Interference Suppression for Next Generation Wireless Systems

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***Abstract*—Future wireless communication systems being characterized by tight frequency reuse, adaptive modulation and coding schemes and diversified data services will be interference limited by the interfering signals of diverse rates and strengths. Frequency reuse factor of 1 being proposed for the next generation mobile systems will lead to 1 or 2 dominant interferers. We propose in this paper a low complexity maximum likelihood (ML) demodulator for interference suppression. The proposed demodulator is also applicable to general multi-stream (spatially multiplexed) MIMO systems where it reduces one complex dimension of the system i.e. the complexity of ML detection reduces from $\mathcal{O}\left(|\chi|^{n}\right)$ to $\mathcal{O}\left(|\chi|^{n-1}\right)$ where n is number of transmit antennas/spatial streams. This is a fundamental result as space and technology constraints shall be restricting future MIMO systems to low dimensionality. Therefore reduction of one complex dimension in detection shall enable implementation of ML detectors at the receivers. We look at the performance of linear MMSE demodulator and the proposed demodulator in the presence of interferers of diverse strengths and rates and simulation results demonstrate much improved performance of the proposed demodulator for interference suppression.**

I. Introduction

To cope with the ever-increasing demands on higher spectral efficiency, a tight frequency reuse will be adopted for future wireless communication systems as 3GPP LTE [1], IEEE 802.11n [2] and IEEE 802.16m [3]. Adaptive modulation and coding schemes will be supported in the next generation wireless systems which combined with the diversified data services will lead to variable transmission rate streams. These system characteristics will overall lead to an interference-limited system. Most state-of-the-art wireless systems deal with the interference either by orthogonalizing the communication links in time or frequency [4] or allow the communication links to share the same degrees of freedom but model the interference as additive Gaussian random process [5]. Both of these approaches may be suboptimal as the first approach entails an *a priori* loss of degrees of freedom in both links, no matter how weak the potential interference is while the second approach treats interference as pure noise while it actually carries information and has structure that can be potentially exploited in mitigating its effect.

3GPP LTE [1] has chosen orthogonal frequency division multiple access (OFDMA) technology for the downlink in order to provide multiple access and eliminate the intracell interference. However frequency reuse factor being 1 will lead to intercell (co-channel) interference impairments. In most cases, dominant interferers will be 1 (near cell boundaries) or 2 (near cell edges). Linear spatial filters as minimum mean square error (MMSE) and zero forcing (ZF) filters are the focus of attention to minimize the level of interference in the former case while nulling out the interference in the latter case. Linear MMSE filters because of their low complexity and comparatively better performance than ZF are being considered as likely candidates for 3GPP LTE [6]. The suboptimality of MMSE for non Gaussian alphabets in low dimensional systems (less number of interferers) is well known [7] and moreover MMSE detection being based on interference attenuation is void of exploiting the interference structure in mitigating its effect. The optimal maximum likelihood (ML) detector because of its associated high complexity prohibits its implementation in the practical systems.

We propose an optimal low complexity demodulator for interference suppression in the single frequency reuse synchronized cellular networks. This demodulator is a reduced complexity version of the max log MAP demodulator and is based on the matched filter outputs. The proposed demodulator has less complexity and more generality than an earlier proposed matched filter based demodulator [8] which was confined to the case of merely one interference. However the new demodulator can be used for any number of interferers (general MIMO system) and it successfully reduces one complex dimension. The proposed demodulator can also be incorporated in the strategies for interference cancellation i.e partial interference cancellation (PIC) and absolute interference cancellation (AIC) [9]. This demodulator can be termed as interference suppression or partial decoding of the interference which is the recommended strategy in the regime of moderate interference [10]. Simulation results demonstrate much improved performance of the proposed demodulator with respect to the other suboptimal linear detectors as MMSE.

Regarding notations, we will use lowercase or uppercase letters for scalars, lowercase boldface letters for vectors and uppercase boldface letters for matrices. $\Re(.)$ or subscript $_R$ indicates real part while $\Im(.)$ or subscript $_I$ indicates imaginary part of a complex number. $(.)^T$ indicates transpose, $(.)^*$ indicates conjugate while $(.)^\dagger$ indicates conjugate transpose. $|.|$ and $\|.\|$ indicate norm of scalar and vector respectively. The paper is divided into five sections. In section II we define the system model and give a brief information theoretic analysis while section III discusses the MMSE and the proposed de-

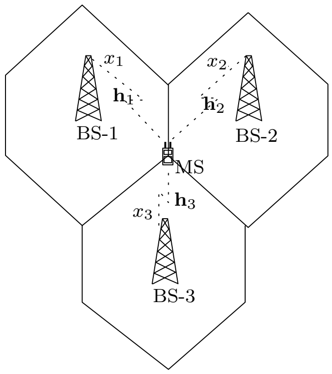

Fig. 1. Interference cancellation in single frequency cellular network. x_1 is the desired signal while x_2 and x_3 are the interference signals.

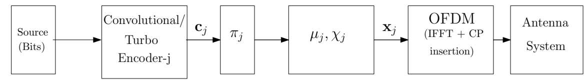

Fig. 2. Block diagram of Transmission chain of BICM OFDM system at j-th BS. π_j denotes random interleaver, μ_j labeling map and χ_j signal set for $\mathbf{x}_j$.

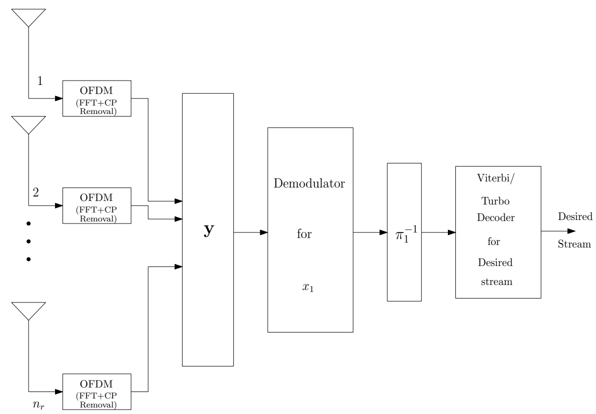

Fig. 3. Block diagram of receiver at MS.

modulator for interference suppression. Section IV is dedicated to the simulation results which is followed by the conclusions.

II. System Model

Consider a single frequency reuse cellular network as shown in Fig. 1. Keeping in view the upcoming wireless standards as LTE [1], IEEE 802.11n [2] and IEEE 802.16m [3], we assume that the base stations (BSs) use bit interleaved coded modulation (BICM) [11] based OFDM system for downlink transmission. We assume antenna cycling at the BS [12] with each stream being transmitted by one antenna in any dimension. The antenna used by a particular stream is randomly assigned per dimension so that each stream sees all the degrees of freedom of the channel. The block diagram of the transmission chain at BS and reception chain at the mobile station (MS) are shown in the Figs. 2 and 3 respectively. We assume receive diversity at the MS with n_r receive antennas. Though the dominant interferers can be 2 in most cases but for generality, we consider $n-1$ interferers. Let the spatial streams arriving at the MS be $\mathbf{x}_1$ (desired stream) and $\mathbf{x}_2, \cdots, \mathbf{x}_n$ (interference streams). x_j is the symbol of $\mathbf{x}_j$ over a signal set $\chi_j \subseteq \mathcal{C}$ of size $|\chi_j| = M_j$ with a Gray labeling map $\mu_j : \{0,1\}^{\log|M_j|} \rightarrow \chi_j$. During transmission, the code sequence $\mathbf{c}_j$ is interleaved by π_j and then mapped onto the signal sequence $\mathbf{x}_j \in \chi_j$. The bit interleaver for the j-th stream can be modeled as $\pi_j : k^{'} \rightarrow (k,i)$ where $k^{'}$ denotes the original ordering of the coded bits $c_{j,k'}$, k denotes the time ordering of the signals $x_{j,k}$ and i indicates the position of the bit $c_{j,k'}$ in the symbol $x_{j,k}$.

We assume that the frequency reuse factor is one and the cyclic prefix (CP) of appropriate length is added to the OFDM symbols at the BSs. We further assume that BSs are synchronized for transmission and there is no channel state information (CSI) at the BS while perfect CSI of the desired and the interference streams is assumed at the MS. Cascading the IFFT at the BS and the FFT at the MS with CP extension, transmission at the k-th frequency tone can be expressed as:-

$$\mathbf{y}_k = \mathbf{h}_{1,k}x_{1,k} + \mathbf{h}_{2,k}x_{2,k} + \cdots + \mathbf{h}_{n,k}x_{n,k} + \mathbf{z}_k, \quad k = 1, 2, \cdots, K \tag{1}$$

where K is the total number of frequency tones and n is the number of signals being received at the MS. We assume that the subcarriers are narrowband and model each subcarrier as a frequency flat fading channel so $\mathbf{h}_{1,k} \in \mathbb{C}^{n_r}$ is the vector characterizing flat fading channel response from first BS to n_r receive antennas at k-th subcarrier. This vector has complex-valued multivariate Gaussian distribution with $E\left[\mathbf{h}_{1,k}\right] = \mathbf{0}$ and $E\left[\mathbf{h}_{1,k}\mathbf{h}_{1,k}^{\dagger}\right] = \mathbf{I}$ i.e. each channel between the BS and the n_r receive antennas is independent while the channels at different subcarriers are also assumed to be independent. Each subcarrier corresponds to a symbol from a constellation map χ_j for the j-th stream. $\mathbf{y}_k, \mathbf{z}_k \in \mathbb{C}^{n_r}$ are the vectors of received symbols and circularly symmetric complex white Gaussian noise of double-sided power spectral density $N_0/2$ at the n_r receive antennas at k-th frequency tone. The complex symbols $x_{1,k}, x_{2,k}, \cdots, x_{n,k}$ of the n streams are also assumed to be independent and of variances $\sigma_1^2, \sigma_2^2, \cdots, \sigma_n^2$ respectively.

To consider the effect of the interference strength and of the rate of interference (alphabet size), we confine to the case of one strong interference. We focus on the mutual information of the desired stream in the presence of one strong interferer. Dropping the frequency index, the mutual information of desired stream is given as [13]

$$I\left(\mathbf{y}; x_1\right) = \log M_1 - \frac{1}{M_1}\sum_{x_1}\int_{\mathbf{y}} p\left(\mathbf{y}|x_1\right)\log\frac{\sum_{x_1} p\left(\mathbf{y}|x_1\right)}{p\left(\mathbf{y}|x_1\right)} d\mathbf{y} \tag{2}$$

Fig. 4 shows the mutual information of the desired stream in the presence of the interference stream. We define the term $\alpha = \sigma_2^2/\sigma_1^2$. Mutual information of the desired stream is a function of the rate as well as the strength of the interference stream. For moderate values of α and when the interference

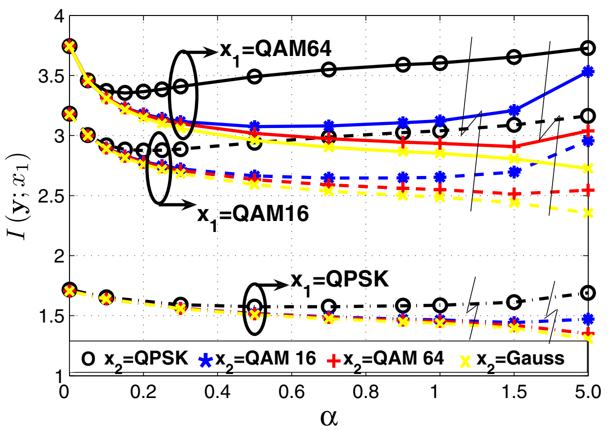

Fig. 4. Capacity of the desired stream x_1 in the presence of the interference stream x_2 for different constellations. SNR is 4.5 dB for x_1=QPSK, 11 dB for x_1=QAM16 and 13 dB for x_1=QAM64. Note that the flash sign indicates a discontinuity of abscissa.

has a lower rate (smaller constellation size) relative to the desired stream, as the interference strength increases, the mutual information of the desired stream increases. However when the interference stream has a higher rate as compared to the rate of the desired stream, this behavior is observed for higher values of α. This can be interpreted as the decoding capability of the MS of the interference in the presence of the desired stream. Once the interference strength and its rate relative to the strength and the rate of the desired stream permits the decoding of the interference, we observe an increase in the mutual information of the desired stream with the increase of α. Fig. 4 also authenticates the well known result of Gaussian being the worst case interference however the gap decreases as the rate of the interference stream increases. This diminution of gap may be related to the proximity of the behavior of large size constellations to Gaussianity as both are characterized by high peak to average power ratios.

III. Demodulators for Interference Suppression

A. MMSE

The frequency domain MMSE filter for $x_{1,k}$ is given as

$$\mathbf{h}_{1,k}^{MMSE} = \left(\mathbf{h}_{1,k}^{\dagger}\mathbf{R}_{1,k}^{-1}\mathbf{h}_{1,k} + \sigma_1^{-2}\right)^{-1}\mathbf{h}_{1,k}^{\dagger}\mathbf{R}_{1,k}^{-1} \tag{3}$$

where $\mathbf{R}_{1,k} = \sigma_2^2\mathbf{h}_{2,k}\mathbf{h}_{2,k}^{\dagger} + \sigma_3^2\mathbf{h}_{3,k}\mathbf{h}_{3,k}^{\dagger} + \cdots + \sigma_n^2\mathbf{h}_{n,k}\mathbf{h}_{n,k}^{\dagger} + N_0\mathbf{I}$. After the application of MMSE filter we get

$$y_k = \alpha_k x_{1,k} + z_k \tag{4}$$

where z_k is assumed to be zero mean complex Gaussian random variable with variance $N_k = \mathbf{h}_{1,k}^{MMSE}\mathbf{R}_{1,k}\mathbf{h}_{1,k}^{MMSE^{\dagger}}$ and $\alpha_k = \mathbf{h}_{1,k}^{MMSE}\mathbf{h}_{1,k}$. Gaussianity has been assumed for post detection interference which increases the suboptimality of MMSE in the case of less number of interferers. Bit metric for the $c_{k'}$ bit on the first stream is given as

$$\lambda_1^i\left(\mathbf{y}_k, c_{k'}\right) \approx \min_{x_1 \in \chi_{1,c_{k'}}^i} \left[\frac{1}{N_k}\left|y_k - \alpha_k x_1\right|^2\right] \tag{5}$$

where $\chi_{1,c_{k'}}^i$ denotes the subset of the signal set $x_1 \in \chi_1$ whose labels have the value $c_{k'} \in \{0,1\}$ in the position i. This metric has computational complexity $\mathcal{O}\left(|\chi_1|\right)$.

B. Reduced Complexity Max Log MAP

Considering the system equation (1), the max log MAP bit metric is given as [11]

$$\lambda_1^i\left(\mathbf{y}_k, c_{k'}\right) \approx \min_{x_1 \in \chi_{1,c_{k'}}^i, x_2 \in \chi_2, \cdots, x_n \in \chi_n} \left\|\mathbf{y}_k - \mathbf{h}_{1,k}x_1 - \cdots - \mathbf{h}_{n,k}x_n\right\|^2 \tag{6}$$

which has computational complexity $\mathcal{O}\left(|\chi_1|\cdots|\chi_n|\right)$. For brevity we drop the frequency index k and the bit position index $k^{'}$ i.e.

$$\begin{aligned}
\lambda_1^i\left(\mathbf{y}, c\right) &\approx \min_{x_1 \in \chi_{1,c}^i, x_2 \in \chi_2, \cdots, x_n \in \chi_n} \left\|\mathbf{y} - \mathbf{h}_1 x_1 - \cdots - \mathbf{h}_n x_n\right\|^2 \\
&= \min_{x_1 \in \chi_{1,c}^i, x_2 \in \chi_2, \cdots, x_n \in \chi_n} \Bigg\{\left\|\mathbf{y}\right\|^2 + \sum_{j=1}^{n}\left\|\mathbf{h}_j x_j\right\|^2 \\
&\quad + 2\Re\sum_{j=1}^{n-1}\sum_{l=j+1}^{n}\left(\mathbf{h}_j x_j\right)^{\dagger}\left(\mathbf{h}_l x_l\right) - 2\Re\sum_{j=1}^{n}\left(\mathbf{h}_j^{\dagger}\mathbf{y}\right)x_j^*\Bigg\} \\
&= \min_{x_1 \in \chi_{1,c}^i, x_2 \in \chi_2, \cdots, x_n \in \chi_n} \Bigg\{\left\|\mathbf{y}\right\|^2 + \sum_{j=1}^{n-1}\left\|\mathbf{h}_j x_j\right\|^2 \\
&\quad + 2\Re\sum_{j=1}^{n-1}\sum_{l=j+1}^{n-1} p_{jl}x_j^* x_l - 2\Re\sum_{j=1}^{n-1} y_j x_j^* \\
&\quad + 2\Re\sum_{j=1}^{n-1} p_{jn}x_j^* x_n - 2\Re y_n x_n^* + \left\|\mathbf{h}_n x_n\right\|^2\Bigg\}
\end{aligned} \tag{7}$$

where $y_k = \mathbf{h}_k^{\dagger}\mathbf{y}$ be the MF output for the k-th stream and $p_{km} = \mathbf{h}_k^{\dagger}\mathbf{h}_m$ be the cross correlation between the k-th and the m-th channel. Breaking some of the terms in their real and imaginary parts with subscripts $_R$ and $_I$ indicating real and imaginary parts of a complex number, we have

$$\begin{aligned}
&= \min_{x_1 \in \chi_{1,c}^i \cdots x_n \in \chi_n} \Bigg\{\sum_{j=1}^{n-1}\left\|\mathbf{h}_j x_j\right\|^2 + 2\Re\sum_{j=1}^{n-1}\sum_{l=j+1}^{n-1} p_{jl}x_j^* x_l - 2\Re\sum_{j=1}^{n-1} y_j x_j^* \\
&\quad + \left(2\sum_{j=1}^{n-1}\left(p_{jn,R}x_{j,R} + p_{jn,I}x_{j,I}\right) - 2y_{n,R}\right)x_{n,R} + \left\|\mathbf{h}_n\right\|^2 x_{n,R}^2 \\
&\quad + \left(2\sum_{j=1}^{n-1}\left(p_{jn,R}x_{j,I} - p_{jn,I}x_{j,R}\right) - 2y_{n,I}\right)x_{n,I} + \left\|\mathbf{h}_n\right\|^2 x_{n,I}^2\Bigg\}
\end{aligned} \tag{8}$$

This equation reduces one complex dimension of the system. For x_n belonging to equal energy alphabets, the bit metric is written as

$$= \min_{x_1 \in \chi^i_{1,c} \cdots x_{n-1} \in \chi_{n-1}} \Bigg\{ \sum_{j=1}^{n-1} \|\mathbf{h}_j x_j\|^2 + 2\Re \sum_{j=1}^{n-1} \sum_{l=j+1}^{n-1} p_{jl} x_j^* x_l - 2\Re \sum_{j=1}^{n-1} \\ - \left| 2 \sum_{j=1}^{n-1} \left(p_{jn,R} x_{j,R} + p_{jn,I} x_{j,I} \right) - 2 y_{n,R} \right| |x_{n,R}| \\ - \left| 2 \sum_{j=1}^{n-1} \left(p_{jn,R} x_{j,I} - p_{jn,I} x_{j,R} \right) - 2 y_{n,I} \right| |x_{n,I}| \Bigg\} \tag{9}$$

For x_n belonging to non-equal energy alphabets, it's real and imaginary part which minimizes (8) are given as

$$x_{n,R} \to -\frac{\sum_{j=1}^{n-1} \left(p_{jn,R} x_{j,R} + p_{jn,I} x_{j,I} \right) - y_{n,R}}{\|\mathbf{h}_n\|^2}$$
$$x_{n,I} \to -\frac{\sum_{j=1}^{n-1} \left(p_{jn,R} x_{j,I} - p_{jn,I} x_{j,R} \right) - y_{n,I}}{\|\mathbf{h}_n\|^2} \tag{10}$$

where $\to$ indicates the quantization process in which amongst the finite available points, the point closest to the calculated continuous value is selected.

This bit metric implies reduction in complexity to $\mathcal{O}\left(|\chi_1| \cdots |\chi_{n-1}|\right)$. The reduction of one complex dimension without any additional processing is a fundamental result of significant importance for lower dimensional systems. Additionally this bit metric is based on matched filter outputs and channel correlations and is therefore simpler for fixed point implementations. The intricacy in the practical implementation of a higher dimensional MIMO system due to space (requisite antenna spacing) and technology constraints underlines the significance of complexity reduction algorithms for lower dimensional systems. The requisites for the proposed demodulator are the knowledge of interference channels and the modulation and coding scheme (MCS) of interfering streams. BSs need to be synchronous with the pilot signals from the adjacent BSs to be orthogonal to meet these requisites. MMSE based demodulators involve computationally complex operations of matrix inversions which are very hard for fixed point implementations. Though the MMSE demodulator only requires the knowledge of interference channels and not their modulation and coding schemes but it additionally needs the knowledge of noise variance.

C. Simulation Results

Moderate and high SNR regime in the interference-limited scenario demands more attention as when the noise is small, interference will have a significant impact on the performance. Low SNR regime is less interesting since here the performance is noise-limited and interference is not having a significant effect. For simulations, we have restricted ourselves to the case of one strong interference. These simulations have been

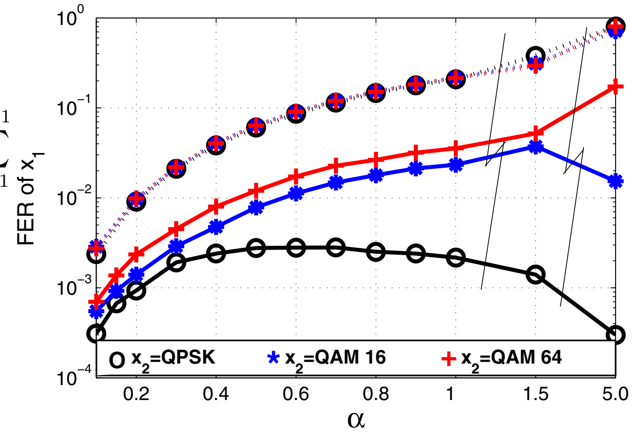

Fig. 5. Desired stream x_1 is QPSK while interference stream x_2 is from QPSK, QAM16 and QAM64. SNR is 4.5 dB. Continuous lines indicate proposed approach while dotted lines indicate MMSE approach. $64-$state, rate $1/2$ Convolutional Code is used

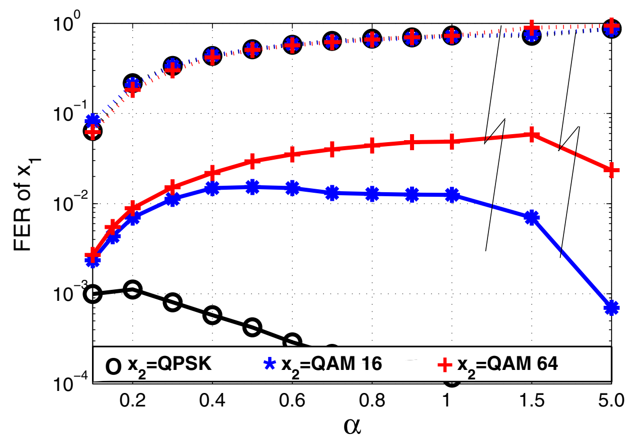

Fig. 6. Desired stream x_1 is QAM 16 while interference stream x_2 is from QPSK, QAM16 and QAM64. SNR is 11 dB. Continuous lines indicate proposed approach while dotted lines indicate MMSE approach. $64-$state, rate $1/2$ Convolutional Code is used.

performed in moderate and high SNR region while the interference strength is being varied. Note that for a fixed SNR, $\alpha = \sigma_2^2/\sigma_1^2$ can be interpreted as the interference strength.

We consider 2 BSs each using BICM OFDM system for downlink transmission using the *de facto* standard, 64 state $(133, 171)$ rate-$1/2$ convolutional encoder of 802.11n standard [2] and the punctured rate $1/2$ turbo code of 3GPP LTE [1][1]. MS has two antennas. We consider an ideal OFDM based system (no ISI) and analyze the system in frequency domain. Due to bit interleaving followed by OFDM, this can be termed as frequency interleaving. Therefore SIMO channel at each sub carrier from BS to MS has iid Gaussian matrix entries with unit variance. Perfect CSI is assumed at the receiver.

[1]The LTE turbo decoder design was performed using the coded modulation library www.iterativesolutions.com

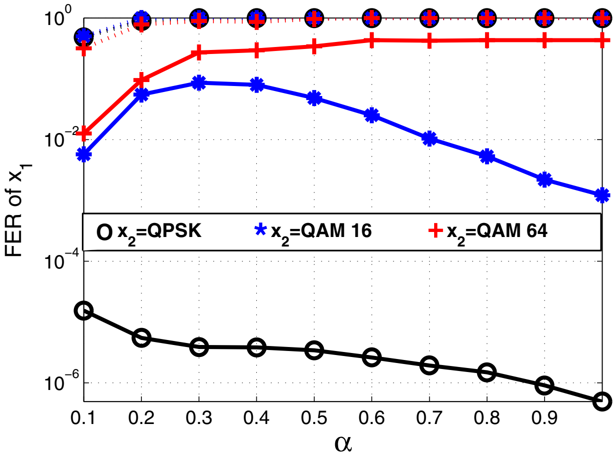

Fig. 7. Desired stream x_1 is QAM 64 while interference stream x_2 is from QPSK, QAM16 and QAM64. SNR is 13 dB. Continuous lines indicate proposed approach while dotted lines indicate MMSE approach. Punctured rate $1/2$ 3GPP turbo code is used with 5 decoding iterations.

Furthermore, all mappings of coded bits to QAM symbols use Gray encoding. We consider the MMSE approach and the proposed approach.

Figs. 5, 6 and 7 show the frame error rates of target stream in the presence of one interference stream. These simulation results show that the dependence of the performance for MMSE detection is insignificant on the rate of the interference stream but its dependence on interference strength is substantial. This can be interperated as a consequence of the attenuation of interference strength at the output of MMSE filter and the subsequent assumption of Gaussianity for its behavior. For the proposed approach, a significant improvement is observed in the performance as the rate of interference stream decreases which is in conformity with the earlier results of mutual information analysis (fig. 4). It is observed that for a given interference level, the performance is generally degraded as the rate (constellation size) of the interfering stream increases. The performance gap with respect to MMSE decreases as the desired and the interference streams grow in constellation size which can be attributed to the proximity to the Gaussianity of these larger constellations due to their high peak to average power ratio and to the optimality of MMSE for Gaussian alphabets.

IV. Conclusions

We have presented in this paper a modified form of the max log MAP demodulator which reduces one complex dimension of the system. We have used this low complexity demodulator for interference suppression. We have also shown that the suboptimality of MMSE increases as the interference streams get stronger. Contrarily the proposed demodulator performs better as the strength of the interference streams increase due to its enhanced ability of interference suppression in case of stronger interferers.

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