Two-Step Resource Allocation for BIC-UFMC Wireless Communications

C. Vitiello, P. Del Fiorentino, V. Lottici,
F. Giannetti and M. Luise
Department of Information Engineering
University of Pisa, Italy

E. Debels and M. Moeneclaey
Department of Telecommunications
and Information Processing (TELIN)
Ghent University, Belgium

***Abstract*—This paper outlines an efficient resource allocation (RA) algorithm for packet-oriented Universal Filtered Multi-Carrier (UFMC) communications with bit interleaved coded (BIC) modulation. Assuming perfect synchronization and perfect knowledge of the channel, the proposed RA strategy optimizes the coding rate, bit loading and power distribution within the subbands and the subcarriers of the BIC-UFMC system maximizing the goodput (GP) metric. The bit loading operation is implemented by means of a *greedy* algorithm in order to reduce the computational complexity of the overall RA. In the presence of a frequency-selective fading channel, we show an improvement of the GP around 0.5 bit/s/Hz in comparison with a uniform bit and power allocation.**

***Index Terms*—UFMC, BIC, Resource, Goodput, Bit Loading**

I. Introduction

Internet of Things (IoT) is revolutionizing wireless communications influencing the future developments of next mobile network [1], changing also traffic characteristics toward sporadic communications of very short packets carrying few information bits. Traditional multicarrier modulation developed for fourth-generation (4G) network, i.e. OFDM, shows advantages in terms of robustness against fading and low complexity modulation and demodulation algorithms, achieving efficiency on high-rate mobile communications. However, OFDM could not be suitable for a typical IoT communications. Indeed, OFDM requires highly complex synchronization algorithms in order to preserve orthogonality, which are in contrast with the energy saving principle of the IoT. Therefore, OFDM could be unsuitable for the future fifth-generation (5G) network, because if the synchronization procedure is not perfect, Inter Carrier Interference (ICI) occurs and the system performance falls down. Furthermore, high sidelobe levels make OFDM sensitive to interference in case of spectrum sharing, cyclic prefix (CP) length decreases spectral efficiency and latency does not allow real-time communication, i.e. for Tactile Internet [1]. For the reasons above, new waveforms have been proposed.Universal Filtered Multi-Carrier (UFMC) groups subcarriers into subbands, in which IFFT and filtering are performed, satisfying 5G requirements, increasing robustness against both time and frequency misalignment and improving energy and spectral efficiency[2], [3]. In parallel with the study of novel waveforms for the next fifth generation (5G), the efficient resource allocation (RA) of the transmission parameters comes out as one of the main challenges to be met in the 5G wireless communications [4], [5]. Moreover, advanced coding schemes such as bit interleaved coded (BIC) modulation demonstrated to the system performance [6]. In this paper, we tackle the RA problem for BIC-UFMC systems, starting from our previous work [7], where a preliminary RA strategy with uniform bit allocation has been introduced, also considering frequency misalignments. In detail, a robust and computationally efficient RA mechanism is presented for a packed-oriented BIC-UFMC transmission over a frequency-selective fading channel under the hypothesis of perfect synchronization and perfect knowledge of the channel. The RA problem is based on the maximization of a particular performance metric, i.e., the goodput (GP) [8]. We exploit the GP, because it is a more suitable metric to evaluate the actual performance of packet-oriented systems employing practical modulation and coding schemes. In particular, the objective function of the RA problem is an estimation of the GP, called *estimated* GP (EGP), which has been presented in [9]. The derived RA technique is divided in two parts. In the first step, the code rate for the BIC block is selected together with the modulation scheme and the power allocation (PA) vector over the UFMC subbands. In the second step, a bit allocation (BA) vector and a PA one are derived for the subcarriers belonging to each subband. Numerical results show the efficiency in terms of GP of the RA for BIC-UFMC, which has been compared with transmission having uniform BA and PA vectors.

Notations. Vectors are in bold, $[\cdot]^{\mathrm{T}}$ is the transpose operator, $\lfloor x \rfloor$ is the lower integer of x and $[x]^+ \triangleq \max\{0, x\}$.

II. System Model

In this paper we consider a packet-based BIC-UFMC communication, which scheme is depicted in Fig. 1, over a frequency-selective fading channel, which is assumed to be stationary for the whole transmission duration of a packet. Each transmitted packet is composed by N_{u} bits, N_{p} of which are information and N_{CRC} are cyclic redundancy check (CRC). These N_{u} bits are encoded with rate $r \in \mathcal{D}_{\mathrm{r}} \triangleq \{r_0, \cdots, r_{|\mathcal{D}_r|}\}$, obtaining N_{c} coded bits that are subsequently randomly interleaved. After the BIC block, the signal passes in the UFMC modulator. The N_{c} bits are transmitted using B subbands, each one composed of D subcarriers, giving a total number of $N = BD$ subcarriers.

At the subcarrier level, therefore, the coded bits are loaded over the N subcarriers and Gray-mapped to complex symbols obtaining the vector $\mathbf{s} \triangleq [s(0), \cdots, s(N-1)]^{\mathrm{T}}$, according to the BA vector $\mathbf{m} \triangleq [m(0), \cdots, m(N-1)]^{\mathrm{T}}$, where $m(n) \in \mathcal{D}_m \triangleq \{0, 2, \cdots, m_{\max}\}$ and $m_{\max}$ represents the maximum of bits per subcarrier. The complex symbols belong to the unit energy constellation $\chi_n \triangleq 2^{m(n)}$-QAM, $\forall n \in \mathcal{N} \triangleq \{0, \cdots, N-1\}$. Subsequently, PA is performed, by element-wise multiplying the complex symbols $\mathbf{s}$ by the square root of $\mathbf{p} \triangleq [p(0), \cdots, p(N-1)]^{\mathrm{T}}$ with constraint $\sum_{n=0}^{N-1} p(n) \leq P_{\mathrm{tot}}$, where P_{tot} is the available power. The BA and PA operations are performed in two phases, called *subband* RA and *subcarrier* RA respectively as shown in Fig. 1. The *subband* RA fixes, for each subband $i \in \mathcal{B} \triangleq \{0, \cdots, B-1\}$, the QAM constellation, obtaining the vector $\tilde{\mathbf{m}} \triangleq [\tilde{m}_0, \cdots, \tilde{m}_{B-1}]^{\mathrm{T}}$, and the subband PA vector $\tilde{\mathbf{P}} \triangleq [\tilde{P}_0, \cdots, \tilde{P}_{B-1}]^{\mathrm{T}}$ with constraint $\sum_{i=0}^{B-1} \tilde{P}_i \leq P_{\mathrm{tot}}$.

The *subcarrier* RA performs first a zero loading (ZL) strategy, where a generic subcarrier $d \in \mathcal{D} \triangleq \{0, \cdots, D-1\}$, belonging to the generic subband i, is either switched off or modulated with constellation size $2^{\tilde{m}_i}$, which is determined by the subband RA. Latter the power $\tilde{P}_i$ is allocated. Therefore, after the *subband* and the *subcarrier* RA, the vectors $\mathbf{p}_i \triangleq [p(iD), \cdots, p(iD+D-1)]^{\mathrm{T}}$ and $\mathbf{s}_i \triangleq [s(iD), \cdots, s(iD+D-1)]^{\mathrm{T}}$ are calculated, which are the i-th partition of vectors $\mathbf{p}$ and $\mathbf{s}$, $\forall i \in \mathcal{B}$. Finally, per-subband processing is performed: $N_{\mathrm{FFT}} - D$ virtual subcarriers are added and then an IFFT with size N_{FFT} is executed, obtaining the block of samples $\mathbf{x}_i$ in time-domain. The vector $\mathbf{x}_i$ is subsequently convoluted with a Dolph-Chebyshev finite impulse response (FIR) filter with 60 dB side lobe attenuation and length L. The generic impulse response $\mathbf{q}$ of the filter is tuned to the i-th subband by a normalized frequency shift defined as $\Delta_i \triangleq \frac{D+1}{2} + iD$, obtaining $\mathbf{q}_i$. The output of each filter, therefore, provides the vector $\mathbf{z}_i$, whose generic u-th element $z_i(n)$ can be expressed as

$$z_i(u) = \sum_{m=0}^{N_{\mathrm{FFT}}-1} x_i(m) q_i(u-m), \tag{1}$$

where $u = 0, \cdots, N_{\mathrm{FFT}} + L - 2$. The vectors of all subbands are then element-wise summed, in order to obtain $\mathbf{z} \triangleq \sum_{i=0}^{B-1} \mathbf{z}_i$. The resulting vector $\mathbf{z}$ is transmitted over a frequency-selective fading channel, with impulse response $\mathbf{h}$. Considering only one UFMC symbol, at the receiver side, after the linear convolution with the channel impulse response that includes path loss, the signal in time-domain is described as

$$r(l) \triangleq \sum_{m=0}^{N_{\mathrm{FFT}}+L-1} z(m) h(l-m) + w(l), \tag{2}$$

where $l = 0, ..., N_{\mathrm{FFT}} + L + L_{\mathrm{CH}} - 3$, L_{CH} is the length of the channel impulse response and $w(l) \in \mathcal{CN}(0, \sigma^2)$ is a complex-valued sample of Gaussian noise. Following the UFMC receiver structure in [2], the first $N_{\mathrm{FFT}} + L - 1$ samples of the received signal (2) are processed by an FFT of size $2N_{\mathrm{FFT}}$, followed by a downsampling by a factor 2 taking only

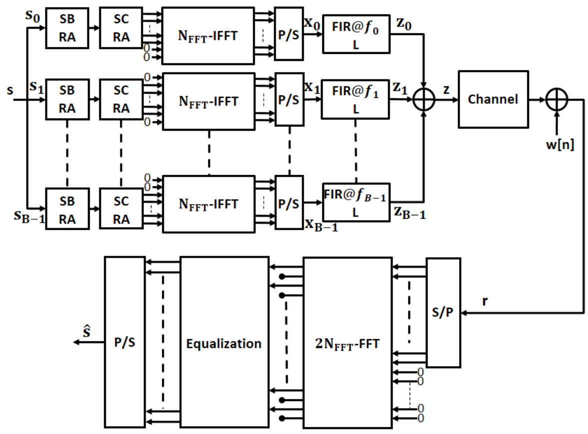

Fig. 1. Architecture of UFMC transceiver

the even-indexed subcarriers. In this way, the k-th subcarrier in the frequency domain is

$$Y(k) = H(k) \sum_{i=0}^{B-1} Q_i(k) X_i(k) + \mathcal{I}(k) + W(k), \tag{3}$$

where $k = 0, 2, ..., 2N_{\mathrm{FFT}} - 2$, $H(k)$, $Q_i(k)$, $X_i(k)$ and $W(k)$ represent the generic k-th element of the $2N_{\mathrm{FFT}}$-FFT output related to channel impulse response $\mathbf{h}$, filter impulse response $\mathbf{q}_i$, symbols $\mathbf{x}_i$ and ambient noise respectively. Moreover $\mathcal{I}(k)$ is the interference contribution defined in [10], which includes the ICI due to the cut of $L_{\mathrm{CH}} - 1$ samples in (2) and the Inter Symbol Interference (ISI) as a result of an overlapping of consecutive transmitted UFMC symbols, because CP is not present. Finally, under the hypothesis of perfect synchronization $X_i(k)$ can be written as

$$X_i(k) \triangleq \begin{cases} \sqrt{p(\frac{k}{2})} s(\frac{k}{2}), & \text{for } \frac{k}{2} = iD, ..., iD+D-1 \\ 0, & \text{otherwise} \end{cases}, \forall i \in \mathcal{B} \tag{4}$$

In this case eq. (3) becomes

$$Y(n) \triangleq H(n) Q_{i(n)}(n) \sqrt{p(n)} s(n) + \mathcal{I}(n) + W(n), \tag{5}$$

with $n \in \mathcal{N}$ and $i(n) \triangleq \lfloor n/D \rfloor$, $i(n) \in \mathcal{B}$. Finally, the signal (5) is equalized by a zero forcing (ZF) equalizer and soft-decoded.

III. Goodput Metric Evaluation

As discussed in Sect. II, practical modulation and coding schemes are employed, therefore the GP, which is defined as the number of correctly received information bits per unit of time [11], is a convenient metric for providing a reliable information about the link performance. In order to get an estimation of the GP, i.e. the EGP, the κESM link performance prediction (LPP) method is exploited [9]. In detail, the EGP function depends on the packet error rate (PER) of the link, but unfortunately, a closed-form expression of PER for coded multicarrier systems over frequency-selective channels is difficult to derive. The solution of this problem is given by the κESM technique [9], which compresses the vector of the received SNRs $\mathbf{\Gamma} \triangleq [\gamma(0), \cdots, \gamma(N-1)]^{\mathrm{T}}$ into a single SNR value γ_{eff}, called *effective* SNR. Given a transmission mode (TM) $\phi \triangleq (r, \mathbf{m}) \in \mathcal{D}_r \times \mathcal{D}_m$ and a PA vector $\mathbf{p}$, γ_{eff} is used to estimate the PER from an equivalent coded BPSK system over the AWGN channel, so that

$$\mathrm{PER}_{\phi}(\mathbf{\Gamma}) \cong \Phi_{\mathrm{r}}(\gamma_{\text{eff}}), \tag{6}$$

where PER_{ϕ} denotes the PER of the coded multicarrier system over frequency-selective channel for a TM ϕ and Φ_r represents the equivalent single carrier coded BPSK system over AWGN channel experiencing the *effective* SNR. The κESM expression is defined as [9]

$$\gamma_{\text{eff}} \triangleq -\log\left[\frac{1}{\sum_{n=0}^{N-1} m_n}\sum_{n=0}^{N-1}\alpha(m_n)\mathrm{e}^{-p(n)\gamma(n)\beta(m_n)}\right], \tag{7}$$

with $\alpha(m_n)$ and $\beta(m_n)$ constant values depending on the size of QAM constellation $m_n, \forall n \in \mathcal{N}$. Exploiting the κESM LPP method, the EGP function can be expressed in bit/s/Hz as

$$\begin{aligned}\zeta(\phi, \mathbf{p}) &\triangleq \frac{N_{\mathrm{FFT}}}{N_{\mathrm{FFT}}+L-1}\frac{T_{\mathrm{s}}}{N}\frac{N_{\mathrm{p}}[1-\Phi_r(\gamma_{\text{eff}})]}{\frac{N_{\mathrm{u}}T_{\mathrm{s}}}{r\sum_{n=0}^{N-1} m_n}} = \\ &= \frac{N_{\mathrm{FFT}}}{N_{\mathrm{FFT}}+L-1}\frac{N_{\mathrm{p}}}{NN_{\mathrm{u}}} r\sum_{n=0}^{N-1} m_n[1-\Phi_r(\gamma_{\text{eff}})],\end{aligned} \tag{8}$$

where $B_{\text{syst}} \triangleq N/T_{\mathrm{s}}$ is the bandwidth, T_{s} is the sample time at the source and $N_{\mathrm{u}}T_{\mathrm{s}}/r\sum_{n=0}^{N-1} m_n$ is the packet transmission time.

IV. Resource Allocation for BIC-UFMC systems

The purpose of the RA technique is to efficiently calculate the optimal PA vector $\mathbf{p}^* \triangleq [p^*(0), \cdots, p^*(N-1)]^{\mathrm{T}}$ and TM $\phi^* \triangleq (r^*, \mathbf{m}^*)$, where $\mathbf{m}^* \triangleq [m^*(0), \cdots, m^*(N-1)]^{\mathrm{T}}$ is the optimal BA vector, by maximizing the EGP function (8).

In detail, the key idea behind the presented RA is to exploit a high degree of flexibility, which is given by the subbands, therefore, the RA problem is divided in two steps. In the first step, a BA procedure selects the optimal coding rate r^* and the BA vector $\tilde{\mathbf{m}}^* \triangleq [\tilde{m}_0^*, \cdots, \tilde{m}_{B-1}^*]^{\mathrm{T}}$ over the subbands, and subsequently a PA strategy allocates the available power P_{tot} over the subbands, obtaining the vector $\tilde{\mathbf{P}}^* \triangleq [\tilde{P}_0^*, \cdots, \tilde{P}_{B-1}^*]^{\mathrm{T}}$. In the second step, a ZL technique and the same PA strategy allocate the vectors $\tilde{\mathbf{m}}^*$ and $\tilde{\mathbf{P}}^*$ over the D subcarriers for each subband, thus the optimal allocation $\mathbf{m}^*$ and $\mathbf{p}^*$ are determined. Furthermore, we approximate the signal (5) as

$$Y(n) \cong H(n)Q_{i(n)}(n)\sqrt{p(n)}s(n) + W(n), \tag{9}$$

where the interference $\mathcal{I}(n)$ was discarded, because it is attenuated by the subband filtering, as shown in [10], making UFMC signal quasi-orthogonal. Consequently, exploiting the approximation (9), the RA problem can be simplified. Accordingly to Eq. (9), the received SNR normalized for power transmission is defined as $\gamma(n) \triangleq \frac{|H(n)|^2|Q_{i(n)}(n)|^2}{\sigma^2}$. In light of the above, the steps of the RA technique are presented in the following sections.

A. Problem Formulation and Solution for Subband RA

The BA strategy over the subbands is performed for calculating r^* and $\tilde{\mathbf{m}}^*$ exploiting an iterative *greedy* algorithm, which has been presented in [12] and it is opportunely adapted here for the BIC-UFMC system. Subsequently, the vector $\tilde{\mathbf{P}}^*$ can be evaluated.

Let us define an average channel in frequency-domain for each subband $\tilde{H}(i) \triangleq \frac{1}{D}\sum_{d=0}^{D-1} H(iD+d)Q_i(iD+d), \forall i \in \mathcal{B}$, and a subband SNR as

$$\tilde{\gamma}(i) \triangleq \frac{|\tilde{H}(i)|^2}{\sigma^2}, \quad \forall i \in \mathcal{B}. \tag{10}$$

Considering uniform power over the subbands, i.e. $\tilde{P}_i \triangleq P_{\text{tot}}/B$, $\forall i \in \mathcal{B}$, the BA optimization problem (OP) can be defined as

$$(\tilde{\mathbf{m}}^*, r^*) = \underset{\tilde{\mathbf{m}}, r}{\arg\max}\left\{\tilde{\zeta}(\tilde{\mathbf{m}}, r)\right\} \tag{11}$$

$$\text{s.t. } \tilde{\mathbf{m}} \in \mathcal{D}_m, \;\; r \in \mathcal{D}_r, \tag{12}$$

where the subband EGP (S-EGP) $\tilde{\zeta}$, which is a particular case of the EGP formulation in (8), is

$$\tilde{\zeta}(\tilde{\mathbf{m}}, r) \triangleq \frac{N_{\mathrm{FFT}}}{N_{\mathrm{FFT}}+L-1}\frac{N_{\mathrm{p}}}{BN_{\mathrm{u}}} r\sum_{i=0}^{B-1}\tilde{m}_i[1-\Phi_r(\tilde{\gamma}_{\text{eff}})], \tag{13}$$

and $\tilde{\gamma}_{\text{eff}}$ is the subband κESM (S-κESM) derived by (7) as

$$\tilde{\gamma}_{\text{eff}} \triangleq -\log\left[\frac{1}{\sum_{i=0}^{B-1}\tilde{m}_i}\tilde{\Omega}(\tilde{\mathbf{m}})\right], \tag{14}$$

with

$$\tilde{\Omega}(\tilde{\mathbf{m}}) \triangleq \sum_{i=0}^{B-1}\alpha(\tilde{m}_i)\mathrm{e}^{-\tilde{P}_i\tilde{\gamma}(i)\beta(\tilde{m}_i)}. \tag{15}$$

The problem (12) can be solved by means to an exhaustive search with complexity $\mathcal{O}(\mathcal{D}_m^B)$, but exploiting the *greedy* algorithm proposed in [12], we are able to reduce the complexity to $\mathcal{O}(B\log B)$. Once the $\tilde{\mathbf{m}}^*$ and r^* are established, the subband PA vector $\tilde{\mathbf{P}}^*$ is calculated. Considering that the S-EGP (13) monotonically increases when $\tilde{\gamma}_{\text{eff}}$ (14) increases, the maximization of S-EGP is equivalent to minimize $\tilde{\Omega}(\tilde{\mathbf{m}})$ (15). Therefore the PA OP is defined as

$$\tilde{\mathbf{P}}^* = \underset{\tilde{\mathbf{P}}}{\arg\min}\left\{\tilde{\Omega}(\tilde{\mathbf{m}}^*, \tilde{\mathbf{P}})\right\} \tag{16}$$

$$\text{s.t. } \sum_{i=0}^{B-1}\tilde{P}_i \leqslant P_{\text{tot}} \;\text{ and }\; \tilde{P}_i \geqslant 0, \forall i \in \mathcal{B} \tag{17}$$

Proposition 1. The OP (17) is a convex optimization problem [13] whose solution, $\forall i \in \mathcal{B}$, is given by

$$\tilde{P}_i^* \triangleq \frac{1}{\tilde{\gamma}(i)\beta(\tilde{m}_i^*)}\left[\log\frac{1}{\theta} - \log\left(\frac{1}{\tilde{\gamma}(i)\alpha(\tilde{m}_i^*)\beta(\tilde{m}_i^*)}\right)\right]^+. \tag{18}$$

The solution (18), already proved in [7] for a different RA approach, is a typical water-filling solution [13], where $\log(1/\theta)$ is the water level, which is such that $\sum_{i=0}^{B-1} \tilde{P}_i^*$ meets the power constraint P_{tot} with equality.

B. Problem Formulation and Solution for Subcarrier RA

In this second step of the RA, we first allocate $\tilde{\mathbf{m}}^*$ and after $\tilde{\mathbf{P}}^*$ over the D subcarrier of each subband, thus obtaining the transmission parameters ϕ^* and $\mathbf{p}^*$.

Recalling that in the ith subband $\tilde{m}_i^*$ specifies the $2^{\tilde{m}_i^*}$-QAM constellation, we perform a ZL strategy with the aim to obtain the BA vector $\mathbf{m}_i^* \triangleq [m^*(iD), \cdots, m^*(iD+D-1)]^{\mathrm{T}}$, where the generic element of the vector $\mathbf{m}_i^*$ can assume the values 0 or $\tilde{m}_i^*$. The ZL is based on the minimization of the subband PER (S-PER) $\Phi_r^{(i)}(\gamma_{\mathrm{eff}}^{(i)})$, where $\gamma_{\mathrm{eff}}^{(i)}$ is the κESM calculated over the D subcarriers in the subband i and it is defined as

$$\gamma_{\mathrm{eff}}^{(i)} \triangleq -\log\left[\frac{1}{\sum_{d=0}^{D-1} m(iD+d)}\Omega(\mathbf{m}_i)\right], \tag{19}$$

with

$$\Omega(\mathbf{m}_i) \triangleq \sum_{d=0}^{D-1} \alpha(m(iD+d))\mathrm{e}^{-p(iD+d)\gamma(iD+d)\beta(m(iD+d))}, \tag{20}$$

and $\mathbf{m}_i \triangleq [m(iD), \cdots, m(iD+D-1)]^{\mathrm{T}}$, which is a partition of the BA vector $\mathbf{m}$. In light of the above, given uniform power over the subcarriers, i.e. $p(iD+d) \triangleq P_{\mathrm{tot}}/BD = P_{\mathrm{tot}}/N$, $\forall d \in \mathcal{D}$, the ZL problem can be formalized as follow

$$\begin{array}{l} \mathbf{m}_i^* = \arg\min\limits_{\mathbf{m}_i} \left\{\Phi_{r^*}^{(i)}(\gamma_{\mathrm{eff}}^{(i)}(\mathbf{m}_i))\right\} \\ \text{s.t.} \quad \mathbf{m_i} \in \mathcal{D}_m \end{array}. \tag{21}$$

As already done in IV-A, we apply the *greedy* algorithm achieving a complexity around to $\mathcal{O}(D\log D)$. The ZL solution is applied to the other subbands and the optimal BA vector $\mathbf{m}^*$ is given. Finally we allocate the subband available power $\tilde{P}_i^*$, $\forall i \in \mathcal{B}$ over the subcarriers. Given the optimal TM ϕ^* the subcarrier PA problem is formulated as

$$\mathbf{p}_i^* = \arg\min_{\mathbf{p}} \{\Omega(\mathbf{m}_i^*, \mathbf{p})\} \tag{22}$$

$$\text{s.t.} \sum_{d=0}^{D-1} p(iD+d) \leqslant \tilde{P}_i^* \text{ and } p(iD+d) \geqslant 0, \forall d \in \mathcal{D} \tag{23}$$

where $\mathbf{p}_i^* \triangleq [p^*(iD), \cdots, p^*(iD+D-1)]^{\mathrm{T}}$ is a partition of the PA vector $\mathbf{p}^*$.

Proposition 2. The OP (23) is a convex optimization problem [13] and the solution, $\forall d \in \mathcal{D}$, is given by

$$p^*(iD+d) \triangleq \frac{1}{\gamma(i)\beta(m_i^*)}\left[\log\frac{1}{\theta} - \log\left(\frac{1}{\gamma(i)\alpha(m_i^*)\beta(m_i^*)}\right)\right]^+ \tag{24}$$

proved in [7]. The OP (23) is evaluated for each subband obtaining the optimal PA vector $\mathbf{p}^*$. Finally, in Tab. I the RA strategy is summarized.

For $i = 0 : B-1$
  Set: $\tilde{P}_i = P_{\mathrm{tot}}/B$;
  Evaluate: $\tilde{\gamma}(i)$ in (10);
End For
Evaluate:
  $(\tilde{\mathbf{m}}^*, r^*) = \arg\max\limits_{\tilde{\mathbf{m}} \in \mathcal{D}_m, r \in \mathcal{D}_r} \{\tilde{\zeta}(\tilde{\mathbf{m}}, r)\}$ in (12);
For $i = 0 : B-1$
  Evaluate: $\tilde{P}_i^*$ in (18);
End For
Collect $\tilde{\mathbf{P}}^* = [\tilde{P}_0^*, \cdots, \tilde{P}_{B-1}^*]^{\mathrm{T}}$;
For $i = 0 : B-1$
  For $d = 0 : D-1$
    Set: $p(iD+d) = P_{\mathrm{tot}}/N$;
  End For
  Evaluate:
  $\mathbf{m}_i^* = \arg\min\limits_{\mathbf{m}_i \in \mathcal{D}_m} \{\Phi_{r^*}^{(i)}(\gamma_{\mathrm{eff}}^{(i)}(\mathbf{m}_i))\}$ in (21);
  Evaluate: $p(iD+d)^*$ in (24);
End For
Collect $\mathbf{p}^* = [\mathbf{p}_0^*, \cdots, \mathbf{p}_{D-1}^*]^{\mathrm{T}}$ and $\mathbf{m}^* = [\mathbf{m}_0^*, \cdots, \mathbf{m}_{D-1}^*]^{\mathrm{T}}$;
Return $\phi^* = \{r^*, \mathbf{m}^*\}$ and $\mathbf{p}^*$;

TABLE I
RA STRATEGY

V. NUMERICAL RESULTS

In this section, a comparison in terms of GP between the proposed RA technique referred as Optimal PA-Optimal BA (OPA-OBA) and classical Uniform PA-Uniform BA (UPA-UBA) strategy both applied to BIC-UFMC modulation is provided. In detail, in the UPA-UBA case, we calculate only the optimal TM $\phi^* = (r^*, \mathbf{m}^*)$, where $m^*(i) = m^*$, $\forall i \in \mathcal{N}$. The comparison is performed on AWGN and extended vehicular A (EVA) [14] channels, evaluated with the modified COST231 Hata path loss (PL) model. The set of simulation parameters is summarized in Tab. II, which are referred to LTE standard with 10MHZ channelization [15]. To simplify the processing of the soft-decoding, convolutional coding is used; however, turbo code can be considered as in [9]. The GP performance is evaluated by averaging the number of bits/s/Hz correctly received on N_{pkt} = 1000 transmitted

Simulation Parameters	*Value*
Information bits (N_{p})	1024
CRC (N_{CRC})	32
Subcarriers (N)	120
Subbands (B)	10
FFT size	1024
Bandwidth (B_{syst})	10 MHz
QAM modulation order (m)	2, 4, 6
Convolutional code-rate (r)	1/2, 2/3, 3/4, 5/6
Noise power in B_{syst} (P_{noise})	-110 dBm
Variance ambient noise (σ^2)	P_{noise}/N
Distance transmitter-receiver (d)	141 m

TABLE II
SYSTEM PARAMETERS

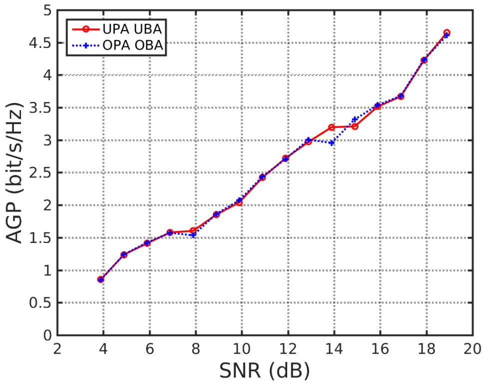

Fig. 2. UPA-UBA vs OPA-OBA on AWGN channel

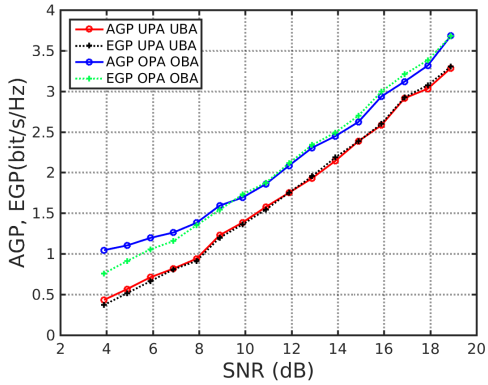

Fig. 3. UPA-UBA vs OPA-OBA on EVA channel

packet for each SNR. Thus, the average GP (AGP) is defined as $\mathrm{AGP} \triangleq \frac{N_{\mathrm{FFT}}}{N_{\mathrm{FFT}}+L-1}\frac{1}{N_{\mathrm{pkt}}}\frac{1}{B_{\mathrm{syst}}}\sum_{s=1}^{N_{\mathrm{pkt}}}\frac{N_{\mathrm{p}}\delta(s)}{T_{\mathrm{pkt}}(s)}$, where $\delta(s)$ equals 1 if the s-th packet is correctly decoded and 0 when it is discarded, $T_{\mathrm{pkt}}(s)$ is the transmission time of the s-th packet. Finally, each transmitted packet experiences independent channel realization and SNR points are calculated as $\mathrm{SNR} \triangleq P_{\mathrm{tot}}/(P_{\mathrm{noise}}\Theta)$ by varying P_{tot}, where Θ represents PL value. Initially, we evaluated the mean square error (MSE), defined as $\mathrm{MSE} \triangleq \sum_{l=1}^{N_{\mathrm{pkt}}} \frac{\sum_{n=1}^{N} \frac{|Y_l(n)-s_l(n)|^2}{N}}{N_{\mathrm{pkt}}}$, which has been calculated in absence of ambient noise, fixing the strategy UPA-UBA, code-rate $r = 1/2$ and normalizing the transmitted signal by the power and PL, in order to only obtain the level of interference $\mathcal{I}$ as a mean over the N subcarriers. The resulting interference value is very low, around -59 dB, therefore the approximation of the received signal in (9) is validated. Subsequently, the RA strategy has been tested on an AWGN channel. As shown in Fig. 2, AGP of OPA-OBA is quite similar to UPA-UBA as expected, because the channel frequency response is flat on the entire bandwidth of the signal and no difference in the RA procedure occurs. Next, we test the proposed RA on EVA channel. In Fig. 3 it is possible to appreciate the performance of OPA-OBA approach and its very few differences between EGP and AGP, except for the first sample due to a slightly degradation of PER approximation (6) at low SNR. Considering the frequency-selective fading EVA channel, relevant differences in terms of SNR between subbands and subcarriers occur, which are exploited by OPA-OBA RA procedure by allocating more power and bits to the best subbands and turning off the rest. For this reason, the AGP performance of the OPA-OBA is constantly above the UPA-UBA benchmark. Indeed, Fig.3 shows that a maximum AGP gain around 0.5 bits/s/Hz is obtained at low SNR exploiting the proposed RA strategy.

VI. Conclusions

In this paper, we presented a GP-based RA strategy based on the maximization of the GP metric for a packet-oriented BIC-UFMC transmission. We have shown a solution for selecting the transmission parameters, i.e. code rate, both BA and PA vectors, first over subbands and subsequently over subcarriers in order to maximize the GP metric. In presence of a frequency-selective fading channel, we achieved a significant GP gain for the proposed RA technique compared to a conventional uniform power and bit allocation. **Acknowledgment**: This work has been partially supported by the PRA 2016 research project 5GIOTTO funded by the University of Pisa.

References

[1] Wunder, G. et all, "5GNOW: non-orthogonal, asynchronous waveforms for future mobile applications", *in Comm. Magazine, IEEE* , Feb. 2014.

[2] Wang, X.; Wild, T.; Schaich, F.; Fonseca dos Santos, A., "Universal Filtered Multi-Carrier with Leakage-Based Filter Optimization", *in European Wireless Conference 2014; Proceedings of*, May 2014.

[3] Schaich, F.; Wild, T., "Waveform contenders for 5G - OFDM vs. FBMC vs. UFMC", *in Communications, Control and Signal Process. (ISCCSP), 2014 6th International Symposium on* , May 2014.

[4] M. N. Tehrani, M. Uysal and H. Yanikomeroglu, "Device-to-device communication in 5G cellular networks: challenges, solutions, and future directions," *in IEEE Comm. Mag.*, May 2014.

[5] Andrews, J.G.; Buzzi, S. et all, "What Will 5G Be?", *in Selected Areas in Comm., IEEE Journal on* , June 2014.

[6] Caire, G.; Taricco, Giorgio; Biglieri, Ezio, "Bit-interleaved coded modulation," *in Inform. Theory, IEEE Trans. on* , May 1998.

[7] Del Fiorentino, P.; Vitiello, C; et all, "A Robust Resource Allocation Algorithm for Packet BIC-UFMC 5G Wireless Communications", *EUSIPCO 2016*, August 2016.

[8] B. Devillers, J. Louveaux, and L. Vandendorpe, "Bit and power allocation for goodput optimization in coded parallel subchannels with ARQ", *IEEE Trans. on Signal Process.*, Aug.2008.

[9] Stupia, I.; Lottici, V.; Giannetti, F.; Vandendorpe, L., "Link Resource Adaptation for Multiantenna Bit-Interleaved Coded Multicarrier Systems", in Signal Process., *IEEE Trans. on* , July 2012.

[10] X. Wang, T. Wild, F. Schaich and S. ten Brink, "Pilot-Aided Channel Estimation for Universal Filtered Multi-Carrier," *Vehicular Technology Conf. (VTC Fall), 2015 IEEE 82nd*, Boston, MA, 2015

[11] D. Qiao, S. Choi and K.G. Shin, "Goodput analysis and link adaptation for IEEE 802.11a wireless LANs", Mobile Computing, *IEEE Transactions on* , Oct-Dec 2002.

[12] I. Stupia, F. Giannetti, V. Lottici and L. Vandendorpe, "A greedy algorithm for goodput-based adaptive modulation and coding in BIC-OFDM systems," *Wireless Conf. (EW), 2010 European*, Lucca, 2010.

[13] S. Boyd and L. Vandenberghe, "Convex optimization", *Cambridge Univ. Press*, 2004.

[14] Extended vehicular A (EVA) channel. Available: http://www.raymaps.com/index.php/lte-multipath-channel-models.

[15] LTE E-UTRA; Requirements for support of radio resource management (3GPP TS 36.133 version 12.10.0 Release 12).