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Evaluation of Ultra-Wideband In-vivo Radio Channel and Its effects on System Performance

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Abstract—This paper presents bit error rate (BER) performance analysis and improvement using equalizers for an in vivo radio channel at ultra-wideband (UWB) frequencies (3.1 – 10.6 GHz). By conducting simulations using a bandwidth (BW) of 50 MHz, we observed that the in vivo radio channel is affected by small-scale fading. This fading results in inter-symbol interference (ISI) affecting upcoming symbol transmission, causing delayed versions of the symbols to arrive at the receiver side and causes increase in BER. A 29 taps channel was observed from the experimentally measured data using a human cadaver and BER was calculated for the measured in vivo channel response along with the ideal additive white Gaussian noise (AWGN) and Rayleigh channel models. Linear and non-linear adaptive equalizers i.e., decision feedback equalizer (DFE) and least mean square (LMS) were used to improve the BER performance of the in vivo radio channel. It is noticed that both the equalizers improve the BER, but DFE has better BER compared to LMS and shows a 2 dB and 4 dB performance gain of DFE over the LMS at Eb/No = 12 dB and at Eb/No = 14 dB, respectively. The current findings will help guide future researchers and designers in enhancing systems performance of an ultra-wideband in vivo wireless systems.

Index Terms—Bit error rate, Equalization, Implantable devices, In vivo communication, Wireless body area networks, Ultra wideband.

I. INTRODUCTION

Implantable devices are under research for a while now, and researchers make it possible to commercialize most of the devices including cardiac pacemakers, drug delivery and defibrillators [1-2]. However, the size of the implantable devices is always an issue, and scientists are continuously working to make it smaller, and design micro antennas [3] for the successful communication with the outside wireless devices plus it will help reduce the size of the implantable device. As any electronic device, these devices need stable power required to work correctly and charging them or changing their batteries is an issue. Due to this implanted nature, power efficient [4] devices need to be designed for the successful communication. Experiments have been performed to check if these devices can be charged wirelessly [5]. In [6] an option is presented which uses human motion to recharge the device. To transfer data to the central system or from one node to another node, those devices commonly use the wireless channel, and as this important data is sent wirelessly, it is vulnerable and at risk of attacks by an outside intruder. The communication between implantable devices must be encrypted and secured [7] enough to safely transfer the data. Semantic wireless attacks are performed and tested in [8], and it is shown that the low power and cheaper medical devices are at high risk. Considering the requirements of low power and high data rates, UWB communication can help provide reasonable bandwidths plus low power consumption. Experimental analysis or UWB path loss model is presented in [9]. Different channel models for implantable medical devices (IMD’s) are discussed in [10], and the commercial deployment of these devices are discussed in [11].

Multiple signal strength and impulse response tests were performed using a perfect human body model [12]. It is observed that the variations are more profound at high frequencies up to received signal strength (RSS) 20 dB. Blood vessels are used as a transport channel for in vivo communication in [13], controlled information transfer through an in vivo nervous system is demonstrated using the neurons of an earthworm in [14]. In communication systems we always face errors on the receiver side, caused by multiple factors including effect of noise, outside environment, climate, multipath especially in indoor structures, reflection, diffraction,
fading and interference due to the nearby wireless devices. These factors affect the signal and result in the degradation of bit error rate (BER). In [15] authors demonstrated the maximum allowed transmitted power from an in vivo device to achieve the desired BER while maintaining the specific absorption rate (SAR). The targeted data rates of 100 Mbps is achieved with maximum SAR which is the amount of radio frequency (RF) absorbed by the body [16]. The SAR provides a metric for the observed power amount in the tissues. Federal Communications Commission (FCC) recommends that the value of SAR must be less than 1.6W/kg taken over a volume having 1g of tissues. To obtain a high data rate communication, UWB is still a good option due to its properties of low power and high bandwidth.

UWB channel is used in different areas of medical science for positioning and communication in the operating room [17] both in live (during operation) and non-live scenarios. It is observed that even within a highly multipath environment UWB provides reasonable results. UWB channel characteristics and system model is discussed in [18]. In communication system there is always a tradeoff to achieve high bandwidth with low power. The use of UWB introduces inter-symbol interference (ISI), which affects the performance of the system and thus results in high BER’s. To avoid or improve the BER performance of UWB in vivo communication, using equalizers [19] is the best option.

There are different types of equalizers present in the literature, they can be divided into two main types, adaptive linear and non-linear equalizers. The most effective linear equalizers are recursive least square (RLS) and least mean square (LMS) equalizers [20]. RLS and LMS can be used in combination by using their properties effectively [21]. On the other hand, the most common non-linear equalizers are decision feedback equalizer (DFE) and maximum likelihood sequence estimator (MLSE) equalizers. There are different types of DFE [22] equalizers available in the literature with their specific functions according to the requirement [23-25], such as to improve the BER and reduce ISI to get the desired signal at the receiver end. Channel coding is also used to improve the BER performance of the system but channels with high ISI first need to be equalized [30-32] to cancel or reduce ISI and then recovered using channel coding. This paper presents the BER analysis of experimental in vivo radio channel with and without equalization. for the first time in the literature as per author’s knowledge. The channel response is compared with AWGN and Rayleigh channel. Furthermore, different types of equalizers (both linear and non-linear) are used to improve the BER performance. Overall, the equalizers improved the performance significantly. According to the results it is clear that non-linear equalizer outperformed the linear equalizer. DFE shows a 2 dB performance gain over the LMS at Eb/No = 12 dB and 4 dB at Eb/No = 14 dB.

The rest of this paper is organized as follows. Section II discussed the experimental setup used to perform experiments using a human cadaver. Section III presents the bit error rate performance of experimental in vivo channel without using equalization and comparing it with ideal additive white Gaussian noise (AWGN) channel and ideal Rayleigh channel. Section III focuses on the equalization techniques and use of equalizer along with the mathematical equations used for all the equalizers. Section IV presents the simulations of a BER performance for in vivo communication and spectrum of the unequalized and equalized signals. Finally, future developments are discussed in the conclusion in section V.

II. EXPERIMENTAL SETUP

A human cadaver was used to perform the experiments as presented in [28, 34, 36]. All the experiments were performed under the assistance of certified medical doctor in a hospital after ethical approvals. In order to generate affective data, fresh organs such as heart, stomach and intestine from a sheep were used to place inside the human cadaver. The complete list of equipment’s can be seen in Table I.

Initially the equipment’s were checked to make sure the system was working properly in order to avoid false reading the cable losses are calculated properly [34]. Two types of antennas were used in the experiments an in vivo and an ex vivo antenna. In vivo antennas [35] was used to place inside the human cadaver in different places including on top of heart, below heart, on top of stomach, inside stomach and below stomach and finally on top of intestine, inside intestine and below intestine. The ex vivo antennas were placed on the center of the body, near the head at the right lateral, left lateral. Extensive experiments were performed to collect the data, each experiment was performed multiple times and the average was taken to be used in simulations and mathematical modeling. A two-meter-long coaxial cable was used to connect the antennas with a vector network analyzer (VNA). The experimental setup can be seen in Figure 1.

![Figure 1 - Experimental setup for UWB In-vivo measurements.](image)

<table>
<thead>
<tr>
<th>Equipment/Subject</th>
<th>Values/quantity</th>
</tr>
</thead>
<tbody>
<tr>
<td>Measurement device</td>
<td>VNA</td>
</tr>
<tr>
<td>Cable</td>
<td>Coaxial</td>
</tr>
<tr>
<td>No of Antennas</td>
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</tr>
<tr>
<td>Antenna Type 1</td>
<td>Ex vivo</td>
</tr>
<tr>
<td>Antenna Type 2</td>
<td>In vivo</td>
</tr>
</tbody>
</table>
III. BER PERFORMANCE OF EXPERIMENTAL IN VIVO CHANNEL

In vivo communication is a highly multipath communication that suffers from fading due to the dense structure of the human body, as in vivo devices must be placed inside the human body. It is also highly location dependent and a slight change in the position of the device can affect the channel and performance of the system and its link budget. We used UWB frequencies between (3.1 – 10.6 GHz) with the central frequency $f_c$ = 6.75 GHz and bandwidth $BW = 50$ MHz for our simulations of the experimental data. The complete list of simulation parameters can be found in Table II. Channel response $h(t)$, is extracted by IFFT using (1) in MATLAB\(^3\). The channel response of the experimental data can be seen in Figure 2 where it lies and which is the closed resembled channel to understand the scenario. A 29 taps channel can be observed a highly multipath channel with high ISI. The mathematical modeling for the channel is presented in [33].

$$X(n) = \frac{1}{N} \sum_{k=1}^{N} X(K) * e^{j2\pi p t + n K_N} \quad (1)$$

Where, $X(K)$ represents the frequency domain samples, $X(n)$ represent the time domain samples $N$ is the size of IFFT and $k$ is 0,1,2,3,……N-1.

The signal is modulated using binary phase shift keying (BPSK) modulation with oversampling factor of 4, and can be define as (2) [26], root raised cosine (RRC) pulse shaping is used with span = 10 and a roll-off factor of 0.25 using (3) [27]. The signal is then convolved with 29 taps in vivo channel response, and AWGN is added as noise. At the receiver end, the signal is again demodulated, and BER is calculated for the in vivo channel without using any equalization which as expected show the worst channel scenario affected by fading and ISI. The channel is further compared with the ideal AWGN and Rayleigh channel to understand the status of in vivo channel and where it lies and which is the closed resembled channel currently available and known in the literature. The BER of BPSK in AWGN is calculated using (4). The Rayleigh channel was selected for the comparison as it is used for highly multipath and non-line of side (LOS) fading scenarios, which has a probability density function given in (5). The BER simulations for in vivo channel, ideal Rayleigh and ideal AWGN are shown in Figure 3. The in vivo BER is plotted without using any performance enhancing tools the aim was to observe the BER in its original form directly from the experimental data and as expected the worst performance can be seen in the simulation which is because of the multipath and non-line of sight scenario in a highly congested environment inside the human cadaver.

$$V_{BPSK}(t) = b(t) \sqrt{2P} \cos 2\pi f_c t \quad (2)$$

Where $0 < t < T$.

And

$$b(t) = +1 \text{ or } -1, f_c \text{ is the carrier frequency, and } T \text{ is the bit duration. The signal has a power } P = \frac{A^2}{2}, \text{ so that } A = \sqrt{2P}, \text{ where } A \text{ represents the peak value of sinusoidal carrier.}$$

$$H(f) = \begin{cases} \sqrt{T} & (0 \leq |f| \leq \frac{1-\beta}{2T}) \\ \frac{T}{2} \left[ 1 + \cos \left( \frac{\pi T}{\beta} \left( |f| - \frac{1-\beta}{2T} \right) \right) \right] & \left( \frac{1-\beta}{2T} \leq |f| \leq \frac{1+\beta}{2T} \right) \\ 0 & \left( |f| > \frac{1+\beta}{2T} \right) \end{cases}$$

(3)

Where $f$ is frequency, $T$ is the symbol time, and $\beta$ is the roll-off factor.

$$p_r(r) = \begin{cases} \frac{1}{\sqrt{\pi}} e^{\frac{-r^2}{2\sigma^2}} & 0 \leq r \leq \infty \\ 1 & 0, r < 0 \end{cases}$$

(5)

### TABLE II

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values/units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bandwidth</td>
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</tr>
<tr>
<td>Central Frequency</td>
<td>6.75GHz</td>
</tr>
<tr>
<td>S-parameters</td>
<td>$S_{21}$</td>
</tr>
<tr>
<td>Time</td>
<td>$\mu$sec</td>
</tr>
<tr>
<td>Channel Response</td>
<td>dB</td>
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<td>Frequency to Time Domain</td>
<td>IFFT</td>
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<tr>
<td>No of Channel Taps</td>
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</tr>
<tr>
<td>Modulation Scheme</td>
<td>BPSK</td>
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<td>Over-Sampling Factor of BPSK</td>
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<tr>
<td>Pulse Shaping</td>
<td>RRC</td>
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<tr>
<td>RRC Span</td>
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</tr>
<tr>
<td>RRC Roll Off Factor</td>
<td>0.25</td>
</tr>
<tr>
<td>Comparison Channel 1</td>
<td>AWGN</td>
</tr>
<tr>
<td>Comparison Channel 2</td>
<td>Rayleigh</td>
</tr>
</tbody>
</table>

![Figure 2 - Channel response of in vivo channel with BW = 50 MHz](image-url)
A highly multipath channel and high BER is observed in Figure 3. In order to improve the BER and obtain our desired performance, equalizer need to be use to improve the BER, this will help us improve the BER by compromising the ISI.

IV. EQUALIZATION

Equalization is the process used to render the frequency component of an electronic signal [29]. It helps to get rid of the ISI in the time dispersive and frequency selective channels; those are the channels in which the signal bandwidth is higher than the coherence bandwidth ($B_c$). ISI introduces distortion in the signal resulting in symbol overlapping which makes it hard for a receiver to distinguish between the desired and undesired symbols causing high BER affecting the performance of the system. An additional reason for an ISI includes multipath scattering environments which are basically non-line of sight signals. Those signals result in the delayed version of a transmitted signal as the signal arrive from a different direction with different power at the receiver and starts interfering with other transmitted symbols. To mitigate the effect of ISI and improve BER, performance equalizers are used to compensate the effect of ISI on a signal.

There are different types of equalization techniques available in the literature to accommodate for different scenarios [23-26]. The most effective ones are the adaptive equalizers distributed as linear and nonlinear equalizers, both of them are used to improve the system performance and to subsequently select the best equalizer for the system. In this article, least mean square equalizer was picked as a linear equalizer combined with recursive least square equalizer and decision feedback equalizer, from the nonlinear equalizers. Some tests were performed using maximum likelihood sequence estimator equalizer (MLSE) equalizer to improve the BER, as in theory MLSE has the best performance, but practically it is the most complicated equalizer [27]. It is observed that for a 29 taps channel MLSE faces memory problems, although with lesser number of taps the MLSE performs better than the LMS and DFE. Considering the number of taps, we only consider LMS and DFE for our tests.

RLS and LMS both are adaptive equalizers, but each of them has its advantages and disadvantages. RLS algorithm converges quickly but the execution takes time, and it is slow as compared to LMS. Where the complexity of the RLS grows roughly with the square of the number of weights which makes it unstable especially in a situation where we need to use higher weights. The RLS algorithm can be summarized with the initialization using (6) [26].

$$w(0)=k(0)=x(0)=0, R^{-1}(0)=\delta I_{NN}$$ (6)

Where, $I_{NN}$ is an N×N identity Matrix, and $\delta$ is a large positive constant. Computing recursively using (7-11) as follows.

$$w(n) = w(n-1) + k(n)e^*(n)$$ (7)

$$k(n) = \frac{R^{-1}(n-1)y(n)}{\lambda + y^T(n)R^{-1}(n-1)y(n)}$$ (8)

$$x(n) = e(n) + \hat{d}(n)$$ (9)

Where

$$\hat{d}(n) = w^T(n-1)y(n)$$ (10)

$$R^{-1}(n) = \frac{1}{\lambda} [R^{-1}(n-1) - k(n)y^T(n)R^{-1}(n-1)]$$ (11)

Where $\lambda$ represent the weighting coefficient.

The block diagram presenting the structure of RLS equalizers is shown in Figure 4.

![Figure 4 - RLS block diagram used for equalizing in-vivo signal.](image)

While on the other hand, LMS algorithm executes quickly but the convergence process is slow, and the complexity of the LMS increases linearly with respect to weights. LMS can be computed using (12-14) [26].

$$\hat{d}_k(n) = w^T_N(n)y_N(n)$$ (12)

$$e_k(n) = x_k(n) - \hat{d}_k(n)$$ (13)

$$w_N(n+1) = w_N(n) - \alpha e_k^*(n)y_N(n)$$ (14)

Where $n$ represents a sequence of iterations, $N$ number of delay stages and $\alpha$ is step size.

The block diagram presenting the structure of LMS is shown in Figure 5.
The recursive least square and least mean square algorithms are used together by exploiting their properties of quick convergence (RLS) and fast execution (LMS), which help the simulation to execute quickly and present better results.

V. SIMULATION AND RESULTS DISCUSSION

The general requirement for implementation of adaptive equalizers consists of a number of taps, step size, signal constellation (BPSK in our case) and an initial set of weights.

The goal of the least mean square equalizer is to minimize the mean square error (MSE) presented in the output of the equalizer. The prediction of error is highly dependent on the tap gain such that the MSE of the equalizer output is the function of weight. The recursive least square on the other hand requires the calculation of tap gain vector so that the cumulative square error can be minimized.

While for the nonlinear equalizer, decision feedback equalizer is used which has \( N_1 + N_2 + 1 \) taps in the feedforward filter and \( N_3 \) taps in the feedback producing an output (15). The main idea of DFE is to estimate and subtract the symbols which will introduce ISI on future symbols [26].

\[
\hat{d}_k = \sum_{n=-N_1}^{N_2} c^*_n y_{k-n} + \sum_{i=1}^{N_3} f_i d_{k-i} \quad (15)
\]

Where \( c^*_n \) and \( y_n \) are tap gain and the input \( f^*_i \) tap gain input for feedback. \( d_i (i < k) \) is the previous decision made on the detected signal. The minimum mean square error (MSE) a DFE can be achieved is (16)

\[
E[|e(n)|^2]_{min} = \exp\left\{\frac{\tau}{2\pi} \int_{-\pi}^{\pi} \ln \left[ \frac{N_p}{|F(e^{j\omega})|^2 + N_0} \right] d\omega \right\} \quad (16)
\]

The mean excess delay can be calculated using (17).

\[
\tau = \frac{\sum_{k} \tau_k P(\tau_k) \tau_k}{\sum_{k} \tau_k P(\tau_k)} \quad (17)
\]

And, The RMS delay spread is defined as (18).

\[
\sigma_t = \sqrt{\tau^2 - (\tau)^2} \quad (18)
\]
Where

\[ r^2 = \frac{\sum_k p(x_k) r_k^2}{\sum_k p(x_k)} \tag{19} \]

\[ B_c \approx \frac{1}{50 \sigma_r} \tag{20} \]

This is not critical for a narrow band (NB) communication but it can cause issues while working with ultra-wideband. As in our case, we are working on the \( BW = 50 \text{ MHz} \) which is much higher than the calculated \( B_c \) and makes the signal frequency selective.

The block diagram presented in Figure 6 is showing the equalizers structure used to improve the BER performance of the in vivo radio channel. The basic limitations with the linear equalizers are their poor performance on the channel having spectral nulls. Decision feedback equalizer is a non-linear equalizer and has the advantage to subtract the distortion on a current pulse that is caused by the previous pulses. In the block diagram of DFE, the forward filter and the feedback filter can be both linear filters and are working as a linear filter while working separately but the non-linearity of decision feedback is because of the non-linear properties of the detector. The main idea of DFE is if the values of the symbols previously detected are known it can be used to cancel the ISI right at the output of the feedforward filter. In order to minimize the mean square error, the weights of the feedforward and feedback filters can be adjusted simultaneously to produce better results.

So if we send a single bit from a transmitter and pass it through the channel, the channels act like a low pass filter and it smears the transmitted bit and introduce inter-symbol interference. The DFE is used to subtract that smearing, which comes from the previous bit. So it slices the bit and delays it by one-unit interval multiply it by a constant that represent the amount of smearing and finally subtract that amount from the next bit, which helps return a voltage closer to the true value better results. RLS converges faster than LMS and LMS can be executed quickly. The parameter values for LMS and DFE are set by using 55 taps linear equalizer and 29 taps feedforward and 29 taps feedback weights for DFE. The detailed list of the parameter are shown in Table III. RLS is used only for the first data block at each Eb/No which helped us rapidly converge the taps and LMS algorithm is used for the remaining data blocks in order to ensure rapid execution speed.

A linear equalizer object is constructed and implemented in the simulations for both LMS and DFE. First, the simulation for the linear equalizer is executed and the equalizer signal spectrum of a linearly equalized signal can be seen in Figure 8. It is observed that as the Eb/No increases the spectrum has a deeper null, which point us to the fact that a linear equalizer must use more taps to get better performance. After that, the DFE equalizer has been executed and the signal spectrum for DFE is plotted as shown in Figure 9. It is observed that at low BER DFE suffers from error bursts but later it shows dynamically improved results compare to LMS. DFE performs much more effectively on the in vivo channel as compared to the LMS by mitigating the channel null better than LMS. The DFE errors bursts were higher compared to LMS that is because of its feedback detection of bits instead of the correct bits.

Two types of power spectrums are extracted using power spectral density (PSD) functions. Power spectral density or power spectrum are used to characterize random processes in frequency domain. The power spectrum \( S(\omega) \) is actually the discrete time Fourier transform (DTFT) of the correlation sequence \( r[k] \) for the process. The power spectral density can be defined as (21).

\[ S(\omega) = \sum_{k=-\infty}^{\infty} r[k] e^{-j k \omega} \tag{21} \]

Or equivalently

\[ r[k] = \frac{1}{2\pi} \int_{-\pi}^{\pi} S(\omega) e^{j k \omega} d\omega \tag{22} \]

In terms of an interpretation, if the power spectrum is integrated between \( \omega_a \) and \( \omega_b \) and called that \( P_{ab} \) and normalized it by \( 2\pi \), the quantity will represent the expected contribution to total power or variance due to components of the random process between theses points \( \omega_a \) and \( \omega_b \), and can be represented as (23)

\[ P_{ab} = \frac{1}{2\pi} \int_{\omega_a}^{\omega_b} S(\omega) d\omega \tag{23} \]

Hence, by finding the area under \( S(\omega) \) between \( \omega_a \) and \( \omega_b \) that’s the power of this portion of the spectrum is expected to contribute the random process, which tells us how the contribution of the power are distributed in frequency.

The power spectrum is actually the density, hence the units in terms of radian frequency will be represented as \( S(f) = \text{power/radian} \) and the frequency that measured the units of hertz will be presented as \( S(f) = \text{power/hertz} \).

The total power can be represented as (24)

\[ r[0] = \frac{1}{2\pi} \int_{-\pi}^{\pi} S(\omega) d\omega = E[x^2[n]] \tag{24} \]
Where the power cannot be zero.

\[ S(\omega) \geq 0 \quad \text{non negative} \]

Hence by comparing the power spectrums, both linearly equalized and decision feedback equalizer power spectrum, it can be seen that the maximum fluctuation observed in un equalized frequency response is between 0 dB to -40 dB. Where the linear equalization reduces it between 0 dB to -30 dB. Although DFE outperformed the LMS and presents the best results by keeping the power spectrum between 0 dB to -10 dB. These results show that a non-linear decision feedback equalizer offers better performance than the recursive least square and least mean square combined. As the power spectrum was improved by 25% using least mean square and 75% using decision feedback equalizer considering the highest fluctuated peaks. Although the analysis is based on the maximum peaks but the overall fluctuation of the power spectrum using Figure 8 for linearly equalized least mean square is between 0 dB to -10 dB and 0 to -20 dB in that case part of the spectrum is recovered between 50% to 75%. While in the case of decision feedback, it remains the same throughout the simulation as shown in Figure 9, which shows better and stable performance of decision feedback compared to least mean square.

Finally, the BER results are shown for both LMS and DFE equalizers in Figure 10 along with the ideal BPSK as a comparison. It can be clearly observed that DFE easily outperformed the BER performance compare to LMS. It can be concluded that non-linear equalizers performed better for in vivo channels as compare to linear equalizers. Although with low Eb/No initially the LMS performance was slightly better than DFE but at Eb/No = 6 dB and so on DFE shows a much better BER results compared to LMS.

The detailed analysis with exact quantitative values for the bit error rate performance comparison between least mean square equalizer and decision feedback equalizer are presented in Table IV.

<table>
<thead>
<tr>
<th>Table IV</th>
<th>BER Quantitative Results</th>
</tr>
</thead>
<tbody>
<tr>
<td>Eb/No</td>
<td>Ideal BPSK</td>
</tr>
<tr>
<td>0</td>
<td>0.07865</td>
</tr>
<tr>
<td>2</td>
<td>0.03751</td>
</tr>
<tr>
<td>4</td>
<td>0.0125</td>
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<tr>
<td>6</td>
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<td>8</td>
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</tr>
<tr>
<td>10</td>
<td>3.872e-6</td>
</tr>
<tr>
<td>12</td>
<td>N/A</td>
</tr>
<tr>
<td>14</td>
<td>N/A</td>
</tr>
</tbody>
</table>

It can be clearly seen from the values that the ideal binary phase shift keying has the best performance in its ideal state. Initially the least mean square equalizer show better performance as compared to the decision feedback algorithm.
The linear mean square was incorporated with recursive least square algorithm in order to converge the system quickly and the LMS further take the operation to execute it rapidly. The results of LMS was slightly better until Eb/No = 6, afterwards the decision feedback equalizer present significant performance compared to the least mean square and dramatically improve the bit error rate.

VI. CONCLUSION

This paper presents a detailed analysis and improvement of in vivo radio channel BER performance with and without different equalizers. The paper shows the similarity between the in vivo channels that do not use any equalizers and the Rayleigh channel. Both channels suffer from severe fading due to the non-line of sight situation. Furthermore, different equalizers were used to test their performance for improving BER performance of an in vivo channel and compromise the effect of ISI caused by using the UWB technology. It is clear that the in vivo channel is a frequency selective channel that will suffer from ISI with ultra-wideband communication. Therefore, equalizers must be used to compensate for the effects of the ISI. This provides a preliminary insight of in vivo communication. The presented analysis offers novel results and findings to obligate the communication and can help guide future research that will acquire more time, resources and challenges in such environment. This study can be used as a stepping stone for comprehensive and thorough studies in this undesired fading environment.

REFERENCES


Muhammad Ilyas (S’18) received his B.S. degree in Computer Engineering from COMSATS IIT Abbottabad, Pakistan, in 2010 and his M.S. degree in Electrical and Computer Engineering from Altinbas University, Istanbul, Turkey in 2014. He is currently pursuing his Ph.D. degree in the Department of Electrical and Computer Engineering, Altinbas University, Istanbul, Turkey. His current research interests include in vivo wireless communication, wireless body area networks and implantable devices.

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