A DS-CDMA Interference Cancellation Technique for Body Area Networks

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***Abstract*—A low complexity Parallel Interference Cancellation (PIC) technique that is applicable to body area networks is presented. Using Direct-Sequence Code Division Multiple Access (DS-CDMA), the technique aims at suppressing interference caused by rapid changes in relative sensors position due to body parts motion as well as interference from adjacent BANs. Interference signal is estimated using a relationship between cyclic correlation of the received signal and interferer code without requiring any knowledge of the channel condition. The codes for the multi-sensor DS-CDMA communication are constructed by using a set of m-sequences. The cyclic correlation is performed using Fast Walsh-Hadamard Transform (FWHT) which exhibits low computational complexity. In addition to low complexity, the proposed technique does not require complicated channel condition estimation and has no convergence issues. The uncoded Bit Error Rate (BER) performance of the proposed interference cancellation over a body-surface channel is calculated and compared with the conventional scheme.**

***Index Terms*— Body area networks, Interference cancellation, Direct-sequence code division multiple access, Fast Walsh-Hadamard transformation**

I. Introduction

Body Area Networks (BANs) consist of multiple wearable (or implantable) sensors that have two-way communication with a controller node that could be either worn or located in the vicinity of the body. With its vast set of intriguing applications, it is conceivable that this technology will be used by many people as part of their daily life. Therefore, interference from multiple BANs could create a serious problem on the reliability of the network operation. Similarly, rapid body parts motion is another reason that could greatly affect the quality of a link between a sensor node and the controller. If a person is wearing multiple sensors, it is possible that link variation of one node creates severe temporary interference on other links.

Spread Spectrum technology (e.g., DS-CDMA) could be an appropriate candidate to provide a robust approach to mitigate such interference. In a Direct-Sequence Code Division Multiple Access (DS-CDMA) system, the desired signal of a node is detected and decoded against interference by using the de-spreading feature of a code sequence corresponding to that node. All other nodes in this system are simply considered as Multi-Access Interference (MAI). The performance could be degraded when total interference increases due to the increasing number of nodes. Even when the number of nodes is not large; the received signal from nearby or high-powered interfering nodes may be too high; resulting in overwhelming the signal from the desired node. This is called the near-far effect which causes performance degradation in detecting or decoding the desired signal. To minimize the near-far effect, an efficient power control mechanism is usually required. However, implementing such mechanisms could be very complicated in fast fading environments such as communication links in Body Area Networks. Recent studies on the variability of BAN channels point to severe signal fluctuations due to different body postures and/or body parts motion [13-15]. These link variations create situations very similar to the near-far effect mentioned earlier. And in this case, controlling the transmit power based on the channel conditions in order to keep the received signal at a desired level (i.e., power control) is virtually impossible. Therefore, using interference cancellation techniques could be very beneficial for body area networks [1].

There are two main problems with the previously proposed interference cancellation techniques especially when it comes to their application for BAN. First is the high complexity of the receiver which makes the implementation of such techniques impractical unless the number of nodes is very small [2, 3]. Complexity is specially a critical issue for nodes in body area networks. As they mainly rely on battery power, prolonging the lifetime of these nodes are of prime importance. Second disadvantage is that some interference cancellation schemes require perfect knowledge of the channel condition (such as attenuation, phase, and delay) between each of the interferers and the receiver [4-6]. Obtaining accurate estimates of the channel condition is extremely difficult for body area networks.

In [12] authors have proposed using multi-sensor detectors to perform interference cancellation for BAN. However, in addition to the remaining complexity issues, only simple Additive White Gaussian Noise (AWGN) and Rayleigh fading channels were considered. In our previous publication, we presented a low complexity interference cancellation technique that does not require channel estimation [11]. The low complexity of this technique makes it an appropriate candidate for low power applications. In this paper, we demonstrate the advantage of this technique for body area networks by evaluating its performance using a standard-based body-surface channel model.

The remainder of this paper is organized as follows. Section II describes the body-surface channel model. Section III provides an outline of the DS-CDMA system. The proposed method of interference estimation and cancellation is presented in section IV. Section V describes the Fast Walsh-Hadamard Transformation (FWHT) for implementation of interference estimation. Finally, simulation results and conclusions are discussed in Sections VI and VII respectively.

II. THE BODY-SURFACE CHANNEL MODEL

Several statistical channel models for body area networks have been considered in the IEEE 802.15.6 standard group. These models are applicable to different usage scenarios in various frequency bands [9]. This paper is mainly focused on body-surface applications using Ultra-Wide Band (UWB) frequency transmission. The multipath impulse response of the body-surface channel (i.e., also referred to as on-body) can be modeled by:

$$h(t,\xi)=\sum_{l=0}^{L-1}\alpha_l\delta(t-\xi_l)e^{j\theta_l} \tag{1}$$

where α_l, θ_l and ξ_l are amplitude, phase and time delay corresponding to l^{th} arrival path. L is the number of arrival paths. The parameters of the body-surface channel model have the following properties [9, 10]. The amplitude α_l follows an exponential decay Γ with a Rician factor γ_0,

$$20\log_{10}|\alpha_l|=\begin{cases}0, & l=0\\ \gamma_0+10\log_{10}\left(e^{-\xi_l/\Gamma}\right)+S, & l\neq 0\end{cases} \tag{2}$$

where S is a zero-mean Gaussian distribution with a standard deviation of σ_s. The phase θ_l is uniformly distributed over $[0,2\pi)$. The path arrival time ξ_l has a Poisson distribution as:

$$p(\xi_l\mid\xi_{l-1})=\lambda e^{-\lambda(\xi_l-\xi_{l-1})} \tag{3}$$

where λ is path arrival rate. The number of arrival path follows a Poisson distribution with the mean $\overline{L}$.

$$p(L)=\frac{\overline{L}^{L}\exp\left(-\overline{L}\right)}{L!} \tag{4}$$

III. SYSTEM DESCRIPTION

In a multi-sensor DS-CDMA BAN, K sensors could be simultaneously transmitting in the same frequency band. The k^{th} sensor is assigned a spreading code sequence, $c_n^k, n=0,1,\dots,N-1,$ with spreading factor N. The transmitted baseband signal from the k^{th} sensor is given by

$$S^k(t)=\sum_i d_i^k C^k(t-iT_s) \tag{5}$$

where d_i^k is the i^{th} symbol transmitted by sensor k, $C^k(t)=\sum_{n=0}^{N-1}c_n^k g(t-nT_c)$, T_c is the chip duration, T_s is a symbol duration ($T_s=N*T_c$) and $g(t)$ is a unit rectangular pulse centered at 0 with width T_c. Thus, for a specific receiver, the received signal from the k^{th} sensor during the i^{th} symbol period is

$$r_i^k(t)=s_i^k d_i^k C^k(t)\otimes h^k(t),\ \text{ for } iT_s\le t<(i+1)T_s \tag{6}$$

where s_i^k is the received signal strength, $h^k(t)$ is the channel impulse response from the k^{th} transmitter to the given receiver and $\otimes$ denotes the convolution operator. The total received baseband signal from all sensors can be described by Eq. (7)

$$r_i(t)=\sum_{k=1}^{K}r_i^k(t-\tau_k)+n(t) \tag{7}$$

where τ_k is the time delay with respect to a reference time and $n(t)$ is the AWGN. For simplicity, we ignore the noise at this time and assume that the receiver is interested in decoding the signal from sensor 1 and $\tau_1=0$. To extract the 1^{st} sensor's signal from the total received signal, one can use cross-correlation between the received baseband signal and the code sequence of sensor 1 as shown in Eq. (8)

$$\begin{aligned}S_i^1&=C^1(t)\bullet r_i(t)=s_i^1 d_i^1\rho^{1,1}(t)\otimes h^1(t)\\&+\sum_{k=2}^{K}s_i^k d_i^k\rho^{1,k}(t,\tau_k)\otimes h^k(t-\tau_k)\end{aligned} \tag{8}$$

where $\rho^{1,k}(t,\tau_k)=C^1(t)\bullet C^k(t-\tau_k)$ is the cross-correlation function of the 1^{st} code sequence with the k^{th} code sequence and $\bullet$ denotes the cross-correlation operator.

The first term on the right side of equation (8) represents the desired signal and the second term is the interference. Now, if the code sequence of sensor 1 is orthogonal to the spreading code sequences of other sensors with any time-shift (i.e. $\rho^{1,k}(t,\tau_k)=0$ when $k\neq 1$), then the effect of interference from sensors 2 to K in Eq. (7) becomes zero. But, in reality, $\rho^{1,k}(t,\tau_k)\neq 0$ when $k\neq 1$, and the interference will be a non-zero component. If $s_i^k>>s_i^1$ for any k, then this interference becomes significant. In general, even compatible levels of interference cause degradation in performance and reduction in system capacity when there is no interference cancellation.

Here, we consider using m-sequences as the spreading codes. The auto-correlation function of an m-sequence, with m being an integer, is close to the desirable impulse function.

Furthermore, the cross-correlation between any two m-sequences is very small as they are almost mutually orthogonal. Therefore, they are an ideal candidate to implement an interference cancellation scheme. Based on this orthogonality property, interference can be estimated using the cross-correlation of the received signal and the m-sequence of the interferer. In [7-8], an interesting relationship between m-sequences and the Hadamard Transform has been discussed. This relationship allows for fast computation of the cross-correlation (as well as its inverse) of the received signal and the m-sequence of the interferer. This is referred to as Fast Walsh-Hadamard Transform (FWHT). In the following two sections, we will describe how to use correlation to estimate and suppress interference; and then, how to use the low-complexity FWHT to implement the interference cancellation technique.

IV. Interference Estimation and Cancellation

Let the discrete received baseband signal at i^{th} symbol duration be $\underline{r}^i = [r^i(1), r^i(2), r^i(3), \ldots r^i(mN)]^t$, where m is the number of samples per chip and the superscript t represents the transpose operation. This vector can also be reshaped to an $(m \times N)$ matrix where the j^{th} row vector of the matrix is

$$\underline{r}_j^i = [r^i(j), r^i(m+j), r^i(2m+j), \ldots r^i((N-1)m+j)]^t, \quad j = 1,2,\ldots,m.$$

Now, consider the interferer to be the k^{th} sensor with code sequence $\underline{c}^k = [c_0^k, c_1^k, \ldots c_{N-1}^k]$. Each of the row vectors is correlated with the code sequence as below.

$$\hat{\underline{r}}_j^{i,k} = \underline{c}^k \bullet \underline{r}_j^i \tag{9}$$

The correlation output can be well-approximated by a cyclic cross-correlation since the m-sequence has a good autocorrelation property. The cyclic cross-correlation can be written in matrix form as:

$$\hat{\underline{r}}_j^{i,k} = M_N^k \cdot \underline{r}_j^i \tag{10}$$

where M_N^k is a circulant m-sequence matrix for the given user code $\underline{c}^k$ as shown below.

$$M_N^k = \begin{bmatrix} c_0^k & c_1^k & \cdots & c_{N-2}^k & c_{N-1}^k \\ c_{N-1}^k & c_0^k & \cdots & c_{N-3}^k & c_{N-2}^k \\ \vdots & \vdots & \ddots & \vdots & \vdots \\ c_2^k & c_3^k & \cdots & c_0^k & c_1^k \\ c_1^k & c_2^k & \cdots & c_{N-1}^k & c_0^k \end{bmatrix} \tag{11}$$

The correlation output contains two components, the signal from sensor k and a noise-like component that is due to all other sensors. As the value of this second component is usually below the signal of sensor k, one can define a threshold η to remove it from $\hat{\underline{r}}_j^{i,k}$ as indicated in Eq. (12)

$$\hat{\underline{r}}_j^{i,k}(l) = \begin{cases} \hat{\underline{r}}_j^{i,k}(l) & \text{if } \hat{\underline{r}}_j^{i,k}(l) \ge r_p / \eta \\ 0 & \text{elsewhere} \end{cases}, \text{ for } l = 0,1,\ldots,L-1, \tag{12}$$

where $r_p = \max_l \left| \hat{\underline{r}}_j^{i,k}(l) \right|$.

Moreover, the inverse of the m-sequence matrix can be used to determine the interference which is due to the k^{th} sensor as specified in Eq. (13).

$$\tilde{\underline{r}}_j^{i,k} = (M_N^k)^{-1} \cdot \hat{\underline{r}}_j^{i,k} \tag{13}$$

$\tilde{\underline{r}}_j^{i,k}$ is the estimated signal of sensor k from the received signal $\underline{r}_j^i$ at the j^{th} sample of a chip. By doing the same procedure for $j=1\ldots,m$, one can obtain an $(N \times m)$ matrix given by:

$$\underline{R}^{i,k} = [\tilde{\underline{r}}_1^{i,k} \quad \tilde{\underline{r}}_2^{i,k} \quad \cdots \quad \tilde{\underline{r}}_m^{i,k}] \tag{14}$$

Each vector $\tilde{\underline{r}}_j^{i,k}$ is an estimate of the interference signal due to sensor k at the j^{th} sample of each chip during i^{th} symbol duration. Knowing this estimate, the undesired k^{th} sensor component can be removed from the total received signal (i.e., $\underline{r}^i$). Similarly, all undesired signals can be removed using the same process.

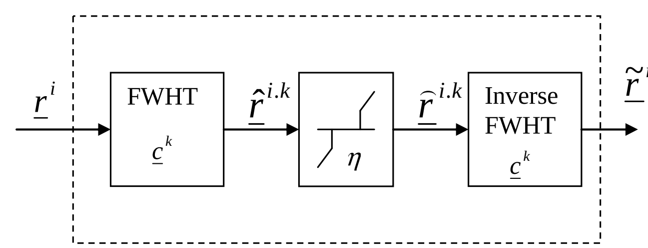

Fig. 1: Block diagram of Interference Estimation (I.E.) using FWHT

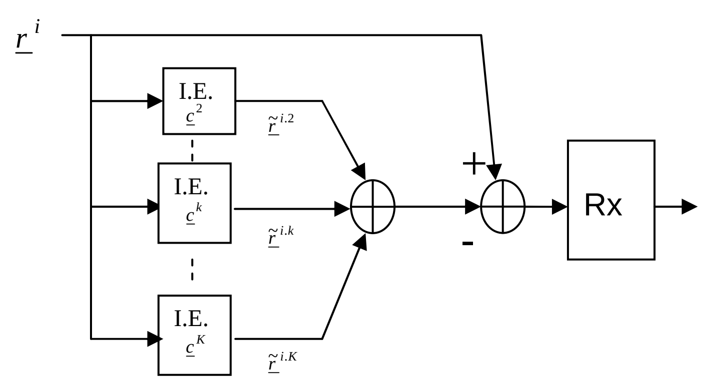

Fig. 2: Block diagram of proposed receiver with interference cancellation

The block diagram of the interference estimation and the receiver with proposed interference cancellation are shown in Figs. 1 and 2 respectively. Note that the interference estimation process requires calculation of the inverse of m-sequence matrix as appears in Eq. (13). The matrix inverse operation could be computationally expensive if the size of the matrix is large. To reduce the computational complexity, we propose using the Fast Walsh-Hadamard Transform. This will be further described in the next section.

V. Cyclic correlation computation using Fast Walsh-Hadarmard transform

Equations (10) and (13) involve an operation with the m-sequence matrix and its inverse, respectively. This section describes a low complexity algorithm for the m-sequence matrix operation through FWHT.

The relationship between the m-sequence matrix in Eq. (11) and the associated Hadamard matrix can be described by the following equation [8].

$$M_L^k = V_R P_w H_{L+1} P_s V_I \tag{15}$$

V_R is a ($N \times N+1$) matrix that reduces the size of the processed vector from $N+1$ to N; similarly, V_I is a ($N+1 \times N$) matrix that increases the size of the processed vector from N to $N+1$. Both matrices require no special computation and are basically extensions of the identity matrix I_N given by:

$$V_R = \begin{bmatrix} \underline{0} & I_N \end{bmatrix} \qquad V_I = \begin{bmatrix} \underline{0} \\ I_N \end{bmatrix} \tag{16}$$

P_w and P_s are permutation matrices used for descrambling a vector. Therefore, their computational complexity is negligible. H_{L+1} is a Walsh-Hadamard matrix with the size ($N+1 \times N+1$). A Walsh-Hadamard matrix of a certain size can be generated recursively by the following methodology:

$$H_1 = [1] \qquad H_i = \begin{bmatrix} H_{i-1} & H_{i-1} \\ H_{i-1} & -H_{i-1} \end{bmatrix} \tag{17}$$

Note that the dimension of these matrices is of the form 2^{i-1} where i is an integer.

The correlation operation in Eq. (10) can be implemented by using Eq. (15) which only involves simple permutations and the FWHT. FWHT is similar to the Fast Fourier Transform, but only requires "addition" and "subtraction" operations. More details on FWHT can be found in [8]. The inverse of the m-sequence matrix in Eq. (13) is therefore obtained by the following equation.

$$\left(M_L^k\right)^{-1} = V_I^t P_s^t \left(H_{L+1}\right)^{-1} P_w^t V_R^t \tag{18}$$

The operation in Eq. (13) can also be performed using similar permutations and inverse FWHT. Fortunately, the inverse FWHT is similar to FWHT except for a constant factor of size $N+1$. In terms of computational complexity, the proposed PIC does not require any multiplication or division operations. The proposed technique only requires $m(N+1)(1+2\cdot\log_2(N+1))$ complex additions or subtractions per bit per interfering sensor, where m is the number of samples per chip and N is the number of chips (i.e., the length of the spreading code). In addition, the complexity grows linearly with the number of sensor as this technique can be implemented in parallel. Due to its complexity advantages, the FWHT provides an efficient and practical way to estimate the interference signal for interference cancellation.

VI. Simulation results

In this section, we present simulation results which are based on the proposed interference cancellation technique discussed earlier. The simulation scenario emphasizes on the situations, where there is significant interference power from a nearby interferer node. Parameters of the on-body channel model based on measurements obtained in an anechoic chamber are $\gamma_0 = -0.48dB$, $\Gamma = 8.88ns$, $\sigma_s = 2.87dB$, $\lambda^{-1} = 6.82ns$, and $\bar{L} = 1.5$ [10]. Other simulation parameters are assumed to be as follows: N=31 (i.e., spreading code length is 31) and m=2 (i.e., 2 samples per chip). Therefore, the processing gain will be around 15 dB. To observe the effectiveness of our interference cancellation technique, two scenarios are considered for the signal-to-interference ratio i.e., -10 dB and -20 dB.

Another parameter is the threshold η in Eq. (12), which is usually difficult to be optimized. In general, the lower η is chosen for low SNR or low SIR environments with a single dominant path or when it is desired to remove interference from the most dominant path in multipath-rich environments. To remove the interference from more number of paths, a higher η may be used at the risk of increased inaccuracy in the estimation process.

Here, we have compared the receiver's uncoded BER performance of our scheme with a conventional PIC technique such as one discussed in [6]. Figure 3 shows the resulting BER performance for SIR=-10 dB and η=1.5. When the conventional PIC has perfect knowledge of the channel conditions including interference power level and channel impulse response, its BER performance is better than the proposed PIC. However, as stated previously, for BAN, these channel conditions are extremely difficult to estimate. As seen in Fig. 3, the proposed PIC outperforms the conventional PIC in the BER performance if the conventional PIC only has knowledge of the interference level (and no knowledge of channel impulse response). As expected, the BER performance without using any interference cancellation technique is worst.

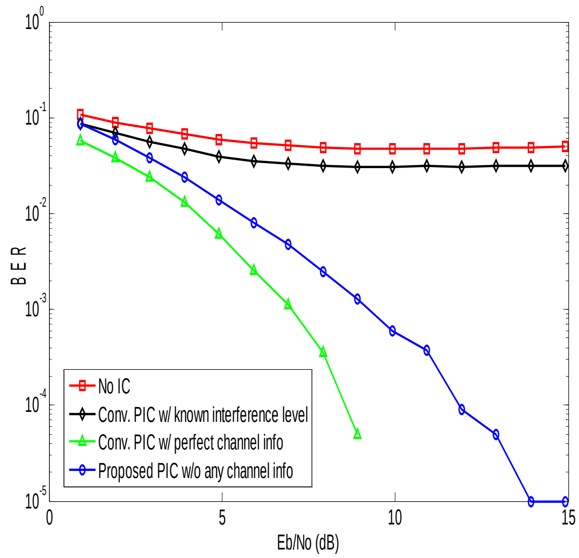

Fig. 3: Uncoded BER at SIR= -10dB over a body surface channel

Figure 4 shows the uncoded BER performance while SIR=-20 dB. For this case, the threshold has been chosen higher (i.e., η=5) in order to accurately estimate multipath interference. Similar to the previous case, the proposed PIC has higher BER compared to the conventional PIC that has perfect knowledge of channel conditions. However, the proposed PIC considerably outperforms the conventional PIC that only has knowledge of interference level (but no knowledge of channel impulse response). In other word, multipath interference significantly degrades the performance of the conventional PIC if it does not have any knowledge of channel impulse response. On the other hand, our proposed PIC removes part of the multipath interference while channel conditions are unknown. If the channel conditions are also known, then the performance of the proposed PIC would match the conventional one in these cases.

VII. CONCLUSIONS

In this paper, we have proposed a low complexity parallel interference cancellation technique suitable for body area networks based on direct sequence code division multiple access. The proposed method does not require any knowledge of the interferer signal power level and its channel condition at the receiver. This property along with low implementation complexity makes this scheme suitable for fast fading environments such as human body-surface where power control cannot be performed efficiently. A performance comparison between a conventional receiver (with and without interference cancellation) and a receiver with the proposed interference cancellation has been presented. The proposed technique demonstrates considerable improvement in BER performance. Authors believe that the extension of this technique will be very effective in combating interference especially for multiple BAN scenarios. Further results on the performance will be presented in future publications.

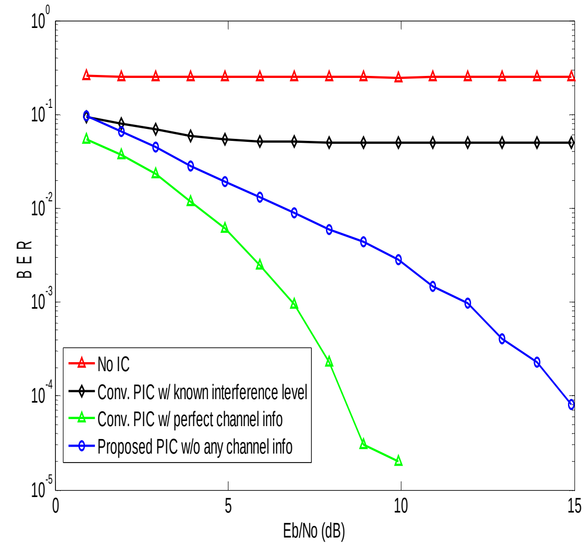

Fig. 4: Uncoded BER at SIR= -20dB over a body surface channel

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