systems was studied [Nikitopoulos, 2004]. Besides phase noise- time-selective fading also destroys the orthogonality among different subcarriers within one Orthogonal Freguency Division Multiplexing (OFDM) symbol and causes Intercarrier interference (ICI) [Russell, 1995]. Similar to single-antenna Orthogonal Freguency Division Multiplexing (OFDM), MIMO-OFDM is also vulnerable to channel time selectivity. Error performance of MIMO-OFDM systems in the presence of time-selective fading without considering phase noise was analysed [Stamoulis, 2002].

Although the issue caused by phase noise and time-selective fading in MIMO-OFDM has been recognized, the exact quantitative effect of the combination of the two has not been well addressed. Phase noise mitigation for MIMO-OFDM in fast time-varying fading environments has not been well studied either. In this chapter, It was analyzed, via mainly an analytical approach, the impact of phase noise to the performance of MIMO-OFDM systems over doubly-selective Rayleigh fading channels. After characterizing Common Phase error (CPE) caused by phase noise and Intercarrier Interference (ICI) caused by phase noise and time-selective fading, an MMSE-based mitigation scheme to effectively minimize the impact of phase noise was derieved. The author also compare four detection schemes, Zero forcing, ZF, MMSE, decorrelating division feeback (DF) and MMSE-DF schemes, and evaluate their SER performance.

II. METHODS AND MATERIAL

A. System Model

Consider a MIMO-OFDM system with N_t transmit antennas, N_r receive an-tennas, and N_s subcarriers in a doubly-selective Rayleigh fading environment. Input data are assumed to be independent variables with zero mean and unit variance. The time domain data sequence is obtained by taking the inverse discrete Fourier transform (IDFT) of the data block for each transmit antenna. A cyclic prefix (CP) with a length longer than the channel length is inserted at the beginning of each of the data sequences. The data sequences with a cyclic prefix (CP) are then trans-mitted through N_t independent antennas. At each receive antenna, the cyclic prefix (CP) is removed and a discrete fourier Transform (DFT) unit is applied. Let $x_k = [x_{k1},\ldots,x_{kNt}]^T$ and $y_k = [y_{k1},\ldots,y_{kNr}]^T$ denote, respectively, the transmitted and received data for all antennas on subcarrier k, where $0 \le k \le N_{s-}1$. The general form of the received signal in MIMO-OFDM over slowly fading channels (the channel is time-invariant over several Orthogonal Freguency Division Multiplexing (OFDM) symbol periods) (OFDM) signal are time and frequency synchronsed to each other, allowing the interference between subcarriers to be carefully controlled. These multiple subcarriers overlap in the frequency domain but do not cause inter-carrier interference (ICI) due to the orthogonal nature of the modulation.

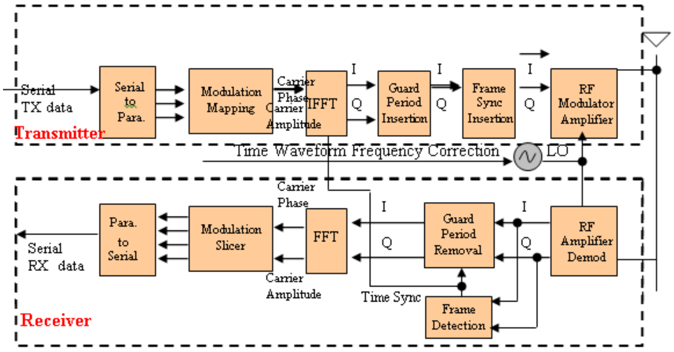

Fig 1a. AN OFDM TRANSCEIVER

B. Space-Time Coded Orthogonal frequency division multiplexing (OFDM) Transmitter

Consider a Mult-Input Multi-Output- Orthogonal frequency division multiplexing (OFDM) system with N_t transmit antennas, N_r receive an-tennas, and N_s subcarriers. The channel is frequency-selective Rayleigh fading and is modeled as quasi-static, allowing it to be constant over an orthogonal frequency division multiplexing (OFDM) block and change independently from one block to another.

Input symbol sequence $\{a(0),\ a(1),\ldots,a(N_tN_s\ -\ 1)\}$ is serial-to-parallel con-verted into N_t sequences, each of length N_s, as

$$a_p(k) = a(k + (p - 1)\,N_s);\ k = 0,\ \ldots,\ N_s - 1,\ p = 1,\ \ldots,\ N_t.$$

Each of the N_s sequences $\{a_1(k),\ a_2(k)\ldots,\ a_{Nt}(k)\}$, $k = 0, \ldots, N_s - 1$, is mapped to a matrix $\boldsymbol{\Psi}_k$ of size N_t x N (N is the number of time burst defined in STC) by using the orthogonal space- time block coding scheme given in [Tarokh ,1999]

$$\{a_1(k),\ a_2(k)\ldots,\ a_{Nt}(k)\} \Rightarrow\ \boldsymbol{\Psi}_k,\quad k = 0,\ \ldots,\ N_s - 1. \quad (3.1)$$

For instance, **if we apply the Alamouti code for a** system with two transmit antennas, $\mathbf{\Psi}_k$ is obtained as

$$\mathbf{\Psi}_k = \begin{pmatrix} a_1(k) & -a^*_2(k) \\ a_2(k) & -a^*_1(k) \end{pmatrix} \quad (3.2)$$

Then we take the inverse discrete Fourier transform (DFT) of $\{\mathbf{\Psi}_0, \Psi_{Ns}^{-1}\}$ as

Where $j = \sqrt{-1}$, to form the transmitted signals represented in a matrix form as

$$\boldsymbol{S}m = \frac{1}{\sqrt{N_s}} \sum_{k=0}^{N_s-1} \mathbf{\Psi}k \cdot e^{j\frac{2\pi}{N_s}mk}, \; \mathbf{m} = 0 \ldots N_s - 1 \quad (3.3)$$

$$\boldsymbol{S} = [\boldsymbol{S}^T_0, \; \boldsymbol{S}^T_1, \; \ldots, \boldsymbol{S}^T_{Ns-1}]^T \quad (3.4)$$

Where Sm is given by

$$\boldsymbol{S}m = \begin{pmatrix} s_{1,0}(m) & \cdots & s_{1,(N-1)}(m) \\ \vdots & \ddots & \vdots \\ s_{Nt,0}(m) & \cdots & s_{Nt,(N-1)}(m) \end{pmatrix} \quad (3.5)$$

It is easy to recognize that $\{s_{p,n}(m)\}$, $p = 1, \ldots, N_t$, $n = 0, \ldots, N-1$, $m = 0, \ldots, N_s - 1$, are transmitted in parallel using the N_s subcarriers and N_t antennas over N time intervals. Thus, each transmitted symbol is coded onto the space, time, and frequency dimensions through the ST-OFDM process.

C. Multi-Input Multi-Output Wireless Channel

In a frequency-selective fading channel with L resolvable paths, there exists mutual interference between adjacent Orthogonal frequency division multiplexing (OFDM) blocks. This interblock interference (IBI) could be cancelled by adding a cyclic prefix (CP) of length $_{Cp}$ ($_{Cp} \geq L$) to each trans-mitted block. At the receiver, the Cyclic Prefix is discarded, leaving InterBlock Interference-free information-bearing signals. The channel matrix H is block-circulant with N_s x N_s blocks expressed as

$$H = \begin{pmatrix} H(0) & \cdots & 0 & \mathrm{H}(L-1) & \cdots & \mathrm{H}(1) \\ \vdots & H(L-1) & \cdots & H(1) & H(0) & 0 \\ \vdots & & & \vdots & & \vdots \\ 0 & \cdots & H(L-1) & \cdots & H(1) & H(0) \end{pmatrix} \ldots (4.1)$$

where $\mathbf{0}$ is a zero matrix of size N_r x N_t. Each nonzero block of $\boldsymbol{H}$ represents the MIMO spatial channel matrix of size for a particular path l and is expressed as

$$\boldsymbol{H}(l) = \begin{pmatrix} h_{1,1}(l) & \cdots & h_{1,Nt}(l) \\ h_{Nr,1}(l) & \cdots & h_{Nr,Nt}(l) \end{pmatrix}, \quad l = 0, \; \ldots, \; L-1 \quad (4.2)$$

where $h_{i,j}(l)$, $1 \leq i \leq N_r$, $1 \leq j \leq N_t$, is zero-mean complex Guassian with unit variance. In a practical scenario, insufficient spacing among antennas will cause spatial correlation

a) The Impact of Intercarrier Interference ICI Caused by Phase Noise and Time-Varying Fading

In the presence of phase noise and time-selective fading, the effective $N_sN_r \times N_sN_t$, spatiotemporal channel matrix H_t during the tth (OFDM) Orthogonal Freguency Division Multiplexing symbol period with the *effects* of phase noise taken into consideration is expressed in APPENDIX C

b) Phase Noise Suppression and Data Detection

As mentioned in Section 4.5, do not hold for MIMO-OFDM systems in the presence of phase noise and time-selective fading. The term $P_{kk}x_k$ carries data symbols, but the distortion P_{kk} is a function of the phase noise process, which is costly to estimate. Additionally, when N_s is large, this term is very small due to the scaling factor $1/N_s$. Therefore, the term $P_{kk}x_k$ will be treated as noise for the derivation of minimum mean-squared error, MMSE-based phase noise mitigation and the third term on the right-hand side in the APPENDIX B, the intercarrier interference ICI term caused by both phase noise and time-selective fading.

For Orthogonal frequency division multiplexing (OFDM) systems over fast fading channels, channel estimates are generally obtained by transmitting pilot symbols as certain positions of the frequency-time grid [Simeone, 2004]. When significant phase noise is also present, a joint scheme to simultaneously estimate Common phase error CPE and channel state information CSI is needed. Such a joint estimation appears to be very challenging because of the mutual coupling effects of phase noise and channel fading processes is out of the scope of this chapter. The author thus assumes perfect channel state information (CSI) at the receiver.

III. RESULTS AND DISCUSSION

Numerical Results

Simulations are carried out based on the “SUI-5” channel model [Falconer, 2002] , which is one of six channel models adopted by IEEE 802.16a for evaluating broadband wireless systems in the 2-11GHz band. The author considered a system with two transmits antennas and three receive antennas which employs QPSK modulation.

The doubly-selective Rayleigh fading channel is assumed to have three resolvable multipath components. These paths are modelled as independent complex Gaussian random variables and have relative delays of 0μs, 5μs, and 10μs. The rms delay spread of the channel is 3.05μs and the maximum Doppler shift of the channel is calculated based on a carrier frequency of f_c = 2GHz.

Table 1 PARAMETER VALUES USED IN THE SYSTEM SIMULATIONS

Cell Geometry	Hexagonal Array with side R = 1000 m
Carrier Frequency	fc = 2 GHz
System Bandwidth	W = 5 MHz
Path Loss Exponent	r = 3.7
Shadow Fading	Lognormal, with Standard Deviation _ = 8 dB
Multipath Fading	Rayleigh (K-factor = 0)
Antenna Pattern	Omnidirectional or Uniform over 120
Thermal Noise Density	N0 = −174 dBm/Hz
Mobile Station's Noise Figure	NF = 8 dB
Transmit Power	PT = 5 W for fc = 2 GHz PT = 31.25 W for fc = 5 GHz
Median Cell-Boundary SNR	p = 20 dB
Transmit Antenna Array (BS) Length	BS = 3 m
Receive Antenna Array (MS) Length	IMS = 0.1 m
AoD Statistics (at the Base Station)	Laplacian Power Angular Spectrum with Angular Spread _BS = 15_ = II/12
AoA Statistics (at the Mobile Station)	Laplacian Power Angular Spectrum with Angular Spread _MS = 45_ = 11/4

TABLE .2: PARAMETER VALUES USED IN THE SYSTEM SIMULATIONS

VEHICLE SPEED VS	30km, 60km, 100km, 200km, 400km
NUMBER OF SUBCARRIERS. Ns	12, 64, 128, 256, 512
DATA SYMBOLE Ts	Ts = 5x10-7, 10-6, 10-5, 10-4, 10-3
PHASE NOISE (dB)	βT_{s} = 10-7, βT_{s} =10-6,, βT_{s} = 3x 10-6, βT_{s} =10-5, βT_{s} =10-4, βT_{s} = 10-3
FREQUENCY	fc = 2.GHZ

TABLE 3, showing the improvement in terms of dB, by using the proposed STBC code structure for different Modulations and for different Channels

DIFFERENT MODULATIONS	FOR AWGN CHANNEL	FOR RAYLEIGH CHANNEL
64-QAM	2.7 dB	5 dB
256-QAM	3.5 dB	3.8 dB
1024-QAM	3.25dB	2.5 dB

Table 4 THROUGHPUTS IN BPS/HZ ACHIEVED (WITH ONLY SIGNALING INEFFICIENCY PRESENT) FOR VARIOUS VALUES OF CHANNEL RMS DELAY SPREAD (NANOSEC) AND NUMBER OF CARRIERS/TONES (N).

	N = 64 N	N = 128	N = 256	N = 512
0 ns	3.84	3.84	3.84	3.84
50 ns	3.81	3.82	3.83	3.84
100 ns	3.75	3.80	3.81	3.83
250 ns	3.63	3.73	3.77	3.81
500 ns	3.44	3.62	3.73	3.78
1000 ns	3.14	3.44	3.63	3.73
2500 ns	2.70	3.04	3.36	3.58
5000 ns	X	2.70	3.03	3.36
10000 ns	X	X	2.70	3.04

Table 5 PERCENTAGE LOSS RELATIVE TO THE FLAT-CHANNEL (WITH ONLY SIGNALING INEFFICIENCY PRESENT), FOR VARIOUS VALUES OF CHANNEL RMS DELAY SPREAD (NANOSEC), AND NUMBER OF CARRIERS/TONES (N).

	N = 64	N = 128	N = 256	N = 512
0 ns	0.00%	0.00%	0.00%	0.00%
50 ns	0.78%	0.52%	0.26%	0.00%
100 ns	2.34%	1.04%	0.78%	0.26%
250 ns	5.47%	2.86%	1.56%	0.78%
500 ns	10.42%	5.73%	2.86%	1.30%
1000 ns	18.23%	10.42%	5.47%	2.86%
2500 ns	**29.69%**	**20.83%**	**12.50%**	**6.77%**
5000 ns	**X**	**29.69%**	**21.09%**	**12.50%**
10000 ns	**X**	**X**	**29.69%**	**20.83%**

MIMO-OFDM system more vulnerable to phase noise or time variations of the channel coefficients. The author has assumed perfect channel state information CSI for all numerical results so far. In practical systems, however, there exist channel estimation errors. It is beyond the scope of this chapter to discuss channel estimation schemes for time-selective fading channels. To access its impact, channel estimation error is emulated by introducing an error with a normalized average MSE defined as $MSE = E\,[\|\,\boldsymbol{H} - \boldsymbol{H}\,\|^2_F]$ /E $[\|\boldsymbol{H}\|^2_F]$, where $\boldsymbol{H}$ has the same form, except that phase noise terms and OFDM symbol index are neglected. The performance results of MIMO-OFDM systems with various MSE values are shown in Fig. 1, where all parameters, except β = 10Hz, are the same as those applied in Fig. 2. The proposed MMSE-based phase noise suppression scheme and the MMSE detection scheme are employed in this simulation. It is observed that the Performance degradation is negligible only when the MSE value of channel estimation errors is small (e.g., 10^{-3}).

The strongest signal refers to the signal with the highest signal-to-noise ration (SNR), and the weakest signal refers to the signal with the lowest SNR. As the number of subcarriers increases, however, system performance deteriorates rapidly.

Discussion

Fig. 2 shows the CIR values as a function of data symbol period T_s, the 3-dB phase noise linewidth β, and the number of subcarriers N_s within one OFDM symbol. These curves are obtained by using the analytical expression given in Eq. (4.1) and simulations based on the maximum Doppler shift under a vehicle speed of v_s = 100Km/h. Simulation results match well with the theoretical results. CIR is found to be inversely proportional to T_s, N_s, and β; thus, increasing β or *Ts* makes the MIMO-OFDM system more vulnerable to phase noise or time variations of the channel coefficients.

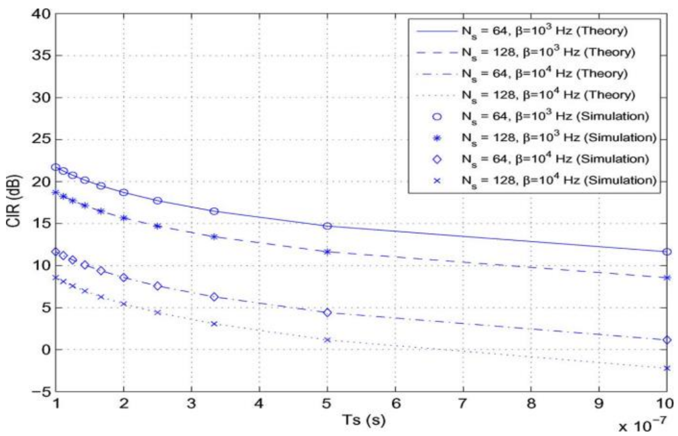

Figure 2: CIR comparisons with different number of subcarriers and phase noise linewidth (vs = 100Km/h).

In Fig. 3, SINR versus E_s/N_0 curves under different values of βT_s and v_s are obtained by using computer simulations. The OFDM symbol is assumed to have N_s = 256 subcarriers, and data symbol period is T_s = 10^{-6} seconds. It is observed that SINR is inversely proportional to βT_s. With a fixed but large value of βT_s (e.g., 10^{-3}), however, the difference between SINR curves corresponding to different vehicle speeds diminishes. This is because when βT_s is large, ICI is dominated by phase noise. On the other hand, with a smaller βT_s value such as βT_s = 10^{-4}, increasing the Doppler shift (or vehicle speed) clearly lowers the SINR value.

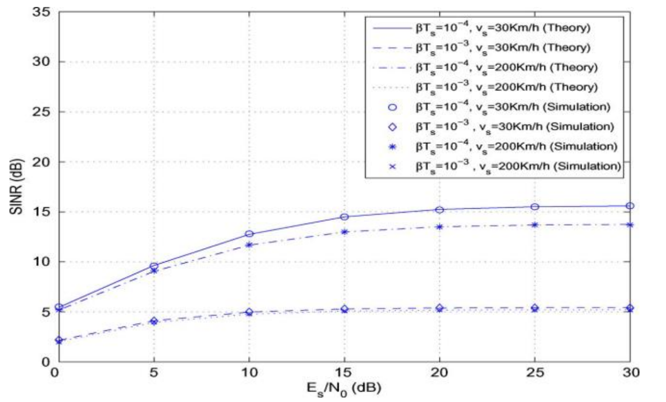

Figure 3. SINR versus Es/N0 for MIMO-OFDM with different vehicle speed and phase noise variance (Ns = 256, Ts = 10-6s).

Fig. 4. shows the SER performance of the proposed MMSE-based phase noise suppression scheme together with those of a phase-noise-free system and a system without phase noise correction when the MMSE detection scheme is considered. System parameters chosen are: N_s = *128*, T_s = 10^{-7}*s*, β =*10Hz*, and v_s = 30Km/h. It is observed that without phase noise correction, even a very mild amount of phase noise (βT_s = 10^{-6}) causes a high error floor. On the other hand, the proposed scheme significantly reduces the effect of phase noise. Note that performance of the proposed scheme does not approach that of the phase-noise-free system because this scheme mitigates only CPE, and it does not eliminate ICI, which is caused by both phase noise and time-selective fading.

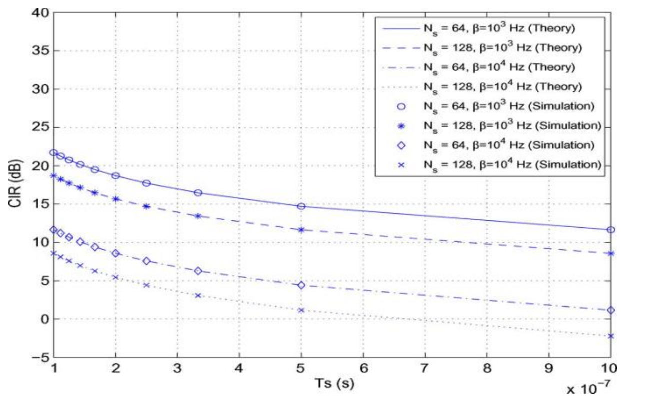

Figure 4 CIR comparisons with different number of subcarriers and phase noise linewidth (vs = 100Km/h).

Shown in Fig. 4 are the simulated SER performances of the system when the proposed MMSE-based phase noise suppression scheme and the MMSE detection scheme described are employed. Other parameters chosen are: N_s = *64*, T_s = 10^{-7}*s*, and v_s = 100Km/h. Performances with different values of the 3-dB phase noise variance (βT_s = 10^{-7}, 10^{-6}, 3×10^{-6}, and 10^{-5}) are compared. The performance curve of a phase-noise-free MIMO-OFDM system is used as the baseline performance. It appears that the scheme works effectively only when βT_s is small.

In Fig. 6.., we compare the performances of four different detection methods: the ZF, MMSE, decorrelating DF, and MMSE-DF schemes when the MMSE – based phase noise suppression scheme applied. Other than that β = 30Hz, all other parameters are the same as those applied for Fig. 4.. Performance of the ML scheme is used as the benchmark for other detection schemes. Since these schemes are not specifically optimized for MIMO-OFDM systems with phase noise over fast time-varying fading channels for which ICI should be dealt with, error floors are observed for all cases. Note that from Eqs. (4.1) and (4.2), the energy of ICI due to the phase noise and time-selective fading is found to spread over all subcarriers, which is different from the assumption that most of ICI on each subcarrier comes from several neighbouring subcarriers. Consequently, ICI suppression for the scenario studied in this chapter becomes more challenging than the case dealt with earlier.

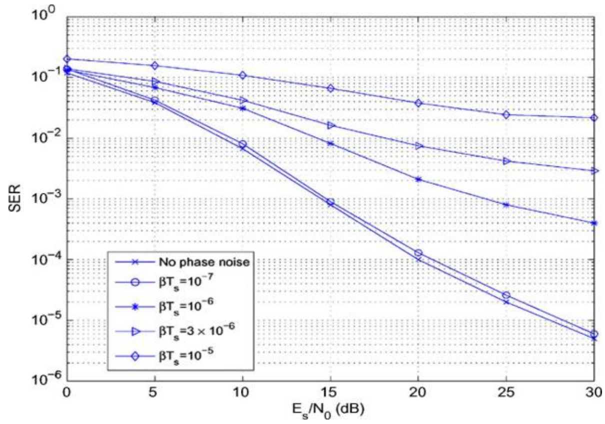

Figure 5. SER versus Es/N0 for MIMO-OFDM with different phase noise variance (Ns = 64, Ts = 10-7s, vs = 100Km/h).

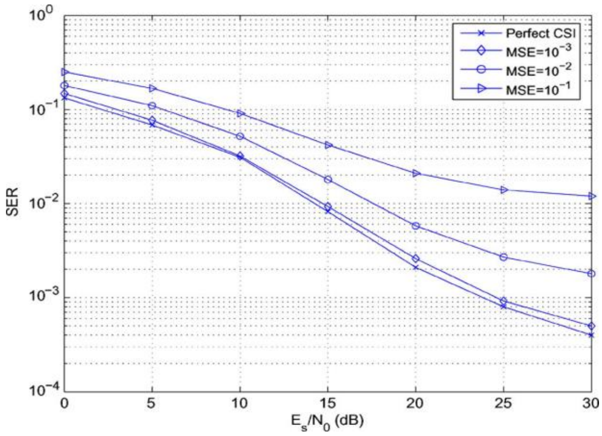

Figure 6 : SER versus E_s/N_0 for MIMO-OFDM with different detection schemes (βT_s = 3 x 10^{-6}, N_s = 64, v_s = 100Km/h).

We have assumed perfect CSI for all numerical results so far. In practical sys-tems, however, there exist channel estimation errors. It is beyond the scope of this chapter to discuss channel estimation schemes for time-selective fading channels. To access its impact, channel estimation error is emulated by introducing an error with a normalized average MSE defined as $MSE = E[\| \boldsymbol{H} - \boldsymbol{H} \|^2_F] / E[\|\boldsymbol{H}\|^2_F]$, where $\boldsymbol{H}$ has the same form as Eq. (4 1), except that phase noise terms and OFDM symbol index are neglected. The performance results of MIMO-OFDM systems with various MSE values are shown in Fig. 4.2., where all parameters, except β = 10Hz, are the same as those applied in Fig. 8.. The proposed MMSE-based phase noise suppression scheme and the MMSE detection scheme are employed in this simulation. It is observed that the performance degradation is negligible only when the MSE value of channel estimation errors is small (e.g., 10^{-3}).

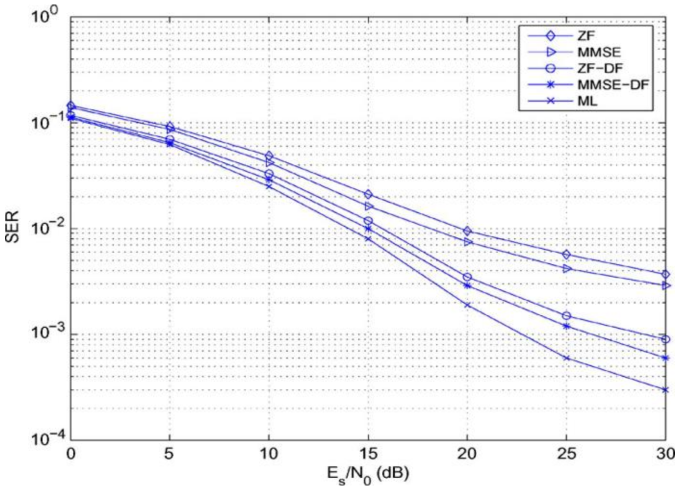

Figure 7. SER versus Es/N0 for MIMO-OFDM with different MSE (βTs = 10-6, Ns = 64, vs = 100Km/h).

IV. CONCLUSION

We have analyzed the impact of phase noise and channel time selectivity on the performance of MIMO-OFDM systems. Specifically, we have quantified ICI caused by phase noise and channel time variations. A phase noise suppression scheme based on the MMSE criterion is proposed, which is shown to effectively reduce the effect of phase noise. Performances of five detection schemes are compared, and it

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