MULTIUSER FREQUENCY DOMAIN EQUALIZER FOR MULTIRATE DS-CDMA SYSTEMS

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***Abstract*— Direct Sequence Spread Spectrum (DS-SS) signals exhibit cyclostationary properties which imply a redundancy between frequency components separated by multiples of the symbol rate. In this paper it is presented a Multiple Access Interference Canceller that explores this property and applies to UMTS-TDD. This linear frequency domain canceller operates in 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) is implemented jointly with a RAKE where the multipath combining is made prior the FSC (POST-FSC) or after (PRE-FSC). The performance is evaluated for POST-FSC and PRE-FSC in standalone and each one concatenated with a PIC. The performance is also evaluated for imperfect estimates of the channel. The results are benchmarked against the performance of the conventional RAKE detector and the conventional PIC detector.**

***Index Terms*— cyclostationary, redundant , frequency shift, PIC, DS-CDMA**

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 well known 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 Frequency Shift Detector can be implemented jointly with a RAKE where the multipath combining could be made prior the FSC or after. This two configurations could be used either as standalone unit or it can be used prior to a PIC where it is intended to produce signals clean enough so that the first decisions of the PIC can be considered reliable enough to be used by the subsequent stages.

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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 it is explored to propose a new MAI canceller.

In previous work [4] the POST-FSC was defined and implemented for a full loaded multirate scenario. The simulation results have shown that a performance close to the limiting case of a single user is achieved for all spreading factors (POST-FSC concatenated with a PIC). It was also presented a complexity evaluation against the time domain MMSE.

In the present paper the work is extended to the performance evaluation of PRE-FSC and the sensitivity of both configurations to imperfect channel estimates.

The paper is outlined as follows. In section two 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 three the architecture and design principles of a MAI canceller that explores this redundancy. In section four it is presented simulation results that illustrate the performance provided by the new canceller, in the several configurations and with imperfect channel estimates. Finally in section five the main conclusions of this work are outlined.

II. Theoretical Background

A DS-SS signal with spreading factor Q_{max} (maximum spreading factor allowed in the system) is represented by

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

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

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

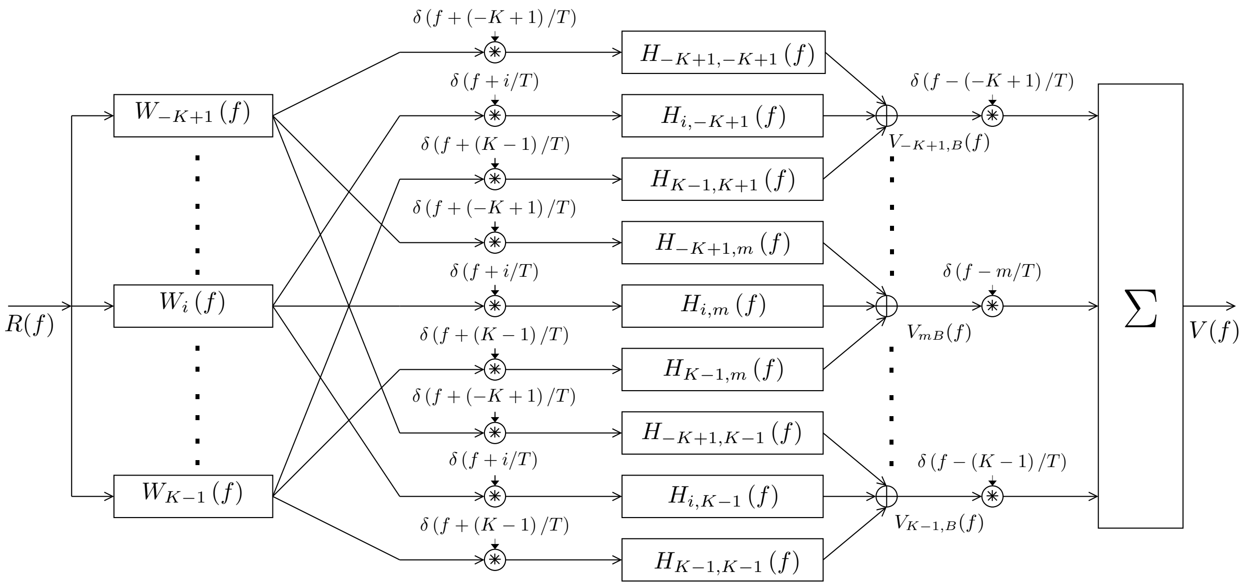

Fig. 1. Conceptual Schematic of the Canceller

From (3) it is easy to verify that

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

Let us define

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

and

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

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

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

Let $m = m_1$ and $m = m_2$ (representing two bands)

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

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

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

This means that it can related the signal information in non-overlapping frequency bands spaced by a multiple of the baud rate 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). In the following derivation it is adopted however a continuous signal notation.

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.

The input signal $R(f)$ 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 $\eta_{in}(f) = E[N(f)N^*(f)]$. 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 filters $W_i(f)$ are rectangular filters of symbol rate bandwidth and centered at $\frac{i}{T}$.

The canceller is first defined to the maximum spreading factor and next then is extended to multirate.

It is considered 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

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

where $G^{(1)}_{mB}(f)$ is defined in (6). The superscript $^{(1)}$ refers to the user of interest. The $\alpha_{i,m}$ are weight factors.

In those conditions the band m, shifted to baseband, of the signal at the output is

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

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

$$\beta^{(u)}_{mB}(f) = \sum_i \alpha_{im} X_{i,m}(f) G^{(u)}_{iB}(f) \tag{13}$$

As can be seen in (12) if it is put the constraint that $\sum_i \alpha_{i,m} = 1$ the user of interest is not distorted. The power spectral density of additive the noise disturbance $N^{'}_{mB}(f)$ is given by

$$\eta_{out_{mB}}(f) = \sum_i |\alpha_{i,m}|^2 |X_{i,m}(f)|^2 \eta_{in_{iB}}(f) \quad (14)$$

The design criteria implies that the weights $\alpha_{i,m}$ are dimensioned so that the Mean Squared Error (MSE)

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

is minimized subject to the condition that $\sum_i \alpha_{im} = 1$. In (15) T_{Bt} is the burst duration time and C correspond to the number of symbols existing in one burst.

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 of the input subbands with the same bandwidth and spaced by multiples of $\frac{1}{T}$ of the output subband. The shorter the subband better the performance.

The UMTS-TDD standard was designed to accommodate several simultaneous transmission rates. The different rates are accomplished by varying the spreading factor. The spreading code is composed by the product between the channelization code and the scrambling code. The channelization code lasts for one symbol and its number of chips is equal to the spreading factor. The scrambling code lasts for Q_{max} chips or during $\frac{Q_{max}}{Q}$ symbols being Q the spreading factor. 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 in the case of $Q \neq Q_{max}$. In the case of UMTS-TDD, $Q_{max} = 16$.

The approach to extend the canceller explained above to multirate is to consider that at the input of the canceller there are a decomposition of each signal with a spreading factor different of sixteen in several signals with spreading factor of sixteen.

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

$$s(t) = \sum_{l=0}^{\frac{Q_{\max}}{Q}-1} \sum_k a^l_k g_l(t - lT - \tfrac{Q_{\max}}{Q} kT) \quad (16)$$

where $\{a^l_k\}$ are the sequences of information symbols, $\frac{1}{T}$ the symbol rate (depends of the spreading factor) 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.

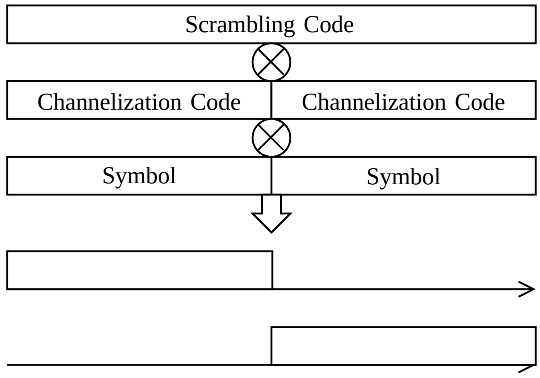

Fig. 2. Decomposition of the signal seen at the transmitter

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

$$s_l(t) = \sum_k a^l_k g_l(t - lT - \tfrac{Q_{\max}}{Q} kT) \quad (17)$$

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 spreading factor of sixteen. The signatures waveforms used in the cancelling process are 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. The proposed configurations are more suitable to be implemented in the uplink because the detectors require knowledge of the spreading codes of the active users. 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.

B. Channel Model

The channel model used in this work was the Geometrical Based Single Bounce Elliptical Model (GBSBEM)

TABLE I
Simulation parameters settings

Number of Taps	2
Velocity	50 Km/h
Path Loss	3.7
Maximum Delay Spread	2.0 μs
Degrees of Freedom of FSC	16
Number of samples per chip	4
Line of Sight Distance	300m

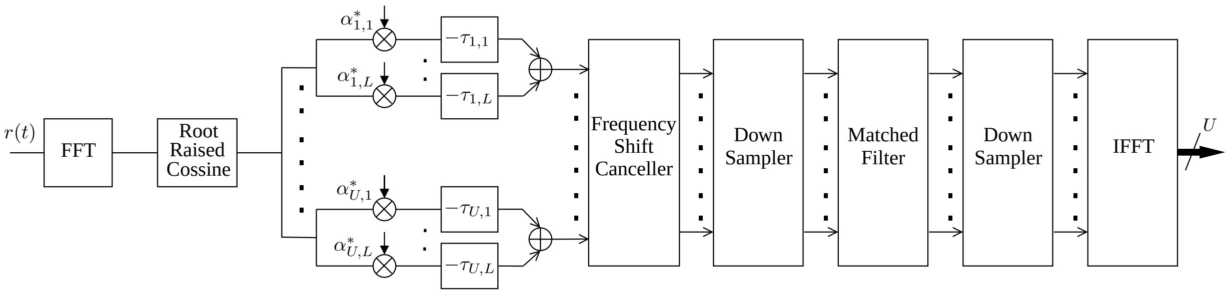

Fig. 3. Configuration with combination prior cancelling (POST-FSC)

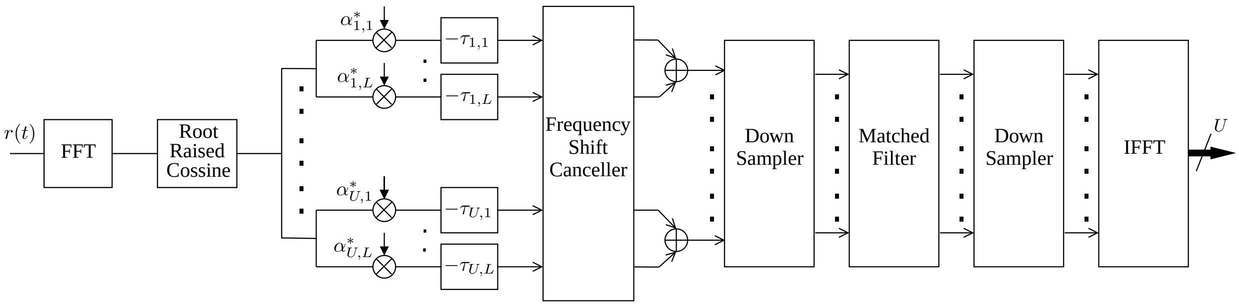

Fig. 4. Configuration with combination after cancelling (PRE-FSC)

proposed by Liberti [5]. This model was developed for microcell and picocell environments. The propagation channel is characterized by L paths for each user, one in line of sight plus $L-1$ arriving from remote reflectors located randomly within an ellipsis where the base station and the mobile unit are at the foci. Each path is characterized by complex constant and a delay. The probability distribuition of the delay of each tap have a maximum at sample zero and increases linearly from the sample one until the maximum delay spread sample. The phase of the complex constant is uniformly distributed in $[0, 2\pi[$. The amplitude of the complex constant follows a Rayleigh distribution. The channel model used takes in account the Doppler Effect for the amplitudes. The channel parameters are assumed to be constant within each burst.

C. Receivers

Figure 3 and 4 depicts the basic configurations for the detector that includes the Frequency Shift Canceller. If the Frequency Shift Canceller block is removed the detectors are conventional RAKEs. All the operations made between the FFT and IFFT blocks are frequency domain operations despite the fact that the names reflect the correspondent time domain operations. The signal in $r(t)$ has a resolution of four samples per chip (in case of evaluation of performance for imperfect channel estimates in the parameter delay the resolution of the signal is sixteen samples per chip) and the first downsampling has the same factor. The second downsampling takes a factor equal to the maximum spreading factor (Q_{max}). The other detector configurations to be evaluated are the detectors composed by POST-FSC or PRE-FSC concatenated with PIC (figure 5).

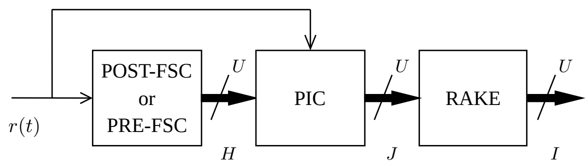

Fig. 5. Receiver including FSC plus PIC

In figure 5 the arrows crossed with letter U means that there are U (number of users) signals.

These two configurations are benchmarked with the conventional RAKE and conventional single stage PIC.

D. Results

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

The results are presented for the following scenario.

- Scenario:
  - Eight users: Four with spreading factor of sixteen, two with spreading factor of eight and two with spreading factor of four (Full load system).
  - The users with spreading factor eight and four have powers above the users of spreading factors of sixteen of 3 and 6dB respectively.

The results presented are only for the users of spreading factors of sixteen because the results for the other users of different spreading factors are similar.

In figure 6 the bit error rate of the several detector configurations assuming perfect channel estimates is presented. Both configurations POST-FSC and PRE-FSC with PIC achieve a performance close to the single user

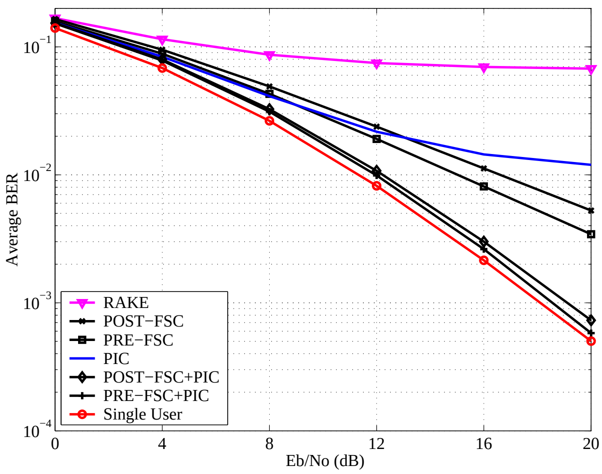

Fig. 6. Performance with perfect channel estimates

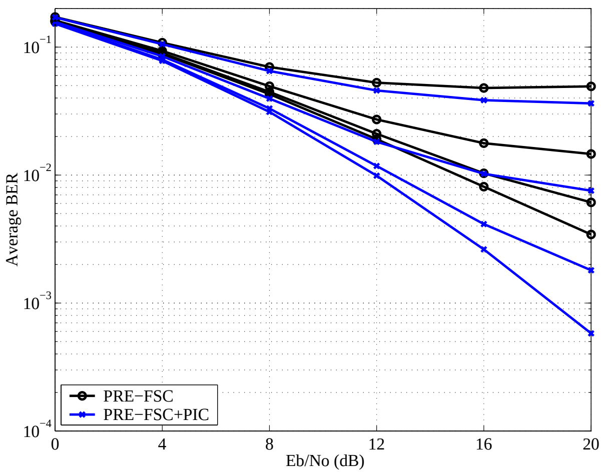

Fig. 7. Performance with Amplitude Standard Deviation of 0%, 5%, 10% and 20% of the medium amplitude of the taps

lower bound. The impact of imperfect channel estimates was assessed and results are presented in figures 7, 8 and 9 for the cases of the PRE-FSC with PIC and with the PRE-FSC in a standalone mode. The performance degradation of the POST-FSC with imperfect channel estimates is similar to the PRE-FSC. For the simulations a Gaussian distribution of the errors around the true value was assumed.

The performance degradation (PRE-FSC with PIC) with Amplitude Standard Deviation of 5% for a BER of 5E-3 is about 1 dB. The performance degradation with Phase Standard Deviation of 10 degrees for a BER of 5E-3 is about 6 dB. The performance degradation with Delay Standard Deviation of 0.05 chip period for a BER of 5E-3 is about 2 dB. For POST-FSC with PIC, this values are similar (graphics not shown).

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 have shown that significant gains

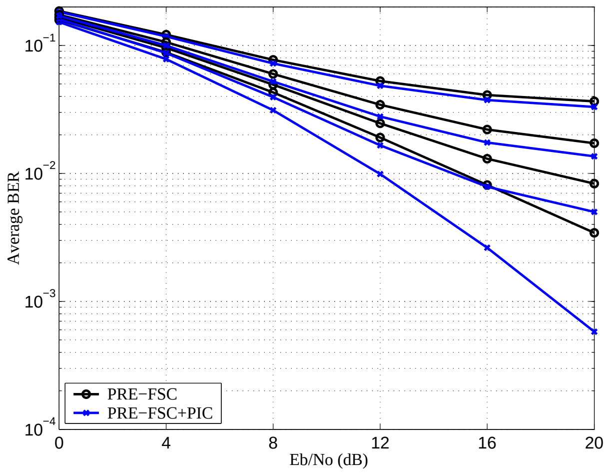

Fig. 8. Performance with Phase Standard Deviation of 0, 10, 15 and 20 degrees

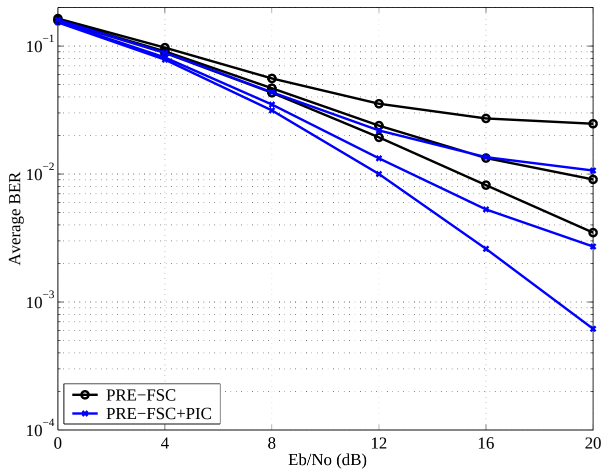

Fig. 9. Performance with Delay Standard Deviation of 0, 0.05 and 0.1 chip period

are obtained either with the POST and PRE-FSC when implemented in a standalone mode, and a performance close to the single user lower bound is achieved when either of the configurations are concatenated with a PIC. The sensitivity of the algorithms to imperfect channel estimates was investigated and the results have shown similar behavior for both configurations with PIC.

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