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Blind Channel Estimation for Frequency Hopping System Using Subspace Based Method

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Abstract**—**Subspace channel estimation methods have been studied widely, where the subspace of the covariance matrix is decomposed to separate the signal subspace from noise subspace. The decomposition is normally done by using either the *eigenvalue decomposition* (EVD) or the *singular value decomposition* (SVD) of the *auto-correlation matrix* (ACM). However, the subspace decomposition process is computationally expensive. This paper considers the estimation of the multipath slow *frequency hopping* (FH) channel using noise space based method. In particular, an efficient method is proposed to estimate the multipath time delays by applying *multiple signal classification* (MUSIC) algorithm which is based on the null space extracted by the *rank revealing LU* (RRLU) factorization. As a result, precise information is provided by the RRLU about the numerical null space and the rank, (i.e., important tool in linear algebra). The simulation results demonstrate the effectiveness of the proposed novel method by approximately decreasing the computational complexity to the half as compared with RRQR methods keeping the same performance.

Keywords**—**Time Delay Estimation, RRLU, RRQR, MUSIC, LS-ESPRIT, LS-ESPRIT, Frequency Hopping.

I. Introduction

In many cellular networks, various techniques have been proposed to improve (i.e., maximize) the overall network capacity of the existing systems. In general, current cellular networks are facing extensive growth, which requires effective application of the existing spectrum. Examples for such techniques include cell splitting, resource reuse (TDMA / FDMA), *adaptive frequency allocation* (AFA), *fixed channel allocation* (FCA), *dynamic channel allocation* (DCA), hierarchal cell structure, *discontinued transmission* (DTX), *power control* (PC), smart antennas, and slow FH.

In this scope, the *frequency hopping spread spectrum* (FHSS) technique is efficiently used for securing radio transmission in military wireless communications and networks. Moreover, it is widely used in *code division multiple access* (CDMA) satellite systems as well as an ambiguity optimization method in *multiple input multiple output* (MIMO) radars. This is due to its well-known capabilities (i.e., Needs less synchronization as compared with *direct sequence spread spectrum* (DSSS) and low probability of detection and interception) [1].

Moreover, commercial networks utilize this technique to improve performance in fading environments. As a result, it is extensively used in modern wireless networks (i.e., Bluetooth networks, FH ad hoc networks, and *wireless personal area networks* (WPAN) [2], [3]). The authors in [4] studied the hybrid DS-FH systems, which is a combination of DSSS and FHSS. Based on that, parameters estimation and detection of FHSS signals are of great deal of interest and have attracted much attention for several decades.

Due to rapid time varying multipath fading in mobile wireless communications, the overall system performance is severely degraded in general. Therefore, the design of the traditional communication systems takes in to account the worst-case channel conditions. This can be achieved by maintaining an acceptable performance for poor channel quality by using a fixed link margin. The transmitter and the receiver are not optimized for the current channel condition. In particular, the potential of the wireless fading channel is not exploited properly in these systems. This is besides the bandwidth and power utilization deficiencies. But this is not good for future wireless communication systems, where high-speed data rates are desired. As a result, improving the bandwidth efficiency with low power constraint for mobile wireless communications is desired. To reach this target, one can adapt the signal transmission due to the channel conditions. If the *channel state information* (CSI) can be precisely predicted at the transmitter, the capacity of fading channel can be approached by adapting the transmission parameters to the exact fading level. The proposed approach converts the delay estimation problem using temporally received packets to a new approach that can estimate the directions of arrival in array processing by applying MUSIC algorithm in which the null space, which is extracted by the RRLU [5].

In general, modeling the channel using the flat fading model is not valid for the narrow band FH system [3], [4], where the bandwidth of interest is close to multipath coherence bandwidth. In this scope, the multipath time delay estimation approach is studied for a SFH system using ESPRIT [6] and SPECC [7] algorithms where it becomes significant at high data rates for frequency selective received signal in SFH systems.

This paper addresses the estimation of the multipath time delay parameters using RRLU factorization which provides more precise rank and numerical null space. Finally, one can estimate the frequencies from the projection matrix by using the MUSIC algorithm. The remainder of the paper is organized as follows. Section II presents general multipath channel model. In Section III, the highly efficient proposed methods and formulation are shown. Next, Section IV

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describes the performance of these methods through MATLAB simulations with comparisons with the previous works [8], [10]. Finally, Section V concludes this paper.

II. Multipath Channel Model

In general, multipath interference phenomenon resulting due to reflections coming from surroundings items (i.e., trees, hills, towers, and buildings) affects the overall behavior of the mobile wireless communication. (e.g., the objects around the wireless signal path cause signal reflection, where the receiver get some of these reflected waves). Each reflected signal has different amplitude and phase since it takes different path as shown in Fig. 1. Therefore, accurate time-varying channel estimates are needed at the receiver in order to provide reliability and relatively high data rates.

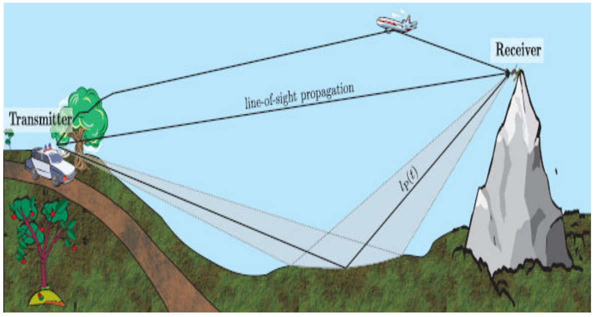

Fig. 1 Multipath Fading Channel

The total received power at the receiver may increase or decrease depending upon the phases of these multiple signals. (i.e., a varying slight the position of the mobile may cause a significant phases difference). Furthermore, sending a single impulse from the transmitter will result in multiple copies to be received at different time instances since different paths are of different lengths as shown in Fig. 2. As a result, channel estimation is an important technique, especially in mobile wireless network systems, where the wireless channel changes instantaneously over time and this is usually caused by the motion of the transmitter and/or receiver at nearly vehicular speeds [1].

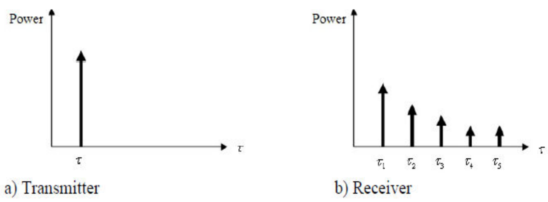

Fig. 2 The power delay profile for the multipath channel of $\tau_i$delay

III. Problem Formulation

In general, the FH received signal baseband model of the sampled version form, in multipath environments, is given by:

$$y^{(n)}(kT) = \sum_{i=1}^{P} \beta_i e^{-j2\pi f_n \tau_i} s(kT - \tau_i; \boldsymbol{b}_n) + w^{(n)}(kT) \quad (1)$$

Here, $y^{(n)}(kT)$ is the n^{th} hop received signal, f_n is the n^{th} hop frequency, T is the sampling period, $s(kT; \boldsymbol{b}_n)$ is the transmitted baseband signal, $w^{(n)}(kT)$ is the white Gaussian noise parameter, $\boldsymbol{b}_n$ is the sequence of the transmitted bits, and the total number of the multipath that are considered in the proposed model is denoted by the parameter P. Furthermore, τ_i and β_i (i.e., which are assumed to be independent of the hop index) are the time delay and the channel gain corresponding to the i^{th} multipath, respectively.

In general, time delay, channel gain, and the sequence of the transmitted bit are considered to be unknowns, while the hop frequency is known. This paper addresses only the estimation of the time delays based on the received signal. In particular, after estimating the time delays, two separate *maximum likelihood* (ML) approaches can be utilized to estimate the parameters β_i and $\boldsymbol{b}_n$.

Next, three more assumptions are made in this paper [3], [4], [6]. The first block of symbols is fixed for all hops, the time delays remain very slow or constant, and no frequency change within one packet. Based on the set of samples that remains constant (first block of symbols that is fixed for all hops), the discrete time version of (1) is given as follows:

$$y^{(n)}[k] = \sum_{i=1}^{P} \beta_i e^{-j2\pi f_n \tau_i} s_i[k] + w^{(n)}[k]$$
$$k = 1,2, \ldots K \quad (2)$$

where $s_i[k]$ is the i^{th} multipath delayed version of the transmitted signal. Considering the part of data that are constant over all hops, the signal samples $s_i[k]$ are independent of n. So far, $s_i[k]$ are still unknown. Let the K samples in (2) are given by:

$$\boldsymbol{y}_n = [y_n(0)\; y_n(1)\; \ldots \ldots y_n(K-1)] \quad (3)$$

Collecting data from a number of hopes and then partitioning the packets of data into $M+1$ subsets $\{F_0, F_1, F_2, \ldots \ldots F_M\}$ each of which is of received frequencies at least of size N as $F_i = \{f_{i1}, f_{i2} \ldots \ldots f_{iN}\}$ where:

$$f_{0i} = f_{1i} + \Delta_f = f_{2i} + 2\Delta_f = \cdots = f_{Mi} + M\Delta_f \quad (4)$$

Note that the constant Δ_f separates any two successive frequency sets. Within the frequency set of interest, let f_{ij} be obtained at the $n_{ij}{}^{th}$ hope.

Now, define i^{th} vector of $\boldsymbol{y}$ as:

$$\boldsymbol{y}_i\,[k] = [y^{n_{i1}}[k], \;\; \ldots, \;\; y^{n_{iN}}[k]]^T \quad (5)$$

$$\boldsymbol{Y}_i = [\boldsymbol{y}_i\,[1], \boldsymbol{y}_i\,[2], \ldots, \boldsymbol{y}_i\,[K]] \quad (6)$$

$$\mathbf{x}[k] = [\beta_1 s_1[k], \;\; \ldots \;\; \beta_P s_P[k]]^T \quad (7)$$

$$\boldsymbol{X} = [\mathbf{x}[1], \mathbf{x}\,[2], \ldots, \mathbf{x}\,[K]] \quad (8)$$

One can easily represent the M subsets as:

$$\begin{aligned} Y_0 &= A\Phi^0 X + W_0 \\ Y_1 &= A\Phi^1 X + W_1 \\ Y_2 &= A\Phi^2 X + W_2 \\ &\vdots \\ Y_M &= A\Phi^M X + W_M \end{aligned} \tag{9}$$

where the matrices $\boldsymbol{A}$ and $\boldsymbol{\Phi}$ are defined as:

$$\boldsymbol{A} = \begin{bmatrix} e^{-j2\pi f_{p1}\tau_1} & \dots & e^{-j2\pi f_{p1}\tau_p} \\ \vdots & \vdots & \vdots \\ e^{-j2\pi f_{pN}\tau_1} & \dots & e^{-j2\pi f_{pN}\tau_p} \end{bmatrix}$$

$$\boldsymbol{\Phi} = diag(e^{-j2\pi\Delta_f\tau_1} \dots \dots e^{-j2\pi\Delta_f\tau_p})$$

and $\boldsymbol{W}_1, \boldsymbol{W}_2, \dots \boldsymbol{W}_M$ are the corresponding noise matrices. The sub-matrices, which are calculated in (9), can be collected in the matrix **Y** as:

$$\boldsymbol{Y} = \begin{bmatrix} \boldsymbol{Y_0} \\ \boldsymbol{Y_1} \\ \vdots \\ \vdots \\ \boldsymbol{Y_M} \end{bmatrix}$$

Moreover, the auto-correlation matrix, $\mathbf{R}_{yy}$, is given by:

$$\mathbf{R}_{yy} = E[\boldsymbol{Y}\,\boldsymbol{Y}^{\boldsymbol{H}}] = \begin{bmatrix} \boldsymbol{R}_{00} & \dots & \boldsymbol{R}_{0M} \\ \vdots & \ddots & \vdots \\ \boldsymbol{R}_{M0} & \dots & \boldsymbol{R}_{MM} \end{bmatrix} + \sigma^2\mathbf{I} \tag{10}$$

where σ^2 is the variance of the noise [11]. Here, $\mathbf{R}_{ij}$ represents the correlation between the data coming from F_i and that coming from F_j. The sub-matrices of the autocorrelation matrix are given by:

$$\mathbf{R}_{ij} = E\left[\boldsymbol{Y_i}\,\boldsymbol{Y_j}^{\boldsymbol{H}}\right] = \boldsymbol{A\Phi^i}\mathbf{R}_{\boldsymbol{xx}}\boldsymbol{\Phi^{*j}A^H} \tag{11}$$

Finally, by applying LU decomposition to the above auto-correlation matrix, the signal and noise subspace separation is achieved.

LU Decomposition Theorem: Let $\boldsymbol{A} \in R^{n\times n}$ and all the leading principal minors $det\,(\boldsymbol{A}(1{:}k;1{:}k) \neq 0, k = 1,2,\dots,n$. Then a unique unit lower triangular **L** exists with diagonal elements all equal to one and a unique upper triangular matrix **U** such that **A** = **LU**, and $de\,t(\boldsymbol{A}) = u_{11}u_{22}\dots u_{nn}$.

Now, applying LU decomposition [9] to the auto-correlation matrix, $\mathbf{R}_{yy}$, it can be expressed as a product of a lower triangular matrix $\boldsymbol{L}$ (with 1's on the diagonal), and an upper triangular matrix $\boldsymbol{U}$ as shown next:

$$\mathrm{R}_{yy} = \boldsymbol{LU} = \begin{bmatrix} \boldsymbol{L}_{11} & \boldsymbol{0} \\ \boldsymbol{L}_{21} & \boldsymbol{L}_{22} \end{bmatrix} . \begin{bmatrix} \boldsymbol{U}_{11} & \boldsymbol{U}_{12} \\ \boldsymbol{0} & \boldsymbol{U}_{22} \end{bmatrix} \tag{12}$$

where the sub matrix $\boldsymbol{U_{11}}$ is an upper triangular matrix of size $(P \times P)$ and the sub matrix $\boldsymbol{U_{12}}$ is of size $\left(P \times (N - L + 1)\right)$. Since $\boldsymbol{U_{22}}$ has small norm, one can easily extract the basis of the noise space form the sub matrix $\widetilde{\boldsymbol{U}} = [\boldsymbol{U_{11}} \quad \boldsymbol{U_{12}}]$.

$$\mathrm{R}_{yy} \approx \tilde{\boldsymbol{L}}\widetilde{\boldsymbol{U}} = \begin{bmatrix} \boldsymbol{L}_{11} \\ \boldsymbol{L}_{21} \end{bmatrix} . [\boldsymbol{U}_{11} \quad \boldsymbol{U}_{12}] \tag{13}$$

Here, the first p columns of the matrix $\boldsymbol{L}$ constructs the matrix $\tilde{\boldsymbol{L}}$. The matrix $\widetilde{\boldsymbol{U}}$ expresses the signal space. The spans of the signal subspace and the matrix **A,** which is shown in (9), is the same. In particular, the following properties must be satisfied for any vector that belongs to the null space, $\mathbf{R}_{yy}^{\perp}$:

$$\mathbf{R}_{yy}.\mathbf{R}_{yy}^{\perp} = \boldsymbol{0} \tag{14}$$

$$\tilde{\boldsymbol{L}}.\widetilde{\boldsymbol{U}}.\mathbf{R}_{yy}^{\perp} = \begin{bmatrix} \boldsymbol{L_{11}} \\ \boldsymbol{L_{21}} \end{bmatrix} . [\boldsymbol{U_{11}} \quad \boldsymbol{U_{12}}] \begin{bmatrix} \boldsymbol{g_1} \\ \boldsymbol{g_2} \end{bmatrix} = \boldsymbol{0}$$

Simply:

$$\widetilde{\boldsymbol{U}}.\boldsymbol{G} = \boldsymbol{0} \tag{15}$$

$$[\boldsymbol{U_{11}} \quad \boldsymbol{U_{12}}] \begin{bmatrix} \boldsymbol{g_1} \\ \boldsymbol{g_2} \end{bmatrix} = \boldsymbol{U_{11}g_1} + \boldsymbol{U_{12}g_2} = \boldsymbol{0}$$

Since $\boldsymbol{U_{11}}$ is a nonsingular matrix, $\boldsymbol{g_1}$ can be written in terms of $\boldsymbol{g_2}$ as:

$$\boldsymbol{g_1} = -\boldsymbol{U_{11}^{-1}U_{12}g_2}$$

then

$$\boldsymbol{H} = \begin{bmatrix} -\boldsymbol{U_{11}^{-1}U_{12}} \\ \boldsymbol{I}_{(L-P)} \end{bmatrix} \boldsymbol{g_2} \tag{16}$$

Here, and in order to satisfy orthonormality, an orthogonal projection onto the subspace is applied to the none-orthonormal columns of the null space, $\boldsymbol{H}$, as:

$$\mathbf{R}_{yy}^{\perp} = \boldsymbol{H}(\boldsymbol{H^H H})^{-1}\boldsymbol{H^H} \tag{17}$$

Furthermore, the following function is utilized to perform the MUSIC like search algorithm, which is used to estimate the frequencies [11]:

$$f(\omega) = \left[\boldsymbol{A}(\omega)^{\boldsymbol{H}}\mathbf{R}_{yy}^{\perp}\boldsymbol{A}(\omega)\right]^{-1} \tag{18}$$

Now one can use the root-MUSIC method instead of searching for the peaks over ω in (18). Here, the frequency time-delays is taken as the values of the angles of the P roots that are closest to the circumstance of the unit circle for the polynomial D(z)

$$D(z) = \textstyle\sum_{i=0}^{L-1} \boldsymbol{V_i}(z)\boldsymbol{V_i^*}(1/z^*) \tag{19}$$

IV. Simulation Results

Extensive MATLAB simulations were done to validate the effectiveness of the proposed model, where three multipath model (P = 3) is considered. Furthermore, the range of transmission was confined to be between 1899 to 1929 MHz, which is the uplink frequency range for the PCS system. In particular, the model considered 75 frequencies with a 400 KHz frequency separation among carriers. Moreover, four QPSK symbols were assumed to structure the header of each packet in the proposed model and the symbol period is assumed to be 4µs. Next, the multipath delays are assumed to have the values of 0.15, 0.33, and 0.83 µs, respectively. The channel gain parameter is assumed to be a complex random. One thousand independent realizations are considered with normalized MSE which is defined as:

$$MSE = E\left\{\sum_{i=1}^{N_t}\sum_{j=1}^{P}\left(\frac{\tau_i-\hat{\tau}_i}{\tau_i}\right)^2\right\} \quad (20)$$

The RRLU method is compared with ESPRIT-LS, ESPRIT-TLS [8] and RRQR [10]. The RRLU shows a great matching with respect to the reference methods (i.e., LS and TLS) and it is very close to the RRQR as shown in Figs. 3 and 4. The normalized MSE of RRLU method versus SNR with different number of blocks is shown in Fig. 5. It clearly indicated that as more packets are involved, a better performance can be achieved.

V. Conclusion

A technique for the multipath delay estimation using frequency hopping system has been proposed in this paper using RRLU factorization. Comparisons with other various existing methods are made. The RRLU factorization with the MUSIC search shows a reasonable and almost similar to RRQR gaining the reduction of the complexity.

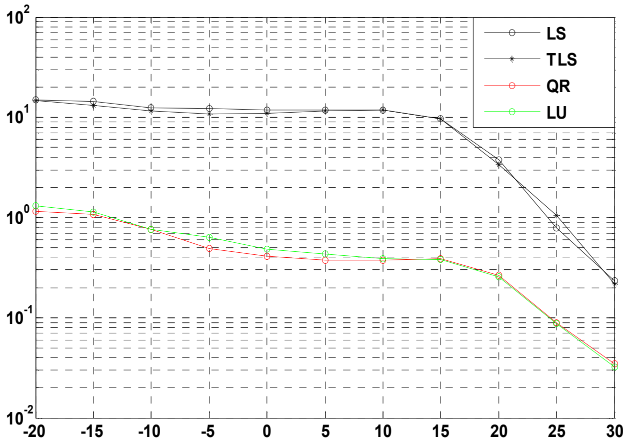

Fig. 3 Normalized MSE versus SNR (K=10)

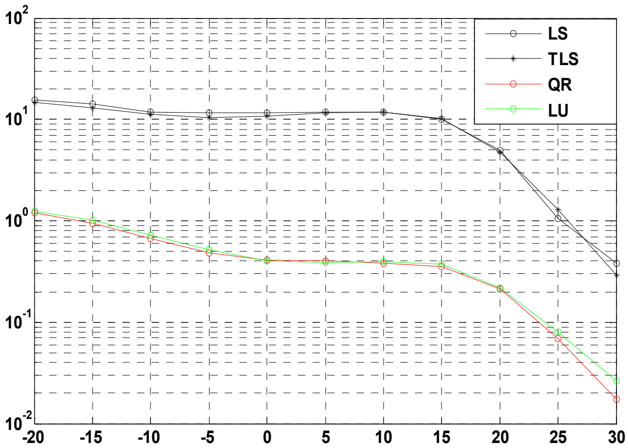

Fig. 4 Normalized MSE versus SNR (K=20)

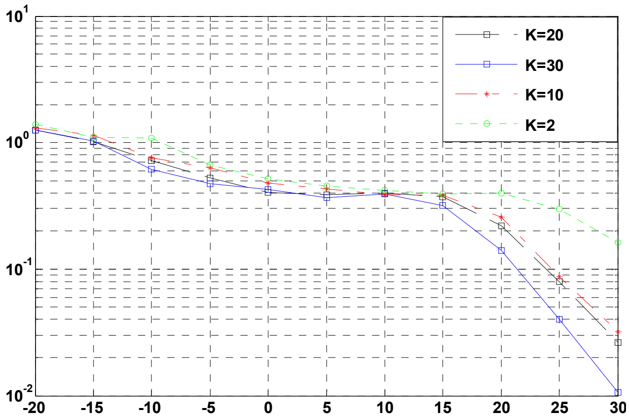

Fig. 5 Normalized MSE of RRLU method versus SNR for different number of blocks

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