Efficient Design of OFDMA-Based Programmable Wireless Radios

S. F. A. Shah and A. H. Tewfik

Department of Electrical & Computer Engineering, University of Minnesota, Minneapolis, MN 55455, USA

Correspondence should be addressed to S. F. A. Shah, sfaisal@umn.edu

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With the increasing demand for efficient spectrum management, programmable wireless radios can potentially play a key role in shaping our future spectrum use. In this paper, we consider the design of low-power programmable wireless radios based on orthogonal frequency division multiple access (OFDMA). To meet the demands of higher data rate communications, we split OFDMA symbols carrying multiuser data across several noncontiguous bands of available spectrum. To relax power consumption in analog-to-digital and digital-to-analog converters, we use a programmable narrowband RF front end comprising of programmable synthesizers and fixed low-pass filters. To perform digital baseband signal processing in an energy efficient manner, we propose efficient designs for the fast Fourier transform (FFT) and inverse FFT (IFFT) modules. Our designs of the FFT/IFFT modules reduce power consumption and chip area, and are capable of handling the dynamic nature of spectrum in programmable radios. To recover data that falls within the transition band of the filters, we propose a combiner similar to maximal ratio combiner. We also present the complete design of programmable wireless radios in accordance with the IEEE 802.22 (draft) standard.

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1. INTRODUCTION

Over the last several years, the demand for spectrum has continued to grow as new and innovative wireless services are being made available to the public. To meet the ever increasing demand for radio spectrum, and to facilitate new technologies and services, the FCC is in the phase of revising its regulations and policies to enable an efficient use of spectrum. In 2003, the FCC issued a notice of proposed rule making to expand the ability of wireless spectrum licensees to enter voluntary transactions to make all or part of their spectrum usage rights available for new uses through “smart” or cognitive radio (CR) systems [1]. These initiatives from the FCC have spurred tremendous research on cognitive radio within academic and industrial forums, and standard bodies. A typical example of under utilized spectrum is the TV band from 54 MHz to 862 MHz. The IEEE 802.22 task group is currently developing specifications for the use of CR technology in the TV band to provide broadband wireless access [2].

Another driving force for CR systems is the Defense Advanced Research Projects Agency’s (DARPA) neXt Genera-
and robustness against multipath fading [5]. Orthogonal frequency division multiple access (OFDMA) extends the idea of OFDM to multiuser environment by assigning multiple subcarriers to different users for simultaneous transmission [6].

The dynamic spectrum access requirement of a CR poses several unique challenges to the low-complexity design of its PWR unit. Our main goal in this paper is to provide a complete design architecture of OFDMA-based PWR's that includes RF/analog front end (AFE) with data converters and digital baseband processor as shown in Figure 1.

To relax power consumption in analog-to-digital converter (ADC) and digital-to-analog converter (DAC), we use multiple programmable narrowband RF front end comprising programmable synthesizers and fixed analog low pass filters (LPFs).

On the digital front, designing a wideband OFDMA system for PWR is challenging due to the high complexity of the FFT/IFFT modules used in an OFDM based multiple access system. The naive approach of nullifying the subcarriers by zero-padding the data blocks does not utilize the resources efficiently [7]. There are several classical papers on efficient computation of the discrete Fourier transform (DFT) when only a subset of input or output has nonzero values (see, e.g., [8] and the references therein). Sorensen and Burrus [8] referred to their approach as *transform decomposition* (TD) that is based on the Cooley-Tukey decomposition with some modifications. Although the TD method is more flexible when compared to the pruning techniques, we observe two major issues with this approach. Firstly, the complexity of the TD approach increases when the nonzero input points are noncontiguous. Secondly, the TD method was developed mainly for software computation of the DFT and did not consider the hardware scalability implications when the number of nonzero values are dynamic as in CR. We will have more to say about these issues and our proposed solution in Section 4.

In the context of CR, the authors in [9] used the pruning technique to design efficient FFT modules when only a subset of outputs is needed. However, it is well known that the pruning of FFT does not provide efficient solutions due to if statements [8]. In another recent work [10], the authors applied the TD method to DFT based OFDM systems but did not address the above mentioned problems associated with the TD method. Furthermore, they failed to observe the fact that the TD approach yields lower complexity results when the number of nonzero inputs are below a certain threshold. For this reason, they did not discuss any implications of using TD approach in CRs. For low-complexity FFT/IFFT modules, we propose a modified Cooley-Tukey decomposition that solves the increased complexity issue of the TD method. We also discuss the hardware implementation from the scalability perspective.

Our main contributions are (i) the complete design of PWR’s that includes RF section and digital baseband processor and the impact of the choice of AFE on baseband processor; (ii) efficient IFFT/FFT modules based on a modified Cooley-Tukey decomposition and their scalability issues; (iii) data recovery during filter transition bands; (iv) a specific example of low-complexity PWR based on the IEEE 802.22 standard (draft). The paper is structured as follows: in Section 2, we emphasize the role of PWR in dynamic spectrum access and discuss OFDMA as a perfect fit for physical layer transmission. We discuss RF front-end alternatives in Section 3 and justify the selection of channelized architecture for PWR. We describe the design of low-complexity IFFT/FFT modules for OFDMA systems in Section 4. We measure the computational complexity of our designs in terms of the number of complex multiplications. This number is a good measure of power consumption and area in digital circuits. We also discuss the scalability issues of the IFFT/FFT modules that are very pertinent in the context of CR for dynamic spectrum access. In Section 5, we discuss the effect of analog filter transition bands on data recovery and propose a solution inspired from maximal ratio combining. In Section 6, we use the IEEE 802.22 standard as an example system and discuss the design of low-complexity PWR for such systems. We provide the conclusions in Section 7.

### 2. PWR AND DYNAMIC SPECTRUM ACCESS

In general, the availability of spectrum is a function of *geographic area* and *time*. With the FCC pushing for more intensive and efficient use of spectrum, the next generation of wireless communication systems is being designed to use the spectrum in a dynamic manner. The term dynamic spectrum access has different connotations in different contexts [11]. In this paper, we are mainly concerned with the scenario...
where a CR dynamically monitors certain bands of the licensed spectrum (as allowed by FCC, e.g., the unlicensed operation in the TV band[12]), finds idle spectrum and uses it as needed. It is mandatory for the cognitive radios not to cause any interference to the primary users of the spectrum. This scheme of dynamic spectrum access is also known as opportunistic spectrum access[11]. A typical example of dynamic spectrum access scenario is shown in Figure 2.

As mentioned earlier, our emphasis is on a design that enables the opportunistic spectrum access in a low-complexity paradigm and we will not consider the subject of spectrum sensing in this paper. We assume that the PWR transmitter has the knowledge of available spectrum bands through its sensing unit. To achieve the goal of dynamic spectrum access in a PWR, we select OFDM as the underlying technology to transmit the information over hostile wireless channels. Due to its robustness against multipath distortion and ease of implementation, OFDM has been successfully used in wireless local area networks based on the IEEE 802.11a and g standards[13]. In particular, OFDM provides a flexibility of assigning the user data to different subcarriers and it is a perfect fit for PWR with dynamic spectrum access.

To support a multiuser environment, traditional methods of statistical multiplexing, like TDMA, can be used in conjunction with OFDM. However, the intrinsic multicarrier nature of OFDM provides a unique opportunity to share the spectrum with multiple users with very fine granularity. The multiuser version of OFDM is commonly known as orthogonal frequency division multiple access (OFDMA)[6]. In OFDMA, single or multiple groups of subcarriers are allocated to each user that can transmit simultaneously. Further, OFDMA has been selected as one of the physical layer multiple access technologies in the recent wireless communication systems [2, 6]. The frequency diversity available in wideband OFDMA systems leads to optimal subcarrier assignment based on propagation and interference conditions.

In general, OFDMA with adaptive modulation and coding allows for fine granulation of bandwidth, traffic and power allocation that result in optimum and flexible resource management.

Before going into the details of the PWR transceiver architecture, we will discuss two possible scenarios involving the available spectrum and their consequences on the transceiver design. In many scenarios it is helpful to think of the spectrum in terms of basic channels. For example, in the IEEE 802.22 case, a basic channel is a TV channel of bandwidth 6 MHz. To meet the desired quality of service (QoS) of many applications, it is often necessary to use a bandwidth that exceeds that of a single basic channel. Furthermore, in some cases, it may not be possible to find a contiguous band of available frequencies of bandwidth equal to that of one or more basic channels. Hence, two scenarios are possible.

1. Contiguous spectrum (channel bonding)

This refers to the scenario when the available spectrum spans a continuous band of frequencies that can be handled by a single RF circuitry. It is implied that the available bandwidth is sufficient to achieve the desired QoS in the environment. This is also known as channel bonding in the literature of 802.22. For example, a proposal submitted to the IEEE 802.22 working group supports channel bonding in the form of three TV channels that can provide up to 18 MHz of bandwidth.

2. Noncontiguous spectrum (channel aggregation)

In situations, when the spectrum occurs in several chunks over a wide range of frequencies that cannot be covered by a single RF circuitry, we call this scenario a noncontiguous spectrum allocation. This is also known as channel aggregation. The need for channel aggregation arises to meet the QoS especially for users that are farther from the base station.

3. Analog/RF front-end for PWR transceiver

In this section, we discuss different choices available for the analog/RF front-end (AFE) of PWR. Our emphasis is on the designs that minimize power consumption and overall system size requirements. We therefore focus on solutions that move complexity from the AFE to the digital section. In Mitola's[14] software defined radio (SDR), the RF signal was digitized upfront at antenna and digital signal processing was then used to achieve flexibility. However, the ADC needed for this purpose is still far from being practical. In contrast to Mitola's wideband digitized architecture, Abidi[15] proposed a narrowband approach and focused on the receiver RF front-end design that can be tuned for a single channel of 20 MHz located anywhere from 800 MHz to 6 GHz. Although Abidi's narrowband receiver[15] fulfills the definition of SDR in a broad sense, it may not be suitable for future license-exempt cognitive radio systems that will try to use any available spectrum possibly in the form of noncontiguous channels of bandwidth 1–20 MHz.

![Figure 2: A typical dynamic spectrum access model (opportunistic spectrum access).](image)
To reduce the sampling frequency requirement of ADCs, the received analog signal can be channelized either in the time domain or frequency domain [16]. In the time domain, the received analog signal is sampled in time-interleaved manner using an array of $M$ ADCs which are triggered successively at $1/M$ the effective sampling rate. This system suffers from ADC mismatches, clock jitter and narrowband interference [17]. The received analog signal can also be channelized into $M$ frequency subbands using a hybrid filter bank (HFB) consisting of analog synthesis filters, ADCs and digital analysis filters [17]. Due to subband filtering, each ADC can now sample the signal at $1/M$ of the effective sample rate. In theory, the synthesis filter is a bandpass filter, but for system-on-chip (SoC) applications it is extremely difficult to implement these bandpass filters with high center frequencies [16]. To overcome this problem, Namgoong modified the HFB approach and used mixers for frequency translation to baseband followed by simple low pass filters [16]. In the following subsections, we present the architecture of a PWR receiver AFE.

### 3.1. PWR receiver AFE architecture

In the light of the above discussion, we follow the subband channelized design of [16] for PWR RF front-end. However unlike the case of Namgoong’s receiver [16], the application of the hybrid filter bank approach to OFDMA systems does not require any analysis filter bank and the symbol detection can be directly performed using the FFT demodulator. The complete architecture of the PWR receiver is shown in Figure 3. The programmable RF front-end consists of multiple chains of mixer, filter and ADC. The multiple chains of the receiver are selectable based on the available spectrum. Each chain is designed to handle narrowband channels of the order of 20 MHz with center frequency of several GHz. The programmability in the RF front-end is introduced through adjustable mixers that use programmable synthesizers [18]. The mixers translate the passband signal to baseband with zero-IF and produce I and Q branches, each containing a fixed LPF and an ADC.

Another advantage of using narrowband RF chains, though not so obvious, is the resulting low-complexity digital baseband processor. With wideband RF front-end, the channel selection in the digital baseband requires larger FFT modules that consumes chip area and power. With minimum possible narrowband RF, the FFT size can be reduced to a viable design. The architecture for the PWR transmitter is very similar to that of the receiver and will not be discussed here.

The digital baseband processing involves error control coding, mapping to constellation, and efficient IFFT/FFT modulation and demodulation. As discussed earlier, for PWR we need to design an OFDMA system that can dynamically modulate and demodulate data on different subcarriers or group of subcarriers depending on the availability of spectrum. The core of OFDM systems that perform this modulation and demodulation are the FFT/IFFT modules that is the subject of the next section.

### 4. EFFICIENT BASEBAND PROCESSING

We base our design of the low-complexity FFT/IFFT modules on Cooley-Tukey decomposition [19] after modifying it to exploit the fact that there are only a subset of nonzero input or output points. As we have pointed out in Section 1, the transform decomposition (TD) method of [8] did not address the increased complexity and scalability issues when used for CR technology. Our proposed design, specially for the IEEE 802.22 standard as in Section 6, addresses these implications of the Cooley-Tukey decomposition when applied to PWR. In the following subsections, we start with the proposed modifications in the Cooley-Tukey decomposition and then discuss the proposed scalability solution. Before going into the mathematical details, we first interpret the proposed modified Cooley-Tukey decomposition pictorially.

#### 4.1. Modified Cooley-Tukey decomposition

The main idea of the Cooley-Tukey FFT/IFFT algorithm is to map a one-dimensional input sequence of length $N = KL$ into a two-dimensional array with $K$ rows and $L$ columns. This $K \times L$ array is filled column-wise with samples from the
input sequence. The next step of the Cooley-Tukey algorithm is to compute the $L$-point DFT of the $K$ rows. This is followed by multiplication with twiddle factors and then finally computing the $K$-point DFT of the $L$ columns. A pictorial representation of 16-point DFT with $K = 4$ and $L = 4$ is shown in Figure 4.

To exploit the fact that only a subset of input or output points are nonzero and to save on number of multiplications, we can absorb the twiddle factor multiplication with the row-wise DFT or column-wise DFT depending on whether zeros are at the input or the output. This approach has been used in the transform decomposition (TD) proposed by Sorensen and Burrus in [8]. However, for the subset of nonzero input, they only considered the case when the first $K$ input values are nonzero [8, page 1194]. But, as we mentioned in Section 2, in CR systems the subcarriers with nonzero data may be noncontiguous. Thus, the Cooley-Tukey decomposition or the TD approach will result in higher complexity. The problem arises due to the fact that the mapping to 2D array in the Cooley-Tukey algorithm distributes the nonzero data in the array. Consequently, the multiplication with the twiddle factors cannot exploit the fact that there is only a subset of multiplications that results in nonzero product. The multiplication with all the twiddle factors ($NL$ in total) increases the complexity far more than the full $N$-point FFT/IFFT.

To circumvent the issues with the original Cooley-Tukey or the TD approach, we propose a change in the sequence of operations that can reduce the twiddle factor multiplications from $NL$ to $N$. Unlike the Cooley-Tukey or TD approach, we first multiply the incoming serial data sequence with $L$ different twiddle factors. The multiplier unit will be enabled only for the $K$ nonzero values of the input sequence. This limits the twiddle factor multiplications to $N$. We then map each of these $L$ one-dimensional sequences into 2-D arrays followed by row-wise accumulation of the array elements. The last step is similar to the Cooley-Tukey decomposition and performs the $K$-point DFT. Hence, the proposed change in the sequence of operations results in savings in twiddle factor multiplications by a factor of $L$.

### 4.2. Efficient IFFT module for OFDMA transmitter

The transmitter of an OFDMA system uses an IFFT to map the data to subcarriers. Here, we describe the design of an IFFT module that can modulate the data to a single or multiple groups of noncontiguous subcarriers. Assume that $N$ is the total number of subcarriers in an analog band (RF chain) and there are total of $K$ subcarriers available for use such that $N = LK$. (For simplicity, we assume that $N$ and $K$ are powers of 2.) In case the number of zero points $K$ is not a multiple of $N$ or is not a power of 2, we can always set it as the nearest power of 2 that is a multiple of $N$. Thus, we need to design an IFFT module that can efficiently compute $N$-point IFFT knowing that only $K$ inputs are nonzero. It is important to note that the arrangement of these $K$ subcarriers is irrelevant to our design. The $N$-point IFFT operation is defined as

$$x(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X(k) W_N^{-nk}, \quad 0 \leq n \leq N - 1,$$  \hspace{1cm} (1)

where the scalar $1/\sqrt{N}$ is used for normalization and we define the twiddle factor as $W_N^{-nk} := e^{-j2\pi nk/N}$. Unlike the original Cooley-Tukey decomposition that replaces both indices, $n$ and $k$, in (1), we replace only index $n$ in (1) with

$$n = n_1L + n_2, \quad n_1 = 0, 1, \ldots, K - 1, \quad n_2 = 0, 1, \ldots, L - 1.$$  \hspace{1cm} (2)

This is consistent with our proposed modification to the Cooley-Tukey decomposition where we do not map the input sequence $X(k)$ to a 2D array initially. Substituting (2) into (1) gives

$$x(n_1L + n_2) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} [X(k) W_N^{-n_1k}] W_K^{-n_2k}.$$  \hspace{1cm} (3)

We define the multiplication of the input sequence with the twiddle factors as

$$\tilde{X}_{n_1}(k) := X(k) W_N^{-n_1k}$$  \hspace{1cm} (4)

and then map the multiplier output $\tilde{X}_{n_1}(k)$ for every $n_2$ to a 2D array of size $K \times L$ using

$$k = k_1K + k_2, \quad k_1 = 0, 1, \ldots, L - 1, \quad k_2 = 0, 1, \ldots, K - 1.$$  \hspace{1cm} (5)

Substituting (4) and (5) into (3), we get

$$x(n_1L + n_2) = \frac{1}{\sqrt{N}} \sum_{k_1=0}^{K-1} \left[ \sum_{k_2=0}^{L-1} \tilde{X}_{n_1}(k_1K + k_2) \right] W_K^{-n_2k_2}. \hspace{1cm} (6)$$

The term inside the brackets represents the row-wise accumulation of the array elements. If we define $\tilde{X}_{n_1}(k_2)$ as the accumulator output, that is,

$$\tilde{X}_{n_1}(k_2) := \sum_{k_1=0}^{L-1} \tilde{X}_{n_1}(k_1K + k_2),$$  \hspace{1cm} (7)

then (6) reduces to

$$x(n_1L + n_2) = \frac{1}{\sqrt{N}} \sum_{k_2=0}^{K-1} \tilde{X}_{n_1}(k_2) W_K^{-n_2k_2}. \hspace{1cm} (8)$$

Now, (8) represents the $K$-point IFFT of $\tilde{X}_{n_1}$. A block diagram depicting the efficient computation of IFFT using the modified Cooley-Tukey decomposition is shown in Figure 6.
4.3. Computation cost

We measure the computation cost of the proposed design in terms of the number of complex multiplications. There are \( K \times L = N \) complex multiplications needed to compute (4) for all values of \( n_2 \). Since (8) is a \( K \)-point IFFT, it requires \((K/2)(\log K - 2)\) complex multiplications for a given \( n_2 \). Hence the total number of complex multiplications performed by the IFFT module is

\[
C_{\text{IFFT}} = \frac{N}{2} \log K. \tag{9}
\]

That is significantly lower than the naive approach of [7] that requires \((N/2)(\log N - 2)\) complex multiplications. It is simple to show that the IFFT computation using (4) and (8) results in lower complexity only when \( K < N/4 \). Otherwise, the full \( N \)-point IFFT should be used. As an example, consider \( N = 2048 \) (a typical value for IEEE 802.22) and \( K = 2 \) to \( K = 2048 \) such that the ratio \( K/N \) goes from 0 to 1. The ratio \( K/N \) represents the fraction of number of input points with zero value. Figure 7 shows the complexity of the IFFT module in terms of number of multiplications. In Figure 7, we also mark the regimes where the proposed efficient IFFT is preferred over a complete \( N \)-point IFFT. The break even point occurs exactly when the ratio \( K/N = 0.25 \). Certainly, there is a clear advantage in using the proposed decomposition for IFFT when \( K/N < 0.25 \). For instance, at the next possible value of \( K/N = 0.125 \) there is almost 11 in multiplications.

4.4. Efficient IFFT with scalability

One important issue that has not been addressed in the literature is the scalability of the TD approach or Cooley-Tukey decomposition. By scalability we mean that how easy it is to scale (4), (7), and (8) for different values of \( K \)? To answer this, let us consider the implementation details of (4), (7), and (8) as shown in Figure 8. It is obvious that the implementation
of (4) requires simple multiplication that can be easily scaled for different values of $K$. Similarly, (7) requires accumulation of each of $N$ samples $L$ times. Thus, the accumulator implementation is independent of $K$ and can be scaled without any modification in the logic. The major problem is in the implementation of (8) for different values of $K$. To solve this problem, we need additional logic to compute the last few stages of the FFT that are needed due to increase in the value of $K$. For example, if we initially design the hardware of the IFFT module for $K = 256$, then to scale it to $K_{\text{new}} = 512$, we need to compute the last stage of a 512-point FFT. This requires multiplication with 256 twiddle factors and radix-2 butterflies. In general, if $K_{\text{new}} = 2^a K$ then the complexity of the scalable IFFT, originally designed for $K$ nonzero points, is

$$C_{\text{IFFT}} = 2^a \frac{N}{2} \log K + K (1 + C_{2a}), \quad (10)$$

where $C_{2a}$ is the number of complex multiplications needed for the computation of a radix-2$^a$ butterfly. For example, a radix-4 butterfly involves three complex multiplications and $C_4 = 3$. Again, we rely on the intelligence of CR to appropriately select the final path in Figure 8.

4.5. Efficient FFT module for OFDMA receiver

For completeness, we briefly discuss the design of an efficient FFT module for the OFDMA receiver. An OFDMA receiver uses an FFT operation to demodulate the data. The FFT operation is defined as

$$X(k) = \sum_{n=0}^{N-1} x(n) W_N^{nk}, \quad 0 \leq k \leq N - 1. \quad (11)$$

Using the Cooley-Tukey decomposition, the FFT operation at the receiver can be written as

$$X(k_1 K + k_2) = \sum_{n_2=0}^{L-1} \left[ \sum_{n_1=0}^{K-1} x(n_1 L + n_2) W_K^{kn_1} \right] W_N^{n_2 k_2}, \quad (12)$$

where $n_1, n_2$ are defined in (2) and $k_1, k_2$ are defined in (5). Combining the outer twiddle factors in (12), we have

$$X(k_1 K + k_2) = \sum_{n_2=0}^{L-1} \left[ \sum_{n_1=0}^{K-1} x(n_1 L + n_2) W_K^{kn_1} \right] W_N^{n_2 k_2}. \quad (13)$$

The term inside the square bracket represents a $K$-point FFT. Defining

$$\tilde{X}_n (k_2) := \sum_{n_1=0}^{K-1} x(n_1 L + n_2) W_K^{n_1}, \quad (14)$$

we can write (13) as

$$X(k_1 K + k_2) = \sum_{n_2=0}^{L-1} \tilde{X}_n (k_2) W_N^{n_2 k}. \quad (15)$$

Since $k = k_1 K + k_2$, we can write $k_2 = (k) K$ where $(\cdot)_K$ represents the modulo $K$ operation. Now, (15) can be expressed as

$$X(k) = \sum_{n_2=0}^{L-1} \tilde{X}_n ((k) K) W_N^{n_2 k}. \quad (16)$$

The computation of (16) requires $K$ complex multiplications for each value of $n_2$. This decomposition is essentially similar to the less familiar zoom-FFT approach [20]. It can be shown that the number of complex multiplications required by the FFT module is also

$$C_{\text{FFT}} = \frac{N}{2} \log K. \quad (17)$$

5. Effect of filter transition bands on data recovery

For practical systems, the LPFs in the channelized RF front end have a nonzero transition band. To utilize the spectrum efficiently, we consider overlapping filter banks in the RF front-end of a PWR receiver as shown in Figure 9. The overlapping filters provide a mean to recover symbols that fall within the transition bands of the filters that is challenging otherwise. For simplicity, here we consider the overlap between two analog bands but the results can be easily extended to the case of multiple bands overlapping. Assume that each of the analog bands can accommodate one complete OFDMA symbol consisting of $N$ subcarriers. For OFDM systems, the use of cyclic prefix (CP) in each analog band converts a multipath fading channel into a set of $N$ parallel flat-frequency channels. We use $h(i)$ to represent the channel (including the transmit and receive analog filters) frequency response at the $i$th subcarrier for analog band 1. The diagonal channel matrix as seen by the received OFDM symbols
in analog band 1 is given by \( \mathbf{H}_1 = \text{diag}[h_1(1), \ldots, h_1(N)] \). Assume that \( \mathbf{u}_i = [u_i(1), \ldots, u_i(N)] \) is an \( N \times 1 \) vector of information symbols drawn from a constellation such that \( u_i(i) \) is used to modulate the \( i \)th subcarrier in analog band 1. In the absence of overlap between the filter transition bands, the received OFDM symbol \( \mathbf{r}_1 = [r_1(1), \ldots, r_1(N)] \) in analog band 1 is given by

\[
\mathbf{r}_1 = \mathbf{H}_1 \mathbf{u}_i + \mathbf{\eta}_i,
\]

where \( \mathbf{\eta}_i \) represents an \( N \times 1 \) vector of additive white Gaussian noise.

Now, assume that there are \( P \) subcarriers that overlap between the two adjacent analog bands 1 and 2. For analog band 1, these subcarriers correspond to index \( i = N - P, \ldots, N - 1 \), and for analog band 2 they correspond to \( i = 0, \ldots, P - 1 \) as shown in Figure 9. To avoid inter-band interference, we use the first half of these interfering subcarriers \( (i = N - P, \ldots, N - (P/2) - 1) \) to transmit data using the OFDMA symbol in analog band 1 and leave the other half \( (i = N - (P/2), \ldots, N - 1) \) unused during transmission in analog band 1. Similarly in analog band 2, we leave the first half \( (i = 0, \ldots, (P/2) - 1) \) and use the other half \( (i = P/2, \ldots, P - 1) \) to transmit data using the OFDMA symbol in analog band 2. The \( P/2 \) unmodulated subcarriers do not cause any inefficiency because OFDM systems, in general, do not use the subcarriers at the edges to compensate for the \( \text{sinc} \) function roll-off. For example, in a proposal submitted to the IEEE 802.22 standard, it has been proposed to leave 549 subcarriers on each edge of an OFDMA symbol of 6144 subcarriers.

With this arrangement, the received symbol in analog band 1 at the \( i \)th subcarrier is given by

\[
r_1(i) = \begin{cases} h_1(i)u_1(i) + \eta_1(i), & i = 0, \ldots, N - \frac{P}{2} - 1, \\ h_1(i)u_2(i) + \eta_1(i), & i = N - \frac{P}{2}, \ldots, N - 1, \end{cases}
\]

where \( u_1 \) and \( u_2 \) belongs to the symbols transmitted in analog band 1 and 2, respectively. Similarly, the received symbol in analog band 2 at the \( i \)th subcarrier can be written as

\[
r_2(i) = \begin{cases} h_2(i)u_1(i) + \eta_2(i), & i = 0, \ldots, \frac{P}{2} - 1, \\ h_2(i)u_2(i) + \eta_2(i), & i = \frac{P}{2}, \ldots, N - 1. \end{cases}
\]

The filter transition band severely attenuates the subcarriers at the edges. Our channelized receiver design is capable of overcoming this difficulty by combining the received symbols at corresponding subcarriers in analog band 1 and 2. Any well-known diversity combining method can be used depending on the system requirements [21]. In this paper, we consider the use of maximal ratio combining (MRC) to
We combine the corresponding subcarriers from the adjacent analog bands. The use of MRC maximizes the output signal to noise ratio but requires the knowledge of channel gains $h_1(i)$ and $h_2(i)$. Thus, to detect the data modulated on the subcarriers falling within the filter transition band, we apply MRC to obtain

$$
\tilde{u}_1(i) = h_1^+(i) r_1(i) + h_2^+(i-N+P) r_2(i-N+P)
$$

$$
i = N - P, \ldots, N - \frac{P}{2} - 1,
$$

(21)

$$
\tilde{u}_2(i) = h_1^+(i+N-P) r_1(i+N-P) + h_2^+(i) r_2(i)
$$

$$
i = \frac{P}{2}, \ldots, P - 1.
$$

To assess the impact of filter transition band on the system performance, we simulate an OFDMA system in a Rayleigh fading channel. We select Chebyshev analog filters and consider the case when $P = 128$ subcarriers belonging to a particular user fall in the filter transition bands. Figure 10 shows the effect of filter transition band on the bit error rate (BER) of the user that receives data from the subcarriers that fall within the transition band. The BER of the user increases significantly when a single analog band is used to detect the data. The increase in BER is more severe for higher order filters due to a sharper filter roll off. The proposed combining scheme uses two adjacent analog bands and combines the corresponding subcarriers using MRC according to (21). As shown in Figure 10, for a 12th order filter the proposed combining scheme offers almost 4 dB advantage in signal to noise ratio over the scheme that does not use filter band combining. The performance gap reduces for lower order filters.

### 6. DESIGN OF PWR FOR 802.22 NETWORKS

In this section, we discuss the architecture of PWR’s for the upcoming IEEE standard for wireless regional area networks (WRANs). In 2004, IEEE chartered the 802.22 working group to develop a fixed wireless access standard based on cognitive radio for WRANs. The preliminary link budget study for WRANs shows that it would be difficult to achieve the required minimum data rate (1.5 Mbps for DL and 384 kbps for UL) for users at the edge of a 33 km radius using a single TV channel of 6 MHz [22]. To meet these requirements, channel bonding (contiguous spectrum) and channel aggregation (noncontiguous spectrum) are being considered for 802.22 standard [2]. To take advantage of these provisions, we propose an IEEE 802.22 compatible PWR receiver with three chains of RF front-end each 18 MHz wide as shown in Figure 11. Each of the RF chains is selectable depending on whether the system is operating...
with channel aggregation or not. In channel aggregation mode, the center frequencies of the oscillator will be adjusted through a programmable synthesizer [18]. When channel bonding is used, a single RF chain can accommodate the full system bandwidth by selecting a proper center frequency of the oscillator. The ADC sampling frequency is in accordance with the baseline system clock of 6.875 MHz that has been approved in the draft of IEEE 802.22 standard for 2K symbols [2]. An RF front-end similar to Figure 11 can be used for transmission with parameters adjusted as per the standard.

For the IFFT/FFT modules, we use the efficient IFFT/FFT architectures described in Section 4 with $K = 512$. Based on the RF front end shown in Figure 11, it is clear that the value of $N$ is fixed at 6144. Although the value of $N$ is not a power of 2, the decomposition discussed in Section 4 is valid without any penalty as opposed to traditional radix-2 implementations that require $N$ to be a power of 2 for lowest complexity [23]. The value $K = 512$ gives us a leverage to be computationally efficient when 802.22 systems operate in the fractional bandwidth mode. In the context of 802.22, the fractional bandwidth is defined as 1 MHz chunks of the spectrum that may be available to CR when narrowband Part 74 devices are operating in a TV channel [2]. With $K = 512$, we can cover 2 MHz of fractional bandwidth with lowest computational cost.

7. CONCLUSIONS

Cognitive radio technology is a new paradigm in wireless communications that promises the efficient utilization of spectrum using opportunistic spectrum access. A CR relies on PWR to adapt its transmission and reception system based on the feedback from the spectrum sensing unit. We described the complete design of OFDMA-based PWR's including RF front-end and digital baseband modem. Use of channelized multiple narrowband RF front end simplifies the sampling requirements of ADCs and their power consumption as well as the complexity of digital baseband processor. We described the low-complexity design of IFFT/FFT modules based on modified Cooley-Tukey decomposition and discussed their scalability features. We discussed the impact of analog filter (LPF) transition bands on data recovery and proposed a combining scheme to compensate for the filter attenuation. We used IEEE 802.22 standard (draft) as an example system and discussed the design of low-complexity PWR's for such systems.

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