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A Wireless Multiuser System Using Diversity

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Abstract – This paper quantifies how diversity and the number of users affect the performance of a wireless multiuser system. The system accommodates N users and one central base station. Only the reverse link is considered. The base station receives the signal at A different antennas. In addition to antenna diversity, bandwidth diversity is used by transmitting signals with larger than Nyquist bandwidth. The receiver consists of an optimum linear MMSE equalizer/combiner. The stationary, frequency selective radio channels between all users and the base station are assumed to be known at the receiver. MMSE expressions are given and numerically evaluated. A tight lower bound is derived for the case when the number of users is larger than the total degree of diversity, referred to as *overpopulated*. Otherwise, the system is *well populated*. It is shown that overpopulated systems are interference limited while well populated systems are noise limited.

I. INTRODUCTION

THE connection of several individual stations to a central unit is a characteristic of many modern communication systems. The rise of multimedia applications requires in addition high data rates, which causes the radio channel to behave frequency selectively. While the latter introduces intersymbol interference (ISI) in the received signals, multiple, simultaneously transmitting stations are the cause of cochannel interference (CCI). It is known that this combined interference is the major limiting factor of both system performance and capacity. Several access schemes – among them TDMA, FDMA and CDMA – can be employed to avoid or mitigate CCI. These schemes are based on bandwidth expansion. More recently, it has been shown that multiple receiver inputs (antenna/spatial diversity, SDMA) have a similar ability to suppress CCI [1]. ISI is very effectively mitigated by equalizers at the receiver.

This paper combines all three methods – a linear MMSE equalizer/combiner, bandwidth and antenna diversity – in an effort to increase the performance and capacity of a bandwidth efficient wireless multiuser system. All transmitted signals are allowed to overlap spectrally and temporally, causing severe interference at the receiver. CDMA, for example, is a special case of this method. Qualitative results, [2], [3], indicate that the combination of spatial and temporal diversity improves the system performance and capacity. For a zero-forcing equalizer/combiner, it was shown that the number of suppressible interferers increases linearly by the product of the number of receive antennas and the system bandwidth relative to the symbol rate [4], [5], [6]. The objective of this paper is to analyze the relationship among diversity, number of system users and performance when a linear discrete-time MMSE equalizer/combiner (E/C) is used.

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II. SYSTEM MODEL

Consider the reverse link of the multiuser system shown in Figure 1. The complex baseband notation is used to describe the system. All signals and impulse responses are in general complex functions. The system consists of N users transmitting data sequences a_i $(i = 1, \ldots, N)$. The data sequences consist of symbols drawn from a finite alphabet of complex numbers. After K times upsampling, the individual sequences are filtered by discrete-time filters q_i. Their outputs are fed into linear impulse generators which modulate the discrete-time sequences with the waveform $p_C(t)$ to produce the transmit signals. Let the clock rate of the impulse generators be $1/T_s$. The symbol period is then given by $T = KT_s$.

The signal of user i travels through the radio channel with impulse response $h_{Cil}(t)$ and is received at antenna l of the base station. Mutually independent, complex AWGN signals ν_{CGl} $(l = 1, \ldots, A)$ with two-

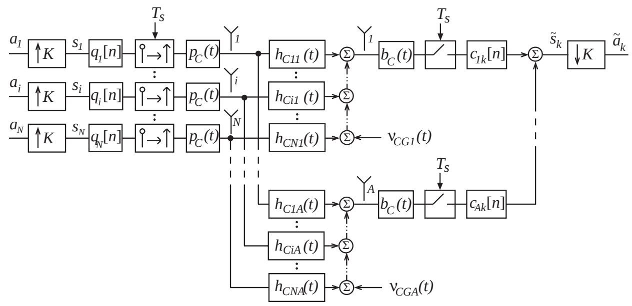

Fig. 1. Block diagram of the multiuser system

sided power spectral density N_0 are added at each of the A base station antennas. The received signals are lowpass filtered ($b_C(t)$), sampled at a rate $1/T_s$, equalized, combined and finally K times downsampled to produce the linear estimate $\tilde{a}_k$ of user k's input sequence a_k. Consider the *net channel* between user i and the l-th receive antenna which is given by the convolution of $p_C(t)$, $h_{Cil}(t)$ and $b_C(t)$:

$$\psi_{Cil}(t) = \iint p_C(u) h_{Cil}(v-u) b_C(t-v)\, du\, dv. \quad (1)$$

For generality, limits are from minus to plus infinity. After sampling, the noise signal components are colored: $\nu_l[n] = \int b_C(\tau)\nu_{CGl}(nT_s - \tau)\, d\tau$. The continuous-time net channel is embedded into the overall discrete-time system. It can be described by equivalent discrete-time impulse responses $\psi_{il}[n] = \psi_{Cil}(nT_s)$. Let us define the *combined channel*

$$x_{il}[n] = \sum_{v=-\infty}^{\infty} q_i[v]\psi_{il}[n-v]. \quad (2)$$

The analysis is done using the D-transform which is defined by $\boldsymbol{u}(D) = \sum_{n=-\infty}^{\infty} \boldsymbol{u}[n]D^n$, where $\boldsymbol{u}$ may be an arbitrary dimensional row vector $\boldsymbol{u} = [u_1, u_2, \dots]$. Let us define the truncated sequence

$$\boldsymbol{u}_M[n] = \begin{cases} \boldsymbol{u}[n] & \text{, for } |n| \le M \\ 0 & \text{, for } |n| > M. \end{cases} \quad (3)$$

Let $\boldsymbol{v}$ be another row vector whose truncated sequence is defined according to Equation (3). The cross-power spectrum $\boldsymbol{S}_{uv}(D)$ of $\boldsymbol{u}[n]$ and $\boldsymbol{v}[n]$ is then equal to

$$E_M[\boldsymbol{u}^H(D^{-*})\boldsymbol{v}(D)] \stackrel{\text{def}}{=} \lim_{M\to\infty} \frac{E[\boldsymbol{u}_M^H(D^{-*})\boldsymbol{v}_M(D)]}{2M+1} \quad (4)$$

where 'E' is the expectation operator, the superscripts 'H', '$*$', '-1' denote the conjugate transpose, complex conjugate and inverse, respectively. The superscript '$-*$' shall be interpreted in the sense $D^{-*} = (D^{-1})^*$.

According to Figure 1 the signals of the system can be expressed as

$$s_i(D) = a_i(D^K) \quad (5)$$

$$y_l(D) = \sum_{i=1}^{N} s_i(D)x_{il}(D) + \nu_l(D) \quad (6)$$

$$\tilde{s}_k(D) = \sum_{l=1}^{A} y_l(D)c_{lk}(D) \quad (7)$$

$$\tilde{a}_k(D) = \frac{1}{K}\sum_{m=0}^{K-1} \tilde{s}_k(D^{\frac{1}{K}} w_K^m) \quad (8)$$

where $w_K = e^{-j2\pi/K}$. Let us define the following row vectors and matrices:

$$\boldsymbol{a}(D) = [a_1(D), a_2(D), \dots, a_N(D)] \quad (9)$$
$$\tilde{\boldsymbol{a}}(D) = [\tilde{a}_1(D), \tilde{a}_2(D), \dots, \tilde{a}_N(D)] \quad (10)$$
$$\boldsymbol{y}(D) = [y_1(D), y_2(D), \dots, y_A(D)] \quad (11)$$
$$\boldsymbol{\nu}(D) = [\nu_1(D), \nu_2(D), \dots, \nu_A(D)] \quad (12)$$
$$\boldsymbol{y}_t(D) = [\boldsymbol{y}(\gamma_0), \boldsymbol{y}(\gamma_1), \dots, \boldsymbol{y}(\gamma_{K-1})] \quad (13)$$
$$\boldsymbol{\nu}_t(D) = [\boldsymbol{\nu}(\gamma_0), \boldsymbol{\nu}(\gamma_1), \dots, \boldsymbol{\nu}(\gamma_{K-1})] \quad (14)$$
$$\boldsymbol{X}(D) = [x_{il}(D)] \quad (15)$$
$$\boldsymbol{C}(D) = [c_{lk}(D)] \quad (16)$$
$$\boldsymbol{X}_t(D) = [\boldsymbol{X}(\gamma_0), \boldsymbol{X}(\gamma_1), \dots, \boldsymbol{X}(\gamma_{K-1})] \quad (17)$$
$$\boldsymbol{C}_t(D) = [\boldsymbol{C}^H(\gamma_0), \boldsymbol{C}^H(\gamma_1), \dots, \boldsymbol{C}^H(\gamma_{K-1})]^H \quad (18)$$

where $\gamma_m = D^{\frac{1}{K}} w_K^m$; $i, k = 1, \dots, N$; $l = 1, \dots, A$ and $[x_{il}]$ defines per convention a matrix whose (i,l)-th component is x_{il}. Using the above definitions, Equations (5) to (8) can be expressed in vector form as

$$\boldsymbol{y}_t(D) = \boldsymbol{a}(D)\boldsymbol{X}_t(D) + \boldsymbol{\nu}_t(D) \quad (19)$$

$$\tilde{\boldsymbol{a}}(D) = \frac{1}{K}\boldsymbol{y}_t(D)\boldsymbol{C}_t(D) \quad (20)$$

where we used the fact that $w_K^{Km} = 1$ if m is an integer. Note that Equations (19) and (20) describe the multiuser system completely. Furthermore we notice that the channel matrix $\boldsymbol{X}_t$ consists of N rows (one for each user) with AK components each. The linear equalizer/combiner (E/C) $\boldsymbol{C}_t$ has AK degrees of freedom (or diversity) to estimate each of the transmitter sequences a_i. Thus, the degree of diversity is given by the product of the oversampling factor K and the number of receive antennas A.

Spectral Correlation of the Noise

Let $\boldsymbol{S}_n$ denote the spectrum of the colored Gaussian noise signal $\boldsymbol{\nu}$: $\boldsymbol{S}_n(D) = E_M[\boldsymbol{\nu}^H(D^{-*})\boldsymbol{\nu}(D)]$. Accordingly, the spectrum of the extended noise signal $\boldsymbol{\nu}_t$ is defined as $\boldsymbol{S}_\nu(D) = E_M[\boldsymbol{\nu}_t^H(D^{-*})\boldsymbol{\nu}_t(D)]$. If spectral components of the mutually independent signals ν_{CGl} separated by at least $\Delta f = 1/T_s$ are uncorrelated, it can be shown that

$$\boldsymbol{S}_n(D) = \frac{N_0}{T_s}\left(\sum_{v=-\infty}^{\infty} |B(f - v/T_s)|^2\right)\boldsymbol{I}_A \quad (21)$$

$$\boldsymbol{S}_\nu(D) = K\,\mathbf{D}\mathrm{iag}\langle \boldsymbol{S}_n(D^{\frac{1}{K}} w_K^m)\rangle,\ m=0,..,K{-}1 \quad (22)$$

with $B(f) = \int_{-\infty}^{\infty} b_C(t) e^{-j2\pi f t}\,dt,\ D = e^{-j2\pi \check{f}}$ and $\check{f} = fT_s$. $\boldsymbol{I}_A$ is the $A\times A$ identity matrix and $\mathbf{D}\mathrm{iag}\langle \boldsymbol{G}_i\rangle$ is a diagonal hypermatrix with diagonal elements $\boldsymbol{G}_i$ $(i = 1, 2, \dots)$, which may be matrices of arbitrary size. Since $\boldsymbol{S}_n(D)$ is a $A\times A$ diagonal matrix with nonnegative elements, (21), $\boldsymbol{S}_\nu(D)$ is a $AK\times AK$ diagonal matrix with nonnegative elements. It is reasonable to assume that all diagonal elements are nonzero. In this case, $\boldsymbol{S}_\nu(D)$ is positive definite.

III. THE OPTIMUM MMSE EQUALIZER/COMBINER

The optimum continuous-time E/C for a multiple input multiuser receiver is derived in [7]. It is shown for colored noise signals that the optimum linear MMSE structure can be achieved with four components. The first is a matrix whitening filter for the noise signals. Second is a matrix filter matched to the transfer functions of the combined channels. This is followed by symbol rate (1/T) samplers at all outputs of the matrix matched filter. The last component is a $N\times N$ discrete-time matrix filter. It is possible to extend the results in [7] to the system described in Section II. The first part of our discrete-time MMSE E/C is the matrix whitening filter $\boldsymbol{S}_\nu^{-1}(D)$ which is the inverse of the noise spectrum matrix in Eqn. (22). The second is the matched filter matrix $\boldsymbol{X}_t^H(D^{-*})$ which is matched to the combined discrete-time channels (2). Following are compressors which downsample all N signals by a factor of K. This is equivalent to the symbol-rate sampling step performed for the system in [7]. The next component is an $N\times N$ discrete-time matrix filter $\boldsymbol{L}(D)$. Finally, each of the N symbol-rate sequences are amplified K times to compensate for the factor $1/K$ in Equation (20). This can be done by an amplifier with transfer matrix $K\boldsymbol{I}_N$. Up to this point the receiver $\boldsymbol{C}_t(D)$ can be expressed in the form

$$\boldsymbol{C}_t(D) = K\,\boldsymbol{S}_\nu^{-1}(D)\boldsymbol{X}_t^H(D^{-*})\boldsymbol{L}(D) \quad (23)$$

where $\boldsymbol{L}(D)$ is a symbol spaced $N\times N$ matrix filter. Let us define the *effective channel* as

$$\boldsymbol{S}_x(D) = \boldsymbol{X}_t(D)\boldsymbol{S}_\nu^{-1}(D)\boldsymbol{X}_t^H(D^{-*}). \quad (24)$$

The system Equations (19) and (20) can then equivalently be written in the form

$$\tilde{\boldsymbol{a}}(D) = \boldsymbol{a}(D)\boldsymbol{S}_x(D)\boldsymbol{L}(D) + \boldsymbol{z}(D)\boldsymbol{L}(D) \quad (25)$$

where $\boldsymbol{z}(D) = \boldsymbol{\nu}_t(D)\boldsymbol{S}_\nu^{-1}(D)\boldsymbol{X}_t^H(D^{-*})$ is the symbol rate discrete-time noise sequence at the input of the E/C $\boldsymbol{L}(D)$. It is easy to show that the spectrum of this noise signal is $\boldsymbol{S}_z(D) = E_M[\boldsymbol{z}^H(D^{-*})\boldsymbol{z}(D)] = \boldsymbol{S}_x(D)$. Interestingly, the spectrum of the noise signal is equal to the transfer matrix of the effective channel. Let $\boldsymbol{S}_a(D) = E_M[\boldsymbol{a}^H(D^{-*})\boldsymbol{a}(D)]$ be the spectrum of the input signal. We can now apply the results in [7] and [8] to find an expression for $\boldsymbol{L}(D)$:

$$\boldsymbol{L}(D) = [\boldsymbol{S}_a(D)\boldsymbol{S}_x(D) + \boldsymbol{I}_N]^{-1}\boldsymbol{S}_a(D). \quad (26)$$

MMSE Expression and Bound

Let us from now on assume that the input signals a_i of different users are mutually independent. Additionally, samples of the same sequence are assumed to be uncorrelated with zero mean and variance $\mathcal{E}_a$:

$$E\left[a_i[n]a_k[m]\right] = \mathcal{E}_a\delta[i-k]\delta[n-m] \quad (27)$$

where $\delta[k]$ is the Kronecker delta sequence. In this case, the spectrum of the input signal reduces to $\boldsymbol{S}_a(D) = \mathcal{E}_a\boldsymbol{I}_N$. An expression for the normalized minimum mean-squared error (NMMSE) of user k can be found in [8]:

$$\sigma_k = \frac{1}{\mathcal{E}_a} E[|\tilde{a}_k[n] - a_k[n]|^2] = \int_0^1 U_k(e^{-j2\pi\check{f}})\,d\check{f} \quad (28)$$

where $U_k(D)$ is the k-th diagonal element of the matrix $\boldsymbol{U}(D)$ which is given by

$$\boldsymbol{U}(D) = [\mathcal{E}_a\boldsymbol{S}_x(D) + \boldsymbol{I}_N]^{-1}. \quad (29)$$

The average MMSE can then be determined to be

$$\sigma = \frac{1}{N}\sum_{k=1}^{N}\sigma_k = \frac{1}{N}\,\mathrm{tr}\left\{\int_0^1 \boldsymbol{U}(e^{-j2\pi\check{f}})\,d\check{f}\right\} \quad (30)$$

where $\mathrm{tr}\{\boldsymbol{A}\}$ is the trace of the matrix $\boldsymbol{A}$.

We consider now explicitly the case $D = D^{-*} = e^{-j2\pi\check{f}}$. For the sake of brevity, we drop the argument in the following but keep in mind that the functions and matrices depend on D. It is assumed that the noise spectrum $\boldsymbol{S}_\nu$ is a positive definite diagonal matrix. Hence, the matrix $\boldsymbol{S}_x$, which is defined in Eqn (24), is Hermitian. Therefore, we can decompose $\boldsymbol{S}_x$ as $\boldsymbol{S}_x = \boldsymbol{Q}\boldsymbol{\Lambda}\boldsymbol{Q}^H$, where $\boldsymbol{Q}$ is a unitary matrix and $\boldsymbol{\Lambda} = \mathbf{D}\mathrm{iag}\langle\lambda_i\rangle$ is a $N\times N$ diagonal matrix whose diagonal elements λ_i $(i = 1, 2, \dots, N)$ are the eigenvalues of $\boldsymbol{S}_x$ [9, p.165]. The matrix $\boldsymbol{U}$ in (29) can then be written as

$$\boldsymbol{U} = \boldsymbol{Q}[\mathcal{E}_a\boldsymbol{\Lambda} + \boldsymbol{I}_N]^{-1}\boldsymbol{Q}^H. \quad (31)$$

Taking the trace of $\boldsymbol{U}$ [9, p.167], we obtain the average MMSE in terms of the eigenvalues of $\boldsymbol{S}_x$:

$$\sigma = \frac{1}{N}\int_0^1 \sum_{i=1}^{N}[\mathcal{E}_a\lambda_i(\check{f}) + 1]^{-1}d\check{f}. \quad (32)$$

The matrix $\boldsymbol{S}_x$ defined in (24) can only be regular if $\boldsymbol{X}_t$ is a full row rank matrix [10, p.164–67]. This is possible only if the number of system users N is smaller or equal to the product of oversampling factor K and the number of receive antennas A. On the other hand, if $N > AK$, the rank of $\boldsymbol{X}_t$ is at most AK. In this case, the $N \times N$ matrix $\boldsymbol{S}_x$ is singular. Its rank is smaller than or equal to AK [10] and its rank deficiency is greater than or equal to the *overpopulation number* $\xi = N - AK$. This means that $\boldsymbol{S}_x$ has at least ξ eigenvalues equal to zero.

A zero-forcing (ZF) E/C will not exist if the system is overpopulated, i.e. $N > AK$ [4]. However, a unique MMSE E/C filter $\boldsymbol{L}$ (26) always exists. This can easily be verified by recognizing that $\boldsymbol{S}_x$ (24) is positive semidefinite and hence $\boldsymbol{U}^{-1}$ (29) is positive definite and thus regular. It is now easy to find a lower bound for the average MMSE of overpopulated systems. For this we consider only the ξ eigenvalues of $\boldsymbol{S}_x$ which are zero and neglect all other eigenvalues. We obtain then from Equation (32)

$$\sigma > \frac{1}{N}\int_0^1 \left(\sum_{i=1}^{\xi} 1\right) d\check{f} = 1 - \frac{AK}{N}, \quad N > AK. \quad (33)$$

IV. NUMERICAL RESULTS

The results described in this section have been obtained for a system with the following parameters: Symbol period $T = 200$ ns, $A = 4$ receive antennas at the base station, oversampling factor $K = 4$. The number of system users N has been varied between 1 and 30. The filters $p_C(t)$ and $b_C(t)$ are identical fifth-order butterworth lowpass filters with cut-off frequency $f_c = K/(2T)$. The discrete-time transmit filters of all users have been set to $q_i[n] = \delta[n]$.

We have used indoor channel impulse response (CIR) measurements obtained at TR*Labs* [11]. The CIR's were measured in an indoor office environment. The measurement system included four stationary transmit antennas and a mobile with four receive antennas. The distance between two adjacent receive antennas was one wavelength of the carrier frequency $f_{\text{car}} = 1.8$ GHz. The stationary antennas were placed in different corners of the office environment. Different impulse responses were obtained by changing the location of the mobile. Each measurement at a certain mobile location yielded four sets of four CIR's between the adjacent mobile antennas and one of the stationary antennas. The four CIR's belonging to one set had the same large scale propagation characteristics because the distances between a certain stationary antenna and each of the four mobile antennas were practically the same. The measurements resulted in a total of 2044 sets or 8176 CIR's. The bandwidth of the measured CIR's was approximately 120 MHz.

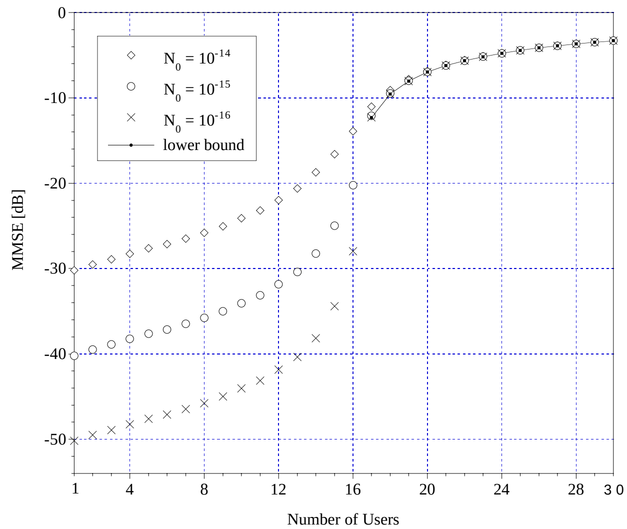

Fig. 2. Average MMSE for identical systems with different received SNR's

The reverse link of the system has been simulated by randomly selecting N of the 2044 CIR sets and calculating the theoretical NMMSE's (28). This procedure has been repeated 100 times for each value of N with different CIR sets. Figure 2 shows the average MMSE, averaged over all system users (Eqn. (30)) and all 100 trials for different noise levels. The three curves have been obtained for identical environments but received SNR's that differ by 10 dB between two adjacent curves. The lower bound for overpopulated systems (Eqn. (33)) is also included. It can be seen that this bound is very tight. The average MMSE ratio in dB for identical systems with different received SNR's is shown in Figure 3. The solid curve corresponds to the MMSE ratio between curves one (diamonds) and three (crosses) of Figure 2. Analogously, the dashed line represents the ratio between curves one and two in Figure 2. The difference in received SNR between the two systems is 20 dB and 10 dB, respectively. It can clearly be seen that there are large MMSE advantages for systems with higher SNR levels if the number of users is smaller or equal to the degree of diversity ($AK = 16$). For overpopulated systems ($N > 16$), larger SNR's have practically no effect on the MMSE performance. This is in agreement with the lower bound for overpopulated systems, Eqn. (33), which depends only on the number of users and degree of diversity but not on the received SNR. Note that the ratio of the average MMSE for $N \leq 14$ is practically identical to the SNR difference between the two systems. The transition of the MMSE ratio

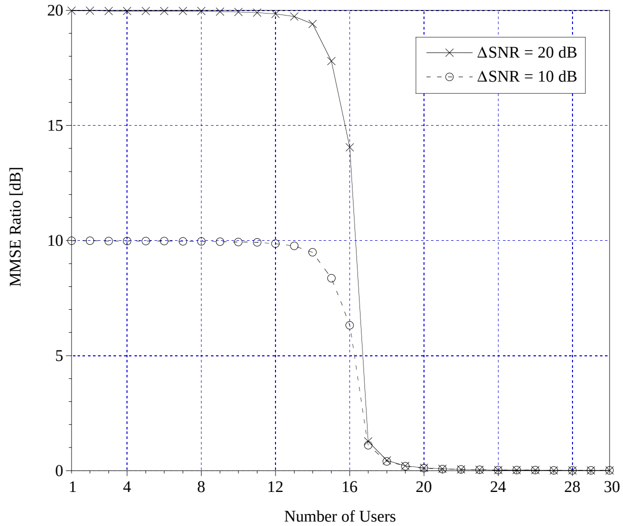

Fig. 3. MMSE ratio between identical systems with different received SNR's

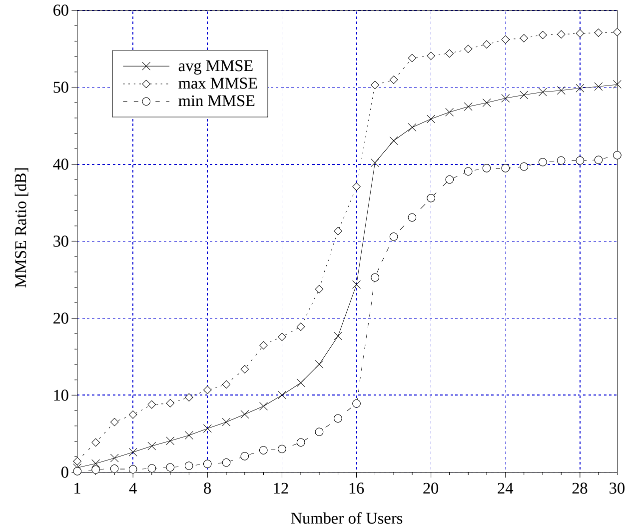

Fig. 4. Maximum, average and minimum of the relative MMSE

from $N \leq 16$ to overpopulated systems is steep but continuous. The transition is influenced by the SNR level. If high SNR systems are compared, the transition will become more abrupt, changing almost step like from the SNR difference ($N \leq AK$) to zero (overpopulated systems). On the other hand, the transition around $N = AK$ becomes the more continuous the lower the SNR of the compared systems is. Figure 4 shows the relative MMSE $\sigma_{k,\mathrm{rel}} = \sigma_k/\sigma_{k,\mathrm{id}}$ which is the ratio between the absolute MMSE (Eqn. (28)) and the matched filter bound [12] $\sigma_{k,\mathrm{id}} = 1/(1+\Phi_k)$, where $\Phi_k = \sum_{l=1}^{A} \Phi_{k,l}$ is the total received SNR from user k and $\Phi_{k,l}$ is the received SNR at base antenna l from user k. The matched filter bound establishes the ultimate performance bound. It can only be reached in optimal systems with neither cochannel nor intersymbol interference. The crosses in Fig. 4 are the average relative MMSE averaged over all system users and trials. The diamond symbols represent the largest individual MMSE $\sigma_{k,\mathrm{rel}}$ that has been found in all trials. Accordingly, the circle symbols mark the smallest relative MMSE that has been determined. There are strong differences between the smallest and largest MMSE. For a moderate number of 4 users, the difference is already 7 dB and increases to 28 dB for $N = 16$. This shows that even if the average MMSE is low, the received signals of certain users may be poor. It can also be seen that the average MMSE is for $N = 1$ user very close to the matched filter bound (horizontal 0 dB line) and increases almost linearly with N for a small number of system users. The increase becomes stronger the more N approaches the system limit of 16 users. A steep MMSE increase appears around $N = 16$. Overpopulated systems show a large difference between the MMSE and the matched filter bound.

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