Multi-Sensor Frequency Domain Multiple Access Interference Canceller for DS-CDMA Systems

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***Abstract*— Direct Sequence Code Division Multiple Access (DS-CDMA) signals exhibit cyclostationary properties which imply a redundancy between frequency components separated by multiples of the symbol rate. In this paper a Multiple Access Interference Canceller (Frequency Shift Canceller) that explores this property is presented. This linear frequency domain canceller operates on the spreaded signal in such way that the interference and noise at its output is minimized (Minimum Mean Squared Error Criterium). The Frequency Shift Canceller (FSC) performance was evaluated for a UMTS-TDD scenario and multisensor configurations, where the cases of diversity and beamforming were considered. All these configurations are evaluated concatenated with a parallel interference canceller (PIC-2D). The results are benchmarked against the performance of the conventional RAKE-2D detector, the conventional PIC-2D detector and single user scenario, and we observe considerable performance gains with the FSC specially for the diversity case and a performance close to the single user case when it was evaluated jointly with PIC-2D.**

***Index Terms*— cyclostationary, redundant, frequency shift, PIC, DS-CDMA, Beamforming, Spatial Diversity.**

I. INTRODUCTION

Direct sequence spread-spectrum (DS-SS) code division multiple access (CDMA) has emerged as one of the most promising techniques to implement various radio communication systems. It presents significant advantages over Time Division Multiple Access (TDMA), namely frequency diversity, multipath diversity and more spectrum efficiency on multicell systems [1], which led to its choice as the technology for third generation cellular systems. The first version of third generation CDMA systems will be based on the conventional RAKE receiver, which is known to be limited by the multiple access interference (MAI) and require a very precise power control. To overcome these limitations and therefore enhance the capacity of CDMA systems, joint detection of the received DS-SS signals has been proposed to be used at the base station (BS) or at the user equipment. The optimum joint detector [2] although giving the best performance requires however a prohibitively high computational complexity, and consequently effort has been made to devise suboptimum algorithms with good compromise between performance and complexity that can be implemented without prohibitive costs in near future CDMA systems. This communication fits in this approach, and aims at presenting a moderate complexity MAI canceller operating on the broadband DS signal.

The DS-SS signal is a particular case of a stationary random pulse amplitude modulation. This kind of signals are known to have cyclostationary properties [3]. Those properties imply redundancy between frequency components separated by multiples of the symbol rate. It is this characteristic that is explored to propose a new MAI canceller.

The paper is outlined as follows. In section II it is shown that in a DS signal, non overlapping frequency bands separated by a multiple of the baud rate are linearly related. This result is used to present in section III the architecture and design principles of a MAI canceller that explores this redundancy. In section IV, for an UMTS-TDD scenario, simulation results for several configurations employing the FSC considering both the beamforming and diversity cases are presented. Finally in section V the main conclusions of this work are outlined.

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II. THEORETICAL BACKGROUND

A DS-SS signal is represented as

$$s(t) = \sum_k a_k g(t - kT) \tag{1}$$

where $\{a_k\}$ is the sequence of information symbols, $\frac{1}{T}$ the symbol rate and $g(t)$ is the signature waveform.

The Fourier Transform of this signal is

$$S(f) = \sum_k a_k G(f) e^{-j2\pi fkT} = G(f)A(f) \tag{2}$$

with

$$A(f) = \sum_k a_k e^{-j2\pi fkT} \tag{3}$$

From (3) it is easy to verify that

$$A(f + \tfrac{i}{T}) = A(f) \ \forall i \in \mathbb{Z} \tag{4}$$

Let us define

$$S_{mB}(f) = S\left(f + \frac{m}{T}\right) rect\left(fT\right) \tag{5}$$

and

$$G_{mB}(f) = G\left(f + \frac{m}{T}\right) rect\left(fT\right) \tag{6}$$

where m is the index of the band. From now on the subscript $_B$ means a signal frequency shifted to baseband. Then using (2)

$$S_{mB}(f) = G_{mB}(f)A(f) \tag{7}$$

Let us consider two values m_1 and m_2 for the band index

$$S_{m_1B}(f) = G_{m_1B}(f)A(f) \tag{8}$$

and

$$S_{m_2B}(f) = G_{m_2B}(f)A(f) \tag{9}$$

Therefore assuming $G_{m_1B}(f)$ has no singularities, it can be

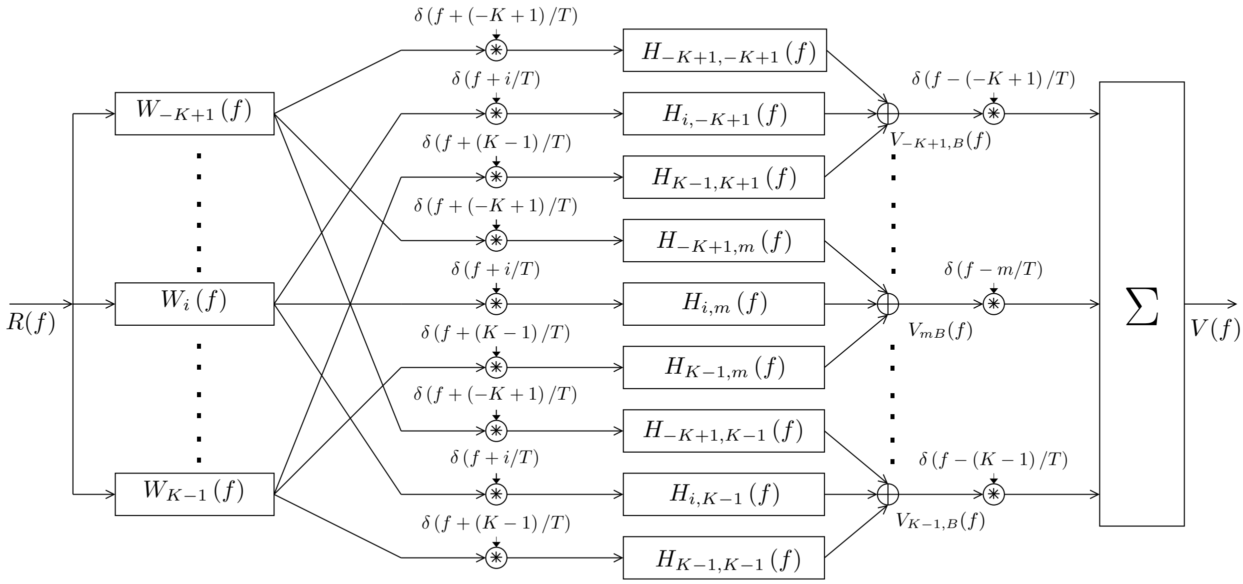

Fig. 1. Conceptual Schematic of the Canceller

concluded that in the interval $f \in \left[-\frac{1}{2T}, \frac{1}{2T}\right[$ we have

$$S_{m_2B}(f) = \frac{G_{m_2B}(f)}{G_{m_1B}(f)} S_{m_1B}(f) \tag{10}$$

This means that the signal information in non-overlapping frequency bands spaced by a multiple of the baud rate are related by a linear transformation.

III. Principles of the Canceller

The canceller operates in the frequency domain and in a practical implementation or in a simulation system, the time to frequency domain conversion is performed digitally through a Fast Fourier Transform (FFT). However in the following derivations we shall use a continuous signal representation.

The architecture of the canceller is shown in figure 1, for a given user. In a base station where all the signals have to be recovered, the canceller consists of the replica of this basic receiver for each user. This canceller could be implemented in the mobile station provided the codes of active users are detected.

The input signal in the frequency domain $R(f)$ in figure 1 is defined as $\sum_{u=1}^{U} S^{(u)}(f) + N(f)$ where U is the number of users and $N(f)$ is stationary noise in the frequency domain with power spectral density $\eta_{in}(f)$.The signal $S^{(u)}(f)$ is the Fourier transform of $s^{(u)}(t)$ given by (1) where the superscript $^{(u)}$ refers to the user. The blocks $W_i(f)$ refer to the transfer function of a rectangular filter with bandwidth equal to the symbol rate and centered at $\frac{i}{T}$.

We shall start by considering the case where all the users have the same SF and then proceed with the generalization for multirate.

Let us consider without loss of generality that user one is the user of interest. The objective and design criteria for the canceller is to minimize the overall disturbance (MAI+noise) subject to the condition that $S^{(1)}(f)$ is not distorted. This constraint implies that the filters in figure 1 which convert the band i to band m (see section II) be of the form

$$H_{i,m}(f) = \alpha_{i,m} \frac{G_{mB}^{(1)}(f)}{G_{iB}^{(1)}(f)} \tag{11}$$

where $\alpha_{i,m}$ is a scalar weight factor, $G_{mB}^{(1)}(f)$ is defined in (6) and superscript $^{(1)}$ refers to the user of interest.

In those conditions the baseband shifts of the output signal $V(f)$ are given by

$$\begin{aligned} V_{mB}(f) &= S_{mB}^{(1)}(f)\left(\sum_i \alpha_{i,m}\right) \\ &+ \sum_{u=2}^{U}\left[\sum_k a_k^{(u)} e^{-j2\pi fkT} \beta_{mB}^{(u)}(f)\right] + N'_{mB}(f) \end{aligned} \tag{12}$$

where $S_{mB}^{(1)}(f)$ is defined in (5) and

$$\beta_{mB}^{(u)}(f) = \sum_i \alpha_{i,m} \frac{G_{mB}^{(1)}(f)}{G_{iB}^{(1)}(f)}(f) G_{iB}^{(u)}(f) \tag{13}$$

From (12) we conclude that the condition that the user of interest is not distorted is verified if $\sum_i \alpha_{i,m} = 1$. The power spectral density of additive the noise disturbance $N'_{mB}(f)$ is given by

$$\eta_{out_{mB}}(f) = \sum_i |\alpha_{i,m}|^2 \left|\frac{G_{mB}^{(1)}(f)}{G_{iB}^{(1)}(f)}\right|^2 \eta_{in_{iB}}(f) \tag{14}$$

The design criteria has the objective of minimizing the Mean Squared Error (MSE), i.e. the weights $\alpha_{i,m}$ are dimensioned so that

$$C\sum_{u=2}^{U}\left[\int_f \left|\beta_{mB}^{(u)}(f)\right|^2 df\right] + T_{Bt}\int_f \eta_{out_{mB}}(f) df \tag{15}$$

is minimized subject to the condition that $\sum_i \alpha_{i,m} = 1$. In (15) T_{Bt} is the burst duration time and C correspond to the number of symbols existing in one burst. The minimization of (15) leads to a linear system of equations involving the variables $\alpha_{i,m}$. The number of degrees of freedom is defined as number of independent parameters $\alpha_{i,m}$ available to minimize the MSE. The matrix to be inverted is Hermitian semi-definite positive of dimension equal to the degrees of freedom minus one and the method used for inverting is the Cholesky decomposition. It can be proved that the minimization of the MSE and so the calculation of $\alpha_{i,m}$ only needs to be performed for one output band (single value of m) because in the other bands $\alpha_{i,m}$ are the same ($\alpha_{i,m}$ do not really depend on m).

Each band of the output signal $V(f)$ depends linearly of the input bands with symbol rate bandwidth ($\frac{1}{T}$) and spaced by multiples $\frac{1}{T}$. The performance of the interference canceller could be enhanced if the input and output bands are divided in subbands of equal bandwidth. Then each output subband depends linearly on the input subbands with the same bandwidth and spaced by multiples of $\frac{1}{T}$ of the output subband. The shorter the subbands better the performance.

Let us consider now the case of users with different SF's, which is of interest for UMTS-TDD. This standard was designed to accommodate multiple symbol rates which can be done by using different SF's. In the standard the spreading code is defined as the product between the channelization and the scrambling code. The channelization code lasts for one symbol and its number of chips is equal to the SF. The scrambling code lasts for Q_{max} chips or during $\frac{Q_{max}}{Q}$ symbols with Q being the SF. Then to construct the spreading code, the channelization code must be repeated $\frac{Q_{max}}{Q}$ and multiplied by the scrambling code. The spreading code extends for more than one symbol when $Q \neq Q_{max}$.

The approach to extend the canceller explained above from single rate canceller to multirate is to consider that at the input of the canceller there is a decomposition of each signal with a SF different of sixteen in several signals with SF of sixteen.

The representation in the time domain of an information sequence spreaded using a SF of Q is then

$$s(t) = \sum_{l=0}^{\frac{Q_{\max}}{Q}-1} \sum_{k} a_k^l g_l \left(t - lT - \tfrac{Q_{\max}}{Q} kT\right) \qquad (16)$$

where $\{a_k^l\}$ are the sequences of information symbols, $\frac{1}{T}$ the symbol rate (depends on the SF) and $g_l(t)$ are the components of signature waveform. To obtain these components the spreading code (length equal to Q_{max}) is divided in $\frac{Q_{max}}{Q}$ sequences of length Q and then each one is pulse shaped (Raised Cosine) and affected by channel estimates.

This signal is then divided in $\frac{Q_{max}}{Q}$ signals

$$s_l(t) = \sum_{k} a_k^l g_l \left(t - lT - \tfrac{Q_{\max}}{Q} kT\right) \qquad (17)$$

with $l \in \{0, 1..., Q_{\max}/Q - 1\}$ having each a signature waveform $g_l(t - lT)$. Figure 2 illustrates this decomposition for the case $\frac{Q_{max}}{Q} = 2$, assuming rectangular shaping.

The cancellation process is repeated to recover each decomposed signal of SF of sixteen. The signature waveforms used in the cancelling process are the ones of the decomposed signals.

IV. Application of the Canceller to UMTS-TDD

In this section some numerical results are presented illustrating the performance of the proposed detector configurations with UMTS-TDD signals. To evaluate the canceller performance a simulation chain was implemented. Basically this simulation chain is composed by a transmitter, a transmission channel and a receiver.

A. Transmitters

The transmitters are compliant with the 3GPP specifications for UMTS-TDD.

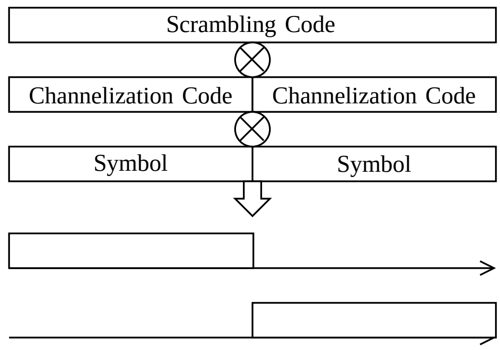

Fig. 2. Decomposition of the signal seen at the transmitter

B. Channel Model

The channel model used in this work was the Geometrical Based Single Bounce Elliptical Model (GBSBEM) proposed by Liberti [4]. This model was developed for microcell and picocell environments. The propagation channel is characterized by one line of sight (LOS) tap and $L-1$ taps for each user arriving from remote reflectors located randomly within an ellipsis where the base station and the mobile unit are at the foci. Each tap (from each user and each burst) is characterized by a complex amplitude and a delay. The delay of each tap including the LOS tap is a random variable characterized by a probability density function whose expression can be found in [4]. After evaluating the delays of the taps, all LOS taps delays of all users are synchronized and the others shifted in according. The phase of the tap is uniformly distributed in $[0, 2\pi[$. The amplitude of the tap is obtained from a constant, dependent of the distance followed by the tap (it is proportional to $1/d_i^{pl}$ where d_i is the distance and pl is the pathloss) and normalized (such that the sum of square of that constant of all taps of that user is equal to one; in case of spatial diversity is equal to the number of antennas instead of one) times a random variable with Rayleigh distribution. The complex amplitude (amplitude and phase of the tap) is Doppler filtered.

In this model the angle of arrival (AOA) of the LOS tap of each user is fixed for one simulation and the angles of others taps are random variables with mean equal to AOA of LOS tap and a distribution given [4]. The channel parameters are assumed to be constant within each burst.

C. Receivers

Fig. 3 and 4 depict two configurations for the detector that include the Frequency Shift Canceller, one for beamforming and one for spatial diversity (*Maximum Ratio+Delay-Combining+FSC*).

TABLE I
Simulation parameters settings

Number of Taps (per user and antenna)	2
Velocity	50 Km/h
Path Loss	3.7
Maximum Delay Spread	2.0 μs
Degrees of Freedom of FSC	18
Number of samples per chip	4
Line of Sight Distance	300m

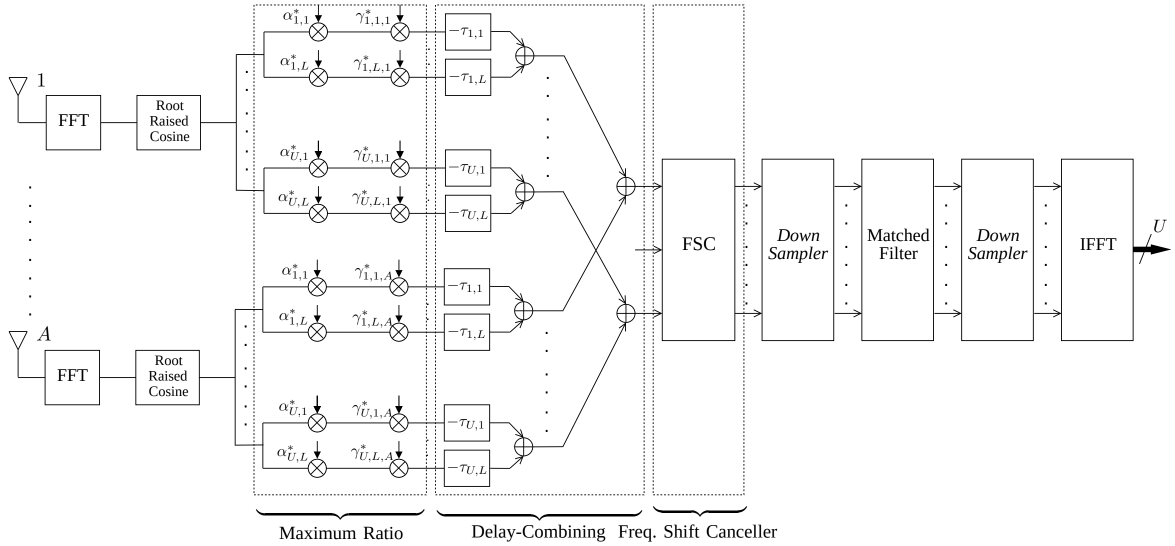

Fig. 3. Detector with FSC. Configuration *Maximum Ratio+Delay-Combining+FSC* with multiple antennas and beamforming. All the operations made between the FFT and IFFT blocks are frequency domain operations despite the fact that some blocks reflect the correspondent time domain operations (delay, downsampling).

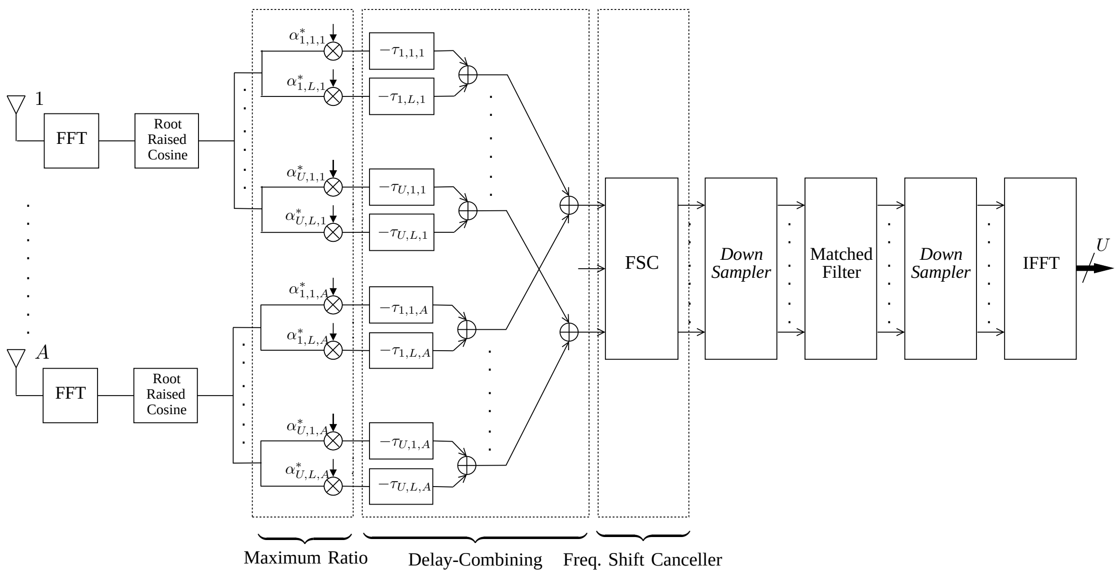

Fig. 4. Detector with FSC. Configuration *Maximum Ratio+Delay-Combining+FSC* with multiple antennas and spatial diversity. All the operations made between the FFT and IFFT blocks are frequency domain operations despite the fact that some blocks reflect the correspondent time domain operations (delay, downsampling).

The other configurations evaluated are the configurations: *Maximum Ratio+FSC+Delay-Combining* and *FSC+Maximum Ratio+Delay-Combining* (each one obtained from the ones in Fig. 3 and 4, interchanging the macroblocks *Maximum Ratio*, *Delay-Combining* and *FSC*). These last two configurations have the same performance because *Maximum Ratio* and *FSC* are linear blocks and their order can be interchanged without functional changes in the behavior. However in terms of complexity for implementation the solution *FSC+Maximum Ratio+Delay-Combining* is usually preferable because it is the number of taps per user and antenna less complex than *Maximum Ratio+FSC+Delay-Combining*.

In the configurations with beamforming, the receptor sensor is a circular array of four antennas with 0.45λ spacing between elements. In figure 3 the coefficients γ^* depend on the array configuration and the direction of arrival of the correspondent user and tap and they are such that for each user and tap the (reception) maximum beam pattern is in the direction of the direction of arrival of the tap [5].

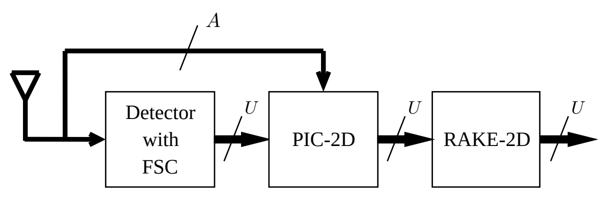

Fig. 5. Receiver including FSC plus PIC-2D. In the text of this article when it is referred to PIC-2D it means PIC-2D+RAKE-2D. U=Number of Users, A=Number of Antennas.

In the configurations with space diversity, the antennas (four) are spaced enough such that the channels are uncorrelated with each other. The average power of each user, across all antennas

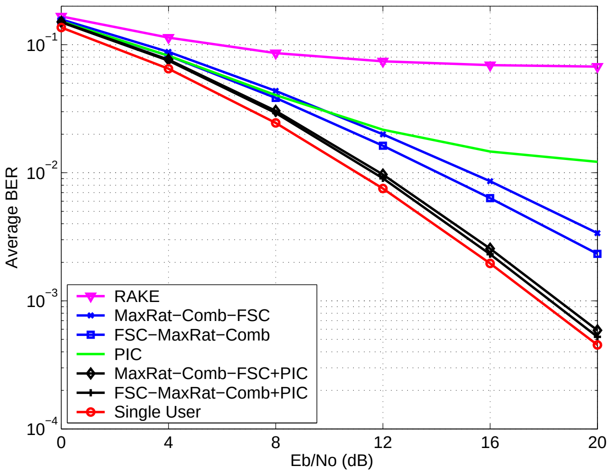

Fig. 6. Single Antenna

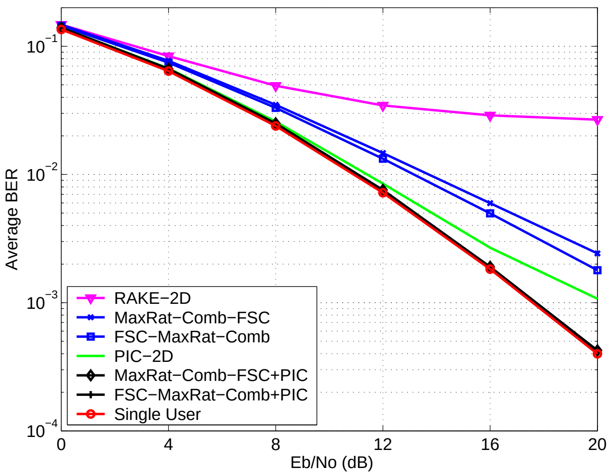

Fig. 7. Beamforming, Circular Array, four antennas (A).

is equal.

In figure 3 and 4 the first downsampling has a factor of the number of samples per chip. The second downsampling takes a factor equal to the maximum SF (in the case of UMTS-TDD, $Q_{max} = 16$). In the frequency domain this operation corresponds to a block frequency averaging operation.

The other detector configurations to be evaluated are the detectors with FSC concatenated with a single stage PIC-2D (fig. 5). The configurations of reference are the conventional RAKE-2D and conventional single stage PIC-2D.

D. Results

The simulations were made with the parameters shown in Table I.

The results are presented for the following scenario.

- Eight users: Four with SF of 16, two with SF of eight and two with SF of 4 (Full load system).
- The users with SF of 8 and 4 have powers above the users of SF of 16 of 3 and 6dB respectively ($\frac{E_b}{N_0}$ equal to all users).

The results presented are only for the users of SF of 16.

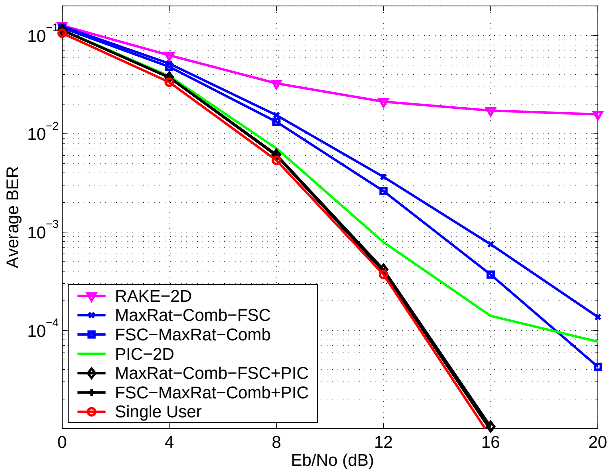

Fig. 8. Spatial Diversity, four antennas (A).

Fig. 6 shows the results with single antenna and Fig. 7 and 8 for the case of multiple antennas considering beamforming and spatial diversity respectively. For the curves of Fig. 7 and 8 the results are presented considering the normalized $\frac{E_b}{N_o}$ i.e removing the antenna gain $10log_{10}(A)$ so that the results can be comparable directly with the ones obtained from single antenna.

The configurations with FSC, if we do not take in account the array gain, there is a gain of 1.5 dB from the single antenna configuration to beamforming configuration and 6 dB to the spatial diversity configuration for a BER of 10^{-2}. For all the configurations with FSC concatenate with PIC-2D is nearly achieved single user performance. For the users with SF's lower than Q_{max} the numerical results have shown identical behavior and also a performance very close to the single user case when using the cascade FSC+PIC-2D.

V. Conclusions

In this communication a new linear multirate canceller operating in the frequency domain that takes advantage of frequency redundancy of the DS-SS signals was presented. The results show that a performance very close to the single user bound is achieved either considering the single or multiple antenna case, when the receiver configuration includes the concatenation of a FSC and PIC-2D. For the channels considered the performance gain (removing the array factor gain) obtained with beamforming is moderate relatively to the case of a single antenna, but significantly improves when the receiving antennas are spaced enough for the channels to be considered uncorrelated.

References

[1] R. Kohno, R. Meidan, and L. B. Milstein, “Spread Spectrum Access Methods for Wireless Communications,” *IEEE Communications Magazine*, January 1995.

[2] S. Verdú, “Minimum Probability of Error for Asynchronous Gaussian Multiple-Access Channels,” *IEEE Transactions of Information Theory*, January 1986.

[3] W. A. Gardner, *Cyclostationarity in Communications and Signal Processing*. IEEE PRESS, 1994.

[4] J. C. Liberti and T. S. Rappaport, *Smart Antennas for Wireless Communications: IS-95 and Third Generation CDMA Applications*. Prentice Hall, 1999.

[5] J. Litva and T. K.-Y. Lo, *Digital Beamforming in Wireless Communications*. Artech House Publishers, First ed., 1996.