Correction of the CFO in OFDM Relay-based Space-Time Codes

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Abstract—In this paper, we analyze the impact of carrier frequency offset (CFO) on the performance of orthogonal frequency division multiplexing (OFDM) transmission employing space-frequency coding over relay channels. The challenge in such systems lies in the difficulty of canceling the interference resulting from the different CFOs that correspond to the relays involved in the transmission. We first analyze the CFO correction schemes and examine their impact on the achievable information rates. Further, we analyze the interference cancellation (IC) technique based on the so-called turbo-principle, that is, which jointly detects and decodes the received data. The increase of the rates achievable thanks to IC is assessed via parametric description of the iterative process. We provide examples that demonstrate the efficacy of the proposed scheme and numerical results are contrasted with theoretical performance limits.

I. INTRODUCTION

In this paper with analyze the relay-based space-frequency coded OFDM transmission. We asses the rates achievable in the presence of the multiple CFOs and evaluate the potential of the rate increase due to iterative interference cancellation.

Cooperative transmission is considered a viable and low-cost solution to improve the coverage (reliability) of wireless communications [1]. In particular, the use of multiple relays that employ space time coding (STC) allow to harvest the diversity inherent in relay channels [2]. However, when employing STC, time synchronization between antennas becomes crucial [3]. To overcome this problem, one may use orthogonal frequency division multiplexing (OFDM), which has been adopted in many wireless communication standards including WiFi, WiMax [5], and LTE [4]. Since then, the code is constructed using space- and frequency-dimensions, such a scheme is called a space-frequency coding (SFC).

A downside of using OFDM is that it is sensitive to frequency synchronization errors [6] [7]. For instance, a slight carrier frequency offset (CFO) results in destroying the orthogonality between subcarriers, leading to significant performance degradation. Therefore, accurate CFO estimation is essential to have a working OFDM system. The most common approach to the CFO-problem in single-antenna transmission relies on the accurate estimation of the CFO and its removal from the received signal through frequency correction [6]. On the other hand, in the case of multiple-relay transmission, the situation is similar to that occurring in orthogonal frequency division multiple access (OFDMA) where the base station receiver has to deal with CFOs that are different for various users [8]. The main difference with the single-user (or single relay) transmission is that the effect of CFO cannot be removed from the signal via simple frequency correction so interference is unavoidable.

Many papers had considered CFO estimation for OFDMA systems [8] [9] or methods to deal with the interference appearing in such cases [10] [11]. These works were based on the simulation of particular modulation/coding setups. Such an approach allows to validate basic concepts but makes it difficult to draw general conclusions, particularly when comparing different transmission strategies. Aiming at a more general results, we evaluate the transmission rates achievable for various strategies the receiver may deploy to mitigate the effects of the CFOs in relay-based SFC transmission. In particular, we assess the performance of the channel-dependent CFO correction and of an iterative (turbo) detection/decoding [12]. Our approach provides a picture that is more complete than any simulation-based study, can be related to particular cases of the coded transmission (that we also show as an example), and at the same time may be meaningfully contrasted with other relay-based transmission strategies.

The paper is organized as follows: the system model is described in Sec. II and in Sec. III we analyze the throughput attainable with Alamouti SFC scheme in CFO-corrupted relay channel. The results obtained with practical coding/modulation schemes are presented in Sec. IV and conclusions – in Sec. V.

II. SYSTEM MODEL

We consider here a two-hop relay channel shown in Fig. 1, where the source terminal is broadcasting information toward the relays. Since the transmission is coded, the relays can reliably decode the broadcasted message and forward it to the destinations. We assume also that –in order to harvest the diversity of the relays-to-destination channels– the relays implement the Alamouti scheme [13]. This simplifies the receiver’s processing as the sent symbols may be then recovered with a simple arithmetic operations.

In order to counteract –via simple processing– the effect of frequency-selective fading (on all links), we assume that the transmission is based on OFDM. This has the advantage of
avoiding the strict time-synchronization requirement between
the relays but its disadvantage is the sensitivity to the CFO at
the destination. This problem of CFO in OFDM/OFDMA
transmission has been extensively treated in the literature, [8].

The Alamouti transmission scheme is thus affected by the
interference due the residual CFO and its mitigation/reduction
strategies are the main focus of this paper.

To establish the notation, we consider the case of single-
antenna OFDM where \( N \) symbols \( \mathbf{a} = [a_0, a_1, ... , a_{N-1}]^T \)
(we will use \((\cdot)^T\) and \((\cdot)^H\) to denote transpose and transpose-
conjugate operators) are passed to a \( N \)-points inverse Discrete
Fourier Transform (DFT). The resulting sequence is extended
with the cyclic prefix, and sent over the frequency-selective
channel whose frequency responses is known (i.e., accurately
estimated). At the receiver, after sampling, and the prefix’
removal, the signal is passed though the DFT which produces the
signal

\[
\mathbf{r} = \mathbf{F} \mathbf{D}(\delta) \mathbf{F}^{-1} \mathbf{H} \mathbf{a} + \mathbf{w}
\]  

(1)

where \( \mathbf{r} = [r_0, r_1, ..., r_{N-1}]^T \), \( \mathbf{w} = [w_0, ..., w_{N-1}]^T \)
is a zero-mean Gaussian noise (AWGN) with autocorrelation
\( \mathbb{E} \{ \mathbf{w} \mathbf{w}^H \} = \mathbf{N}_0 \) (\( \mathbf{I} \) is \( N \times N \) identity matrix),
\( \mathbf{F} \) and \( \mathbf{F}^{-1} \) are \( N \)-points DFT and inverse DFT matrices,
\( \mathbf{H} = \text{diag} \{ H_0, H_1, ..., H_{N-1} \} \) is the diagonal matrix
whose entries are channel frequency responses \( H_n = \sum_{n=0}^{N-1} h_n \exp(-2\pi kn/N) \) (where \( h_n, n = 0, ..., N-1 \) is the
impulse response of the channel), and

\[
\mathbf{D}(\delta) = \text{diag} \{ 1, e^{2\pi \delta/N}, e^{2\pi \delta/2N}, ..., e^{2\pi \delta(N-1)/N} \}
\]  

(2)

is the diagonal matrix representing the effect of the CFO \( \delta \).

If \( \delta = 0 \), then \( \mathbf{D}(\delta) = \mathbf{I} \) and (1) reduces to \( \mathbf{r} = \mathbf{H} \mathbf{a} + \mathbf{w} \),
otherwise symbols \( s_n \) interfere with each other, i.e., introduce
the so-called inter carrier interference (ICI).

After some algebra it can be shown that [8]

\[
r_n = H_n f(\delta) a_n + ICI_n(\delta) + w_n
\]  

(3)

where

\[
f(\delta) = \frac{\sin(\pi \delta)}{N \sin(\pi \delta/N)} e^{j\pi \delta N/2} \]  

(4)

\[
ICI_n(\delta) = \sum_{p=0}^{N-1} H_p a_p f(\delta + p - n)
\]  

(5)

We now extend the above model to the two-relay cooperative
transmission scheme shown in Fig. 1. We are interested
in the relay-destination communication, so we assume that
the symbols \( \mathbf{s} = [s_0, ..., s_{N-1}]^T \) are received at the relays
without errors. This is a reasonable assumption as we consider
here a coded transmission. The symbols \( s_n \) are then forwarded
towards the destination. Let \( a_i \) denote the sequence transmitted
from the relay \( i \) \( (i = 1, 2) \). As such, the two sequences form
a space-frequency code based on the Alamouti scheme. It can be
expressed in a matrix form as

\[
\mathbf{A} = \begin{bmatrix} \mathbf{a}_1^T \\ \mathbf{a}_2^T \end{bmatrix} = \begin{bmatrix} s_1 & -s_2 & s_3 & -s_4 & ... & s_{N-1} & -s_N \\ s_2 & s_1 & s_4 & s_3 & ... & s_N & s_{N-1} \end{bmatrix}
\]  

(6)

Consequently, the received signal after processing is in the
following form

\[
\mathbf{r} = \mathbf{F} \cdot \mathbf{D}(\delta) \cdot \mathbf{F}^{-1} \cdot \mathbf{H}_1 \cdot \mathbf{a}_1 + \mathbf{F} \cdot \mathbf{D}(\delta - \Delta) \cdot \mathbf{F}^{-1} \cdot \mathbf{H}_2 \cdot \mathbf{a}_2 + \mathbf{w}
\]  

(7)

where \( \delta \) is the CFO between the destination and the first relay,
\( \Delta \) is the CFO between the second and the first relay, and \( \mathbf{H}_i \),
\( i = 1, 2 \) is the frequency responses of the channels between
the \( i \)-th relay and the destination.

From (7) we see that, in general, no matter what strategy is
chosen by the receiver for CFO compensation, some interference
is unavoidable: it is possible to eliminate the effect of the
CFO for the first relay \( (\delta = 0) \) or for the second one \( (\delta = \Delta) \),
but the effect of both CFOs cannot be removed simultaneously
(unless \( \Delta = 0 \), i.e., when there is no CFO between the relays).

Before considering any effect of the CFO-related interference,
the receiver applies the conventional ST decoding to the
received signals at time \( n \) and \( n + 1 \) which have the following
form [14], [15]

\[
r_n = H_{1,n} \cdot f(\delta) \cdot s_n + H_{2,n} \cdot f(\delta - \Delta) \cdot s_{n+1} + ICI_{1,n}(\delta) + ICI_{2,n}(\delta - \Delta) + w_n
\]  

(8)

\[
r_{n+1} = - H_{1,n+1} \cdot f(\delta) \cdot s_n^{*} + H_{2,n+1} \cdot f(\delta - \Delta) \cdot s_{n+1}^{*} + ICI_{1,n+1}(\delta) + ICI_{2,n+1}(\delta - \Delta) + w_{n+1}
\]  

(9)

The signals are gathered into a vector

\[
\begin{bmatrix} r_n \\ r_{n+1}^{*} \end{bmatrix} = \begin{bmatrix} H_{1,n} f(\delta) & H_{2,n} f(\delta - \Delta) \\ H_{2,n+1}^{*} f(\delta - \Delta) & -H_{1,n+1}^{*} f(\delta) \end{bmatrix} \begin{bmatrix} s_n \\ s_{n+1} \end{bmatrix} + \begin{bmatrix} ICI_{1,n}(\delta) + ICI_{2,n}(\delta - \Delta) \\ ICI_{1,n+1}(\delta) + ICI_{2,n+1}(\delta - \Delta) \end{bmatrix} + \begin{bmatrix} w_n \\ w_{n+1}^{*} \end{bmatrix},
\]  

(10)
and, next, the combining characteristic of the Alamouti scheme is applied
\[
\begin{bmatrix}
    y_n \\
    y_{n+1}
\end{bmatrix} = \mathbf{B}_n \begin{bmatrix}
    r_n \\
    r_{n+1}^*
\end{bmatrix}
\] (11)

Since all the transformations are linear, they may be written as
\[
y = \mathbf{G}s + \mathbf{G}'s^* + \mathbf{Q}w
\] (12)

where the form of \( \mathbf{G} \) and \( \mathbf{G}' \) may be deduced from (5), (10) and (11). The matrix \( \mathbf{Q} \) is block-diagonal composed of \( \mathbf{B}_n^{\mathbf{H}} \) so the covariance of the noise-related term, given by \( \mathbb{E}\{\mathbf{Q}\mathbf{Q}^\mathbf{H}\}N_0 \) is also block diagonal with entries of \( \mathbf{B}_n^{\mathbf{H}}\mathbf{B}_nN_0 \). If we assume that \( H_{i,n} \approx H_{i,n+1} \), matrix \( \mathbf{B}_n^{\mathbf{H}} \) is unitary, thus the covariance matrix' diagonal elements are given by \( g_n = |H_{1,n}|^2|f(\delta)|^2 + |H_{2,n}|^2|f(\delta - \Delta)|^2 \).

The observation \( \mathbf{y}_n \) in (12) contains both the desired symbol \( \pi_n \) as well as its conjugate \( \pi_n^* \), and is affected by all other symbols and their conjugates (ICI). The detection is possible at this stage but, in the following we will use the symbols \( s_n \) taken from the normalized constellation \( \mathcal{Y} \) that is separable into identical real and imaginary parts, i.e., \( \mathcal{Y} = \mathcal{X} \times (\mathcal{J}\mathcal{X}) \). This happens, e.g., when \( M \)-ary quadrature amplitude modulation (\( M \)-QAM) is used. The constellation \( \mathcal{Y} \) is zero-mean \( \sum_{x \in \mathcal{Y}} x = 0 \) and has unitary energy \( E_{\mathcal{Y}} = \sum_{x \in \mathcal{Y}} |x|^2 = 1 \), so the energy of the symbols \( s_n \) taken from \( \mathcal{X} \) is \( E_{\mathcal{X}} = \frac{1}{2} \).

Consequently, it is convenient to re-write (12) as
\[
\mathbf{y} = \begin{bmatrix}
    y_R \\
    y_I
\end{bmatrix} = \begin{bmatrix}
    \mathbf{G}_R + \mathbf{G}'_R & \mathbf{G}'_I - \mathbf{G}_I \\
    \mathbf{G}_I + \mathbf{G}'_I & \mathbf{G}'_R - \mathbf{G}_R
\end{bmatrix}\cdot \begin{bmatrix}
    s_R \\
    s_I
\end{bmatrix} + \begin{bmatrix}
    \mathbf{Q}_R - \mathbf{Q}_I \\
    \mathbf{Q}_I + \mathbf{Q}_R
\end{bmatrix}\cdot \begin{bmatrix}
    w_R \\
    w_I
\end{bmatrix}
\] (13)

\[
= \mathbf{Q}\mathbf{G}s + \mathbf{Q}\mathbf{w}
\] (14)

where \( (\cdot)_R \) and \( (\cdot)_I \) denote, respectively, the real and imaginary parts and \( (\cdot) \) denotes concatenation of the vectors/matrices appearing in (13).

III. PROCESsing AT THE RECEIVER

A. Performance Criterion

Since the receiver converts the signals \( \mathbf{y}_n \) into reliability metrics that will be used by the channel decoder, it is relevant to use the mutual information (MI) between the metrics and the corresponding coded bits as the performance measure; it will provide us with information about the maximum achievable rate (or “constrained capacity”) of the composite channel which comprises all the elements of the channel: modulation, DFT processing, and SF-decoding and that takes into account the CFO effects.

We assume that the interference can be approximated as Gaussian so the signal-to-interference-and-noise ratio
\[
\gamma_n = \frac{E_{\mathcal{X}} \cdot |\mathbf{y}_n|}{E_{\mathcal{X}} \cdot \sum_{l \neq n} |\mathbf{y}_n| + \frac{1}{2} N_0} \quad n = 0, \ldots, 2N - 1
\] (15)
defines the mutual information for the \( n \)-th element of \( \mathbf{y} \) as \( I_{\mathcal{X}}(\gamma_n) \), where the function \( I_{\mathcal{X}}(\gamma) \) is acquired through numerical integration for a given \( \mathcal{X} \) as explained in [16]. Here, \( \mathbf{Q}_n \) are the diagonal elements of \( \mathbf{Q} \mathbf{Q}_n^{-1} \). If we assume that \( \mathbf{B}_n \) are unitary, we can use \( \bar{\mathbf{y}}_n + 1 = \| \mathbf{y}_n \|^2 \) for \( n = 0, \ldots, N - 1 \).

The estimate of the MI at the output of the receiver, providing information about the maximum attainable transmission rate is thus given by
\[
I_{\text{tot}} = \frac{1}{N} \sum_{n=0}^{2N-1} I_{\mathcal{X}}(\gamma_n).
\] (16)

The maximum achievable rate \( I_{\text{tot}} \) is channel-dependent so to characterize the property of the transmission scheme in fading (variant) channel we will use the so called outage rate\(^1\) \( I_{\text{out},1-\epsilon} \) defined through
\[
\text{Pr}\{I_{\text{tot}} < I_{\text{out},1-\epsilon}\} = \epsilon.
\] (17)

This measure is convenient to evaluate the performance of the transmission over quasi-static channels, that is, where the coding and modulation are carried out in the same channel state (here: realizations of the channels’ frequency responses \( \mathbf{H}_1 \) and \( \mathbf{H}_2 \)). In fact, the outage rate is a much more practical measure that the so-called ergodic rate which requires the transmission to be sufficiently long to encompass all possible channel states. In the following examples the channels between the relays and the destination are assumed to have two zero-mean circularly-symmetric, complex Gaussian, equal-power, independently fading taps, that is, \( \mathbb{E}\{|h_0|^2\} = \mathbb{E}\{|h_1|^2\} = 0.5 \). The number of sub-carriers is given by \( N = 256 \).

B. CFO Compensation

We are interested here in the effect of the CFOs \( \delta \) and \( \delta - \Delta \) on the capacity of the relay channel. First we analyze the impact of the selection of \( \delta \) on the attainable rate \( I_{\text{tot}} \).

In Fig. 2 we show \( I_{\text{tot}} \) as a function of \( \delta \) for arbitrarily selected realization of channels \( \mathbf{H}_1 \) and \( \mathbf{H}_2 \). We also show the horizontal lines which denote the rate when only one relay (the first or the second) is used. In such a case, the CFO effect might be perfectly removed (that is why the capacity line is independent of \( \delta \)) and, for a fair comparison with two-relay transmission, the transmission power from the transmitting relay is doubled.

The simplest strategy is to set \( \delta = \frac{1}{2}\Delta \), so that, on average, both channels are affected in the same way. However, ignoring the knowledge of the channels \( \mathbf{H}_1 \) and \( \mathbf{H}_2 \) has a price: for example, if \( \Delta = 0.6 \), setting \( \delta = \frac{1}{2}\Delta \) yields the rate \( I_{\text{tot}} \approx 0.9 \) while optimizing \( \delta \) (setting \( \delta = 0.1\Delta \)) yields \( I_{\text{tot}} \approx 1.2 \).

The case shown in Fig. 3 is slightly different as the optimized CFO (\( \delta \approx 0.3\Delta \)) would yield rates similar to \( \delta = \frac{\text{frac}}{12}\Delta \). The importance of the optimization of \( \delta \) is elucidated further in Fig. 6 which gathers all the results.

\(^1\)It has the same meaning as the outage capacity, but the distinction is made to take into account the particular constellation \( \mathcal{X} \).
C. Interference Cancellation (IC)

Up to now we assumed that the output of the ST combining is directly converted into reliability metrics. We note that the relationship (12) might be exploited to diminish the level of interference via linear, e.g., minimum mean-square error (MMSE), filtering of the output $y$. We let this option aside, nevertheless, it may be evaluated using the proposed methodology. Instead, we focus on the removal of interference using the information provided by the channel decoder as shown in Fig. 4.

If the decoder uses itself the iterative decoding, such an approach would not increase the processing complexity significantly, i.e., might be almost seamlessly included in the decoding process. On the other hand, linear filtering of $y$ would first require solving a $2N$-dimensional linear equation (to design the filter). In practice, the decoder implements a soft-input soft-output processing which provides not only the estimates of the information bits but also the so-called extrinsic reliability metrics for all the coded (transmitted) bits $b$. These are usually expressed in the form of logarithmic likelihood ratios (LLR) $L = \log\{\Pr(b = 1)/\Pr(b = 0)\}$.

The LLRs obtained from the channel decoder become a priori information which lets us to decrease the interference’ energy. First, we calculate the expected values of the symbols

$$\hat{s}_n = E\{s_n\} = \sum_{x \in X} x \cdot Pr(s_n = x)$$  \hspace{1cm} (18)

where $Pr(x)$ are obtained from $L$ as shown e.g. in [19].

Next, the effect of the symbols’ estimates is subtracted from the received signal

$$y' = y - (G - \text{diag}(G))\hat{s}$$  \hspace{1cm} (19)

which increases the effective SNR to

$$\gamma'_n = \frac{E_X \cdot |g_{n,n}|^2}{\sum_{l \neq n} |g_{n,l}|^2 \cdot \text{Var}\{s_l\} + |g_{n,n}|^2 \frac{1}{2} N_0}$$  \hspace{1cm} (20)

where $\text{Var}\{s_n\} = E\{(s_n - \hat{s}_n)^2\}$.

Since $\text{Var}\{s_n\} \leq E_X$, $\gamma'_n \geq \gamma_n$.

D. Performance Limits of IC

In order to establish the limits of the proposed interference cancellation scheme, we rely on the parametric description of the iterative turbo process. Namely, we will use the so-called EXIT charts [17] to evaluate the achievable rates of the transmission. In particular, we will find the area below the
EXIT function of the detector [18], which approximates well the maximum achievable transmission rate.

The EXIT function, characterizing the behavior of the detector in the iterative process, is obtained calculating the MI $I_{\text{tot}} = I_{\text{tot}}(I^a)$ as a function of a-priori MI $I^a$ defined between the a priori reliability metrics $L$ and the corresponding bits $b$. This is done as described in [17]: the metrics $L$ are assumed to have Gaussian distribution with variance $\sigma_L^2$ and the mean $(b - \frac{1}{2}) \cdot \sigma_L^2$ conditioned on the bit’s value $b$. In other words, $L$ is treated as the outcome of a binary phase-shift keying (BPSK) transmission over AWGN channel with SNR defined as $SNR = \frac{1}{2} \sigma_L^2$, thus the a priori MI is given by

$$I^a = I_{\text{BPSK}} \left( \frac{1}{4} \sigma_L^2 \right).$$

(21)

As conjectured by [18] the transmission rate for the receivers employing the iterative processing is bounded by the area below the EXIT function $I_{\text{tot}}(I^a)$, i.e.

$$\hat{I}_{\text{tot}} = \int_0^1 I_{\text{tot}}(I^a) dI^a.$$  

(22)

Caution must be taken when using this interpretation of the area below the EXIT function as it is based on the assumption that the decoder’s EXIT function is “matched” to the detector’s function. That is, that both function do not cross each-other and the area between them is negligibly small. In practice, the decoder is channel-independent so some loss with respect to the predicted rate should be expected.

We note that, in general, the EXIT function may be obtained via numerical simulations, but here, we are able to deduce it knowing the relationship between $I^a$ and $\text{Var}\{s_n\}$. It can be found via numerical integration as shown in [19] and in a particular case of $X=\text{BPSK}$ we will consider here, the average variance of the symbols is given by

$$\text{Var}\{s\} = 1 - \int_{-\infty}^{\infty} \Phi(\lambda; \sigma_L^2) \cdot \tanh^2(0.5 \cdot \lambda) d\lambda$$  

(23)

where $\Phi(\lambda; \sigma_L^2) = \frac{1}{\sqrt{2\pi}\sigma_L} \exp \left( -\left( \lambda - 0.5 \cdot \sigma_L^2 \right)^2/(2\sigma_L^2) \right)$.

Thus, the EXIT function may be obtained efficiently (without Monte-Carlo simulations) if we use (21) to relate $\sigma_L^2$ with $I^a$, apply (23), replace $\text{Var}\{s_n\}$ with $\text{Var}\{s\}$ in (20), and use $\gamma_n$ instead of $\gamma_n'$ in (16).

Figure 5 shows the EXIT function $I_{\text{tot}}(I^a)$ for a particular channel realization and different values of $\Delta$, $\delta = \frac{1}{2} \Delta$, and $SNR = \frac{1}{N_0}$. The area (22) is then easily calculated. Note that the EXIT function, $I_{\text{tot}}(0)$ corresponds to $I_{\text{tot}}$ with “one-shot” detection, that is, when no iterative IC is employed. It is thus easy to observe that the improvement due to iterative IC when compared to non-iterative approach will be particularly important for large values of $\Delta$ (as expected because in such a case the interference is significant).

All the results are gathered in Fig. 6 where the outage rates $I_{\text{out},0.99}$ and $I_{\text{out},0.99}$ are shown for the CFO compensation explained in Sec. III-B, with and without IC explained in Sec. III-C. For comparison, we show the results of the single-relay transmission and the relay-selection results. The latter is feasible only if the receiver deciding which relay-transmission provides the largest rate, is capable of feeding this information back to the relays. We may appreciate that the optimal selection of $\delta$ provides only a slight rate increase when comparing to that obtained setting $\delta = \frac{1}{2} \Delta$; this frees us from from doing this optimization at the receiver. On the other hand, the iterative IC may provide significant gains which, as expected, are particularly important with growing value of $\Delta$. We note also that the achievable rate for $\Delta = 0.2$ is practically the same as in the case of relay-selection, independently whether IC is used or not. This is because with Alamouti SFC scheme, unlike in the case of relay-selection, we take also advantage of the frequency diversity, that is, amplitude-response of $H_1$ and $H_2$ is taken into consideration. As we saw on the example shown in Fig. 3, this may make the Alamouti scheme perform better than any of the relays taken individually.
Δ iteration for two different values of four OFDM blocks within which the channel is kept constant. From Fig. 6 we may read $I$ is used which corresponds to the target transmission rate of 4-QAM with and without interference cancellation at the receiver.

Fig. 7. Bit- and block error rates (BER and BLER) versus transmission SNR for 4-QAM with and without interference cancellation at the receiver.

IV. PRACTICAL CODED-MODULATION SCHEME

Finally, to illustrate the analysis we show an example of the coded transmission using 1360 information bits encoded with rate $\rho = \frac{3}{7}$ parallel concatenated convolutional codes (PCCC), i.e., turbo code, consisting of two identical component convolutional codes with generator $\{7, 5\}_8$. 4QAM modulation is used which corresponds to the target transmission rate of $I = 2\rho = 1.33$. The coded bits are modulated and sent over four OFDM blocks within which the channel is kept constant.

In Fig. 7, we show the bit error rate (BER) at the third iteration for two different values of $\Delta$ with and without the interference cancellation (IC). From Fig. 6 we may read that when $\Delta = 0.6$, the transmission rate $I = 1.33$ is not achievable with “one-shot” receiver but may be achieved when iterative IC is used. This is reflected in the simulation results as the BER curves tends to show the error-floor with “one-shot” processing. On the other hand, iterative IC, yields the regular decrease of the BER/BLER when SNR increases.

According to Fig. 6, when $\Delta = 0.4$ the rate $I = 1.33$ may be achieved regardless of the processing employed but IC provides the SNR gain of roughly 2dB. A similar SNR-improvement is observed in the simulations.

V. CONCLUSION

In this paper, we analyzed the impact of carrier frequency offset (CFO) on the performance of orthogonal frequency division multiplexing (OFDM) transmission employing space-frequency coding over relay channels. From the point of view of achievable transmission rates, we evaluated the CFO correction schemes and assessed the gains obtained when the received is dotted with the interference cancellation (IC) based on the turbo-principle. For the two-tap model of the transmission channel we evaluate the outage of the two-relay Alamouti SFC scheme and compared it to a single-relay transmission. The results indicate that a) the channel-dependent CFO correction provides only marginal performance gain, b) despite severe CFO the relay-based transmission may provide significant diversity gains over one-relay transmission, and c) the performance may be significantly increased via iterative interference cancellation. The conclusions drawn from the proposed analysis were illustrated with numerical simulations.

REFERENCES