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MIMO-OFDM COMPARISON WITH RESPECT TO CHANNEL ASSESSMENT

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ABSTRACT

With the ever increasing number of wireless subscribers and their seemingly “greedy" demands for high-data-rate services, radio spectrum becomes an extremely rare and invaluable resource for all the countries in the world. Efficient use of radio spectrum requires that modulated carriers be placed as close as possible without causing any ICI and be capable of carrying as many bits as possible. Optimally, the bandwidth of each carrier would be adjacent to its neighbors, so there would be no wasted bands. In practice, a guard band must be placed between neighboring carriers to provide a guard space where a shaping filter can attenuate a neighboring carrier's signal. These guard bands are waste of spectrum. In order to transmit high-rate data, short symbol periods must be used. The symbol period *Tsym* is the inverse of the baseband data rate *R* ($R = 1 = Tsym$), so as *R* increases, *Tsym* must decrease. In a multipath environment, however, a shorter symbol period leads to an increased degree of ISI, and thus performance loss. OFDM addresses both of the two problems with its unique modulation and multiplexing technique. OFDM divides the high-rate stream into parallel lower rate data and hence prolongs the symbol duration, thus helping to eliminate ISI. It also allows the bandwidth of subcarriers to overlap without ICI as long as the modulated carriers are orthogonal. OFDM therefore is considered as a good candidate modulation technique for broadband access in a very dispersive environment**.**

KEYWORDS: Channel estimation, Transmission paths, normalization factor, gain capacity.

INTRODUCTION

Relying solely on OFDM technology to improve the spectral efficiency gives us only a partial solution. At the end of 1990s, seminal work by Foshini and Gans and, independently, by Teltar showed that there is another alternative to accomplish high-data-rate over wireless channels: the use of multiple antennas at the both ends of the wireless link, often referred to as MA (multiple antenna) or MIMO in the literature. The MIMO technique does not require any bandwidth expansions or any extra transmission power. Therefore, it provides a promising means to increase the spectral efficiency of a system. In his paper about the capacity of multi-antenna Gaussian channels, Telatar showed that given a wireless system employing N_t TX (transmit) antennas and N_r RX (receive) antennas, the maximum data rate at which error-free transmission over a fading channel is theoretically possible is proportional to the minimum of N_t and N_r (provided that the $N_t N_r$ transmission paths between the TX and RX antennas are statistically independent). Hence huge throughput gains may be achieved by adopting $N_t X N_r$ MIMO systems compared to conventional 1 *x* 1 systems that use single antenna at both ends of the link with the same requirement of power and bandwidth. With multiple antennas, a new domain, namely, the spatial domain is explored, as opposed to the existing systems in which the time and frequency domain are utilized.

Now let's come back to the previous question: what can be done in order to enhance the data rate of wireless communication systems? The combination of MIMO systems with OFDM technology provides a promising candidate for next generation fixed and mobile wireless systems. In practice for coherent detection, however, accurate channel state information in terms of channel impulse response (CIR) or channel frequency response (CFR) is critical to guarantee the diversity gains and the projected increase in data rate.

The channel state information can be obtained through two types of methods. One is called blind channel estimation, which explores the statistical information of the channel and certain properties of the transmitted signals. The other is called training-based channel estimation, which is based on the training data sent at the transmitter and known *a priori* at the receiver. Though the former has its advantage in that it has no overhead loss, it is only applicable to slowly time-varying channels due to its need for a long data record. Our work in this paper focuses on the training-based channel estimation method, since we aim at mobile wireless applications where the channels are fast time-varying. The conventional training-based method is used to estimate the channel by sending first a sequence of OFDM symbols, so-called preamble which is composed of known training symbols. Then the channel state information is estimated based on the received signals corresponding to the known training OFDM symbols prior to any data transmission in a packet. The channel is hence assumed to be constant before the next sequence of training OFDM symbols. Drastic performance degradation then arises if applied to fast time-varying channels. In optimal pilot-tone selection and placement were presented to aid channel estimation of single-input/single-output (SISO) systems. To use a set of pilot-tones within each OFDM block, not a sequence of training blocks ahead of a data packet to estimate the time-varying channel is the idea behind our work. However direct generalization of the channel estimation algorithm in to MIMO-OFDM systems involves the inversion of a high-dimension matrix due to the increased number of transmit and receive antennas, and thus entails high complexity and makes it infeasible for wireless communications over highly mobile channels. This becomes a bottleneck for applications to broadband wireless communications. To design a low-complexity channel estimator with comparable accuracy is the goal of this chapter.

The bottleneck problem of complexity for channel estimation in MIMO-OFDM systems has been studied by two different approaches. The first one shortens the sequence of training symbols to the length of the MIMO channel, as described in, leading to orthogonal structure for preamble design. Its drawback lies in the increase of the overhead due to the extra training OFDM blocks. The second one is the simplified channel estimation algorithm, as proposed in that achieves optimum channel estimation and also avoids the matrix inversion. However its construction of the pilot-tones is not explicit in terms of space-time codes (STC). We are motivated by both approaches in searching for new pilot-tone design. Our contribution in this chapter is the unification of the known results of in that the simplified channel estimation algorithm is generalized to explicit orthogonal space-frequency codes (SFC) that inherit the same computational advantage as in, while eliminating their respective drawbacks. In addition, the drastic performance degradation occurred in is avoided by our pilot-tone design since the channel is estimated at each block. In fact we have formulated the channel estimation problem in frequency domain, and the CFR is parameterized by the pilot-tones in a convenient form for design of SFC.As a result a unitary matrix, composed of pilot-tones from each transmit antenna, can be readily constructed. It is interesting to observe that the LS algorithm based on SFC in this paper is parallel to that for conventional OFDM systems with single transmit/receive antenna. The use of multiple transmit/receive antennas offers more design freedom that provides further improvements on estimation performance.

SYSTEM DESCRIPTION

Basically, the MIMO-OFDM transmitter has N_t parallel transmission paths which are very similar to the single antenna OFDM system, each branch performing serial-to-parallel conversion, pilot insertion, *N*-point IFFT and cyclic extension before the final TX signals are up-converted to RF and transmitted. It is worth noting that the channel encoder and the digital modulation, in some spatial multiplexing systems, can also be done per branch, not necessarily implemented jointly over all the N_t branches. The receiver first must estimate and correct the possible symbol timing error and frequency offsets, e.g., by using some training symbols in the preamble as standardized in. Subsequently, the CP is removed and N-point FFT is performed per receiver branch. In this thesis, the channel estimation algorithm we proposed is based on single carrier processing that implies MIMO detection has to be done per OFDM subcarrier. Therefore, the received signals of subcarrier k are routed to the k-th MIMO detector to recover all the N_t data signals transmitted on that subcarrier. Next, the transmitted symbol per TX antenna is combined and outputted for the subsequent operations like digital demodulation and decoding. Finally all the input binary data are recovered with certain BER. As a MIMO signaling technique, N_t different signals are transmitted simultaneously over $N_t X N_r$ transmission paths and each of those N_r received signals is a combination of all the N_t transmitted signals and the distorting noise. It brings in the diversity gain for enhanced system capacity as we desire. Meanwhile compared to the SISO system, it complicates the system design regarding to channel estimation and symbol detection due to the hugely increased number of channel coefficients.

SIGNAL MODEL

To find the signal model of MIMO-OFDM system, we can follow the same approach as utilized in the SISO case. Because of the increased number of antennas, the signal dimension is changed. For instance, the transmitted signal on the k-th subcarrier in a MIMO system is a $N_t X 1$ vector, instead of a scalar in the SISO case. For brevity of presentation, the same notations are used for both the SISO and MIMO cases. But they are explicitly defined in each case. There are Nt transmit antennas and hence on each of the N subcarriers, Nt modulated signals are transmitted simultaneously. Denote $\vec{s}\,(m)$ and $\vec{S}(mN+k\,)$ as the m-th modulated OFDM symbol in frequency domain and the k-th modulated subcarrier respectively as

$$\vec{s}\,(m) = \begin{bmatrix} \vec{S}\,(m) \\ \vdots \\ \vec{S}\,(mN+N-1) \end{bmatrix}$$

$$\vec{S}(mN+k) = \begin{bmatrix} S_1\,(mN+k) \\ \vdots \\ S_{N_t}(mN+K) \end{bmatrix} \tag{1.1}$$

Where $\vec{S}(mN+k)$ represents the k-th modulated subcarrier for the m-th OFDM symbol transmitted by the j-th antenna. And it is normalized by a normalization factor KMOD so that there is a unit normalized average power for all the mappings. Taking IFFT of $\vec{S}\,(m)$ as a baseband modulation, the resulting time-domain samples can be expressed as

$$\vec{S}\,(m) = \begin{bmatrix} \vec{S}\,(m) \\ \vdots \\ \vec{S}\,(mN+N-1) \end{bmatrix}$$

$$\vec{S}(mN+n\,) = \begin{bmatrix} S_1\,(mN+n) \\ \vdots \\ S_{N_t}(mN+n) \end{bmatrix} \tag{1.2}$$

$$=\frac{1}{N}(F_N^H \otimes I_{N_t})\,\vec{S}(m)$$

Here IFFT is a block-wise operation since each modulated subcarrier is a column vector and the generalized NN_t-point IFFT matrix is a Kronecker product of FN and IN_t. This is just a mathematical expression. In the real OFDM systems, however, the generalized IFFT operation is still performed by N_t parallel *N*-point IFFT. To eliminate the ISI and the ICI, a length-*Ng* (*Ng* >= *L*) CP is prepended to the time-domain samples per branch. The resulting OFDM symbol $\vec{u}(m)$ is denoted as

$$\vec{u}(m) = \begin{bmatrix} \vec{u}(mN_{tot}) \\ \vdots \\ \vec{u}(mN_{tot} + N_{tot} - 1) \end{bmatrix}$$

$$\acute{u}(mN_{tot} + n) = \begin{bmatrix} u_1(mN_{tot} + n) \\ \vdots \\ u_{N_t}(mN_{tot} + n) \end{bmatrix} \quad (1.3)$$

In a matrix form, there holds

$$\vec{u}(m) = A_{CP}\,\vec{S}(m) \quad (1.4)$$

Where

$$A_{CP} = \begin{bmatrix} 0 & I_{N_g} \\ I_{N-N_g} & 0 \\ 0 & I_{N_g} \end{bmatrix} \otimes I_{N_t}$$

The time-domain samples denoted by $\vec{u}(m)$ may be directly converted to RF for transmission or be up-converted to IF first and then transmitted over the wireless MIMO channel. For the MIMO channel, we assume in this thesis that the MIMO-OFDM system is operating in a frequency-selective Rayleigh fading environment and that the communication channel remains constant during a frame transmission, i.e., quasi-static fading. Suppose that the channel impulse response can be recorded with *L* time instances, i.e., time samples, then the multipath fading channel between the *j*-th TX and *i*-th RX antenna can be modeled by a discrete-time complex base-band equivalent (*L* - 1)-th order FIR filter with filter coefficients $h_{ij}(l, m)$, with $l \subseteq \{0, \dots, L-1\}$ and integer *m* > 0. As assumed in SISO case, these CIR coefficients $\{h_{ij}(0, m), \dots, h_{ij}(L-1, m)\}$ are independent complex zero-mean Gaussian RV's with variance $\frac{1}{2}P_l$ per dimension. The total power of the channel power delay profile $\{P_0, \dots, P_{L-1}\}$is normalized to be $\sigma_c^2 = 1$. Let h_m be the CIR matrix and denote $h_{l,m}$ as the *l*-th matrix-valued CIR coefficient.

$$h_m = \begin{bmatrix} h_{0,m} \\ \vdots \\ h_{L-1,m} \end{bmatrix}$$

$$h_{l,m} = \begin{bmatrix} h_{11}(l,m) & \cdots & h_{1N_t}(l,m) \\ \vdots & \ddots & \vdots \\ h_{N_r1}(l,m) & \cdots & h_{N_rN_t}(l,m) \end{bmatrix} \quad (1.5)$$

In addition, we assume that those $N_t N_r$ geographically co-located multipath channels are independent in environments full of scattering. In information-theoretic point of view, it guarantees the capacity gain of MIMO systems. For the practical MIMO-OFDM systems, it enforces a lower limit on the shortest distance between multiple antennas at a portable receiver unit. If the correlation between those channels exists, the diversity gain from MIMO system will be reduced and hence system performance is degraded.

At the receive side, an N_r-dimensional complex baseband equivalent receive signal can be obtained by a matrix-based discrete-time convolution as

$$\vec{r}\,(mN_{tot} + n) = \sum_{l=0}^{L-1} h_{l,m}\vec{u}(mN_{tot} + n - l) + \vec{v}\,(mN_{tot} + n) \tag{1.6}$$

Where

$$\vec{r}\,(mN_{tot} + n) = \begin{bmatrix} r_1\ (mN_{tot} + n) \\ \vdots \\ r_{N_r}\ (mN_{tot} + n) \end{bmatrix} \qquad \vec{v}\,(mN_{tot} + n) = \begin{bmatrix} v_1\ (mN_{tot} + n) \\ \vdots \\ v_{N_r}\ (mN_{tot} + n) \end{bmatrix}$$

Note that $\vec{v}\ (mN_{tot} + n)$ is assumed to be complex AWGN with zero mean and variance of $\frac{1}{2}\ \sigma_v^2$ per dimension. Therefore, the expected signal-to-noise ratio (SNR) per receive antenna is $\frac{N_t}{\sigma_t^2}$. In order to have a fair comparison with SISO systems, the power per TX antenna should be scaled down by a factor of N_t. By stacking the received samples at discrete time instances, $\vec{r}\ (m)$ can be described by

$$\vec{r}\,(m) = \begin{bmatrix} \vec{r}\,(mN_{tot} \\ \vdots \\ \vec{r}\,(mN_{tot} + N_{tot} - 1) \end{bmatrix}$$

To combat the ISI, the first $N_g N_r$ elements of $\vec{r}(m)$ must be removed completely. The resulting ISI-free OFDM symbol $\vec{y}(m)$ is

$$\vec{y}(m) = \begin{bmatrix} \vec{y}(mN) \\ \vdots \\ \vec{y}(mN + N - 1) \end{bmatrix} = A_{DeCP}\vec{r}(m),$$

Where

$$A_{DeCP} = [0\ \ I_N]\ \otimes\ I_{N_r}$$

By exploiting the property that $\vec{u}(m)$ is a cyclic extension of $\vec{s}(m)$so that cyclic discrete-time convolution is valid, the relation between $\vec{s}(m)$ and $\vec{y}(m)$ can be expressed as

$$\vec{y}\,(m) = h_{m,Cir}\vec{S}\,(m) + A_{DeCP}\vec{v}\,(m), \tag{1.8}$$

Where $h_{m,Cir}$ is a $NN_r\ X\ NN_t$ block circulant matrix. In general, a $NN_r\ X\ NN_t$ block circulant matrix is fully defined by its first $NN_r\ X\ NN_t$block matrices. In our case, $h_{m,Cir}$ is determined by

$$\begin{bmatrix} h_{0,m} \\ \vdots \\ h_{L-1,m} \\ 0_{(N-L)N_r \times N_t} \end{bmatrix}$$

Finally taking FFT on the $\vec{y}(m)$ at the receiver, we obtain the frequency domain MIMO-OFDM baseband signal model

$$\vec{Y}(m) = (F_N \otimes I_{Nr})\vec{y}(\mathrm{m})$$
$$= (F_N \otimes I_{Nr})(h_{m,Cir}\vec{S}(m) + A_{DoCP}\vec{v}(\mathrm{m}))$$
$$= (\tfrac{1}{N})((F_N \otimes I_{Nr})h_{m,Cir}(F_N^H \otimes I_{Nt})\vec{S}(\mathrm{m}) \qquad + ((F_N \otimes I_{Nr})A_{DeCP}\vec{v}(m)$$
$$= H_{m,diag}\vec{S}(m) + \vec{V}(m)$$

In the above expression, $\vec{V}(m)$resents the frequency domain noise, which is i.i.d. (independent and identically distributed) zero-mean and complex Gaussian random variable with variance $\frac{1}{2}\sigma_v^2$ per dimension, and $H_{m,diag}$ is a block diagonal matrix which is given by

$$H_{m,diag} = \begin{bmatrix} H_{0,m} & & \\ & \ddots & \\ H_{N-1,m} & & \end{bmatrix}$$

The *k*-th block diagonal element is the frequency response of the MIMO channel at the *k*-th subcarrier and can be shown to be $H_{k,m} = \sum_{l=0}^{L-1} h_{l,m}e^{-j\frac{2\pi}{N}kl}$. So for that subcarrier, we may write it in a simpler form

$$\vec{y}(mN + k) = H_{k,m}\,\vec{S}(mN + k) + \vec{V}(mN + k) \qquad (1.7)$$

Where

$$H_{k,m} = \begin{bmatrix} H_{11}(k,m) & \cdots & H_{1N_t}(k,m) \\ \vdots & \ddots & \vdots \\ H_{N_r1}(k,m) & \cdots & H_{N_rN_t}(k,m) \end{bmatrix}$$

This leads to a flat-fading signal model per subcarrier and it is similar to the SISO signal model, except that $H_{k,m}$ is a $N_r\ X\ N_t$ matrix.

CONCLUSIONS

Based on those assumptions such as perfect synchronization and block fading, we end up with a compact and simple signal model for both the single antenna OFDM and MIMO-OFDM systems. Surely it is an ideal model that says, considering first a noise free scenario, the received signal on the *k*-th subcarrier is just a product (or matrix product for MIMO case) of the transmitted signal on the *k*-th subcarrier and the discrete-time channel frequency response at the *k*-th subcarrier. Noise in frequency domain can also be modeled as an additive term. When it comes to channel estimation for OFDM systems, this model is still valid since there is no ICI as we assume.

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