1. Introduction

Wireless data traffic has grown significantly since 3G communication networks were widely deployed. The growth rate is expected to be even higher when the 4G technologies become more mature, more data hungry wireless services become available, and affordable smart phones loaded with data demanding applications start to deeply penetrate the handset market [1–3]. It is clear that exclusive usage of the current commercial spectrum will not be sufficient to support such a growth. The spectrum scarcity will become a main roadblock for the expansion of the overall wireless industry. To meet the future market demand, regulators and government agencies are actively seeking for new spectra and, at the same time, strongly encouraging the industry and academia to develop technologies that support efficient spectrum sharing [4–6].

Spectrum sharing can be characterized by the level of coordination in different layers of the system, and can be achieved in multiple dimensions such as frequency, time, space, code, and power [7]. Some of the legacy multi-access technologies used in current systems that rely on accurate synchronization may not be directly applicable in future air interface designs. To fully utilize the potential benefits of spectrum sharing, a new set of enabling technologies for channel access, resource management and spectrally agile air interfaces need to be developed. Cognitive radio (CR) has been developed as such a technology to achieve these goals [8].

The general framework of the CR technologies has been developed over the past fifteen years [9,10]. Some of the CR concepts have been utilized in wireless communication standards that use TV white space as the transmission spectrum [11]. Still, a few fundamental technology areas are yet to be developed so that truly cognitive systems...
can be standardized and commercialized. One such area is selection of the baseband waveform that would enable opportunistic transmission on any available spectrum fragments. This requires dynamic spectrum sharing at the level of almost arbitrary spectral granularity. To achieve this goal, in this paper, we focus on waveforms with multicarrier modulation (MCM).

MCM techniques enable simultaneous transmissions of a set of data over multiple narrow band subcarriers. It can be shown that MCM combined with an advanced modulation and coding scheme (MCS) can achieve very high spectral efficiency in frequency selective channels [12]. As one of the simplest MCM schemes, Orthogonal Frequency Division Multiplexing (OFDM) has been widely used in modern wireless communication systems with tightly controlled synchronization. Orthogonal Frequency Division Multiple Access (OFDMA) based on OFDM has been successfully used in LTE systems. However, OFDMA may not be suitable for CR systems due to its high out-of-band emission (OOBE) originated from large side lobes of the rectangular pulse shape. In CR systems, different transmitter–receiver links between primary and secondary users or among secondary users may not be synchronized. In this case, high OOBE of OFDMA results in significant inter-carrier interference (ICI) at victim users, especially when nonlinear power amplifiers (PA) are used at transmitters. Thus, victim users may suffer from significant adjacent channel interference (ACI), if their receivers do not have proper ACI rejection capability. Some of the common methods considered to reduce ACI are lowering the transmit power and creating large enough guard bands. Clearly, the former limits the coverage area and the latter leads to low spectral utilization. Other methods without sacrificing the coverage and spectral utilization would require advanced signal processing technologies.

Filter bank multicarrier (FBMC) modulation is a family of MCM techniques proposed as an alternative to OFDM to reduce ACI. OFDM-Offset QAM (OFDM-OQAM) is a popular implementation of FBMC schemes [13–20]. In OFDM-OQAM, adjacent subcarriers of the signal overlap to achieve a high spectral efficiency. Different from OFDM, the real and imaginary parts of the QAM symbols are processed separately with 2× symbol rate. Besides OFDM-OQAM, Filtered Multitone (FMT) [21,22], Cosine Modulated Filter Bank (CMFB) [23,24], and Exponentially Modulated Filter Bank (EMFB) [25], are also some of the commonly considered FBMC schemes in literature. In FBMC, a prototype filter needs to be carefully designed to minimize or zero out inter-symbol interference (ISI) and ICI while keeping the side lobes small. Even though FBMC schemes have much better spectral containment, their implementation in practical systems poses several challenges such as high complexity and latency. Also, since FBMC schemes do not use a cyclic prefix (CP), they require more complex equalization algorithms than OFDM, especially in doubly dispersive channels [26].

A more recently proposed MCM scheme with reduced OOBE is the generalized frequency division multiplexing (GFDM) [27,28] in which the signal transmitted on separate frequency chunks are filtered individually and summed before transmission. GFDM has a block based structure; a block consists of several symbols and a CP is added to a block instead of to each symbol as in OFDM. Tail biting is used at the transmitter to reduce the required CP length and convert the filtering operation into circular convolution to achieve a low-complexity transceiver. To fully utilize the advantages of OFDM with CP in terms of robustness to multipath fading and implementation simplicity, several other OFDM-based techniques with lower OOBE have also been developed [29]. Among them, a simple and commonly used technique is to filter the OFDM signal, and is referred to as Filtered-OFDM. However, when available spectrum fragments are not contiguous, filtering becomes challenging since a separate filter needs to be dynamically designed and used for each fragment.

Combining the benefits of FBMC and Filtered-OFDM, in this paper, we propose a new MCM transceiver technique. In this technique, the available spectrum fragments are divided into chunks of contiguous subcarriers, referred to as resource blocks (RB). And the signal transmitted on each RB is generated and filtered individually. The proposed MCM is called RB Filtered-OFDM (RB–F–OFDM). It has the advantage of being modular and scalable since the same transmit (or receive) module is used for all RBs. In addition, some of the peak-to-average power ratio (PAPR) reduction techniques [30–32] can be naturally adapted to the RB–F–OFDM design. The above features are very important for efficient implementation of the signal processing blocks in hardware and modern power consumption.

We show that the complexity of the proposed transceiver can be reduced by utilizing the polyphase implementation that is also used in FMT. Moreover, orthogonality is still maintained between subcarriers, and RB–F–OFDM modulates each data symbol in the same time and same subcarrier as in OFDM. Therefore, one can also show that the RB–F–OFDM modulated signal can be directly demodulated by the legacy OFDM receiver, and that its receiver could also demodulate legacy OFDM signals. This feature makes the system based on RB–F–OFDM backwards compatible with the legacy OFDM systems.

It should be noted that although RB–F–OFDM is similar to GFDM in the sense that both schemes divide the available bandwidth into frequency chunks and filter each chunk individually, there are fundamental differences between the two. One important difference is that orthogonality between subchannels is not maintained in GFDM and a more complex receiver is required to cancel the created self-interference, and as such, it is not backwards compatible to legacy OFDM systems. In addition, due to the block based structure and tail biting, GFDM is suitable for packet based wireless systems with low mobility. Finally, tail biting causes a significant increase in the OOBE since the filtered signal is distorted. Padding zeros to end of a source block may be used to reduce the increased OOBE at the expense of a spectral efficiency loss.

Note that our RB–F–OFDM idea was first presented in [42]. Much more details in the RB–F–OFDM design, especially in the aspects of complexity reduction and PAPR reduction, are described in this paper. This paper is organized as follows: In Section 2, we introduce the system model. Section 3 presents a general introduction of the proposed RB–F–OFDM scheme with PAPR reduction. An efficient polyphase implementation of the RB–F–OFDM is also presented in this section. In Section 4, we present a detailed evaluation of the performance of RB–F–OFDM. Specifically,
the proposed scheme’s spectral containment, bit error rate (BER) and PAPR performance is evaluated and complexity and latency analysis is presented. Finally, Section 5 draws the conclusions.

**Notations.** \(0 \times a \times b\) signifies an all zero matrix of dimension \(a \times b\). \(\max(a[n])\) and \(E(a[n])\) stand for the maximum and expectation of a sequence \(a[n]\), respectively. \([a]\) and \(\lfloor a\rfloor\) stand for the maximum integer not greater than real value \(a\) and the minimum integer not smaller than real value \(a\), respectively. \(\text{lcm}(a, b)\) stands for the least common multiple of two integers \(a\) and \(b\). \(a \mod b\) stands for the remainder of the Euclidean division of \(a\) by \(b\). \(\delta[n]\) is the Dirac delta function.

### 2. System model

In an MCM system, usually a group of contiguous subcarriers is assumed to be the smallest dynamically available transmission band (the smallest granularity of frequency resource). All the assigned transmission bands or opportunistically detected frequency bands are consisting of one or multiple of such a group of contiguous subcarriers. We call such a group of contiguous subcarriers as resource block (RB). Therefore, the entire spectrum that a system or device detects or utilizes could be considered uniformly divided into RBs. As depicted in Fig. 1, a non-contiguous spectrum may be divided into RBs of the same size.

Assume each RB is consisting of \(D\) subcarriers. The RBs are label the 0th, 1st, 2nd, ..., \(k\)th, ..., \((K - 1)\)th RBs, where \(K\) is the maximum number of available RBs. The normalized center frequency of the \(k\)th RB is \(f_k = m_k/L\), where \(m_k\) is the “center subcarrier index” of the \(k\)th RB and is not necessary an integer. \(L\) is a power of 2, and is related to the sampling rate of the transmit signal \(f_s = L/T\), where \(T\) is the symbol duration. Define the \(n\)th data symbol vector (an \(M \times 1\) vector) as

\[
\mathbf{S}[n] = [S_0[n] \ S_1[n] \ \cdots \ S_K[n]]^T
\]

(1)

where the \(n\)th data symbol vector for the \(k\)th RB is a \(D \times 1\) vector

\[
\mathbf{S}_k[n] = [S_{k0}[n] \ S_{k1}[n] \ \cdots \ S_{kD-1}[n]]^T.
\]

(2)

When the \(k\)th RB is available for transmission and has data loaded, \(\mathbf{S}_k[n] \neq 0_{D \times 1}\). Otherwise, \(\mathbf{S}_k[n] = 0_{D \times 1}\). In this paper, we assume that through perfect feedback in control channel, both the transmitter and receiver have the knowledge of the available RB list assigned or detected. In practice, each RB can have its own properties, such as power loading, MCS, and hybrid automatic repeat request (HARQ).

### 3. Resource block Filtered-OFDM (RB-F-OFDM)

In this section, we present our RB-F-OFDM design for spectrally agile and power efficient systems.

#### 3.1. General description

The design criteria of RB-F-OFDM are as follows: Firstly, it provides a spectral containment improvement of OFDM and Filtered-OFDM for contiguous and non-contiguous spectrums. Secondly, its transmit signal is compatible with the existing OFDM receiver in current standards and systems. And the RB-F-OFDM receiver could demodulate OFDM signals in current systems. Lastly, the RB-F-OFDM transceiver enables phase rotation based PAPR reduction techniques to improve the power efficiency of the system. The block diagrams of the proposed RB-F-OFDM transmitter and receiver are illustrated in Figs. 2 and 3, respectively.

The main idea of the RB-F-OFDM is to generate the spectrally contained per-RB filtered CP-OFDM signal independently and obtained the transmit signal as their phase rotated sum. The phase rotation enables PAPR reduction.

---

**Fig. 1.** Non-contiguous spectrum divided into RBs.

**Fig. 2.** RB-F-OFDM transmitter block diagram.
To lower the computational complexity, the per-RB CP-OFDM signal is generated in low rate and upscaled. The per-RB transmit filter with good frequency selectivity not only eliminates the spectral images due to upsampling but also provides good spectral containment to the per-RB signal. The detailed ideas of RB-F-OFDM are as follows:

Firstly, in the RB-F-OFDM transmitter ("RB-F-OFDM Tx") in Fig. 2, the data of each RB, $S_k[n]$, $k = 0, 1, \ldots, K-1$, is generated independently to form a per-RB modulated signal $x_k[m]$. The transmit signal is the phase rotated sum of all the per-RB signals. It is given as

$$x[m] = \sum_{k=0}^{K-1} e^{j\phi_k} x_k[m].$$

(3)

In this way, the transceiver structure could enable phase rotation based PAPR reduction techniques, such as partial transmit sequences (PTS) [30,31] and selective mapping (SLM) [32]. The signal of each RB (resp., the signal of groups of RBs), instead of the signal of each subcarrier (resp. the signal of groups of subcarriers), are dealt with in SLM (resp. PTS). At the transmitter, by properly optimizing the phase rotations, $\phi_k$, $k = 0, 1, \ldots, K - 1$, the PAPR of the transmit signal could be minimized. At the receiver, the complex conjugates of the phase rotations $e^{-j\phi_k}$, $k = 0, 1, \ldots, K - 1$ are applied. If PAPR reduction is disabled, $e^{j\phi_k} = 1, k = 0, 1, \ldots, K - 1$. The details in PAPR reduction will be described in Section 3.3, and for simplicity, ignored in the following discussions.

Secondly, module based direct implementations could be used at transmitter and receiver. At the transmitter, a per-RB transmit module (the red shaded block in Fig. 2) could be used to modulate $S_k[n]$ (the nth data symbol vector in the kth RB) into $x_k[m]$ (the modulated signal of the kth RB) for all RBs. In the per-RB transmit module, $S_k[n]$ is first modulated as a CP-OFDM signal of possibly lower sampling rate $f_{s1} = N/T$, where $N$ is a power of 2, and $N \leq L$. This is done as in the per-RB CP-OFDM transmit module in Fig. 4 (i.e., the “OFDM Tx” block in Fig. 2) by using an N-point inverse fast Fourier transform (IFFT), CP insertion, and a parallel-to-serial (P/S) conversion. And then, as in Fig. 2, the low-rate CP-OFDM signal is upscaled by $Q = L/N$, and filtered by a lowpass per-RB transmit filter $p_T[n]$. After filtering, the RB modulation modulates the filtered signal into the frequency band of the kth RB according to the RB’s “central subcarrier index” $m_k$, to form the per-RB signal

$$x_k[n] = \left( \sum_m b_k[m] p_T[n - mQ] \right) e^{j2\pi m_k t_k}. $$

(4)

On the other hand, at the RB-F-OFDM receiver ("RB-F-OFDM Rx"), a per-RB Filtered-OFDM receive module (the red shaded block in Fig. 3) has “reverse” operations of the per-RB Filtered-OFDM transmit module. It could be used in a module based direct implementation to demodulate $\hat{S}_k[n]$ (the channel and filter corrupted nth data symbol vector in the kth RB) from $y[n]$ (the received signal) for
all RBs. In the per-RB receive module, $y[n]$ is demodulated from frequency band of the $k$th RB to baseband according to the RB’s “central subcarrier index” $m_k$, to form an RB demodulated signal. Then, the RB demodulated signal goes through the per-RB receive filter $p_R[n]$ to reject out-of-RB signals. After filtering, the filtered signal is downsampled by $Q$. The resulting low-rate signal is demodulated as a CP-OFDM signal, to form the demodulated streams $\hat{S}_k[n]$. This is done in the per-RB CP-OFDM receive module in Fig. 4 (i.e., the “OFDM Rx” block in Fig. 3) by using a serial-to-parallel ($S/P$) conversion, CP removal, and $N$-point fast Fourier transform (FFT).

Thirdly, the per-RB transmit and receive filters are properly designed to achieve good spectral agility in both contiguous and non-contiguous spectrums. As in the power spectral density (PSD) plot in Fig. 5, a properly designed per-RB transmit filter should eliminate all spectral images due to the upsampling operation and minimize OOB. The signal overlap between RBs does not create ICI due to bi-orthogonality between subcarriers in CP-OFDM systems in different RBs. Therefore, the minimum requirement of $p_T[n]$ to maintain orthogonality between RBs is that its passband bandwidth is $D$ subcarriers, i.e., $BW_{pass} = D/T$. Each side of transition band has the bandwidth $BW_{trans} = (N-D)/T$, and the stopband has attenuation of at least 55 dB. In addition to this minimum requirement, we require that each per-RB modulated signal only has signal overlapping its adjacent RBs but not the RBs beyond its adjacent RBs, i.e., $x_k[m]$ only has overlap with $x_{k-1}[m]$ and $x_{k+1}[m]$ in the frequency domain. The signal leakage of a per-RB signal to its non-adjacent RBs is negligible. In this way, the RB-F-OFDM could achieve low OOB for both contiguous and non-contiguous spectrums. The per-RB receive filter is also crucial in the sense that it should have good frequency selectivity to reject out-of-RB signals. Therefore, the minimum requirement of $p_R[n]$ is that it is a lowpass filter with passband bandwidth of $D$ subcarriers, i.e., $BW_{pass} = D/T$. For simplicity, in this paper, we assume that the per-RB transmit and receive filters are identical, i.e., $p_T[n] = p_R[n] = p[n]$. The filter design is not the interest of this paper. We simply assume to use equal ripple filters with passband attenuation of 0.75 dB, stopband attenuation of 58 dB, and stopband slope of 20.

Lastly, the RB-F-OFDM transmitter essentially modulates each data symbol to the same subcarrier as it is modulated in CP-OFDM using an $L$-point IFFT, subject to a phase modulation. This phase modulation is due to the fact that CP insertion in RB-F-OFDM is before the RB modulation, while CP insertion in CP-OFDM is after the subcarrier modulation. The phase modulation in RB-F-OFDM for the $n$th data symbol in the $k$th RB is $e^{-j2\pi m_k L_{cp} n/L}$, where $L_{cp}$ is the number of CP samples in sampling rate $f_s$. A CP-OFDM receiver could be used to demodulate the signal. The channel estimation (CHEST) in CP-OFDM could be used to estimate the equivalent channel, including the transmit and receive filtering and multipath channel. And the 1-tap frequency domain equalizer (FDE) in CP-OFDM could be used to equalize the equivalent channel and perform phase demodulation (i.e., applying a multiplier of $e^{j2\pi m_k L_{cp} n/L}$ for the $n$th data symbols in the $k$th RB). Alternatively, the phase demodulation may be applied to the source symbols prior
to the RB-F-OFDM transmitter. Note that similarly, the RB-F-OFDM receiver with phase modulation, along with the CHEST and 1-tap FDE, as in Fig. 6, could demodulate OFDM signals in current systems as well. The realistic CHEST and more advanced equalizer designs are not of interest in this paper.

3.2. Efficient polyphase implementation

In the above direct implementation of RB-F-OFDM, the computational complexity could be very high since it scales by the number of available RBs. Note that the RB-F-OFDM transmitter could be represented as the per-RB CP-OFDM signals going through a synthesis filter bank (SFB) of the FMT modulation. On the other hand, the RB-F-OFDM receiver could be represented as that received signal passes through an analysis filter bank (AFB) of the FMT demodulation followed by the per-RB CP-OFDM receive modules. Note that the orthogonality between the subchannels in the SFB-AFB is maintained due to the bi-orthogonality in CP-OFDM systems.

In this section, we derive the efficient polyphase implementation of the RB-F-OFDM transmitter and receiver to further reduce the complexity when the number of available RBs is large. Note that in this section, it is assumed that PAPR reduction is disabled, i.e., the RB is large. Without loss of generality, let $m_k = kD$. And we further assume that $D$, the number of subcarriers per RB, is a power of 2, so that $C = L/D$ is a power of 2, and the RB modulation and demodulation multipliers become $e^{j2\pi kn/D} = e^{j2\pi k/C}$. In the next two sub-sections, we derived the efficient polyphase implementations of the SFB and AFB based on [33–35].

3.2.1. Synthesis filter bank (SFB)

The SFB is the component with inputs of $b_0[n], b_1[n], \ldots, b_{K-1}[n]$, and outputs of $x[n]$. Substituting (4) into (3), the RB-F-OFDM transmit signal could be expressed as

$$x[n] = \sum_{k=0}^{K-1} \left( \sum_{m} b_k[m] p[n - mQ] \right) e^{j2\pi k\frac{n}{C}}$$

which is exactly the expression of an FMT transmit signal, where $C \geq Q, C \geq K$, and $lcm(Q, C) = C$. After exchanging the order of summation, we obtain

$$x[n] = \sum_{m} \left( \sum_{k=0}^{K-1} b_k[m] e^{j2\pi k\frac{n}{C}} \right) p[n - mQ].$$

Express $n$ as

$$n = \left\lfloor \frac{n}{Q} \right\rfloor C + w, \quad w = 0, 1, \ldots, C - 1.$$  (7)

Substituting (7) into (6), we obtain

$$x[n] = \sum_{m} u_w[m] p[n - mQ]$$

where the signal

$$u_w[m] = \sum_{k=0}^{K-1} b_k[m] e^{j2\pi k\frac{m}{Q}}, \quad w = 0, 1, \ldots, C - 1$$

could be obtained through a $C$-point IFFT. Re-express $n$ as

$$n = \rho Q + v, \quad \rho = \left\lfloor \frac{n}{Q} \right\rfloor, \quad v = 0, 1, \ldots, Q - 1.$$  (10)

We define the $v$th polyphase component (with respect to $Q$) of filter $p[n]$ as

$$p_{Q,v}[n] = p[nQ + v], \quad v = 0, 1, \ldots, Q - 1,$$

and obtain $Q$ such polyphase filters of $p[n]$. Therefore, substituting (10) into (8), we obtain

$$x[n] = \sum_{m} u_w[m] p[(\rho - m)Q + v]$$

$$= u_w[n] \ast p_{Q,v}[n]$$

which is the convolution of the IFFT output sequence $u_w[n]$ and the $v$th polyphase filter $p_{Q,v}[n]$, and the values of $w$ and $v$ depend on $n$, and are from (7) and (10), respectively. Since $lcm(Q, C) = C$, we obtain

$$v = w \bmod Q, \quad v = 0, 1, \ldots, Q - 1.$$  (13)

Therefore, each IFFT output sequence $u_w[n]$ corresponds to a unique polyphase filter $p_{Q,v}[n]$. The value of the pair $(w, v)$ changes periodically with a period of $C$. Thus, a multiplexer (MUX) could be used to choose the samples from the polyphase filtered sequences, based on the sample index. The resulting polyphase implementation of the SFB and the overall RB-F-OFDM transmitter block diagram are depicted in Fig. 7.

3.2.2. Analysis filter bank (AFB)

Similar to the SFB, the polyphase implementation of the AFB could also be derived. Define the $C \times 1$ output vector from the AFB in the $m$th instance as

$$a[m] = \begin{bmatrix} a_0[m] & a_1[m] & \cdots & a_{K-1}[m] \end{bmatrix}^T a_k[m]$$

where $a_k[m]$ is the input to the per-RB CP-OFDM receive module in the $k$th RB. Note that only the first $K$ sequences are the useful outputs, and the last $C - K$ sequences are discarded. The input to the CP-OFDM receive module in the $k$th RB could be expressed as

$$a_k[m] = \left[ \left( y[n] e^{-j\frac{2\pi km}{C}} \right) \ast p[n] \right]_{n=mQ}$$

$$= \sum_{l} y[l] e^{-j\frac{2\pi kl}{C}} p[mQ - l].$$  (15)

Express $l$ as

$$l = \gamma C + z, \quad \gamma = \left\lfloor \frac{l}{C} \right\rfloor, \quad z = 0, 1, \ldots, C - 1.$$  (16)

Substituting (16) into (15), we obtain

$$a_k[m] = \sum_{z=0}^{C-1} y[\gamma C + z] e^{-j\frac{2\pi k}{C}} p[mQ - \gamma C - z].$$  (17)
which is exactly the expression of an FMT received signal at the kth subchannel. After exchanging the order of sum-
mation, we obtain
\[
a_k[\gamma] = \sum_{c=0}^{C-1} \left( \sum_{p=0}^{mQ} y[\gamma C + z] p[mQ - \gamma C - z] \right) e^{-j\frac{2\pi kq}{m}}
\]
(18)
where the signal is
\[
\hat{u}_k[m] = \sum_{\gamma} y[\gamma C + z] p[mQ - \gamma C - z]
\]
(19)
with the C-downsampled received signal
\[
y_{z}[\gamma] = y[\gamma C + z], \quad z = 0, 1, \ldots, C - 1.
\]
(20)

Therefore, the output vector from the AFB could be obtained through a C-point FFT, where each FFT input se-
quency \( \hat{u}_k[m] \) comes from the convolution of \( y_{z}[\gamma] \) with a possible time varying filter. The C-downsampled received signal \( y_{z}[\gamma] \) could be obtained by using an S/P conversion from \( y[n] \). Let
\[
mQ - z = \zeta C + c, \quad \zeta = \left\lceil \frac{mQ - z}{C} \right\rceil,
\]
(21)
c = (mQ - z) \mod C, \quad m \in Z.

Also, define the cth polyphase component (with respect to C) of filter \( p[n] \) as
\[
p_{c,c}[n] = p[nc + c],
\]
(22)
c = 0, 1, \ldots, C - 1, n = 0, 1, \ldots

and obtain C such polyphase filters of \( p[n] \). Therefore, substituting (21) and (22) into (19), after some math manipu-
lations, we obtain
\[
\hat{u}_k[m] = \sum_{\gamma} y_{z}[\gamma] p[(\zeta - \gamma) C + c]
\]
(23)
where the values of \( \zeta \) and \( c \) depend on \( m \), and are from (21).

Therefore, the polyphase filter \( p_{c,c}[\zeta] \) for the FFT input se-
quency \( \hat{u}_k[m] \) is a time-varying filter that changes period-
ically with a period of \( q = C/Q \). For notation simplicity, define the periodic time-varying filter for the FFT input se-
quency \( \hat{u}_k[m] \) as
\[
g_{z}[\beta][n] = p_{c, (\beta q - z) \mod C}[n], \quad \beta = m \mod q.
\]
(24)
The resulting polyphase implementation of the AFB and the overall RB-F-OFDM transmitter block diagram are depicted in Fig. 8. Each time-varying filter could be implemented using the tap-delayed line, filtering in each branch, and a MUX to choose the output signal, as in Fig. 9.

3.3. Peak-to-average power ratio (PAPR) reduction

PAPR is a measure of the envelope variation of a waveform and is the peak amplitude of the waveform divided by the root-mean-square value of the waveform. In general, MCM transmit signals have large PAPR, which requires PAs to have a very large linear range. Otherwise, the nonlinearity leads to signal distortion, which causes larger OOBE and BER. A lot of PAPR reduction methods in litera-
ture for MCM systems could be applied to the RB-F-OFDM.

In particular, the distortionless phase rotation based PAPR reduction techniques, e.g., PTS [30,31] and SLM [32] (a spe-
cial case of PTS when there is 1 RB/group), are natural choices for RB-F-OFDM, as depicted in Fig. 2. Note that if the efficient polyphase implementation of RB-F-OFDM trans-
mitter is used, the sum signal of each RB group could be generated using one efficient implementation of the RB-F-
OFDM transmitter.

In PTS, \( K \) per-RB modulated signals are divided into \( [K/n_g] \) groups, each having at most \( n_g \) RBs. The same phase rotation is used in each group, while the 0th group has phase rotation value of 1. The phase rotations of \( [K/n_g] \) groups form one phase vector. The side information of \( [K/n_g] - 1 \) phase rotations may be sent to the receiver. In this case, the resulting spectral efficiency loss depends on the number of groups, the nature of the phase vector (e.g., quantized or not), and the time interval of generating phase vectors. Alternatively, the phase rotation could be
Fig. 8. RB-F-OFDM transmitter block diagram using polyphase implementation of AFB.

Fig. 9. Time-varying filter block diagram at AFB.

Table 1
Modified LTE-like MCM systems.

<table>
<thead>
<tr>
<th>System parameters</th>
<th>CP parameters</th>
<th>Design parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>OFDM</td>
<td>Symbol duration: $T = 1/\Delta f$</td>
<td>One subframe: 2 slots/subframe</td>
</tr>
<tr>
<td></td>
<td>Subcarrier spacing: $\Delta f = 15$ kHz</td>
<td>One slot: 7 data symbols (7T) and half additional symbol duration (0.5T) used for CP, with CP duration $T_{cp1} = 80T / 1024$ (i.e., $L_{cp1} = 80L / 1024$) for the first symbol and $T_{cp2} = 72T / 1024$ (i.e., $L_{cp2} = 72L / 1024$) for the other symbols. Average CP length: $L_{cp} = L / 14$</td>
</tr>
<tr>
<td>Filtered-OFDM</td>
<td>Number of samples per symbol duration: $L = 1024$</td>
<td></td>
</tr>
<tr>
<td>RB-F-OFDM</td>
<td>Number of subcarrier per RB: $D = 12$</td>
<td>Per-RB FFT size and equal ripple filter length: $N = 128$ &amp; $L_p = 53$</td>
</tr>
<tr>
<td></td>
<td>Maximum number of active RBs: $K = 50$</td>
<td>Low-rate CP length: $[L_{cp}/Q]$</td>
</tr>
<tr>
<td></td>
<td>Maximum number of active subcarriers: $M = KD$</td>
<td></td>
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</tbody>
</table>

considered part of the equivalent multipath channel and detected in CHEST. The phase rotation is recovered at the receiver, possibly in the 1-tap FDE. Therefore, the computational complexity is only added to the transmitter. In this case, enough pilot symbols in each RB group are required to enable reliable CHEST. If the pilot density in each RB group is not high enough, the CHEST may rely on interpolation between consecutive RB groups. In such a case, a pilot based phase estimator can be utilized for each RB group. Thus, spectral efficiency loss may occur due to additional pilots. In practical systems, each phase vector is used for a few consecutive “OFDM” symbols. This results in an inherent delay in optimizing the phase vector.

The phase vector optimization has been studied extensively in literature [30,31,36–39]. We consider the simple random PPTS (RPTS) and quantized PPTS (QPTS). In RPTS, besides the all one vector, $n_r$ phase vectors are randomly generated and each phase value is uniformly distributed in $[0, 2\pi)$. The output signal for each phase vector is obtained, and the one achieves the lowest maximum PAPR value in a certain time interval is chosen. On the other hand, QPTS is similar to RPTS, except that each phase value is randomly chosen from a size-$n_{pts}$ candidate set $\{0, 2\pi/n_{pts}, 4\pi/n_{pts}, \ldots, 2\pi(n_{pts} - 1)/n_{pts}\}$. To reduce computational complexity in optimizing the phase vector, a PAPR threshold could be set to allow early termination of search when the PAPR threshold is achieved.

4. Performance evaluation

In this section, the performance of RB-F-OFDM is evaluated and compared with OFDM and Filtered-OFDM in terms of complexity, latency, spectral containment, PAPR, and BER. The performance comparison of MCMs is based on the modified LTE-like systems specified in Table 1, if not otherwise specified. Note that in Table 1, the CP lengths of OFDM, Filtered-OFDM, and RB-F-OFDM are the same. Therefore, they have the same spectral efficiency.

4.1. Computational complexity

In this sub-section, computational complexity is evaluated in terms of number of real multiplications per “OFDM
symbol” (i.e., at most M complex data symbols) for each MCM system. But multiplications with ±1 and ±j are not included since they are merely flips of sign and/or flips of real and imaginary parts. The normalized complexity of one MCM is the complexity of that MCM divided by that of OFDM. A pair of Lc-point FFT and IFFT (via Split Radix FFT [40]) with complexity
\[ \mu_{\text{FFT/IFFT}}(L_c) = 2(L_c \log_2 L_c - 3L_c + 4) + 4M \]
for \( M_c \) (where \( M_c \leq L_c \)) QAM symbols (subcarriers) is used as the component in the efficient implementations of all MCMs, where \( L_c \) is a power of 2. A 1-tap FDE with complexity
\[ \mu_{1\text{-tap FDE}}(M_s) = 4M_s \]
for \( M_s \) QAM symbols (subcarriers) is assumed for all MCMs. Therefore, the complexities of OFDM and Filtered-OFDM (either using transmit filter only or using both transmit and receive filters), including the 1-tap IFFT, are summarized in Table 2. In Filtered-OFDM in [29], it is assumed that the transmit filtering and adding CP could be combined such that the filtering is only performed once for the CP samples.

### 4.1.1. RB-F-OFDM complexity

In the direct implementation, similar to Filtered-OFDM in [29], it is assumed that the transmit filtering and adding CP could be combined such that the filtering is only performed once for the CP samples; it is also assumed that the receive filtering, CP removal, and downsampling could be combined such that filtering is performed only for the needed samples to yield the downsampled signals. Note that the per-RB filter is a real filter. The complexities of different components for one RB are summarized in Table 3.

In the polyphase implementation of RB-F-OFDM, the complexity of the SFB and AFB is due to the C-point IFFT, C-point FFT, and 4q sets of polyphase filters of \( p[n] \). Assume that the number of active RBs is \( n_{RB} \), where \( n_{RB} \leq K \). The complexities of different components are summarized in Table 4.

The total complexities of different implementations of RB-F-OFDM are summarized in Table 5.

#### 4.1.2. Evaluation and comparison with other MCMs

Examples of modified LTE-like MCM systems using the parameters in Table 1 are considered here. But we change the number of subcarriers per RB from \( D = 12 \) in LTE to \( D = 2^d \), where \( d = 4, 5, \) or 6. Consequently, the maximum number of RBs in a 10 MHz bandwidth should be reduced from \( K = 50 \) RBs to \( K = 640/D \) RBs.

The normalized complexities of the OFDM and Filtered-OFDM are summarized in Table 6. And the normalized complexities of RB-F-OFDM with different RB sizes \( D \) and different number of RBs \( n_{RB} \) are summarized in Table 7 for different implementations. From the comparison between these tables, we have the following observations: Firstly, in RB-F-OFDM, the direct implementation is more efficient when \( n_{RB} \) is small, while the polyphase implementation is more efficient when \( n_{RB} \) is large. Secondly, by increasing \( D \) (decreasing the frequency granularity), the complexity of RB-F-OFDM for a fixed total bandwidth could be highly reduced. Thirdly, when there are 16 subcarriers/RB, the complexity of RB-F-OFDM with polyphase implementation is 17–20 times the complexity of OFDM, depending on \( n_{RB} \). It is much higher than the complexity of OFDM, and is up to twice that of Filtered-OFDM with both transmit and receive filters. Yet it has much better spectral containment than Filtered-OFDM in non-contiguous spectrum. Lastly, when there are 32 or 64 subcarriers/RB, the complexity of RB-F-OFDM could be reduced to less than that of Filtered-OFDM with both transmit and receive filters.

#### 4.2. Latency

The latency discussed in this sub-section is the inherent latency introduced in the MCM’s structure from the input of the QAM symbols to the output of the estimated QAM symbols. The latency due to the arithmetic operations (e.g., multiplication) is ignored since they depend on hardware or digital signal processing (DSP) implementation. Let the sample duration be \( T_s = T/L \). Note that Filtered-OFDM and RB-F-OFDM essentially have the same latency as OFDM with CP. In OFDM, latency comes from the \( L : 1P/5 \) and

**Table 2**

<table>
<thead>
<tr>
<th>MCM</th>
<th>Number of real multiplications per “OFDM symbol” (M complex data symbols)</th>
</tr>
</thead>
<tbody>
<tr>
<td>OFDM</td>
<td>( \mu_{\text{OFDM}} = 2(L \log_2 L - 3L + 4) + 4M )</td>
</tr>
<tr>
<td>Filtered-OFDM</td>
<td>( \mu_{\text{Filtered-OFDM, point}} = 2(L \log_2 L - 3L + 4) + 2L_{f} + 4M )</td>
</tr>
<tr>
<td></td>
<td>( \mu_{\text{Filtered-OFDM, total}} = 2(L \log_2 L - 3L + 4) + 2L_{f} + 2(L + L_{cp})L_{f} + 4M )</td>
</tr>
</tbody>
</table>

**Table 3**

<table>
<thead>
<tr>
<th>N-point FFT/IFFT</th>
<th>Tx/Rx filtering</th>
<th>1-tap FDE</th>
<th>RB modulation/demodulation</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \mu_{\text{FFT/IFFT}}(N) = 2(N \log_2 N - 3N + 4) )</td>
<td>( \mu_{\text{filtering}} = 2(2N + \lceil L_{cp}/Q \rceil)L_{f} )</td>
<td>( \mu_{1\text{-tap FDE}}(D) = 4D )</td>
<td>( 8(L + L_{cp}) )</td>
</tr>
</tbody>
</table>

**Table 4**

<table>
<thead>
<tr>
<th>Per-RB processing</th>
<th>Non per-RB processing</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \mu_{\text{FFT/IFFT}}(N) = 2(N \log_2 N - 3N + 4) )</td>
<td>( \mu_{\text{SFB/AFB}} = \mu_{\text{FFT/IFFT}}(C) + 4L_{cp} = 2(C \log_2 C - 3C + 4 + 2q_{\text{cp}}) )</td>
</tr>
<tr>
<td>( \mu_{1\text{-tap FDE}}(D) = 4D )</td>
<td>( n_{RB} )</td>
</tr>
<tr>
<td>( \mu_{\text{SFB/AFB}} = \mu_{\text{FFT/IFFT}}(C) + 4q_{\text{cp}} )</td>
<td>( N + \lceil L_{cp}/Q \rceil )</td>
</tr>
</tbody>
</table>

**Table 5**

**Table 6**

**Table 7**

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In this section, we evaluate the spectral containment of RB-F-OFDM and compare it with OFDM and Filtered-OFDM. The spectral containment could be evaluated in terms of the PSD, out-of-band (OOB) power versus the OOB frequency ($f_{out}$ in MHz), as well as the required minimum guard band to meet certain OOB power requirement, in systems with 5 MHz bandwidth (i.e., using 300 subcarriers). We consider the required OOB power to be either $-50$ dB or $-55$ dB. The OOB power is measured with 20 kHz spacing, and each OOB power of a particular OOB frequency is the average measured power of a 100 kHz bandwidth centered at that OOB frequency.

4.3. Spectral containment

In this section, we investigate how the RB-F-OFDM transmit signal's spectral containment depends on the per-RB FFT size $N$ and equal ripple filter length $L_p$. The investigation allows us to find tradeoffs between the guard band size (to meet OOB power requirement) and other performance measures that depend on the per-RB FFT size $N$ and filter length $L_p$.

In Fig. 10, the OOB power (in dB) is plotted versus the OOB frequency ($f_{out}$ in MHz) for RB-F-OFDM transmit signals with different per-RB FFT sizes and filter lengths. And the minimum guard bands for the RB-F-OFDM transmit signals to meet the OOB power thresholds are summarized in Table 9.

We have the following observations: Firstly, the stopband attenuation of the transmit signal depends mainly on the per-RB FFT size. Due to the equal ripple filter design with fixed passband and stopband deviations requirements, the stopband attenuation of all equal ripple filters of different lengths are very close. Therefore, as in Fig. 10, with the same per-RB FFT size but different filter lengths, the stopband attenuations of RB-F-OFDM transmit signals are very close. When the per-RB FFT size is larger, the upsampling factor $Q$ is smaller, and thus, in each RB, the spectral images due to upsampling are further away from the desired passband and have larger signal attenuation after filtering. Therefore, the stopband attenuation of the transmit signal is larger. When the per-RB FFT size is large enough, e.g., 512, and when a properly designed per-RB filter, e.g., the 37-tap or 53-tap equal ripple filter, is used, all spectral images could be almost eliminated and the OOB power of the transmit signal could be as small as $-90$ dB. Secondly, the transmit signal's transition band behavior depends mainly on the equal ripple filter length. Shorter filter has faster roll off and larger transition band, while longer filter has faster roll off and smaller transition band, as long as all spectral images fall in the stopband of the per-RB transmit filter.

Therefore, in the design of RB-F-OFDM systems, the per-RB FFT size and filter length could be jointly chosen to meet the OOB power requirement, and to provide tradeoffs between OOB and other performance measures. For example, as in Table 9, if the design requirement of spectral containment is to have guard bands no greater than
Table 9
Minimum guard bands for 5 MHz RB-F-OFDM systems.

<table>
<thead>
<tr>
<th>Per-RB FFT size</th>
<th>32</th>
<th>64</th>
<th>128</th>
<th>128</th>
<th>256</th>
<th>512</th>
<th>512</th>
<th>1024</th>
</tr>
</thead>
<tbody>
<tr>
<td>Filter length</td>
<td>85</td>
<td>69</td>
<td>53</td>
<td>69</td>
<td>53</td>
<td>37</td>
<td>53</td>
<td>37</td>
</tr>
<tr>
<td>Guard band for −50 dB</td>
<td>370 kHz</td>
<td>328 kHz</td>
<td>416 kHz</td>
<td>309 kHz</td>
<td>394 kHz</td>
<td>565 kHz</td>
<td>400 kHz</td>
<td>567 kHz</td>
</tr>
<tr>
<td>Guard band for −55 dB</td>
<td>&gt; 1600 kHz</td>
<td>957 kHz</td>
<td>466 kHz</td>
<td>356 kHz</td>
<td>439 kHz</td>
<td>634 kHz</td>
<td>443 kHz</td>
<td>637 kHz</td>
</tr>
</tbody>
</table>

500 kHz for OOB power threshold −55 dB, RB-F-OFDM systems with \((N, L_p)\) being (128, 53), (128, 69), (256, 53), and (512, 53) all satisfy the requirement. Complexity wise, the one with (128, 53) achieves the lowest complexity, and may be chosen in the design.

4.3.2. Comparison with other MCMs

In Figs. 11 and 12, respectively, the OOB power is plotted versus the OOB frequency \(f_{\text{out}}\) in MHz for different MCM systems without or with a nonlinear PA that is modeled using a 5th order polynomial. The minimum guard bands for different MCM systems without nonlinear PA are as in Table 10. We have the following observations: Firstly, OFDM is most robust to nonlinearity of PA, because its OOB after the nonlinear PA does not highly increase, compared to its OOB without the nonlinear PA. But due to its large OOB, it always has the worst spectral containment. Secondly, even though Filtered-OFDM with a transmit filter could largely reduce the OOB, compared to OFDM, this is only true if PA nonlinearity is not taken into account. Filtered-OFDM is quite sensitive to PA nonlinearity. When there is a nonlinear PA, Filtered-OFDM’s OOB is much larger, and increasing the transmit filter length does not help to reduce OOB. Lastly, compared to Filtered-OFDM, RB-F-OFDM is more robust to PA nonlinearity. It only has around 10 dB OOB increase when a nonlinear PA is applied. This is in part due to the fact that PA nonlinearity largely increases the power in transition band and stopband of a filtered signal but not the power in passband. As in Fig. 5, the filtered spectral images of the passband are the dominant interference in the stopband of a per-RB signal in RB-F-OFDM. Even though PA nonlinearity increases the power in transition bands and stopbands between the spectral images, the stopband of the entire RB-F-OFDM transmit signal would not have large power increase.

We conclude that RB-F-OFDM still provides considerable small OOB when a nonlinear PA is applied. It is much better than OFDM and Filtered-OFDM. Certain PA linearization, e.g., digital pre-distortion (DPD), could be used to further reduce the OOB increase due to nonlinear PA.

In Fig. 13, the PSDs of OFDM, Filtered-OFDM, and RB-F-OFDM are compared in a system with 5 MHz bandwidth when the available spectrum is fragmented and non-contiguous, and it illustrates the advantage of RB-F-OFDM over Filtered-OFDM in this scenario. In this figure, two available spectrum fragments are separated by an unavailable frequency band that is occupied by others. When Filtered-OFDM is used with a filter over the entire frequency band, we can see that the OOB to the unavailable frequency band between the available fragments is the same as that of OFDM. However, RB-F-OFDM can utilize the available fragments with less OOB to the unavailable frequency band in the middle. RB-F-OFDM needs a much larger number of guard bands than OFDM or Filtered-OFDM.
**Table 10**
Minimum guard bands for 5 MHz MCM systems without nonlinear PA.

<table>
<thead>
<tr>
<th>MCM Parameters</th>
<th>OFDM</th>
<th>Filtered-OFDM</th>
<th>RB-F-OFDM</th>
</tr>
</thead>
<tbody>
<tr>
<td>Guard band for $-50,\text{dB}$</td>
<td>$&gt;1600,\text{kHz}$</td>
<td>465 kHz</td>
<td>279 kHz</td>
</tr>
<tr>
<td>Guard band for $-55,\text{dB}$</td>
<td>1142 kHz</td>
<td>608 kHz</td>
<td>435 kHz</td>
</tr>
</tbody>
</table>

Fig. 11. OOB power of MCM transmit signals versus OOB frequency without nonlinear PA.

Fig. 12. OOB power of MCM transmit signals versus OOB frequency with nonlinear PA.

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4.4. PAPR performance

In this section, we compare the PAPR performance of RB-F-OFDM with OFDM and Filtered-OFDM, and evaluate the PAPR performance of RB-F-OFDM after PAPR reduction. The PAPR performance is evaluated in terms of the complementary cumulative distribution function (CCDF) of PAPR for MCM signals. All results are generated for 10 MHz systems with 600 subcarriers and 10,000 subframes.

In general, the PAPR observation interval can be very long. But in practice, it is usually chosen as the effective symbol duration. In this paper, the PAPR measure for the $n$th “OFDM” symbol is defined as

$$\text{PAPR}_n = \max_{l \in \mathbb{N}} \frac{|x[l]|^2}{E(|x[l]|^2)} \cdot \begin{cases} \frac{1}{n(L + L_{cp})} \cdot \frac{n(L + L_{cp})}{n(L + L_{cp}) - 1} \\ \end{cases}.$$  \tag{28}

Note that for the filtering based MCM, e.g., Filtered-OFDM and RB-F-OFDM, due to the overlapping of symbols in time, the observed PAPR values in the beginning and end of a continuous transmission are discarded for statistics purpose.

4.4.1. Without PAPR reduction

In Fig. 14, we observe that the CCDF curves of PAPR of the OFDM, Filtered-OFDM, and RB-F-OFDM signals are very close when no PAPR reduction techniques are applied.

4.4.2. With PAPR reduction

In this sub-section, we evaluate the performance of RB-F-OFDM with RPTS and QPTS using different numbers of randomly generated phase vectors $n$, (e.g., 8, 16, 32, 64) and different RB group sizes $n_g$. For QPTS, $n_{pts} = 8$ quantized phase values with the candidate set $\{0, \pi/4, \pi/2, \ldots, 7\pi/4\}$, or $n_{pts} = 4$ quantized phase values with the candidate set $\{0, \pi/2, \pi, 3\pi/2\}$ are used. “RPTS $\left( n_r, n_g \right)$” means RPTS with $n_r$ random phase vectors and $n_g$ RB/group; “QPTS $\left( n_r, n_g, n_{pts} \right)$” means QPTS with $n_r$ random phase vectors, $n_g$ RB/group, and $n_{pts}$ quantized phase values.

The CCDF curves of PAPR for RB-F-OFDM, without PAPR reduction (black solid line), with RPTS (colored solid lines) and QPTS (dotted lines for $n_{pts} = 4$) are plotted in Figs. 15 and 16, for $n_r = 8$ and $n_r = 64$, respectively. The PAPR values for RPTS and QPTS at CCDF of $10^{-4}$ are summarized in Tables 11–13, respectively, for different combinations of parameters.

We have the following observations for both RPTS and QPTS: Firstly, for the same number of randomly generated phase vectors, smaller RB group size (i.e., more RB groups) leads to lower PAPR. The PAPR performances of 1, 2, or 5 RB/group are very close, and they are relatively close to that of 10 or 13 RB/group. This suggests that large RB group size and less RB groups could be used to achieve relatively good PAPR performance while keeping minimum amount of side information. Secondly, for the same RB group size, larger number of random phase vectors inefficiently leads to lower PAPR. Thus, to lower the computational complexity in PAPR reduction, the number of random phase vectors could be kept to a relatively small number, e.g., 8 or 16. Thirdly, tradeoffs could be made to achieve same PAPR performance while minimizing the complexity and amount of side information. Lastly, with the minimum complexity increase and amount of side information in our investigation, i.e., by using 8 random phase vectors and 25 RB/group, the PAPR values at CCDF of $10^{-4}$ are 10.85 dB for RPTS (8, 25), 10.83 dB for QPTS (8, 25, 8), and 10.94 dB for QPTS (8, 25, 4), and have at least 1 dB gain, compared to the case without PAPR reduction (see Fig. 14).

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Comparing RPTS and QPTS, when the RB group size is small (i.e., the number of RB groups are large), or when the number of random phase vectors is small, their PAPR performances are similar. Only when the number of RB groups is small and when the number of random phase vectors is large enough, e.g., \((n_r, 25, n_{pts})\) where \(n_r = 16, 32,\) or \(64\) and \(n_{pts} = 8\) or \(4\), and \((n_r, 17, 4)\) where \(n_r = 32\) or \(64\), QPTS has higher PAPR than RPTS. In particular, in this case, larger quantized phase candidate set provides better PAPR reduction in QPTS.

4.5. Bit error rate (BER) performance

In this section, we evaluate the BER performance of RB-F-OFDM and compare it with OFDM and Filtered-OFDM. The BER is evaluated versus the \(E_b/N_0\) (the energy per bit to
Fig. 16. CCDF of PAPR of RB-F-OFDM signals using RPTS/QPTS with $n_r = 64$.

Table 11
PAPR of different RB-F-OFDM signals with RPTS ($n_r$, $n_g$) at CCDF of $10^{-4}$.

<table>
<thead>
<tr>
<th>$n_r$</th>
<th>1</th>
<th>2</th>
<th>5</th>
<th>10</th>
<th>13</th>
<th>17</th>
<th>25</th>
</tr>
</thead>
<tbody>
<tr>
<td>8</td>
<td>10.05</td>
<td>10.09</td>
<td>10.07</td>
<td>10.26</td>
<td>10.33</td>
<td>10.53</td>
<td>10.85</td>
</tr>
</tbody>
</table>

Table 12
PAPR of different RB-F-OFDM signals with QPTS ($n_r$, $n_g$, 8) at CCDF of $10^{-4}$.

<table>
<thead>
<tr>
<th>$n_r$</th>
<th>1</th>
<th>2</th>
<th>5</th>
<th>10</th>
<th>13</th>
<th>17</th>
<th>25</th>
</tr>
</thead>
<tbody>
<tr>
<td>8</td>
<td>10.10</td>
<td>10.07</td>
<td>10.08</td>
<td>10.19</td>
<td>10.31</td>
<td>10.42</td>
<td>10.83</td>
</tr>
<tr>
<td>16</td>
<td>9.79</td>
<td>9.84</td>
<td>9.82</td>
<td>9.95</td>
<td>10.14</td>
<td>10.23</td>
<td>10.76</td>
</tr>
<tr>
<td>64</td>
<td>9.48</td>
<td>9.47</td>
<td>9.51</td>
<td>9.65</td>
<td>9.73</td>
<td>10.02</td>
<td>10.68</td>
</tr>
</tbody>
</table>

Table 13
PAPR of different RB-F-OFDM signals with QPTS ($n_r$, $n_g$, 4) at CCDF of $10^{-4}$.

<table>
<thead>
<tr>
<th>$n_r$</th>
<th>1</th>
<th>2</th>
<th>5</th>
<th>10</th>
<th>13</th>
<th>17</th>
<th>25</th>
</tr>
</thead>
<tbody>
<tr>
<td>8</td>
<td>10.03</td>
<td>10.05</td>
<td>10.04</td>
<td>10.14</td>
<td>10.32</td>
<td>10.45</td>
<td>10.94</td>
</tr>
<tr>
<td>16</td>
<td>9.80</td>
<td>9.79</td>
<td>9.87</td>
<td>9.98</td>
<td>10.11</td>
<td>10.38</td>
<td>10.87</td>
</tr>
</tbody>
</table>

noise power spectral density ratio) in dB. The moderately selective “EVA5” channel model (Extended Vehicular A channel model with maximum Doppler frequency 5 Hz) from [41] is used. It is assumed that perfect channel knowledge is known at the receiver. The 1-tap minimum mean square error (MMSE) FDE is used at the receiver of all MCMs. If not otherwise specified, only the results using 16QAM modulation is presented here due to page limit. Each BER point is obtained by averaging over 10,000 subframes.

We investigate two cases. In the case without ACI, there is only one transmitter and one receiver, and no interferers in adjacent frequency bands. On the other hand, in the case with ACI, except for the desired transmitter and receiver, there is an interfering transmitter transmitting in an (non-overlapping) adjacent frequency band of the same bandwidth as the desired system. The two transmitters are perfectly synchronized in time. The edge subcarriers of the two systems are as close as spacing $\Delta F$ being 1.5 subcarrier spacing away from each other. The ACI power $\Delta P$ in dB is defined as the difference between the average power of the interfering system’s passband and that of the desired system’s passband, and $\Delta P$ is a parameter in simulations. The two systems are as depicted in Fig. 17. At the desired receiver, the ACI is treated as noise, and the ACI power is not measured or accounted for in the MMSE equalization.

Except for Section 4.5.1, 128-point per-RB FFT and 53-tap equal ripple per-RB filter is used in RB-F-OFDM.

4.5.1. RB-F-OFDM with different parameters

In this section, we present the BER performance of RB-F-OFDM with various per-RB FFT sizes and filter lengths in the EVA5 channel. “RB-F-OFDM ($N$, $L_{p,tx}$, $L_{p,rx}$)” means RB-F-OFDM with $N$-point per-RB FFT, $L_{p,tx}$-tap per-RB transmit filter, and $L_{p,rx}$-tap per-RB receive filter. When $L_{p,rx}$ is “N/A”, it means that an $L$-point CP-OFDM receiver is used instead of the RB-F-OFDM receiver and there is no receive filtering.

In Fig. 18, the raw BER performance of RB-F-OFDM systems with 10 MHz bandwidth (i.e., $K = 50$ RBs and $M = 600$ subcarriers) and without ACI is shown. Almost all RB-F-OFDM curves have some BER degradation, compared to
OFDM, since their filters contribute to the selectivity of the equivalent channel. The BER performance mainly depends on the filter length. RB-F-OFDM with shorter overall filters provide better BER performances, while the ones with longer overall filters (compared to CP length) have larger degradation in medium to high signal-to-noise-ratio (SNR). Note that when the per-RB FFT size is large enough, e.g., 128 or above, for the same filter length, larger per-RB FFT does not show advantage in BER performance.

In Fig. 19, the raw BER performance of RB-F-OFDM systems with 5 MHz bandwidth (i.e., $K = 25$ RBs and $M = 300$ subcarriers) and with ACI $\Delta P = 0$ dB is shown. Note that RB-F-OFDM with 32-point per-RB FFT and 85-tap transmit filter and a CP-OFDM receiver does not show any performance gain over OFDM when ACI is present. Receive filtering is needed to reject ACI, since the CP removal breaks the continuity of the signal, resulting in much larger OOB of the interfering signal. This phenomenon also holds for Filtered-OFDM. For RB-F-OFDM systems with both transmit and receive filters, their performance depend on the per-RB FFT sizes as well as the filter lengths. When the ACI power is 0 dB, only RB-F-OFDM with 53-tap or 37-tap filters show performance gains over OFDM since the performance is more sensitive to the equivalent channel selectivity than ACI. Note that they also have quite close performance in the $[10^{-2} \text{ } 10^{-1}]$ raw BER range, which is the most meaningful raw BER range to achieve the highest throughput when channel coding is used in practical systems. Taking into account the spectral containment and complexity perspectives, the one with 128-point per-RB FFT and 53-tap transmit and receive filters is chosen in the design and evaluated in the next sub-sections.

4.5.2. Without adjacent channel interference (ACI)

In Fig. 20, the raw BER performance of the MCM systems with 5 MHz bandwidth (i.e., $K = 25$ RBs and $M = 300$ subcarriers) is shown in EVA5 channel without ACI. All MCMs perform identically at low to medium SNR, including the meaningful $[10^{-2} \text{ } 10^{-1}]$ raw BER range. At high SNR, OFDM performs the best as long as no ACI is present. Filtered-OFDM suffers slight BER degradation at high SNR, because its 35-tap transmit and receive filters contribute to the selectivity of the equivalent channel. As for RB-F-OFDM, it has largest BER degradation at high SNR, because the overall equivalent channel is much longer than CP.

4.5.3. With ACI

In Figs. 21 and 22, respectively, the raw BER performance of the MCM systems with 5 MHz bandwidth (i.e., $K = 25$ RBs and $M = 300$ subcarriers) is shown in EVA5 channel with ACI powers 10 and 30 dB, respectively, for 16QAM modulation. In Fig. 23, the raw BER performance of the MCM systems is shown in EVA5 channel with ACI power 30 dB for 64QAM modulation.

We have the following observations: Firstly, OFDM is most vulnerable to ACI. When ACI other than channel selectivity becomes a more dominant factor for performance degradation, Filtered-OFDM and RB-F-OFDM all outperform OFDM, because they could mitigate certain levels of ACI. Secondly, the relatively short filters enable Filtered-OFDM to outperform RB-F-OFDM in the moderately selective EVA5 channel, when channel selectivity is a more dominant factor for performance degradation. However,
Filtered-OFDM exhibits to be relatively more sensitive to ACI power increase, compared to RB-F-OFDM, because the RB-F-OFDM receiver has better ACI rejection capability. In Fig. 11, compared to RB-F-OFDM, Filtered-OFDM (with 35-tap transmit filter) transmit signal has smaller transition band. But this does not imply that Filtered-OFDM have better BER performance under ACI. At the receiver, the per-RB receive filter of RB-F-OFDM has much smaller passband and transition band, and could reject ACI much better than the loose 35-tap receive filter used in Filtered-OFDM. Therefore, the RB-F-OFDM receiver structure, which filters each RB individually with a much narrower per-RB filter, is more capable of ACI rejection. When ACI power is relatively large, e.g., 30 dB, RB-F-OFDM outperforms Filtered-OFDM. The difference is more profound for higher order symbol modulation, e.g., 64QAM.

In fact, preliminary results show that by adding very small guard bands, RB-F-OFDM could mitigate ACI much better and provide much better BER performance. This part of investigation, along with the investigation of the performance in non-contiguous spectrum, will be left for future work.

5. Conclusion and future work

In this paper, we have presented a new MCM scheme called RB-F-OFDM, and presented an efficient polyphase implementation of the transceiver. The performance of the proposed MCM scheme has been compared with OFDM and Filtered-OFDM in the moderately selective EVA5 channel under various ACI conditions. Without ACI, OFDM offers the best performance while the performance of Filtered-OFDM and RB-F-OFDM degrades at high SNR compared to OFDM due to the fact that the CP length is not enough to remove the ISI completely when filters are used. With ACI, the RB-F-OFDM receiver structure, which
Fig. 21. BER performance of MCM systems in EVA5 channel with ACI 10 dB for 16QAM modulation.

Fig. 22. BER performance of MCM systems in EVA5 channel with ACI 30 dB for 16QAM modulation.

Fig. 23. BER performance of MCM systems in EVA5 channel with ACI 30 dB for 64QAM.
filters each RB individually with a much narrower per-RB filter, is more capable of ACI rejection. Therefore, the performance of the proposed RB-F-OFDM is least sensitive to ACI power increase, and as the ACI power increases, RB-F-OFDM eventually outperforms Filtered-OFDM.

When compared to Filtered-OFDM, the proposed RB-F-OFDM enables utilization of non-contiguous spectrum due to very low achievable OOB E between the utilized channels, and more effective ACI rejection at the receiver. Therefore, RB-F-OFDM is more spectrally agile. Moreover, the RB-F-OFDM transmit signal is backwards compatible with legacy OFDM receivers and the RB-F-OFDM receiver could also demodulate legacy OFDM signals. In addition, we have shown that the PAPR of the proposed MCM scheme can significantly be reduced by utilizing SLM and PTS based techniques. Therefore, RB-F-OFDM presents itself as a viable candidate for future wireless communication networks.

Future work includes the filter optimization, theoretical analysis of ergodic capacity and error probability, BER and throughput comparison of the MCM schemes with non-contiguous spectrum and/or small guard bands and/or higher mobility environment, and performance evaluation in 802.11 based systems with longer CP.

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