EUSIPCO 2013 1569743441

Analytical Models for Channel Aging and Synchronization Errors for Base Station Cooperation

Konstantinos Manolakis[1], Christian Oberli[2], Lurys Herrera[2], Volker Jungnickel[3]
[1] Technische Universität Berlin, Department of Telecommunication Systems, Berlin, Germany
[2] Pontificia Universidad Católica de Chile, Department of Electrical Engineering, Santiago, Chile
[3] Fraunhofer Heinrich Hertz Institute, Berlin, Germany
Email: konstantinos.manolakis@tu-berlin.de; obe@ing.puc.cl; volker.jungnickel@hhi.fraunhofer.de

***Abstract*—Base station cooperation is a powerful technique for eliminating inter-cell interference and enhancing spectral efficiency in cellular networks. However, impairment effects due to channel time variance, channel estimation, feedback quantization and imperfect synchronization among base stations are limiting the potential gains. In this paper, we present a comprehensive signal model for multi-user multi-cellular systems with cooperating base stations, which includes those essential impairments. The effect of each impairment is captured by its mean square error (MSE), for which exact analytical expressions and accurate approximations are derived and numerically verified. Those MSE expressions can be used for link-layer abstraction, system-level evaluation and signal to interference ratio (SIR) analysis. We evaluate the spectral efficiency of cooperative networks considering a real-world scenario and find that channel aging has the largest impact on performance degradation.**

I. Introduction

Cellular networks aim at reducing interference by means of multiple-input multiple-output (MIMO) techniques. Unlike traditional single-antenna techniques, they serve multiple data streams in parallel on the same time-frequency resource in order to enhance spectral efficiency. However, multi-cell networks still suffer from interference between adjacent cells limiting the overall performance.

Base station cooperation, also known as joint transmission coordinated multi-point (JT CoMP), is a generalized idea of MIMO where, in the downlink, antennas of multiple distributed base stations are considered as inputs and antennas of multiple terminals in those cells are considered as outputs of a distributed MIMO system. Joint signal processing allows for eliminating inter-cell interference between the cells [1]. In the simplest case of zero-forcing (ZF) precoding, in the downlink, data signals are multiplied with the channel pseudo-inverse [2]. In this way, transmission to different users becomes orthogonal and each user receives only its own signal.

In real-world systems, however, there are impairments causing a mismatch between the precoder and the channel over which the transmission is realized. As shown in Fig. 1, in frequency division duplex (FDD) systems, there is a feedback delay, equal to the time between channel estimation and when this estimate is used for precoding the downlink transmission. As the precoder is computed from quantized estimates of the channel, noise and quantization also contribute to inaccuracy. Finally, base stations are driven by local oscillators with individual synchronization parameters.

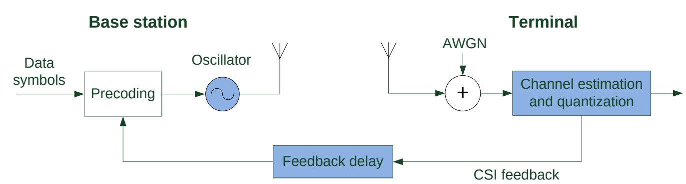

Fig. 1. Sources of imperfect channel knowledge are colored blue: channel estimation and feedback quantization at the terminal, channel aging during the feedback delay and individual synchronization parameters of the base stations.

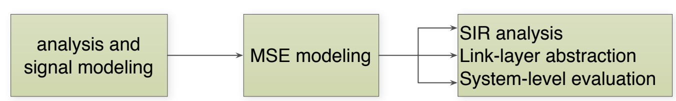

Fig. 2. Starting from exact signal modeling, impairments are captured by their MSE, vital for SIR analysis, link-layer abstraction and performance evaluation.

In [3] and [4], signal models for JT CoMP with carrier and sampling frequency offsets at transmitter and receiver sides are provided. It is found that accurate synchronization among base stations (BSs) is essential. Precoded transmission with imperfect channel knowledge and expressions for spectral efficiency are investigated in [5]. In [6], the importance of transmit-side channel knowledge is addressed and performance degradation due to channel variance is evaluated. In [7], the authors introduce the channel MSE as an interface between exact signal modeling and an evaluation framework, as shown in Fig. 2. The relation between MSE and SIR was not yet clear. Meanwhile, analytical SIR expressions for JT CoMP have been derived in [8]. The SIR is inverse to the MSE and depends on the ratio of the number of BSs and terminals.

Those findings motivated our work on MSE modeling under the perspective of SIR analysis, link-layer abstraction and system level evaluation, see Fig. 2. In this paper, analytical expressions and accurate approximations are derived for the MSE due to each impairment, and are numerically verified. Spectral efficiency is further evaluated for JT CoMP.

The paper is as organized as follows. In Section II, a signal model including impairments is given for JT CoMP. Analytical MSE expressions and the resulting SIR are derived for each impairment. In Section III, the MSE and spectral efficiency are evaluated. Conclusions are drawn in Section IV.

II. Signal Model and MSE Analysis

A multi-cellular network with N_b BSs serving N_u single-antenna terminals with JT CoMP is considered. A backhaul network enables fast data and channel state information (CSI) exchange between the BSs. The narrow-band channel matrix $\mathbf{H}$ has size $N_u \times N_b$. Considering orthogonal frequency division multiplexing (OFDM), $\mathbf{H}$ denotes the frequency response on a single subcarrier. For ZF precoding, the $N_b \times N_u$ precoding matrix $\mathbf{W}$ is calculated by the right-hand Moore-Penrose pseudo-inverse of the channel matrix $\mathbf{H}$, i.e. $\mathbf{W} = \mathbf{H}^H(\mathbf{H}\mathbf{H}^H)^{-1}$, which we assume that exists for $N_b \geq N_u$. The Hermitian operator is denoted by $(\cdot)^H$.

We introduce a mismatch between the channel information used for precoding and the channel faced by the downlink transmission. Sources of inaccuracy are highlighted in Fig. 1: channel aging, channel estimation and quantization and imperfect synchronization. Independently of the impairment origin, the precoding mismatch can be modeled by an additive channel error. The precoder $\mathbf{W}$ is calculated from $\mathbf{H}$, while data are transmitted over channel $\hat{\mathbf{H}} = \mathbf{H} + \boldsymbol{\Delta}$. Note that in this model, the mean power of the channel used for transmission is $\sigma^2_{\hat{h},j} = \sigma^2_h + \sigma^2_{\delta,j}$, while the mean power of the ideal channel is σ^2_h. Considering additive white Gaussian noise (AWGN) with zero mean and variance σ^2_n at the receiver, we reach

$$\mathbf{y} = \mathbf{s} + \boldsymbol{\Delta}\mathbf{W}\mathbf{s} + \mathbf{n}. \tag{1}$$

As observed, the channel error matrix $\boldsymbol{\Delta}$ breaks the inverse relation between channel and precoder and causes inter-user interference (IUI). Considering a channel matrix with independent, identical distributed (i.i.d.) complex Gaussian random variables (Rayleigh fading channel) with zero mean and variance σ^2_h, it was shown by random matrix theory that the mean SIR (ratio of mean signal to mean IUI power) which a user j observes is given by

$$\mathrm{SIR}_j = \frac{1}{\mathrm{MSE}_j} \cdot \frac{N_b - N_u}{N_u - 1} + \frac{1}{N_u - 1} \tag{2}$$

for $N_b > N_u \geq 2$. The ratio $\sigma^2_{\delta,j}/\sigma^2_h = \mathrm{MSE}_j$ describes the imperfect knowledge which the BSs have about the channel of user j, normalized to the mean channel power. Equation (2) reveals that for ZF, the mean SIR of a single-antenna terminal is inversely proportional to the normalized MSE and depends on the ratio of the number of base stations and terminals.

In [8] it is also shown that the SIR upper bound, which corresponds to a channel matrix with orthonormal channel vectors to all users, is given by

$$\mathrm{SIR}_{j,\max} = \frac{1}{\mathrm{MSE}_j} \cdot \frac{N_b}{N_u - 1} + \frac{1}{N_u - 1}. \tag{3}$$

In what follows, the main contribution of this work is presented, by specializing (1) for each impairment and deriving analytic MSE expressions, which can be used for obtaining the SIR from (2) and (3).

A. Imperfect channel knowledge

In this sub-section the network is assumed to be perfectly synchronized in terms of sampling and carrier frequency. The precoder is calculated from imperfectly estimated, quantized and outdated CSI, which is modeled as follows.

Channel aging: In real-world systems there is always time elapsed between channel estimation at $t = 0$ and when this estimate is used for precoding the downlink transmission. This delay ensues from technical limitations of the equipment involved in the feedback loop from the terminals to the BSs. In [9], the sources of feedback delay were identified and evaluated in our multi-cellular experimental system. Results indicate that the main delay is due to terminal-side procedures, from multi-cell channel estimation until CSI packet construction. The CSI reconstruction at the BSs is also a source of delay, while transmission of the compressed CSI over the air is the fastest part of the loop.

When observing the mobile radio channel for time intervals equal to the feedback delay, we can safely assume that large-scale parameters such as path loss and shadow fading do not change and that any variation is only related to small-scale fading. Therefore, we consider those parameters as constant and, without loss of generality, exclude them from our channel model. Following the analysis in [10], the time-variant channel coefficient h_t at time t can be modeled by

$$h_t = \rho_t h_0 + v_t \ , \tag{4}$$

where ρ_t is the autocorrelation function of the channel h_t. The complex Gaussian variable v_t has zero mean and variance $(1-\rho_t^2)\sigma^2_h$ and is uncorrelated to the channel h_0, which refers to time $t = 0$. Considering Jake's fading model we have $\rho_t = J_0(2\pi f_D t)$, where J_0 is the zero-order first kind Bessel function and $\rho_t \in [-1, 1]$. The maximum Doppler frequency f_D according to terminals velocity v and radio frequency f_c is given by $f_D = v/c \cdot f_c$ (c denotes the speed of light).

Extending to MIMO matrix notation and assuming equal velocity for all terminals, (4) reaches

$$\mathbf{H}_t = \rho_t \mathbf{H}_0 + \mathbf{V}_t. \tag{5}$$

Channel estimation and CSI quantization: Those effects apply at $t = 0$ and can be modeled as

$$\hat{\mathbf{H}}_0 = \mathbf{H}_0 + \mathbf{E}_0 + \mathbf{Q}_0 \ , \tag{6}$$

where $\mathbf{E}_0$ and $\mathbf{Q}_0$ are channel error matrices due to channel estimation and quantization, respectively. Entries of $\mathbf{E}_0$ are i.i.d. complex Gaussian zero-mean variables with variance σ^2_e, while entries of $\mathbf{Q}_0$ are also zero-mean with mean power σ^2_q, respectively. The (non-ideal) ZF precoder is

$$\hat{\mathbf{W}}_0 = \hat{\mathbf{H}}_0^H(\hat{\mathbf{H}}_0\hat{\mathbf{H}}_0^H)^{-1}. \tag{7}$$

The downlink system equation including channel aging and channel estimation and quantization is

$$\mathbf{y}_t = \mathbf{H}_t\hat{\mathbf{W}}_0\mathbf{s}_t + \mathbf{n}_t. \tag{8}$$

By imposing (5) and (7) into (8) and considering the inverse relation between $\hat{\mathbf{H}}_0$ and $\hat{\mathbf{W}}_0$, the downlink equation reaches

$$\mathbf{y}_t = \rho_t \mathbf{s}_t + \mathbf{V}_t \hat{\mathbf{W}}_0 \mathbf{s}_t - \rho_t (\mathbf{E}_0 + \mathbf{Q}_0) \hat{\mathbf{W}}_0 \mathbf{s}_t + \mathbf{n}_t \ . \quad (9)$$

The first term on the right-hand side of (9) is useful signal. The second and third terms are IUI due to channel aging and channel estimation/quantization, respectively. The third term also has a time-varying nature: it takes its maximum absolute value at $t = 0$, when $\rho_0 = 1$, and decreases over time.

In what follows, we derive MSE expressions for each channel impairment separately. Starting with channel aging and disregarding channel estimation and quantization, the third term in (9) becomes zero. We separate useful signal from interference and following the MSE definition as in (2) and (3), we reach

$$\mathrm{MSE}_t = \frac{\mathbb{E}\{|v_t|^2\}}{\mathbb{E}\{|\rho_t h_0|^2\}} = \frac{1-\rho_t^2}{\rho_t^2}. \quad (10)$$

The expectation operator is denoted by $\mathbb{E}\{\cdot\}$. Expression (10) provides the exact MSE for channel model (4). In order to gain insight into how the MSE depends on essential parameters, we simplify the correlation coefficient ρ_t. It is known that the Bessel function can be defined by the Taylor series expansion [11]. For small arguments $x = 2\pi f_D t$, it can be very well approximated by the second order polynomial $J_0(x) \approx 1 - 0.25x^2$, which used in (10) leads to

$$\mathrm{MSE}_t(x) \approx 0.5x^2 \cdot \frac{1-0.125x^2}{(1-0.25x^2)^2}. \quad (11)$$

The second order Taylor series expansion at $x = 0$

$$\frac{1+\alpha x^2}{(1-\beta x^2)^2} \approx 1 + (\alpha + 2\beta)x^2 \quad (12)$$

is applied to the fractional part at the right-hand side of (11) and the MSE reaches

$$\mathrm{MSE}_t(x) \approx 0.5x^2 + 0.1875x^4. \quad (13)$$

For $x \ll 1$, the second term on the right-hand side of (13) is much smaller than the first term and can be thus neglected. If the maximum allowed error is set to 10 % with respect to the exact MSE expression (10), we can use

$$\mathrm{MSE}_t \approx \begin{cases} 2\pi^2 (f_D t)^2, & f_D t < 0.1 \\ 2\pi^2 (f_D t)^2 + 3\pi^4 (f_D t)^4, & 0.1 \le f_D t < 0.2 \end{cases} \quad (14)$$

For the range of validity of (14) this means that, considering a maximum feedback delay of $t = 10$ ms, the second order approximation can be used for velocities up to $v = 3.7$ km/h (carrier frequency is $f_c = 2.65$ GHz) and the MSE is then around -18 dB. The fourth order approximation can be used for up to $v = 8$ km/h, while for larger velocities or feedback delays, higher order approximations shall be required.

Modeling of channel estimation and feedback quantization errors is slightly different than for channel aging. As their physical origin is at the terminal, they already flow into the precoder calculation. In order to use the following expression in (2) and (3), the MSE needs to be normalized to the mean power of the transmission channel $\sigma^2_{\hat{h},j}$. For $t = 0$ we obtain

$$\mathrm{MSE}_{eq} = \frac{\sigma_e^2 + \sigma_q^2}{\sigma_h^2 + \sigma_e^2 + \sigma_q^2}. \quad (15)$$

In OFDM systems the channel is estimated by interpolation over channel observations at pilot tones. The mean power of the channel estimation error σ_e^2 is equal to the receiver AWGN power, divided by estimator gain G. In [12] it is found that in a multi-path Rayleigh fading channel with L channel taps, N_s OFDM subcarriers and a pilot spacing of d subcarriers, the estimator gain G is given by the following expressions:

$$\sigma_e^2 = \frac{\sigma_n^2}{G} \ , \text{ with } \quad G = \frac{N_s}{d \cdot L}. \quad (16)$$

Finally, as found in [13], for linear CSI quantization with B bits and quantization interval X_{max}, the mean error power is

$$\sigma_q^2 = \frac{1}{3} \cdot X_{max}^2 \cdot 2^{-2B}. \quad (17)$$

B. Synchronization impairments

Here we assume perfect channel knowledge and analyze the effect of synchronization impairments. It is known from [14] that OFDM systems are sensitive to carrier frequency offset (CFO). The impact of sampling frequency offset (SFO) is very small compared to CFO and can be thus neglected. In the frequency domain, CFO causes a common phase error on all subcarriers, which can be estimated and compensated as part of the channel frequency response. Additionally, the CFO destroys orthogonality among the subcarriers and causes thus inter-carrier interference (ICI), which is AWGN-like and is not easy to compensate [14].

In JT CoMP, distributed BSs are driven by their individual oscillators. The downlink transmission is therefore impaired by multiple carrier frequency offsets. Modeling and evaluation of the IUI in [3] shows that it is much higher than the ICI. In what follows, ICI shall be therefore neglected.

Each base station, denoted by subindex i, has its own carrier frequency f_i, whereas the ideal carrier frequency is f_c. The CFO ($f_i - f_c$) is modeled as a zero-mean random variable with equal variance σ_f^2 for all BSs. For a single link, the frequency domain representation of the CFO effect is given by

$$\phi_i(t) = e^{j\theta_i(t)}, \text{ with } \ \theta_i(t) = 2\pi(f_i - f_c)t. \quad (18)$$

A closed-form expression has been derived in [3] for $\phi_i(t)$, covering the general case with CFOs and SFOs at both transmitter and receiver sides. Considering here only CFO at the BSs, we see that phase term $\theta_i(t)$ increases linearly with time t.

In the downlink, phase drifts are applied to precoded data symbols before transmission, see Fig. 1. We therefore introduce diagonal matrix $\mathbf{\Phi}_t = \mathrm{diag}(\phi_1(t), ...\phi_i(t), ...\phi_{N_b}(t))$ into the downlink equation and develop (1) as

$$\mathbf{y} = \mathbf{H}\mathbf{\Phi}_t \mathbf{W}\mathbf{s} + \mathbf{n}. \quad (19)$$

Matrix $\mathbf{\Phi}_t$ is reset to zero each time the precoder is updated. Residual phase terms are considered as part of the channel. Equation (19) is thus re-formulated as

$$\mathbf{y} = \mathbf{s} + \underbrace{\mathbf{H}(\mathbf{\Phi}_t - \mathbf{I})}_{\triangleq \mathbf{\Delta}_\Phi} \mathbf{W}\mathbf{s} + \mathbf{n} \tag{20}$$

where $\mathbf{\Delta}_\Phi$ can be interpreted as an *equivalent* channel error due to the CFO. Matrix $\mathbf{\Phi}_t$ is diagonal; column i of $\mathbf{H}$ is multiplied with $\phi_i(t) - 1$. Entries of $\mathbf{\Delta}_\Phi$ are thus $\delta_{\phi,ji} = h_{ji}(\phi_i(t) - 1)$. Their mean power equals to $\mathbb{E}\{|\delta_{\phi,ji}|^2\} = \mathbb{E}\{|h_{ji}|^2\} \cdot \mathbb{E}\{|\phi_i(t) - 1|^2\}$, as h_{ji} and $\phi_i(t)$ are uncorrelated. The normalized MSE results into

$$\begin{aligned} \text{MSE}_\phi &= \mathbb{E}\{|\phi_i(t) - 1|^2\} \\ &= 2 \cdot (1 - \text{Re}\{\bar{\phi}(t)\}). \end{aligned} \tag{21}$$

with $\mathbb{E}\{\phi_i(t)\} = \bar{\phi}(t)$, $\{\cdot\}^*$ the complex conjugate and $\text{Re}\{\cdot\}$ the real part of a complex number. We now simplify by applying small angle approximation $e^{jx} \approx 1 + jx$ to $\phi_i(t)$:

$$\phi_i(t) \approx 1 + \jmath\, \theta_i(t) \quad \text{and} \quad \mathbf{\Phi}_t \approx \mathbf{I} + j\mathbf{\Theta}_t. \tag{22}$$

Imposing (22) into (19), the received signal reaches

$$\mathbf{y} \approx \mathbf{s} + j\mathbf{H}\mathbf{\Theta}_t\mathbf{W}\mathbf{s} + \mathbf{n}. \tag{23}$$

The second term is the inter-user interference due to non-orthogonal transmission to different users, caused by imperfect synchronization among base stations.

From (23) and (20) it is straightforward that $\mathbf{\Delta}_\Phi \approx j\mathbf{H}\mathbf{\Theta}_t$ and $\delta_{\phi,ji} \approx jh_{ji}\theta_i$. As h_{ji} and θ_i are uncorrelated and both zero-mean, we have $\mathbb{E}\{|\delta_{\phi,ji}|^2\} = \sigma_h^2\sigma_\theta^2$, with phase variance $\sigma_\theta^2 = (2\pi)^2\sigma_f^2t^2$. Finally, the normalized MSE is

$$\text{MSE}_\phi \approx (2\pi)^2\sigma_f^2t^2. \tag{24}$$

Equation (24) reveals that the MSE is proportional to the CFO variance σ_f^2 and grows quadratic with time. If the approximation error is desired to remain under 10 % for a feedback delay up to $t = 10$ ms, σ_f needs to be less than 10 Hz. This requires an oscillator accuracy of $\text{Oc} = \sigma_f/f_c = 4 \cdot 10^{-9}$, which is close to the 3GPP reference of $\text{Oc} = 5 \cdot 10^{-9}$. The resulting MSE is around -17.5 dB.

Since for a certain range of normalized delay ($f_Dt < 0.1$) and synchronization accuracy ($\text{Oc} \leq 4 \cdot 10^{-9}$), both MSE grow quadratic with time, their ratio remains constant over time:

$$R_{MSE} = \frac{\text{MSE}_t}{\text{MSE}_\phi} \approx \frac{f_D^2}{2 \cdot \sigma_f^2}. \tag{25}$$

From a system design point of view, it is reasonable to target both MSE to be in the same order of magnitude, i.e. $R_{MSE} = 1$. This means that, given mobility, synchronization accuracy does not need to be higher than (25). Thus, the overall MSE is 3 dB higher than the MSE due to each of the impairments.

III. Evaluation of MSE and Spectral Efficiency

For main impairments of channel aging and CFO, the MSE approximations are validated with respect to the exact analytical expressions and numerical results. Additionally, spectral efficiency in JT CoMP is evaluated for a practical scenario.

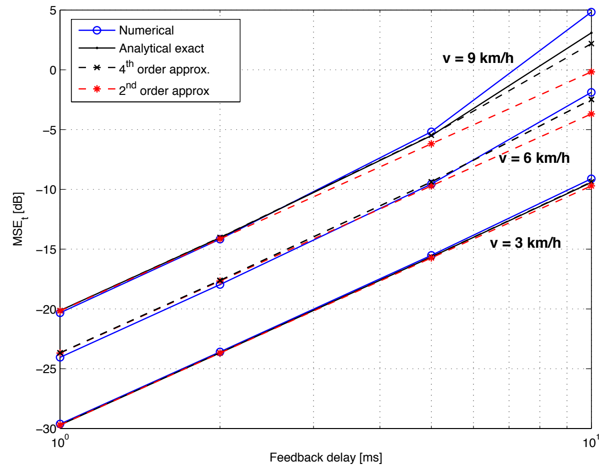

Fig. 3. MSE due to channel aging according to numerical evaluation and exact analytic expression (10), second and fourth order approximation (14).

A. Mean square error evaluation

Fig. 3 evaluates MSE due to channel aging according to the exact expression (10) (analytically and numerically) and approximation (14), for a terminal velocity of v = 3, 6 and 9 km/h and feedback delay up to 10 ms. For numerical evaluation, 10^4 i.i.d. Rayleigh channel initializations were observed over the feedback delay time, and $\mathbb{E}\{|v_t|^2\}$ and ρ_t were statistically evaluated. As observed, doubling the feedback delay or velocity increases the MSE by around 6 dB.

Fig. 4 evaluates MSE due to the BSs CFO according to the exact expression (21) and approximation (24). The exact expression is numerically evaluated over 10^4 i.i.d. CFOs, considering typical oscillators with RMS values $\sigma_f = \text{Oc} \cdot f_c$. It can be observed that by reducing the oscillator accuracy by one order of magnitude, the MSE increases by 20 dB. Doubling the feedback delay also increases the MSE by 6 dB.

Approximations are very tight to the exact expressions. Results indicate that channel aging has a larger impact on MSE: already for pedestrian speed $v = 3\ km/h$, MSE is around 5 dB higher than the one due to an average-quality oscillator with $\text{Oc} = 10^{-9}$.

The MSE due to channel estimation and quantization depends on pilot density and number of quantization bits, respectively. By proper parametrization, the MSE remains under a desired level, which does not need to be lower than the already existent MSE due to channel aging and synchronization.

B. Spectral efficiency in JT CoMP

Spectral efficiency can be calculated by expression (26), provided by [15]. Fig. 5 reveals that spectral efficiency in JT CoMP is very sensitive to impairments; it decreases quickly with feedback delay time already for pedestrian mobility.

$$\text{c} = 0.9 \cdot \log_2(1 + 0.85 \cdot \text{SIR}) - 0.18. \tag{26}$$

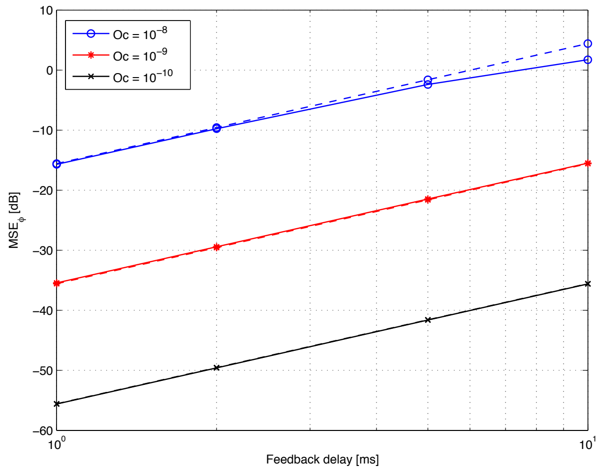

Fig. 4. Equivalent channel MSE due to oscillator CFO. *solid*: numerical evaluation of exact expression (21); *dashed*: analytical approximation (24).

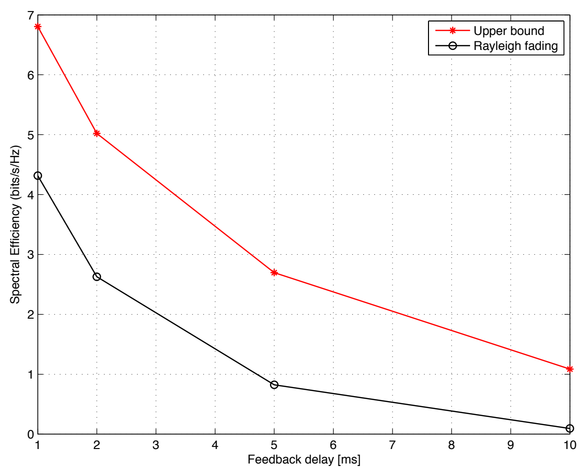

Fig. 5. Spectral efficiency for JT CoMP with ZF precoding under the presence of channel aging and synchronization errors. Parameters are given in Table I.

IV. Conclusion

We have presented a signal model for the downlink of cooperative networks using zero-forcing precoding, including channel aging, channel estimation, feedback quantization and imperfect synchronization of base stations. We derived exact expressions and accurate approximations for the mean square error, which were validated numerically. Analytical SIR expressions were used for calculating the practical spectral efficiency. Results indicate that short feedback delay times and channel prediction or other compensation techniques shall be required, in order to support higher mobilities. Accurate synchronization is also essential for cooperative base stations.

TABLE I
Evaluation Parameters

Number of base stations (N_b)	7
Number of mobile users (N_u)	6
Terminal velocity (v)	6 km/h
Oscillator accuracy (Oc)	10^{-9}
Carrier frequency (f_c)	2.65 GHz
Overall MSE (dB)	-23.4; -17.4; -9.2; -2.2

Acknowledgments

This work was funded by the Deutsche Forschungsgemeinschaft (DFG) under project Nr. JU 2793/3-1 and JU 2793/4-1 and by projects from CONICYT, Departamento de Relaciones Internacionales “Programa de Cooperación Científica Internacional” CONICYT/DFG-622 and FONDECYT 1110370.

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