Frequency Reversal Alamouti Code-Based FBMC With Resilience to Inter-Antenna Frequency Offsets

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Abstract—Transmit diversity schemes for filter bank multicarrier (FBMC) are known to be challenging. No existing schemes have considered the presence of inter-antenna frequency offset (IAFO), which will result in performance degradation. In this letter, a new transmit scheme based on the frequency reversal Alamouti code (FRAC)-based structure to address the issue of IAFO is proposed and is proven to inherently cancel the inter-antenna inter-carrier interference (ICI) while preserving spatial diversity. Moreover, the proposed FRAC structure is applicable in frequency-selective channels. Numerical results show that the proposed scheme undergoes negligible bit error rate (BER) degradation even with considerable IAFOs.

Index Terms—FBMC (filter bank multicarrier), Alamouti code, IAFO (inter-antenna frequency offset), transmit diversity.

I. INTRODUCTION

FILTER bank multicarrier (FBMC) is a promising candidate modulation of beyond-5G systems [1]. However, the implementation of the MIMO technique on filter bank multicarrier (FBMC) is always a challenging issue due to the presence of interference terms, which do not exist in some other multicarrier schemes. For example, in [2], while a low complexity channel estimation scheme was proposed for MIMO-FBMC, the error floor caused by the intrinsic inter-antenna interference of MIMO-FBMC was not resolved. In addition, MIMO techniques that exploit transmit diversity, such as Alamouti codes and more generally space-time block codes (STBC), have also been considered difficult for MIMO-FBMC systems. To overcome this obstacle, two different kinds of approaches have been studied in the literature. The first approach, dedicated to FBMC/OQAM, is to apply blockwise Alamouti coding in the time domain (e.g., [3], [4], [5]) or in the frequency domain [6], which has shown to inherently cancel intrinsic inter-antenna ICI in the Alamouti combining stage. The second approach, using QAM-based FBMC system (e.g., [7]), basically gets rid of intrinsic ICI even before the Alamouti combining stage and is therefore free of inter-antenna ICI.

However, in some scenarios of cell systems, e.g., distributed antenna systems (DAS) [8], and coordinated multipoint (CoMP) [9], the geographically distributed transmit (TX) antennas have different frequency offsets. This is mainly because the individual TX antennas experience different RF chains with different local oscillators that cannot be synchronized by a phase-locked loop (PLL) given the distance [10], [11], [12], [13]. Moreover, different Doppler frequencies will occur due to different signal propagation paths. As a result, even when the receiver is able to recover the carriers from each of the transmitters (e.g., through a PLL), the inter-antenna frequency offset (IAFO) is inevitable [10], [11]. Unfortunately, both of the aforementioned existing approaches [3], [4], [5], [6], [7] do not consider the IAFO and therefore cannot overcome the inter-antenna ICI in the presence of IAFOs. This leads to severe bit error rate (BER) performance degradation. In [14], many types of interference (IBI, ISI, and ICI) in FBMC/OQAM systems have been studied and their effects were thoroughly analyzed. However, the interference caused by IAFO was not studied.

In this letter, considering a situation where the signals from the two transmitters have different frequency offsets, we adopt the frequency reversal Alamouti code-based-FBMC (FRAC-FBMC) scheme proposed in [6] and [15] and extend it to multiple subblocks. The contribution of this letter is three-fold:

- A new FRAC-FBMC transmit and processing method is proposed specifically for tackling the effect of IAFO. The multi-subblock design makes the scheme applicable even in frequency-selective channels.
- The proposed FRAC-FBMC is shown, through rigorous mathematical derivations, to cancel the inter-antenna ICI caused by IAFO, without any cancellation algorithm.
- Intensive comparisons with the state-of-the-art Alamouti-coded FBMCs are provided to show the superiority of the proposed FRAC-FBMC.

II. IAFO-FREE FRAC-FBMC

A. Transmit Signal Structure

We consider an $N$-subcarrier FBMC-OQAM system with two transmit antennas and one receive antenna. Assume the two antennas, namely antenna $A$ and antenna $B$, transmit the
FBMC-OQAM-modulated signals as

\[ s_a(t) = \sum_{l=1}^{N} \sum_{m=-\infty}^{\infty} a_{l,m} \epsilon^{(a)}_{l,m}(t) \left( t - \frac{mT}{2} \right) e^{j2\pi t}, \]

\[ s_b(t) = \sum_{l=1}^{N} \sum_{m=-\infty}^{\infty} b_{l,m} \epsilon^{(b)}_{l,m}(t) \left( t - \frac{mT}{2} \right) e^{j2\pi t}, \]

where \( T \) is the FBMC symbol duration, \( l \) is the subcarrier index, \( m \) is the half-symbol (time) index, and \( p(t) \) is the pulse shaping prototype filter. The information-bearing symbols, denoted \( a_{l,m} \) and \( b_{l,m} \) for antennas \( A \) and \( B \), respectively, are real-valued. The terms \( \epsilon^{(a)}_{l,m} \) and \( \epsilon^{(b)}_{l,m} \) are phase shift terms whose properties will be described later in this subsection.

The \( N \) subcarriers are partitioned into \( K \) subblocks of size \( NF_t \), i.e., \( K = N / NF_t \). The \( \nu \)th subblock is indexed from \( iNF_t \) to \( (i+1)NF_t \) for all \( i = 0, 1, \ldots, K-1 \). The \( l \)th subblock in the \( n \)th FBMC subsymbol, we divide it into two halves and exploit \( L_n \) null subcarriers to the left of each half-subblock, i.e., \( a_{k+iNF_t,n} = b_{k+iNF_t,n} = 0 \) for all \( k = 1, \ldots, L_n \) and \( k = NF_t/2 + 1, \ldots, NF_t/2 + L_n \). The rest of the subcarriers in the subblock are encoded by \( NF_t - 2L_n \) real-valued information-bearing symbols \( \epsilon^{(i)}_{k,n} \) and \( \epsilon^{(i)}_{k,n} \), \( k = L_n + 1, \ldots, NF_t/2 \), according to the following frequency-reversal Alamouti coding:

\[
\begin{bmatrix}
a_{k+iNF_t,n} & b_{k+iNF_t,n} \\
b_{k+iNF_t,n} & a_{k+iNF_t,n}
\end{bmatrix}_{k+N} = \begin{bmatrix}
\epsilon^{(i)}_{k,n} - \epsilon^{(i)}_{k,n} \\
\epsilon^{(i)}_{k,n} - \epsilon^{(i)}_{k,n}
\end{bmatrix}
\]

for all \( i = 0, \ldots, K-1 \). Fig. 1 illustrates the idea of the FRAC symbol mapping described above for the special case \( L_n = 1 \). The phase shift terms \( \epsilon^{(a)}_{l,m} \) and \( \epsilon^{(b)}_{l,m} \) alternate between 1 (or -1) and \( f \) (or \(-f\)) in both of time and frequency axes, and follow the rules similar to those of [15, eqs. (25) and (26)], i.e., \( \epsilon^{(a)}_{l,m} = (l+iNF_t, m) = \chi^{(a)}_{l+iNF_t, m} \) and \( \epsilon^{(b)}_{l,m} = (l+iNF_t, m) = \chi^{(b)}_{l+iNF_t, m} \) for all \( l, m \), \( 1 + L_n \leq l \leq NF_t/2 \) and \( \chi = \pm f \) (or \( \pm f \)), where the superscript \( (\cdot)^* \) denotes the complex conjugate.

**B. Received Signal and Demodulation**

We assume there exist carrier frequency offsets among the two transmitters and the receiver. Denote \( f_a \) and \( f_b \) as the carrier frequency offsets of transmitters \( A \) and \( B \), respectively, relative to the carrier frequency of the receiver. Let the transmit signals from transmitters \( A \) and \( B \) pass through frequency-selective channels \( h_a(t) \) and \( h_b(t) \), respectively. Then, the received signal [11] defined as

\[ r(t) = e^{j2\pi f_a t} \int_{0}^{\Delta} h_a(\tau)s_a(t-\tau)d\tau + e^{j2\pi f_b t} \int_{0}^{\Delta} h_b(\tau)s_b(t-\tau)d\tau + n(t), \]

where \( \Delta \) is the maximum delay spread of the channels, and \( n(t) \) is an additive complex white Gaussian noise with power spectral density \( N_0/2 \).

We assume that \( f_a, f_b, h_a(t), \) and \( h_b(t) \) are known to the receiver. Next, we consider the \( k \)th subcarrier of the \( q \)th subblock in the \( n \)th FBMC subsymbol, with the received samples \( r_{k,n}^{(a,q)}, r_{k,n}^{(b,q)} \) being calculated by FBMC demodulation with respect to frequency offsets \( f_a, f_b, \) and phase shift terms \( \epsilon^{(a)}_{l,m}, \epsilon^{(b)}_{l,m} \), respectively:

\[ r_{k,n}^{(a,q)} = \int_{-\infty}^{\infty} e^{-j2\pi f_a t} r(t) e^{j2\pi f_a t} dt \]

\[ r_{k,n}^{(b,q)} = \int_{-\infty}^{\infty} e^{-j2\pi f_b t} r(t) e^{j2\pi f_b t} dt \]

for \( k = 1, 2, \ldots, NF_t, q = 0, 1, \ldots, K-1 \). We define \( H_{a,k}^{(q,k)} \) (or \( H_{b,k}^{(q,k)} \)) as the channel response of \( h_a(t) \) (or \( h_b(t) \)) at the \( k \)th subcarrier of the \( q \)th subblock, i.e.,

\[ H_{a,k}^{(q,k)} = \int_{0}^{\Delta} h_a(t) e^{-j2\pi(k+qNF_t)/T} dt \]

where \( * \) can be \( a \) or \( b \). We assume that \( \Delta \) is sufficiently small so that a quasi-flat fading assumption holds, i.e., \( H_{a,k}^{(q,k)} \) and
\(H_b(q, k)\) can be regarded as constant within the \(q\)th subblock. In other words, the subblock size \(N_F\) is chosen so that it spans a frequency range that is smaller than the channel coherence bandwidth (\(\approx 1/\Delta f\)) [16]. Therefore, we define \(H_b(q) \approx H_b(q, k)\) and \(H_b(q) \approx H_b(q, k)\) next, the decision variables for data symbols \(x_{k,n}\) and \(y_{k,n}\) with \(k = L_n + 1, L_n + 2, \ldots, N_F/2\), and \(q = 0, 1, \ldots, K - 1\), are obtained by combining the two received samples as follows:

\[
d_k(x, q) = H_a(x_a, q, k_n) + H_b(x_{N_F-1+1+L_n}, q, k_n),
\]

(8)

\[
d_k(y, q) = H_a(x_b, q, k_n) - H_a(x_{N_F-1+1+L_n}, q, k_n),
\]

(9)

where \(H[x]\) denotes the real part of \(x\). In the following developments, for simplicity, we only consider the 0th subblock (i.e., \(q = 0\)), noting that cases in other subblocks can be easily extended and will obtain identical results. In this regard, we let \(H_a = H_a(0), H_b = H_b(0), x_k = d_k(x, 0), d_k(y, 0), x_k = d_k(x, 0), y_k = d_k(y, 0), x_k = x_{k,n}, y_k = y_{k,n}, r_k = r_{k,n}\), and \(r_k = r_{k,n}\).

We will show that \(d_k(x, 0)\) and \(x_{k,n}\) or \(y_{k,n}\) have a relationship of diversity gain, i.e.,

\[
d_k(x, 0) = (|H_a|^2 + |H_b|^2)x_{k,n} + r_k(x),
\]

(10)

\[
d_k(y, 0) = (|H_a|^2 + |H_b|^2)y_{k,n} + r_k(y),
\]

(11)

where operator \(\cdot\) denotes the absolute value. To see the validity of (10), we substitute (1) and (2) into (4) and then into (5).

After performing some variable substitutions, we write \(r_{k,n}\) as follows:

\[
r_{k,n} = H_a \sum_{l=-L_m-M}^{L_m-M} s_{l+k,m+n}^{(a)} F_{l,m,n}^{(0)} + H_b \sum_{l=-L_m-M}^{L_m-M} s_{l+k,m+n}^{(b)} F_{l,m,n}^{(b)},
\]

(12)

where \(F_{l,m,n}^{(c)}\) is defined as

\[
F_{l,m,n}^{(c)} = e^{j\pi(x+\Delta f)T} \int_{-\infty}^{\infty} p(t) e^{j2\pi xT} e^{j2\pi \Delta fT} dt,
\]

(13)

\(w_{k,n}\) is a zero-mean complex Gaussian noise term, and \(L\) and \(M\) are small positive integers determined by the localization of the pulse \(p(t)\) in the time-domain and frequency-domain, respectively [17]. Note that \(e^{j2\pi \Delta fT}\) with \(\Delta f = f_b - f_a\) represents the residual carrier term due to IAFO. We assume that \(L_n\) is chosen sufficiently large such that \(L_n \geq L\). Noting that \(r_{k,n}\) as in (6) can be treated similarly, we express \(x_{k,n}\) and \(y_{k,n}\) as follows:

\[
x_{k,n} = H_a \sum_{l=-L_m-M}^{L_m-M} s_{l+k,m+n}^{(a)} F_{l,m,n}^{(k,n)} + H_b \sum_{l=-L_m-M}^{L_m-M} s_{l+k,m+n}^{(b)} F_{l,m,n}^{(k,n)},
\]

(14)

\[
y_{k,n} = H_a \sum_{l=-L_m-M}^{L_m-M} s_{l+k,m+n}^{(a)} F_{l,m,n}^{(k,n)} + H_b \sum_{l=-L_m-M}^{L_m-M} s_{l+k,m+n}^{(b)} F_{l,m,n}^{(k,n)} + w_{k,n},
\]

where (14) and (15) contain the inter-antenna ICI caused by IAFO. Now, substituting (14) and (15) into (8), we have

\[
d_k(x, 0) = |H_a|^2 x_{k,n} + d_k(x, 0) + |H_b|^2 y_{k,n} + r_k(y),
\]

\[
n_k = |H_a|^2 + |H_b|^2 + d_k(x, 0) + |H_b|^2 + r_k(y).
\]

(16)

Due to the property of \(\delta(t)\) and the alternating phase shift terms, we can verify the following property:

\[
\delta(t) \int U(t) dt = \delta(d_t) U(t) = \delta(t) \int U(t) dt
\]

(17)

\[
U(t) = \int Z(t) dt = x_{k,n}.
\]

(18)

\[
W^* = -V
\]

(19)

\[
V^* = -U
\]

(20)

where in the third equality, we had used the fact that \(L_n \geq L\) and \(a_{N_F-k+1+L_n,n} = -b_{k,n}\) for any \(k\) with \(1 - L_n \leq k \leq N_F/2 - L\), which is justified by (3). Finally, by substituting (22) and (23) into (20), the achieved diversity gain without inter-antenna interference as shown in (10) is verified. Since (11) can also be verified using a similar approach, the inter-antenna ICI-free characteristics of the proposed scheme are proven.
In the simulation of this letter, we use the PHYDYAS prototype filter in [17] with an overlapping factor $\lambda = 4$, and the corresponding localization factors $L = 1$ and $M = 4$. Some more parameters for the simulations are summarized in Table I.

### III. Simulation Result

In the simulation of this letter, we use the PHYDYAS prototype filter in [17] with an overlapping factor $\lambda = 4$, and the corresponding localization factors $L = 1$ and $M = 4$. Some more parameters for the simulations are summarized in Table I.

#### A. BER Versus Different IAFOs and Half-Subblock Size

In Figs. 2(a)–2(d), we investigate the dependency of the proposed scheme on IAFO and simulate its BER performance according to different half-subblock sizes $N_F/2$ and different IAFOs under flat fading channel, ITU-R-Pedestrian A (ITU-R-PA), ITU-R-Vehicular A (ITU-R-VA), and ITU-R-Pedestrian B (ITU-R-PB) multipath channels [18]. In this section, we set $N = 256$, $L_n = 1$, and the subcarrier spacing $\epsilon (= 1/T)$ is set to 15 kHz. At each simulated BER point, a total of $4 \times 10^4$ Monte Carlo trials have been performed. Figs. 2(a)–2(d) reveal that even with nonzero IAFOs, the proposed scheme maintains the identical BER to that with zero IAFO. Fig. 2 corresponds to the case of flat fading, with negligible change in the BER curves along the $N_F/2$-axis have negligible change. On the other hand, in Figs. 2(b)–2(d), with the channel selectivity increasing, choosing a large half-subblock size tends to violate the quasi-flat fading assumption. Thus, the inter-antenna ICI can not be canceled, resulting in the degradation of BER performance.

In practical applications, the choice of $N_F/2$ will be a compromise between bandwidth efficiency and BER degradation: choosing a large subblock size $N_F$ implies a saving in bandwidth efficiency but at the expense of a potential BER performance degradation, especially when the channel frequency selectivity is severe. Therefore, in Figs. 2(b) and 2(c), where the subcarrier spacing is $\epsilon = 15$ kHz, we recommend to choose $N_F/2$ to be 8 and 4, respectively.

#### B. Performance Comparison With the Other Schemes

In this subsection, we compare the impact of IAFO on various existing Alamouti-coded FBMC schemes. In Fig. 3(j), the BER curves for TR-FBMC in [3], FBMC-QAM in [7], and the proposed scheme, are plotted under flat channel (Fig. 3a-c), ITU-R-PA channel (Fig. 3d-f), and ITU-R-VA channel (Fig. 3g-i), respectively. The half-subblock size $N_F/2$ for the proposed scheme in Figs. 3(c), 3(f), and 3(i) are set to 128, 8, and 4, respectively. A complete set of simulation parameters is listed in Table II. It is clearly observed that under all channels, the two existing FBMC schemes suffer BER degradation when the IAFO increases from 0 to 0.3 subcarrier spacing, as shown in Figs. 3ab, 3de, and 3gh, respectively. This is because TR-FBMC and FBMC-QAM can not discard the interference of IAFO when we compensate for specific carrier frequency offset and thus it leads to the retention of inter-antenna ICI. However, as shown in Figs. 3c, 3f, and 3i, the proposed scheme only has negligible changes in BER with increasing IAFOs, demonstrating the proposed scheme’s unprecedented advantages in the presence of a high IAFO.

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### TABLE I

PARAMETERS FOR THE SIMULATION RESULTS OF FIG. 2

| Parameter                  | Value                                      |
|----------------------------|--------------------------------------------|
| Subcarriers ($N$)          | 256                                        |
| Prototype filter           | PHYDYAS prototype filter [17]              |
| Null subcarriers ($L_n$)   | 1                                          |
| Modulation                 | QPSK                                       |
| Channel models             | Flat fading, ITU-R-PA, ITU-R-VA, ITU-R-PB  |
| Subcarrier spacing ($\epsilon$) | 15 kHz                                    |
| SNR                        | $\{0: 2: 20\}$ dB                        |
| Inter-antenna frequency offset (IAFO) | $(0,0.01,0.1,0.2,0.3)\epsilon$   |
| Monte Carlo trials         | 40000                                      |
| Symbols per Monte Carlo trials | 40                                      |

### TABLE II

PARAMETERS FOR THE SIMULATION RESULTS OF FIG. 3

| Parameter                  | Proposed Method | TR-FBMC [3] | FBMC-QAM [7] |
|----------------------------|-----------------|-------------|--------------|
| Subcarriers ($N_F$)        | 256             |             |              |
| Prototype filter           | PHYDYAS prototype filter [17] |             |              |
| Modulation                 | QPSK            | QPSK        | QPSK         |
| Channel models             | Flat fading, ITU-R-PA, ITU-R-VA         |             |              |
| Subcarrier spacing ($\epsilon$) | 15 kHz         |             |              |
| SNR                        | $\{0: 2: 20\}$ dB |             |              |
| Inter-antenna frequency offset (IAFO) | $(0,0.01,0.1,0.2,0.3)\epsilon$ |             |              |
| Half-subblock size ($N_F/2$) | 128 (Fig. 3c), 8 (Fig. 3d), 4 (Fig. 3i) |     | N/A          |
| Monte Carlo trials         | 40000           |             | N/A          |
| Symbols per Monte Carlo trials | 40            |             |              |
IV. CONCLUSION

In this letter, a new FRAC-FBMC scheme that has inter-antenna ICI-free characteristics even with nonzero IAFOs was proposed. By allowing the use of multiple subblocks, the proposed scheme works well even in frequency-selective channels, with only a slight sacrifice in bandwidth efficiency. Simulation results confirmed the resilience of the proposed scheme to the IAFO and showed that even if the IAFO is as high as 0.3 times subcarrier spacing, the proposed scheme outperforms other schemes for any frequency-selective channels.

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