Radio Resource Unit Allocation and Rate Adaptation in Filter Banks Multicarrier System

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***Abstract*— In this paper the effect of the effective channel gain and the SNR using improved joint resource unit (RU) allocation and Bit loading (JRAB) in a filter banks multicarrier (FBMC) system are analyzed. Computer simulations have been performed assuming a hypothetical WiMAX scenario in which an FBMC system substitutes OFDM, maintaining as much physical layer compatibility as possible. It has been demonstrated that it is possible to upper-bound the maximum delay for delay-sensitive applications (rtPS and nrtPS) using above mentioned effective metrics in a FBMC system.**

Keywords-RRM, Filter bank, WiMAX, JRAB

I. Introduction

Adaptive modulation is considered as one of the main techniques to increase the data rate that can be reliably transmitted over a fading channel. Many forms of adaptive modulation and transmission techniques have been proposed and implemented in recent wireless systems [1] [4] [5], however, Adaptive bit loading is one of the key features of very recent wireless communication systems (i.e., WiMAX, etc), and its importance will increase in the near future (i.e., LTE). Adaptive modulation technique is possible thanks to adjustability of many system parameters according to the channel fading state variability, the transmit power, the data rate, and channel coding rate.

Two adaptive approaches are widely considered. The first is the bite error rate (BER) (or the packet error rate -PER) which is bounded while the maximum throughput is attained by allocating different transmit powers into different users. This method has been treated in [1] [2] and is usually referred as rate adaptation (RA) process. The second approach is referred as margin adaptation (MA). The MA method is based on the use of minimum transmit power while the minimum required quality of service (QoS) is guaranteed (i.e., BER bound). It is also possible to generalize the (*multiuser*) link adaptation process using either the RA or the MA objective functions. The outcomes of these adaptations are the resources assigned to each user in the time, frequency or space domains, the transmitted power per user, and the optimum modulation and coding scheme (MCS). Using an orthogonal frequency division multiplexing (OFDM) scheme [2] resources can be efficiently assigned to different users without the need to use guard bands or time gaps (when perfect synchronization between the mobile station (MS) and the base station (BS) is assumed). The whole frequency and time domains are segmented into different resource units (RU) which can be arbitrarily (or using specific policies) assigned to different users. The minimum RU is a single symbol, the resource allocation algorithm inputs are the channel state information (CSI) of all users, and the maximum allowed transmission power. The main considered output is the power and MCS assigned to each RU.

To the best knowledge of the authors, this is the first time the radio resource unit allocation is treated using both the effective channel gain and the effective signal to noise ratio (SNR_{eff}) concept in a filter banks multicarrier (FBMC) system. The FBMC communication scheme has been subject of intense researches during last year's mainly within the ICT [1] European project PHYDYAS **(**Physical layer for dynamic spectrum access and cognitive radio**)** [9].

The remainder of this paper is organized as the following: in Section II, the model and main features of the used FBMC system is summarized. In Section III, the expression of the effective channel transfer functions of both OFDM and FBMC systems is derived, and the resource unit (RU) capacity is calculated in Section IV. The main results obtained by simulation are presented in Section V, and finally main conclusions are outlined in Section VI.

II. Introduction to FBMC System

Filter banks multicarrier system can be realized via a digital transmultiplexer configuration, where a synthesis filter bank (SFB) is used at the transmitter side while an analysis filter banks (AFB) is used at the receiver side [3] [9]. In FBMC applications, the use of a critical sampled filter banks would be problematic, since the aliasing effect would make it difficult to compensate the channel imperfections' effects by processing the sub-channel signals only after the AFB. Therefore, a factor of two oversampling is commonly applied into the sub-channel signals at the AFB [3]. In this paper a uniform modulated filter banks is assumed where the prototype filter $p[m]$ of length L is shifted to cover the

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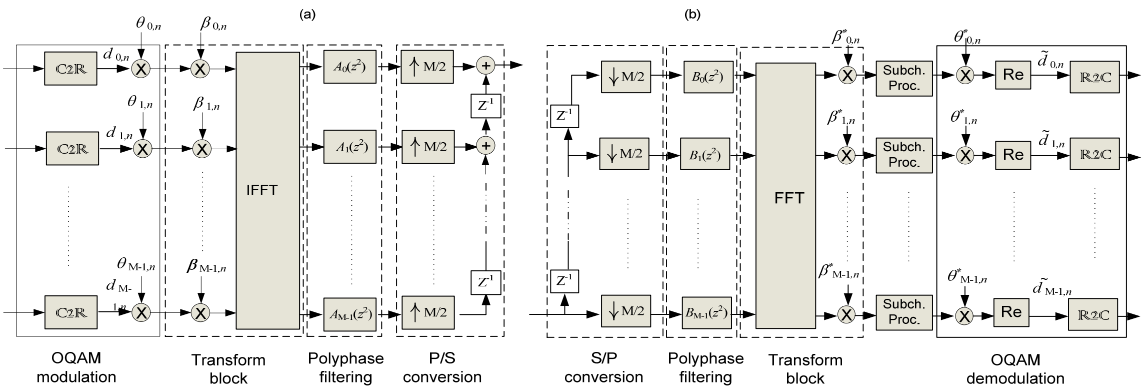

Figure 1: Multicarrier polyphase filter banks for SISO case, a) Synthesis filter banks (SFB), b) Analysis filter banks (AFB

whole system bandwidth. The output signal from the synthesis filter bank is defined as,

$$s[m] = \sum_{k=0}^{M-1} \sum_{n=-\infty}^{\infty} d_{k,n}\theta_{k,n}\beta_{k,n} p\left[m - \frac{nM}{2}\right] e^{j\frac{2\pi}{M}km} \quad (1)$$

where

$$\beta_{k,n} = (-1)^{kn} e^{-j\frac{2\pi k}{M}\frac{L-1}{2}} \quad (2)$$

where M is the total number of subcarriers (IFFT/FFT size), $d_{k,n}$ is the real-valued symbol (of rate $2/T$) modulated over the k-th subcarrier and the n-th time symbol interval. The time signaling interval T is defined as the inverse of the subcarrier spacing Δf . The symbols $d_{k,n}$ and $d_{k,n+1}$ can be seen as the in-phase and quadrature (I/Q) components of a complex-valued symbol $c_{k,l}$ (of rate $1/T$) chosen from a multilevel quadrature amplitude modulation (M-QAM) alphabet. Note that the sign of the sequence $\{\theta_{k,n} = j^{k+n}\}$ sequence can be chosen arbitrarily, but the pattern of the real and the imaginary samples have to follow definition (1) and (2) to maintain (near) orthogonality [3] [9]. L is the length of the prototype filter $p[m]$ and is equal to the product of the filter bank size M and the overlapping factor K (L=KM) [3]. The "C2R" and "R2C" blocks depicted in Figure 1, indicate the conversion from complex to real form, and the inverse operation respectively.

As it can be observed in (1), the synthesized signal is a composite of M sub-channel signals each one is a linear combination of time-shifted (by multiples of $T/2$) and the overlapped impulse response of the prototype filter weighted by the respective data symbol $d_{k,n}$. When a real (imaginary) part of a subcarrier symbol is used (to carry an information symbol) the unused imaginary (real) part is at the receiver a fairly complicated function of surrounding data symbols effect.

III. Effective OFDM and FBMC Channel Transfer Functions

In order to obtain the optimal power and the bit rate adaptation we need the bit error rate (*BER*) expression in additive white Gaussian noise (AWGN) which is easily invertible in terms of bit rate and power. Unfortunately, for most of non-binary modulation techniques, e.g., multi-level QAM (MQAM), and multi-level phase shift keying (M-PSK), an exact expression for the *BER* is hard to obtain. Often, the *BER* with Gray bit mapping at high SNRs can be approximated as the symbol error rate (*SER*) divided by number of bits per symbol [4]. The equivalent subcarrier approach developed by C. Tang *et. al*., in [5] allows a group of subcarriers containing spread data symbols to be represented by a single equivalent subcarrier to handle the bit and power loading mechanism in a more compact and simpler way.

To better understand this concept let first consider an OFDM system where the M-QAM bit error estimation approximation developed in [4] is used to obtain the *BER* of the k-th equivalent subcarrier,

$$BER_k \approx 0.2\,exp\left(\frac{-1.6\left|H_{eff,k}\right|^2 P_k}{\left(2^{b_k}-1\right)\sigma^2}\right) \quad (3)$$

where BER_k is the approximate *BER*, $\left|H_{eff,k}\right|^2$ means the square magnitude of the effective channel transfer function, P_k is the transmit power atthe k-th subcarrier, b_k is the number of transmitted bits, and σ^2 is the additive white Gaussian noise (AWGN) power. Using this expression as equality instead of an approximation, the expressions for the assigned power and number of bits may be solved as in [4] by,

$$P_k = \frac{-\left(2^{b_k}-1\right)\sigma^2}{1.6\left|H_{eff,k}\right|^2}\ \ln\left(\frac{BER_k}{0.2}\right) \text{ for } BER_k < 0.2 \quad (4)$$

$$\max b_k = \max\left\{\log_2\left(1-\frac{1.6\left|H_{eff,k}\right|^2 P_k}{\sigma^2 \ln\left(\frac{BER_k}{0.2}\right)}\right)\right\} \quad (5)$$

Subject to: $BER_k < \overline{BER_k}$

Subject to: $\sum_{k=1}^{M} P_k - P_T \le 0$

where $\overline{BER_k}$ is the upper bound *BER*, and P_T is the maximum allowed total transmit power. Knowing the value $\left|H_{eff,k}\right|^2$ of each active user, the number of bits that can be loaded without exceeding a certain *BER* threshold is estimated by

$$b_k = \log_2\left(1-\frac{1.6\left|H_{eff,k}\right|^2 P_k}{\sigma^2 \ln\left(\frac{BER_k}{0.2}\right)}\right) \quad (6)$$

The first step of the adaptation process is the calculation of the effective channel gain over the M subcarriers and transmitted frames (such parameters will be fixed in the simulation Section e.g., by using WiMAX standard specifications [8]). For purpose of an easier comprehension we started calculating the effective channel gain of an OFDM system, and after the FBMC system.

A. *Calculation of Effective Channel Transfer Function: OFDM Case*

We assume the single user case, where the modulated data symbol of the active user at the k-th carrier is S_k. By removing the cyclic prefix at the demodulator the k-th received data symbol in frequency domain is

$$Y_k = S_k H_k + \eta_k \quad (7)$$

where H_k means the channel gain, and η_k is the AWGN component at the k-th subcarrier index. After using a zero-forcing (ZF) equalization, the enhanced noise term at the k-th subcarrier is given by the term η_k / H_k. We assume that the noise power is equally distributed over all the subcarriers with a value σ^2. Therefore, the power of the enhanced noise term at k-th subcarrier is $\sigma^2 / \left|H_k\right|^2$. If we define $P_s = \mathrm{E}\left[S_k S_k^*\right]$ as the total power transmitted by a single subcarrier, the instantaneous signal to noise ratio will be equal to:

$$SNR_k = \frac{P_s}{\sigma^2 / \left|H_k\right|^2} \quad (8)$$

The effective channel power attenuation for the modulated symbol is,

$$\left|H_{eff,k}\right|^2 = \left|H_k\right|^2 \quad (9)$$

Substituting (9) into (8) yields to the following

$$SNR_k = \frac{P_s}{\sigma^2}\left|H_{eff,k}\right|^2 \quad (10)$$

Knowing the CSI values, the base station (BS) can use the calculated power in (9) to define the bit and power loading values for transmission. Then the packet scheduler is in charge of formatting the symbols to fit into one or several RUs. The MCS of the burst frame is fixed based on the effective *SNR* (SNR_{eff}) of the sub-channel and the symbol where the burst is allocated. The effective *SNR* function (SNR_{eff}) is a function of different instantaneous *SNR*s, and is defined as

$$SNR_{eff} = \phi^{-1}\left\{\frac{1}{M}\sum_{k=1}^{M}\phi\left(SNR_k\right)\right\} \quad (11)$$

B. *Calculation of Effective Channel Transfer Function: FBMC Case*

In contrast to the OFDM scheme where complex valued symbols are transmitted at a given symbol rate the FBMC transmits real symbols at twice the OFDM's rate. Therefore, FBMC is a scheme that preserves the spectral efficiency and even allows the optimization of the carrier pulse shape according to the channel characteristics [3] [9].

Using the Zero-Forcing Equalization: The frequency-time pair (k,n) denotes the subcarrier k and symbol time n (with $T/2$ spacing time) position respectively within a a transmitted frame (see Figure 2). In FBMC every frequency-time position suffers interference from neighboring sub-channels (Figure 2). For an ideal channel this interference affects only the sub-channels of the imaginary symbols, while the real part of the symbols yields the originally transmitted symbol $S_{k,n}$. Note that this interference could be considered as a (sometimes close to zero) random variable that depends on the transmitted symboles around the symbol position (k,n).

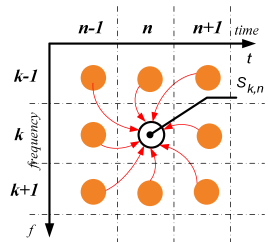

Figure 2: First time order neighbors in time-frequency representation for PHYDYAS FBMC system [9].

From (1) and (2), and with some mathematical arrangements we obtain in (12) the received FBMC signal expression

$$y_{k,n} = H_{k,n} S_{k,n} + j \sum_{(p,q)\neq(k,n)} H_{p,q} w_{p,q} S_{p,q} + \eta_{k,n} \qquad (12)$$

where the $\{w_{p,q}\}$ values are the constant filter banks coefficients depicted in Table 1, their effect constitute the main interference component in (12) (right summation value) on every transmitted data symbol.

TABLE 1: Transmultiplexer response of the FBMC system used in PHYDYAS project [9]. Rows represent time direction and columns the frequency direction.

0.0006	0.0001	0	0	0	0.0001	0006
- 0.0429j	0.1250	**0.2058j**	**0.2393**	**0.2058j**	0.1250	0.0429j
-0.0668	0.0002	**0.5644**	**1.000**	**0.5644**	0.0002	0.0668
0.0429j	0.1250	**0.2058j**	**0.2393**	0.2058j	0.1250	0.0429j
0.0006	0.0001	0	0	0	0.0001	0.0006

The neighborhood set of positions that affects a given frequency-time (k,n) position of a transmitted symbol is defined as,

$$\Omega_{\Delta k,\Delta n} = \left\{ (p,q), |p| \le \Delta k, |q| \le \Delta n, H_{k+p,n+q} \approx H_{k,n} \right\} \qquad (13)$$

We define the set $\Omega^*_{\Delta k,\Delta n}$ such that $\Omega^*_{\Delta k,\Delta n} = \Omega_{\Delta k,\Delta n} - (k,n)$. Note that both Δk and Δn should be chosen taking into account the channel time coherence T_c, and the bandwidth B_c. It is worth mentioning that when B_c decreases the value of Δk also decreases. The same can concluded for T_c and Δn. Having a well-dimensioned real system, B_c encompass few subcarriers ($\Delta k \ge 1$) while T_c is generally larger than T ($\Delta n \ge 1$). This allows us to rewrite the received signal in (12) using zero-forcing (ZF) equalization as,

$$\begin{aligned} \frac{y_{k,n}}{H_{k,n}} &= S_{k,n} \\ &+ j \sum_{(p,q)\in\Omega^*_{\Delta k,\Delta n}} \frac{H_{k+p,n+q}}{H_{k,n}} w_{k+p,n+q} S_{k+p,n+q} \\ &+ j \sum_{(p,q)\notin\Omega_{\Delta k,\Delta n}} \frac{H_{k+p,n+q}}{H_{k,n}} w_{k+p,n+q} S_{k+p,n+q} + \frac{\eta_{k,n}}{H_{k,n}} \end{aligned} \qquad (14)$$

If we consider that the generated filter banks prototype is well-localized in time and frequency domain, we consequently have

$$\begin{aligned} &\left| \sum_{(p,q)\notin\Omega_{\Delta k,\Delta n}} \frac{H_{k+p,n+q}}{H_{k,n}} w_{k+p,n+q} S_{l+p,n+q} \right| \\ &\ll \left| \sum_{(p,q)\in\Omega^*_{\Delta k,\Delta n}} \frac{H_{k+p,n+q}}{H_{k,n}} w_{k+p,n+q} S_{k+p,n+q} \right| \end{aligned} \qquad (15)$$

Therefore, (14) can be rewritten as (see weights' value at columns 2 and 6 in Table 2)

$$\frac{y_{k,n}}{H_{k,n}} \approx S_{k,n} + j \sum_{(p,q)\in\Omega^*_{\Delta k,\Delta n}} \frac{H_{k+p,n+q}}{H_{k,n}} w_{k+p,n+q} S_{k+p,n+q} + \frac{\eta_{k,n}}{H_{k,n}} \qquad (16)$$

using $(p,q) \in \Omega^*_{\Delta k,\Delta k}$ as the summation range in (16) means that we can approximate the channel gain at $(k+p, n+q)$ position by that experienced at (k,n). Therefore, using the ZF equalizer the received symbol at the k-th sub-carrier and n-th time is equal to,

$$\hat{S}_{k,n} = S_{k,n} + j \sum_{(p,q)\in\Omega^*_{\Delta k,\Delta n}} w_{k+p,n+q} S_{k+p,n+q} + \frac{\eta_{k,n}}{H_{k,n}} \qquad (17)$$

$$\hat{S}_{k,n} = S_{k,n} + jI_{k,n} \qquad (18)$$

From (15), the enhanced noise term after equalization is

$$I_{k,n} = \sum_{(p,q)\in\Omega^*_{\Delta k,\Delta n}} w_{k+p,n+q} S_{k+p,n+q} - j\frac{\eta_{k,n}}{H_{k,n}} \qquad (19)$$

where the power term here is,

$$\mathrm{E}\left[I_{k,n} I^*_{k,n} \right] \approx \sum_{(p,q)\in\Omega^*_{\Delta k,\Delta n}} w^2_{k+p,n+q} P_{k+p,n+q} + \frac{\sigma^2}{\left|H_{k,n}\right|^2} \qquad (20)$$

If $P_{k,n} = \left|S_{k,n}\right|^2$ is the total transmitted power by a single carrier, then, the instantaneous signal to noise ratio is

$$SNR_{k,n} = \frac{P_{k,n}\left|H_{k,n}\right|^2}{\sigma^2 + \sum_{(p,q)\in\Omega^*_{\Delta k,\Delta n}} \left|H_{k,n}\right|^2 w^2_{k+p,n+q} P_{k+p,n+q}} \qquad (21)$$

where $P_{k+p,n+q}$ is the total transmitted power by each symbol belongs to the set $\Omega^*_{\Delta k,\Delta n}$. From (20), it can be seen that the effective power for the modulated symbol is:

$$\left|H_{eff,k,n}\right|^2_{FBMC} = \frac{\left|H_{k,n}\right|^2}{\left(1 + \frac{\left|H_{k,n}\right|^2 \sum_{(p,q)\in\Omega^*_{\Delta k,\Delta n}} w^2_{k+p,n+q} P_{k+p,n+q}}{\sigma^2} \right)} \qquad (22)$$

Then the SNR_{eff} is calculated and is equal to,

$$SNR_{eff,k,n} = \frac{P_{k,n}}{\sigma^2} \left|H_{eff,k,n}\right|^2_{FBMC} \qquad (23)$$

Note that $\left|H_{eff,k,n}\right|^2_{FBMC}$ has a random behavior as it's strongly dependent of the interference part $\left\{ \sum_{(p,q)\in\Omega^*_{\Delta k,\Delta n}} w^2_{k+p,n+q} P_{k+p,n+q} \right\}$.

IV. FBMC RESOURCE UNIT ALLOCATION USING THE SNR_{eff}

In time division duplex (TDD) approach the communication frame consists on N_s symbols of duration T_{frame} seconds. The numbers of downlink and uplink OFDM/FBMC symbols usually follow the ratio 2:1 or 3:1. However, this parameter can be adjusted at the BS according to user demands and the available resources. The total system bandwidth BW consists of N_c subcarriers where only a limited number equal to N_{used} are active, while the remaining carriers are used as guard tones. Active subcarriers include both pilot subcarriers and data subcarriers, which will be mapped over different sub-channels according to specific subcarrier permutation schemes.

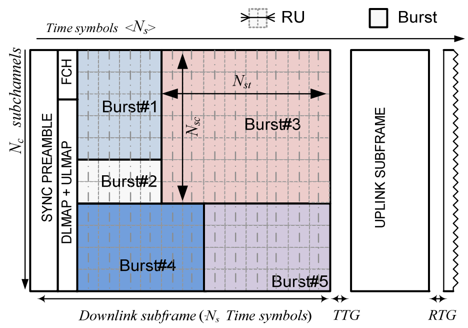

Figure 3: FBMC frame in TDD mode like burst structures based on IEEE 802.16e standard [8].

For the full usage of subcarriers (FUSC) scheme, pilot subcarriers are allocated first, and the remainders are grouped into sub-channels, where the data subcarriers are mapped. On the other hand, in partial usage subcarriers scheme (PUSC), and in adjacent subcarrier permutation schemes (usually referred as band AMC) map first all the pilots and the data subcarriers into the sub-channels, and therefore, each sub-channel contains its own set of pilot subcarriers.

For the FUSC and PUSC modes, the assigned subcarriers to each sub-channel are distant in frequency, whereas for AMC scheme the subcarriers belonging to one sub-channel are adjacent. Note that both FUSC and PUSC increase the frequency diversity and average the interference effect, whereas the AMC is more convenient for bit loading and beamforming as an increase in multiuser diversity is demanded. As depicted in Figure 3, and similar to WiMAX standard [8], the minimum RU assigned to any data stream within a frame has a two dimensional shape constructed by at least one sub-channel and two symbols[2].

We define a RU as a resource unit formed by a set of $N_{sc} \times N_{st}$ subcarriers and FBMC symbols, respectively. Once the size of the RU is defined it's possible to obtain the total number of RUs per frame $Q \times T$, where $Q = N_c / N_{sc}$ is the number of sub-channels and $T=N_s/N_{st}$ is the number of time slots. Note that both the RU and the data region always follow a rectangular shape structure. In the IEEE 802.16 standard, the specific size of the RU varies according to the permutation scheme, concretely for the AMC scheme the RU may take the sizes; 9×6, 18×3 or 27×2, where one ninth of the subcarriers are dedicated to pilots [8]. By analogy, and taking into account that the OFDM symbol duration is twice that of the FBMC symbol (Figure 2) for the AMC scheme, the FBMC RU may takes the sizes 9×12, 18×6 or 27×4.

The effective SNR_{eff} for FBMC is given by (23). The effective SNR merges the SNR from the different subcarriers, i.e., in a sub-channel or in a burst. Therefore, the BER, the packet error rate or the channel capacity can be obtained directly by assuming an AWGN channel from an equivalent SNR equal to the SNR_{eff}. The power is assumed uniformly distributed over all the subcarriers. Therefore, the effective channel gain can be obtained via the geometric mean of the subcarriers gain (considering the FBMC structure in the capacity calculation described in [4]) the total capacity C_{RU} within each RU is calculated by

$$C_{RU} = \sum_{i=1}^{N_{st}/2} \Delta f \; N_{sc} \left\lfloor \log_2 \left(1 - \frac{1.6 H_{eff}^2 \, SNR}{\ln\left(\frac{BER}{0.2}\right)} \right) \right\rfloor \tag{24}$$

where H_{eff}^2 is the geometric mean of the effective channel experienced over each i={1,2,…, $N_{st}/2$}, and the SNR_{eff} is the geometric mean of the SNR values over all the N_{sc} sub-channels of the RU unit.

Improved Joint RU Allocation and Bit Loading (JRAB) by Scheduling: Two competing aspects exist during the RU allocation and the scheduling process. These are

- the guarantee of the different service QoS constraints, and
- the maximization of the spectral efficiency.

The packet scheduling functions described by Shakkottai *et. al*. in [10] maximize both the spectral efficiency (therefore the bit loading) and the delay effect based on the CSI values. Furthermore, according to described functions in [10] the prevalence of the channel over the distribution of the RUs (to maintain the QoS) or the opposite is difficult to assure. From a system administrator perspective, this approach might be difficult to implement, and furthermore, it has been shown that when the physical layer is in charge of the bit allocation process of each active user the spectral efficiency increase [6]. For these reasons, the scheme proposed in [11]

[2] one OFDM symbol in case of OFDM scheme

has tackled the problem from a different perspective. It is usually unavoidable that the packets might be fragmented to fit into the physical layer burst.

We assume in this paper that any packet can be arbitrarily fragmented as many times as necessary (obviously this will affect the spectral efficiency due to the fragmentation of headers). Based on this assumption, and assuming that each packet is delivered within a certain time interval (no matter which class of service it belongs to, with either constant bit rate (CBR) or variable bit rate (VBR) [11]), by following the development in [12], we can obtain the minimum number of bits $b_k^{(u)}$ that the system should assign to each active FBMC user during each frame by

$$b_k^{(u)} = \begin{cases} T_{\text{frame}} \sum_{p=1}^{P} \frac{L_p^{(u)}}{\tau_{max}^{(u)} - \Delta\tau - \tau_p^{(u)}}, & \text{if } \forall p' \rightarrow \tau_{p'}^{(u)} < \left(\tau_{max}^{(u)} - \Delta\tau\right) \\ T_{\text{frame}} \left(\sum_{p=1}^{P} \frac{L_p^{(u)}}{\tau_{max}^{(u)} - \Delta\tau - \tau_p^{(u)}} + \sum_{p'} L_{i,p'} \right), & \text{otherwise} \end{cases} \quad (25)$$

where T_{frame} is the frame time period (in seconds), and $L_p^{(u)}$ is the number of bits still queued from the p-th packet. Then, if $b_k^{(u)}$ bits are allocated during each frame to each active user u and the *k-th* RU, the delay is certainly under its upper bound. Note that if any p' packet has waited period more than $\left(\tau_{max}^{(u)} - \Delta\tau\right)$ all the remained bits of the packet will be considered for transmission in the following frame.

The RU allocation and the bit loading problem can be solved for the a minimum rate $R^{(u)}$ based on the allocated bits $b_k^{(u)}$ as defined in (25). $b_k^{(u)}$ is also used to determine the priority assigned to each user within each RU. Hence, for the k-th sub-channel of the u-th user the hereafter priority assignment is defined as

$$\varphi_k^{(u)} = \begin{cases} \min\left(\frac{b_k^{(u)}}{b_{\max}}, 1\right) \frac{\bar{\omega}_k^{(u)}}{\bar{\omega}_{\max}}, & \text{if } \forall p' \rightarrow \tau_{p'}^{(u)} < \left(\tau_{\max}^{(u)} - \Delta\tau\right) \\ P_{\text{urgency}} \frac{\bar{\omega}_k^{(u)}}{\bar{\omega}_{\max}}, & \text{otherwise} \end{cases} \quad (26)$$

where $b_{\max}$ is a normalization factor and is equal to the maximum number of bits that can be transmitted within a frame using the highest MCS scheme. Furthermore, when a packet from the u-th user/service flow is close to exceed its maximum delay, the term $b_k^{(u)} / b_{\max}$ in (26) is substituted by an urgency factor P_{urgency}, which is a fixed constant satisfying $P_{\text{urgency}} > \left(\bar{\omega}_{\min} / \bar{\omega}_{\max}\right)^{-1}$ inequality. As a result, the packets close to their maximum delay are put ahead in the allocation process in order to avoid the packet drops due to the excessive packet delay. The $\bar{\omega}_k^{(u)}$ is the achievable throughput or the rate from the u-th user on the k-th RU, which is obtained based on the SNR_{eff} and the available MCSs.

V. SIMULATION RESULTS

Table 2 summarizes the simulation parameters used to verify the results of the RU allocation using the JRAB for FBMC system.

TABLE 2: FBMC Air Interface and System Level configuration parameters

Parameter	Values/ Quantities
Carrier Frequency, Bandwidth	3.5GHz, 20MHz
Sampling Frequency	22.857Msps
Subcarrier Permutation	Band AMC
FFT Length	2048
# of Used Subcarriers	1728
# of Subcarriers per Sub-Channel	18
# of FBMC Time Symbols per RU	6
# of Data Symbols per RU	48
Modulation	{4,16,64}-QAM
Channel Coding	Punctured Convolutional with coding rates1/2, 2/3
Channel Model	Pedestrian B
MS Velocity	10Km/h
Channel Estimation and CQI	Ideal
Shadowing Standard deviation	5dB
BS Transmit Power	49dBm

Figure 4 shows the cumulative density function of the packet delay for 50 and 100 active users. Let first focus on the case when the number of users is K=50. Figure 4 demonstrates that all the schemes achieve a delay lower than the maximum (50ms), in fact the 99th percentile is achieved at 25ms using the JRAB procedure and for the PFS (Proportional Fair Scheduler) [12]. Furthermore, the packet loss rate performance for each scheme is almost zero for the JRAB, and about 1.6×10^{-5} for the PFS.

For the case K =100, we can observe that the PFS is the algorithm that achieves the lower packet delays whereas the JRAB sends the packets mainly when the urgency factor is applied. During the simulations, the guard time $\Delta\tau$ is fixed equal to $0.2\,\tau_{\max}^{(u)}$, and thus the urgency factor is activated when $\tau_p^{(u)} > \left(\tau_{\max}^{(u)} - \Delta\tau\right) = 0.4$ ms . Now, for K =100 the packet loss rate for each scheduling function and bit loading procedure is 0.0824 and 0.1375 for the JRAB and the PFS respectively. Therefore, although most of the packets are sent when they are near to expire with the JRAB, a lower packet loss rate is achieved.

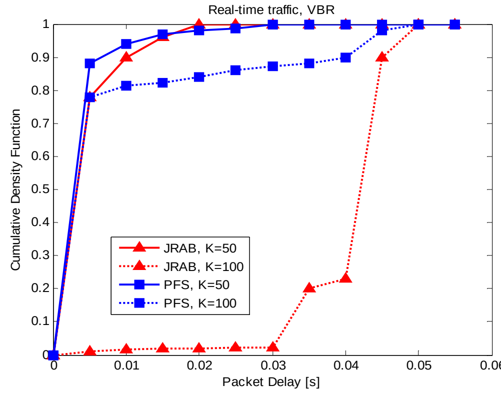

Figure 4: Cumulative density functions of the packet delay for the PFS, and JRAB algorithms with K=50 and 100 active users

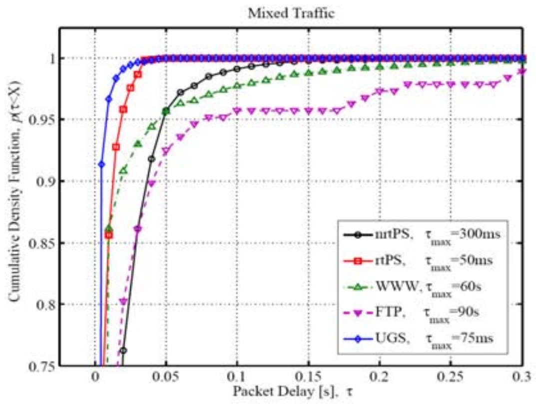

Figure 5: Cumulative density functions of the packet delay for the JRAB algorithm with mixed traffic and K=50 users.

The performance of the JRAB in case of mixed traffic is shown in Figure 5. In this scenario K=50 users are simulated, where ten users require non-real time test service (*nrtPS*), 13 users require real-time test service (*rtPS*), ten users are browsing internet files (World Wide Web service), five users are downloading files using the File Transfer Protocol (FTP), and 12 users require UGS connections for applications such as Voice over IP. The delay for the www and the FTP services has been assumed as τ_{max}=60s and τ_{max}=90s respectively.

It is clearly shown in Figure 5 that each traffic type achieves a maximum packet delay lower than its defined maximum value. The 99*th* percentile for the delay sensitive applications is found to be 100ms, 35ms and 20ms for the *nrtPS*, the *rtPS* and the UGS, respectively, much lower than the fixed maximum values.

VI. CONCLUSIONS

The analysis of using the effective channel information and the SNR_{eff} metrics in each RU using the JRAB algorithm has shown that it is possible to upper-bound the maximum delay for delay sensitive applications (*rtPS* and *nrtPS*) in a FBMC system. This was achieves despite the interference effect experienced at each subcarrier (or sub-channel) (k,n) position due to the proper characteristics of filter banks prototype. Besides the higher achieved efficiency by using FBMC compared to OFDM ([3] [7]), it is even possible to obtain an extra spectral efficiency margin by exploiting the multiuser diversity on those unallocated resources.

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