

Agenda item:	17, MIMO
Source:	Lucent
Title:	Proposed MIMO channel metric for PARC system
Document for:	Discussion

1. INTRODUCTION

The spatial channel model ad hoc group (SCM AHG) has nearly completed the specification of the harmonized spatial channel model [1] to be used for MIMO discussions in both 3GPP and 3GPP2. With its completion, we can return to our discussion of MIMO system design issues.

One of the important outstanding MIMO issues is to determine what the channel metrics (channel quality indicator) should be. The channel metrics give a reliable indicator of the UE performance at a given data rate. In conventional single antenna HSDPA transmission, the channel metric is simply an estimate of the C/I. However in MIMO systems, a different metric is required to account for the interactions among the channels from different transmit antennas [2].

In this contribution, we propose a signal-to-noise-plus interference (SINR) metric for a per-antenna rate control (PARC) MIMO system [3]. We show the performance of this metric for flat fading channels and for dispersive channels modeled by the SCM text [1].

In Section 2, we review the PARC transmission and detection algorithms. In Section 3, we describe the proposed SINR channel metric. In Section 4, we show numerical results to demonstrate the feasibility of this metric.

2. REVIEW OF PARC TRANSMISSION AND DETECTION

A block diagram of PARC transmission is shown in Figure 1. The high-speed data stream is first demultiplexed among T transmit antennas. The number of bits assigned to each antenna may be different depending on the rate assignment. Following demultiplexing, the individual substreams for each antenna are coded, interleaved, and mapped to symbols. These symbols are further demultiplexed among C orthogonal spreading codes with spreading factor F . Note that in PARC transmission, code reuse occurs because a given code modulates data for all of the antennas. Different coding and modulation can be used for each transmit antenna. The goal of the metric is to accurately predict the frame error rate corresponding to a given data rate for each antenna.

At the receiver, we assume a minimum mean-squared error (MMSE) detector [4] without interference cancellation. This detector has been proposed as a baseline detector for HSDPA MIMO [5] because it has shown to provide more reliable performance in dispersive channels compared to a space-time rake and is relatively easy to implement. Figure 2 shows the MMSE detector. The received signal is a complex R -dimensional vector where R is the number of receive antennas. For the flat channel, the vector is multiplied by a T -by- R matrix representing the linear transformation, which minimizes the mean-squared error between the transmitted chip-level data symbols and the output of the transformation. Because the spreading codes are

orthogonal, the transformation is applied on each chip period, and its matrix is not dependent on the spreading codes. A generalization for the dispersive channel is described in the following section when discussing the proposed metric in detail. For each chip period, the output of the linear transformation is a T -dimensional vector. Letting F be the number of chips per symbol, we collect F consecutive output vectors corresponding to a given symbol, and the t th collection ($t = 1 \dots T$) of F samples are correlated with the C spreading codes. For the t th antenna, the N despread signals are multiplexed, symbols are detected, demapped, deinterleaved, and decoded.

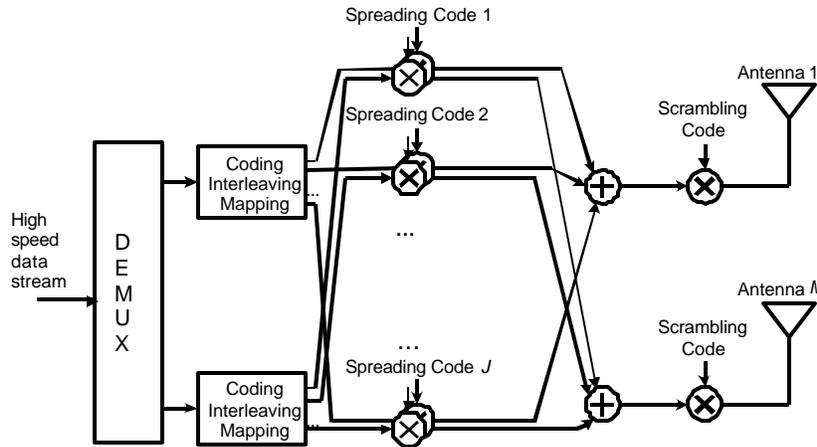


Figure 1. Block diagram of PARC transmitter

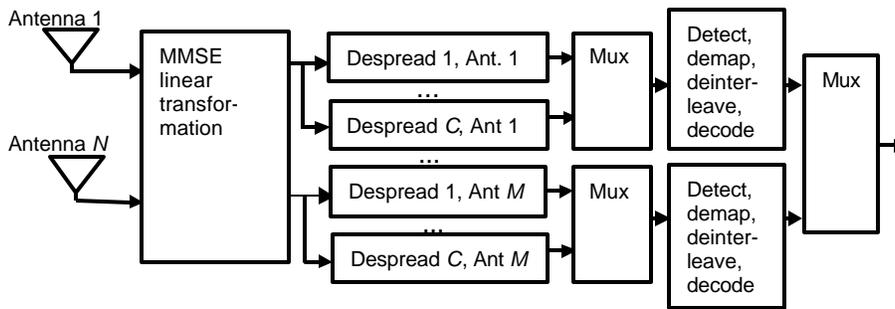


Figure 2. Block diagram of MMSE receiver

3. TRANSMITTED SIGNAL

A given block of information is demultiplexed into M lower rate streams, where M is the number of transmitters. Because of PARC, the number of information bits for each of these streams may be different. Each stream is coded, punctured, interleaved, mapped to symbols and demultiplexed into J equal-rate substreams where J is the number of orthogonal spreading codes of spreading factor F . Let $b_{j,m}$ denote the symbol from the m th antenna ($m = 1 \dots M$) spread by the j th code ($j = 1 \dots J$). For a given block of information, while the number of information bits for each of the M streams may be different, the number of coded symbols is the same. The transmitted signal from the m th antenna during this symbol period is

$$\mathbf{t}_m = \sum_{j=1}^J \mathbf{s}_j b_{j,m} = \mathbf{S} \mathbf{b}_m. \quad (1)$$

where \mathbf{s}_j is the j th spreading code, $\mathbf{S} \stackrel{\Delta}{=} [\mathbf{s}_1 \cdots \mathbf{s}_J]$ is the F -by- J spreading code matrix, and $\mathbf{b}_m \stackrel{\Delta}{=} [b_{1,m} \cdots b_{J,m}]^T$ is the vector of data from antenna m .

4. RECEIVED SIGNAL MODEL

We first consider a system with a single transmit antenna and single receive antenna. Let $\bar{\mathbf{x}} \stackrel{\Delta}{=} [x(1) \ x(2) \ x(3) \ \cdots]^T$ denote the vector of data (chip modulated data symbols) transmitted over a given frame. The components of $\bar{\mathbf{x}}$ correspond to the components of \mathbf{t}_1 in (1) taken over successive symbol intervals, and the length of $\bar{\mathbf{x}}$ is the number of chips per frame. Let L be the delay spread of the channel measured in units of the chip period, and let P be the oversampling factor. The channel coefficient corresponding to the l th chip and p th oversample is $h_p(l-1)$ ($l = 1 \dots L$, and $p = 1 \dots P$). The coefficients are obtained by convolving the channel impulse response with the transmit and receive pulse-shaping filters. We define $y_p(k)$ as the received signal sample obtained when the k th chip of $\bar{\mathbf{x}}$ is multiplied by the last sample $h_p(L-1)$ of the channel response:

$$\begin{aligned} y_p(k) &\stackrel{\Delta}{=} x(k)h_p(L-1) + x(k+1)h_p(L-2) + \dots + x(k+L-1)h_p(0) + n_p(k) \\ &= \sum_{j=0}^{L-1} x(k+j)h_p(L-1-j) + n_p(k) \end{aligned}$$

where $n_p(k)$ is the additive noise on the p th sample of the k th chip. This noise component is a zero-mean, complex Gaussian random variable with variance $\sigma_n^2/2$ per complex dimension. Let E be the span of the equalizer measured in units of the chip period, and let $\mathbf{x}(k)$ be the $(E+L-1)$ -dimensional subvector of $\bar{\mathbf{x}}$ starting with the k th term $x(k)$ and ending with the term $x(k+E+L-2)$. Then the received signal vector $\mathbf{y}(k)$ can be written as:

$$\underbrace{\begin{bmatrix} y_1(k) \\ \vdots \\ y_p(k) \\ \vdots \\ y_1(k+E-1) \\ \vdots \\ y_p(k+E-1) \end{bmatrix}}_{\mathbf{y}(k)} = \underbrace{\begin{bmatrix} h_1(L-1) & \cdots & h_1(0) \\ \vdots & & \vdots \\ h_p(L-1) & \cdots & h_p(0) \\ & & \ddots \\ & & h_1(L-1) & \cdots & h_1(0) \\ & & \vdots & & \vdots \\ & & h_p(L-1) & \cdots & h_p(0) \end{bmatrix}}_{\Gamma} \underbrace{\begin{bmatrix} x(k) \\ x(k+1) \\ x(k+2) \\ \vdots \\ \vdots \\ x(k+E+L-2) \end{bmatrix}}_{\mathbf{x}(k)} + \underbrace{\begin{bmatrix} n_1(k) \\ \vdots \\ n_p(k) \\ \vdots \\ n_1(k+E-1) \\ \vdots \\ n_p(k+E-1) \end{bmatrix}}_{\mathbf{n}(k)}$$

where $\mathbf{y}(k)$, Γ , $\mathbf{x}(k)$, and $\mathbf{n}(k)$ are respectively size PE -by-1, PE -by- $(E+L-1)$, $(E+L-1)$ -by-1, and PE -by-1.

To generalize this model for M transmit and N receive antennas, we let $\bar{\mathbf{x}}_m \stackrel{\Delta}{=} [x_m(1) x_m(2) x_m(3) \cdots]^T$ denote the vector of data transmitted over a given frame over the m th antenna ($m = 1 \dots M$) corresponding \mathbf{t}_m in (1). Let $\mathbf{x}_m(k)$ be the $(E + L - 1)$ -dimensional subvector of $\bar{\mathbf{x}}_m$ starting with the k th term $x_m(k)$ and ending with the term $x_m(k + E + L - 2)$. We define $h_{n,m,p}(l)$ as the channel coefficient between the n th transmitter ($n = 1 \dots N$) and m th receiver ($m = 1 \dots M$) corresponding to the l th chip and p th oversample. Let $\Gamma_{n,m}$ denote the channel matrix analogous to Γ composed of channel coefficients $h_{n,m,p}(l)$, $l = 0 \dots L - 1$, $p = 1 \dots P$. Let $n_{n,p}(k)$ be the additive Gaussian noise at the n th antenna on the p th sample of the k th chip. Defining $\mathbf{n}_n(k) \stackrel{\Delta}{=} [n_{n,1}(k) \cdots n_{n,p}(k) \cdots n_{n,1}(k + E - 1) \cdots n_{n,p}(k + E - 1)]^T$, the received signal vector at the n th antenna can be written

$$\mathbf{y}_n(k) = \sum_{m=1}^M \Gamma_{n,m} \mathbf{x}_m(k) + \mathbf{n}_n(k)$$

By stacking the received vectors and generalizing the definition of $\mathbf{y}(k) \stackrel{\Delta}{=} [\mathbf{y}_1^H(k) \cdots \mathbf{y}_N^H(k)]^H$, we can write

$$\underbrace{\begin{bmatrix} \mathbf{y}_1(k) \\ \vdots \\ \mathbf{y}_N(k) \end{bmatrix}}_{\mathbf{y}(k)} = \underbrace{\begin{bmatrix} \Gamma_{1,1} & \cdots & \Gamma_{1,M} \\ \vdots & & \vdots \\ \Gamma_{N,1} & \cdots & \Gamma_{N,M} \end{bmatrix}}_{\Gamma} \underbrace{\begin{bmatrix} \mathbf{x}_1(k) \\ \vdots \\ \mathbf{x}_M(k) \end{bmatrix}}_{\mathbf{x}(k)} + \underbrace{\begin{bmatrix} \mathbf{n}_1(k) \\ \vdots \\ \mathbf{n}_N(k) \end{bmatrix}}_{\mathbf{n}(k)}. \quad (2)$$

The matrix Γ and the vectors $\mathbf{x}(k)$ and $\mathbf{n}(k)$ have also been generalized and redefined for the multiple antenna case. The sizes of $\mathbf{y}(k)$, Γ , $\mathbf{x}(k)$, and $\mathbf{n}(k)$ are respectively PEN -by-1, PEN -by- $M(E+L-1)$, $M(E+L-1)$ -by-1, and PEN -by-1.

5. MMSE EQUALIZER DERIVATION

Let the components of $\mathbf{x}_d(k)$ be a M -dimensional vector whose m th component is the transmitted signal from the m th antenna with delay d with respect to sample k : $\mathbf{x}_d(k) \stackrel{\Delta}{=} [x_1(k+d) \cdots x_M(k+d)]^T$. Given $\mathbf{y}(k)$, the minimum mean-square error (MMSE) equalizer $\mathbf{W}_d \in \mathbb{C}^{M \times PEN}$ minimizes the mean-square error between the equalizer output $\mathbf{W}_d \mathbf{y}(k)$ and the desired M -dimensional output vector $\mathbf{x}_d(k)$: $E[\|\mathbf{W}_d \mathbf{y}(k) - \mathbf{x}_d(k)\|^2]$. Assuming that the noise $\mathbf{n}(k)$ and the desired vector $\mathbf{x}_d(k)$ are independent, the Wiener solution is given by

$$\begin{aligned}
\mathbf{W}_d &= E\left[\mathbf{x}_d(k)\mathbf{y}^H(k)\right]\left\{E\left[\mathbf{y}(k)\mathbf{y}^H(k)\right]\right\}^{-1} \\
&= E\left[\mathbf{x}_d(k)\mathbf{x}^H(k)\Gamma^H + \mathbf{x}_d(k)\mathbf{n}^H(k)\right]\left\{E\left[(\Gamma\mathbf{x}(k) + \mathbf{n}(k))(\Gamma^H\mathbf{x}^H(k) + \mathbf{n}^H(k))\right]\right\}^{-1} \\
&= \mathbf{s}_x^2 \mathbf{E}_d \Gamma^H \left\{ \mathbf{s}_x^2 \Gamma \Gamma^H + \mathbf{R}_n \right\}^{-1} \\
&= \mathbf{E}_d \Gamma^H \left\{ \Gamma \Gamma^H + \frac{1}{\mathbf{s}_x^2} \mathbf{R}_n \right\}^{-1}
\end{aligned} \tag{3}$$

where we have used

$$E\left[\mathbf{x}_d(k)\mathbf{x}^H(k)\right] = E\left\{ \begin{bmatrix} x_1(k+d) \\ \vdots \\ x_M(k+d) \end{bmatrix} \begin{bmatrix} \mathbf{x}_1^H & \cdots & \mathbf{x}_M^H \end{bmatrix} \right\} = \mathbf{s}_x^2 \underbrace{\begin{bmatrix} \mathbf{e}_d & \mathbf{z} & \cdots & \mathbf{z} \\ \mathbf{z} & \mathbf{e}_d & & \vdots \\ \vdots & & \ddots & \mathbf{z} \\ \mathbf{z} & \cdots & \mathbf{z} & \mathbf{e}_d \end{bmatrix}}_{\mathbf{E}_d} \tag{4}$$

where $\mathbf{s}_x^2 = E[x_m(k)x_m^*(k)]$ is the chip power (independent of antenna m and time k), $\mathbf{e}_d = [\underbrace{0 \cdots 0}_d 1 \underbrace{0 \cdots 0}_{E+L-2-d}]$ is a $E+L-1$ dimensional unit vector, \mathbf{z} is a $E+L-1$ dimensional vector of zeroes, \mathbf{E}_d is the M -by- $M(L+E-1)$ matrix defined above in (4), and $\mathbf{R}_n = E[\mathbf{n}(k)\mathbf{n}^H(k)]$ is the noise covariance matrix. If we assume that the noise is white and uncorrelated among antennas, $\mathbf{R}_n = \mathbf{s}_n^2 \mathbf{I}_{PNE}$ where \mathbf{I}_j denotes the j -by- j identity matrix. Note that because the MMSE equalizer is only dependent on the power of the chip sequences and not their actual values, \mathbf{W}_d is independent of the time index k . Hence it needs to be recomputed at the rate of significant channel variations. Using an alternative MMSE detector design which depends on the spreading sequences, the equalizer taps would have to be updated whenever either the spreading codes change or the channel changes significantly. In systems with long spreading codes, the spreading codes change ever symbol, resulting in an enormous computational burden. Note that for a single antenna system ($M = N = 1$), the equalizer in (3) reduces to the conventional single antenna MMSE equalizer for dispersive channels [5].

6. SINR CHANNEL METRIC

We write the MMSE equalizer matrix \mathbf{W}_d in terms of its row vectors:

$$\mathbf{W}_d = \begin{bmatrix} \mathbf{w}_{d,1} \\ \vdots \\ \mathbf{w}_{d,M} \end{bmatrix}$$

so that the minimum mean-squared error $E\left[|\mathbf{w}_{d,m}\mathbf{y} - x_m(k+d)|^2\right]$ ($m = 1, \dots, M$) is minimized.

We write the channel matrix Γ in terms of its $M(E+L-1)$ column vectors:

$$\Gamma = [\Gamma_{1,1} \cdots \Gamma_{1,d+1} \cdots \Gamma_{1,E+L-1} \cdots \Gamma_{M,1} \cdots \Gamma_{M,d+1} \cdots \Gamma_{M,E+L-1}].$$

Then the equalizer output for the m th transmit antenna can be written as

$$\begin{aligned} \mathbf{w}_{d,m} \mathbf{y} &= \mathbf{w}_{d,m} \sum_{j=1}^M \sum_{l=1}^{E+L-1} \Gamma_{j,l} x_j(k+l-1) + \mathbf{w}_{d,m} \mathbf{n} \\ &= \mathbf{w}_{d,m} \Gamma_{m,d+1} x_m(k+d) + \underbrace{\mathbf{w}_{d,m} \sum_{j=1, j \neq m}^M \sum_{l=1}^{E+L-1} \Gamma_{j,l} x_j(k+l-1) + \mathbf{w}_{d,m} \sum_{l=1, l \neq d+1}^{E+L-1} \Gamma_{m,l} x_m(k+l-1)}_{\text{interference}} + \mathbf{w}_{d,m} \mathbf{n}. \end{aligned}$$

The first term above is the desired signal term, the second term is the interference from other antennas, the third term is the self-interference from the m th antenna, and the last term is the contribution from filtered noise. Because the data symbols The signal to interference ratio (SINR) for the m th antenna is

$$SINR_m = \frac{|\mathbf{w}_{d,m} \Gamma_{m,d+1}|^2}{\sum_{j=1, j \neq m}^M \sum_{l=1}^{E+L-1} |\mathbf{w}_{d,m} \Gamma_{j,l}|^2 + \sum_{l=1, l \neq d+1}^{E+L-1} |\mathbf{w}_{d,m} \Gamma_{m,l}|^2 + M \mathbf{w}_{d,m} \mathbf{R}_n \mathbf{w}_{d,m}^H}. \quad (5)$$

In general, the channel metric should give a reliable prediction of the FER for a given channel realization. It should be used in the actual implementation of a MIMO system, and the information should be fed back to the Node B from the UE for rate determination. It should also be used in system level simulations so that link level performance in terms of FER can be predicted for a given channel realization. In order for the metric to give a reliable prediction of FER, different channel realizations which result in the same FER should be mapped to the same metric value. In the next section, we perform numerical simulations to demonstrate that this is indeed the case with the SINR metric.

7. NUMERICAL RESULTS

In this section we show the reliability of the SINR metric in predicting the FER for a given channel realization. We assume that the transmitter uses $T = 4$ antennas and the receiver uses $R = 4$ antennas. We consider 4QAM and 16QAM constellations and rate 1/2 and 3/4 coding for each transmit antenna.

To generate the performance of the metric, the following procedure is used:

1. For a given channel realization, the metric for the four transmit antennas is calculated according to (5).
2. Data rates for each antenna are chosen according to the metric. The minimum required SINR is shown in Table 1 below. For example, if the SINRs are 4.0, -1.0, 10.8, and 6.5 dB, then the respective coding and modulation formats are (4QAM, 1/2), (4QAM, 1/4), (16QAM, 3/4), and (16QAM, 1/2), respectively.
3. A link simulation is run for this channel realization and the assigned data rates over 100 or 200 frames respectively for the flat and dispersive channels. We assume that 80% of the power is for the data channels, and that 10 out of 16 codes are used. The resulting FER for the 4 branches are plotted versus SINR.

4. Steps 1 through 3 are repeated for different channel realizations.

We consider both flat and dispersive channels. The flat channel is characterized by the 4 -by-4 matrix \mathbf{H} whose elements are IID complex Gaussian random variables with zero mean and unit variance. The dispersive channel is taken from the suburban macrocell environment described in the SCM text [1]. This channel is characterized by spatial correlation, and 6 multipath components with random delays. The average SNR ($\mathbf{s}_x^2 / \mathbf{s}_n^2$) for the two channels is slightly different (12.5 dB for flat and 13.5 dB for dispersive) in order to get a uniform distribution of SINR per branch over the range of interest.

The FER versus SINR for single antenna transmission are shown for the 4 possible coding and modulation combinations as solid lines in Figures 3 and 4. For both channel types, the SINR metric is very accurate in predicting the FER performance over the entire range of SINR.

data rate (Mbps)	modulation	code rate	required SINR (dB)
2.4	QPSK	1/2	-Infinity
3.6	QPSK	3/4	3.5
4.8	16QAM	1/2	5.8
7.2	16QAM	3/4	9.5

Table 1. MCS and required SINRs

Channel type	Flat	Dispersive, SCM text
number of multipaths	1	6
spatial correlation	no	yes
average SNR $\mathbf{s}_x^2 / \mathbf{s}_n^2$	12.5 dB	13.5 dB

Table 2. Channel characteristics

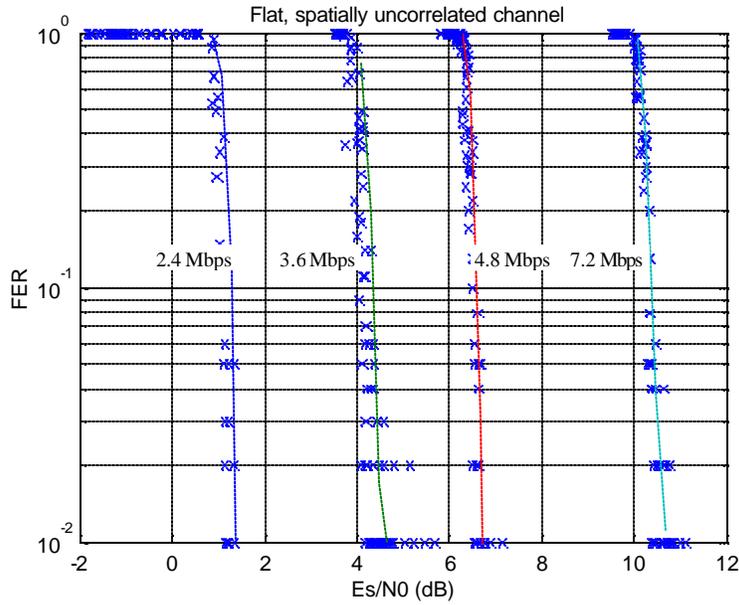


Figure 3. FER versus SINR metric for flat channel

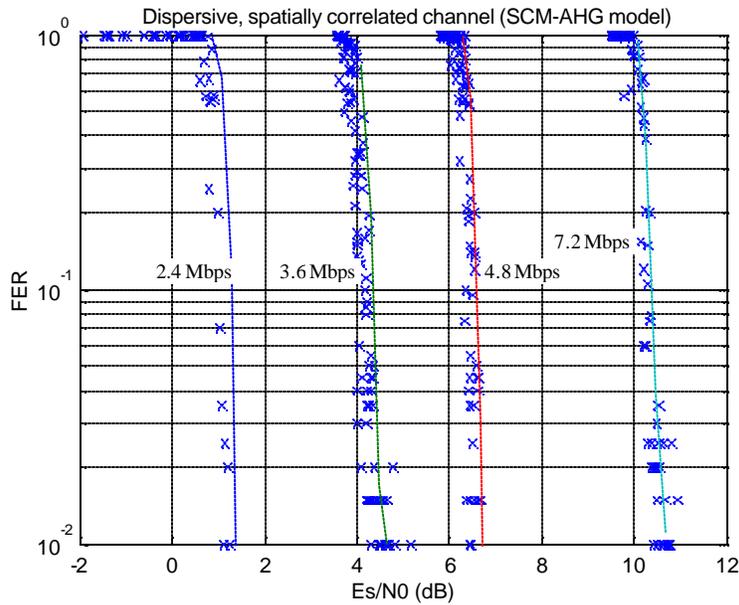


Figure 4. FER versus SINR metric for dispersive channel (SCM-AHG model)

8. CONCLUSIONS

An SINR metric has been proposed for predicting link FER performance in MIMO systems. It has been shown to be an extremely reliable predictor for a variety of coding and modulation schemes and a variety of channel conditions including dispersive and spatially correlated channels generated from the SCM AHG model. In the future we will show the performance of this metric using channel estimates and propose methods for accounting for doppler fading.

9. REFERENCES

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