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Title: **Fundamentals of CPS-OFDM Waveform for 5G New Radio**

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## 1 Introduction

New radio (NR) waveform design is based on OFDM with potential support of non-orthogonal property and multiple access [1]. In this document, a circularly pulse-shaped (CPS) precoding technique for OFDM-based waveforms is proposed to facilitate 5G fragmented spectrum usage. We call this proposed scheme CPS-OFDM, which is characterized by precoder flexibility with multiple access support.

### 1.1 Contributions

The details on OFDM-based transceiver structure, CPS-OFDM waveform realization, used parameters, and performance evaluations are provided according to the agreements in the last meeting [2]. Furthermore, we analyze the transmitter complexity and derive the closed-form expression of the power spectrum for the proposed CPS-OFDM waveform so as to justify that the implementation is very efficient and the out-of-subband (OOSB) emission can be lowered without block extension, respectively. Simulation results show that the proposed scheme using the CPS precoder has even better peak-to-average power ratio (PAPR) and bit-error rate (BER) compared to multicarrier waveforms using subband filtering techniques. More evaluation results of CPS-OFDM for Case 1a, 1b, 2, 3, and 4 can be found in our related documents [3]-[4].

### 1.2 Literature Review

Non-orthogonal multicarrier waveforms featuring low out-of-band emission and insensitivity of synchronization mismatch are considered as alternatives to OFDMA and SC-FDMA applied to 4G LTE technology [5]. FBMC [6] doing subcarrier filtering highly restricts the spectral leakage at the costs of long transmission time and complicated equalization, which are often known as MIMO-unfriendly. In [7]-[10], f-OFDM and UF-OFDM are proposed. However, both of them using subband filtering techniques usually results in inter-block interference (IBI) at the receiver. On the other hand, GFDM proposed by [11] is regarded as a precoding technique without block extension, but its modulation performing on the whole band makes multiple access difficult. Although many waveform options arise in recent years [12], a multicarrier waveform characterized by precoder flexibility has not been found yet.

In our previous work [13]-[15], the idea of confining the GFDM modulation matrix to a subband has already been introduced with some preliminary performance evaluations.

### 1.3 Notations

Boldfaced lower case letters such as  $\mathbf{x}$  represent column vectors, boldfaced upper case letters such as  $\mathbf{X}$  represent matrices, and normal letters such as  $X$  represent scalars. Superscripts as in  $\mathbf{X}^T$  and  $\mathbf{X}^H$  denote the transpose and transpose-conjugate operators, respectively, of a matrix or a vector. The matrices  $\mathbf{I}_N$  and  $\mathbf{0}_{N \times M}$  are the  $N \times N$  identity matrix and  $N \times M$  all-zero matrix, respectively. The  $N$ -point normalized discrete Fourier transform (DFT) matrix denoted by  $\mathbf{W}_N$  is defined as that the  $(k, n)$ -th entry of  $\mathbf{W}_N$  is  $\frac{1}{\sqrt{N}}e^{-j2\pi(k-1)(n-1)/N}$ . The  $(N \times M)$ -dimensional complex and real fields are expressed as  $\mathbb{C}^{N \times M}$  and  $\mathbb{R}^{N \times M}$ , respectively. Calligraphic upper case letters such as  $\mathcal{I}$  represent sets of discrete indices. The  $(i + 1)$ -th entry

of  $\mathbf{x}$  is denoted by  $[\mathbf{x}]_i$ . The sub-matrix of  $\mathbf{X}$  formed by the column vectors with the indices given in  $\mathcal{I}$  is denoted by  $[\mathbf{X}]_{\mathcal{I}}$ . The magnitude response of a complex number is obtained by the operator  $|\cdot|^2$ . The expectation operator is denoted by  $\mathbb{E}\{\cdot\}$ . The operators of a linear convolution and an  $S$ -period circular convolution are denoted by  $*$  and  $\otimes_S$ , respectively. The decibel representation of a real number  $C$  is shown as  $C|_{\text{dB}}$ . The operation of  $i$  modulo  $S$  is abbreviated to  $((i))_S$ .

## 2 OFDM-Based Transceiver Structure

In this section, we provide a generalized OFDM-based transceiver model as shown in Figure 1, which can address the realizations of CPS-OFDM, f-OFDM, UF-OFDM, OFDMA (i.e., OFDM with CP), SC-FDMA (i.e., DFT-s-OFDM with CP), and etc.

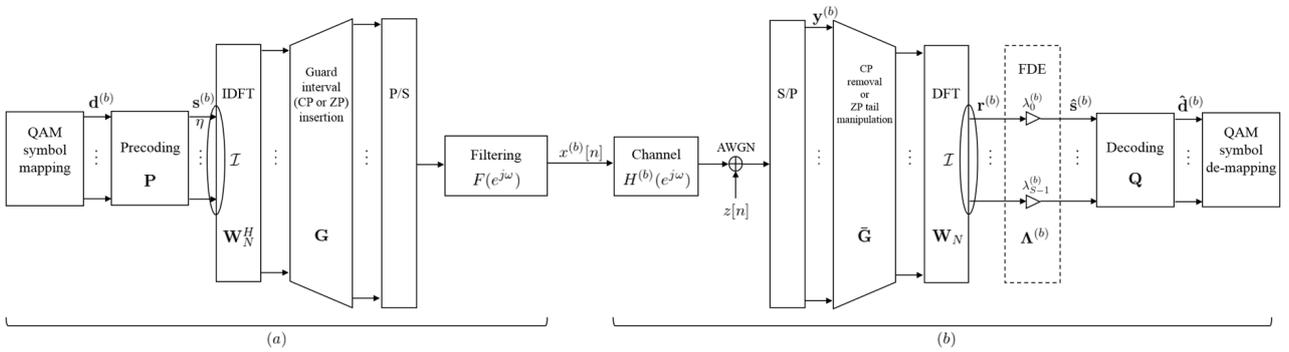


Figure 1: Transceiver model (a) waveform generation based on a generalized OFDM framework (b) OFDM-FDE receiver structure.

### 2.1 Generalized OFDM-Based Transmitter

The foundation of waveform generation using digital implementation based on a generalized OFDM framework is illustrated in Figure 1 (a). Multicarrier transmission is realized by the inverse DFT matrix  $\mathbf{W}_N^H$ . The insertion of guard interval is represented by the matrix  $\mathbf{G}$ , which is usually chosen to be either

$$\mathbf{G}_{\text{cp}} = \begin{bmatrix} \mathbf{0}_{G \times (N-G)} & \mathbf{I}_G \\ & \mathbf{I}_N \end{bmatrix} \quad \text{or} \quad \mathbf{G}_{\text{zp}} = \begin{bmatrix} \mathbf{I}_N \\ \mathbf{0}_{G \times N} \end{bmatrix} \quad (1)$$

for cyclic prefix (CP) or zero padding (ZP) to prevent the IBI stemmed from multipath propagation whose delay spread order is  $L$ ,  $\mathbf{G} \in \mathbb{R}^{(N+G) \times N}$ ,  $G \geq L$ . The precoding matrix  $\mathbf{P} \in \mathbb{C}^{S \times S}$  and the  $L_f$ -order finite impulse response (FIR) filter  $\mathbf{h}_f \in \mathbb{C}^{(L_f+1) \times 1}$  can be designed in the frequency domain and the time domain, respectively. The filtering operation by doing a linear convolution can be further modeled as a banded Toeplitz matrix denoted by  $\mathbf{F}$ . Thus, the transmitted signal of the  $b$ -th block can be formulated as

$$\mathbf{x}^{(b)} = \mathbf{F}\mathbf{G} [\mathbf{W}_N^H]_{\mathcal{I}} \mathbf{P} \mathbf{d}^{(b)}, \quad (2)$$

where  $\mathcal{I}$  denotes the set of  $S$  subcarrier indices allocated for the user and  $\mathbf{d}^{(b)}$  represents the corresponding data vector consisting of  $D$  quadrature amplitude modulation (QAM) symbols and  $Z$  null symbols,  $S = D + Z$ . The coefficients on the subcarriers indexed by  $\mathcal{I}$  are provided by  $\mathbf{s}^{(b)} = \mathbf{P} \mathbf{d}^{(b)}$ . The synthesis matrix  $\Phi = \mathbf{G} [\mathbf{W}_N^H]_{\mathcal{I}} \mathbf{P}$  is required to be a full column rank matrix for the symbol detection at the receiver.

The performance of OOSB emission is evaluated by knowing the power spectrum [16]. Suppose that the data vector follows

$$\mathbb{E} \left\{ \mathbf{d}^{(b)} \mathbf{d}^{(b')H} \right\} = \begin{cases} E_s \mathbf{I}_S, & b = b' \\ \mathbf{0}_{S \times S}, & \text{otherwise} \end{cases} \quad (3)$$

where  $E_s$  is the energy of each symbol. The power spectrum of transmitting signal  $x^{(b)}[n] = [\mathbf{x}^{(b)}]_n$  is given as

$$\tilde{S}(e^{j\omega}) = \frac{E_s}{N+G} \sum_{s \in \mathcal{I}} |\Phi_s(e^{j\omega})|^2 |F(e^{j\omega})|^2, \quad (4)$$

where the constant  $1/(N+G)$  is due to the cyclo wide sense stationary (CWSS) [16] property of  $x^{(b)}[n]$ ,  $F(e^{j\omega})$  is the discrete-time Fourier transform (DTFT) expression of the filter  $\mathbf{h}_f$ ,  $\Phi_s(e^{j\omega})$  represents the synthesis filter with respect to the column vector  $[\Phi]_{s-\eta}$ , and  $\eta$  is the starting index of the occupied  $S$  subcarriers.

## 2.2 OFDM-FDE Receiver and Implementation Aspects

A multicarrier signal with CP or ZP redundancy via a multipath channel can be efficiently processed by the DFT matrix  $\mathbf{W}_N$  and the one-tap frequency-domain equalization (FDE) as revealed in Figure 1 (b), in which the matrix  $\bar{\mathbf{G}}$  is chosen to be either

$$\bar{\mathbf{G}}_{\text{cp}} = [\mathbf{0}_{N \times G} \quad \mathbf{I}_N] \quad \text{or} \quad \bar{\mathbf{G}}_{\text{zp}} = \begin{bmatrix} \mathbf{I}_G & \mathbf{0} & \mathbf{I}_G \\ \mathbf{0} & \mathbf{I}_{N-G} & \mathbf{0} \end{bmatrix} \quad (5)$$

for CP removal or ZP tail manipulation. Let  $\mathbf{h}^{(b)} \in \mathbb{C}^{(L+1) \times 1}$  denote the channel impulse response (CIR) for the  $b$ -th block transmission. The received signal is expressed as

$$\mathbf{y}^{(b)} = \mathbf{H}_{\text{toep}}^{(b)}(e^{j\omega}) \mathbf{G} [\mathbf{W}_N^H]_{\mathcal{I}} \mathbf{s}^{(b)} + \tilde{\mathbf{z}}, \quad (6)$$

where  $\mathbf{H}_{\text{toep}}^{(b)}(e^{j\omega})$  can be viewed as a pseudocirculant Toeplitz matrix [16] formed by the equivalent CIR  $\tilde{\mathbf{h}}^{(b)} = \mathbf{h}^{(b)} * \mathbf{h}_f \in \mathbb{C}^{(L+L_f+1) \times 1}$  and  $\tilde{\mathbf{z}}$  is a white noise vector. Suppose that the IBI can be completely eliminated at the receiver (i.e.,  $G \geq L + L_f$ ), the received frequency-domain signal is written as

$$\begin{aligned} \mathbf{r}^{(b)} &= [\mathbf{W}_N^T]_{\mathcal{I}}^T \bar{\mathbf{G}} \mathbf{y}^{(b)} \\ &= [\mathbf{W}_N^T]_{\mathcal{I}}^T \mathbf{H}_{\text{circ}}^{(b)} [\mathbf{W}_N^H]_{\mathcal{I}} \mathbf{s}^{(b)} + \mathbf{z} \\ &= \mathbf{H}^{(b)} \mathbf{P} \mathbf{d}^{(b)} + \mathbf{z}, \end{aligned} \quad (7)$$

where  $\mathbf{H}_{\text{circ}}^{(b)} = \bar{\mathbf{G}} \mathbf{H}_{\text{toep}}^{(b)}(e^{j\omega}) \mathbf{G}$  is a circulant matrix of the equivalent channel,  $\mathbf{H}^{(b)} \in \mathbb{C}^{S \times S}$  is a diagonal matrix whose diagonal elements corresponding to the frequency response of  $\tilde{\mathbf{h}}^{(b)}$ , and  $\mathbf{z} = [\mathbf{W}_N^T]_{\mathcal{I}}^T \bar{\mathbf{G}} \tilde{\mathbf{z}}$  is still a zero-mean Gaussian random vector. The received data vector can then be obtained as follows

$$\begin{aligned} \hat{\mathbf{d}}^{(b)} &= \mathbf{Q} \mathbf{\Lambda}^{(b)} \mathbf{r}^{(b)} \\ &= \mathbf{d}^{(b)} + \mathbf{Q} \mathbf{\Lambda}^{(b)} \mathbf{z}, \end{aligned} \quad (8)$$

where the diagonal matrix  $\mathbf{\Lambda}^{(b)}$  and the decoding matrix  $\mathbf{Q}$  are used for the FDE and compensating  $\mathbf{P}$ , respectively, in zero forcing (ZF) or minimum mean square error (MMSE) manner.

Accordingly, no matter what the precoder  $\mathbf{P}$  and the filter  $F(e^{j\omega})$  are chosen, the orthogonality among different users holds in the absence of IBI and synchronization mismatch. However, the symbol detection may be subject to the noise amplification caused by the decoder  $\mathbf{Q}$ , which is not a unitary matrix. Besides, the block extension due to the  $L_f$ -order filter usually results in significant IBI especially in typical urban scenarios.

## 2.3 Precoder Design Problem

We aim at finding an  $S \times S$  precoding matrix  $\mathbf{P}$  that determines the waveform characterizations. The filtering technique is not considered,  $\mathbf{h}_f = \mathbf{1}$ , and no IBI occurs ( $G \geq L$ ). The power spectrum given in Equation (4) is rewritten as

$$S(e^{j\omega}) = \frac{E_s}{N+G} \sum_{s \in \mathcal{I}} |\Phi_s(e^{j\omega})|^2. \quad (9)$$

At the receiver, the modulation symbols can be recovered according to Equation (7) and Equation (8).

## 3 Proposed CPS-OFDM Waveform

The proposed CPS precoding matrix is obtained by

$$\mathbf{P} = \mathbf{W}_S \mathbf{A}, \quad (10)$$

where every column of  $\mathbf{A}$  is derived from the prototype pulse vector  $\mathbf{a}_{0,0}$  given by

$$\begin{aligned} \mathbf{A} &= [\mathbf{a}_{0,0} \cdots \mathbf{a}_{0,M-1} \mid \cdots \mid \mathbf{a}_{k,0} \cdots \mathbf{a}_{k,m} \cdots \mathbf{a}_{k,M-1} \mid \cdots \mid \mathbf{a}_{K-1,0} \cdots \mathbf{a}_{K-1,M-1}] \\ &= [\mathbf{A}_0 \mid \cdots \mid \mathbf{A}_k \mid \cdots \mid \mathbf{A}_{K-1}], \end{aligned} \quad (11)$$

$$a_{k,m}[n'] = [\mathbf{a}_{k,m}]_{n'} = a_{0,0} [((n' - (m + \mu)K))_S] e^{j2\pi \frac{k+\beta}{K} n'}, \quad n' = 0, 1, \dots, S-1, \quad S = KM. \quad (12)$$

In Equation (12), two parameters  $\mu$  and  $\beta$  are introduced to shift the output pulses and the intra-subband subcarriers, respectively.

This way enables that the time-frequency numerology in each subband can be flexibly determined by  $K$  shaped subcarriers and  $M$  sub-blocks in terms of GFDM. Orthogonal multiple access of different users allocated on different subbands is still supported, although the intra-subband subcarriers of each user may be non-orthogonal due to the time-domain CPS modulation of  $\mathbf{A}$ . We call this proposed scheme CPS-OFDM.

In the subsequent subsections, the index  $b$  is omitted for simplicity. Every sub-vector of the data vector follows the same subscript index as given in Equation (11), i.e.,

$$\mathbf{d} = [\mathbf{d}_0^T \mid \cdots \mid \mathbf{d}_k^T \mid \cdots \mid \mathbf{d}_{K-1}^T]^T. \quad (13)$$

### 3.1 CPS-OFDM Transmitter Complexity Analysis

The implementation of the CPS precoder in Equation (10) is analyzed as follows. Firstly, Equation (12) can be rewritten as

$$a_{k,m}[n'] = \{\delta[n' - (m + \mu)K] \otimes_S a_{0,0}[n']\} e^{j\frac{2\pi}{S}(k+\beta)Mn'} \quad (14)$$

as shown in [17]. Equation (14) can be further interpreted as

$$\text{IDFT}_S \{ \text{DFT}_S \{ \delta[n' - (m + \mu)K] \} \cdot \gamma_i \} e^{j\frac{2\pi}{S}(k+\beta)Mn'},$$

where  $\gamma_i = \text{DFT}_S \{ a_{0,0}[n'] \}$ . Hence, benefited from the equidistance of picking  $M$  symbols, the  $k$ -th sub-matrix of  $\mathbf{A}$  can be expressed as

$$\mathbf{A}_k = \mathbf{W}_S^H \mathbf{C}_{(k+\beta)M} \mathbf{\Gamma} \mathbf{R} \mathbf{W}_M, \quad (15)$$

where  $\mathbf{R} = [\mathbf{I}_M \cdots \mathbf{I}_M]^T$  is a repetition matrix formed by the  $K$  identity matrices,  $\mathbf{\Gamma} = \text{diag}(\gamma_0, \gamma_1, \cdots, \gamma_{S-1})$  is a diagonal matrix, and  $\mathbf{C}_{(k+\beta)M}$  is a permutation matrix defined as

$$\mathbf{C}_0 = \mathbf{I}_S, \quad \mathbf{C}_{(k+\beta)M} = \begin{bmatrix} \mathbf{0} & \mathbf{I}_{(k+\beta)M} \\ \mathbf{I}_{S-(k+\beta)M} & \mathbf{0} \end{bmatrix}. \quad (16)$$

Similarly, the  $k$ -th sub-matrix of  $\mathbf{P}$  is then known as

$$\mathbf{P}_k = \mathbf{W}_S \mathbf{A}_k = \mathbf{C}_{(k+\beta)M} \mathbf{\Gamma} \mathbf{R} \mathbf{W}_M, \quad (17)$$

which can be depicted in Figure 2 in block diagram form. Finally, the coefficients on the subcarriers indexed by  $\mathcal{I}$  are given by

$$\mathbf{s} = \sum_{k=0}^{K-1} \mathbf{P}_k \mathbf{d}_k. \quad (18)$$

Based on the above derivations, Figure 3 displays the equivalent low-complexity realization of the CPS precoder.

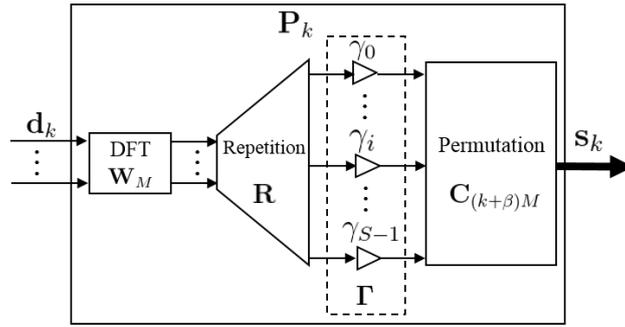


Figure 2: Building blocks in  $\mathbf{P}_k$  for efficient implementation.

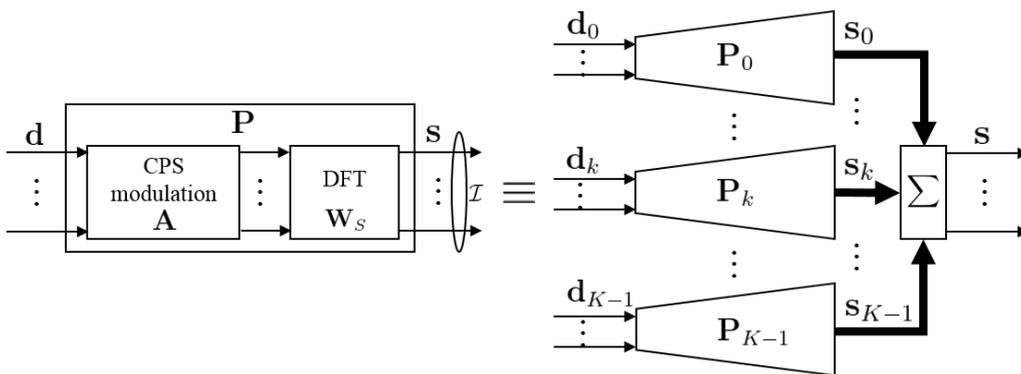


Figure 3: Equivalent low-complexity realization of the CPS precoder.

**Observation 1:** *CPS precoder implementation is very efficient since each shaped subcarrier is equivalently realized by a small size DFT, a repetition operation, an element-wise multiplication, and a permutation operation.*

The transmitter complexity of CPS-OFDM is evaluated by computing its number of complex multiplications. The OFDM modulation based on an  $N$ -point inverse fast Fourier transform (IFFT) has  $N \log_2 N$  complex multiplications. The CPS precoder is made up of the  $K$  submatrices  $\mathbf{P}_0, \mathbf{P}_1, \dots, \mathbf{P}_{K-1}$ , each of them executes an  $M$ -point DFT and an  $S$ -tap frequency-domain weighting containing  $M \log_2 M$  and  $KM$  complex multiplications, respectively. Note that the repetition and the permutation operations can be actualized without multiplier units. Compared to the  $S$ -point DFT precoder used for the legacy SC-FDMA, the proposed CPS precoder slightly increases the complexity by  $KM(K - \log_2 K)$  complex multiplications. Compared to the filtering techniques used for f-OFDM and UF-OFDM whose complexity increases in the order of  $N$ , the proposed CPS precoding technique is much more efficient in the transmitter implementation.

In practice, we can utilize the real-valued raised-cosine function (RC) represented by the shaping coefficients  $\gamma_0, \gamma_1, \dots, \gamma_{S-1}$  with the roll-off factor  $\alpha$  dominating the extent of non-orthogonal overlapping. Some of the shaping coefficients are zeros. The realization of the CPS precoder can be substantially simplified.

### 3.2 CPS-OFDM Power Spectrum Analysis

The power spectrum of the CPS-OFDM waveform is theoretically derived for OOSB emission evaluation. According to Equation (17), each column vector of the synthesis matrix can be found by

$$\bar{\phi}_{k,m} = \mathbf{G} [\mathbf{W}_N^H]_{\mathcal{I}} \mathbf{C}_{(k+\beta)M} \mathbf{\Gamma} \mathbf{R} [\mathbf{W}_M]_m \quad (19)$$

with the same subscript indices  $k$  and  $m$  used for  $\mathbf{a}_{k,m}$ . Assuming  $\mathbf{G} = \mathbf{G}_{\text{cp}}$  whose CP length is now denoted by  $G'$ , we expand Equation (19) with the delay effect of  $\mu$  and then obtain the generalized synthesis function

$$\phi_{k,m}[n] = [\bar{\phi}_{k,m}]_n = \frac{1}{\sqrt{NM}} \sum_{i=0}^{S-1} \gamma_{((i-(k+\beta)M))_S} e^{-j\frac{2\pi}{M}i(m+\mu)} e^{j\frac{2\pi}{N}(\eta+i)(n-G')}, \quad (20)$$

$n = 0, 1, \dots, N + G' - 1$ , where the constant  $1/\sqrt{NM}$  comes from  $\mathbf{W}_N^H$  and  $\mathbf{W}_M$ . The corresponding magnitude response of the synthesis filter is

$$|\Phi_{k,m}(e^{j\omega})|^2 = \frac{1}{NM} \left| \sum_{i=0}^{S-1} \gamma_{((i-(k+\beta)M))_S} e^{-j2\pi i \left( \frac{m+\mu}{M} + \frac{G'}{N} \right)} W \left( e^{j \left[ \omega - \frac{2\pi}{N}(\eta+i) \right]} \right) \right|^2, \quad (21)$$

where

$$W \left( e^{j \left[ \omega - \frac{2\pi}{N}(\eta+i) \right]} \right) = \sum_{n=0}^{N+G'-1} e^{-j \left[ \omega - \frac{2\pi}{N}(\eta+i) \right] n}. \quad (22)$$

When choosing ZP as the guard interval whose length is  $G$ , the alternative expressions of Equation (20) and Equation (21) can be simply obtained by setting  $G' = 0$ . Refer to Equation (9), the closed-form expression of the CPS-OFDM power spectrum is thus written as

$$S(e^{j\omega}) = \frac{E_s}{N+G} \sum_{k \in \mathcal{K}} \sum_{m \in \mathcal{M}} |\Phi_{k,m}(e^{j\omega})|^2, \quad (23)$$

where  $\mathcal{K}$  and  $\mathcal{M}$  are two sets of indices indicating the used subcarriers and sub-blocks in the subband, respectively.

Figure 4 (a)-(b) provide two illustrations of the synthesis functions and the synthesis filters in Equation (20) and Equation (21), respectively. The parameters are chosen as follows:  $K = 9$ ,  $M = 4$ ,  $N = 128$ ,  $G = 8$ ,  $G' = 0$ ,  $\mu = 0$ , RC ( $\alpha = 0.5$ ), and  $\mathcal{K} = \{4\}$ . The magnitude responses are normalized to 0 dB. Obviously, the first sub-block ( $m = 0$ ) located in the block boundary exhibits the worst leakage because of its steep truncation. A common way to avoid this issue is to let  $[\mathbf{d}_k]_0 = 0, \forall k$ . For CP utilization, the policy of assigning null symbols depends on  $G'$  and  $\mu$ . An example is shown in Figure 4 (c)-(d) with  $G' = 8$  and  $\mu = -1/9$ . Equation (23) is then verified with the simulation result in the next section.

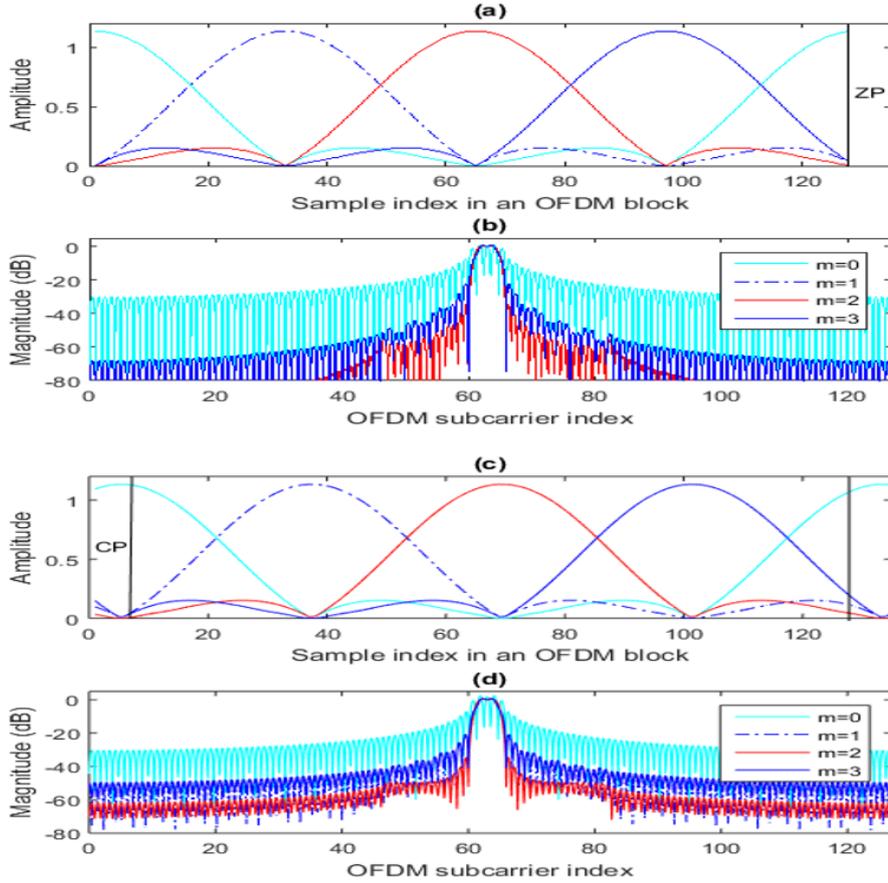


Figure 4: Illustration of synthesis functions and synthesis filters: (a) and (b) for ZP utilization, (c) and (d) for CP utilization.

## 4 Simulation Results and Performance Comparisons

The performance evaluation results of applying the proposed CPS-OFDM waveform are given in this section. In the simulations, practical system parameters are chosen according to [20], in which the carrier frequency is 4 GHz, the subcarrier spacing is 15 kHz, the FFT size  $N$  is 1024, the guard interval length  $G$  is 72, the number of subcarriers allocated for one user  $S$  is 48, the starting subcarrier index  $\eta$  is 212, uncoded 16QAM is adopted, perfect synchronization and channel estimation are assumed, single transmit antenna and single receive antenna are equipped, and the two channel models are ETU for 3 km/hr and EVA for 120 km/hr mobility. The detailed transceiver structure refers to Figure 1, where the received data vector is obtained by MMSE equalization. Here PA nonlinearity is not considered. For CPS-OFDM, we set  $K = 3$ ,  $M = 16$ ,  $[\mathbf{d}_0]_0 = [\mathbf{d}_1]_0 = [\mathbf{d}_2]_0 = 0$ ,  $D = 45$ , RC prototype pulse ( $\alpha = 0.5$ ), and ZP as the guard interval.

Three multicarrier waveforms: OFDMA, f-OFDM, and UF-OFDM are used for performance comparisons. OFDMA ( $\mathbf{P} = \mathbf{I}_S$ ,  $\mathbf{h}_f = 1$ ) serves as a benchmark to justify the gain of new waveform utilization. For f-OFDM, the FIR filter  $\mathbf{h}_f$  is generated by imposing the raise-cosine window<sup>1</sup> on the sinc impulse response with 3 tone-offset subcarriers on each side [21]-[23]. Two filter orders  $L_f = G$  and  $L_f = N/2$  are taken into account. For UF-OFDM, the additional precoding matrix is a diagonal matrix whose diagonal entries are to compensate the passband decay caused by the Dolph-Chebyshev filter with the attenuation factor 75 dB and  $L_f = G$  [24]-[25]. Note that the guard intervals of f-OFDM and UF-OFDM are CP and ZP, respectively. In Figure 5, we point out the used function blocks of OFDMA, CPS-OFDM, f-OFDM, and UF-OFDM for their waveform generations.

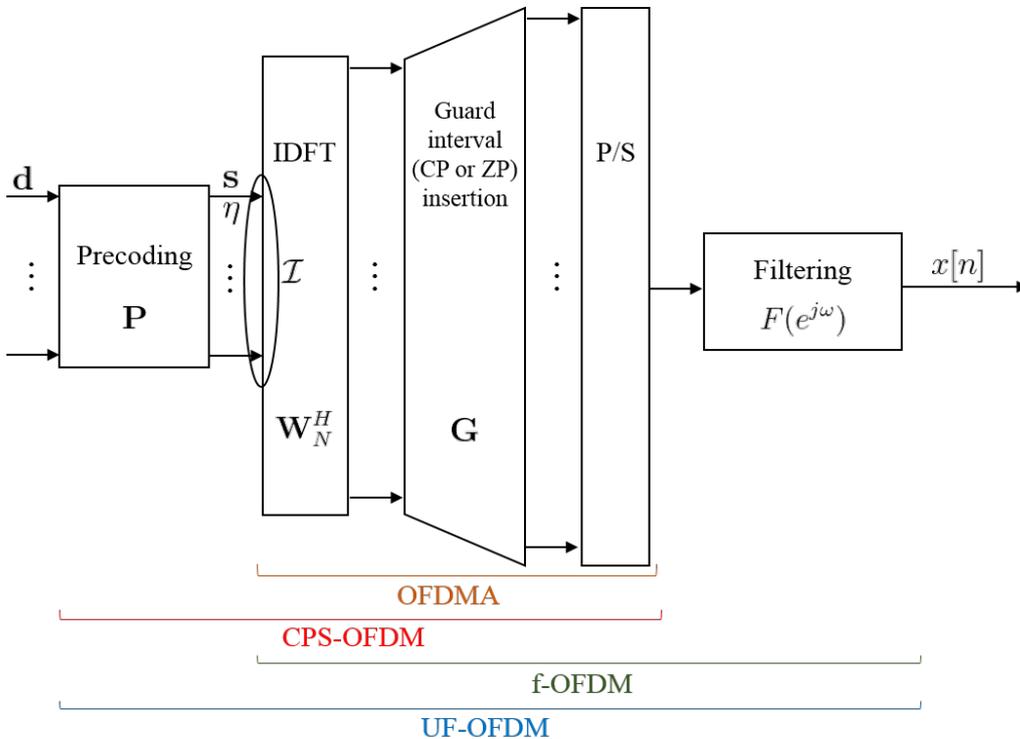


Figure 5: Generalized OFDM-based waveform generation model

<sup>1</sup> $w[l_f] = \left(0.5 \left(1 + \cos\left(\frac{2\pi l_f}{L_f}\right)\right)\right)^{0.6}$ ,  $-\lfloor \frac{L_f+1}{2} \rfloor \leq l_f \leq \lfloor \frac{L_f+1}{2} \rfloor$

## 4.1 OOSB Emission Evaluation

The power spectrum results shown in Figure 6 validate the derivation result of Equation (23) as delineated by the markers “+”, and reveal that the CPS-OFDM, the f-OFDM, and the UF-OFDM signal transmissions all meet the ACLR<sup>2</sup> requirements in the measured region. While using the CPS precoding technique cannot achieve extremely low OOSB emission like using the filtering techniques, it basically prevents IBI and timing offset sensitivity increase due to block extension. In the CPS-OFDM passband, the 48 OFDM subcarrier are indeed transformed into 3 RC shaped subcarriers. Observed from the f-OFDM passband, the filter orders  $L_f = G$  is not enough to guarantee the flatness for such narrowband transmission.

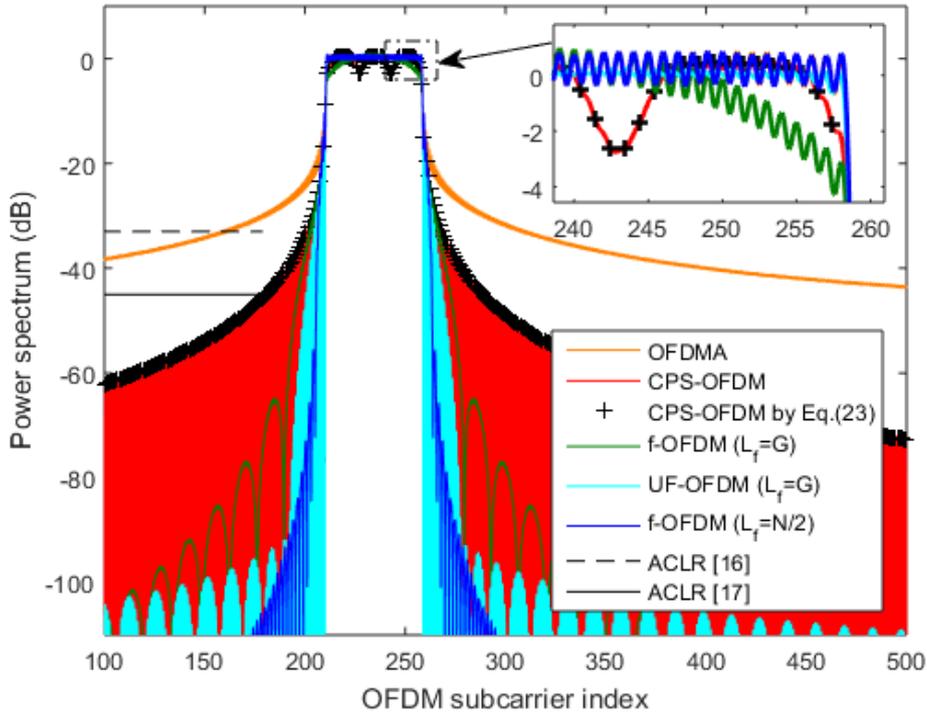


Figure 6: Power spectrum simulation results in which Equation (23) is validated and the new waveforms meet the low OOSB emission requirement.

**Observation 2:** *CPS-OFDM waveform has a lower OOSB emission compared to the legacy OFDMA. UF-OFDM and f-OFDM have even lower OOSB emission, when PA nonlinearity is not considered.*

## 4.2 PAPR Evaluation

The PAPR results shown in Figure 7 indicate that the CPS-OFDM waveform has the best power amplifier (PA) efficiency. The reason is that the embedded CPS configuration can turn a group of the OFDM subcarriers into a widened subcarrier by intrinsic DFT spreading as illustrated in Figure 2 and Figure 3. Given  $L_f = G$ , f-OFDM has better PAPR performance than UF-OFDM because of CP utilization. When applying the filtering techniques, the PAPR rises as the filter length increases.

<sup>2</sup>abbreviation for adjacent channel leakage power ratio

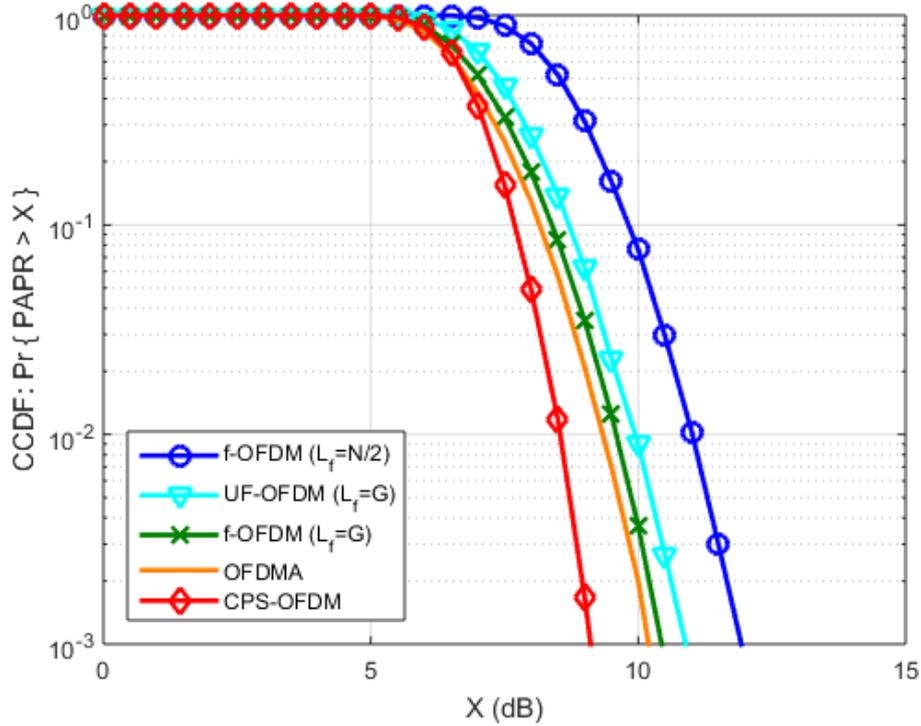


Figure 7: PAPR evaluation results in which CPS-OFDM shows the best performance because of the intrinsic DFT spreading of the CPS precoder.

### 4.3 Uncoded BER Evaluation

Figure 8 and Figure 9 present the uncoded BER results for ETU and EVA channel environments, respectively. Obviously, CPS-OFDM without IBI impairment exhibits better detection performance than f-OFDM and UF-OFDM. Compared to OFDMA, the performance of CPS-OFDM may be slightly worse since the decoding matrix  $\mathbf{Q}$  is now a non-unitary matrix that causes noise amplification. In high signal-to-noise ratio (SNR) region, the effect of the decoding matrix against the channel spectral nulls [26] becomes evident so that CPS-OFDM can achieve a more reliable transmission than OFDMA. For f-OFDM with  $L_f = N/2$ , an extended CP or a complicated equalizer may be needed in order to deal with the significant IBI.

**Observation 3:** *CPS-OFDM waveform has better PAPR performance than f-OFDM, UF-OFDM, and the legacy OFDMA.*

**Observation 4:** *Due to IBI-free reception, CPS-OFDM has better BER performance than f-OFDM and UF-OFDM.*

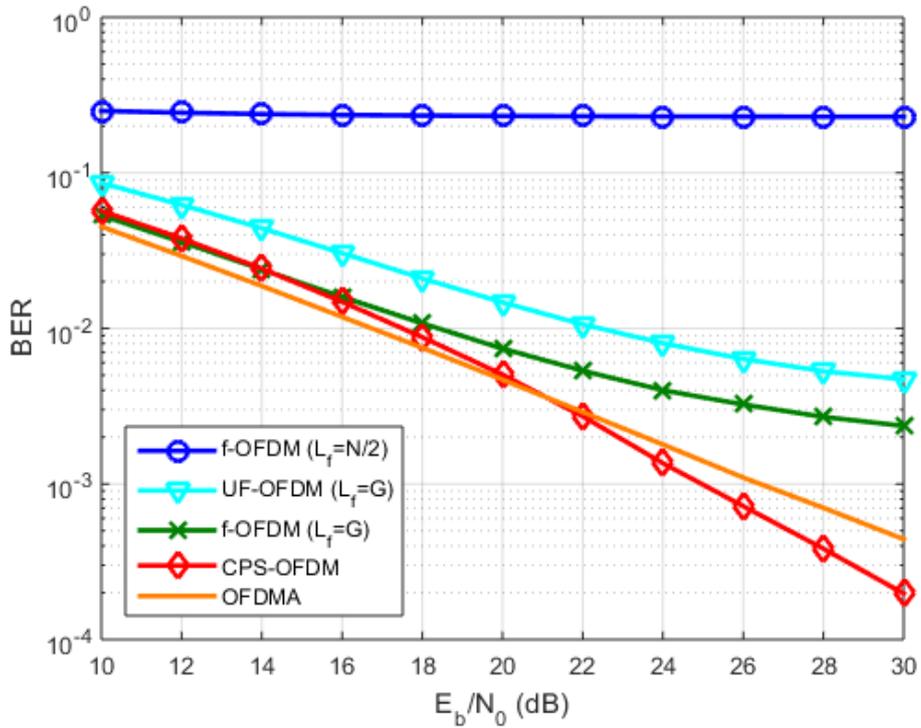


Figure 8: BER for ETU channel with 3 km/hr mobility showing the merit of CPS-OFDM whose decoding matrix is against channel spectral nulls.

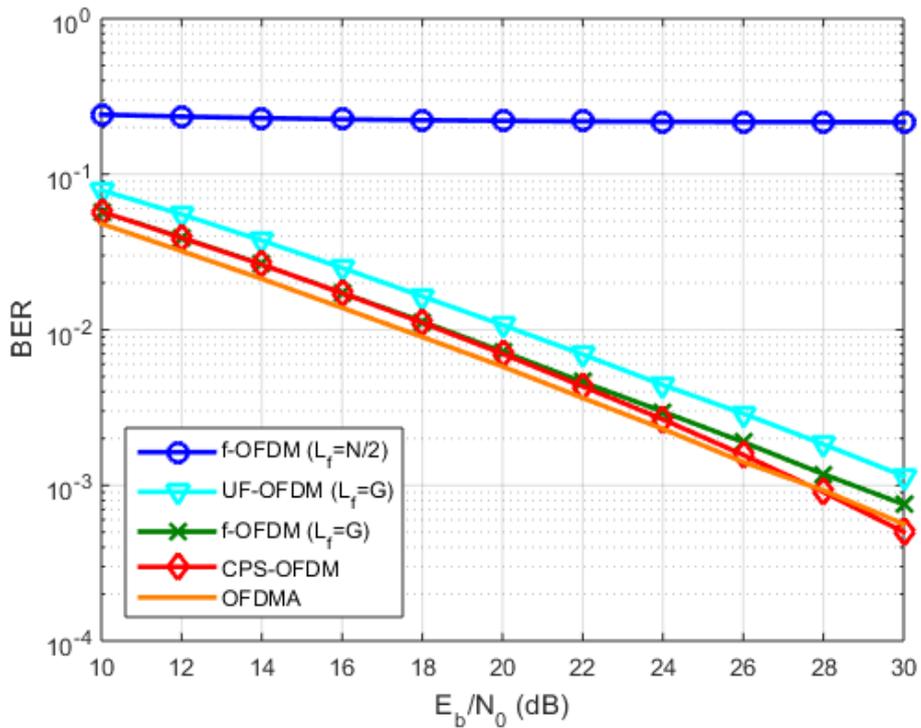


Figure 9: BER for EVA channel with 120 km/hr mobility showing that the IBI-free reception of the CPS-OFDM system ensures the reliability.

## 5 Summary

CPS-OFDM waveform characterized by precoder flexibility with multiple access support is designed to facilitate 5G fragmented spectrum usage. Based on the analysis and the evaluations in this document, we summarize the observations and then make two proposals as follows:

**Observation 1:** *CPS precoder implementation is very efficient since each shaped subcarrier is equivalently realized by a small size DFT, a repetition operation, an element-wise multiplication, and a permutation operation.*

**Observation 2:** *CPS-OFDM waveform has a lower OOSB emission compared to the legacy OFDMA. UF-OFDM and f-OFDM have even lower OOSB emission, when PA nonlinearity is not considered.*

**Observation 3:** *CPS-OFDM waveform has better PAPR performance than f-OFDM, UF-OFDM, and the legacy OFDMA.*

**Observation 4:** *Due to IBI-free reception, CPS-OFDM has better BER performance than f-OFDM and UF-OFDM.*

**Proposal 1:** *Precoding techniques applied to OFDM-based waveforms shall be further studied in order to prevent block extension causing inter-block interference (IBI)*

**Proposal 2:** *CPS-OFDM can be considered as one of highly-potential waveforms in NR.*

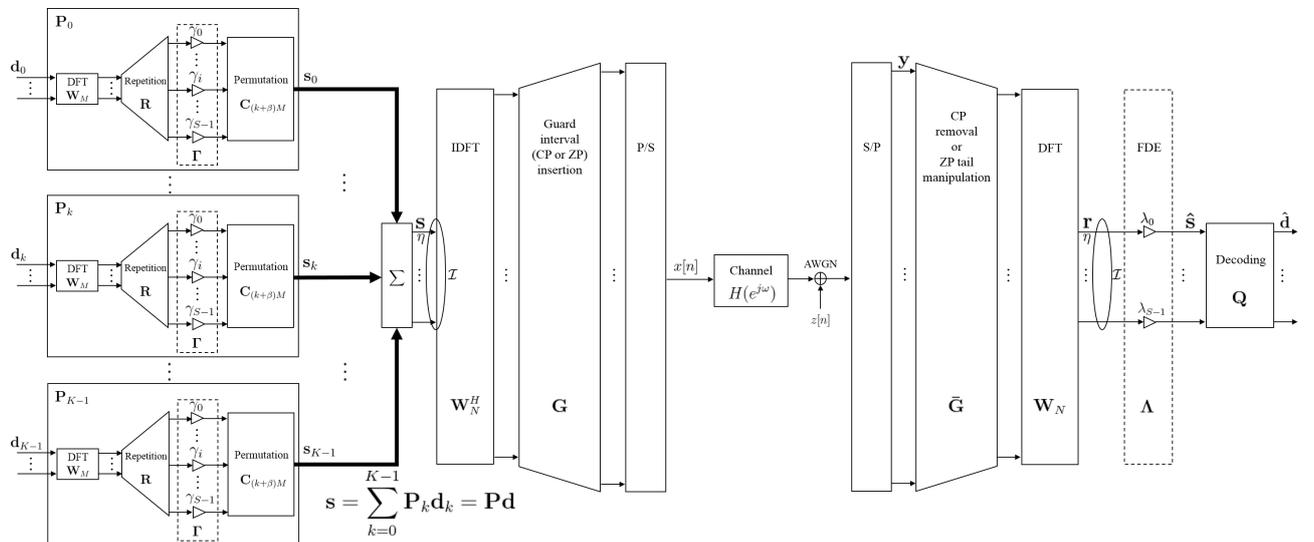


Figure 10: CPS-OFDM transceiver model

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