Time-Reversal Techniques for MISO and MIMO Wireless Communication Systems

A. E. Fouda, M. E. Yavuz, and F. L. Teixeira

Abstract

We consider the application of different time-reversal (TR) signal processing and beamforming techniques to multiple-input single-output (MISO) and multiple-input multiple-output (MIMO) wireless communication systems. Conventional TR beamforming provides spatial focusing at the intended receiver; however, it does not yield perfect channel equalization. Time-reversed pilot can be normalized to provide perfect equalization at the expense of power level. This equalization is particularly important for high data rates where the bit error rate performance is dominated by internal noise due to intersymbol interference. To increase physical layer covertness, TR beamforming is combined with the multiple-signal-classification (MUSIC) technique to produce null fields at eavesdroppers. This technique is also applied to MIMO setups to eliminate interuser interference and hence increase system capacity. Differential TR is used to obtain and update pilot signals for passive moving receivers, i.e. those that cannot (or do not) transmit pilot signals. Time-reversed differential backscattered signal is able to provide satisfactory spatial and temporal focusing at the moving receiver.

Index Terms

Communication covertness, differential time-reversal, interuser interference, pre-equalization, spatial focusing, temporal compression, time-reversal beamforming, time-reversal MUSIC.

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I. INTRODUCTION

With the continuous increase in demand for higher data rates in wireless communications, ultrawideband (UWB) systems are becoming more popular [1]. Despite their advantages, UWB systems are very sensitive to delay spreads caused by multipath in rich scattering environments and require complex and expensive receivers to compensate for delays of different replicas of the desired received signal. The application of time-reversal (TR) techniques to UWB wireless communication systems has gained increasing interest recently [2], [3], [4], [5], [6], [7], [8], [9]. TR provides a simple and cost-efficient solution to the problem of delay spread in UWB systems. It moves the burden of equalization from the receiver side to the transmitter and channel sides, thus allowing for simpler and cheaper receivers at the expense of some added complexity to the transmitter [7].

TR provides both spatial and temporal focusing of the signal at the receiver [10], [11]. Not only does it mitigate negative effects of multipath, but also it is capable of actually harnessing multipath to achieve better focusing resolution (superresolution) both in time and space [11], [2], [3], [12], [13]. The temporal focusing property amounts to a type of pre-equalization procedure that reduces (or cancels) intersymbol interference (ISI) at the receiver [14]. Spatial focusing combats channel fading [10], maximizes delivered power to the intended receiver, and therefore enables power saving at the transmitter side and/or increasing channel capacity and communication range. Spatial focusing also reduces power leakage to other locations. This is very important to reduce interuser interference (IUI) in multiuser configurations, which in turn allows for a more effective use of space-division multiple access (SDMA) to boost the system capacity [5], [8], [9], [15]. Spatial focusing also adds a degree of physical layer security (covertness) to the systems, making it hard for eavesdroppers away from the intended receiver’s location to decode the signal using traditional decoding techniques [14], [16].

In this paper, we study the application of different TR techniques to wireless communication systems. We start with a brief review on the basics of TR communication and its performance metrics. Then, we consider multiple-input single-output (MISO) configurations, where there exists one intended receiver and possibly one or more eavesdroppers. We introduce and study the applicability of three TR techniques, namely, (i) equalized TR beamforming, (ii) TR beamforming with multiple-signal-classification (MUSIC), and (iii) differential TR, in both free-space and
rich scattering environments. Equalized TR beamforming yields perfect channel equalization on the expense of spatial focusing. TR beamforming combined with MUSIC produces null fields at eavesdroppers locations for increased physical layer covertness. Differential TR is used to extract the pilot signal of passive moving receivers (scatterers) from array acquisitions [17]. These passive receivers may represent, for example, receivers in one-way communication links, sensors such as passive RFIDs or intercepting eavesdroppers. We highlight relative strengths and limitations of these different techniques, and compare their bit error rate performances under high and low data rates operations. We also compare the performance of equalized TR beamforming in rich scattering scenarios with that of conventional (non-equalized) TR and conventional (non TR-based) beamforming. Finally, we consider multiple-input multiple-output (MIMO) configurations, where TR beamforming with MUSIC is shown to significantly reduce undesired interuser interference.

II. TIME-REVERSAL COMMUNICATIONS AND PERFORMANCE METRICS

Consider a MISO communication link between an $N$ elements transmitter array and a receiver (let us call it receiver $A$). To start the link, the receiver transmits a signal that is recorded by the transmitter array elements. These recorded $N$ signals constitute the steering vector (column vector of Green’s functions or equivalently impulse responses) between receiver $A$ and each element in the array. The steering vector (or processed versions of it, as discussed later) is used as a pilot for information signal transmission from the array to receiver $A$. From the time-reversal invariance of the wave equation, when the $N$ signals of the pilot are time-reversed and simultaneously transmitted, they tend to automatically focus at the intended receiver location (regardless of the intervening medium which is, in general, not known to the transmitter array) and produce a compressed pulse in time as well [3], [2]. After time-reversed backpropagation, the received signal at location $i$ in the frequency domain is given by the following inner product

$$H_i(\omega) = \langle g_i(\omega), p_A(\omega) \rangle$$

where $p_A(\omega)$ is the pilot vector of receiver $A$, $g_i(\omega) = [G_{(i,1)}(\omega), ..., G_{(i,N)}(\omega)]^T$ is the steering vector of location $i$, where $G_{(i,n)}(\omega)$ is the Green’s function between location $i$ and the $n^{th}$ element of the array, and $\langle a, b \rangle = b^\dagger a$ denotes the inner product between $a$ and $b$ where $\dagger$ represents a conjugate transpose.
The received signal in the time domain is obtained by taking the inverse Fourier transformation \( h_i(t) = \mathcal{F}^{-1}\{H_i(\omega)\} \). We refer to \( h_i(t) \) as the equivalent channel impulse response after TR, or simply the channel impulse response (CIR). The CIR peak at location \( i \) is defined as \( \eta_i = \max_t |h_i(t)| \). This will be used as a measure for the spatial distribution of energy [18].

The intersymbol interference (ISI) is defined as the ratio of the value of the CIR at integer multiples of period \( T \) to the peak value of CIR as follows

\[
ISI_i = \frac{\sum_{n=-N,n\neq 0}^{N} |h_i(\tau_i + nT)|}{|h_i(\tau_i)|}
\]

where \( T \) is the reciprocal of the bandwidth \( B \) [19]. So if \( H_i(\omega) \) is flat with frequency, \( ISI_i = 0 \) which is the case of perfect equalization.

An information signal \( S(\omega) \) is filtered using the time-reversed (phase conjugated in the frequency domain) pilots. The transmitted signal vector in the frequency domain intended to reach receiver \( A \) is given by \( t_A(\omega) = S(\omega)p_A^\dagger(\omega) \), where * denotes complex conjugation. For transmission over the physical channel, the transmitted vector is converted into the time domain by simply taking inverse Fourier transformation \( t_A(t) = \mathcal{F}^{-1}\{t_A(\omega)\} \).

The received signal in the frequency domain can be written as \( R_A(\omega) = p_A^\dagger(\omega)g_A(\omega)S(\omega) \) or equivalently in the time domain as \( r_A(t) = s(t) *_t h_A(t) + n(t) \), where *\(_t\) denotes convolution and \( n(t) \) is the additive noise at the receiver.

III. TR Beamforming Techniques for Wireless Communications

In this section, the exact steering vector of the receiver, also known as the channel state information (CSI) between the array and the receiver, is assumed to be perfectly known to the transmitters. This implies that the time period that the transmitter array takes to record the receiver’s steering vector is long enough to capture (mostly) all the multiple scattering in the medium. In addition, the sampling rate is equal to or higher than the Nyquist sampling rate corresponding to the bandwidth of operation. In other words, we are assuming that TR does not add intrinsic bandwidth limitation, i.e. if the original communication system (before using TR) is capable of generating signals with certain bandwidth \( B \), we assume that it will also be capable of generating and sampling the TR pilot over the same bandwidth. Deviations from ideality due to some hardware limitations are discussed in Section IV.
A. TR Beamforming

1) Conventional TR (without equalization): The simplest form of the pilot is to coincide with the steering vector of the intended receiver [4], that is

$$p_A(\omega) = g_A(\omega)$$

This choice exhibits high spatial focusing performance that is further enhanced in rich scattering scenarios, as will be shown in the next section. However, the equivalent CIR at the intended receiver is proportional to $$\|g_A(\omega)\|^2$$, where $$\|g_A(\omega)\|$$ is the norm of $$g_A$$ given by $$\|g_A\| = \sqrt{\langle g_A, g_A \rangle}$$. This means that the CIR is not flat with frequency, i.e. the equivalent channel transfer function is not perfectly equalized. This gives rise to undesired intersymbol interference which hampers performance in case of high data rates, as shown in the next section.

2) Equalized TR: In order to achieve perfect equalization at the receiver, the steering vector can be normalized by the square of its norm as follows

$$p_A(\omega) = \frac{g_A(\omega)}{\|g_A(\omega)\|^2}$$

This guarantees a flat CIR at the receiver, as can be easily deduced from (1). This choice, however, implies inferior spatial focusing performance as compared with conventional TR. In addition, rich scattering scenarios do not result in better spatial focusing, in general.

B. Conventional Beamforming

In conventional (non TR-based) beamforming, the information available to the array about the receiver can be its direction (which can be estimated using some direction of arrival (DOA) algorithms [20]), or, in the best-case scenario, its position with respect to the array. So if we assume that the array knows the receiver location, it can generate an approximate pilot, which is the steering vector of the receiver location based on free-space assumption. Note that although the array may know the receiver’s location, it does not know its exact CSI. Following the above discussion, the pilot vector of conventional beamforming can be written as

$$p_A(\omega) = \frac{\tilde{g}_A(\omega)}{\|\tilde{g}_A(\omega)\|^2}$$

where $$\tilde{g}_A$$ is the steering vector of receiver A based on some background medium assumption that may not correspond to the actual one.
C. TR Beamforming with Nulling at Eavesdroppers

If one of the concerns of the communication system is to minimize information leakage to eavesdroppers (or interuser interference in case of MIMO configurations), TR beamforming can be combined with MUSIC algorithm to produce null fields at eavesdroppers. Assume the presence of \( M \) eavesdroppers \( M < N \), whose steering vectors are known to the array. The pilot of receiver \( A \) can be orthogonalized to each and everyone of the steering vectors of the \( M \) eavesdroppers as follows

\[
p_A(\omega) = \frac{g_A(\omega)}{\|g_A(\omega)\|^2} - \sum_{i=1}^{M} \left( \frac{g_A(\omega)}{\|g_A(\omega)\|^2}, \hat{g}_i(\omega) \right) \hat{g}_i(\omega)
\]

where

\[
\hat{g}_i(\omega) = \frac{g_i(\omega)}{\|g_i(\omega)\|}
\]

For the above processing to produce ideal nulling at the \( M \) locations, the \( M \) steering vectors \((g_i(\omega), i = 1, \ldots, M)\) must be mutually orthogonal. The pilot given by (6) does not produce perfectly equalized CIR at the intended receiver. For perfect equalization, the pilot is normalized as follows

\[
\overline{p}_A(\omega) = \frac{p_A(\omega)}{\langle g_A(\omega), p_A(\omega) \rangle^*}
\]

D. Extracting Pilot Signals from Array Acquisitions

1) Differential TR: For scenarios involving receivers that are unable to transmit pilots, like, for example, passive RFIDs or receivers in one-way communication links, differential TR techniques [17] can be used to extract an approximate version of the receiver’s pilot. Assuming that the initial location of the receiver (denoted by location 0) is known to the array, differential TR proceeds as following: the array transmits a beam \( g_0(\omega) \) twice at two close time instants. The beam illuminates the moving receiver (target) at two adjacent locations (locations 1 and 2), and backscatterings from both illuminations \((s_1(\omega) \text{ and } s_2(\omega))\) are recorded and subtracted to yield the differential backscattering vector, written as \( d(\omega) = s_2(\omega) - s_1(\omega) \). It was shown in [17] that when \( d(\omega) \) is time-reversed and backpropagated, it focuses somewhere between locations 1 and 2. Therefore, \( d(\omega) \) can be used as a pilot for communication. To achieve better equalization at the receiver, \( d(\omega) \) can be normalized as follows

\[
p_A(\omega) = \frac{d(\omega)}{\langle g_0(\omega), d(\omega) \rangle^*}
\]
Fig. 1: Equalized TR beamforming in free-space using linear array (indicated by white circles). The intended receiver is indicated by “o” and the eavesdropper is indicated by “x”. (a) Normalized CIR peak distribution (in dB). (b) ISI (in dB).

2) Filtered MDM: As mentioned before, for the TR beamforming with MUSIC algorithm defined in (6) to produce simultaneous null fields at $M$ eavesdroppers, the steering vectors of the eavesdroppers must be mutually orthogonal. One way to obtain a set of orthogonal steering vectors corresponding to the existing eavesdroppers is through the decomposition of the time-reversal operator (DORT, under its French acronym) [21], [22], [23], [24] obtained from array acquisition of the multistatic data matrix (MDM) [25]. The MDM at frequency $\omega$ is given by $K(\omega) = [k_1(\omega), ..., k_N(\omega)]$, where $k_i(\omega) = [K_{i1}(\omega), ..., K_{Ni}(\omega)]^T$ is the vector of scattered fields received by the array elements resulting from an illumination by the $i^{th}$ element. DORT provides a set of significant eigenvalues/vectors pairs each pertaining to a point-like scatterer. If we are interested in obtaining the set of eigenvectors of the eavesdroppers, the MDM can first be filtered to eliminate the contribution of the intended receiver. Assuming that the steering vector of the intended receiver is known, columns of the filtered MDM are obtained via $\tilde{k}_i = k_i - \langle k_i, \tilde{g}_A \rangle \tilde{g}_A$, where $\tilde{g}_A = g_A/\|g_A\|$. 
Fig. 2: Same as Fig. 1, but in the presence of surrounding walls with relative permittivity equal to 4 and thickness equal to 60 cm.

IV. MISO CONFIGURATION: SIMULATION RESULTS AND PERFORMANCE COMPARISON

A. Problem Setup

In this section, we assess and contrast the performances of the proposed TR techniques applied to MISO configurations. MISO here refers to the configuration where a multiple-input antenna array is communicating with one single intended receiver. To assess the covertness property offered by TR, we assume the presence of one eavesdropper, and compare the performance of the wireless channel at both the intended receiver and the eavesdropper. Our goal is to maximize the signal “quality” at the receiver, while achieving maximum covertness (low probability of intercept (LPI)) at the eavesdropper [14].

The simulation domain is either extended free-space or a 3.6 m × 3.6 m room surrounded by 60 cm thick walls with relative permittivity of 4. The transceivers array consists of eight point sources in 2-D. Two array deployments are considered: (i) A dense linear array with inter-element spacing of 15 cm and total aperture of 105 cm as shown in Fig. 1. (ii) A sparse square-shaped full-aspect array as shown in Fig. 3. Pilots are extracted from UWB pulses generated as the first derivative of Blackmann Harris (BH) pulse [26] with center frequency of 1.25 GHz and useful bandwidth covering from DC up to 2.5 GHz. All simulations are carried out using the finite-difference time-domain (FDTD) method [27].
TABLE I: Summary of the considered MISO setups.

<table>
<thead>
<tr>
<th>Acronym</th>
<th>TR technique</th>
<th>Array configuration</th>
<th>Background medium</th>
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<tbody>
<tr>
<td>ETRLFS</td>
<td>Equalized TR beamforming</td>
<td>Linear</td>
<td>Free-space</td>
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<tr>
<td>ETRLW</td>
<td>Equalized TR beamforming</td>
<td>Linear</td>
<td>with walls</td>
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<tr>
<td>ETRFAFS</td>
<td>Equalized TR beamforming</td>
<td>Full-aspect</td>
<td>Free-space</td>
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<td>ConvLW</td>
<td>Conventional beamforming</td>
<td>Linear</td>
<td>with walls</td>
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<td>TRLFS</td>
<td>Conventional TR beamforming</td>
<td>Linear</td>
<td>Free-space</td>
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<tr>
<td>TRLW</td>
<td>Conventional TR beamforming</td>
<td>Linear</td>
<td>with walls</td>
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<tr>
<td>MLFS</td>
<td>TR MUSIC w/o equalization</td>
<td>Linear</td>
<td>Free-space</td>
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<tr>
<td>EMLFS</td>
<td>TR MUSIC w/ equalization</td>
<td>Linear</td>
<td>Free-space</td>
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<td>DTRLFS</td>
<td>Differential TR beamforming</td>
<td>Linear</td>
<td>Free-space</td>
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<tr>
<td>DMLFS</td>
<td>Differential TR MUSIC w/o equalization</td>
<td>Linear</td>
<td>Free-space</td>
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B. Spatial Focusing and Temporal Compression

Different setups are summarized in Table 1. Pilots of all setups are normalized to have equal total (unity) energies. We start with equalized TR beamforming in free-space and use it as a reference for comparison with other setups. Spatial distributions of the peak value of the CIR and the ISI are plotted in Fig. 1. These plots are for linear array operating in free-space. It is obvious from the figure that the spatial focusing of this setup is not so good. The power is distributed over a wide region of space, and the eavesdropper (indicated by “\(x\)” in the figure) is receiving a significant amount of power. Also, the ISI at the eavesdropper is comparable to that at the receiver. Consequently, the eavesdropper will be receiving a high quality signal that allows it to intercept the signal.

Better spatial focusing is obtained when the same linear array is operating in the presence of rich scattering environment, such as including the surrounding walls shown in Fig. 2. This is a result of the way TR exploits multipathing, where frequency components of the CIR add (only) incoherently at all locations in space except at the receiver’s location where different frequency components add coherently [28], [11], [3], [2]. At these other locations, the peak value of the CIR drops compared with its value at the receiver and it loses its temporal compression. In this
case, the covertness capability of TR becomes more evident.

CIR peak values and ISI at the intended receiver and eavesdropper for all setups are plotted in Fig. 8. It is interesting to note that, despite the spatial focusing in the presence of walls evident in Fig. 2, the CIR peak at the receiver actually decreased. As a consequence of the input energy normalization, the ratio between the CIR peak at the receiver in the presence of walls to that in free-space is

$$\sqrt{\frac{\int_\omega \| g_{A,f}(\omega) \|^2 d\omega}{\int_\omega \| g_{A,w}(\omega) \|^2 d\omega}}$$

where the subscripts $f$ and $w$ refer to free-space and with walls respectively. This ratio is not necessarily larger than unity. A natural question arises as to where the energy in the presence of walls goes. Note that the spatial distributions in Fig. 1 and 2 are for the CIR peak value not the total energy. The energy in the presence of walls goes into the side lobes of the CIR at locations other than the intended receiver. This is evident in the increased ISI (degraded the temporal compression) at the eavesdropper in the presence of walls.

Full-aspect deployment is considered in Fig. 3. Such deployment is convenient for indoors communication systems, where fixed transmitters can be mounted on the surrounding walls. Comparing the performance of the full-aspect array with that of linear array with the same number of elements, Fig. 3 shows that better spatial and temporal focusing around the receiver are achievable using full-aspect arrays.

The performance of conventional (non-equalized) TR in free-space and in the presence of walls is considered in Fig. 4. Comparing Fig. 4 with Fig. 1 and 2 for the equalized TR
Fig. 4: Normalized CIR peak distribution of conventional TR beamforming (a) in free-space. (b) in the presence of surrounding walls.

case shows that conventional TR exhibits better spatial focusing. Moreover, Fig. 8 shows that conventional TR provides for higher CIR peaks at the expense of increased ISI, as expected. The ratio between the CIR peak in the presence of walls to that in free-space is given by
\[ \sqrt{\int \omega g_{A,w}(\omega) d\omega / \int \omega g_{A,f}(\omega) d\omega}, \]
which is always larger than unity. This shows that conventional TR makes use of multipath to deliver more power to the intended receiver while scrambling the signal at other locations. Although multipath serves to increase the CIR peak to temporal sidelobe ratio at the intended receiver [3], it increases the temporal span of the sidelobes, which increases the overall ISI as shown in Fig. 8(b) [10].

At this point, it is useful to compare the performance of TR-based beamforming with that of conventional beamforming. Conventional beamforming, despite estimating the receiver’s DOA or even its location, is unable to compensate for delays due to multipathing and hence unable to achieve satisfactory equalization. Therefore it yields higher ISI than the equalized TR case as shown in Fig. 8(b).

The CIR peak images of TR MUSIC without and with equalization for the case of linear array in free-space are shown in Fig. 5(a) and (b), respectively. It is interesting to compare these images with Fig. 1(a) to see how MUSIC produces null field at the eavesdropper while delivering power to the intended receiver. This is a common feature to both equalized and non-equalized cases. Equalization, however, spreads out the power into larger region of space. This means that,
Fig. 5: Normalized CIR peak distribution of TR MUSIC using a linear array (a) without equalization. (b) with equalization.

given equal input powers, equalization results in less power reaching the receiver as shown in Fig. 8.

As explained in the previous section, differential TR can be used to extract the pilot signal of passive moving receivers, as that shown in Fig. 6(a). The resulting CIR peak image is very close to that obtained using TR beamforming in Fig. 1(a). The magnitude of the CIR in the frequency domain at the receiver (being at location 2) is plotted in Fig. 6(b). This figure shows that satisfactory equalization is attainable using differential TR pilots. Differential TR can also be used to extract the steering vectors of moving eavesdroppers to be used in the MUSIC algorithm. This application is important to increase system covertness, since intruding eavesdroppers usually do not communicate with the array. In Fig. 7(a), differential TR is used to obtain the steering vector of the moving eavesdropper, which is then used in the MUSIC algorithm to produce null field along the eavesdropper’s path. Note that the null in this case is not perfect because differential TR provides only an approximate version of the steering vector. Finally, in Fig. 7(b), DORT applied to the filtered MDM is used to obtain the eigenvectors of the two indicated eavesdroppers. MUSIC is then used to produce simultaneous nulls at the eavesdroppers, as shown.
Fig. 6: Differential TR beamforming using a linear array. The intended receiver is moving downwards in the direction of the arrow. (a) Normalized CIR peak distribution (in dB). (b) Magnitude of the CIR in the frequency domain at the intended receiver.

Fig. 7: Eavesdropper pilot retrieval from array acquisitions. (a) Differential TR MUSIC using a linear array. The intended receiver is indicated by the white circle and the eavesdropper is moving along the path indicated by the black arrow. (b) TR MUSIC using eigenvectors obtained from the filtered MDM to simultaneously nullify on the two eavesdroppers indicated by square boxes.

C. Bit Error Rate Performance

In this section, we compare the bit error rate (BER) performance of the discussed techniques under high and low data rates operations. In the high data rate case, bits are represented by
Fig. 8: Performance comparison of the proposed MISO setups under equal input power assumption. Acronyms are summarized in Table 1. (a) CIR peak and (b) ISI, at the intended receiver and the eavesdropper.

Fig. 9: Transmitted bit stream. (a) High data rate time-domain signal. (b) High data rate spectrum. (c) Low data rate time-domain signal. (d) Low data rate spectrum.
Nyquist pulses whose spectrum covers the entire useful bandwidth of operation (UWB pulses) as shown in Fig. 9(a) and (b). In this particular example, a maximum bit rate of 5 Gbps is accommodated, which is equal to twice the bandwidth. For the low data rate case, we use binary phase shift keying (BPSK) passband modulation as shown in Fig. 9(c) and (d). In this example, a bit rate of 0.125 Gbps is used, which corresponds to a $-10$ dB bandwidth of 0.178 GHz.

Received bit streams at the intended receiver and at the eavesdropper are obtained by convolving the input stream with the corresponding equivalent CIR. Additive white Gaussian noise (AWGN) is added, and the noisy stream is decoded. Assuming perfect synchronization at the decoder, Nyquist pulses stream is decoded by simply sampling the stream at the bit rate, whereas BPSK stream is decoded using coherent detection [29]. Finally, Monte Carlo simulation is used to compute the BER of the decoded stream. For the computed BER to represent a statistically stable measure for the probability of error, the process is repeated over a large number of noise realizations and the results are averaged. In our simulations, we generate streams of 1000 bits and average over 100 noise realizations.

To make comparison easy, considered setups are divided into four groups. The first group includes equalized TR techniques as shown in Fig. 10. Note that to provide a consistent measure for comparing different setups, the bit energy $E_b$ in the abscissa of the BER curves represents the pilots energy at the transmitter array (which is equal for all setups) rather than the bit energy at the receiver (which varies from one setup to the other). Since these setups are equalized (have negligible ISI), the BER performance at the receiver depends on the CIR peak value at the receiver. That is why the full-aspect case exhibits better performance. At the eavesdropper, however, the CIR is not perfectly equalized. Therefore, the performance at high data rate is hampered by the internal noise due to ISI, especially for low external noise levels (high $E_b/N_0$), as evident from the BER saturation in Fig. 10(b). At low data rates, the effect of ISI is less pronounced, especially in free-space, which allows the eavesdropper to intercept the signal.

The second group compares conventional TR and conventional beamforming with equalized TR as shown in Fig. 11. At high data rates, the performance of equalized TR is superior to other techniques. On the contrary, at low data rate the performance is controlled by the power level at the receiver. Therefore, when ISI at the receiver is not the limiting factor, conventional TR operating in rich scattering environments yields the best performance since it provides the highest CIR peak at the receiver and the highest ISI at the eavesdropper.
Fig. 10: BER performance of equalized TR setups. For high data rate: (a) at the intended receiver, (b) at the eavesdropper. For low data rate: (c) at the intended receiver, (d) at the eavesdropper.

The third group compares TR beamforming with MUSIC techniques with equalized TR as shown in Fig. 12. Using MUSIC significantly increases the BER at the eavesdropper (thus increases coveryness) especially at low data rates. However, because of the reduced CIR peak provided by equalized MUSIC, it yields unsatisfactory high BER at the receiver for both high and low rates. Only at very high $E_b/N_0$ (beyond the limits of Fig. 12(a)), equalized MUSIC performance can outperform that of non-equalized MUSIC, which saturates at high $E_b/N_0$ due to ISI.

Finally, the fourth group compares differential TR techniques with equalized TR. As shown in Fig. 13(a) and (c), both differential TR techniques yield satisfactory performance at the receiver, comparable with that of equalized TR. At the eavesdropper, differential TR MUSIC serves to increase the coveryness by increasing the BER, but with slightly less efficiency than TR MUSIC in Fig. 12, as expected.
Fig. 11: BER performance of conventional beamforming and conventional TR. For high data rate: (a) at the intended receiver, (b) at the eavesdropper. For low data rate: (c) at the intended receiver, (d) at the eavesdropper.

D. Effect of Hardware Limitations

So far, we have studied the performance of different techniques without considering any practical limitations that may be imposed by the utilized hardware. In practice, the time allocated for recording the steering vector of the receiver might not be long enough to capture all multiple scatterings in rich scattering media [30]. This produces truncated pilots, and therefore imperfectly equalized equivalent CIR. To demonstrate this effect, consider for example the case of equalized TR technique using linear array operating in the presence of surrounding walls. A typical response at the receiver array for a BH pulse derivative transmitted by the receiver in shown in Fig. 14(a). The recorded signal is truncated at 34 ns which corresponds to 90% of the total energy. The CIRs in the frequency domain for both truncated and full responses are plotted in Fig. 14(b). Imperfect equalization is evident as a result of truncation. This increases the ISI by 100%.

Another limitation may arise from the dynamic range of the array transceivers. Pilots energy distribution among array elements for different setups are plotted in Fig. 15. Most techniques result in almost uniform energy distribution among array elements with dynamic range less than
5 dB. Equalized TR MUSIC, however, requires a stringent power distribution with dynamic range of 25 dB. This problem can be mitigated by muting array elements whose energies fall below a certain threshold.

V. Multiuser MIMO Configuration

In multiuser MIMO configurations, the array attempts to communicate with multiple receivers simultaneously. If the interuser interference (IUI) among receivers is adequately mitigated, communication with different users can be achieved in the same time and frequency slots, which is the principle of space division multiplexing. The capacity of the MIMO system is mainly limited by the amount of co-channel interuser interference (IUI) [5], [31], [8], [32], [15]. In this section we will show how TR beamforming with MUSIC can be used to mitigate IUI.

Consider the presence of two receivers $A$ and $B$, the transmitted signal vector in the frequency domain can be written as

$$t(\omega) = S_A(\omega) p_A^*(\omega) + S_B(\omega) p_B^*(\omega)$$

\[ (10) \]
Fig. 13: BER performance of differential TR beamforming and MUSIC. For high data rate: (a) at the intended receiver, (b) at the eavesdropper. For low data rate: (c) at the intended receiver, (d) at the eavesdropper.

Fig. 14: Effect of receiver channel response truncation at the array. (a) A typical response at the receiver array for a BH pulse derivative transmitted by the receiver in the presence of walls. (b) CIR in the frequency domain at the intended receiver.

where $S_A(\omega)$ and $S_B(\omega)$ are the information signals to be transmitted to receiver $A$ and $B$ respectively. The received signal at receiver $A$ is given by

$$r_A(t) = s_A(t) * h_{AA}(t) + s_B(t) * h_{AB}(t) + n(t)$$ (11)
where $h_{AA}(t)$ is the CIR of receiver $A$ computed at receiver $A$, whereas $h_{AB}(t)$ is the CIR of receiver $B$ computed at receiver $A$. Similarly, the received signal at $B$ can be written as

$$r_B(t) = s_B(t) * h_{BB}(t) + s_A(t) * h_{BA}(t) + n(t)$$  \hspace{1cm} (12)$$

Obviously, the second terms in the r.h.s. of the above two equations represent undesired interuser interference. TR beamforming with MUSIC can be used to deliver the signal to the intended receiver while imposing null on the other receiver, therefore, it sets the cross terms $h_{AB}(t)$ and $h_{BA}(t)$ to zero, and the self terms become the CIR of the TR with MUSIC technique. As an example, consider the setup in Fig. 1, where the eavesdropper now represents receiver $B$. The BER performance of non-equalized TR with MUSIC is compared with that of equalized TR in Fig. 16. Note that at high data rate both techniques suffer from saturation with $E_b/N_0$. Equalized TR saturates because of IUI, whereas TR with MUSIC saturates because of ISI. Nevertheless, TR MUSIC still shows better performance especially at low data rates where IUI is stronger than ISI.

VI. CONCLUSION

We have extended existing TR-based wireless communication strategies by introducing three techniques that satisfy different performance criteria. We have applied them to both MISO and MIMO configurations using both linear and full-aspect arrays. BER performances of different techniques under various operational scenarios were compared for high and low data rates. Two main factors affect the BER: (i) ISI (internal noise), and (ii) received power level relative
Fig. 16: BER performance of MIMO setups. For high data rate: (a) at receiver $A$, (b) at receiver $B$. For low data rate: (c) at receiver $A$, (d) at receiver $B$.

to external noise. Equalized TR beamforming was introduced to eliminate the ISI that limits the performance at high data rates; however, it was shown to possess inferior power focusing compared with conventional TR. Focusing resolution was shown to depend significantly on the array configuration, where full-aspect arrays were capable of providing better focusing than linear arrays. TR beamforming using linear arrays typically does not allow for sufficient degree of covertness, and especially when an eavesdropper is closer to the array than the receiver. In this case, TR beamforming combined with MUSIC becomes very beneficial in producing null field at the eavesdropper location. TR beamforming with MUSIC is also useful in reducing the IUI in MIMO configurations, and therefore increasing the system capacity. In case of passive receivers that can not send pilots, approximate versions of the pilots can be obtained from sequential array acquisitions. Some effects of hardware limitations on the performance, such as truncated impulse response and transmitters dynamic range, were also considered.

REFERENCES


