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Toward Digital UWB Radios: Part II – A System Design to Increase Data Throughput for a Frequency Domain UWB Receiver

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ABSTRACT:

Our UWB receiver, which is based on the frequency domain approach, samples spectral components of the received signal at the pulse repetition rate, and process the signal in the frequency domain [1]. Our receiver effectively increases the sampling rate through signal processing. In this paper, we investigate efficient synchronization and equalization schemes to mitigate intersymbol interference, which is critical for high data rate applications. The proposed schemes improve the performance over a conventional analog correlator type receiver by 1 to 4dB.

1. INTRODUCTION

We investigated a new UWB receiver architecture based on the frequency domain processing [1]. Our receiver extracts spectral components of the received signal and increases the effective sampling rate through digital signal processing. Our receiver eliminates the need for a high speed analog-to-digital converter (ADC) and can employ multiple narrowband LNAs (Low Noise Amplifiers), which makes our receiver more CMOS friendly. One drawback of the receiver is that it resolves multipaths in the time domain first and then obtains a template in the frequency domain. The process requires translation of the frequency domain signal into the time domain using inverse fast Fourier transform (IFFT) and the reverse operation using fast Fourier transform (FFT). The translations and subsequent time-domain operations are complex in hardware to result in large power dissipation.

This paper addresses the shortcoming, and the entire signal processing including synchronization, channel estimation, matched filtering, and equalization are performed in the frequency domain. So it eliminates the need for an IFFT block and a FFT block to reduce the hardware complexity of the proposed UWB receiver. The front-end circuit structure of our receiver employing 1-bit ADCs is described in the companion paper [2]. This paper, Part II, focuses on the baseband signal processing in the frequency domain to mitigate intersymbol interference (ISI), which is critical for high data rate applications.

We briefly explain the ISI problem for UWB signals in Section 2. Section 3 describes the proposed digital signal processing schemes to mitigate ISI, and Section 4 presents the simulation results and observations. Section 5 concludes the paper.

2. BACKGROUND

High data rate requires a short pulse repetition interval (PRI), but it causes ISI due to the long channel delay spread of UWB signals. It is observed in [3] that the delay spread of UWB signals may last as long as 250 ns, which results in significant ISI for high data rate.

The channel response $h(t)$ is expressed as the channel impulse response $c(t)$ and the impulse response of the transmitter's pulse shaping filter $g(t)$ and is given in (1)

$$h(t) = c(t) * g(t), \tag{1}$$

where the operator * represents the convolution operation. Therefore, the received signal for an α_{th} single impulse without ISI is expressed as

$$r_\alpha(t) = I_\alpha h(t) + \eta(t), \qquad \alpha = 1, 2, 3, \ldots \tag{2}$$

where I_α is either 1 or –1 for BPSK (Binary Phase Shift Keying) and $\eta(t)$ the additive white Gaussian noise (AWGN). However, as shown in Figure 1, signals are overlapped due to ISI, and the overlapped signals are expressed as in (3).

$$r(t) = \sum_\alpha I_\alpha h(t - \alpha PRI) + \eta(t) \tag{3}$$

Suppose that we observe each pulse only during the period of the time window as indicated in Figure 1. Then, the time domain expression in (3) is equivalent to (4) in the frequency domain within a time window.

$$R = \sum_{\alpha} I_{\alpha} H_{\alpha} + \Gamma \qquad \alpha = 1, 2, 3, \dots \qquad (4)$$

where $H_{\alpha} = \mathrm{F}\{h(t - \alpha PRI)|_{0 \le t \le T_{window}}\}$ is the frequency domain channel response, and Γ is the expression of the noise in the frequency domain.

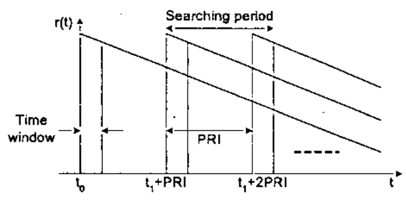

Figure 1: Channel condition under ISI

3. PROPOSED BASEBAND SIGNAL PROCESSING

The most critical baseband signal processing for high data rate UWB applications is to mitigate the ISI effect. The processing includes synchronization, channel estimation, matched filter operation, and equalization. The objective of our research is to perform the operations solely in the frequency domain, so that our receiver can further reduce the hardware complexity as well as power dissipation compared with our previous receiver design in [1].

First, we propose a synchronization operation in the frequency domain followed by a channel estimation operation. The results of the channel estimation are used to configure a frequency domain matched filter and a frequency domain equalizer.

A. Synchronization

For an impulse based UWB receiver based on an analog correlation, the synchronization should be precise in time to capture narrow pulses. However, the synchronization method must address a slightly different issue for a frequency domain receiver, as explained below.

Suppose that one pulse sent by the transmitter constitutes one symbol. A received UWB signal within a time window composed of the current symbol overlapped with multiple previous symbols. If a receiver can resolve each symbol, it is possible to accumulate the signal energy from each symbol to maximize the recovered signal energy. Thus, the criterion adopted for our synchronization scheme is to find the position of a time window within a pulse repetition interval (PRI) with the greatest signal energy.

A packet contains a known preamble sequence, which is used to perform synchronization and channel estimation. *We assume that the autocorrelation of the preamble sequence is a delta function,* so that any shifted version of the preamble sequence results in low (ideally zero) autocorrelation. This assumption is usually valid and allows parallel search of the preamble sequence.

Suppose that the length of the preamble sequence is m. Then, there are m possible circular shifted sequences. Let CS_i be the sequence obtained through the circular shift-right operation of the original preamble sequence by i positions. The proposed approach for synchronization is as follows. We extract the spectral components of the received signal $r(t)$ and correlate them with each individual sequence CS_i, $i = 0, 1, \dots m-1$. The correlation process is illustrated in Figure 2. Note that the length of the preamble sequence is five in the figure.

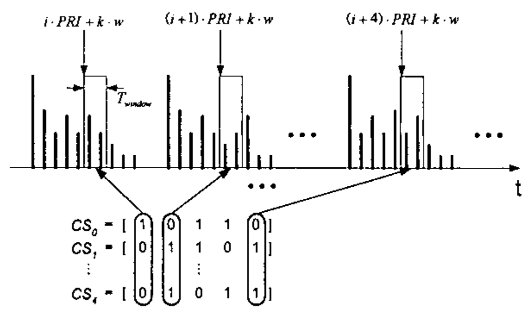

Figure 2: Correlation of the received signal with CS_i

Next, the energy of m correlated values is computed and recorded. Note that the correlation operation in the frequency domain is multiplication. Then, we advance the time by w called time step and repeat the same process. After we sweep the entire period of the preamble sequence, the time position which yields the maximum energy is the tuned time. We describe the process below.

Suppose that we start the synchronization operation at an arbitrary reference time T_{ref}, and suppose that the received signal $r_{n,k}(t)$ under consideration for correlation is apart from the reference time T_{ref} by $n \times PRI + k \times w$ expressed in (5).

$$\begin{aligned} & r_{n,k}(t)|_{T_{ref} + n \cdot PRI + k \cdot w \le t \le T_{ref} + n \cdot PRI + k \cdot w + T_{window}} \\ & = r_{n,k}(T_{ref} + n \cdot PRI + k \cdot w + t)|_{0 \le t \le T_{window}} \end{aligned} \qquad (5)$$

The spectral components of the signal $R_{n,k}$ is obtained by taking Fourier transform of the signal as in (6) and this is equivalent to the output data from the frequency domain sampler.

$$R_{n,k} = \mathrm{F}\{r_{n,k}(T_{ref} + n \cdot PRI + k \cdot w + t)_{0 \le t \le T_{window}}\}. \qquad (6)$$

Let us denote that $CS_i(n)$ is the n_{th} bit (which is 0 or 1; positive or negative in signal polarity) of the sequence CS_i. The energy of the received signal $r_{n,k}(t)$ after correlation with CS_i is expressed in (7).

$$E_{i,k} = e\{\mathrm{E}\{R_{n,k} CS_i(n)|_{n=0,1,\dots,m-1}\}\} \qquad (7)$$

in which E{*} represents the mean function, and the function e\{*\} calculates the energy of the function inside the bracket. Note that $e\{H(f)\} = |H(f)H^*(f)|$. *The objective of the synchronization process is to determine the values of i and k in (7),* which is described below.

The optimal value of k denoted as k_o represents the time window position with the maximum signal energy, and it is obtained as the combined energy of individual energy.

$$k_o = \max_k \left(\sum_{i=0}^{m-1} E_{i,k} \right) \qquad (9)$$

The function $\max_k(*)$ returns the value k giving the maximum value. The optimal value of i denoted as i_o represents the starting point of the preamble sequence that results in the highest correlation at the time window position k_o, and it is obtained as below.

$$i_o = \max_i \left(E_{i,k_o} \right) \qquad (8)$$

Resultantly, the synchronized starting time T_{synch}, which gives the highest signal correlation energy, is expressed as below.

$$T_{synch} = T_{ref} + k_o \cdot w - i_o \cdot PRI \qquad (10)$$

Let us consider the searching time. The search period is the minimum time period in which there exists a time window T_{window} with the maximum energy. As shown in Figure 1, it can be seen readily that the search period is $(PRI + T_{window})$. which is the search period. As the search time step is w, the term $(PRI + T_{window})/w$ in (11) is the required number of repetitions to sweep the entire search period. The correlation period with the preamble sequence with length m is $(m \times PRI + w)$. Note that the extra term w is the required time to slide the time window by w to adjust new time window for next correlation period after completing a correlation. Hence, the search period to sweep the entire search period is expressed as

$$T_{search} = (PRI + T_{window})(m \cdot PRI + w)/w \qquad (11)$$

Using the above expression enables, a designer can decide an appropriate synchronization time to meet performance requirements by adjusting the time step w and the length m of the preamble code sequence.

Finally, we elaborate the reliability of the correlation energy. If the correlation energy achieved from a certain shifted preamble sequence has less energy than the noise energy, it, in fact, degrades the receiver performance. Therefore, it is good strategy to eliminate such erroneous terms. The noise energy in the received signal is defined as the variation of the received signal as expressed in (12).

$$N_{i,k} = E\left\{ \left(R_{n,k} CS_i(n) \Big|_{n=0,1,\dots,m-1} \right)^2 \right\} - E\left\{ R_{n,k} CS_i(n) \Big|_{n=0,1,\dots,m-1} \right\}^2 \qquad (12)$$

In (12), $N_{i,k}$ is the noise energy introduced to the i_{th} previous symbol at the k_{th} time window position. Note that the i_{th} shifted preamble sequence is applied to the surviving ISI of the symbol transmitted i PRIs previously. Therefore, the survival correlation energy terms are selected by comparing them with the noise energy as in (13).

$$E_{i,k} = \begin{cases} E_{i,k} & \text{if } E_{i,k} \ge N_{i,k} \\ 0 & \text{others} \end{cases} \qquad (13)$$

The receiver ignores correlation terms with less energy than the noise.

B. Channel Estimation

During synchronization, the receiver performs a channel estimation operation on the known data in the preamble. ISI causes frequency selective gains, non-linear phase response, and time dispersion; therefore, an equalizer and a matched filter are indispensable to achieve high performance, and channel estimation is an essential component of each of these.

The channel estimation process extracts the response of the channel to an input pulse, and we propose to estimate the frequency domain channel response H. From (3), one can notice that the channel response $h(t)$ describes the signal waveform of the received signal when the information data is '1' while ignoring the noise term. Therefore, the frequency domain channel response H represents the frequency spectrum corresponding to the time domain signal waveform within a time window when receives positive pulse (bit '0'). As a result, averaging the products of the received signal and preamble code over preamble period can derive the frequency domain channel response H. Note that the averaging operation is necessary in order to minimize the noise effect. The averaging operation during the preamble is given as

$$H_{i,k} \cong \mathrm{E}\left\{ R_{n,k} CS_i(n) \Big|_{n=0,1,\dots,m-1} \right\} \qquad (14)$$

Thus, the receiver needs to store the mean values of the received signal for each k during the preamble and then selects a proper channel response according to k_o. The receiver can also perform the channel estimation operation after synchronization, which eliminates the need for storage of each k at the cost of a slightly increased preamble length. We adopt the former method for our receiver. Hereby, we will omit the term k because k is fixed to the constant value k_o after accomplishing the synchronization.

The strongest signal from all received signals in a time window is from the current symbol denoted H_0. The multipaths of this current symbol may last well into future time windows. Therefore, to minimize ISI, the receiver subtracts H_0 from each future time window as follows.

$$H_0 \cong \mathrm{E}\left\{ R_n CS_0(n) \Big|_{n=0,1,\dots,m-1} \right\} \qquad (15.1)$$

$$H_i \cong \mathrm{E}\left\{ CS_i(n) \{ R_n - H_0 CS_0(n) \} \Big|_{n=0,1,\dots,m-1} \right\}_{i \neq 0} \qquad (15.2)$$

Since the preamble data in the future time windows is known, the receiver can estimate the channel after successively extracting each H_0.

C. Matched Filter and Equalization

Once the channel response is known, it can be applied to form a matched filter and an equalizer. The equalization is an extended correlation operation. Therefore, the channel profile (or channel response for the given time window) used for the equalization is conceptually equivalent to the template signal used for the correlation operation. A matched filter provides the inverse of the channel response without ISI. Ignoring the ISI, the response of the matched filter in the time domain is defined as $h^*(-t)$, and its frequency domain expression is H^*. Assuming that $|H|^2 = 1$, the output of the matched filter for one symbol without ISI is given as

$$\tilde{I} \cong RH^{*}, \tag{16}$$

Note that (16) is the same as correlation operation in frequency domain by replacing H with the template signal as explained in companion paper Part I.

Since the matched filter does not consider ISI in the received signal, the ISI effect should be suppressed before the matched filtering. To cancel the post-cursor multipath interference, estimated symbol is fed back to input of the matched filter with a proper delay and subtracted from the received signal containing future symbols. This approach is similar to a decision feedback equalization (DFE) method in the time domain [3]. The result of the feedback subtraction and matched filtering are shown in (17) and (18), respectively.

$$\tilde{R}_r = R_r - \sum_{d=1}^{L-1} \tilde{I}_{r-d} H_d \tag{17}$$

$$\tilde{I}_r = \tilde{R}_r H^{*}_0 \tag{18}$$

where $\tilde{R}$ is the intermediate received signal before matched filtering. Note that L in (17) is the maximum number of available channel responses including any dummy channel responses with less energy than noise energy in order to process signal in consecutive way.

As shown in Figure 1, a symbol is spread over several time windows. If we treat the operation of (17) and (18) as a finger operation, we can combine fingers to form a rake structure to accumulate the spread symbol signal while mitigating ISI as in (19) and (20). The rake structure operates over multiple PRIs and mitigates ISI, so we term it a multi-PRI rake. It should be noted that a multi-PRI rake enhances our earlier rake structure given in [1], which harvests energy multipath energy within one time window without considering ISI.

$$\tilde{R}_{r+j} = R_{r+j} - \sum_{d=1}^{L-j-1} \tilde{I}_{r-d} H_{j+d} \tag{19}$$

$$\tilde{I}_r = \sum_{j=0}^{L-1} \tilde{R}_{r+j} H^{*}_j \tag{20}$$

A DFE structure based on the above analysis is shown in Figure 2, and it combines effective fingers; as described in (19) and (20); to form a multi-PRI rake structure. The circuit structure is shown in Figure 3, which is a DFE structure, and it shows the method of combining fingers to form the multi-PRI rake structure.

However, the processed signals still suffer from pre-cursor multipath interference except the term of $j=0$ in (19) and (20). Therefore, we can minimize the pre-cursor ISI effect by eliminating the pre-cursor ISI by predicting future symbols. As a result, (19) is modified to

$$\tilde{R}_{r+j} = R_{r+j} - \sum_{d=1}^{L-j-1} \tilde{I}_{r-d} H_{d+j} - \sum_{g=1}^{j} \hat{I}_{r+g} H_{j-g} \tag{21}$$

with the definition of the predicted symbol $\hat{I}$ as

$$\hat{I}_{r+g} = \mathrm{sgn}\left(R_{r+g} H^{*}_0 \Big|_{g=1,2,\ldots,L-1}\right), \tag{22}$$

where the function sgn(*) represents the hard decision.

Since it requires a feedforward structure, the improved scheme is named decision feedforward-backward equalization (DFBE). The improved structure is shown in Figure 4.

For the decision operation in either structure, one can refer to the companion paper Part I.

4. SIMULATION AND OBSERVATIONS

Figure 5 shows the estimated channel response by our channel estimation method with a PRI of 10 ns that causes significant ISI. The preamble uses a sequence length of m=25. The proposed channel estimation method effectively suppresses ISI effect so that the estimated channel response closely matches the channel response of single pulse without ISI.

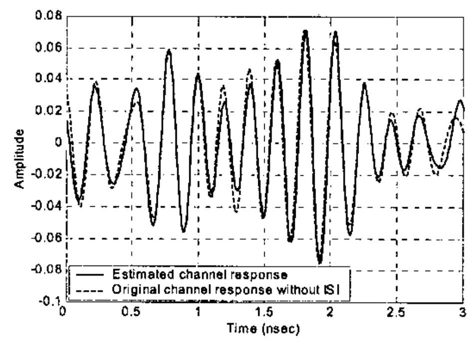

Figure 5: Estimated Channel Response in Time domain

To evaluate the performance of the proposed equalization scheme, we compare it with the analog correlator without ISI. To obtain an upper bound on performance, we also consider a fictional receiver that can capture as much energy from the multipath signal as a frequency domain receiver captures during a time window. Thus, from the theoretical performance of BPSK, we can expect the performance of the fictional receiver to be

$$P_b = Q\left(\sqrt{\frac{2T_{window}}{PRI}\varepsilon_b} \Big/ N_0\right), \tag{23}$$

where P_b is the average probability of error, ε_b is the transmitted signal energy in the UWB pulse, and T_{window}/PRI is the portion of energy captured by the frequency domain receiver. This fictional receiver does not encounter any ISI.

Figure 6 displays the performance of our frequency domain receiver with the synchronization, channel estimation, and equalization. It is compared to the two receivers; the analog correlator and the fictional receiver. The simulation environment applied in this paper is the same as in the companion paper Part I. The number of channel responses L = 5. The analog correlator receiver in this comparison has PRI enough long to ignore ISI effect, in which all multipath signals die out before the next pulse is received.

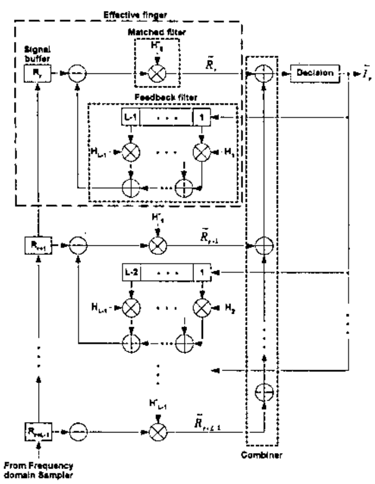

Figure 3: Decision Feedback Structure

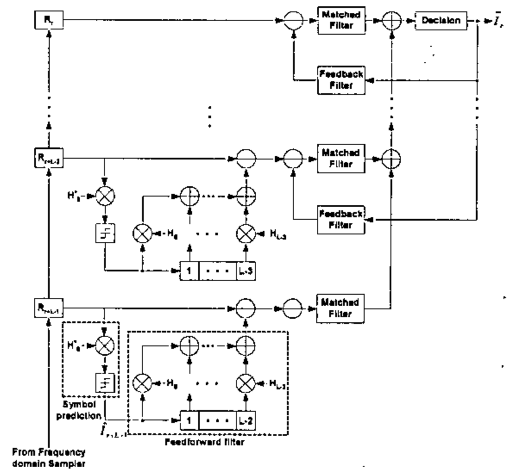

Figure 4: Decision Feedforward-backward Structure

According to the simulation results, the frequency domain receiver with 1-bit ADCs degrades the performance by around 3dB for both DFBE and DFE, compared to the frequency domain receiver with ideal ADCs. However, the frequency domain receivers with 1-bit ADCs still give better performance than the conventional analog correlator receiver by 1dB at BER=10^{-1} even though the analog correlator receiver does not suffer from ISI. If consider the increased circuit complexity and processing cost, the small improvement of performance with DFBE probably is not worth so that we choose DFE scheme for the frequency domain receiver with 1-bit ADCs in Part I.

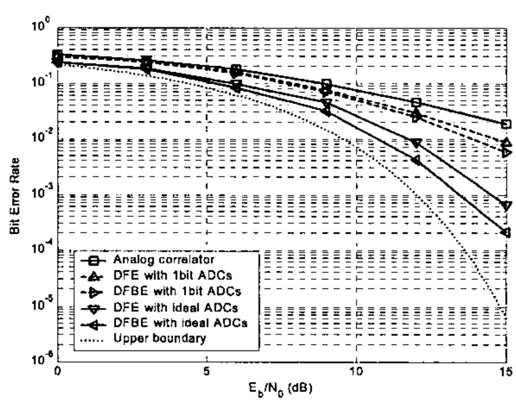

Figure 6: Performance Simulation Results

There are two factors that degrade performance from the upper boundary. The first is that the receiver cannot fully capture the available signal energy due to the non-equal distribution of the signal energy over the delay spread. The second source of degradation occurs when the wrong decision or wrong prediction is fed through the DFE or DFBE.

5. CONCLUSION

In this paper, we introduced a synchronization and equalization scheme for the frequency domain receiver structure. The synchronization finds a time window position with greatest energy. The channel estimation scheme finds the channel response at this position. The matched filter and the equalizer use the results of the channel estimation to minimize ISI. The proposed schemes result in a multi-PRI rake structure over several time windows. The synchronization scheme provides the proper time window position for the individual effective fingers. Therefore, the frequency domain approach does not need to perform synchronization for the individual fingers. This is the one of the advantage of the frequency domain approach over a conventional analog correlator receiver. The analog correlator receiver must perform the synchronization for each individual finger to form the rake structure. Even though we derived the proposed schemes by assuming ideal ADCs, the proposed schemes can be applied to the frequency domain receiver with any multi-bit ADCs and even 1-bit ADCs.

6. REFERENCE

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