SPACE-TIME PROCESSING FOR WIRELESS COMMUNICATIONS

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ABSTRACT

This paper reviews space-time signal processing in mobile wireless communications. Space-time processing refers to the signal processing performed in the spatial and temporal domain on signals received at or transmitted from an antenna array, in order to improve performance of wireless networks. We focus on antenna arrays deployed at the base stations since such applications are of current practical interest.

1. INTRODUCTION

Cellular radio signal processing includes modulation and demodulation, channel coding and decoding, equalization and diversity combination. Current cellular modems do not address the problem of co-channel interference that arises from cellular frequency re-use, limiting the quality and capacity of wireless networks. Improved modem technology that can combats co-channel interference can have significant impact on overall network performance. One way to mitigate co-channel interference is to exploit the spatial dimension and design modems that operate with multiple antennas in receive and in transmit. Such *space-time (ST)* processing can improve network capacity, coverage and quality by reducing co-channel interference while enhancing diversity and array gain.

A space-time receive modem operates simultaneously on all the antennas, processing signal samples both in space and time. This extra dimension enables interference cancelation in a way that is not possible with single antenna modems. The desired signal and co-channel interference almost always arrive at the antenna array (even in complex multipath environments) with distinct and well separated spatial signatures, thus allowing the modem to exploit this difference to reduce co-channel interference. Likewise, the space-time transmit modems can use spatial selectivity to deliver signals to the desired mobile while minimizing interference for other mobiles.

The spatial dimension can also be used to enhance other aspects of modem performance. In the receiver, the antennas can be used to enhance array gain and improve signal to thermal noise ratio, increase diversity gain and even suppress inter-symbol interference. In the transmitter, the spatial dimension can enhance array gain, improve transmit diversity and reduce delay spread at the subscriber unit.

The paper is organized as follows. In Section 2 we present a description of a large (macro) cell propagation channel. In Section 3, we develop a signal model incorporating channel effects. In Section 4 we discuss different algorithmic approaches to reverse-link ST processing. Finally, Section 5 contains a summary of the paper.

2. THE WIRELESS CHANNEL

The propagation of radio signals on both the forward (base-station to subscriber unit) and reverse (subscriber unit to base-station) links is affected by the physical channel in several ways. In this section we review such effects and develop a detailed model to describe channel behavior. Multipath propagation results in dispersion of the signal in different dimensions. These are the delay (or time) spread, Doppler (or frequency) spread and angle spread. These spreads have significant effects on the signal, and are described below.

2.1. Space-Time channel model

A multipath channel is illustrated in Figure 3. Typical path amplitude, delay and fading statistics can be obtained from air interface standards. The signal from the mobile travels through a number of paths, each with its own fading and delay. [3] The fading can be either Rayleigh or Rician, and has a Doppler spectrum that is flat or classical. These paths arrive at the receive antenna array with different angles of arrival. The composite multipaths induce a different multipath channel at each antenna due to differences in relative phasing of the paths.

2.2. Doppler spread - time selective fading

Having seen the probability density of fading, we also need to understand its spectral characteristics. Fast fading results in a *Doppler spread*, i.e., a pure CW tone propagation in channel is spread over a non-zero spectral bandwidth. If one assumes uniformly distributed scatterers, then the baseband power spectrum of the vertical electrical field has the following form [3]:

$$S(f) = \frac{3\sigma}{2\pi f_m}\left[1-\left(\frac{f}{f_m}\right)^2\right]^{-1/2} \qquad (1)$$

where $f_m = v/\lambda$ is the maximum Doppler shift (v is the mobile velocity). The Doppler spectrum described by (1) is often called *classical* spectrum.

Again, when there is a direct line-of-sight path, the above spectrum is modified by an additional line at frequency corresponding to relative velocity between the base and the mobile. The spectrum will now be

$$S(f) + B\delta(f - f_D) \qquad (2)$$

where B is a measure of direct to scattered path energy and f_D is the Doppler shift of the direct path. Doppler spread causes time selective fading and can be characterized by the *coherence time* of the channel. The larger the Doppler spread, the smaller the coherence time.

2.3. Delay spread - frequency selective fading

Due to the multipath propagation, several time-shifted and scaled versions of the transmitted signal will arrive at the receiver. Typically a double negative exponential model is observed: the delay separation between paths decreases exponentially with path delay, and the path amplitudes also fall off exponentially with delay. This spread of path delay is called *delay spread.* Delay spread causes frequency selective fading and is also measured in terms of *coherence bandwidth.* The larger the delay spread, the smaller the coherence bandwidth.

2.4. Angle spread - space selective fading

Angle spread on receive refers to the spread of arrival angles of the multipaths at the antenna array. Likewise, angle spread in transmit refers to the spread of departure angles of the multipaths. The angle of arrival (or departure) of a path can be statistically related to the path delay. Using a constant delay ellipse scattering model, it can be shown that angle spread is proportional to delay spread and inversely proportional to the transmitter-receiver separation. Angle spread causes space selective fading and is characterized by the *coherence distance.* The larger the angle spread, the shorter the coherence distance.

2.5. Multipath propagation in large cells

Multipath scattering underlies the three spreading effects described above. The Doppler spread requires subscriber unit motion as well. It is important to understand the types of scatterers and their contribution to channel behavior. Our description below refers to the reverse link channel but applies equally well to the forward channel.

2.5.1. Scatterers local to mobile

Scattering local to the mobile is caused by buildings in the vicinity of the mobile (a few tens of meters). Mobile motion and local scattering give rise to Doppler spread which causes time-selective fading. For a mobile traveling at 65 mph, the Doppler spread is about 400 Hz in the 1900 MHz band. While local scattering contributes to Doppler spread, the delay spread will usually be insignificant because of the small scattering radius. Likewise, the angle spread will also be small.

2.5.2. Remote scatterers

The wavefront emerging from the local scatterers may travel directly to the base or get scattered towards the base by remote *dominant scatterers,* giving rise to specular multipath. These remote scatterers can be terrain features or high rise building complexes. Remote scattering can cause significant delay and angle spreads.

2.5.3. Scatterers local to base

Once these multiple wavefronts reach the base station, they may be scattered further by local buildings or other structures that are in the vicinity of the base. Such scattering will be more pronounced for low elevation and below roof-top antennas. The scattering local

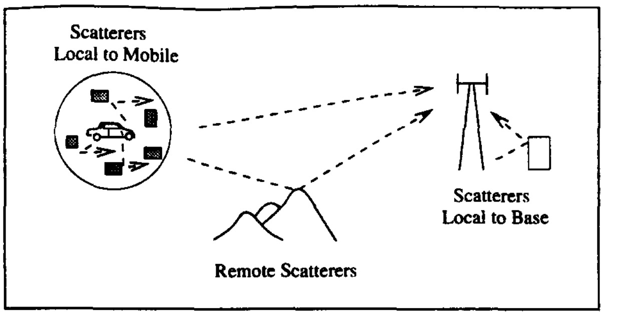

Figure 1. Multipath propagation has three distinct scattering sources, each of which gives rise to different channel effects.

to the base can cause severe angle spread which in turn, causes space-selective fading. This fading is time invariant, unlike the time varying space-selective fading caused by remote scattering.

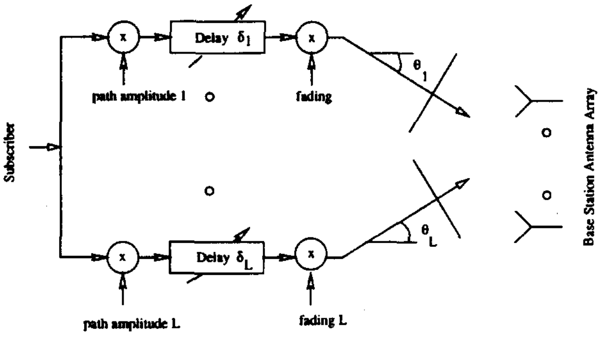

Figure 2. Multipath model

The forward link channel is affected in similar ways by these scatterers, but in a reverse order.

A typical example of a GSM macro-cellular channel in a hilly terrain is shown in Figure 3. We plot the frequency response at each antenna. Since the channel bandwidth is high (200 kHz), the channel is highly frequency-selective in a hilly terrain environment where delay spreads can reach 10 to 15 μ secs. Also, the large angle spread causes variations of the channel from antenna to antenna. The channel variation in time depends upon the Doppler spread. Note that since GSM uses a short time slot, the channel variation during the time slot is negligible.

2.6. Forward Link Model

The principle of reciprocity implies that the channel is identical on the forward and reverse links as long as the channel is measured at the same frequency and time instants (for time varying channels). In time division duplexing (TDD) systems, the principle of reciprocity clearly applies as long as the "ping-pong" time is small compared to the channel coherence time. In frequency division duplexing (FDD) systems (most wireless systems are FDD), the separation between the forward and reverse link frequencies is about 5 % of the mean frequency. One can assume that the paths used by both links are identical. Therefore the number of paths, the path delays and paths angles (arrival/ departure) are the same for both links. However the path amplitudes and phases will not be the same on the two links in FDD and they will in fact be different for the forward and reverse links. Therefore both in FDD and in TDD, there are limitations to predicting the forward link from the reverse link. The direct approach to estimating the forward link is by using feedback from the mobile. The channel is probed using multiple excitation inputs and the channel can be estimated from data fed back from the mobile.

3. RECEIVED SIGNAL MODEL

The noiseless baseband signal transmitted by a single user and received at the base station with multiple antennas can be expressed as

$$\mathbf{x}(t) = \sum_{l=1}^{L} \mathbf{a}(\theta_l)\alpha_l^R(t)u(t-\tau_l) \quad (3)$$

If we have signals from multiple users arriving at the antenna array, the total received signal is given by

$$\mathbf{x}(t) = \sum_{q=1}^{Q}\sum_{l=1}^{L_q} \mathbf{a}(\theta_{lq})\alpha_{lq}^R(t)u_q(t-\tau_{lq}) \quad (4)$$

where we have indexed each user signal, and corresponding path delay, angle and fading parameters by the user index q.

3.1. Discrete time signal model

The channel model described above uses physical parameters such as path gain, delay and angle of arrival. When the received signal is sampled at the symbol (or higher) rate, it is more convenient to use a "symbol response" channel model.

Such a discrete-time signal model can be easily obtained as follows. Let the continuous-time output from the receive antenna array $\mathbf{x}(t)$ be sampled at the symbol rate at instants $t = t_o + kT$.

Then the output may be written as

$$\mathbf{x}(k) = \mathbf{H}\mathbf{s}(k) + \mathbf{n}(k) \quad (5)$$

where $\mathbf{H}$ is the symbol response channel (an $m \times N$ matrix that captures the effects of the array response, symbol waveform and path fading - m is the number of antennas and N is the channel length). $\mathbf{H}$ is assumed to be time invariant by making the fading α^R constant. $\mathbf{s}(k)$ containing the transmitted symbols is defined as

$$\mathbf{s}(k) = \begin{bmatrix} s(k) \\ \vdots \\ s(k-N+1) \end{bmatrix} \quad (6)$$

The $(ij)^{th}$ element of the matrix is given by

$$[\mathbf{H}]_{ij} = \sum_{l=1}^{L} a_i(\theta_l)\alpha_l^R g((M_d+\delta-j)T-\tau_l), \quad i=1\ldots,m \ ; \ j=1,\ldots,N \quad (7)$$

where M_d is the maximum path delay and $2\delta T$ is the waveform duration.

The formulation (5) allows the use of several techniques developed for linear systems. Note that (7) shows how the multipath model can be used to find $\mathbf{H}$. Determining the multipath parameters from $\mathbf{H}$ is more complex and is discussed in Section 6.

It is often convenient to handle signals in blocks. Therefore we may collect M consecutive snapshots of $\mathbf{x}(\cdot)$ corresponding to time instants $k,\ldots,k+M-1$, and neglecting interference, we get

$$\mathbf{X}(k) = \mathbf{H}\mathbf{S}(k) + \mathbf{N}(k) \quad (8)$$

3.2. ST algorithms

We begin with the single-user case where we are only interested in demodulating one signal of interest. We therefore treat interference from other users as unknown additive noise.

3.2.1. ML and MMSE

One criterion for optimality used in ST processing is Maximum Likelihood (ML). *Maximum Likelihood Sequence Estimation* (MLSE) seeks to estimate the data sequence which is most likely to have been sent given the received signal vector. Another frequently used criterion is *Minimum Mean Square Error* (MMSE). In MMSE we obtain an estimate of the transmitted signal as a space-time weighted sum of the received signal and seek to minimize the mean square error between the estimate and the true signal. We present MLSE and MMSE in a form which is a ST extension of the well known MLSE and MMSE algorithms.

MLSE With the channel model described by Eq. (8), we assume that the noise $\mathbf{N}$ is spatially and temporally white and Gaussian, and that there is *no interference*. The MLSE problem reduces to finding $\mathbf{S}$ so as to satisfy the following criterion

$$\min_{\mathbf{S}} \|\mathbf{X} - \mathbf{H}\mathbf{S}\|_F^2 \quad (9)$$

where the channel $\mathbf{H}$ is assumed to be known and $\|.\|_F$ denotes Frobenius norm. This is a generalization of the standard MLSE problem, as now the channel is defined in space and in time. We can therefore use a ST generalization of the well known Viterbi algorithm (VA) to carry out the search in Eq. (9) efficiently.

MMSE In applications where interference is present, we will not, in general, have knowledge of the interference statistics. We can no longer use MLSE and one alternative approach is MMSE. In MMSE we seek to find a space-time filter (STF) that linearly combines the array output such that the difference between its scalar output and the true signal is minimized. The MMSE criterion is therefore

$$\min_{W} E\left\|W^H X(k) - s(k-D)\right\|_2^2 \quad (10)$$

where D is a delay chosen to center the STF. The solution to this LS problem follows from the well known projection theorem

Defining the space-time $mN \times mN$ covariance matrix $\mathbf{R}_{XX} = E\left(XX^H\right)$, W is given by

$$W = \mathbf{R}_{XX}^{-1}\overline{H} \quad (11)$$

If we explicitly calculate $\mathbf{R}_{XX}^{-1}$ and then use $\overline{H}$ to find W, the approach is called Sample Matrix Inversion (SMI). There are a number of trade-offs between SMI, LMS, and RLS approaches that span numerical stability, convergence speed, and computational complexity.

3.2.2. *Blind and non-blind methods*

In MLSE and MMSE as proposed above we assume that the channel $\mathbf{H}$ is known. In practice $\mathbf{H}$ may not be known, but is estimated by using training signals. An alternative approach consists of "blind" methods that do not use training signals, and instead use the properties of the received signal to determine $\mathbf{H}$ and $\mathbf{S}$. We briefly describe some such properties here. [4]

- Toeplitz: This is a powerful structure that captures the underlying convolution processes, usually constraining $\mathbf{X}$, $\mathbf{H}$ or $\mathbf{S}$ to be of block Toeplitz form.
- Low Rank: This property arises from oversampling in space or time. It constrains $\mathbf{H}$ to be a tall, full column rank matrix, and $\mathbf{X}$ to be low rank. (except for special conditions)
- Finite alphabet
  An important temporal structure in all mobile communication signals is *finite alphabet* (FA). The modulated signal is a linear or nonlinear map of an underlying set of finite alphabets.
- Constant modulus
  In many wireless applications, the transmitted waveform has a constant envelope (e.g., in FM modulation). An example of a constant-envelope waveform is the GMSK modulation used in the GSM cellular system.
- Space - Time Manifold
  We recall that $\mathbf{H}$ as defined in Eq. (7) includes $\mathbf{a}(\theta_l)$, the array response vector. While we do not know $\mathbf{a}(\theta_l)$, we often know the manifold on which $\mathbf{a}(\theta_l)$ lies. The so-called *array manifold* (AM), is a set of all possible array response vectors indexed by θ.
  Similarly, while $g((M_d{+}\delta{-}j)T - \tau_l)$ in Eq. (7) is not known, it is parameterized by τ_l. Therefore, $g((M_d{+}\delta{-}j)T - \tau_l)$ lies on a time manifold defined by the symbol waveform $g(t)$ and indexed by τ_l.

Figure 4 depicts a graphical form of the structures described above.

Blind algorithms exploit one or more of these structures. [5]

4. SUMMARY

Application of space-time processing is emerging as a powerful tool for improving performance of cellular wireless networks. STP can improve cell coverage, enhance link quality and increase system capacity. The rapidly fluctuating mobile channel with large multipath delay and angle spreads offers a significant challenge to space-time processing. Forward link space-time processing poses special challenges since the forward channel cannot be directly observed. Effective space-time processing solutions have to be tailored to specific air interface standard and propagation environments. More work is required to develop robust space-time processing techniques and characterize their performance. See [1, 2] for a review of the current state of the art in smart antennas technology.

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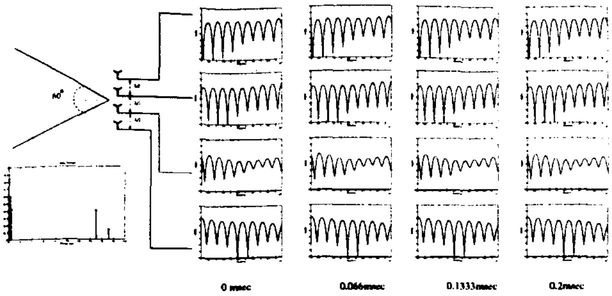

Figure 3. Channel frequency response at four different antennas for GSM in a typical hilly terrain at 1800 MHz. Mobile speed is 100 KPH. The response is plotted at four time instances spaced 1200 μ *secs*.

X	H	S
• Toeplitz • Low-rank	• Toeplitz • Full column rank • S-T Manifold	• Toeplitz • FA • CM

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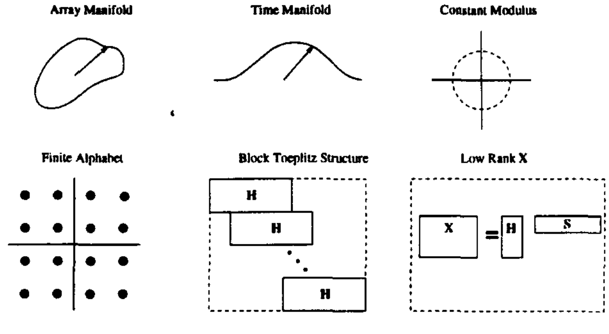

Figure 4. Signal structures for blind S-T processing