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Spatial Filtering Approach for Dynamic Range Reduction in Cognitive Radios

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***Abstract*— Cognitive radios (CRs) receive several users simultaneously. Therefore, the ADC of an CR requires a large dynamic range (DR) to guarantee adequate resolution per user. Thus far the ADC DR requirements have been prohibitive for the wide spread introduction of CR in the hand held market. The power consumption of an ADC reduces with an order of magnitude per decade. In the absence of disruptive new technologies, we expect this power trend to continue for the foreseeable future. Therefore, we propose an analog spatial filtering technique and present iterative methods to alleviate the ADC DR requirements and accelerate the overall power reduction of a CR. Simulation results indicate that for realistic scenarios the ADC resolution can be reduced by 4 bits per ADC, reducing the overall ADC power consumption with more than 90%.**

I. INTRODUCTION

As the number of users in the crowded pre-allocated parts of the radio spectrum continues to increase, a more efficient use of the available spectrum is required. Cognitive radio (CR) promises to vastly improve the efficiency of spectral use. A CR continuously senses and uses unoccupied channels in a wide band spectrum, alleviating congestion and improving the overall throughput.

In order to adequately receive and reconstruct the transmitted data from the desired signal, both coordinated and random interfering signal sources need to be filtered. The filtering of interference can occur in combinations of domains, such as time domain filtering, code domain filtering, frequency domain filtering, and spatial domain filtering.

We observe that in most systems the radio frequency (RF) front end and analog-to-digital converter (ADC) handle the interfering users via extra dynamic range (DR) and linearity, which make them power hungry [1]. The DR and over sample frequency of the ADC are a combination of the interference power level with which the receiver needs to cope, both in and outer band, and a trade off with the analog filter drop off [2].

However, modern MIMO systems make it feasible to include spatial filtering techniques in the analog front end. Since the channel is selective in the angular domain, such techniques have the potential to vastly reduce the interference power and thus the DR and linearity requirements of both the RF front end and ADC. Therefore, a spatial filtering technique can be the basis for ultra low power high data rate wireless receivers, and has the potential to mitigate the DR requirements of cognitive radios.

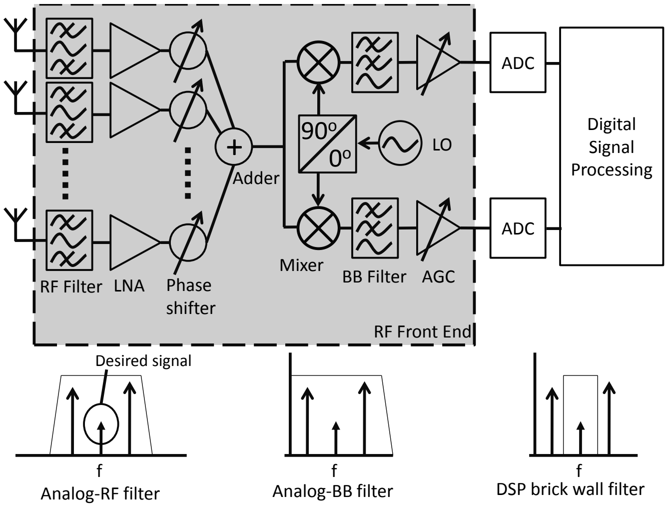

Fig. 1: *Receiver Model.*

In this paper we will focus on the ADC requirements, since the power requirements of ADCs have proven to be prohibitive for the widespread use of CR. In related work we focus on the RF front end throughput optimization for limited available circuit power [3]. Currently, the power consumption of ADCs reduces approximately with an order of magnitude every decade [4] [5]. In the absence of disruptive new technologies, we expect this power trend to continue for the foreseeable future. Therefore, the mitigation of DR requirements has the potential to accelerate the wide spread use of cognitive radios in the hand held market.

Since the wireless channel is selective in the angular domain, we propose an analog spatial filtering approach to mitigate the power requirements of the ADC. The aim of our analysis is to determine the potential benefits of spatial filtering in terms of increased throughput, and to quantize the potential power savings in terms of ADC bits.

II. SYSTEM MODEL

In a typical CR scenario there are multiple primary and secondary user transmissions over a wide bandwidth. The desired user only occupies a small portion of the band [6]. The entire band is filtered, down converted, and finally sampled by an ADC. The user selection occurs in the digital domain.

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Therefore, the first part of the analog signal conditioning (ASC) part of the receiver, has to deal with a broadband RF signal, which contains both interfering users and the desired signal. To prevent clipping and excessive spill over of interfering signals into the desired signal band, highly linear large DR receivers are required. After analog channel selection via mixing and non ideal analog filtering, the signal presented to the analog-to-digital converter (ADC) consist of the desired user, interfering users, and left over adjacent channel interference products. To adequately quantize the desired signal, over sampling to prevent aliasing of the adjacent channel interferer into the desired channel is necessary, and headroom bits are required to guarantee adequate resolution for the desired signal. Finally the baseband (BB) digital signal processing (DSP) brick wall filters the adjacent channel interferers in the digital domain. A co-channel interferer can be filtered digitally using a combination of coding, timing, and frequency hopping schemes. In multiple input multiple output (MIMO) antenna systems the BB DSP can also use beamforming techniques to spatially filter co channel interferers, but this requires a power hungry large DR ADC per antenna. Given that we strive to minimize ADC power consumption, digital spatial filtering is currently considered unpractical for hand held devices.

A. Architecture

Figure 1 depicts the architecture of the receiver. The considered system architecture consists of a commonly used zero IF architecture combined with several receiver antennas, each with their own RF-filter, LNA, and phase shifter, to allow for analog spatial filtering. The outputs of the phase shifters are combined and presented to an in quadrature I and quadrature Q mixer which down converts the combined RF frequency band to baseband with the signal provided by the local oscillator (LO). The I and Q baseband signals containing all users are low pass filtered, set to an adequate gain for the ADC by an automatic-gain-control (AGC), and than quantized by an ADC. Final channel selection occurs in the digital domain via digital signal processing.

B. Analog-to-Digital Converter

The ADC is required to quantize several different users with various received power levels. To guarantee adequate resolution and dynamic range for the desired user, additional headroom bits are required.

1) Power Model: Unfortunately, the power scaling of an ADC is exponential in the number of bits

$$P_{\text{ADC}} \sim \kappa_t F_s 2^{b_{\text{ADC}}}, \tag{1}$$

here F_s is the sample frequency of the ADC, κ_t a technology constant and b_{ADC} the number of ADC bits [4] [5].

2) Headroom Bits: In a cognitive radio several users with various power levels are presented to the baseband ADC. The receiver experiences the undesired users as interference.

Commonly, a receiver is specified to adequately receiver the desired user even when power levels of the interferers are substantially larger. Typical differences in power levels can be up to 40 dB. To maintain sufficient resolution for the desired user, in the presence of such large interferers, additional bits for the ADC are required. Not only does the ADC have to deal with interferers inside the desired frequency range, but also with left over interferer products outside of the desired frequency range. The DR and over sample frequency of the ADC are a combination of the interference power level with which the receiver needs to cope, both in and outer band, and a trade off with the analog filter drop off [2]. The number of additional headroom bits therefore depends on the standard and the practical implementation. In this paper the focus is on in band interference and ADC resolution.

C. Signal Model

All signals within the considered wide bandwidth are modeled as Orthogonal frequency-division multiplexing (OFDM) signals with a smaller bandwidth. OFDM is a commonly used modulation method in wireless communication due to its ability to cope with severe channel conditions. In OFDM the channel is divided in several narrow band sub carriers, each with their own data symbol. The OFDM signal is constructed as such that all the sub carriers are orthogonal to one another. This allows for simplified signal processing in the digital domain and robustness against frequency selective fading due to multipath. For channels with a large delay spread a sufficiently large guard interval is required to maintain orthogonality. The baseband equivalent of an OFDM signal is expressed as

$$x(t) = \sum_{k=-\frac{K}{2}}^{\frac{K}{2}-1} L_k e^{i2\pi \frac{kB}{K} t}, \tag{2}$$

here t is time, B is the bandwidth of the signal, K is the number of OFDM frequencies and L_k is the data symbol of the k^{th} OFDM frequency.

D. Channel Model

Since all users use an OFDM scheme we can subdivide the wide bandwidth into a set of narrow band signals. The channel for a single OFDM sub carrier can be modeled as narrow band Raleigh fading. However, by modeling each sub carrier separately with a narrow band model the angular information across the sub carriers is not accounted for. To account for the angular information we propose the use of a ray tracer. In this section we will first introduce the Rayleigh Fading model for each subcarrier. Next we will introduce a ray tracer model based on a wide band channel model that accounts for the angular information in the wide band channel.

1) Narrow Band Rayleigh Fading: Due to the use of an OFDM signal each sub carrier can be modeled as a narrow band signal. Consider a transmission system that consists of N_t transmit antennas and N_r receive antennas. If a narrow band complex transmitted signal $\mathbf{s}$ is transmitted, the received signal $\mathbf{r}$ can be expressed as

$$\mathbf{r} = \mathbf{H}\mathbf{s} + \mathbf{n}, \tag{3}$$

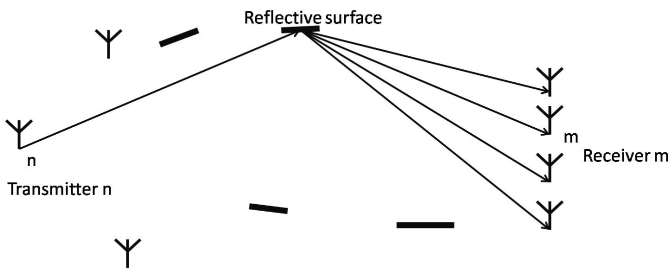


Fig. 2: *Wide band channel model consisting of N_t transmit antennas, each with P reflectors, and an array of N_r receive antennas.*

where **H** is a $N_r \times N_t$ complex channel-gain matrix and **n** is a complex N_r-dimensional additive white Gaussian noise (AWGN) vector. For uncorrelated Rayleigh fading, the entries in **H** are independent and identically distributed (i.i.d.), complex, zero-mean Gaussian with unit magnitude variance.

2) Wide Band Angles of Arrival: Spatial filtering exploits the common angle-of-arrival (AOA) of various multipath components across a wide frequency band. By modeling each channel of an OFDM sub carrier as uncorrelated Rayleigh fading this quality is lost. Therefore, a wide band channel model [7] is proposed based on a ray tracer model [8]. The proposed model is depicted in Figure 1. The model consist of N_t transmit antennas, each with P reflectors and a receiver array consisting of N_r receive antennas. The base band equivalent of the received signal at the m^{th} antenna of the wide band channel model is given by

$$y_m(t) = \sum_{p=1}^{P} a_{n,m,p} e^{-i2\pi f(t-\tau_{n,m,p})} x_n(t) \qquad (4)$$

were $x_n(t)$ is the transmitted signal of the n^{th} transmit antenna, P is the number of multipath components, and $\{a_{n,m,p}\}$ and $\{\tau_{n,m,p}\}$ are the random complex amplitude and random arrival time coefficient sequences of the multipath components between transmit antenna n and receive antenna m. In the proposed ray tracer channel model each transmit antenna has a unique set of reflectors. The transmit antennas and corresponding reflectors are randomly placed in a two dimensional space according to a gaussian distribution with a variance of σ_n^2 and $\sigma_{n,p}^2$ in each dimension, respectively. The center of the receiver array is randomly placed in a two dimensional space according to a gaussian distribution in each dimension with a variance of σ_m^2 and the antennas are spaced in a lineair array according to an inter element distance of half the wavelength of the carrier frequency λ_{car}. The reflection coefficients are defined as

$$a_{n,m,p} = r_{n,p} \left(\frac{c}{f 4\pi (d_{n,m,p})} \right) \qquad (5)$$

and the delays as

$$\tau_{n,m,p} = \frac{d_{n,m,p}}{c} \qquad (6)$$

were $r_{n,p}$ is complex, zero-mean Gaussian with unit magnitude variance, c is the speed of light, f is frequency, $d_{n,m,p}$ is the distance between transmitter n and reflective surface p plus the distance between reflective surface p and receive antenna m. In case the p^{th} multipath is a line of sight (LOS) signal, $d_{n,m,p}$ is the distance between transmitter n and receive antenna m and $r_{n,p} = 1$.

The baseband equivalent of the received OFDM signal is now expressed as

$$y_m(t) = \sum_{k=-\frac{K}{2}}^{\frac{K}{2}-1} \sum_{p=1}^{P} L_k a_{n,m,p} e^{-i2\pi \frac{kB}{K}(t-\tau_{n,m,p})} \qquad (7)$$

were $a_{n,m,p}$ is a function of k and the carrier frequency.

III. SPATIAL FILTERING APPROACH

Now we have defined the signal and channel model we can use them as a basis to calculate the maximum achievable throughput. Since we are focussing on a system in which the ADC is the bottleneck, due to the required resolution, we need to account for the quantization effects in the throughput calculations. By exploiting the common AOA for several multipath components, we strive to increase the effective number of available ADC levels for the desired user. This effect can be quantified via a throughput calculation, were any increase in throughput relates to the increase in number of effective available ADC levels for the desired user.

First we will model the quantization effects, second we will use the quantization model in the throughput calculation, and finally we will introduce two iterative methods to exploit the AOA information.

A. Problem statement

To mitigate the effect of interference power on dynamic range requirements of an ADC we propose to use spatial filtering. We strive to quantize the improvement in dynamic range in terms of the increase in throughput for the desired user.

1) Quantization Noise: The input signal is quantized in the time domain, while the OFDM symbols are extracted after an FFT of the input signal. The received signal at the antenna is now modeled in the frequency domain for all K frequency components as

$$\mathbf{r_k} = \mathbf{U_k H_k x_k} + \mathbf{U_k H_k n_{int,k}} + \mathbf{n_{th,k}} \qquad (8)$$

Here $\mathbf{r_k}$ is the received signal, $\mathbf{x_k}$ is the desired user, $\mathbf{n_{int,k}}$ are the interfering users, $\mathbf{H_k}$ is the current channel state matrix, $\mathbf{U_k}$ is the receiver vector, and $\mathbf{N_{th,k}}$ is the thermal noise at the k^{th} frequency. Due to the central limit theorem we can model the quantization noise in the frequency domain as AWGN [9] [10]. To simplify the mathematics, the AGC sets the variance of the combined input signals to be equal to the variance of the idealized input signal for an idealized ADC. Due to considerations such as large peak to average power ratios and clipping prevention the input signal could have a different scaling. With a unit magnitude scaling the boundary condition for the AGC is

$$\sum_{k=1}^{K} \mathbb{E}\left[\mathbf{r_k}\mathbf{r_k}^*\right] = \rho_{\mathrm{ADC}} \sum_{k=1}^{K} \mathbb{E}\left[\mathbf{n_{q,k}}\mathbf{n_{q,k}}^*\right], \tag{9}$$

were K is the total number of unique OFDM frequencies of $\mathbf{x}$ and $\mathbf{n_{int}}$, $\mathbf{n_{q,k}}$ is the quantization noise at the k^{th} frequency, and $\rho_{\mathrm{ADC}} = 2^{2b_{\mathrm{ADC}}} - 1$ is the SNR of an idealized complex input ADC.

2) Throughput model: By modeling the quantization noise as AWGN we can use the capacity equation to calculate the throughput. Due to the approximation of the quantization effect this throughput should not be interpreted as the Shannon capacity. Lacking a better model we use the capacity equation as a measure of achievable throughput. The approximation of the throughput T is given by

$$\begin{aligned} T = & \\ \max_{\mathbf{K}_x:\mathrm{Tr}[\mathbf{K}_x]\le P} \mathbb{E}_{\mathbf{H}} & \Bigg[\log\det\Bigg(\mathbf{I}_{N_r} + \sum_{k=k_1}^{k_{\mathrm{K}}}(\mathbf{U_k}\mathbf{H_k}\mathbf{K_{x,k}}\mathbf{H_k^*}\mathbf{U_k^*}) \\ & \Bigg[\sum_{k=k_1}^{k_{\mathrm{K}}}(N_{th,k}\mathbf{U_k}\mathbf{U_k^*}\mathbf{I}_{N_r} + N_{q,k}\mathbf{I}_{N_r})\Bigg]^{-1}\Bigg)\Bigg] \end{aligned} \tag{10}$$

here $k_1 \ldots k_{\mathrm{K}}$ are the frequencies components of the OFDM symbols of $\mathbf{x}$, $\mathbf{K_{x,k}} = \mathbb{E}[\mathbf{x_k}\mathbf{x_k^*}] = P/K\mathbf{I}_K$ is the expected transmit power matrix, $N_{th,k} = \mathbb{E}[\mathbf{n_{th,k}}\mathbf{n_{th,k}^*}]$ is the expected thermal noise,

$$\begin{aligned} N_{q,k} = \frac{1}{\rho_{ADC}}\frac{1}{K}\sum_{k=1}^{K}(&\mathbf{U_k}\mathbf{H_k}\mathbf{K_{x,k}}\mathbf{H_k^*}\mathbf{U_k^*} \\ &+ \mathbf{U_k}\mathbf{H_k}\mathbf{K_{int,k}}\mathbf{H_k^*}\mathbf{U_k^*} + N_{th,k}\mathbf{U_k}\mathbf{U_k^*}\mathbf{I}_{N_r}) \end{aligned} \tag{11}$$

is the expected quantization noise, and $\mathbf{K_{int,k}} = \mathbb{E}[\mathbf{n_{int,k}}\mathbf{n_{int,k}^*}]$ is the expected interference power of the k^{th} sub carrier. The throughput is defined for complex input signals. Finding an C-OFDM error correction encoding strategy that is suitable for the fading throughput per subcarrier is out of scope of this paper. An ADC can only sample a real input signal. In a zero force architecture this is solved by using two ADCs, of which one receives a 90 degrees phase shifted version of the input signal compared to the other. Thus, complex analog to digital conversion is achieved by splitting the RF signal into an orthogonal I and Q path, each with its own ADC (Figure 1). Another commonly used method is to use an adequately large intermediate frequency (IF) receiver and use a single ADC which samples at, at least, twice the wide bandwidth of the input signals.

B. Proposed Algorithms

We have now defined the achievable throughput for a given receive vector $\mathbf{U_k}$. We can use this throughput to quantify the potential gain of spatial filtering in terms of bits/s/Hz, and use it to indicate the potential bit reduction, and thus via (1) the potential power reduction of the ADC. Next we will propose two algorithms which strive to exploit the phase information in the narrow band channel and the AOA information of the wide band channel.

1) Narrow Band: Two scenarios are considered for the narrow band case; each user has a single unique sub carrier and all N_t users have the same sub carrier. Further we assume that $N_r \geq N_t$. The $N_r \times N_t$ size channel matrix $\mathbf{M}$ is stacked by current channel state $\mathbf{H}$ vectors, for each transmitter, of size $N_r \times 1$ according to $\mathbf{M} = [\mathbf{H_1}, \mathbf{H_2}, \ldots, \mathbf{H_{N_t}}]$. Via a singular value decomposition (SVD) the Eigen values are derived. The SVD is given by $\mathbf{M} = \mathbf{U_M}\mathbf{\Lambda_M}\mathbf{V_M^*}$, here $\mathbf{U_M}$ and $\mathbf{V_M^*}$ are unitary matrices, and $\mathbf{\Lambda_M} = \mathrm{diag}(\lambda_1, \lambda_2, \ldots, \lambda_{N_r})$, were λ_k^2 is the k^{th} Eigen values of $\mathbf{MM}^*$. From the complex transpose of $\mathbf{U_M}$, $\mathbf{U_M^*}$, the $N_r \times 1$ vector is chosen for vector $\mathbf{U_k}$ which corresponds to the Eigen value of the desired user. The receiver array is now beamforming into the direction of the desired user, while nulling the other users.

2) Wide Band: For the wide band case it is assumed all OFDM sub-carriers of $\mathbf{x}$ and $\mathbf{n_{int}}$ each have a unique frequency. The aim is to find an appropriate setting for a common vector $\mathbf{U_k}$ across all frequency components such that the throughput T is maximized. In order to achieve this we propose two different estimation algorithms.

The first algorithm (WB Method 1) is a least squares algorithm which tries to maximize the power of the desired user and mitigate the power of the interfering users. The algorithm iteratively solves the equality

$$|\mathbf{AU} - \mathbf{b}|_2^2 = 0 \tag{12}$$

here $\mathbf{A}$ is a $K \times N_r$ size matrix containing the channel components of $1 \times N_r$ size channel vector $\mathbf{H_k}$ of each k^{th} frequency, $\mathbf{b}$ is a $K \times 1$ vector containing zeros on the frequency components of $\mathbf{n_{int}}$ and ones on the frequency components of $\mathbf{x}$, $\mathbf{U}$ is a complex $N_r \times 1$ size vector and is used as a common $\mathbf{U_k}$ in (10).

The second algorithm (WB Method 2) is a least squares algorithm which first estimates a $\mathbf{U}$ for each user via (12), were K is now the number of frequency components of the considered user and vector $\mathbf{b}$ is a vector containing only ones. The separate U's are stacked in a matrix $\mathbf{M} = [\mathbf{U_1}, \mathbf{U_2}, \ldots, \mathbf{U_{N_t}}]$. As in the narrow band case, the Eigen values are derived via a singular value decomposition (SVD). The SVD is given by $\mathbf{M} = \mathbf{U_M}\mathbf{\Lambda_M}\mathbf{V_M^*}$, here $\mathbf{U_M}$ and $\mathbf{V_M^*}$ are unitary matrices, and $\mathbf{\Lambda_M} = \mathrm{diag}(\lambda_1, \lambda_2, \ldots, \lambda_{N_r})$, were λ_k^2 is the k^{th} Eigen values of $\mathbf{MM}^*$. From the complex transpose of $\mathbf{U_M}$, $\mathbf{U_M^*}$, the $N_r \times 1$ vector is chosen for vector $\mathbf{U_k}$ which corresponds to the Eigen value of the desired user. The receiver array is now beamforming into the direction of the desired user, while nulling the other users.

IV. NUMERICAL RESULTS

For simulations we will start with a narrow band example, using the commonly used Rayleigh fading model. Secondly we will use the wide band channel model to show the potential

TABLE I: *SNR versus number of bits required to achieve a given throughput $T = 2$bits/s/Hz.*

SNR [dB]	**Analog Array [bits]**	**SISO [bits]**	**Difference [bits]**
10	6.8	11.0	4.2
20	3.1	7.2	4.1
30	2.7	6.8	4.1
40	2.6	7.0	4.4

benefit of spatial filtering in a wide band regime and to quantify the potential gain in terms of ADC power reduction and ADC bits. In simulations the SNR at the receiver is defined as the sum of the received power by all receive antennas of the desired user, divided by the power of the corresponding channel noise

$$\text{SNR} = \frac{\sum_{k=k_1}^{k_K} \text{Tr}[\mathbf{K}_{\mathbf{x},\mathbf{k}}]}{\sum_{k=k_1}^{k_K} N_{th,k}}, \tag{13}$$

here $k_1, \ldots, k_K$ are the frequency components of the OFDM symbols of $\mathbf{x}$, were the channel noise components are defined as $N_{th,k} = \frac{1}{k_K - k_1} N_{th}$, with $N_{th} = k_b \text{T} B$, here T is the temperature, B is the bandwidth of the user, and Boltzmann's constant $k_b = 1.38 \cdot 10^{-23}$. Further, the interference to carrier ratio (ICR) of the i^{th} individual interferer is defined as

$$\text{ICR} = \frac{\sum_{k=k_{i,1}}^{k_{i,K}} \text{Tr}[\mathbf{K}_{\mathbf{int},\mathbf{k},\mathbf{i}}]}{\sum_{k=k_1}^{K_K} \text{Tr}[\mathbf{K}_{\mathbf{x},\mathbf{k}}]}, \tag{14}$$

here $k_{i,1}, \ldots, k_{i,K}$ are the frequency components of the OFDM symbols of $\mathbf{n}_{\mathbf{int},\mathbf{k},\mathbf{i}}$, were $\mathbf{K}_{\mathbf{int},\mathbf{k},\mathbf{i}} = \mathbb{E}[\mathbf{n}_{\mathbf{int},\mathbf{k},\mathbf{i}} \mathbf{n}^*_{\mathbf{int},\mathbf{k},\mathbf{i}}]$, and $\mathbf{n}_{\mathbf{int},\mathbf{k}} = \sum_{i=1}^{N_t - 1} \mathbf{n}_{\mathbf{int},\mathbf{k},\mathbf{i}}$.

A. Narrow Band Channel

We start with an example of the in-channel interference case. We assume $N_t = 3$, $N_r = 3$, the ADCs have 12 bits, and the ICR=40dB for interferer 1 and ICR=30dB for interferer 2. Since our receiver has two ADCs, one in the I-path and one in the Q-path, the combined resolution is 24 bits per sample. Figure 3a and 3b depict the results averaged over 100 Monte Carlo simulations. As can be seen in the graphs, beamforming into the Eigen value corresponding to the desired user, yields a result which is close to the non interference case. Figure 3b depicts the results of the co-channel interferer simulations. Here the interferers are on adjacent OFDM sub carriers. Therefore, the ADC is sampling at three times the rate as in the in-channel case. The desired user can benefit from the over sampling ratio and achieve a higher throughput. The throughput for the desired user can be above 24 bits/s/Hz, because the quantization noise is spread over more sub carriers than are used by the desired user.

B. Wide Band Channel

Now we will give several examples of the wide band channel. First we will present results of the proposed algorithms for various SNR, secondly the ICR is varied, and finally the number of bits of the individual ADCs is swept.

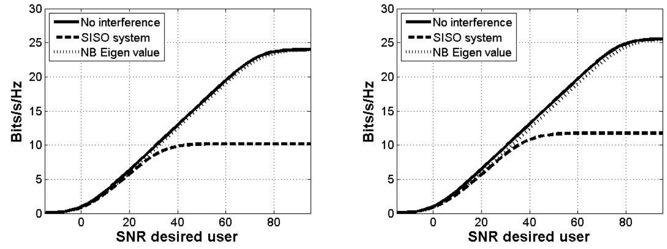
1) Impact of SNR: We assume $N_t = 3$, $N_r = 3$, the ADC has 8 bits, and the ICR=40dB for interferer 1, ICR=30dB for interferer 2. Each user has 20 OFDM symbols, of which the 2 on either side of the spectrum contain zeros and the 16 OFDM sub carriers in the middle contain OFDM data symbols, the carrier frequency is 2.45GHz, the bandwidth $B = 5$MHz per user, $\sigma_n = 10$m, $\sigma_{n,p} = 5$m, $\sigma_m = 10$m, $c = 3 \cdot 10^8$m/s, T = 295K, and the number of multipaths $P = 5$, including a LOS. Figure 4a depicts the results for an 8 bit, and Figure 4b for a 12 bit ADC per I and Q path for 100 Monte Carlo simulations. As can be seen in Figure 4, WB method 2 outperforms WB method 1, and vastly improves the throughput at higher SNR, when compared to the SISO case, up to 8 bits/s/Hz. At lower SNR WB method 1 outperforms WB method 2 and the SISO case, and is close to the interference free scenario, improving the throughput with up to 2 bits/s/Hz. WB method 1 performs well when thermal noise is dominant, because it mainly beam forms in the direction of the desired user, which increases the SNR and thus throughput. WB method 2 is a good strategy in the quantization noise limited regime, because it beamforms in the direction of the desired user while attempting to null the interferers. Decreasing the power of the interferers in the quantization noise limited regime, results in more ADC resolution for the desired user, and thus a higher throughput.

2) Impact of ICR: Compared to the previous section we will now vary the ICR and assume both interferers have an equal ICR. Further, we assume the SNR= 40dB, and the ADCs each have 8 bits. Figure 5a shows the results for 100 Monte Carlo simulations. In Figure 5a it can be seen that relatively small interference can be prohibitive even at large SNR. Furthermore, analog beamforming has the potential to improve the throughput both in the presence of small and large interferers, when compared to a SISO system.

3) Impact of Number of ADC Bits: Compared to the previous section we will now vary the number of ADC bits and assume both the I and Q ADC have an equal number of bits. Further, we assume the ICR= 40dB for interferer 1 and 2. Figure 5b shows that the throughput gain compared to a SISO system is largest in a quantization noise limited scenario. From Figure 5b we can derive that at a Throughput of 2 bits/s/Hz, an analog spatial filtering system can equal a SISO system with 4 bits per ADC less (Table I). We believe this result is significant since this indicates that the power savings for the ADC can potentially be an order of magnitude. Since the power of ADC converters reduces with an order of magnitude every decade, analog spatial filtering has the potential to accelerate the reduction of overall system power reduction.

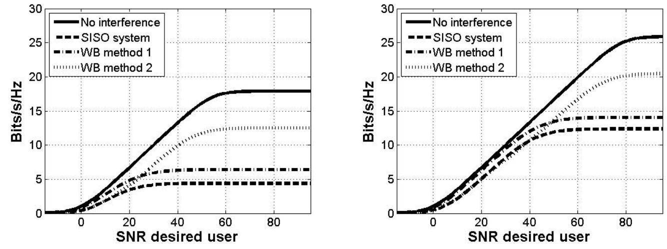
V. CONCLUSIONS

Analog spatial filtering can achieve similar system throughput with significantly less quantization bits. Simulations show that for realistic scenarios a receiver using analog spatial filtering can achieve similar throughput with 4 ADC bits less, corresponding to a power reduction of over 90%. Since ADC power consumption reduces at a rate of an order of magnitude



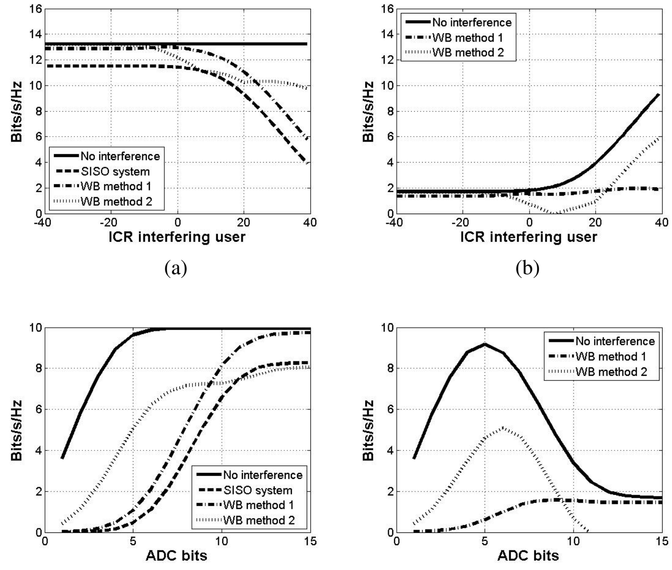
(a) (b)

Fig. 3: *Throughput compared to a SISO system. For two 12 bit ADCs in the I and Q path. Narrow band channel model. ICR of interferer one is 40 dB and of interferer two is 30 dB. For in channel (a), and co channel (b) interference.*



(a) (b)

Fig. 4: *Throughput compared to a SISO system. For two 8 bit (a), and two 12 bit (b) ADCs in the I and Q path. Wide band channel model. ICR of interferer one is 40 dB and of interferer two is 30 dB.*



(c) (d)

Fig. 5: *a & c. Throughput, b & d. difference in Throughput, compared to a SISO system. For two 8 bit ADC in the I and Q path and SNR=* $40dB$ *(a & b). For two interferers with ICR=* $40dB$ *and SNR=* $30dB$ *(c & d). Wide band channel model.*

per decade, we believe analog spatial filtering can contribute to overal system power reduction, and accelerate the introduction of CR in the hand held market.

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