Complexity and implementation aspects of filter bank multicarrier

A potential technology enabler of next generation radio systems

Nikolaos Bartzoudis (CTTC), Vincent Berg (CEA), Leonardo Gomes Baltar (Intel), Oriol Font (CTTC), Kilian Roth (Intel), Miquel Payaró (CTTC), Michael Färber (Intel)

***Abstract*—Multicarrier modulations considered for 5G include filtered orthogonal frequency division multiplexing (f-OFDM), filter bank multicarrier (FBMC) and cyclic convolution based FBMC. In these modulations, to achieve better spectral containment compared to conventional OFDM, individual or groups of subcarriers are combined and shaped with a filter. This increases the computational complexity compared to OFDM. In this contribution, a comparison of this complexity is evaluated for these different modulation schemes. In addition to the complexity analysis, the suitability of FBMC for specific 5G use-cases is also demonstrated in this work via two experimental implementations. Both use-cases aim at exploiting the unused fragmented spectrum in bands below 1 GHz. FBMC exploits efficiently the underutilized fragmented spectrum while guaranteeing superior interference protection to coexisting primary transmissions. These key characteristics can satisfy critical requirements existing 4G and future 5G systems.**

Keywords—multi-carrier modulations, 5G, fragmented spectrum; efficient-agile waveforms

I. Introduction

The next generation mobile networks (NGMN) Alliance highlights in [1] the necessity to make more spectrum available in the existing sub-6GHz radio bands and introduce new agile waveforms that exploit the existing underutilized fragmented spectrum, in order to satisfy specific fifth generation (5G) operating scenarios [1]. Towards this end, new spectrally efficient waveforms are proposed whose goal is to complement or replace the current 4G long term evolution (LTE) technology used mobile broadband networks. LTE is based on orthogonal frequency division multiplexing (OFDM) and is spectrally inefficient due to certain features such as the cyclic prefix (CP), the high exposure to frequency offsets and the high sensitivity to interferences in fragmented spectrum use cases.

Different variants of filter bank multi carrier (FBMC) schemes have been revisited in the past years. A number of sources in the literature discusses the advantages and shortcomings of such candidate 5G waveforms [2], which contrary to CP-OFDM, require higher digital signal processing (DSP) complexity and feature an increased symbol period. These two factors have been mainly holding off the adoption of such waveforms in broadband mobile radio communications. However, advances in baseband processor technologies make it every time more cost and energy effective to adopt such new waveforms. On top of that, systematic research efforts aim at optimizing the DSP complexity of candidate 5G waveform without compromising the anticipated performance.

In this paper we first present a comparison of different 5G candidate waveforms in terms of DSP computational complexity. This study offers a useful insight towards this end. Furthermore, the implementation feasibility and performance validation of two FBMC schemes is demonstrated with two practical experimental prototypes.

In the first implementation, a secondary downlink (DL) FBMC transmission, exploits the unused spectrum in the 400 MHz band, where a primary professional mobile radio (PMR) system operates. The performance of the PMR system is characterized under different mobile channels and interference-levels resulting from the secondary FBMC transmission. A similar experimental validation is used, with a secondary system having LTE characteristics. The impact of the secondary systems (FBMC or OFDM) on the primary one - in terms of interference - is quantified.

The second experimental implementation focused on efficient spectrum usage in TV white space bands. In this context, opportunistic communication systems have to coexist in the same band with digital terrestrial TV (DTT) broadcast transmissions, but must guarantee that no harmful interference is incurred. When an OFDM-based system is acting as the secondary transmission in a shared spectrum usage, certain limitations in terms of dynamic range apply. The adjacent carrier leakage ratio (ACLR) is also a limiting factor in a LTE type system, while alternative approaches based on FBMC seem more suitable. As far as the baseband implementation is concerned, it is shown that on the transmitter side an additional filter is required on top of OFDM, and that the computational complexity of the FBMC receiver remains comparable to the one of an OFDM receiver.

II. Overview of 5G Candidate Waveforms and Complexity Analysis

Multicarrier modulations considered for 5G include filter bank based multicarrier with offset quadrature amplitude modulation (FBMC-OQAM), filtered OFDM (FOFDM), a.k.a. universal filtered multicarrier (UFMC), and cyclic convolution based FBMC, a.k.a. generalized frequency division multiplexing (GFDM). In this section, a comparison of the computational complexity is evaluated for the different transmission schemes. The complexity is quantified in terms of the total number of real multiplications and additions.

We consider the signal processing operations involved in the generation of the MC signal, as well as the recovery of the subcarrier signals and equalization in the presence of multipath propagation. We do not consider here the operations involved in channel estimation or calculation of the equalizer coefficients. Moreover, we assume that all the systems are perfectly synchronized.

A. Cyclic Prefix OFDM

We assume that the total of M subcarriers are available out of which M_f are occupied with symbols. We will consider first the number of real valued multiplications and additions to transmit one block of M_f symbols.

Starting with the fast Fourier transform (FFT), the number of real multiplications and additions of a M-point FFT/ inverse FFT (IFFT) using a split-radix algorithm are given by:

$$C_{\mathrm{FFT}}(M) = M(\log_2(\mathrm{M}) - 3) + 4,$$
$$A_{\mathrm{FFT}}(M) = 3M(\log_2(\mathrm{M}) - 1) + 4.$$

Since the transmitter (Tx) of a CP-OFDM system is basic built with one single IFFT we get:

$$C_{\mathrm{OFDM}}^{\mathrm{Tx}} = C_{\mathrm{FFT}}(M)$$
$$A_{\mathrm{OFDM}}^{\mathrm{Tx}} = A_{\mathrm{FFT}}(M).$$

For the demodulation and recovery of the subcarrier signals two processing tasks are necessary: FFT and single tap equalization per subcarrier. The complexity is the given by:

$$C_{\mathrm{OFDM}}^{\mathrm{Rx}} = C_{\mathrm{FFT}}(M) + 4M_{\mathrm{f}}$$
$$A_{\mathrm{OFDM}}^{\mathrm{Rx}} = A_{\mathrm{FFT}}(M) + 2M_{\mathrm{f}},$$

for the MC demodulation and equalization.

B. FBMC-OQAM

Assuming a FBMC-OQAM system where the prototype filter has length KM, two approaches can be adopted for the generation and recovery of the MC signal: the polyphase based and the frequency spread based structure.

We consider first the complexity of FBMC-OQAM implemented with a structure based on the polyphase decomposition of the prototype filter and using a direct form realization of the polyphase components (PC) [3]. The Tx is composed of 3 steps after the OQAM modulation:

- Phase rotations to get linear phase filters in each subcarrier
- IFFT
- Polyphase filtering followed by block overlapping of 50%

At the Rx side similar operations in the inverted order are implemented including one more step: polyphase filtering, FFT, multitap channel equalization per sub-carrier with an equalizer of length L_{eq} and the OQAM demodulation. The phase rotations at the receiver side can be embedded in the equalizer coefficients.

The total number of real valued multiplications is then given by:

$$C_{\text{PC-FB}}^{\mathrm{Tx}} = 2C_{\mathrm{FFT}}(M) + 4MK + 4M_{\mathrm{f}},$$
$$C_{\text{PC-FB}}^{\mathrm{Rx}} = 2C_{\mathrm{FFT}}(M) + 4MK + 4L_{\mathrm{eq}}M_{\mathrm{f}},$$

where we have taken into account that the IFFT and the polyphase network work double of the QAM symbol rate and that the coefficients of the prototype are real valued. The total number of real valued additions is given by:

$$A_{\text{PC-FB}}^{\mathrm{Tx}} = 2A_{\mathrm{FFT}}(M) + 8M(K-1) + 2M_{\mathrm{f}},$$
$$A_{\text{PC-FB}}^{\mathrm{Rx}} = 2A_{\mathrm{FFT}}(M) + 8M(K-1) + \left(4L_{\mathrm{eq}} - 2\right)M_{\mathrm{f}}.$$

The second approach is a frequency domain filtering, a.k.a. frequency spread [4] based FBMC, featuring also a general prototype with length KM. In this case, the structure changes drastically. The subcarrier signals have to be spread over KM frequency domain samples and each of them multiplied by one of the prototype frequency domain coefficients. The overlapping parts in frequency domain are all added and then transformed with the IFFT of size KM and finally an overlap and add of size $M/2$ is performed to generate the time domain signal. At the Rx side, the inverse operations are done resulting in the following complexity

$$C_{\text{FS-FB}}^{\mathrm{Tx}} = 2C_{\mathrm{FFT}}(KM) + 4M_{\mathrm{f}}(KM-1),$$
$$C_{\text{FS-FB}}^{\mathrm{Rx}} = 2C_{\mathrm{FFT}}(KM) + 8M_{\mathrm{f}}(KM-1),$$

where we have taken into account that the equalizers coefficients can be incorporated in the frequency domain coefficients of the filters.

Additions for the general prototype structure

$$A_{\text{FS-FB}}^{\mathrm{Tx}} = 2A_{\mathrm{FFT}}(KM) + 2(M_{\mathrm{f}} - 1)(2KM - M_{\mathrm{f}})$$
$$A_{\text{FS-FB}}^{\mathrm{Rx}} = 2A_{\mathrm{FFT}}(KM) + 4M_{\mathrm{f}}(KM-1)$$

Frequency spread based FBMC with a prototype of length KM and designed using the frequency sampling approach with only $2(K-1)$ non-zero coefficients has the following complexity:

$$C_{\text{FS-FB}}^{\mathrm{Tx}} = 2C_{\mathrm{FFT}}(KM) + 8M_{\mathrm{f}}(K-1)$$
$$C_{\text{FS-FB}}^{\mathrm{Rx}} = 2C_{\mathrm{FFT}}(KM) + 16M_{\mathrm{f}}(K-1)$$
$$A_{\text{FS-FB}}^{\mathrm{Tx}} = 2A_{\mathrm{FFT}}(KM) + 4M_{\mathrm{f}}(K-1)$$
$$A_{\text{FS-FB}}^{\mathrm{Rx}} = 2A_{\mathrm{FFT}}(KM) + 2M_{\mathrm{f}}(2K-1)$$

C. UFMC/UF-OFDM/Filtered CP-OFDM

The UFMC system can be parametrized between two extremes: in one end, one single CP-OFDM signal is filtered by one filter to reduce the out-of-band radiation. At the other end each resource block is transformed with the IFFT and filtered with its own filter. In an UFMC system with maximum granularity, B resource blocks each with M_B subcarriers require B FFTs of size M, where each of them has only M_B non-zeros inputs.

The modulation is performed in the following steps: First the signal of each subband is spread over the whole symbol length and transformed into the frequency domain. Then the filtering is performed in the frequency domain and the sum of all subbands is converted into the time domain [7].

Instead of filtering and then transforming, a non-matched filtering is applied in the frequency domain [6]. The Rx has then 3 steps:

- Windowing in the time domain
- FFT transformation of size $2M$ with zero padding and half of the outputs thrown away
- Frequency domain filtering and equalization.

The total number of multiplications is then given by:

$$C_{\text{UFMC}}^{\text{Tx}} = B(C_{\text{FFT}}(2M) + C_{\text{FFT}}(M) + 12M) + C_{\text{FFT}}(2M)$$

$$C_{\text{UFMC}}^{\text{Rx}} = C_{\text{FFT}}(2M) + 8M + 4M_{\text{f}}$$

And the total number of additions by:

$$A_{\text{UFMC}}^{\text{Tx}} = B\big(A_{\text{FFT}}(2M) + A_{\text{FFT}}(M)\big) + 2(B-1)2M + A_{\text{FFT}}(2M)$$

$$A_{\text{UFMC}}^{\text{Rx}} = A_{\text{FFT}}(2M) + 4M + 2M_{\text{f}}$$

D. *Generalized frequency division multiplexing (GFDM)*

The GFDM modulation scheme is based on circular convolving each subcarrier in a block of data with a filter kernel. In contrast to OFDM a cyclic prefix is added per block and not per symbol [8] [9]. Since a circular convolution can be calculated as a multiplication of two vectors in frequency domain the transmitter and receiver can be efficiently implemented using the FFT.

Out of a total number of subcarriers M only M_f are used. N symbols per subcarrier are combined to form one transmission block. In total NM_f data symbols can be transmitted per block. The prototype filter is designed to overlap with M_a adjacent subcarriers and it is typically chosen to be an RRC filter $M_a=2$.

As described in [8], excluding the trivial operations like reordering, the following signal processing tasks need to be performed at the transceiver:

- Transformation of the data signal of each subcarrier into the frequency domain
- Filtering in the frequency domain
- Transformation of the signal into the time domain

The complexity at Tx is given by:

$$C_{\text{GFDM}}^{\text{Tx}} = M_{\text{f}}C_{\text{FFT}}(N) + 4M_{\text{f}}M_{\text{a}}N + C_{\text{FFT}}(NM) \;,$$

$$A_{\text{GFDM}}^{\text{Tx}} = M_{\text{f}}A_{\text{FFT}}(N) + A_{\text{FFT}}(NM) + \big((M_{\text{f}} - 2M_{\text{a}})(2M_{\text{a}} - 1) + M_{\text{a}}(M_{\text{a}} + 1)\big)N + 2M_{\text{f}}M_{\text{a}}N.$$

The details of the corresponding receiver are described in [9]. It is important to mention that since the subcarriers are overlapping it is necessary to cancel this interference to achieve a sufficient performance. In [9] the authors use the detected symbols to subtract the interference to adjacent subcarriers in an iterative fashion. For a constellation as large as 64QAM it was shown that $J = 8$ iterations are sufficient. The receiver can be divided into the following signal processing tasks:

- Transformation of the signal into the frequency domain
- Channel equalization
- Filtering in the frequency domain
- Iterative interference cancelation

The complexity at the receiver (Rx) is then given by:

$$C_{\text{GFDM}}^{\text{Rx}} = C_{\text{FFT}}(NM) + 4\big(M_{\text{f}} + 2(M_{\text{a}} - 1)\big)N + 4M_{\text{f}}M_{\text{a}}N + JM_{\text{f}}(2C_{\text{FFT}}(N) + 4N),$$

$$A_{\text{GFDM}}^{\text{Rx}} = A_{\text{FFT}}(NM) + \Big((M_{\text{f}} - 2M_{\text{a}})(2M_{\text{a}} - 1) + M_{\text{a}}(M_{\text{a}} + 1) + \big(M_{\text{f}} + 2(M_{\text{a}} - 1)\big)\Big)2N + 2M_{\text{f}}M_{\text{a}}N + JM_{\text{f}}(2A_{\text{FFT}}(N) + 3N).$$

E. *Numerical evaluation*

Table 1 shows some examples of the total number of operations involved in each of the MC waveforms presented above. The multiplications and additions are presented per transmission block. The column operations per sample show the total amount of operations (sum of multiplications and additions) required to generate a single time domain sample. The parameters were chosen as follows

$$M = 2048; M_{\text{f}} = 1224; L_{\text{CP}} = M/8\,;$$
$$K = 4; L_{\text{eq}} = 4; M_{\text{B}} = 12; B = M_{\text{f}}/M_{\text{B}}\,;$$
$$L = L_{\text{CP}} + 1; N = 4; M_{\text{a}} = 2; J = 8;$$

Table 1

MC waveform	**Total number of real valued operations**		
	Multiplications	***Additions***	***Operations per sample***
CP-OFDM	42381	141003	79.59
PC FBMC	155568	372704	257.95
FS FBMC high	120637104	78364160	97168.59
FS FBMC low	415824	1211488	794.59
UFMC	8070748	21298380	12747.02
GFDM	418472	1143040	184.83

III. Experimental Implementations

A. *Implementation of an FBMC-based PMR system for broadband PPDR applications*

Radio communications is a critical service enabler for public protection and disaster relief (PPDR) users such as the police and the fire and rescue services. The evolution of PPDR radio systems incorporates every time more data-centric end-use applications, demanding likewise larger data capacities, spectral agility, flexible operation and field-reconfigurability. Specific applications include situational awareness, video streaming, facial recognition, rapid access to databases, collection of sensor data, context awareness, real-time picture retrieval-distribution and complex remote management procedures.

The PPDR are critical services that require a premium radio propagation and a rich coverage. Historically the 400 MHz band has been selected and used exactly to satisfy these needs, since it is tailored for stringent propagation scenarios in dense urban areas. Moreover, the 400 MHz band enables the deployment of large cells in rural areas, helping to reduce the number of sites required to guarantee a given coverage.

Taking into account that the licensed spectrum of the 400 MHz band is fragmented and underutilized, major PPDR players move towards a spectrum reuse approach in order to accommodate broadband PPDR services. There are two strategies available to achieve the latter; in the first re-farming of existing fragmented spectrum is applied to make available contiguous unused spectrum; a second way that does not involve the lengthy and cumbersome spectrum re-farm procedure, is to utilize waveforms that efficiently occupy the spectral gaps left from existing primary narrowband PMR systems, without provoking interference to the adjacent transmissions.

In this paper we focus on the second strategy and demonstrate the benefits of new spectral agile waveforms, based on FBMC design principles, applied to a realistic fragmented spectrum cohabitation scenario. The FBMC waveform is compared with CP-OFDM (LTE), which is currently on pilot use in multi-standard base station platforms in the 400 MHz band, combining PMR with LTE technology. Such network elements enable PMR networks to add broadband services to their existing TETRA/TETRAPOL systems. The FBMC DL prototype is based on a real-time implementation that made use of two different software defined radio (SDR) design frameworks and was validated using a complex hardware setup.

1) Specifications

The structure of the DL FBMC frame features a certain degree of similarity with the specifications of the CP-OFDM LTE (release 9). A bandwidth of 1.4 MHz was selected to reduce the processing load. Each FBMC symbol comprises 72 active subcarriers with a 15 kHz spacing. This results in a 10 ms long radio frame consisting of 10 subframes. The frame thus includes a total of 150 symbols for the FBMC waveform and 120 for the CP-OFDM one. In the FBMC case, the first 3 symbols in a frame contain the preamble with a specific structure that allows for frame synchronization under fragmented spectrum scenarios. The pilot pattern in the FBMC frame resembles the MBSFN structure of LTE. An “auxiliary pilot” was employed to eliminate the influence of the surrounding data on the received pilot. The FBMC design is based on fast convolution, featuring a short transform length of 8 FFT bins per subcarrier spacing (i.e., 16 points) combined with a long transform length of 1024 points. A non-overlapping block length of 10 samples ensured good in-band and out-of-band interference levels. The PHYDYAS prototyping filter [5] was employed for this FBMC scheme.

2) Implementation

The FBMC DL transmitter and receiver were developed using two SDR programming approaches. The broadband DL FBMC receiver was built using a conventional SDR framework, in which a RF transceiver unit is interfaced with a high-end general purpose computer (GPP) implementing the baseband processing functions. The transmitter was implemented using a hardware-assisted SDR framework targeting a FPGA-based SoC device. Special focus is given on this SDR programming approach, since it combines the best of the two worlds (i.e., hardware and software) and constitutes a

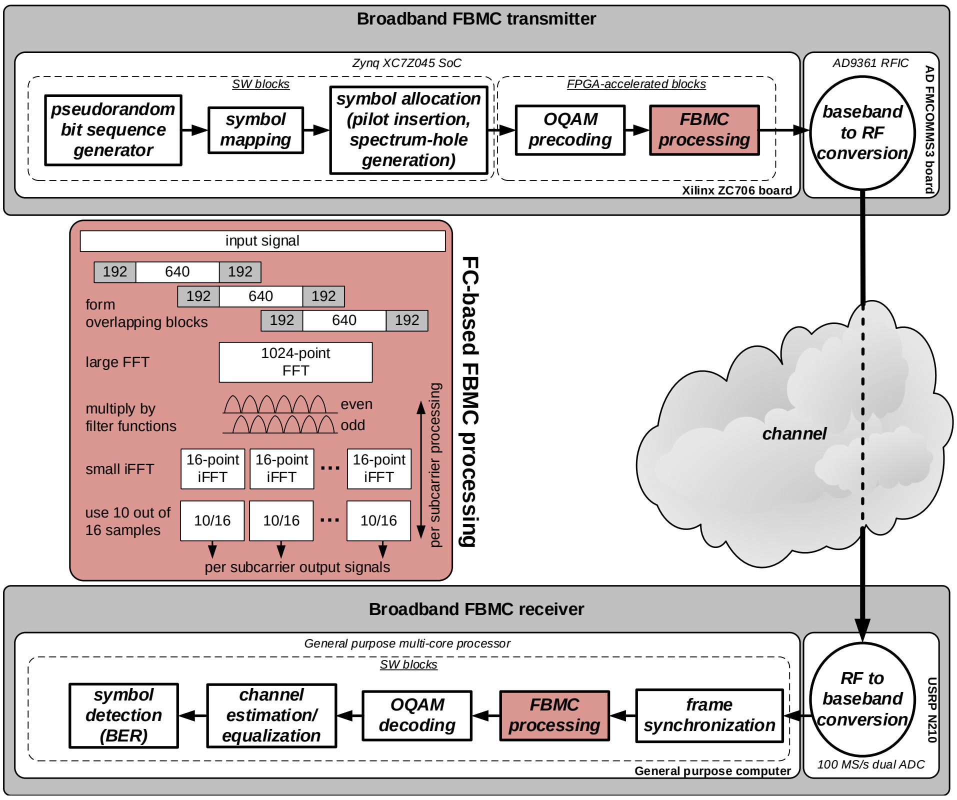

Fig. 1. A high-level block diagram of the DL FBMC SDR implementation.

principle target to architect future processing demanding SDR systems with programming agility requirements. A high-level block diagram of the functions comprising the implemented FBMC transmitter and receiver is shown in Fig. 1.

A balanced partitioning of the baseband functions allowed to exploit the benefits of the SoC device, which apart from the DL FBMC transmitter also hosted a custom implementation of a CP-OFDM (LTE) transmitter. The bit-intensive DSP functions were implemented in the FPGA area of the device (e.g., synthesis filter-bank), whereas the functions related to the generation of the dynamically reconfigurable frame were implemented in the embedded microprocessor. The latter runs an embedded Linux distribution. A custom digital design employing advanced register transfer level (RTL) design techniques such as latency-aware storage, clock-gating and resource sharing, allowed to minimize the FPGA-area processing and the utilization of embedded memory elements. Resource sharing was enabled by applying time division multiplexing, which allowed to reuse a single instance of a 16-point FFT engine and a single instance of a 1024-point IFFT engine. The real-time implementation targeted the Xilinx XC7Z045 SoC device and the resulting utilization metrics are shown in table 2. Details of the SDR implementation of the receiver can be found on [10].

Table 2

System	**Xilinx Zynq-7000 AP SoC (xc7z045ffg900-2)**			
	Slices	***DSP48E1***	***RAMB18E1***	***RAMB36E1***
LTE	6%	2%	1%	1%
FBMC	6%	5%	1%	4%

A hardware-accelerated SDR implementation strategy could be adopted to exploit the massive computational capacity provided by purpose-built FPGA co-processors, in order to alleviate the burden of computationally demanding SDR PHY-layer functions. The implementation of the FBMC and CP-OFDM (LTE) transmitter requires only a small fraction of the FPGA resources embedded within the target SoC device.

3) Hardware setup

The implemented broadband DL FBMC system, or its CP-OFDM (LTE) counterpart, coexists with a narrowband PMR transmission at the 400 MHz. The hardware setup that allowed the testing and validation of the two systems is shown in Fig. 2 and comprises the following equipment, devices and boards:

Transmitters:

- A vector signal generator (Agilent E4438C ESG) emulated a real-time PMR transmitter, by cyclically reproducing a known TETRAPOL I/Q frame-sequence and providing the RF signal centered at 382.53 MHz.
- The Xilinx ZC706 board hosted both the DL FBMC and LTE real-time baseband implementations. The Xilinx board carried the AD-FMCOMMS3 RF transceiver from Analog Devices, which provided the FBMC or LTE RF signal centered at 382.88 GHz.

Channel:

- The RF signals were fed to the Elektrobit Propsim C8 real-time channel emulator, which recreated realistic channel propagation conditions and applied mobility effects. The signals at the output of the channel emulator were driven through a series of RF step attenuators to control the power ratio between the two coexisting transmissions. The RF signals were then combined and two replicas of the combined RF signals fed to the FBMC and PMR receivers.

Receivers:

- A TETRAPOL terminal (M9620SG2) was connected through a custom cable to a computer, where a software application calculated the average bit error rate (BER) observed at the terminal for a given number of frames.
- The Ettus Research USRP N210 device was used as the RF receiver and signal converter providing the complex baseband samples to a computer. The FBMC receiver SDR implementation ran on a laptop, where the frame-based BER metrics were calculated and stored.

4) Results

A measurement campaign was carried out with the aim to experimentally evaluate the performance of the primary narrowband TETRAPOL system when coexisting with a secondary broadband in-band transmission (LTE or FBMC). The signal produced by the real-time broadband transmitter was adapted at run-time, to leave an adequately-sized spectral-hole where the TETRAPOL signal was located; switching

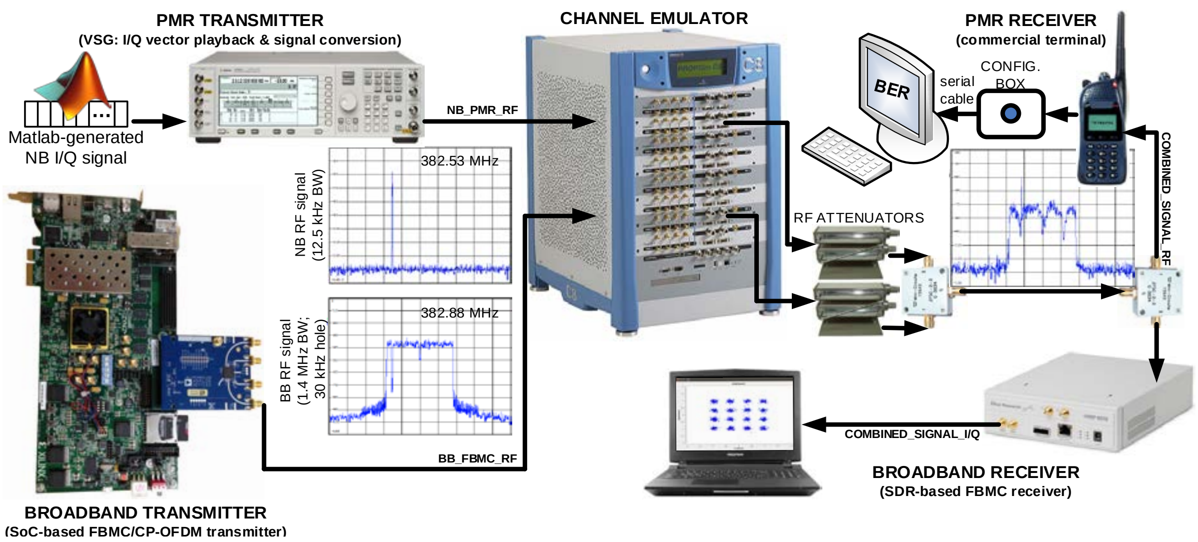

Fig. 2. A representation of the complete experimental setup.

between the FBMC or CP-OFDM waveforms was also enabled. Furthermore, in order to cover a wide range of realistic operating conditions, the hardware setup was precisely calibrated to yield a precise subcarrier power ratio between the two coexisting DL transmissions. The channel emulator was configured with different channel models, including among others, the ITU vehicular A (VA) channel at 50 km/h, the 3GPP extended pedestrian A (EPA) channel model at 3 km/h and the 3GPP typical urban (TU) channel at 50 km/h.

To evaluate the performance of the narrowband TETRAPOL receiver the BER was calculated by averaging the captured values of 10.000 TETRAPOL frames for each of the selected test configurations. In more detail, each BER value represents a different combination of a time varying channel with a given subcarrier power ratio between the coexisting signals, for a fixed power of the received signal of -98 dBm at the TETRAPOL terminal. Moreover, the broadband transmission featured a 30 kHz spectrum hole, which provided an optimum tradeoff between spectral efficiency and interference protection, as a single 12.5 kHz TETRAPOL DL communication was considered.

As it can be observed in Fig. 3, under the considered operating conditions, the FBMC system offers superior interference protection, while transmitting at a much higher subcarrier power-level (up to 29 dB) when compared to its CP-OFDM counterpart. It is important to highlight that the observed gains of the FBMC waveform when compared with the LTE one in terms of sensitivity to baseband interferences measured at the TETRAPOL terminal, were maintained under all evaluated channel conditions (i.e., model, speed).

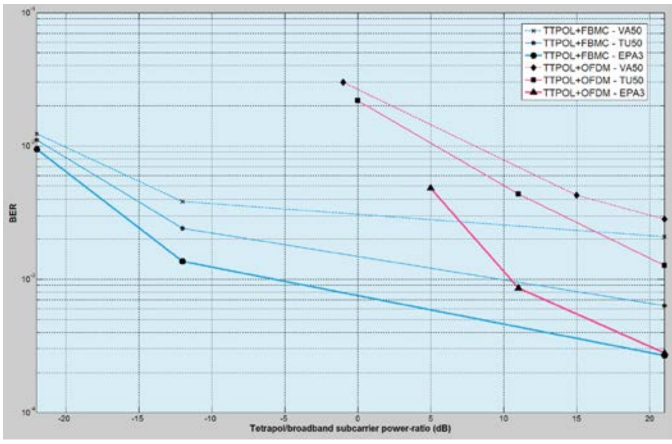

Fig. 3. Performance of the PMR terminal evaluated in relation to the interference received from the in-band broadband transmissions

B. *A flexible radio transceiver for TVWS based on FBMC*

Dynamic spectrum sharing has been proposed to improve spectrum usage. The digital switch over (DSO) in TV bands, which has resulted in making the so-called TV white space (TVWS) UHF spectrum available, was the first actual example where such a mechanism has been allowed. In 2009, the US radio regulator – the federal communication commission (FCC) – authorized opportunistic unlicensed operation in the TV bands [11]. The initiative was recently followed by the UK regulator (Ofcom) [12] and in Japan where the Ministry of Internal affairs and Communications has published rules for secondary operation in TVWS [13]. In this context, opportunistic communication systems have to coexist with incumbent systems, i.e.: TV broadcast signals. The coexistence scheme is enforced with a priority mechanism where opportunistic systems must guarantee that no 'harmful interference' will be incurred to the incumbents.

Harmful interference is defined in a twofold way. Firstly, co-channel communication between incumbent and opportunistic systems is prohibited. This means that opportunistic systems must be able to assess the presence of incumbent signals and access only channels vacant from any incumbent. Besides, opportunistic systems have a limited amount of time to evacuate the channel when an incumbent is switched on. Secondly, the ACLR is limited in order to prevent an opportunistic system from interfering with an incumbent operating in another channel, and in particular in adjacent channels. In [11], ACLR is restricted to be at least 55dB. Similar requirements are considered in other countries (e.g. in the UK [12]). Such a high ACLR requirement is specific to the TVWS context. For instance, ACLR requirement is 10dB stronger than required for LTE systems [14].

Based on the analysis of the standards available at that time, it was concluded that there are some missing points that could be addressed by a new standardization effort [15]. Most white space standards are based on CP-OFDM physical layer (ECMA392, IEEE802.22, IEEE802.11af). These standards are the result of adaptations of CP-OFDM based standard to the white space constraints; however, because of the very stringent requirement on ACLR and the dynamic nature of dynamic spectrum access (DSA), alternative approaches based on FBMC have been proposed [16]. These requirements of flexibility and stringent ACLR have lead IEEE DYSPAN Standard Committee to identify the necessity to develop a new standard defining radio interface for white space radio systems: IEEE 1900.7 standard [17]. The standard is based on a Filter Bank Multi Carrier (FBMC) PHY.

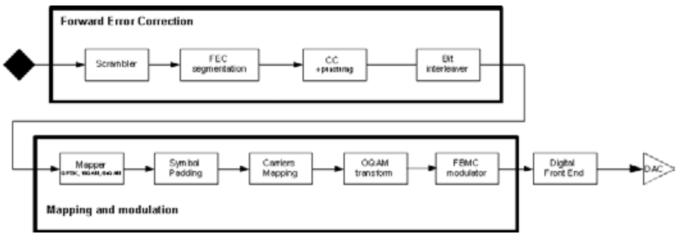

Fig. 4. Block diagram of an IEEE 1900.7 baseband transmitter.

The IEEE 1900.7 standard specifies two different sizes for the prototype filter: K=3 or K=4 governing the level of protection of adjacent channels. The block diagram of the IEEE 1900.7 transmitter is shown in Fig. 4. The transmitter architecture is composed of two main elements: forward error correction (FEC) block and data mapping and modulation block. FEC is implemented using a standard convolutional encoder. The code may be punctured to support variable encoding rates. The convolutional code is segmented by blocks of fixed size. The output of the encoder is forwarded to a bit interleaver of size multiple of the output length of the encoder.

The second block maps and modulates the encoded bits to the multicarrier modulation. The coded data are mapped to QPSK, 16QAM or 64QAM modulation symbols, depending on the desired data rate. Symbols are then padded to make the transmitted burst length an integer multiple of multicarrier symbols. The generated block of QAM symbols is mapped to active carriers and modulated to an offset-QAM before being transformed into a time domain signal using FBMC waveform.

Through an implementation on a flexible hardware TVWS transmitter, [18] showed that FBMC modulation can meet ACLR levels prescribed by the FCC's coexistence requirements. The actual implementation was aimed at assessing the performance under real hardware impairment conditions, such as limited dynamic Digital to Analog Convertors (DAC).

One of the main shortcomings of FBMC was supposed to be its implementation complexity. However, recent results have shown that a flexible approach was possible with very limited complexity overhead. First, on the transmitter side, it has been shown that an additional filter is required on top of CP-OFDM and shall be implemented as a digital filter for the sake of flexibility. In this context, [19] concluded that the complexity of the filtered CP-OFDM TX could be as high as or even higher than the one of the FBMC TX, which does not require this additional filtering.

Similarly, the complexity of a flexible FBMC receiver has been compared to the one of CP-OFDM using the same carrier spacing and multicarrier symbol duration as for the 10MHz LTE mode. The complexity has been evaluated for a Xilinx Kintex-7 FPGA and is given by the amount of Slice Registers, LUT and DSP48E1 cells used by the different modules of the receiver design. Slice Registers correspond to the amount of register cells used, while LUT to the amount of combinatorial logic in the design. DSP48E1 cells are DSP cells dedicated to multiplication and accumulation (MAC) operations.

Table 3. Complexity comparison for FBMC implementation on Xilinx FPGA.

	Complexity Evaluation			
Architecture	***Slice Regs***	***LUTs***	***DSP48E1***	***RAM BLKS***
OFDM Transmitter	10262	6752	14	14
FBMC Transmitter	11300	7990	30	19
FBMC Transmitter Complexity Overhead	10 %	18 %	114 %	36 %
OFDM Receiver	42574	39600	97	49
FBMC Receiver	54970	50096	155	171
FBMC Receiver Complexity Overhead	29 %	27 %	60 %	249 %

The results of [19] have been reported in Table 3 and have shown that the computational complexity of the FBMC transmitter is very similar to the OFDM transmitter one in this context. Furthermore, the receiver was only around 30% higher than the one of the CP-OFDM RX. In addition, the proposed block-wise processing approach requires FBMC symbols to be stored, which impacts the size of the memory (2.5x the one of an equivalent CP-OFDM RX). However, such memory sizes can be implemented at a very limited footprint and cost on current silicon technology nodes.

IV. Conclusion

Different technologies are currently considered to be adopted in the forthcoming generation of mobile broadband. Among them, new spectrum agile waveforms and efficient use of underutilized fragmented spectrum (featuring opportunistic radio use) are two key 5G enablers that are likely to be adopted due to the concrete interest of major industry stakeholders [10]. Despite the disadvantages of CP-OFDM radio solutions, the latter have been widely adopted until now, due to their reasonable DSP complexity, their simplicity and low cost. However, the constant advances in baseband processor technologies, offer nowadays massive processing parallelism and highly flexible operation, satisfying strict energy budgets at a reduced silicon size, a fact that opens up the way to more complex and spectrally efficient waveforms. In addition, various MC schemes based on FBMC have recently been proposed or revisited having as a goal to optimize the DSP complexity.

In this paper we have presented results that reveal the feasibility of opting for FBMC as a candidate 5G waveform, especially when planning to deploy secondary systems in existing underutilized (and spectrally fragmented) bands, where spectral efficiency and interference protection of primary transmissions is a must. Apart from the analytical results, two different DL FBMC systems were implemented and validated when coexisting at the same band either with primary PMR or TVWS transmissions. Likewise, it was demonstrated in practical terms the suitability of FBMC for such use cases and also it was proved that the top-up DSP cost of FBMC - when compared with CP-OFDM -, is not significant enough to prevent the adoption of this technology.

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ABOUT THE AUTHORS

Nikolaos Bartzoudis (CTTC) is a senior researcher and head of the PHYCOM department at CTTC. He received his Ph.D. degree in electronic engineering from Loughborough University (UK 2006). Before joining CTTC (January 2008), Nikolaos has worked as Senior Research Officer (University of Essex, 2005-2007). He is currently involved (technical leader) with baseband development and experimental validation of spectral and energy efficient 4G-5G wireless communication technologies. His research interests include high-speed digital design, hardware-software co-design and reconfigurable SDR. In the past he was the principal investigator in 2 R&D projects and participated in 10 others (industrial, FP7/H2020, EPSRC). Currently he coordinates the AEThER project.

Vincent Berg (CEA) is a research engineer at CEA-LETI working on physical layer development and implementation. He graduated from Supélec, a leading French "grande école" of engineering, and obtained a M.Sc. in Microwaves from University College London in 1995. After graduating, he joined the industry on the development of transceivers for wire and wireless applications. Since 2010, he is working for CEA-LETI at the Wireless and Security Communications Department where he is currently leading the wireless solutions and digital prototyping laboratory. He actively participates to the FBMC physical layer architecture and implementation proposals.

Leonardo Gomes Baltar (Intel) is a research engineer at Intel Deutschland GmbH. He received B.Sc. and M.Sc. degrees in Electrical Engineering from the Federal University of Rio de Janeiro (UFRJ), Brazil, in 2004 and 2006, respectively. From 2006 to 2015 he was at the Technical University of Munich (TUM) as a research and teaching assistant. He has been involved in the European FP7 projects PHYDYAS and EMPHATIC and in projects with industrial partners. His research activity has been mainly in the area of signal processing techniques applied to wireless and wired communications, with special focus on multicarrier transmission, multirate systems, filter banks and block transforms.

Oriol Font (CTTC) received his M.Sc. degree in Computer Engineering from Universitat Autònoma de Barcelona (UAB) in 2004 and the Ph.D. at the Department of Signal Theory and Communications of the Universitat Politècnica de Catalunya (UPC) in 2013. In 2006 he also obtained a Master in Design of Intregrated Circuits from UAB. He joined the CTTC in November 2006, where he currently holds a Researcher position in the PHYCOM department. He has participated in several national and European R&D projects. His research interests include FPGA-assisted SDR, hardware-software co-design and dynamically reconfigurable PHY-layer implementations towards energy and spectral efficient 5G systems

Kilian Roth (Intel) is a Ph.D. student at Intel Deutschland GmbH. He graduated from Munich University of Applied Sciences (MUAS) and obtained a M.Sc. at the Technical University of Munich (TUM) in 2014. Since 2014, he is pursuing a Ph.D. degree at the Chair of Circuit Theory and Signal Processing at TUM. His main research interests are signal processing algorithms and systems concepts to enable energy efficient mobile broadband systems at millimeter wave frequencies.

Miquel Payaró (CTTC) received the Ph.D. degree from the Universitat Politècnica de Catalunya in 2007. From February 2007 to December 2008 he was a Research Associate at the Hong Kong University of Science and Technology. Since January 2009 he is with CTTC where, in 2013, he became a Senior Researcher and was appointed the Head of the Communications Technologies Division. Miquel has led several research contracts with the industry and also has participated in publicly funded research projects (FP7, H2020), e.g., BuNGee, BeFEMTO or Newcom#. From July 2015, Miquel is serving as Technical Manager in the 5G-PPP project Flex5Gware.

Dipl.-Ing. Michael Färber (Intel) received his degree from “FH der DBP” 1985, and joined Siemens as RF engineer. Michael started 1990 in ETSI SMG2 on GSM standardization. 1995 he did GPRS concept work and was elected vice-chairman of SMG2. 1998 he was chairman of the EDGE group. He served 10 years as vice-chairman for the GSM physical layer group. Since 2007 he was with NSN, active for HSPA evolution and LTE. Michael participated in EU FP7 ARTIST4G as WP leader. Michael joined Intel 2012. Michael is Intel’s Director EU wireless research for 5G EU activities, and coordinator of the FP 8 Flex5Gware project.