Low-Complexity Channel Estimation For The UHF MIMO-RFID Systems With Optimal Training

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Abstract

The issue of channel estimation is investigated for the UHF MIMO-RFID systems over pinhole channel. A training sequences aided linear minimum mean square error (LMMSE) channel estimation is proposed for the whole pinhole channel at the reader. Based on the mean square error (MSE) criterion, the MSE bound and the optimal training scheme with this bound are derived. By exploiting the optimal training scheme, an optimal low-rank LMMSE channel estimator is proposed to reduce the computational complexity via singular value decomposition (SVD). Furthermore, the Chu sequence is employed as the optimal training sequence to implement with easy realization and reduce computational complexity at the reader. Simulation results demonstrated the effectiveness of the proposed low-complexity channel estimation method and the superiority of the derived optimal training scheme for the MIMO-RFID systems.

Keywords: RFID, MIMO, Channel Estimation, MMSE

1. Introduction

Radio Frequency Identification (RFID) is a wireless identification technology[1,2,3]. Distinguishing active and passive RID systems is depending on the source of power supply of RFID tags in the Ultra High Frequency (UHF) band. In the passive systems, the tags absorb energy from an electromagnetic field provided by the RFID reader and use backscatter modulation for communication. Due to the multipath propagation of the UHF RFID systems, the provided reliability of these systems hampers their ubiquitous introduction to the market. Bertocco et al. [1] recently demonstrated that with commercial RFID equipment only a fraction of tags is accessible in real world scenarios. Similar results have been shown in [4, 5].

In order to deal with the multipath propagation environment, the reader of Feig Electronics [6] employ multiple receive antennas with Random Antenna Selection (RAS, equivalently named antenna switching). In [7], it was shown that pinhole diversity is available in a rich scattering pinhole channel with multiple antennas. Pinhole diversity can lead to the reduction in the power required to maintain a certain BER. Due to the low cost and complexity of the tag, tags generally don’t use multiple antennas and the readers also don’t use too many antennas when transmitting.

Almost all the existing work has assumed perfect channel state information (CSI) at both the tag and/or the reader. However, under the assumption of coherent detection, the fading channel coefficients need to be estimated and then used in the detection process. Channel estimation in the MIMO RFID systems can be very different from the traditional point-to-point transmission. First, designing the training sequence not only needs to consider channel of the return link, but also the channel of the forward link and their correlation. Secondly, the whole channel is comprised of the multiple subchannels, so it needs much more expenditure of designing training sequences. Angerer et al. [8] utilized I/Q plane to map the constellation for channel estimation in the flat-fading channels. Kim al. [9] analyzed the interrogation range influenced by the channel estimation error and mentioned the method for channel estimation, but didn’t estimate the channel.

In this work, the training-sequences-aided LMMSE channel estimation method for MIMO-RFID systems over pinhole channel is studied. First, the mean square error (MSE) bound on the proposed method is derived and the optimal training scheme is also given. Then by exploiting the inherent orthogonal characteristic of the optimal training scheme, an optimal low-rank channel estimator based on singular value decomposition (SVD) is introduced to reduce the complexity of channel estimation. Furthermore, the Chu sequence is employed as the training sequence at the reader to achieve the minimum MSE estimation performance.
2. System model and channel model

Considering the MIMO-RFID system with $N$ reader transmit antenna, one RF-tag antenna and $M$ reader receive antennas is in a hot-spot area of RFID. Through this paper, the reader and tag are assumed to synchronize perfectly. The symbol transmitted by the $ith$ antenna of reader at time instant $k$ is denoted $s_i(k)$. For the tag, the received signal at the distance of $d$ m from a reader is given as:

$$z_i[k] = \sqrt{P_{tx}} G_{tx} G_{tag} P_0 d^{-\delta} \sum_{m=0}^{L-1} h_i^r(m)s_i(k-m) + n_i(k)$$

where $P_{tx}$ denotes the total transmit power, $P_0$ is the reference path loss at the distance of 1 m, and $G_{tx}$ and $G_{tag}$ are the gain of the transmit antenna at the reader and the gain of the tag antenna, respectively. $n_i(k)$ is an additive white Gaussian noise (AWGN) signal. $h_i^r(n)$ and $h_i^l(n)$ are the channel response of the forward link and return link respectively.

For transmitting information to the reader, tag uses backscatter modulation. The tag changes from absorbing energy to the reflecting energy, by mismatching their antenna input impedance. Thus, the reflection coefficient of the RF tag load keeps unchange per frame. The reflection coefficients at the tag and describes the time-varying modulation of the carrier signal. The channel response of the forward link and return link respectively, $h_i^f$ and $h_i^r$ are assumed to follow the Rayleigh fading distribution which the discrete impulse response is:

$$h_i^r[n] = \sum_{l=0}^{L_r-1} h_i^{r,l} \delta(n-l)$$

where $L_r$ and $L_a$ represents the number of resolvable paths for the forward and reverse link respectively, $h_i^{r,l}$ and $h_i^{r,l}$ denotes the channel gain of the path $l$ with variance $\sigma_i^{r,l}$, $\sigma_i^{r,l}$ respectively.

All fading channels are independent for the RFID systems, $\sum_{l=0}^{L_r-1} \sigma_i^{r,l} = 1$, and $\sum_{l=0}^{L_a-1} \sigma_i^{r,l} = 1$.

3. Low-complexity LMMSE channel estimation method

3.1. LMMSE channel estimation method
This subsection proposes a training-based method for channel estimation for MIMO-RFID systems. The number of the training sequence with unit power is $K$. As the first $L$ training symbols are influenced by unknown previous symbols both at the forward link and return link, only the last $K - L$ received training symbols can be used for channel estimation.

With the matrix $S_i$ of dimension $(K - L) \times L$ at the $ith$ transmit antenna of the reader, which contains the transmitted training symbols, given by:

$$S_i = \begin{bmatrix} s_i(L) & \cdots & s_i(1) \\ s_i(L+1) & \cdots & s_i(2) \\ \vdots & \vdots & \vdots \\ s_i(K) & \cdots & s_i(K-L) \end{bmatrix} \quad (5)$$

The receive vector of $jth$ antenna at time instant $k$ at the reader is:

$$y_j = \begin{bmatrix} y_j(L), \cdots, y_j(K) \end{bmatrix}^T = \begin{bmatrix} \rho E_{E_{\text{tag}}} s(L) h_j + n(L), \cdots, \rho E_{E_{\text{tag}}} s(K) h_j + n(K) \end{bmatrix} \quad (6)$$

Stacking the receive signals of each antenna into a vector:

$$y = [y_1^T, \cdots, y_M^T]^T = Ah + n \quad (7)$$

where $A = \rho h S \otimes I_d = [A_1, \cdots, A_M]$, $S = [S_1, \cdots, S_M]$, $h = [h_1^T, \cdots, h_d^T]^T$, $h_j = [h_j^T, \cdots, h_j^T]^T$, $h_j = h_j^T \otimes h_j^T$, $n = [n_1^T, \cdots, n_d^T]^T$. Note that, the channel vector $h$ is identifiable if and only if $A$ has full column rank, which occurs when: $$(K - L) \geq \sum_{i=1}^{N} ML_i.$$ It is found that the training length should not be less than the channel tap number of all concatenation links; otherwise, the channel vector $h$ would be unidentifiable.

The simplest algorithm for the channel estimation using (7) is the LS estimator. However, it is intractable to perform MMSE channel estimation for RFID systems because the total channel $h$ is non-Gaussian. Therefore, the suboptimal LMMSE channel estimator is proposed and analyzed. Exploiting the non-correlation property of the channels of the different antennas, the autocorrelation matrix of the channels vector $h$ is:

$$C_h = E[hh^H] = \text{diag} \left( \sigma_{1,1}^2 \otimes \sigma_{1,1}^2, \cdots, \sigma_{d,1}^2 \otimes \sigma_{d,1}^2, \sigma_{1,2}^2 \otimes \sigma_{1,2}^2, \cdots, \sigma_{d,d}^2 \otimes \sigma_{d,d}^2 \right) \quad (8)$$

The autocorrelation matrix of the receive signal $x$ is:

$$C_x = E[yy^H] = AC_h A^H + C_a \quad (9)$$

where $C_a = \sigma_a^2 I_{(K-L)ML}$ is the autocorrelation matrix of the additional noise. Based on the LMMSE criterion [10], the estimated channel can be written as:

$$\hat{h} = C_x A^H C_h^{-1} y \quad (10)$$

And the autocorrelation matrix of estimation error is:

$$C_e = E[ee^H] = \left( C_h^{-1} + A^H C_a^{-1} A \right)^{-1} \quad (11)$$

When $C_h$ is rank deficient, a small value can be added to the diagonal of $C_h$. Therefore, the average MSE of the LMMSE channel estimator can be represented as:

$$J_e = \frac{1}{NML} tr[C_a] = \frac{1}{NML} tr \left[ \left( C_h^{-1} + A^H C_a^{-1} A \right)^{-1} \right] \quad (12)$$

**Lemma 1** For positive definite $M \times M$ matrix $B$ with its $mth$ diagonal element given by $b_m$, the following inequality holds: $tr[B^2] \geq \sum_{m=1}^{M} \frac{b_m^2}{b_m}$, where equality holds if and only if $B$ is diagonal[10, page 65]. Based on this lemma, diagonal matrix $C_h$ is positive defined, the minimum of (12) is achieved if and only if $A^H A$ is diagonal. Therefore, the optimal training scheme is:

$$A^H A = c_i P_i \chi_g \chi_x (G_{\chi_g} P_i d^H E_{\chi_g})^H (K - L) M_{\chi_x}, \quad \forall i \in \{1, \ldots, M\} \quad (13)$$

$$A^H A = 0_{L \times L}, \quad \forall m, n \in \{1, \ldots, M\}, \text{with } m \neq n \quad (14)$$

By substituting (8), (13) and (14) into (12), the MSE bound of this channel estimation method is:
3.2. Low-complexity LMMSE channel estimator

The LMMSE channel estimator (10) is of considerable complexity because a matrix inversion is involved. To simplify this estimator, exploiting the optimal training scheme (14, 15) to get an optimal low-complexity LMMSE channel estimator is based on SVD in [11].

If \( V_1 \in \mathbb{C}^{nxn} \) has orthonormal columns, then there exists \( V_2 \in \mathbb{C}^{(k-L)\times n} \), such that \( V=[V_1\ V_2] \) is orthogonal[12, page 69]. Based on this lemma, there exists \( K \times (K-N) \) matrix \( Q \) to make \( K \times K \) matrix \( U=[X\ Q] \) ensure \( U^*U=\text{diag}(u) \), because the training matrix \( A \) shown in (14, 15) has orthonormal columns. Denote the diagonal entry of \( A^*A \), \( C_x \), and \( C_y \) as \( \alpha \), \( \lambda \) and \( \gamma \). Let the \((K-L)\times 1 \) vector \( \beta=[\lambda \ 0_{(K-L-N+1)}] \). Then, the Hermitian matrix \( AC_xA^* \) can be rewritten as:

\[
AC_xA^* = U \text{diag}(\beta)U^* = P \text{diag}(\sqrt{\beta}) \text{diag}(\sqrt{u}) P^* = P \text{diag}(u \odot \beta) P^* \tag{16}
\]

where \( P \) is a unitary matrix from \( U \). Substituting (16) into \( \hat{h} \) yields:

\[
\hat{h} = \text{diag}(\hat{eta}/(\alpha \odot \lambda + \gamma))A^*y \tag{17}
\]

Using (16) and (17), LMMSE channel estimator (10) can be rewritten as(The proof is neglected):

\[
\hat{h} = \text{diag}(\hat{eta}/(\alpha \odot \lambda + \gamma))A^*y \tag{18}
\]

From (18), the optimal low-rank LMMSE channel estimator avoids the matrix inverse calculation, so the computation complexity is significantly reduced compared with (13). Condition (14, 15) is also the optimal training scheme for LS channel estimator based on the similar deduction of minimizing MSE. So the performance of the proposed channel estimator is equal to the Winer-filtered LS channel estimator. The LS channel estimator can be used to obtain initial channel estimates and these estimates can be further used to estimate second channel statistics \( \lambda \) and the noise power \( \gamma \).

3.3. Design of the optimal training

For channel estimation there are \( NML \) unknowns in the whole channel, as remarked before. There have to be leastwise as many training symbols as unknowns to estimate the channel and as only the last \( K-L \) training symbols may be used for the estimation, the number of training symbols \( K \) per frame has to be least:

\[
\sum_{i=1}^{N} k_i (k) = P_{TX}^2 = \text{constant} \geq (K-L) \tag{19}
\]

should be constant and equal. According to (14, 15), (5) and \( A = \rho S \otimes I_M \), the matrix \( S \) has orthogonal columns and is cyclic. A cyclic \( S \) with orthogonal columns can be constructed by writing a perfect root-of-unity sequence (PRUS) into the first row of \( S \) and then by filling any next row with the one element right shifted version of the previous row[13].

A perfect root-of-unity sequence \( s(k) \) can be constructed for any length \( N \) by the Chu-sequences which are of the form[14]:

\[
s(k) = \begin{cases} 
    e^{jMK^2/N} & \text{for } N \text{ even} \\
    e^{jMK/N} & \text{for } N \text{ odd}
\end{cases} \tag{20}
\]

where \( M \) is a natural number greater than zero and needs to be comprime to \( N \). The length \( N-L \) of required perfect root-of-unity is equal to the number of columns of \( S \). The sequence \( s(k) \) is written into the first row of \( S \). Any next row is the one element right shifted version of the previous row. Moreover, since the Chu sequence exists for any finite length, for any finite number of total channel paths, this training scheme can always achieve the minimum MSE estimation performance.
4. Performance analysis

In this section, the performance of proposed channel estimation as well as the optimal training designs is studied under various scenarios for the MIMO-RFID systems. In the simulation, it is assumed that the reader supports a 50 kb/s data rate, and tag uses FM0 coding and the 250KHz channel bandwidth specified by China regulations[14]. Moreover, $P_0$ and $\gamma$ in the path loss model are set to -31.6 dB and 3.0[9], respectively. The pinhole channel consists of forward link channel and return link channel. Both channels are assumed to follow independent Rayleigh fading with an exponential power delay profile, that is:

$$E\left[|h_l(t)|^2\right] = \sum_{k=0}^{L} e^{-\alpha k}, \quad l = 0,\ldots,L$$

where $\alpha$ is the exponent coefficient. Perfect synchronization among reader and tag is assumed to observe the channel estimation alone. The performance of channel estimation will be measured by the normalized MSE (NMSE), which is defined as $\{E\left[|h_l(t)|^2\right] - E\left[|\hat{h}_l(t)|^2\right]\}/E\{|h_l(t)|^2\}$. The reader applies maximum ratio combining (MRC) and uses Gaussian sequences as random sequences, the optimal diversity combing technique for the identical signaling scheme. The transmission power at the tag is normalized according to China regulation. All other parameters used in the simulation are summarized in Table 1.

<table>
<thead>
<tr>
<th>parameters</th>
<th>values</th>
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<tr>
<td>Frequency band</td>
<td>917-920.6 MHz</td>
</tr>
<tr>
<td>Channel bandwidth</td>
<td>250 KHz</td>
</tr>
<tr>
<td>Tag’s data rate(FM0)</td>
<td>50 Kbps</td>
</tr>
<tr>
<td>Transmit power $P_T$</td>
<td>33 dBm</td>
</tr>
<tr>
<td>Reference path loss $P_0$</td>
<td>-31.6 dB</td>
</tr>
<tr>
<td>Tag’s power ref.coeff. $E_{tag}$</td>
<td>0.1 dB</td>
</tr>
<tr>
<td>Reader antenna gain, $G_t$, $G_R$</td>
<td>6 dBi</td>
</tr>
<tr>
<td>Tag antenna gain, $G_{TAG}$</td>
<td>-1 dBi</td>
</tr>
<tr>
<td>Reader-tag dist. $d$</td>
<td>6 m</td>
</tr>
</tbody>
</table>

Figure 1 illustrates the MSE of the proposed LMMSE channel estimation for different numbers of transmit antennas when both length-64 Chu and random sequences are used with the same receive antennas. And the MSE bound is also illustrated. According to figure 1, the optimal training scheme mentioned in section 3 indeed attains the MSE bound and outperforms substantially random training sequences. Besides, the MSE bound is below $10^{-3}$ in moderate to high SNR, indicating good channel estimation performance. Note that as the number of transmit antennas increases from $N = 2$ to $N = 4$, the estimator’s performance degrades according for different mean SNR for the same training sequence. This can be explained by the fact that as $N$ increases, the transmit power for each antenna must also be reduced accordingly by a factor of $N$ in order to maintain constant total transmit power to the tag. Hence, the effective SNR is also reduced. This is also expected from (15).

Figure 2 illustrates the MSE of the proposed LMMSE channel estimation for different numbers of receive antennas when both length-64 Chu and random sequences are used with the same transmit antennas. According to figure 2, it can be seen that as the number of receive antennas increase from $M = 2$ to $M = 4$, the estimator’s performance changes little accordingly for different mean SNR for the same training sequence. In other words, increasing reception diversity does not offer any advantage to the proposed MIMO channel estimators. This can be explained from (15), as $M$ increases, the MSE does not change much. In fact, the estimator has to actually estimate more paths as result of additional number of received antennas. So the next sector just analyzes the different transmit antennas situation with the same receive antennas.
Figure 3 displays the impact of the length-32, length-64 and length-128 optimal training on the MSE performance with the same receive antennas. Note that the longer training sequences lead to the higher MSE performance for the same antennas. This is expected from (15) since the transmitting energy in the training section is linear with the training length $K$. It is also seen from this figure that the length-32 channel estimator would not work when the transmit antenna number increased beyond 4. The reason for this phenomenon is because the antenna number that can be supplied by this channel estimator is bounded by the identifiable condition. Thus, to avoid this phenomenon, it is crucial to make the training length $K$ not less than the channel tap number of all concatenation links.

Figure 4 plots the BER performance corresponding to the length-64 optimal and the suboptimal training schemes when different numbers of transmit antennas are employed with the same receive antennas.
antennas. As expected from the MSE performance comparison results, a substantial BER performance gain of the optimal training scheme over the suboptimal one is observed. The BER performance of perfect CSI is also given as a benchmark. From the figure, the BER performance of the optimal training scheme is very close to the perfect CSI case, which confirms the accuracy of the proposed channel estimation method, while the performance gap increases when transmit antennas are involved. This can be explained by the fact that the MSE performance decreases as the number of antennas increases. However, since spatial diversity is dominant in the BER performance relative to the channel estimation error, more antennas provide better BER performance.

![Fig. 3 Impact of the different length of training sequences on the MSE performance](image1)

![Fig. 4 BER performance comparison of the proposed channel estimation method using training sequences](image2)
The description of the proposed channel estimation method in Section 3 shows that the overall complexity comes from complex matrix operations at the reader. Exploiting the optimal training scheme to derive a low-rank LMMSE channel estimator (18) is based on SVD, where the performance is essentially preserved. Therefore, the complex matrix inverse calculation in the destination terminal can be avoided. To conclude, only complex multiplications \((K + 1)MNL\) and \((K - 1)MNL\) complex additions are required to obtain the accurate time-domain CSI in the MIMO-RFID systems.

5. Conclusions

This work mainly studies the training-sequences-aided LMMSE channel estimation method for the MIMO-RFID systems over the pinhole channel. Exploiting the inherent orthogonal characteristic of the optimal training scheme to simplify the LMMSE channel estimator and to propose low-rank LMMSE channel estimator based on SVD has low-complexity where the performance is essentially preserved. In addition, the Chu sequence is employed as the training sequences to achieve the minimum MSE performance. The simulation results have verified the performance of the proposed low-complexity channel estimation method in the MIMO-RFID systems.

6. References