MISO time reversal and delay spread compression for FWA channels at 5GHz

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***Abstract*-- Fixed wireless access channels in the 5GHz band have been measured with 8 element uniform linear antenna arrays at both ends, thus providing an 8x8 configuration. The measurements were performed at 3 different locations in downtown Toronto. The application of the time reversal (TR) technique in a multiple input-single output (MISO) can reduce the delay spread of the channel impulse response by a factor of 2-3, depending on the power allocation scheme.**

***Index terms*-- fixed wireless access, multiple antenna systems, time reversal.**

I. Introduction

Fixed wireless access (FWA) is an alternative to cable connection in places where fiber has not yet been deployed, or where very fast deployment or capacity upgrade is desired. Wireless optics provides coverage in line-of-sight (LOS) conditions, but in the more obstructed non-LOS case wireless radio connections have to be employed.

The modem designs that are suitable for secure and reliable high data rate operation depend on the characteristics of the radio channel. The delay spread of the channel limits the allowable data rates, as has been demonstrated in [1]. OFDM and equalization techniques have been proposed as strong contenders to counteract the effects of frequency selective fading. However such advanced modulation/ demodulation schemes come at the cost of hardware design and power consumption, and require accurate operation of the local oscillators. It would be desirable to build a system that is less sensitive to such effects, while keeping the complexity and therefore the cost of the receiver units low.

In this paper we present a scheme that is based on the concept of time-reversal (TR) [2]. While operable in wideband environments, this technique removes the complexity from the receiver and requires only a more complicated transmitter instead.

The paper is structured as follows. Section II describes the concept of TR in a MISO configuration, and Section III discusses the measurements. Section IV presents the cumulative results for the entire measurement set, and Section VI concludes this work.

II. MISO Time Reversal

Throughout this paper we discuss the complex base-band representation of the signals and assume that perfect sampling, filtering and down-conversion have been performed. For the purpose of our analysis we neglect the time variation of the channel. This simplification is reasonable for FWA situations, where the channel stays relatively constant.

If the channel impulse response between the transmitter and the receiver is a function $h(\tau)$ of the delay τ, the delay spread is given as the second moment of the average power delay profile pdp_h (averaged over the channel realizations) [3]

$$\left.\begin{aligned} pdp_h(\tau) &= \sqrt{E\left[|h(\tau)|^2\right]} \\ \sigma_h^2 &= \frac{1}{\|pdp_h\|^2}\int_{-\infty}^{\infty}(\tau-\bar{\tau})^2|pdp_h(\tau)|^2 d\tau \\ \bar{\tau} &= \frac{1}{\|pdp_h\|^2}\int_{-\infty}^{\infty}\tau|pdp_h(\tau)|^2 d\tau \end{aligned}\right\} \quad (1)$$

where $\|\bullet\|^2$ denotes the Frobenius norm of the argument $\bullet$ (in our case $\|\bullet\|^2 = \int_{-\infty}^{\infty}|\bullet|^2 d\tau$).

The configuration of the proposed MISO system is shown in Fig. 1, where there are M transmit antennas and one receive antenna. Let $g_m(t)$ denote the filter employed at the mth transmitter antenna, and $x(t)$ represent the data stream destined for the intended user. The signal $s_m(t)$ that is transmitted from antenna m can be written as

$$s_m(t) = g_m(t) \otimes x(t) \quad (2)$$

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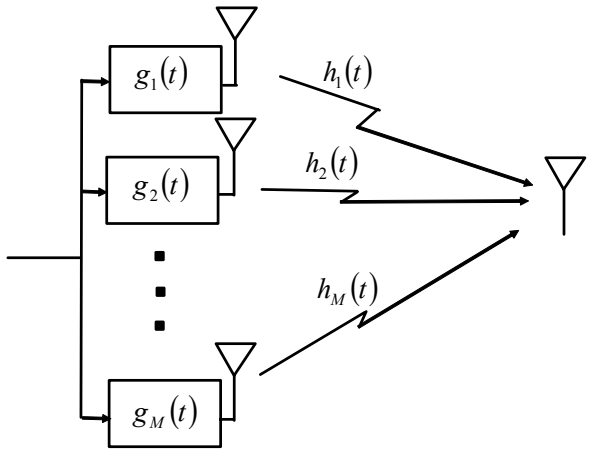

Figure 1: System diagram

The concept of applying filters at the transmitter is familiar in the context of pre-rake systems [4].

Let the filter g_m be given by the scaled time reversed version of the channel impulse response $h_m(t)$ between transmitter antenna m and the intended receiver, i.e. let

$$g_m(t) = A_m h_m^*(-t) \tag{3}$$

The implicit assumption is that the transmitters have channel state information (CSI), i.e. know the channel impulse response $h_m(t)$. This can be obtained via channel estimation on the uplink, or feedback of CSI from the intended receiver. Both techniques have been implemented in various forms in current systems (beam forming and adaptive techniques are examples of cases where CSI is fed back). Their performance is inherently limited by the rate of change of the channel (in time for TDD systems or frequency for FDD systems). For the purpose of demonstration of principle, we assume that the transmitter has perfect CSI.

The signal received by the user is given by

$$y(t) = \sum_{m=1}^{M} h_m(t) \otimes s_m(t) + n(t) = \underbrace{\left(\sum_{m=1}^{M} h_m(t) \otimes g_m(t)\right) \otimes x(t)}_{\text{Signal}} + \underbrace{n(t)}_{\text{noise}} \tag{4}$$

In this equation the first term is the useful signal, whereas the second term represents receiver noise, which is assumed to be additive white Gaussian noise.

The equivalent channel impulse response $h_{eq}(t)$ is

$$h_{eq}(t) = \sum_{m=1}^{M} h_m(t) \otimes g_m(t) = \sum_{m=1}^{M} A_m h_m(t) * h_m^*(-t) = \sum_{m=1}^{M} A_m R_{h_m h_m}(t) \tag{5}$$

We observe that it is a scaled sum of the autocorrelations of the channel impulse responses $h_m(t)$.

Similarly to (1), we can define the delay spread of the equivalent channel impulse response by replacing h by h_{eq} as

$$\left.\begin{aligned} pdp_{h_{eq}}(\tau) &= \sqrt{E\left[\left|h_{eq}(\tau)\right|^2\right]} \\ \sigma_{h_{eq}}^2 &= \frac{1}{\left\|pdp_{h_{eq}}\right\|^2} \int_{-\infty}^{\infty} \left(\tau - \bar{\tau}\right)^2 \left|pdp_{h_{eq}}(\tau)\right|^2 d\tau \\ \overline{\tau_{h_{eq}}} &= \frac{1}{\left\|pdp_{h_{eq}}\right\|^2} \int_{-\infty}^{\infty} \tau \left|pdp_{h_{eq}}(\tau)\right|^2 d\tau \end{aligned}\right\} \tag{6}$$

We are going to investigate the effect of TR as a means of temporal compression. Our criterion is the reduction of the delay spread of the channel, i.e. $\sigma_{h_{eq}} / \sigma_h$. We would like this to be as low as possible.

The scaling factors A_m describe various power allocation situations. We assume that the TR transmission filters introduce a total gain of unity such that the transmit power depends only of the power of the signal x(t), and we are going to investigate three schemes that correspond to different practical system considerations.

a. Simple time reversal

Under the simplest form of time reversal, the factors A_m normalize the total filter gain and are given by

$$A_m^{\text{no PA}} = \frac{1}{\sqrt{\sum_{m=1}^{M} \left\|h_m\right\|^2}} \tag{7}$$

b. Equal power allocation

The equal power allocation scheme corresponds to the situation where the RF transmitted power per antenna is constrained by hardware limitations, and should therefore be kept constant. This scheme is the simplest in the sense that the scaling factors are individually adjusted at each transmitter antenna without knowledge of the other channel impulse responses.

$$A_m^{EP} = \frac{1}{\sqrt{M}\sqrt{\left\|h_m\right\|^2}} \tag{8}$$

c. Optimal power allocation

In discrete form, let the channel impulse responses be sampled every T_s seconds and let the maximum length of the impulse response be N_s samples (the impulse responses have compact support in the time interval $[0\ N_s T_s]$). Then the length of the autocorrelation functions $R_{h_m h_m}$ is $(2N_s-1)$ samples.

We define the vector **a** of the scaling factors, the vectors $\mathbf{r}_{mk}$ of the autocorrelations, and the matrix **R** as

$$\begin{aligned} \mathbf{r}_m &= \left[R_{h_m h_m}(-(N_s-1)T_s) \quad \cdots \quad R_{h_m h_m}(0) \quad \cdots \quad R_{h_m h_m}((N_s-1)T_s)\right]^T \\ \mathbf{R} &= \left[\mathbf{r}_1 \quad \cdots \quad \mathbf{r}_M\right] \\ \mathbf{a} &= \left[A_1 \quad \cdots \quad A_M\right]^T \end{aligned} \tag{9}$$

The equivalent channel impulse response can be written in vector form as $\mathbf{h}_{eq} = \mathbf{Ra}$.

Given that the correlations are complex conjugate symmetric and therefore $\overline{\tau_{h_{eq}}} = 0$, the delay spread can be calculated as

$$\sigma_\tau^2 = \frac{\|\mathbf{Th}_{eq}\|^2}{\|\mathbf{h}_{eq}\|^2} T_s^2 = \frac{\|\mathbf{TRa}\|^2}{\|\mathbf{Ra}\|^2} T_s^2 \tag{10}$$

where

$$\mathbf{T} = diag([-(N_s - 1) \quad \cdots \quad (N_s - 1)]) \tag{11}$$

Making the observation that the minimal delay spread channel is the Dirac delta function ($\boldsymbol{\delta} = [0 \quad \cdots \quad 1 \quad \cdots \quad 0]^T$), and that the errors (deviations from the delta function) are scaled by the matrix **T**, the weighted least squares problem of the estimation of the vector **a** that minimizes the delay spread can be expressed as: $\mathbf{TRa} = \boldsymbol{\delta}$ (12)

In this form, the problem is ill-posed. To guarantee the stability of the problem, the solution is given as

$$\mathbf{a}_{opt} = \lim_{\varepsilon \to 0} \left\{ ((\mathbf{T} + \varepsilon \mathbf{I})\mathbf{R})^+ \boldsymbol{\delta} \right\} \tag{13}$$

where **I** is the identity matrix of dimensions $(2N_s\text{-}1)\text{x}(2N_s\text{-}1)$, and $(\bullet)^+$ denotes the pseudo-inverse of the argument $\bullet$. Using the additional constraint of equal total transmit power

$$A_m^{opt} = \frac{\mathbf{a}_{opt}(m)}{\|\mathbf{Ra}_{opt}\|} \tag{14}$$

III. Measurements

A. Measurement equipment

The measurements were taken with a system comprising 8 transmitters and 8 receivers by Avendo Wireless[1]. The modem operated in CDMA mode to provide simultaneous measurements of all 8x8=64 channel links. The carrier frequency was 5 GHz, and the bandwidth of the measurements was 20MHz. The transmit power was set to 100mW.

B. Transmitting/ Receiving arrays

8 element uniform linear arrays (ULA) were used at each end of the communication link, with a spacing of 3cm (λ/2) between adjacent elements. The arrays were vertically polarized and either vertically (V) or horizontally (H) oriented. Therefore there were 4 combinations of transmitter/ receiver orientation: a. V_{tx} - V_{rx}, b. H_{tx} – H_{rx}, c. V_{tx} – H_{rx} and d. H_{tx} - V_{rx}. For reasons of illustration of principle, in this paper we present results from measurements with horizontally oriented transmitter and receiver arrays only.

C. Measurement environment

The measurement campaign was performed in downtown Toronto. Figure 2 shows the layout of the measurement locations.

[1] Currently Waverider.

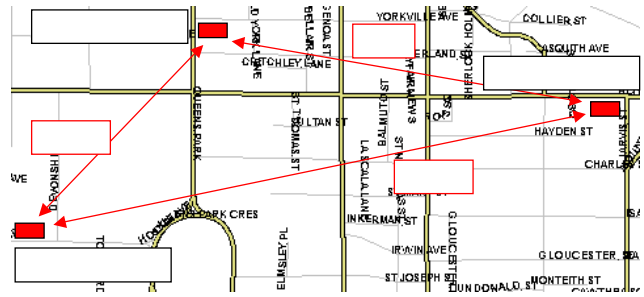

Figure 2: Measurement location layout in downtown Toronto

Location A is a hotel (Four Seasons) that has 29 floors with balconies facing SW and NE. Measurements were taken from the 5th, the 19th and the 28th floor. Location B is an apartment building with 28 stories and measurements were taken on the 19th (with the receiver facing West) and the 29th floor (roof). Location C is the Chemistry Faculty building of the University of Toronto. It is a 7-storey building and measurements were taken from the rooftop.

The link between locations A and B has a length of approximately 2km, and is blocked by a cluster of tall buildings (25-30 floors) approximately midway. The link between locations A and C is approximately 1.2 km long. In the direction of building C, location A is heavily obstructed in the near proximity (50m) by a tall building (Hyatt Hotel), whereas the area surrounding location C is clear. The link between locations B and C is 2.6km, and the environment is similar to the link between A and B.

D. Measurement process

For every measurement location, 512 time consecutive records of the channel impulse were taken for all 8x8 transmitter-receiver pairs simultaneously. The time difference between two consecutive channel records was 26.2112ms. The 512 time records were averaged over in order to derive the power delay profile of the channel as in (1). In order to eliminate the measurement noise, the measurements were clipped in the post-processing phase at twice the noise variance. This clipping level guarantees that 90% of the noise samples have been suppressed (under the assumption of additive white Gaussian noise).

These measurements were originally presented in [5] in the context of their MIMO potential.

IV. Delay Spread Compression Results

It is interesting to observe that the measurements were taken with multiple antenna arrays at each end of the communication link, and that the channel is reciprocal in either direction of transmission. Hence, TR can conceivably be applied in either direction of the communications link. TR filters were used on all 8 antennas of the acting transmitting array, and the equivalent channel impulse response on the intended receiver was then averaged over the 8 such possible intended receivers. We study the efficiency of TR depending on the scattering situation around the acting transmitters and receivers. The cluttered to clear curves describe the situation where the intended

receiver is in a lower scattering environment than the acting transmitter array, whereas the clear to cluttered curves show the results when the intended receiver is in the clutter. Figure 3 shows the efficiency of TR for the various power allocation schemes described in II.

As a general comment, the dissimilarity of the clear-to-cluttered and cluttered-to-clear curves does not violate the reciprocity principle: Reciprocity holds for point-to-point connections (one transmitter-one receiver), whereas the links under investigation are multi point-to-point (several transmitters- one receiver).

If pure TR is used, the perceived delay spread of the channel is reduced by a factor of roughly 2. The advantage of applying MISO TR increases with the separation between the two ends of the communication link when the intended receiver is in the clear and the acting transmitters are in the higher clutter situation. The opposite occurs when the receiver is in the cluttered situation and the acting transmitters in the clear. This trend becomes pronounced when large distances are taken into consideration, and can easily be explained in terms of the shower-curtain effect.

If the transmitted power per antenna is kept constant, the benefit of TR is reduced relative to the pure TR case but is still in the order of a factor of 2. Moreover, it does not clearly depend on distance in the clear-to-cluttered or cluttered-to-clear cases, but stays relatively constant.

If the antenna weights are optimally adjusted, a significant reduction in delay spread is observed. For both the clear-to-cluttered and cluttered-to-clear cases, an improvement of a factor of 3 is observed, and is maintained for all distances.

Independently of the power allocation scheme, MISO TR is an efficient means of reducing the perceived delay spread of the channel by a factor of 2 on average. With optimal power allocation, the advantage increases to a factor of 3.

V. Conclusions and Future Work

MISO TR has been demonstrated here as a powerful way to decrease the perceived delay spread of the signal. Experimental data in an urban environment for a bandwidth of 20MHz, under various separation and scattering situations, show a delay spread reduction of up to a factor of 3. This advantage can be used to simplify the receiver design and reduce the device cost. Increasing the channel bandwidth can result in further delay spread reduction, as demonstrated in [6], although ultra wideband operation comes at a loss of range.

The application of TR techniques requires knowledge of the channel transfer function at the transmitter location. This can be achieved with the use of CSI feedback (if the intended receiver itself estimates the channel), or with knowledge obtained in the uplink. The efficiency depends on the accuracy of this knowledge, and, obviously, more advanced power allocation schemes suffer more from outdated channel state information.

Conceivably TR can be extended to multi-user operation, whereby signals to different receive antennas go through the corresponding TR filters and get transmitted. In that case, the performance depends on the resulting interference levels.

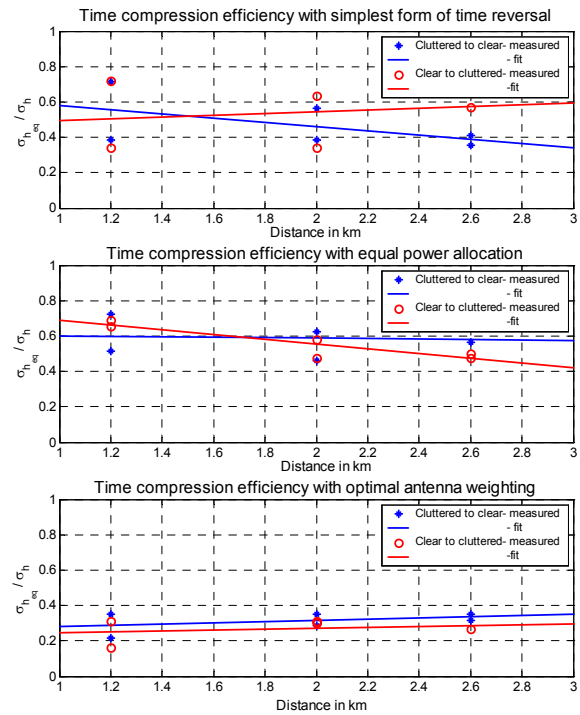

Figure 3: Delay spread compression for various power allocation schemes (Optimal weighting achieves a delay spread reduction by a factor of 3, while simpler schemes achieve delay spread reduction by a factor of 2)

VI. Acknowledgments

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