UWB MISO Time Reversal With Energy Detector Receiver Over ISI Channels

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Abstract—This paper investigates a multiple input single output (MISO) time reversal system for ultra-wideband (UWB) communication over inter-symbol interference (ISI) channels. Time reversal takes advantage of rich scattering environments to achieve signal focusing, which enables the use of simple receiver structures without sacrificing performance. On-off-keying (OOK) modulation and energy detection (square law) are considered for the purpose of low complexity at the receiver. This research is motivated by the need for high-data-rate wireless network with simple receive nodes. The discrete channel models and bit error rate (BER) formulas for the energy-detector receiver over ISI channels are derived. Performance is evaluated based on measured data, considering practical signal waveforms. One reason to use our own measured data is that there is no proper correlation demodulation (ACD) formula for the energy-detector receiver. Another reason is that time reversal needs additional signal processing, which is not straightforward. In particular, these suboptimal schemes' equivalent discrete channel structures without remarkably affecting signal energy collection, which is greatly in favor of some low-complexity suboptimal receivers, such as the ACD receiver and energy-detector receiver.

Motivated by all of these facts, in this paper we study and the principle can be seen in [7][15][20]-[25]. It takes advantage of rich scattering environments to achieve signal focusing both temporally and spatially. Generally speaking, the temporal focusing feature can soften the impact of ISI, while the spatial focusing feature can be utilized to transmit information to an intended location with limited signal leakage at other locations.

UWB time reversal is an emerging technique that takes advantage of the unique properties of UWB channels. A few transmitter-side enhancements were proposed to reduce multipath spread impact. Channel shortening is a technique that was introduced in 1970’s and aims at using an optimal filter at the transmitter to reduce channel memory and simplify equalizer at the receiver [26]. Mathematically, a channel inverse pre-filter is necessary to fully eliminate multipath effect. Although a time reversal pre-filter does not approach the optimum, practically the overall time reversal processing is much simpler than doing inverse filtering.

Normally a high data rate means a system with high complexity thus more expensive. One difficulty at high data rate is in dealing with ISI. Traditional ISI mitigation techniques include equalization, RAKE and OFDM, and all of them are expensive solutions that use coherent detection and require channel estimation at the receivers. However, by using time reversal and employing non-coherent detection at the receiver, the cost-vs-data-rate issue can be softened to some extent. Thanks to time reversal's temporal focusing mechanism that condenses signal and reduces ISI impact, it is possible to use a simple receiver to communicate at high data rate with insignificant performance degradation caused by reduced impact of multipath. Note that time reversal needs additional signal processing resource, but this processing is at the transmitter side and the processing complexity can be reduced [15].

Take downlink transmission in a centralized network as an example, the transmitter is in the base station thus can be very powerful. In addition, a sharpened signal would enable narrow-window integration that reduces noise accumulation without remarkably affecting signal energy collection, which is greatly in favor of some low-complexity suboptimal receivers, such as the ACD receiver and energy-detector receiver.

I. INTRODUCTION

Stimulated by the FCC’s move that allows UWB waveforms to overlay over other systems’, ultra-wideband (UWB) radio has received tremendous interests in the past couples of years [11]-[19]. Mainly due to potentially low implementation complexity, suboptimal reception strategies, such as transmitted reference (TR) [8]-[12], autocorrelation demodulation (ACD) [13]-[15] and energy detection [16]-[19], become attractive for complexity and cost constraint UWB applications. But these types of systems suffer from huge performance degradation. In particular, these suboptimal schemes’ equivalent discrete channels exhibit nonlinear behavior [10][15], which implies that when multipath delay spread exceeds a certain value such that received symbol waveforms overlap and inter-symbol interference (ISI) occurs, normal linear equalization techniques may not work properly.

One unique characteristic that differentiates a UWB system from a “narrow” band system is the UWB propagation channel. The UWB channel impulse response (CIR) contains a large number of resolvable components coming through different paths, especially in indoor environments. However, making good use of these signal components is not straightforward. In this paper we consider a signal focusing technique called time reversal that can turn multipath (and even ISI) into benefit without remarkably affecting signal energy collection, which is greatly in favor of some low-complexity suboptimal receivers, such as the ACD receiver and energy-detector receiver.

Motivated by all of these facts, in this paper we study
a UWB multiple input single output (MISO) time reversal scheme with consideration of measured channels as well as major practical issues. Two reasons to do measurements are: (1) it is intended to obtain solid and convincing results, thus practical waveforms and an end-to-end radio frequency (RF) chain need to be considered; (2) existing UWB channel models, such as the IEEE 802.15 channel models, do not have location information, so that they are not suitable for studying antenna array related issues. We are interested in the bottom line performance of a suboptimal receiver with time reversal enhancement, and an energy-detector receiver (see Fig.1) in conjunction with the MISO time reversal technique is considered in this paper. Highlighted below are the contributions of this paper:

- Practical channels based on measurements in the office are considered;
- A UWB MISO time reversal scheme coupled with on-off keying (OOK) and energy detection is proposed and examined;
- Bit error rate (BER) performance of the energy-detector receiver over ISI channels is analyzed.

The paper continues with Section II that describes the system in detail. Discrete channel models and BER analysis are provided in Section III. Some of our measurement results as well as analytical results are presented in Section IV, followed by concluding remarks in Section V.

II. SYSTEM DESCRIPTION

To isolate issues we limit our discussion to a single user scenario and assume the channel remains static during a data burst (say 100 $\mu$s [5]). We further assume that the forward channel and reversal channel are reciprocal, i.e., their CIRs are identical. An ideal low-pass filter with one-sided bandwidth $W$ is placed at the receiver’s front-end, where $W$ is chosen such that the impairment on the received signal due to filtering is negligible. OOK modulation and energy-detector receiver are considered.

A. Baseline Scheme

The transmitted signal with OOK modulation is

$$S_{tx}(t) = \sum_{j=-\infty}^{\infty} d_j w_{tx}(t - jT_b),$$

where $T_b$ is the symbol duration, $w_{tx}(t)$ is the transmitted symbol waveform defined over $[0, T_b)$, and $d_j \in \{0, 1\}$ is $j$-th transmitted bit. Without loss of generality, assume the minimal propagation delay is equal to zero. The received noise-polluted signal at the output of the receiver front-end is

$$r(t) = h(t) \otimes S_{tx}(t) + \nu(t)$$

$$= \sum_{j=-\infty}^{\infty} d_j w_{rx}(t - jT_b) + \nu(t),$$

where $h(t)$ is the CIR that takes into account the overall effect of the RF front-end circuits at both the transmitter and receiver, $\nu(t)$ is a low-pass additive zero-mean Gaussian noise with one-sided bandwidth $W$ and one-sided power spectral density $N_0$, and $w_{rx}(t)$ is the received symbol-“1” waveform given by

$$w_{rx}(t) = h(t) \otimes w_{tx}(t).$$

An energy-detector receiver performs squaring operation, integration over a given time window, and threshold decision. Corresponding to the time index $k$, the $k$-th decision variable at the output of the integrator is given by

$$z_k = \int_{kT_b}^{(k+1)T_b} r^2(t) dt.$$  

B. Time Reversal Scheme

Time reversal does not work without knowing the channel. As mentioned in previous section, the channels in both links are assumed to be reciprocal, and the node sends a pilot signal via uplink channel prior to data transmission. Upon receiving a train of pilot waveforms, the base station performs waveform estimation to reduce noise impact. To focus on central issues, we further assume perfect pilot waveform estimation. A MISO time reversal configuration is conceptually illustrated in Fig.2, where there are $M$ transmit antenna elements targeting at one receive antenna, $c_m(t)$ is the pre-filter connected with the antenna element $m$, and $h_m(t)$ is the CIR for the channel associated with the $m$-th transmit antenna element and the receive antenna. The received waveform is given by

$$r(t) = \sum_{j=-\infty}^{\infty} d_j \sum_{m=1}^{M} c_m(t - jT_b) \otimes h_m(t) + \nu(t)$$

$$= \sum_{j=-\infty}^{\infty} d_j y_{rx}(t) + \nu(t),$$

where $y_{rx}(t)$ is the received symbol-“1” waveform defined as

$$y_{rx}(t) = \sum_{m=1}^{M} c_m(t - jT_b) \otimes h_m(t).$$

Let $p_{tx}(t)$ be the transmitted pilot pulse and $p_{rx,m}(t) = p_{tx}(t) \otimes h_m(t)$ the pilot signal received from the antenna element $m$ at the base station. The pre-filter can be set as

$$c_m(t) = a_m w_{tx}(t - T_0) \otimes p_{rx,m}(T_0 - t)$$

$$= a_m w_{tx}(t - T_0) \otimes p_{tx}(T_0 - t) \otimes h_m(T_0 - t),$$

where $a_m$ is the weight with respect to the antenna element $m$, $T_0$ is a constant that results in a time shift. Let $T_m$ be the maximum multipath excess delay of all $M$ CIRs, then $T_0 \leq T_m$ would guarantee that $c_m(t)$ is causal. $w_{tx}(t)$ in (7) functions as a filter that takes care of all other impacts along the transmitter-receiver chain. If we consider equal transmit power distribution among the $M$ transmit antenna elements, the weights can be determined by

$$a_m \propto \sqrt{\int_{-\infty}^{\infty} [w_{tx}(t - T_0) \otimes p_{rx,m}(T_0 - t)]^2 dt},$$

$$1 \leq m \leq M.$$
III. PERFORMANCE ANALYSIS

Assume the effect of a single input bit \( d_k \) lasts \( N = \text{ceil}(T_{in}/T_b) \) symbols.

A. Baseline Receiver

Substituting (2) into (4), and taking into account a multipath excess delay of \( N \) symbols, it follows that

\[
z_k = \int_{kT_b}^{(k+1)T_b} \left[ \sum_{j=k-N+1}^{k} d_j w_{rx}(t-jT_b) + \nu(t) \right]^2 dt
= \int_{0}^{T_b} \sum_{j=0}^{N-1} d_{k-j} w_{rx}(t+jT_b) \nu(t+kT_b) + \nu^2(t+kT_b) dt
= \int_{0}^{T_b} \sum_{j=0}^{N-1} d_{k-j} w_{rx}(t+jT_b) \nu(t+kT_b) dt + \eta_k,
\]

where \( \eta_k \) is a noise term given by

\[
\eta_k = \int_{0}^{T_b} 2 \sum_{j=0}^{N-1} d_{k-j} w_{rx}(t+jT_b) \nu(t+kT_b) + \nu^2(t+kT_b) dt
\]

Define matrices \( C \) as

\[
C = \begin{pmatrix}
c_{0,0} & c_{0,1} & \cdots & c_{0,N-1} \\
c_{1,0} & c_{1,1} & \cdots & c_{1,N-1} \\
\vdots & \vdots & \ddots & \vdots \\
c_{N-1,0} & c_{N-1,1} & \cdots & c_{N-1,N-1}
\end{pmatrix}
\]

\[
c_{i,j} = \frac{1}{\sqrt{E_b T_b}} \int_{0}^{T_b} w_{rx}(t+iT_b) w_{rx}(t+jT_b) dt
= c_{j,i}.
\]

then Equation (10) can be rewritten as

\[
z_k = \tilde{d}^T C \tilde{d} + \eta_k
\]

\[
\tilde{d} = (d_k, \ldots, d_{k-N+1})^T
\]

which means that the signal part in the output of the equivalent discrete channel (represented by \( \tilde{d}^T C \tilde{d} \)) is a nonlinear function of data vector \( \tilde{d} \). As a matter of fact, the equivalent discrete channel is a special case of second-order Volterra model [10]. The decision variable \( z_k \) contains a desired signal \( d_k^T c_{0,0} \), a non-Gaussian noise term \( \eta_k \), and a nonlinear ISI component that cannot be well handled by normal linear equalization techniques.

Let \( V_T \) be the decision threshold. We are interested in two types of erroneous probabilities for a given data vector:

\[
P_0(\tilde{d}, d_k = 0) = \text{Pr}(z_k > V_T | \tilde{d}, d_k = 0),
\]

\[
P_1(\tilde{d}, d_k = 1) = \text{Pr}(z_k \leq V_T | \tilde{d}, d_k = 1).
\]

The BER can be calculated based on the above two probabilities by averaging over all possible combinations of previous transmitted bits:

\[
P_b = \frac{1}{2^N} \left[ \sum_{d_k=0} P_0(\tilde{d}, d_k = 0) + \sum_{d_k=1} P_1(\tilde{d}, d_k = 1) \right],
\]

where equal probability of sending “0” bit and “1” bit has been used. It is well-known that the decision variable \( z_k \) has a Chi-square distribution with \( 2T_b W \) degree of freedom [27] (pp.298). \( z_k \) is a central Chi-square random variable when \( \tilde{d}^T C \tilde{d} = 0 \), and it is a noncentral Chi-square random variable when \( \tilde{d}^T C \tilde{d} \neq 0 \). A relevant traditional problem is to calculate probability of false alarm and probability of detection. A number of approximate approaches to this issue can be found in the literature [28][17][18]. Among these approximating methods is Park’s model that is suitable for all ranges of \( T_b W \), and it will be employed in this paper to compute BER. To apply Park’s model, define a probability of false alarm by

\[
P_f = \text{Pr}(z_k > V_T | \tilde{d}^T C \tilde{d} = 0).
\]

From the definition (14), \( P_0(\tilde{d}, d_k = 0) \) is a probability of detection if \( \tilde{d}^T C \tilde{d} \neq 0 \), and it is the probability of false alarm \( P_f \) if \( \tilde{d}^T C \tilde{d} = 0 \). On the other hand, \( P_1(\tilde{d}, d_k = 1) \) is a probability of missing if \( \tilde{d}^T C \tilde{d} \neq 0 \), and it is equal to \( 1 - P_f \) if \( \tilde{d}^T C \tilde{d} = 0 \). According to Park’s model, \( P_0(\tilde{d}, d_k = 0) \) and \( P_f \) are approximately associated with a signal-to-noise ratio (SNR) \( \text{SNR}(\tilde{d}, d_k = 0) \):

\[
P_0(\tilde{d}, d_k = 0) = Q \left( Q^{-1}(P_f) - \sqrt{\text{SNR}(\tilde{d}, d_k = 0)} \right),
\]

\[
\text{SNR}(\tilde{d}, d_k = 0) = \frac{2T_b W \left( \tilde{d}^T C \tilde{d} \right)^2}{2 + \tilde{d}^T C \tilde{d}},
\]

where the Q-function is given by

\[
Q(x) = \frac{1}{\sqrt{2\pi}} \int_{x}^{\infty} e^{-y^2/2} dy,
\]

\( Q^{-1}(x) \) is its inverse function, and for a given data vector \( \tilde{d} \), \( \tilde{d}^T C \tilde{d} \) represents signal part including ISI contribution. Similarly, \( P_1(\tilde{d}, d_k = 1) \) can be approximately expressed as

\[
P_1(\tilde{d}, d_k = 1) = 1 - Q \left( Q^{-1}(P_f) - \sqrt{\text{SNR}(\tilde{d}, d_k = 1)} \right)
= Q \left( \sqrt{\text{SNR}(\tilde{d}, d_k = 1)} - Q^{-1}(P_f) \right).
\]
paper to set a threshold based on two worst signal cases corresponding to

\[ P_{0,\text{max}} = \max_{\tilde{d}, d_k = 0} P_0(\tilde{d}, d_k = 0) \]  

(21)

and

\[ P_{1,\text{max}} = \max_{\tilde{d}, d_k = 1} P_1(\tilde{d}, d_k = 1), \]  

(22)

or equivalently, corresponding to

\[ SNR_{0,\text{max}} = \max_{\tilde{d}, d_k = 0} SNR(\tilde{d}, d_k = 0) \]  

(23)

and

\[ SNR_{1,\text{min}} = \min_{\tilde{d}, d_k = 1} SNR(\tilde{d}, d_k = 1). \]  

(24)

The threshold is chosen such that \( P_{d,\text{max}} = P_{m,\text{min}}, \) or

\[ Q \left( Q^{-1}(P_f) - \sqrt{SNR_{0,\text{max}}} \right) = Q \left( \sqrt{SNR_{1,\text{min}}} - Q^{-1}(P_f) \right). \]  

(25)

Solving the above equation for \( Q^{-1}(P_f) \) yields

\[ Q^{-1}(P_f) = \frac{\sqrt{SNR_{0,\text{max}}} + \sqrt{SNR_{1,\text{min}}}}{2}. \]  

(26)

Thus (17) and (20) can be rewritten as

\[ P_0(\tilde{d}, d_k = 0) = \frac{Q \left( \sqrt{SNR_{0,\text{max}}} \right)}{Q \left( \sqrt{SNR_{0,\text{max}}} + \sqrt{SNR_{1,\text{min}}} \right)} - \frac{\sqrt{SNR(\tilde{d}, d_k = 0)}}{2}, \]  

(27)

\[ P_1(\tilde{d}, d_k = 1) = \frac{Q \left( \sqrt{SNR(\tilde{d}, d_k = 1)} - \sqrt{SNR_{0,\text{max}}} \right)}{Q \left( \sqrt{SNR(\tilde{d}, d_k = 1)} - \sqrt{SNR_{1,\text{min}}} \right)} - \frac{\sqrt{SNR_{0,\text{max}}}}{2}. \]  

(28)

### B. Receiver for Time Reversal

Note that the overall CIR would last at most \( T_0 + T_m \) seconds and all \( M \) peaks would appear at time \( T_0 \). The lower-end and upper-end boundaries for \( k \)-th received symbol are \( kT_b + T_0 - T_b/2 \) and \( kT_b + T_0 + T_b/2 \), respectively. To count the impact of symbol overlapping, define

\[ N_1 = \text{ceil} \left( \frac{T_0 - T_b/2}{T_b} \right), \quad N_2 = \text{ceil} \left( \frac{T_m - T_b/2}{T_b} \right). \]  

(29)

A received symbol waveform covers \( N_1 + N_2 + 1 \) signal peaks (or symbols) with \( T_b \) being repetition interval; \( N_1 \) and \( N_2 \) are the number of received peaks before and after the peak time of current symbol. Denoted by \( T_f \) the integration window size, then (10)-(14) can be slightly modified for the time-reversal-enhanced receiver:

\[ z_k = \int_{kT_b + T_0 - T_b/2}^{kT_b + T_0 + T_b/2} \left[ \sum_{j=-N_1}^{N_2} d_{k-j}y_{rx}(t + jT_b) \right]^2 dt + \eta_k, \]  

(30)

\[ \eta_k = \int_{kT_b + T_0 - T_b/2}^{kT_b + T_0 + T_b/2} \left[ 2 \sum_{j=-N_1}^{N_2} d_{k-j}y_{rx}(t + jT_b)\nu(t + kT_b) + \nu^2(t + kT_b) \right] dt, \]  

(31)

\[ C = \begin{pmatrix} c_{-N_1,-N_1} & \cdots & c_{-N_1,0} & \cdots & c_{-N_1,N_2} \\ \vdots & & \vdots & & \vdots \\ c_{0,-N_1} & \cdots & c_{0,0} & \cdots & c_{0,N_2} \\ \vdots & & \vdots & & \vdots \\ c_{N_2,-N_1} & \cdots & c_{N_2,0} & \cdots & c_{N_2,N_2} \end{pmatrix}, \]  

(32)

\[ c_{i,j} = \frac{1}{\sqrt{E_bT_b}} \int_{kT_b + T_0 - T_b/2}^{kT_b + T_0 + T_b/2} y_{rx}(t + iT_b)y_{rx}(t + jT_b)dt \]  

(33)

and

\[ z_k = \tilde{d}^T C \tilde{d} + \eta_k, \quad \tilde{d} = (d_{k+N_1}, \ldots, d_{k-N_2})^T. \]  

(34)

The same method discussed in the last subsection can be used to calculate BER, but notice that \( \tilde{d} \) is equal to \( (d_{k+N_1}, \ldots, d_{k-N_2})^T \), instead of \( (d_k, \ldots, d_{k+N_1})^T \), and the matrix \( C \) is defined in a slightly different way.

### IV. Measurement and Analytical Results

Measurements are necessary since there is no proper UWB channel model for multiple antenna related study. A time-domain channel sounding system was used for measurements in our office. The sounding pulse has an RMS width of 250 ps. The office is a typical indoor area with wooden and metallic furniture (chairs, desks, bookshelves and cabinets). Distances between the transmitter and the receiver are over six meters and no line of sight between the two antennas. The transmit antenna is a 4-element linear array with 20-cm separation spacing, and each antenna element has dimension 5.5 cm × 11 cm. This channel sounding system is considered as an end-to-end RF chain in evaluating the proposed communication system’s performance. Shown in Fig.3 are the waveforms measured at the receive antenna. From the experiments it has been observed that the downlink and uplink UWB channels between two sites are very symmetric and static, which is truly in favor of time reversal applications. Performance evaluations are made based on the measured data, assuming \( w_{tx}(t) = p_{tx}(t) \). In addition, we assume one pulse per data symbol and \( W = 2 \) GHz.
Ideally the received single-pulse waveform can be represented by \[ |h(t) \otimes h(t)| \otimes |p_{tx}(-t) \otimes p_{tx}(t)| \], implying the main lobe in the received waveform almost has the same profile as the autocorrelation of the single-pulse composite waveform \[ p_{tx}(-t) \otimes p_{tx}(t) \]. Indeed, the main-lobe profile is not greatly affected by undesired distortion. The integration window size \( T_I \) is chosen such that most of the main-lobe energy is captured, and in this paper it is set to \( T_I = 6 \) ns.

Temporal focusing characteristic can be quantified by an energy ratio of main-lobe energy to total energy. The measured energy ratios are given in Table I, where “element” refers to an element of the transmit antenna array. The results suggest that the use of antenna array at the transmitter does enhance signal focusing. For MISO configuration with a constant total transmit power, it is also shown that the main-lobe amplitude is roughly in proportion to \( \sqrt{M} \), which is because of coherent signal combining around time \( t = T_m \).

A BER performance comparison for \( T_I = 20 \) ns is presented in Fig.4, where a curve labeled with “element” refers to a single input single output (SISO) configuration using antenna element \( m \). In the plot \( E_b/N_0 \) is the per-bit SNR at the received side. Note that if the per-bit SNR is measured at the transmitter side, approximately a MISO system would have an additional gain of \( 10 \log_{10} M \) dB over a SISO system. From these results we can conclude that 1) time reversal can effectively reduce ISI impact; 2) the effectiveness of SISO time reversal is location dependent and the use of MISO can increase the system robustness; 3) the inverse filter results in the best performance at the cost of increasing system complexity. It can be expected that as the number of antenna elements increases, the MISO system performance would be comparable to the inverse filter performance bound.

V. CONCLUSION

The UWB MISO time reversal scheme combined with the energy-detector receiver is examined. The equivalent discrete channel models and the BER formulas are derived. BER performance is evaluated considering practical end-to-end RF chain including propagation environment. Our results suggest that the proposed combination of MISO time reversal and energy-detector receiver is very promising for wireless applications in the UWB band, especially when complexity/cost, performance and security are highly concerned.

Channel stability and reciprocity are an essence for applying UWB time reversal [25]. Because of ultra fine multipath resolution, verified by our tests, the UWB channels can be viewed as being static for burst mode communication. It is an open issue to prove reciprocity of wireless channels. Our recent tests (conducted in the Wireless Networking System Laboratory, Tennessee Technological University, Cookeville, TN) showed that perfect reciprocity is difficult to achieve. The degree of reciprocity can vary depending on what types of antennas are employed and how circuits are set up. In terms of the correlation between the downlink and uplink waveforms received at the two sites, waveform similarity around 95 percent was demonstrated in the laboratory. This correlation actually determines profile sharpness of the received waveform in a time reversal system. The measured result suggests that UWB channel reciprocity is imperfect, but this imperfection would be negligible for applying UWB time reversal.

For the system employing time reversal, a narrow integration window and accurate timing can help rejecting noise and ISI. An inverse filter (to replace the time reversal pre-filter) can perfectly form an overall channel with only one significant path. Obviously implementing such inverse filter would be infeasible at present. Comparing to the inverse filter solution, the use of MISO time reversal is much more practical and affordable. Signal focusing can be continuously improved as the number of antenna elements increases. Robustness of UWB time reversal also increases as the number of antenna elements increases, which is very similar to diversity improvement in the fading environments.

REFERENCES


TABLE I. Energy ratio (T1 = 6 ns).

<table>
<thead>
<tr>
<th>Element</th>
<th>SISO</th>
<th>MISO</th>
</tr>
</thead>
<tbody>
<tr>
<td>Energy ratio</td>
<td>53.00%</td>
<td>53.63%</td>
</tr>
</tbody>
</table>

![Fig. 1. Energy detector receiver.](image1)

![Fig. 2. MISO configuration.](image2)

![Fig. 3. Waveforms at the receive antenna’s output.](image3)
Fig. 4. BER comparison.