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SVD-based Blind Estimation of M-Sequence with Modulated DS-CDMA Signals

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***Abstract*-An Singular Value Decomposition (SVD) based method was proposed and developed in this work to estimate the m-sequence of modulated Direct-Sequence Code Division Multiple Access (DS-CDMA) signals with carrier frequency offset. By building eigenvalues based objections, a proper carrier frequency range can be determined, so that the m-sequence can be blindly estimated by SVD in this carrier frequency range. Closed-form models and simulation results suggest that the proposed method is efficient in modulating signals.**

Keywords-Blind estimation of m-sequence; SVD; carrier frequency offset

I. INTRODUCTION

Direct-Sequence Code Division Multiple Access (DS-CDMA) has been extensively adopted for civilian and military purposes such as GPS, low orbit satellite communication systems, wireless sensors [1] and 3G networks. Spread spectrum sequences play an important role in affecting DS-CDMA performance, as well as the blind estimation of m-sequence is critical in civilian information monitoring [2] and military communication countermeasure. When analyzing the low power spectrum density of Direct Sequence Spread Spectrum (DSSS) signals, methods for blind estimation of m-sequence is always challengeable and hence becomes a hot topic in modern communication research. However, existing estimation methods all assume that the carrier frequency is already known [3][4], which, in reality, is hidden. Therefore, existing technologies fail to provide completely accurate results [5].

This work proposes a new m-sequence estimation method based on Singular Value Decomposition (SVD), which has been widely used in data sciences for factorial analysis. By analyzing the eigenvalues and eigenvectors from the SVD of the received signal covariance, and projecting the received signals onto the eigenvectors, the frequency can be modulated. The proposed SVD based method can estimate the m-sequence with a residual (i.e. error) frequency. Furthermore, a Phase Locked Loop (PLL) has been used to remove the residual frequency and accurately estimate the m-sequence.

II. SYSTEM MODEL

The received signal is assumed to have a Binary Phase-Shift Keying (BPSK) modulated spread spectrum, so that it can be expressed as:

$$r(t) = \rho b(m) s(t - mT_b - dT_c)\cos(2\pi ft + \varphi) + n(t) \qquad (1)$$

Where ρ is Signal to Noise Ratio (SNR), $b(m)$ is the m-th data, $s(t) \in \{\pm 1\}$ is user's spreading code, T_b is information bit duration , T_c is chip rate, d is signal propagation delay time, f is residual carrier frequency and φ is phase. The length of the spread spectrum code is C, and $T_b = CT_c$. Here we assume T_c and C are known, since they can be estimated by using algorithms such cyclic spectrum estimation [6] and high order statistical power spectral estimation [7]. $n(t)$ is the Gaussian noise, which is assumed as an Independent and Identically Distributed (IID) and follows the standard distribution.

When the sample frequency is $1/T_c$, and a continuous C samples is obtained to be composed of a C by 1 vector,

$$\boldsymbol{r}_m = [r(mC), r(mC+1), ..., r((m+1)C-1)]^T \qquad (2a)$$

Define a length of $2C$ vector,

$$\boldsymbol{q}_m = [\boldsymbol{r}_m^T, \boldsymbol{r}_{m+1}^T]^T \qquad (2b)$$

Inserting (1) and (2a) in (2b), the received signal vector q_m can be rewritten as,

$$\begin{aligned}\boldsymbol{q}_m = \rho\{&b_{(m-1)}\cos(\Delta_m + \Delta_1)\boldsymbol{g}^E \boldsymbol{g}_c^E + b_m \cos(\Delta_m + \Delta_2)\boldsymbol{g}^F \boldsymbol{g}_c^F \\ &+ b_{(m+1)}\cos(\Delta_m + \Delta_3)\boldsymbol{g}^L \boldsymbol{g}_c^L\} + \{b_{(m-1)}\sin(\Delta_m + \Delta_1)\boldsymbol{g}^E \boldsymbol{g}_s^E \\ &+ b_m \sin(\Delta_m + \Delta_2)\boldsymbol{g}^F \boldsymbol{g}_s^F + b_{(m+1)}\sin(\Delta_m + \Delta_3)\boldsymbol{g}^L \boldsymbol{g}_s^L\}\end{aligned} \qquad (3)$$

Where $\Delta_1 = \varphi$, $\Delta_2 = \varphi - dT_c 2\pi f$, $\Delta_3 = \varphi - CT_c 2\pi f$, $\Delta_m = 2\pi(m-1) fT_b$.

f can be expressed as $f = \pm f_b + \Delta f$, where i=0, 1, 2 ... Thus $\Delta_m = 2\pi(m-1)\Delta f T_b$. $\boldsymbol{g}^E$, $\boldsymbol{g}^F$, and $\boldsymbol{g}^L$ can be expressed as:

$$\boldsymbol{g}^E = [s(C-d+1)...s(C)0...0]^T \qquad (4a)$$

$$\boldsymbol{g}^F = [0...0s(1)...s(C)0...0]^T \qquad (4b)$$

$$\boldsymbol{g}^L = [0...0, s(1)...s(C-d)]^T \qquad (4c)$$

The length of the three vectors above are all $2C$. $\boldsymbol{g}^F$ contains a complete spread spectrum code of the user. According to the subspace structure, $\boldsymbol{g}_c^E$, $\boldsymbol{g}_s^E$, $\boldsymbol{g}_c^F$, $\boldsymbol{g}_s^F$, $\boldsymbol{g}_c^L$ and $\boldsymbol{g}_c^L$ can be expressed as,

$$\boldsymbol{g}_c^E = \mathrm{diag}[\cos(\mathrm{wT_c}),...,\cos(\mathrm{dwT_c}),0,...,0] \tag{4d}$$

$$\boldsymbol{g}_s^E = \mathrm{diag}[\sin(\omega \mathrm{T}_c),...,\sin(\mathrm{dwT}_c),0,...,0] \tag{4e}$$

$$\boldsymbol{g}_c^F = \mathrm{diag}[0,...,0,\cos(\mathrm{wT_c}),...,\cos(\mathrm{CwT_c}),0,...,0] \tag{4f}$$

$$\boldsymbol{g}_s^F = \mathrm{diag}[0,...,0,\sin(\mathrm{wT_c}),...,\sin(\mathrm{CwT_c}),0,...,0] \tag{4g}$$

$$\boldsymbol{g}_c^L = \mathrm{diag}[0,...,0,\cos(\mathrm{wT_c}),...,\cos((\mathrm{C-d})\,\mathrm{wT_c})] \tag{4h}$$

$$\boldsymbol{g}_s^L = \mathrm{diag}[0,...,0,\sin(\mathrm{wT_c}),...,\sin((\mathrm{C-d})\,\mathrm{wT_c})] \tag{4I}$$

Where the above six matrix are $2C$ by $2C$ diagonal matrix, and $\omega=2\pi f$. Equation (3) can be written as a subspace matrix,

$$\boldsymbol{q_m} = \boldsymbol{B_m G} + \boldsymbol{N_m} \tag{5}$$

Where B_m can be expressed as,

$$\begin{aligned}\boldsymbol{B_m} = [&\boldsymbol{b}_{(m-1)}\cos(\Delta_m+\Delta_1), \boldsymbol{b_m}\cos(\Delta_m+\Delta_2), \boldsymbol{b}_{m+1}\cos(\Delta_m+\Delta_3),\\ &\boldsymbol{b}_{(m-1)}\sin(\Delta_m+\Delta_1), \boldsymbol{b_m}\sin(\Delta_m+\Delta_2), \boldsymbol{b}_{(m+1)}\sin(\Delta_m+\Delta_3)]\end{aligned} \tag{6}$$

$$\boldsymbol{G} = \left[\rho \boldsymbol{g}_c^E \boldsymbol{g}^E, \rho \boldsymbol{g}_c^F \boldsymbol{g}^F, \rho \boldsymbol{g}_c^L \boldsymbol{g}^L, \rho \boldsymbol{g}_s^E \boldsymbol{g}^E, \rho \boldsymbol{g}_s^F \boldsymbol{g}^F, \rho \boldsymbol{g}_s^L \boldsymbol{g}^L\right] \tag{7}$$

Where N_m is a $2C$ by 1 Gaussian noise vector, the received DS-CDMA signal has been constructed as a Blind Source Separation (BSS) form as shown in (5). Thus, a SVD based method will be proposed to estimate the m-sequence for the modulated CDMA signal.

III. SVD BASED ESTIMATE OF M-SEQUENCE AND DATA

Assume the number of the sample data is M, and the sample covariance matrix of can be expressed as,

$$\boldsymbol{R}_x = \mathrm{E}\{\boldsymbol{q}_m^T \boldsymbol{q}_m\} = \lim_{M\to\infty} \frac{1}{\boldsymbol{M}} \sum_{m=1}^{M} \boldsymbol{q}_m{}^T \boldsymbol{q}_m \tag{8}$$

Inserting (5) into (8), we can get

$$\boldsymbol{R}_x(\mathrm{M}) = \frac{1}{\mathrm{M}}\sum_{\mathrm{m}=1}^{\mathrm{M}}(\boldsymbol{G}^T \boldsymbol{B}_\mathrm{m}{}^T \boldsymbol{B}_\mathrm{m}\boldsymbol{G}) = \boldsymbol{G}^T \frac{1}{\mathrm{M}}(\sum_{\mathrm{m}=1}^{\mathrm{M}} \boldsymbol{B}_\mathrm{m}^T \boldsymbol{B}_\mathrm{m})\boldsymbol{G} + \boldsymbol{I}_{2C} \tag{9}$$

Where $\boldsymbol{I}_M$ denotes a $2C$ by $2C$ unit matrix. Since the source data is statistically independent from its delay, we can have

$$E(b_i b_j) = \frac{1}{M}\sum_{m=1}^{M} b_{i+m} b_{j+m} = \begin{cases} 0, i \neq j \\ 1, i = j \end{cases}, i, j \in \{0,1,2\}. \tag{10a}$$

It is obvious from (10a) that different delay vectors in B is irrelevant. The analysis of the relationship between the delay vectors will be shown as follows. If the number of sampling data is M, the correlative quantity between the same delay vectors can be calculated as:

$$\begin{aligned}&|E(b_i\cos(\Delta_m+\Delta_i)b_i\sin(\Delta_m+\Delta_i))|\\ &=|\frac{1}{2M}\sum_{i=1}^{M}\sin(4\pi i\Delta f / f_b + 2\Delta_i)|\\ &\approx|\frac{1}{2MT_b}\int_{i=1}^{MT_b}\sin(4\pi\Delta f / f_b + 2\Delta_i)dt|\\ &=|\frac{1}{2M}\frac{2\sin(2\pi\Delta f MT_b)\sin(2\pi\Delta f MT + 2\Delta_i)}{2\pi\Delta f T_b}|\\ &<\frac{f_b}{2M\pi\Delta f}, i = 1,2,3\end{aligned} \tag{10b}$$

Also the correlation coefficient of B_m vector can be calculated as:

$$\begin{aligned}&E((b_i\cos(\Delta_m+\Delta_i))^2)\\ &=\frac{1}{2}+\frac{1}{2M}\sum_{i=1}^{M}\cos(4\pi i f / f_b + 2\Delta_i)\\ &\approx\frac{1}{2}(1+\frac{f_b\sin(2\pi\Delta f MT_b)\cos(2\pi\Delta f MT + 2\Delta_i)}{M\pi\Delta f}), i = 1,2,3\end{aligned} \tag{10c}$$

$$\begin{aligned}&E((b_i\sin(\Delta_m+\Delta_i))^2)\\ &=\frac{1}{2}-\frac{1}{2M}\sum_{i=1}^{M}\cos(4\pi i f / f_b + 2\Delta_i)\\ &\approx\frac{1}{2}(1-\frac{f_b\sin(2\pi\Delta f MT_b)\cos(2\pi\Delta f MT + 2\Delta_i)}{M\pi\Delta f}), i = 1,2,3\end{aligned} \tag{10d}$$

When $\Delta f > \frac{50 f_b}{M\pi}$, the result of (10b) is smaller than 0.005 but the results of (10c) and (10d) are approximate equal to 0.5. So when $\Delta f > \frac{100 f_b}{M\pi}$ and $\mathrm{E}(B^\mathrm{T}{}_\mathrm{m} B_\mathrm{m})\approx 1/2\mathrm{I}$, (9) can also be expressed as,

$$\boldsymbol{R}_x(M) = \frac{1}{2}\boldsymbol{G}^T\boldsymbol{G} + \boldsymbol{I}_M \tag{11}$$

By SVD process of $\boldsymbol{R}_x(M)$, it can be expressed as,

$$\boldsymbol{R_x} = \boldsymbol{U}\lambda\boldsymbol{U}^T \tag{12}$$

Where the U is a 2C by 2C orthogonal matrix, λ=diag[λ1, λ2, ...,λ2C] is a 2C by 2C diagonal matrix , which has its eigenvalues in decreasing order. The SVD process can accurately estimate the received signals [8]. If the column vectors gcFgF, gsFgF and all the other columns of G are orthogonal. The first and second signal subspace (U1, U2) are corresponding to {gcFgF, gsFgF}. According to (5b), the

two column vectors gcFgF and gsFgF, which are modulated by sine and cosine wave respectively, contain the complete spread spectrum sequence.

By taking a close look at matrix G, it can be found that gcFgF and gsFgF are orthogonal with the other four columns. Analysis of orthogonal coefficients between gcFgF and gsFgF is shown as follows:

$$\begin{aligned}(\boldsymbol{g}_c^F \boldsymbol{g}^F)^T \boldsymbol{g}_s^F \boldsymbol{g}^F &= \frac{1}{2}\sum_{m=1}^{C}\sin(2m\omega T_c) \\ &\approx \frac{1}{4T_c}\int_0^{CT_c}\sin(2\omega t)dt \\ &= \frac{1-\cos(2\omega T_b)}{\omega T_c} = \frac{2\sin^2(2\pi\Delta f T_b)}{2\pi f T_c}\end{aligned} \tag{13a}$$

Similarly we can get:

$$\begin{aligned}|\boldsymbol{g}_c^F \boldsymbol{g}^F|^2 &= \frac{1}{2}(C-\sum_{m=1}^{C}\cos(2m\omega T_c+2\Delta_1)) \\ &= \frac{1}{2}(C-\frac{\sin(2\pi\Delta f T_b)}{2\pi f T_c})\end{aligned} \tag{13b}$$

$$\begin{aligned}|\boldsymbol{g}_s^F \boldsymbol{g}^F|^2 &= \frac{1}{2}(C+\sum_{m=1}^{C}\cos(2m\omega T_c+2\Delta_1)) \\ &= \frac{1}{2}(C+\frac{\sin(2\pi\Delta f T_b)}{2\pi f T_c})\end{aligned} \tag{13c}$$

We define the correlation coefficient between gcgF and gsgF is ρ_g ,

$$\rho_g = \frac{(\boldsymbol{g}_c^F \boldsymbol{g}^F)^T \boldsymbol{g}_s^F \boldsymbol{g}^F}{|\boldsymbol{g}_c^F \boldsymbol{g}^F|^2 + |\boldsymbol{g}_s^F \boldsymbol{g}^F|^2} = \frac{\sin^2(2\pi\Delta f T_b)}{\pi f T_b} \tag{14}$$

According to the property of SVD [9], we define ρg<0.05=ρ0 and gcFgF and gsFgF could be considered mutually orthogonal. Considering the value range of Δf and Nyquist sampling theorem, the range of frequency of gcFgF and gsFgF can be expressed as:

$$f_b + \frac{100 f_b}{M\pi} < |f| < i f_b + \frac{\arcsin(\sqrt{\rho_0 \pi f / f_b})}{2\pi} f_b \qquad i = 0,1,2,...(C-1)/2 \tag{15}$$

From (15), we can find that if the observation M becomes bigger, the range of f that can be used to separate $g_c^F g^F$ and $g_s^F g^F$ will be larger. For some special carrier frequency offset, by SVD process of the received sample covariance matrix, the m-sequence with frequency offset can be estimated. In order to recover the m-sequence, a phase lock loop (PLL) is used to extract g^F from the subspace U_1 and U_2.

In the process of PLL, the frequency ω can also be obtained. In order to estimate the source data, the received signal vector q_m will be projected onto $U_1=g_c g^F$ and $U_2=g_s g^F$ respectively,

$$b_c(m) = \boldsymbol{q}_m(\boldsymbol{U}_1) = b_m \cos(2\pi\Delta f m T_b + \Delta_2) \tag{16}$$

$$b_s(m) = \boldsymbol{q}_m(\boldsymbol{U}_2) = b_m \sin(2\pi\Delta f m T_b + \Delta_2) \tag{17}$$

$b_c(m)$ and $b_s(m)$ are the estimated source data with frequency offset. PLL can also be used to recover the source data from $b_c(m)$ and $b_s(m)$.

IV. SVD AND PLL BASED M-SEQUENCE BLIND ESTIMATION

Fig. 1 illustrates the block diagram of the SVD and PLL based method to blind estimation of m-sequence in DS-CDMA signals. The received signal is first demodulated by the local frequency and then sent to a low-pass filter. The filtered signal will be constructed as subspace vector q_m, and the sample covariance matrix of q_m will be calculated. With the SVD process, one can get signal subspace eigenvector $U_1=g_c g^F$, which is the m-sequence with carrier frequency offset modulation. If the frequency offset f does not meet the condition in (15), the local frequency, f_2 has to be adjusted according to the proposed objection function. Since the frequency offset modulated m-sequence and data can be estimated by the process of SVD, the PLL is used to extract the m-sequence and data from the modulated signals.

V. SIMULATION RESULTS

In this section, simulations results are presented to illustrate the performance of the proposed method. BPSK modulation is used, and T_b=1/100, C=63, g(x) = X^6+X+1, sample data length M=400. SNR is 5dB.

Fig. 2 shows the first 12 eigenvalues of the singular value decomposition when the frequency f is 106Hz. It shows that there are 6 larger eigenvalues, as is consistent with the model predictions in (5).

Fig. 3(a) and Fig. 3(b) show the estimated frequency modulated m-sequence $\boldsymbol{g}_s^F \boldsymbol{g}^F$ and $\boldsymbol{g}_c^F \boldsymbol{g}^F$. Fig. 3 (c) shows the estimated m-sequence $\boldsymbol{g}^F$ obtained by a PLL. Compared with the original m-sequence, it can be obtained that the estimated m-sequence is an accurate estimation.

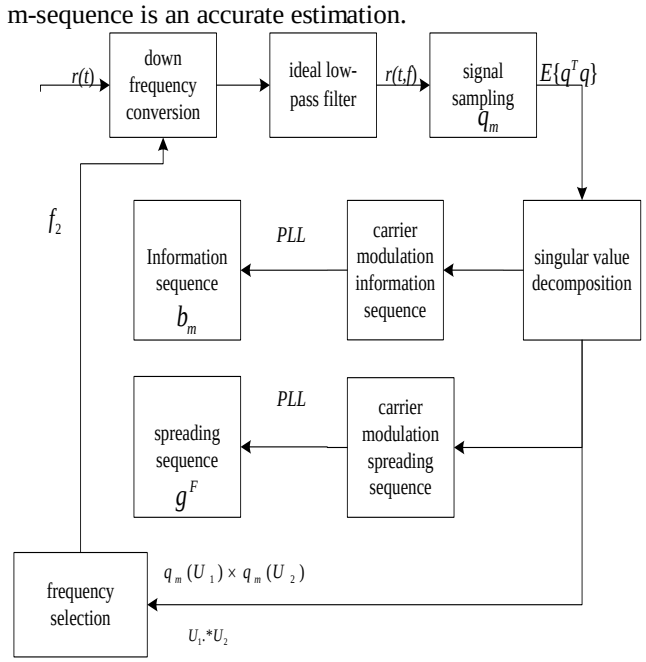

Figure 1. Blind estimation of m-sequence based on SVD and PLL

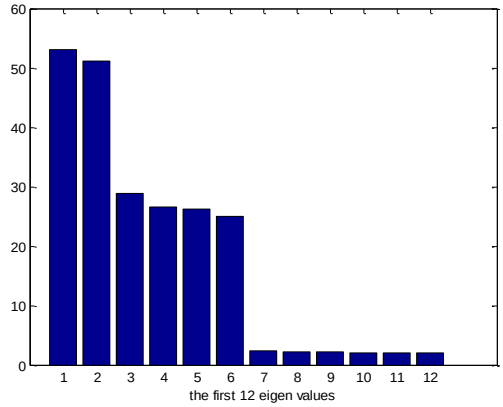

Figure 2. The simulated eigenvalue of SVD.

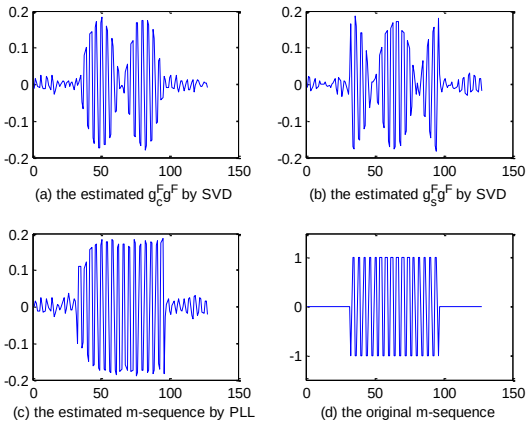

Figure 3. The estimated m-sequence compared to original m-sequence.

Fig. 4(a) and Fig. 4(b) show the estimated frequency modulated signals, $b_m \sin(\Delta_m + \Delta_2)$ and $b_m \cos(\Delta_m + \Delta_2)$. Fig. 4(c) shows the estimated signal b_m, which is also obtained by a PLL. Compared with the original signals, it can be concluded that this estimation has a good performance.

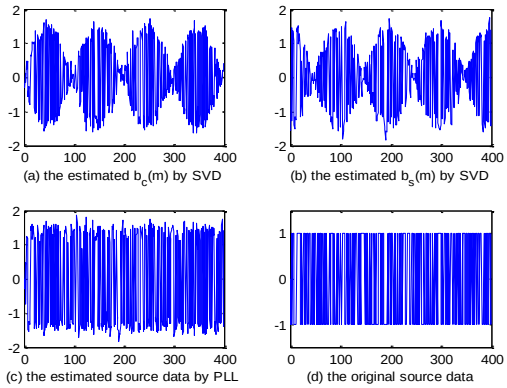

Figure 4. The estimated signal data compared with the original signal data.

VI. CONCLUSION

An SVD and PLL based method to blindly estimate the spread spectrum sequence in modulated DS-CDMA signals without the information of carrier frequency has been proposed and validated. By performing the SVD of the received subspace model, eigenvalues and eigenvectors can be determined. According to the changes of eigenvalues, the appropriate frequency can be found for blind estimation of the m-sequence. With PLL, the data and the carrier frequency offset can also be estimated efficiently.

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