Robust Detection for the Uplink of Broadband Wireless Systems with Strong ACI Levels

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Abstract - This paper considers the uplink of wireless systems using SC-FDE schemes (Single-Carrier with Frequency-Domain Equalization) with OQPSK (Offset Quaternary Phase Shift Keying) modulations. We have an asynchronous scenario where only a coarse synchronization between mobile terminals and the base station is required.

We present an iterative frequency-domain receiver for joint equalization, residual carrier synchronization and ACI (Adjacent Channel Interference) suppression. Our performance results show that the proposed receiver is able to cope with strong ACI levels, even with only a coarse synchronization between users.[1]

I. Introduction

Future broadband wireless systems will have very high transmission rates, combined with high power and spectral efficiencies. This presents a considerable challenge since it is desirable to have low-complexity MTs (Mobile Terminals) and the complexity of the BS (Base Station) should not be too high. For this reason, block transmission techniques combined with frequency-domain processing have been proposed for these systems [1], [2]. It is generally accepted that OFDM schemes are recommendable for the downlink. Since an efficient and low-cost power amplification is mandatory at the MT, SC-FDE (Single-Carrier with Frequency Domain Equalization) preferable for the uplink. For this reason they were proposed for the uplink of 3GPP LTE (Long Term Evolution).

To allow high spectral efficiencies, the transmitted signals are usually designed to have compact spectrum, with bandwidths just slightly above the symbol rate (e.g., by employing a square-root raised-cosine filtering with small roll-off factor). However, the adoption of very compact spectra leads to implementation difficulties, since it is difficult to design the filters for channel separation and the synchronization requirements are stricter. Moreover, the transmitted signals have higher envelope fluctuations when the roll-off factor is very low, leading to amplification difficulties. Therefore, the transmitted signals usually employed have bandwidth higher than the symbol rate and the different frequency channels are separated accordingly to avoid ACI (Adjacent Channel Interference), decreasing the overall spectral efficiency of the system. A promising technique was proposed in [3] where the separation between frequency channels is equal to the symbol rate, regardless of the transmission bandwidth associated to each channel. To cope with the high interference levels that can result form having significant overlap between adjacent channels an iterative frequency-domain receiver with joint equalization and ACI suppression was proposed in [3]. That receiver takes advantage of spectral correlations for the separation of cyclostationary signals [4], [5], [6] and allows excellent performance. However, perfect synchronization between users was required. Since it is very difficult to maintain a good synchronization between different users, some time and, especially, frequency misalignments are almost unavoidable. especially if we want to reduce the cost of the MT.

For very-low-cost MT it is desirable to have grossly nonlinear power amplifiers, which are simpler to implement and have higher amplification efficiency and output power. However, these amplifiers are only recommendable if the signal at its input has an almost constant envelope. An interesting option is to employ OQPSK-type signals (Offset Quaternary Phase Shift Keying), which can have reduced envelope fluctuations or even a constant envelope in the case of MSK signals (Minimum Shift Keying).

In this paper, we consider the uplink of broadband wireless systems where SC-FDE schemes (Single-Carrier with Frequency-Domain Equalization) are employed by low-cost MT. We consider the use of OQPSK signals and different frequency channels are separated by the symbol rate, allowing very high spectral efficiency. Moreover, system is quasi-asynchronous, since only a coarse synchronization between MT is required. We present an iterative frequency-domain receiver for joint equalization, residual carrier synchronization and ACI.

II. System Description and Transmitted Signals

We consider an FDMA (Frequency Division Multiple Access) scheme where an SC-FDE modulation is employed for each frequency channel. We have P channels, with the carrier frequency of the pth frequency channel denoted by f_p, $p = 1, 2, \ldots, P$, with $f_p - f_{p-1} = \delta f$, i.e., the frequency channels are equally spaced. It is assumed that the blocks transmitted by each frequency channel have the same dimensions.

The length-M data block to be transmitted by the pth channel is $\{s_{m,p}; m = 0, 1, \ldots, M-1$, where $s_{m,p} = \pm 1 \pm j$

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is the mth data symbol (a Gray mapping rule is assumed). The corresponding signal is

$$s_p(t) = \sum_{m=-M_G}^{M-1} \text{Re}\{s_{m,p}\} h_T(t - mT_S) + \\ + \sum_{m=-M_G}^{M-1} j\text{Im}\{s_{m,p}\} h_T(t - mT_S - \Delta T_S), \quad (1)$$

with T_S denoting the symbol duration, M_G denoting the number of samples at the CP (Cyclic Prefix) (it is assumed that $s_{-m,p} = s_{M-m,p}$) and $h_T(t)$ is the transmitted pulse shaping filter response. (1) corresponds to a QPSK signal for $\Delta T_S = 0$ and an OQPSK signal for $\Delta T_S = T_S/2$.

The most common option with respect to $h_T(t)$ is to adopt a square-root raised-cosine filtering with a given roll-off factor β. This allows zero ISI (Inter-Symbol Interference) at the matched filter output for an ideal AWGN channel, together with compact spectrum, since the bandwidth of $s_p(t)$ is $B = (1+\beta)R_S$ ($R_S = 1/T_S$ is the symbol rate). The major problem is that the transmitted signals have relatively high envelope fluctuations, even for OQPSK schemes (although the dynamic range of the envelope is much lower for OQPSK signals since there are no zero crossings (see fig. 1), a clear advantage when grossly nonlinear power amplifiers are employed).

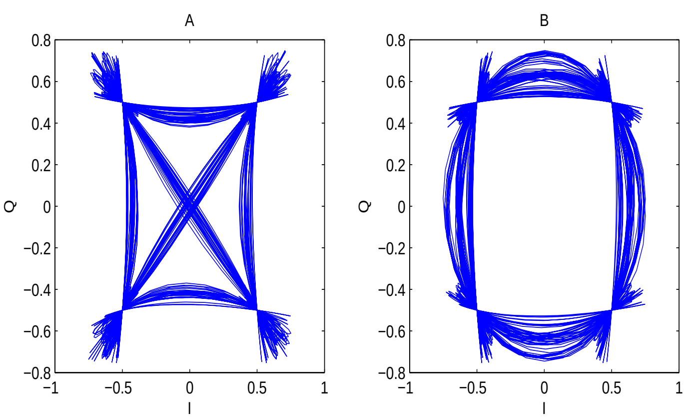

Fig. 1. I-Q diagram for QPSK (A) and OQPSK (B) schemes when a square-root raised-cosine filtering with roll-off $\beta = 1$ is adopted.

If grossly nonlinear power amplifiers are employed then the signal at its input should have constant envelope. This means that a square-root raised-cosine filtering is not suitable, even for OQPSK schemes. In that case, we could employ MSK signals, which can be regarded as constant-envelope OQPSK schemes where $h_T(t) = \cos(\pi t/T_S)\,\text{rect}(t/T_S)$. The major problem with MSK signals is that they have infinite bandwidth. Even if we just consider the main lobe and the first side lobe (where 99% of the power is concentrated), the corresponding bandwidth is $B = 5R_S$, 5 times the minimum bandwidth (achieved with a square-root raised-cosine filtering with roll-off $\beta = 0.0$). This means that the ACI levels might be very high, unless $\delta f >> R_S$ (a situation that is not desirable due to the inherently low spectral efficiency), precluding the separate detection of each frequency channel. In that case, we need to jointly detect all frequency channels, as proposed in [3].

III. Basic Receiver Structure

Let us assume that $\delta f = R_S < B$. To cope with ACI, we consider a receiver similar to the one proposed in [3]. Let us first assume perfect synchronization between different MT and the BS. The extension to asynchronous scenarios will be considered in sec. IV.

We employ a wideband reception filter and the resulting signal is sampled with an oversampling factor J, large enough to avoid aliasing effects, leading to the time-domain received block $\{y_m; m = 0, 1, \ldots, JM-1\}$. The corresponding frequency-domain block, obtained after an appropriate size-JM DFT operation (Discrete Fourier Transform), is $\{Y_k; k = 0, 1, \ldots, JM-1\}$, where Y_k can be written as

$$Y_k = \sum_{p=1}^{P} S_{k-\Delta M_p,p} H_{k,p} + N_k, \quad (2)$$

with $H_{k,p}$ denoting the overall channel frequency response for the kth frequency of the pth frequency channel, and N_k denoting the corresponding channel noise. The block $\{S_{k,p}; k = 0, 1, \ldots, M-1\}$ is the size-M DFT of $\{s_{m,p}; m = 0, 1, \ldots, M-1\}$ (it is assumed that $S_{k,p}$ is periodic, with period M, i.e., $S_{k,p} = S_{k+M,p}$ for any k) and ΔM_p is an appropriate circular shift accounting for the difference between the carrier frequency of the pth channel and the reference frequency of the local oscillator at the receiver.

Our receiver is based on the multichannel detection scheme proposed in [3], where each iteration consists of P detection stages, one for each frequency channel. We have one length-JM frequency-domain feedforward filter and P length-M frequency-domain feedback filters, one for each channel (the receiver implementation can be simplified since most feedforward and feedback coefficients are zero).

Since $S_{k,p} = S_{k+lM,p}$ for any l, the kth frequency-domain sample associated with the pth frequency channel can be given by

$$\tilde{S}_{k,p} = \sum_{l=0}^{J-1} F_{k+lM+\Delta M_p,p} Y_{k+lM+\Delta M_p,p} - \\ - \sum_{p'=1}^{P} B_{k,p}^{(p')} \overline{S}_{k+\Delta M'_p,p'} \quad (3)$$

where $F_{k,p}$ ($k = 0, 1, \ldots, JM-1$) denote the feedforward coefficients and $B_{k,p}^{(p')}$ ($k = 0, 1, \ldots, M-1$; $p = 1, 2, \ldots, P$) denote the feedback coefficients. The block $\{\overline{S}_{k,p'}; k = 0, 1, \ldots, M-1\}$ is the DFT of the block $\{\overline{s}_{m,p'}; m = 0, 1, \ldots, M-1\}$, where $\overline{s}_{m,p'}$ is the average value of $s_{m,p'}$ conditioned to the latest FDE output. For QPSK (and OQPSK) constellations

$$\overline{s}_{m,p'} = \tanh\left(\frac{\tilde{s}_{m,p'}^I}{\sigma_i^2}\right) + j\tanh\left(\frac{\tilde{s}_{m,p'}^Q}{\sigma_i^2}\right), \quad (4)$$

where

$$\sigma^2 = \frac{1}{2}E[|s_{m,p'} - \tilde{s}_{m,p'}|^2] \approx \frac{1}{2M}\sum_{m=0}^{M-1}|\hat{s}_{m,p'} - \tilde{s}_{m,p'}|^2, \quad (5)$$

with $\hat{s}_{m,p'}$ denoting the hard decision associated to $\tilde{s}_{m,p'}$.

The optimum feedforward coefficients can be written as $F_{k,p} = \breve{F}_{k,p}/\gamma_p$, with

$$\gamma_p = \frac{1}{M}\sum_{l=0}^{J-1}\sum_{k=0}^{M-1}\breve{F}_{k+lM,p}H_{k+lM,p} \quad (6)$$

and $\breve{F}_{k+lM,p}$ ($l = 0, 1, \ldots, J-1$ and $k = 0, 1, \ldots, M-1$) obtained from the set of J equations:

$$(1-\rho_p^2)H^*_{k+lM,p}\sum_{l'=0}^{J-1}\breve{F}_{k+l'M,p}H_{k+l'M,p}$$
$$+\sum_{p'\neq p}(1-\rho_{p'}^2)H^*_{k+lM,p'}\sum_{l'=0}^{J-1}\breve{F}_{k+l'M,p}H_{k+l'M,p'} +$$
$$+\frac{\breve{F}_{k-lM,p}}{SNR_p} = (1-\rho_p^2)H^*_{k-lM,p}, \;\; l = 0, 1, \ldots, J-1, \quad (7)$$

where $SNR_p = E[|S_{k,p}|^2]/E[|N_k|^2]$ and the correlation coefficient ρ_p is given by (see [7])

$$\rho_p = \frac{1}{2M}\sum_{m=0}^{M-1}(|\mathrm{Re}\{\overline{s}_{m,p'}\}| + |\mathrm{Im}\{\overline{s}_{m,p'}\}|). \quad (8)$$

The optimum feedback coefficients are

$$B^{(p)}_{k,p} = \sum_{l=0}^{J-1}F_{k+lM,p}H_{k+lM,p} - \gamma_p\delta_{p,p'}, \quad (9)$$

($\delta_{p,p'} = 1$ for $p = p'$ and 0 for $p' \neq p$).

Since $F_{k,p} = 0$ when $H_{k,p} = 0$, we just need to compute MB/R_S) coefficients for each frequency channel and each iteration. (7) suggests that we need to solve a system of J equations for obtaining each feedforward coefficient. However, since the bandwidth of the transmitted signals associated to each frequency channel is B, the multiplicity of each $S_{k,p}$ is upper-bounded by $\lceil B/R_S \rceil$ ($\lceil x \rceil$ denotes "smaller integer greater than x"). Therefore, we just need to solve a system of $\lceil B/R_S \rceil$ equations for each of the M frequencies. For a square-root raised cosine filtering this means solving a system of two equations whenever there is ACI ($B = (1+\beta)R_S > \delta f$. For MSK signals the number of equations is much higher; however, we just need so solve a system of five equations if we assume that all power is concentrated in the main lobe and the first side lobe of the spectrum.

The receiver described above considers QPSK constellations. The extension to OQPSK constellations is straightforward. We just need to employ two FDE, one associated to the in-phase component and the other, operating on a delayed version of the signal, associated to the quadrature component. The FDE coefficients are identical. Under perfect synchronization, we just need the real part of the output for the first FDE and the imaginary part of the output for the second FDE. However, if we have carrier synchronization errors, as in the next section, it is better to have the complex outputs for each FDE and to remove the quadrature (or in-phase) interference when detecting the in-phase (or quadrature) component.

IV. Asynchronous System

The receiver structure considered in the previous section requires perfect time and frequency synchronization between the BS and the MTs. Typically, this synchronization is ensured though a feedback channel, from the BS to each MT. However, some residual time and/or frequency errors between BS and the MTs is unavoidable.

Let us assume that there is a time misalignment ΔT_p on the block associated to the pth MT. If the CP is long enough to cope with the length of the channel impulse response plus $\max_{p\neq p'}|\Delta T_p - \Delta T_{p'}|$ then it can easily be shown that the corresponding received frequency-domain block is $\{Y_k^{(\Delta T)}; k = 0, 1, \ldots, N-1\}$, with

$$Y_k^{(\Delta T)} = \sum_{p=1}^{P}S_{k-\Delta M_p,p}H_{k,p}\exp(-j2\pi k\Delta T_p/T) + N_k^{(\Delta T)} \quad (10)$$

($k = 0, 1, \ldots, N-1$), where the equivalent noise component $N_k^{(\Delta T)}$ has the same statistical properties of N_k. Clearly,

$$Y_k^{(\Delta T)} = \sum_{p=1}^{P}S_{k-\Delta M_p,p}H^{(\Delta T_p)}_{k,p} + N_k^{(\Delta T)}, \quad (11)$$

with the equivalent channel frequency response given by $H^{(\Delta T_p)}_{k,p} = H_{k,p}\exp(-j2\pi k\Delta T_p/T)$. Therefore, provided that we have accurate channel estimation and the CP is long enough, we can easily deal with timing errors (these errors are absorbed in the overall channel frequency response associated to each user).

With respect to the frequency errors, it can be shown that the samples at the input of the decision device associated to the pth user are approximately given by

$$\tilde{s}^{\delta f_p}_{m,p'} \approx \tilde{s}_{m,p'}\exp(j2\pi\delta f_m nT/M) \quad (12)$$

($m = 0, 1, \ldots, M-1$), with T denoting the duration of the useful part of the block and δf_p denoting the frequency error between the local oscillator at the pth MT and the local oscillator at the BS (this is a good approximation provided that $\delta f_p T < 1/2$ and $M >> 1$). From (12), it is clear that we have a progressive phase rotation along the data symbols, we just have to estimate δf_p and compensate the phase rotation before the decision device.

The frequency offset can be estimated using specially designed reference blocks. However, since the frequency error is different for different users, we would need to estimate all frequency errors. This means P reference blocks (multiplexed in time and/or frequency), which might lead to significant overhead. To avoid this problem, we will consider a modified version of our receiver that combines user separation with iterative, DD (Decision-Directed) estimation and compensation of frequency errors, as proposed in [8] for scenarios without ACI.

For each iteration, the detection of the pth user can be made using the structure depicted in fig. 2. Δf_p is estimated as follows:

$$\widehat{\delta f_p} = \frac{M}{2\pi \Delta M T} \arg \left\{ \sum_{m=0}^{M-\Delta M-1} \frac{\tilde{s}^{\Delta f}_{m+\Delta M,p} \tilde{s}^{\Delta f *}_{m,p}}{\hat{s}_{m+\Delta M,p} \hat{s}^{*}_{m,p}} \right\}, \quad (13)$$

with $\Delta M \approx 2M/3$. To avoid high error rates in $\hat{s}_{m,p}$, the frequency error can not be too high (say, $\Delta f_p T < 0.1$ or 0.2).

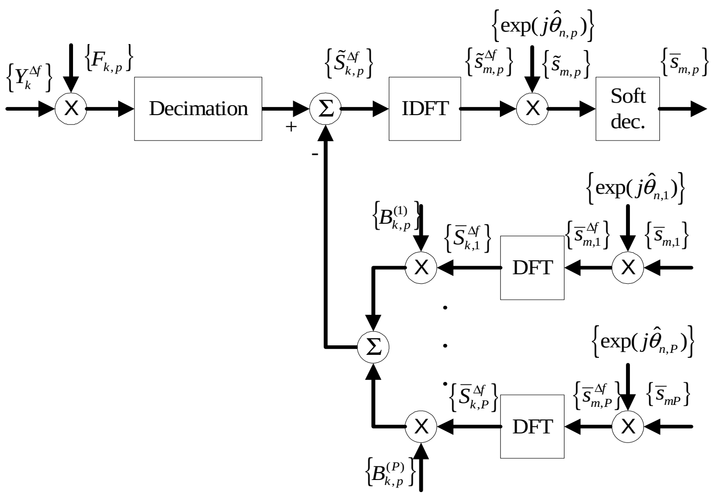

Fig. 2. Detection of the pth user in the presence of frequency errors ($\theta_{n,p} = 2\pi \delta f_p n T/M$).

V. Performance Results

In this section we present a set of performance results concerning the proposed receiver for SC-FDE schemes with strong ACI levels. We consider the uplink transmission with $P = 5$ frequency channels, each one transmitting blocks of $M = 256$ data symbols (similar results were observed for other values of M, provided that $M >> 1$). We consider both OQPSK signals with square-root raised-cosine filtering with roll-off $\beta = 1$ and MSK signals. The propagation channel for each frequency channel is characterized by the power delay profile type C for the HIPERLAN/2 (HIgh PERformance Local Area Network) [9], with uncorrelated Rayleigh fading between frequency channels and on the different paths (similar results were observed for other strongly frequency-selective channels). The duration of the useful part of the data blocks (M symbols) is 4μs and the CP has duration 1.1μs. The power amplifier of each MT is assumed to be linear for the dynamic range of the signal at its input (this dynamic range is very small for square-root raised-cosine pulses if the roll-off factor is not too small; for MSK signal a grossly nonlinear amplifier can be employed, since they have constant envelope). At the receiver, the average power associated to each frequency channel is the same. The BS has perfect channel knowledge for each user and a coarse estimate of the corresponding carrier frequency, with the residual frequency error δf_p for the pth user.

We employ the proposed receiver, with the frequency channels detected in a sequential order (i.e., for each iteration we detect first the frequency channel with lowest carrier frequency and proceed sequentially up to the frequency channel with highest frequency). This sequential order is not necessarily optimal, since the channels with lowest interference levels should be detected first. However, after a few iterations the performance is almost the same regardless of the detection order. For square-root raised-cosine pulses we need to solve a system of 2 equations for each frequency; for MSK signals, we considered a relevant bandwidth $B \approx 5R_S$, i.e., we need to solve a system of 5 equations for each frequency.

Let us assume first perfect carrier synchronization. The average BERs, averaged over all channels, for raised-cosine pulses and MSK signals are presented in figs. 3 and 4, respectively. We present both the performance without ACI ($\Delta f \geq B$) and the performance when $\Delta f = R_S$, allowing high spectral efficiency but with strong ACI levels. For the sake of comparisons, we included the MFB performance (Matched Filter Bound). Clearly, our receiver allows good ACI suppression, with performance close to the performance without ACI. Moreover, the BER can be close to the MFB after just three or four iterations. The performances are slightly better for MSK signals due to the larger diversity inherent to the wider bandwidth.

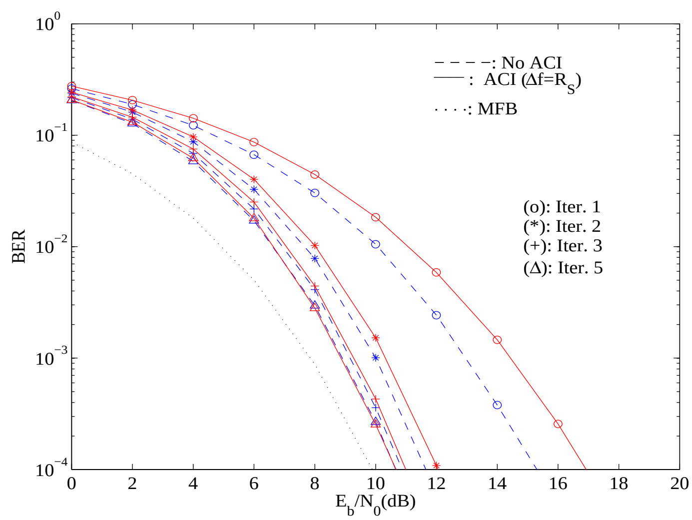

Fig. 3. Average BER performance with perfect carrier synchronization, for square-root raised-cosine pulses with $\beta = 1$.

Let us consider now the case where there is a residual frequency error in the carrier estimates at the BS. For the sake of simplicity, we assume that $\delta f_p = \delta f$, i.e., we have the same frequency error for all users. Since we perform the DD frequency error estimation separately for each user, the performance of a given user is almost independent of the frequency error for the adjacent channels.

Figs. 5 and 6 present the BER performance for MSK signals when $\delta f T = 0.05$ or 0.1, respectively, without frequency error estimation, with our DD frequency error estimation and without frequency error (or with perfect frequency er-

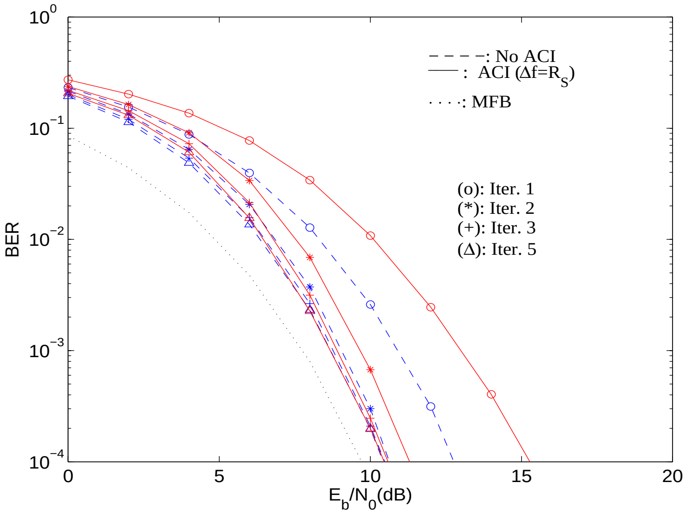

Fig. 4. Average BER performance with perfect carrier synchronization, for MSK signals.

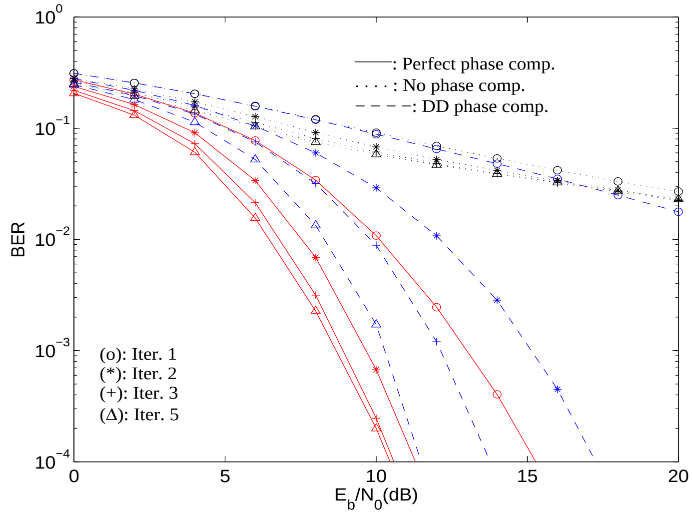

Fig. 6. Average BER for MSK signals, when $\delta f T = 0.1$.

ror compensation). Similar results were observed for square-root raised-cosine pulses. Clearly, our DD frequency error estimation technique allows almost ideal performance when $\delta f = 0.05$. Moreover, it is powerful enough to with residual frequency errors as large as $\delta f_p T = 0.1$. This means that the local oscillators at the MT do not need to have very high stability.

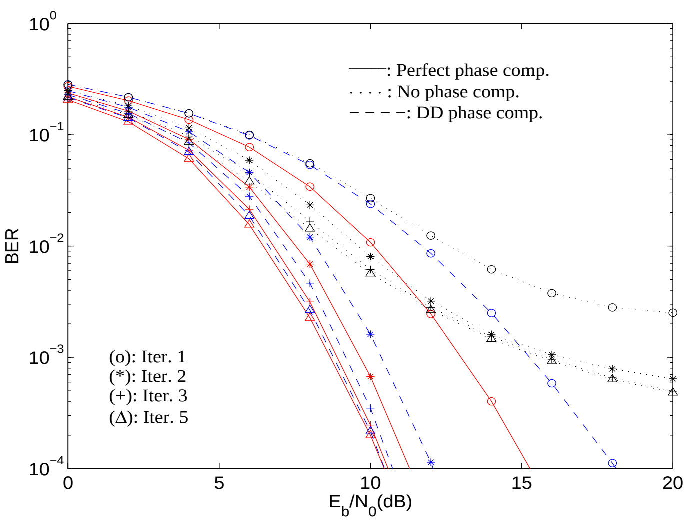

Fig. 5. Average BER for MSK signals, when $\delta f T = 0.05$.

VI. Conclusions

In this paper, we considered the uplink of broadband wireless systems where SC-FDE schemes are employed by low-cost MT. MSK signals with frequency channels separated by the symbol rate were assumed. We considered an asynchronous scenario where only a coarse synchronization between MT and the BS is required.

An iterative frequency-domain receiver for joint equalization, residual carrier synchronization and ACI suppression was presented and evaluated. This receiver is able to cope with strong ACI levels, even with only a coarse synchronization between users. This means we can employ low-cost local oscillators at the MT, since they do not need to have very high stability. Our performance results also show that it is possible to employ constant-envelope signals such as MSK signals, without compromising the overall spectral efficiency of the system, since spectrums can be separated by the symbol rate. This is an excellent candidate for the uplink if we want to have very-low-cost MT.

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