective countermeasure we have proposed to address this issue is to use error detecting codes, the advantages of which are to decrease the probability of undetectable errors (i.e., to reduce the false identification rate of the tag ID) and to automatically issue a retransmission in case of errors being detected. As demonstrated in the numerical results, the employment of error-detecting codes can significantly diminish the false identification rate using only a reasonable number of retransmissions/interrogations; and the BAC dedicated codes, for instance, the constant weight codes and the Berger-Freiman codes perform better than the CRC codes in this scenario. A final remark is that the proposed framework does not work for the harsh applications where the tags are subjected to severe bending or even crumbling so that some bits might be read always erroneously. It is proposed in [22] to use the error-correcting codes to encode the RFID tags to tackle this challenge.

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