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CP-Less OFDM With Alignment Signals for Enhancing Spectral Efficiency, Reducing Latency, and Improving PHY Security of 5G Services

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ABSTRACT Although orthogonal frequency-division multiplexing (OFDM) is a widely accepted waveform in many standards and is expected to keep its dominance in future 5G systems with various types of parameterized waveforms, its performance in terms of spectral efficiency as well as transmission latency is usually degraded due to the excessive usage of cyclic prefix (CP). Particularly, in highly dispersive channels, CP rate might be very large in order to maintain the low-complex frequency-domain equalization. In this paper, we propose a novel method that can fit the low latency and high spectral efficiency requirements of future 5G wireless services by eliminating the need for inserting CP between successive OFDM symbols while keeping the whole detection process the same at the receiver side. In order to achieve that, we utilize specially designed alignment signals that can cancel the interference of one symbol on the other and add an additional signal component that makes the signal circularly convolved with the channel at the receiver side. Simulation results prove the superiority of the proposed scheme in terms of enhancing spectral and power efficiency, reducing latency, and improving physical-layer security against eavesdropping while using a low-complexity one-tap frequency-domain equalizer. These numerous, simultaneous, and desirable advantages have the potential to make the proposed technique a suitable fit for future 5G wireless services and applications including Internet of Things-based massive machine-type communication, ultra-reliable and low-latency communication, and enhanced mobile broadband.

INDEX TERMS OFDM, cyclic prefix (CP), guard period, 5G, spectral efficiency, latency, reliability, physical layer (PHY) security, IoT, mMTC, URLLC, eMBB, PAPR, OOBE, MIMO, complexity.

I. INTRODUCTION

Due to its numerous advantages, orthogonal frequency division multiplexing (OFDM) [1] has been the most dominantly used transmission waveform in the vast majority of the currently available standards, such as WiFi, WiMax, DVB, LTE, and NB-IoT. Moreover, it should not be a surprise that OFDM waveform (with its new parametrized waveforms to meet the diverse requirements of different emerging applications) is expected to maintain its dominance in future 5G systems [2]–[4]. It has been adopted due to its desirable features, including higher spectral efficiency, robustness to multipath with simple equalization in the frequency domain, easy integration with MIMO systems, and multi-user diversity where time and frequency resources are flexibly scheduled among users based on their requirements and channel conditions [2].

Nonetheless, OFDM has several major drawbacks, such as high peak-to-average power ratio (PAPR), spectral leakage, strict synchronization requirements and frequency offset sensitivity. These issues have been heavily studied in the literature and many solutions were proposed to mitigate their side effects [2], [5], [6]. In addition, it should be mentioned that in 5G new radio, different window based time-domain filtering schemes have been proposed such as universal filtered OFDM (UF-OFDM or UFMC) [7], [8], filtered-OFDM (f-OFDM) [9], [10], time domain windowed overlap-and-add (WOLA) based CP-OFDM [11], etc. Also, flexible and effective frequency-domain filtering scheme, based on fast-convolution (FC), was proposed for filtered OFDM [12], [13].

Besides the aforementioned issues, there are other issues which have not been studied much in the literature, but have started gaining the attention of researchers due to their significant importance for future dynamic, low latency, spectral and power efficient 5G wireless networks. Among these issues are the ones related to the non-optimal design of OFDM, which can be summarized in two aspects: 1) the excessive usage of cyclic prefix (CP) as guard times between OFDM symbols to prevent inter-symbol-interference (ISI) while providing circular convolution; and 2) the inability to optimally collect and effectively combine the leaked energy (due to channel dispersion) of the OFDM symbol to the guard time owing to the fixed waveforms length of OFDM transceiver structure.

For the issue of collecting the energy leakage, Hamamreh and Arslan [14], [15] designed a new waveform that can optimally collect the dispersive signal energy that leaks to the guard time part of the signal instead of just discarding it as is the case in standard OFDM. The designed waveform in [14], which is named as orthogonal transform division multiplexing (OTDM), results in a better reliability compared to standard OFDM due to accumulating the dispersive energy and combining it optimally with the data symbols. Also, the design yields physical layer secrecy [16] as an extra advantage that comes for free due to making the waveform channel-dependent.

On the other hand, the issue of using excessive CPs between OFDM symbols makes OFDM suffer from a significant spectral efficiency loss as well as latency increase due to using a fraction of the available time resources as guard intervals instead of utilizing them for data transmission. This motivates designing new novel solutions that can completely release and free the need for using CPs between OFDM symbols while maintaining good reliability performance as that of the standard CP-OFDM. This issue is substantially important and worth to be practically solved as it causes a huge waste of resources in some specific emerging 5G scenarios and services such as massive machine type communication (mMTC) and enhanced mobile broadband (eMBB) [17], resulting in an inability to enhance the total efficiency and transmission rate of the wireless systems. Besides, CP also causes extra unnecessary transmission delay for latency-sensitive applications such as ultra reliable and low latency communication (URLLC) services [3], [17].

To realize the significant effect of this issue and how much important it is to properly address this challenge, we give the following explanatory example. In mMTC scenarios [18], massive amount of Internet of Things (IoT) devices often receive and transmit sporadic short packets of control and sensed data to 5G base station (BS) whose resources are usually limited and may become insufficient when the BS tries to serve all these massive devices simultaneously. Let us assume we have 10000 of these IoT devices, each requires on average 100 OFDM symbols of 512 sub-carrier resources. With current standard OFDM waveform, where the CP length can be as large as one fourth (0.25) of the whole OFDM symbol, the BS would sacrifice around $10000 \times 100 \times 0.25 \times 512 = 128,000,000$ units of time resources. This is an extremely huge amount of loss in the total spectral efficiency of just one BS as these resources could be saved and then used to serve other users and IoT devices in the network.

Moreover, mitigating the CP length (denoted by R) or totally removing it will certainly help meet the requirements of enhanced mobile broadband (eMBB) services that not only require high spectral efficiency but also low transmission latency (a very desirable feature for URLLC services as well) to increase peak throughput. Particularly, latency can significantly be reduced if the net OFDM symbol duration is made equal to only the useful data part of the symbol (i.e., data subcarriers N) instead of being equal to the sum of data and guard time parts. Consequently, in low latency applications, the CP becomes a considerable overhead which not only degrades the system spectral efficiency by $N/(N+R)$, but also causes latency and energy efficiency loss. Motivated by these practical facts, it would be extremely very advantageous if one could develop practical methods that can totally remove the need for inserting CPs between OFDM symbols while maintaining the same performance as that of standard OFDM.

In the literature, there are several techniques introduced to mitigate the spectral efficiency loss and latency due to CP by either reducing (shorting) or totally removing (eliminating) the CP guard period while maintaining reasonable performance. Among the techniques used to reduce the required guard time are channel shortening schemes that are capable of reducing the CP length by either utilizing a time-domain equalizer to shorten the effective channel impulse response as in [19], or frequency-domain equalizer as in [20], or by designing proper precoding matrices as in [21] and [22]. Besides channel shortening, Taheri *et al.* [23] proposed an asymmetric window which provides a significant CP reduction without increasing out-of-band-emission (OOBE) nor causing ISI/ICI. Lorca [24] explored a scheme based on the concept of multiple symbol encapsulation, which was originally introduced in [25] to reduce CP rate. Lorca [24] showed that the scheme can noticeably mitigate the CP overhead and hence increase the throughput by using only a single CP to a group (block) of adjacent OFDM symbols instead of inserting many CPs between consecutive OFDM symbols in each transmission block. At the receiver, a frequency domain equalization technique that applies FFT and IFFT on the whole block was proposed for cancelation of any ISI/ICI resulting from multipath channels.

All the aforementioned techniques can be classified under the category of CP-short approach. On the other hand, there are other techniques that can be classified under the category of CP-free approach. These CP-free OFDM techniques aim to totally remove the CP guard times, and hence achieving the maximum gain in terms of higher spectral efficiency and lower latency. In [26], a Nyquist pulse shaping filter bank was utilized to OFDM for CP elimination. In [27], a CP-free OFDM transmission scheme, called overlapping based OFDM (Ov-OFDM), was proposed to shorten the total OFDM symbol length without guard overhead. The scheme utilizes an overlapping minimum mean square error (MMSE) frequency domain equalization to eliminate ISI between OFDM symbols.

In [28], a low-complexity frequency-domain equalizer that utilizes redundancy in the frequency domain was proposed to completely suppress ISI and ICI when no guard intervals are inserted between symbols. In [29], a different equalizer based on a multi-antenna generalized side-lobe canceler (GSC) was also proposed to suppress ISI for high-rate SIMO-OFDM systems without CP. The method basically depends on the block representation of the OFDM transmission and exploits the subspace of the ISI structure in the multi-antenna scenarios. Wang and Li [30] introduced a non-orthogonal and CP-free scheme based on maximum-likelihood sequence detection in frequency domain to enhance spectrum and power efficiency from one side and alleviate synchronization requirement from another side. The detection algorithm is termed as frequency-domain maximum-likelihood sequence detection (MLSD).

In [31], a multi-carrier detection algorithm for OFDM systems without guard time by utilizing successive interference cancelation with decision feedback was also introduced. Most recently, Liu *et al.* [32] proposed a scheme called symbol cyclic shift equalization (SCSE) to implement a CP-free OFDM system. The scheme is based on performing decision feedback equalization (DFE) before FFT operation for removing the ISI between the overlapped OFDM symbols. The scheme also uses CP restoration mechanism at the receiver to convert linear shift into circular shift, thereby allowing the use of FFT transform at the receiver. Despite its effectiveness, this scheme comes with three major demerits. First, it results in an excessive, unaffordable complexity at the receiver,[1] making it unsuitable for low complexity, battery-limited devices (such as IoT). Second, it makes the design incompatible with current and future expected devices, where minimal or even no changes in the receiver terminal are required to reduce the cost of making new reception designs. Third, its bit-error rate performance is worse than that of OFDM for certain channel types. Due to these issues, SCSE appears to be not a very good fit for future 5G services and specifically for mMTC applications with massive IoT devices, which are expected to be compact in size, light in processing, and limited in battery power consumption.

As inferred from the literature, most of the proposed CP-free and CP-short approaches are based on either utilizing equalization technologies such as MMSE, ZF, DFE, SIC and so forth; or exploiting some redundancy in the spectral or spatial domain to mitigate or eliminate CP overhead at the expense of extra complexity at the receiver or wasting other resources.

In this paper, we propose **a novel power domain[2] based-scheme, called CP-less OFDM (or CP-free OFDM),** that removes the requirement of inserting CP between successive OFDM symbols, while keeping the whole detection process the same at the receiver side without the need to use any complex equalizers. This is achieved by adding an alignment signal on top (in the power domain) of each transmitted OFDM symbol to guarantee achieving two goals simultaneously: 1) canceling the interference of one symbol on the other adjacent one, and 2) maintaining the circularity property of the received signals before the low complexity frequency domain equalization process. The obtained results demonstrate the superiority of the proposed CP-less OFDM scheme in terms of spectral efficiency, reduced latency, power efficiency, and physical layer secrecy compared to standard CP-OFDM while maintaining low complexity using simple one tap frequency domain equalizer, making the design inherently compatible with current and future systems, and also highly capable of meeting the needs of 5G and beyond wireless services.

The main advantages of the proposed scheme can be emphasized and summarized as follows:

- Enhancing spectral efficiency and throughput due to removing the time resources required by CP.
- Enhancing power efficiency as a result of saving the power consumed by CP, where the power of the alignment signal in our proposed design is less than that of CP in OFDM.
- Reducing transmission latency due to decreasing the net time duration required for the transmission of each symbol.
- Improving physical layer security [14], [16] as a consequence of making the alignment signal function of the legitimate user's channel, which is different from that of eavesdroppers located at different locations. Thus, extra interference will be caused to the eavesdroppers.
- The scheme does not require any changes or extra processing at the receiver thanks to the proper design of the added alignment signal, making it compatible with current and future networks. In particular, all the processing is done at the transmitter, making it very suitable for the downlink transmission scenario (from BS to low complexity IoT devices).

A highly critical and **emerging 5G scenario** that our proposed CP-less OFDM method is particularly suited for is URLLC type of services. This is so due to the fact that the OFDM symbol durations in URLLC services are required to be much smaller compared to the symbol durations currently being used in 4G and WiFi systems in order to provide low latency and much faster communication. However, this

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[1] Note that Liu *et al.* [33] tried to reduce the complexity of SCSE scheme by using pulse amplitude modulation (PAM) instead of M-ary quadrature amplitude modulation (M-QAM). This unfortunately can not be applied to current and future systems that mainly adopt M-QAM.

[2] The use of the term "power domain" is inspired from the power domain non-orthogonal multiple access (NOMA) concept, where the signals of different users sharing the same time and frequency resources are superimposed on top of each other in the power domain. Similarly, in our proposed scheme, since the alignment signal is superimposed on top of the time domain data signal, the multiplexing domain is termed as power domain.

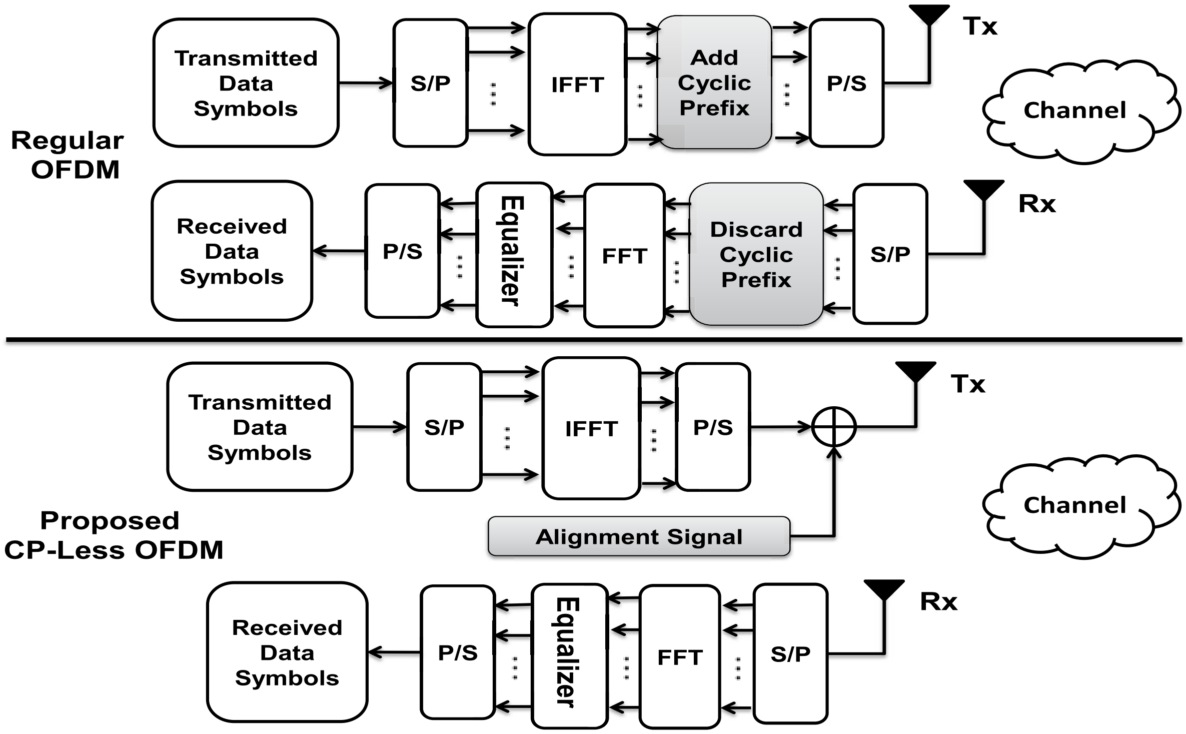

FIGURE 1. Transceiver structures of the conventional, regular OFDM and the proposed CP-less OFDM.

increases the CP overhead for a given fixed channel delay spread if conventional CP-OFDM scheme is used. Therefore, removing the CP requirement with the proposed method can result in a highly efficient transmission system. Our scheme can even become more beneficial for the scenarios of stationary channels (e.g., access point communicating with a non-moving wireless terminal in a static environment where all reflectors are almost fixed and immobile during the period between two channel estimation) where the receiver in this case does not require to frequently send pilot signals to the transmitter to update the channel estimation process at the transmitter as the channel is almost static (i.e., not varying much).

These highly desirable merits are achieved at the **expense** of a slight increase in the processing complexity at the base station alongside the need for acquiring channel state information (CSI) at the transmitter [14], [34]–[37].

The rest of the paper is organized as follows. The system preliminaries and assumptions are described in Section II. The details of the devised CP-less OFDM waveform by using two different methods are revealed in Section III. Numerical results including BER, spectral efficiency, transmission latency, power saving, secrecy performance, robustness to imperfect channel, PAPR, OOBE, complexity; are all presented, discussed and explained in Section IV, which is followed by conclusion and future works in Section V.

Notations: Vectors are denoted by bold-small letters, whereas matrices are denoted by bold-large letters. The convolution operator is indicated by $(*)$. The transpose, hermitian (conjugate transpose) and inverse are symbolized by $(\cdot)^{\mathrm{T}}$, $(\cdot)^{\mathrm{H}}$ and $(\cdot)^{-1}$, respectively. $\mathbf{I}$ is the $N \times N$ identity matrix.

II. PRELIMINARIES AND ASSUMPTIONS

A regular single-input single-output (SISO) OFDM system as depicted in the upper part of Fig. 1 is considered as a reference baseband system model upon which we develop our proposed scheme. In the regular OFDM system, the M-ary modulated frequency data symbols are first converted from serial to parallel and then passed through an IFFT block, resulting in a time domain signal vector, whose samples are coming from the contribution of all the frequency data symbols. Then, a CP of length equal to or greater than the channel spread length is appended to the time domain signal to avert ISI and provide channel circularity. The signal is then sent serially from the transmitter (Tx) to the receiver (Rx) over a multi-path frequency selective and slowly varying fading channel denoted by $\mathbf{h} = \left[h_0 \; h_1 \; \ldots \; h_R\right]^T \in \mathbb{C}^{[(R+1)\times 1]}$, where $(R+1)$ is the number of taps. The channel impulse response has $R+1$ taps, each experiences Rayleigh fading distribution. At the Rx side, the received samples are first converted from serial to parallel, and then the CP is cut and discarded. Afterwards, the samples are passed through an FFT block, followed by one tap frequency domain channel equalization to recover the M-ary modulated frequency data symbols.

Unlike the aforementioned described classical OFDM design, in the proposed CP-less OFDM scheme as portrayed in the lower part of Fig. 1, there is neither CP addition at the Tx nor CP removal at the Rx. Instead, a well-designed

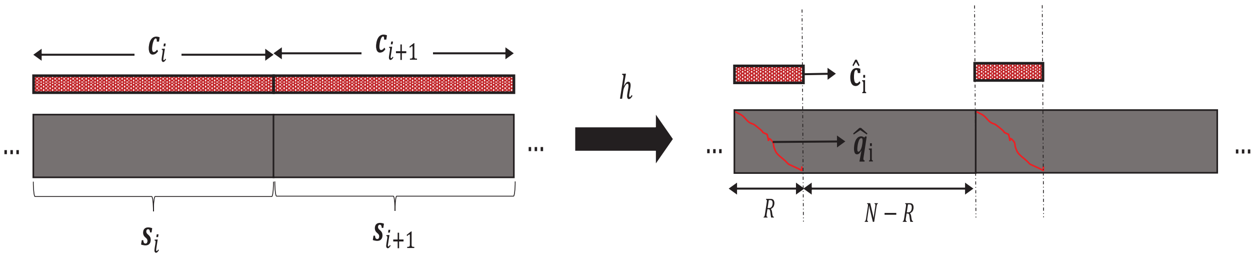

FIGURE 2. **Visualization of the designed CP-less OFDM with alignment signal superposition for two consecutive OFDM symbols.**

alignment signal (AS) is added on top of the time domain CP-free OFDM symbol in order to both cancel the ISI caused between adjacent OFDM symbols due to channel dispersion, and convert linear convolution into circular one so that low complexity frequency domain equalization can be used instead of the extremely complex time domain equalization [32].

In this work, the channel reciprocity property is adopted, where the downlink channel can be estimated from the uplink one in a time division duplexing (TDD) or hybrid systems (TDD with FDD) [34]. The channel realizations (i.e., the impulse response including its explicit complex taps and delays) are assumed to be known at the transmitter [34], [38] so that a special AS can be properly designed to enhance the spectral efficiency and reduce latency (with low complexity) of the system by eliminating the need for inserting CPs between successive OFDM symbols.

For the secrecy performance evaluation, we also consider an eavesdropper, called Eve, who tries to intercept the communication between the legitimate Tx and Rx parties, called Alice and Bob, respectively. Eve also experiences a multipath channel of independent realizations from the one experienced by the legitimate receiver as the channel de-correlates for different locations that are distant from each other by half-wave length or more [14], [16], [34].

III. PROPOSED CP-LESS OFDM DESIGN

The key concept underlying the transmission mechanism of the proposed CP-less OFDM with AS, whose transceiver structure is briefly illustrated in the lower part of Fig. 1, is clearly visualized in Fig. 2. As exhibited, a specially designed AS (colored in red) is superimposed with the OFDM time domain signal (colored in gray) in the power domain at the Tx side. After passing through the channel and reaching the Rx side, the AS accumulates itself and aligns over the interference part of the OFDM signal to cancel ISI and provide circularity. In this section, we provide the mathematical and physical meaning details of the proposed scheme at both the Tx and Rx sides.

At the Tx, the total number of frequency data symbols to be sent is N, where each transmitted OFDM block is represented as $\mathbf{x} = \left[x_0\ x_1\ \ldots\ x_{N-1}\right]^{\mathrm{T}} \in \mathbb{C}^{[N\times 1]}$. Each OFDM block $\mathbf{x}$ is then passed through an IFFT process $\mathbf{F}^{\mathrm{H}} \in \mathbb{C}^{[N\times N]}$, which basically maps the frequency data points to orthogonal subcarriers, where $\mathbf{F}$ is the discrete Fourier transform matrix. Thus, the time domain samples of each OFDM symbol can be given as $\mathbf{s} = \mathbf{F}^{\mathrm{H}}\mathbf{x}$. At this point, no CP addition is used, instead, a specially designed alignment signal (AS) of size equal to $N \times 1$ and denoted by $\mathbf{c} = \left[c_0\ c_1\ \ldots\ c_{N-1}\right]^{\mathrm{T}} \in \mathbb{C}^{[N\times 1]}$ is added on top of each CP-less OFDM signal. In other words, the AS is superimposed to the power domain of the time domain signal of the original CP-less OFDM symbol. Thus, the total resulting i^{th} CP-less OFDM signal to be transmitted can be modeled as

$$\mathbf{t_i} = \mathbf{s_i} + \mathbf{c_i} = \mathbf{F}^{\mathrm{H}}\mathbf{x_i} + \mathbf{c_i} \in \mathbb{C}^{[N\times 1]}. \tag{1}$$

Now, the key idea behind the proposed design is based on the fact that each added AS [35], [36], [39] at the Tx can be designed to perfectly align to a specific target part of the OFDM symbol after passing through the channel and reaching at the receiver side as shown in Fig. 2. Since the AS can be aligned at any specific part, it would be beneficial to align the AS to the dispersive part (due to which we need CP in standard OFDM) of the OFDM signal and try to design the value of the AS in such away that can cancel the interference that appears in that specific part of the signal. Note that the length of the dispersive part (i.e., the region of ISI) of OFDM symbol is equal to the number of channel samples, which is denoted by R. Additionally, in order to have the same functionality as that provided by CP in standard CP OFDM, an aligned AS at the receiver, $\hat{\mathbf{c}}_i \in \mathbb{C}^{[R\times 1]}$, should correspond to the summation of both an interference canceling signal $\hat{\mathbf{q}}_i \in \mathbb{C}^{[R\times 1]}$ to compensate the ISI between successive OFDM symbols and a circularity providing signal $\hat{\mathbf{z}}_i \in \mathbb{C}^{[R\times 1]}$ to maintain circularity and have simple one tap frequency domain equalization without the need to use complex time domain equalizers. Let us assume a channel impulse response of $R+1$ taps with exponentially decaying function. These taps are first estimated using channel sounding techniques [37] and then used to construct a Toeplitz matrix of size $N+R$ by N (to be used for calculating the value of the alignment signal). Then, the interference term on i^{th} signal can be calculated as

$$\hat{\mathbf{q}}_i = \mathbf{H_p}\mathbf{t_{i-1}} \in \mathbb{C}^{[R\times 1]}, \tag{2}$$

where $\mathbf{t_{i-1}}$ is the previously transmitted CP-less OFDM symbol and $\mathbf{H_p} \in \mathbb{C}^{[R\times N]}$ is the dispersive channel part, which can be given as

$$\mathbf{H_p} = \begin{bmatrix} 0 & \cdots & h_R & \cdots & h_0 \\ \vdots & 0 & \ddots & \ddots & \vdots \\ 0 & \cdots & 0 & \cdots & h_R \end{bmatrix}. \tag{3}$$

Note that $\mathbf{H_p}$ represents the interference part of the Toeplitz matrix of the channel impulse response with $R+1$ taps for a block-based transmission system. More precisely, $\mathbf{H_p}$, which is basically the last R rows of the channel Toeplitz matrix, causes dispersion and inter-symbol-interference between OFDM blocks when CP is not used. The Toeplitz matrix is denoted by $\mathbf{H} \in \mathbb{C}^{[(N+R)\times N]}$, and can be given as

$$\mathbf{H} = \begin{bmatrix} \mathbf{H_u} \\ \mathbf{H_p} \end{bmatrix} = \begin{bmatrix} h_0 & 0 & \cdots & \cdots & 0 \\ \vdots & h_0 & 0 & \ddots & \vdots \\ \vdots & \vdots & h_0 & \ddots & \vdots \\ h_R & \vdots & \vdots & \ddots & 0 \\ 0 & h_R & \vdots & \ddots & h_0 \\ \vdots & 0 & h_R & \ddots & \vdots \\ \vdots & \vdots & 0 & \ddots & \vdots \\ 0 & 0 & \cdots & \cdots & h_R \end{bmatrix}, \tag{4}$$

where $\mathbf{H_u}$ is the interference-free channel convolution matrix (i.e., $\mathbf{H}$ without the dispersion-responsible part represented by the last R rows).

On the other hand, the circularity providing signal for the i^{th} signal can be obtained as

$$\hat{\mathbf{z}}_i = \mathbf{H_p s_i} \in \mathbb{C}^{[R\times 1]}. \tag{5}$$

Then, $\hat{\mathbf{c}}_i$ which is the required signal at the receiver for interference cancelation and circularity assurance can be given as

$$\hat{\mathbf{c}}_i = \hat{\mathbf{z}}_i - \hat{\mathbf{q}}_i = \mathbf{H_p s_i} - \mathbf{H_p t_{i-1}} \in \mathbb{C}^{[R\times 1]}, \tag{6}$$

which can further be simplified into the below form

$$\hat{\mathbf{c}}_i = \mathbf{H_p}\left(\mathbf{s_i} - \mathbf{s_{i-1}} - \mathbf{c_{i-1}}\right) \in \mathbb{C}^{[R\times 1]}. \tag{7}$$

In order to design such an AS, we firstly need to find the null space of the multiplication of the interference-free channel convolution matrix (i.e., $\mathbf{H_u}$) and an imaginary CP removal matrix that creates nulls at the beginning of each OFDM symbol. Now, the real valued, virtual CP removal matrix is defined as if it removes the first R samples of each transmission symbol. Let us give that as $\mathbf{B} \in \mathbb{C}^{[(N-R)\times N]}$. Then the required null space can be calculated as

$$\mathbf{P} = \ker\left(\mathbf{BH_u}\right) \in \mathbb{C}^{[N\times R]}, \tag{8}$$

where ker( ) corresponds to kernel extracting operation, by which we can find the null space of the effective channel matrix $\mathbf{BH_u}$ so that we can use this null space to align the added AS into the specific part of the OFDM symbol. Note that any vector $\mathbf{w_i} \in \mathbb{C}^{[R\times 1]}$ multiplied by $\mathbf{P}$ and passed through the channel will give an AS aligning on the first R samples of the OFDM symbol. Note that this AS also leaks to the next symbol but we consider that as interference and will be canceled by the next symbol's AS. Therefore, the AS at the transmitter can be calculated as

$$\mathbf{c_i} = \mathbf{Pw_i} \in \mathbb{C}^{[N\times 1]}. \tag{9}$$

By doing so, we guarantee the alignment of the added AS over the desired portion of the OFDM signal at the receiver side. In the next step, we will calculate the values of $\mathbf{w_i}$ in order to obtain the desired signal, $\hat{\mathbf{c}}_\mathbf{i}$.

The i^{th} received baseband signal vector $\mathbf{y_i} \in \mathbb{C}^{[(N+R)\times 1]}$ can be given as

$$\begin{aligned} \mathbf{y_i} &= \mathbf{h} * \mathbf{t_i} + \mathbf{n_i} \\ &= \mathbf{h} * (\mathbf{s_i} + \mathbf{c_i}) + \mathbf{n_i}, \end{aligned} \tag{10}$$

where $\mathbf{y} = \left[y_0\ y_1\ \ldots\ y_{N+R}\right]^{\mathrm{T}}$, in which $y_i = \sum_{l=0}^{R} h_l t_{(i-l)} + n_{(i)}$, and $\mathbf{n_i} \in \mathbb{C}^{[(N+R)\times 1]}$ is the zero-mean complex additive white Gaussian noise (AWGN) at the Rx. The above convolution form can also be rewritten in a matrix form as

$$\mathbf{y_i} = \mathbf{Ht_i} + \mathbf{n_i} = \mathbf{H}(\mathbf{s_i} + \mathbf{c_i}) + \mathbf{n_i}. \tag{11}$$

The net received signal including the interference coming from the previous symbol can be given as

$$\begin{aligned} \mathbf{y_i} = \underbrace{\mathbf{Hs_i}}_{\text{Desired signal plus interference}} + \underbrace{\mathbf{Hc_i}}_{\text{Alignment signal}} \\ + \underbrace{\hat{\mathbf{H}}_p \mathbf{t_{i-1}}}_{\text{Interference signal}} + \mathbf{n_i}, \end{aligned} \tag{12}$$

where $\hat{\mathbf{H}}_p$ can be given as

$$\hat{\mathbf{H}}_p = \begin{bmatrix} \mathbf{H_p}^{[R\times N]} \\ \mathbf{0}^{[N\times N]} \end{bmatrix}. \tag{13}$$

In order to achieve a CP free transmission, the second term of $\mathbf{y_i}$, $\mathbf{Hc_i} = \hat{\mathbf{c}}_i$ should be $\hat{\mathbf{H}}_p\,(\mathbf{s_i} - \mathbf{t_{i-1}})$ as mentioned before. We can interpret $\hat{\mathbf{c}}_i$ as a column vector whose first R elements cancel ISI and introduce circularity, and the other N elements are supposed to be perfectly zero. The second term of $\mathbf{y_i}$ can be given as

$$\hat{\mathbf{c}}_i = \mathbf{Hc_i} = \mathbf{HPw_i}. \tag{14}$$

It can be seen that the last N elements of $\hat{\mathbf{c}}_i$ will be zero. Then, we can focus the multiplication of $\mathbf{HPw_i}$ which should give us the required values as the first R elements of $\hat{\mathbf{c}}_i$. Thus, $\mathbf{w_i}$ can readily be calculated as

$$\mathbf{w_i} = (\mathbf{HP})^{-1}\,\hat{\mathbf{c}}_i \tag{15}$$

$$= \left((\mathbf{HP})^{\mathrm{H}}\,(\mathbf{HP})\right)^{-1}(\mathbf{HP})^{\mathrm{H}}\,\hat{\mathbf{c}}_i. \tag{16}$$

By the time the properly designed AS passes through the channel and reaches the Rx, the ISI will be naturally canceled by the added AS. Thus, the received N interference-free samples of the effective OFDM symbol along with the circularity-providing signal component, can be given as

$$\mathbf{y_i} = \mathbf{H_u s_i} + \begin{bmatrix}\mathbf{H_p}\\ \mathbf{0}\end{bmatrix}\mathbf{s_i} + \hat{\mathbf{n}}_i \in \mathbb{C}^{[N\times 1]}, \tag{17}$$

where $\hat{\mathbf{n}}_i$ has the first N samples of the original vector $\mathbf{n_i}$. One can see that the above effective received OFDM symbol can be written in terms of a circular channel matrix denoted by $\mathbf{H_c} \in \mathbb{C}^{[N\times N]}$ as follows

$$\mathbf{y_i} = \mathbf{H_c s_i} + \hat{\mathbf{n}}_i \in \mathbb{C}^{[N\times 1]}. \tag{18}$$

Afterwards, the Rx applies fast Fourier transform by using the matrix $\mathbf{F}$ to get the frequency data symbols of each OFDM symbol, which can be given as

$$\begin{aligned}\hat{\mathbf{y}}_\mathbf{i} &= \mathbf{F H_c s_i} + \mathbf{F}\hat{\mathbf{n}}_i\\ &= \mathbf{F H_c F^H x_i} + \mathbf{F}\hat{\mathbf{n}}_i\\ &= \mathbf{H_f x_i} + \mathbf{F}\hat{\mathbf{n}}_i.\end{aligned} \tag{19}$$

Lastly, the Rx uses the diagonal channel frequency response $\mathbf{H_f} = diag\left[H_0\ H_1\ \ldots\ H_N\right]$ to perform simple one tap zero-forcing equalization process on $\hat{\mathbf{y}}_\mathbf{i}$ to get the final equalized data symbols block $\hat{\mathbf{x}}$, which can be given as

$$\begin{aligned}\hat{\mathbf{x}}_\mathbf{i} &= \mathbf{H_f^{-1} H_f x_i} + \mathbf{H_f^{-1} F}\hat{\mathbf{n}}_i\\ &= \mathbf{x_i} + \mathbf{H_f^{-1} F}\hat{\mathbf{n}}_i.\end{aligned} \tag{20}$$

It should be mentioned that the previous proposed design is based on the concept of CP-OFDM, which is the most commonly used OFDM scheme in wireless systems. In the following, we present an alternative design inspired by the concept of zero padding (ZP)-OFDM with overlap addition (OLA) receiver structure. This second proposed design inspired by ZP is equivalent to the first proposed design inspired by CP in the sense that both of them have the same system performance. The key difference between the two designs (methods) is that, in the aforementioned presented CP-inspired design, the added AS tries to cancel the interference coming from its proceeding adjacent symbol; whereas in the design inspired by ZP-OFDM with OLA, the AS tries to cancel the interference caused by each symbol onto its later adjacent symbol. Thus, instead of making the AS function of both the current and previous symbols as presented before, in this new alternative design, the AS will be function of only the current symbol.

Without loss of generality, in this second proposed method, one can design two signals, namely, an interference-canceling signal $\mathbf{u_i}$ to cancel the ISI caused by each i^{th} OFDM symbol onto the other right-adjacent symbol given be $\hat{\mathbf{u}}_i$, and a circularity-providing signal for the i^{th} symbol $\mathbf{z}_i$ to convert linear convolution to circular one at the receiver $\hat{\mathbf{z}}_i$, thereby enabling the use of simple one tap frequency domain equalization by performing FFT operation. Particularly, the ISI signal at the end of each OFDM symbol at the receiver can mathematically be given as

$$\hat{\mathbf{u}}_i = \mathbf{H_p s_i} \in \mathbb{C}^{[R\times 1]}, \tag{21}$$

whereas the circularity providing signal at the beginning of each OFDM symbol at the receiver can be given as

$$\hat{\mathbf{z}}_i = \mathbf{H_p s_i} \in \mathbb{C}^{[R\times 1]}, \tag{22}$$

According, the alignment signal that should be added at the transmitter to cancel ISI can be given as

$$\mathbf{H u_i} = \begin{bmatrix}\mathbf{0}^{[N\times 1]}\\ -\hat{\mathbf{u}}_i{}^{[R\times 1]}\end{bmatrix} \tag{23}$$

$$\mathbf{u_i} = \left(\mathbf{H^H H}\right)^{-1}\mathbf{H^H}\begin{bmatrix}\mathbf{0}^{[N\times 1]}\\ -\hat{\mathbf{u}}_i{}^{[R\times 1]}\end{bmatrix}, \tag{24}$$

whereas the alignment signal that should be added at the transmitter to provide circularity can be calculated as

$$\mathbf{H z_i} = \begin{bmatrix}\hat{\mathbf{z}}_\mathbf{i}{}^{[R\times 1]}\\ \mathbf{0}^{[N\times 1]}\end{bmatrix} \tag{25}$$

$$\mathbf{z_i} = \left(\mathbf{H^H H}\right)^{-1}\mathbf{H^H}\begin{bmatrix}\hat{\mathbf{z}}_\mathbf{i}{}^{[R\times 1]}\\ \mathbf{0}^{[N\times 1]}\end{bmatrix}. \tag{26}$$

Thus, the total resulting i^{th} signal to be transmitted can be modeled and calculated as

$$\begin{aligned}\mathbf{t_i} &= \mathbf{s_i} + \mathbf{u_i} + \mathbf{z_i} \in \mathbb{C}^{[N\times 1]}\\ &= \mathbf{F^H x_i} + \left(\mathbf{H^H H}\right)^{-1}\mathbf{H^H}\begin{bmatrix}\mathbf{0}^{[N\times 1]}\\ -\hat{\mathbf{u}}_i{}^{[R\times 1]}\end{bmatrix}\\ &\quad + \left(\mathbf{H^H H}\right)^{-1}\mathbf{H^H}\begin{bmatrix}\hat{\mathbf{z}}_\mathbf{i}{}^{[R\times 1]}\\ \mathbf{0}^{[N\times 1]}\end{bmatrix}.\end{aligned} \tag{27}$$

It is worth mentioning that the merits of the proposed design are not limited to just enhancing spectral efficiency and reducing latency without introducing any extra or complex processing at the Rx side, but also in providing physical layer security against eavesdropping [40], which is a very desirable merit of the proposed scheme for future inherently secure services and IoT applications. Secrecy can be provided because the AS is designed to be a function of the intended user channel, which is usually different from that of the eavesdropper [14]. Thus, investigating and quantifying the exact secrecy performance of the proposed scheme and its coexistence with other schemes and layers can be a subject of future research as it is beyond the main scope of this paper.

Besides, it should be clearly pointed out that our above-proposed algorithms are significantly different from the works reported in [41] and [42] in the following perspectives: 1) ours totally removes the CP, while [41] and [42] do not, but rather reduce it; 2) ours provides physical layer security against eavesdropping as an extra advantage due to the utilization of channel knowledge at the transmitter side, whereas [41] and [42] do not; 3) ours does not impose any changes on the receiver structure, where all the processing is put at the base station (transmitter side), making it backward compatible with the 3GPP wireless standards, in which there is no need for any modification at the mobile handset side.

Finally, the proposed concept is different from full channel pre-equalization that can be used for channel shortening and

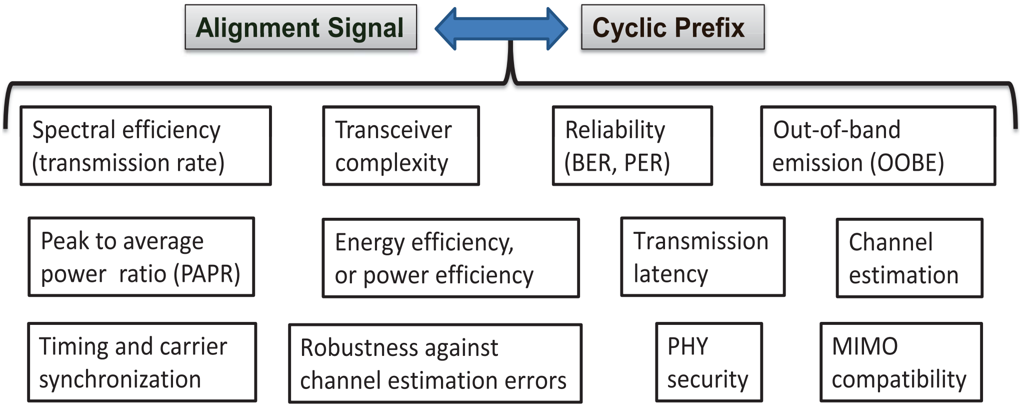

FIGURE 3. Different performance aspects and measures that can be investigated for providing thorough and comprehensive comparison between CP-less OFDM featured by AS and regular OFDM featured by CP.

thus CP-reduction in the following aspects: 1) direct, full pre-equalization is highly complex and causes significant PAPR increase due to power distribution changes; 2) our proposed scheme does not actually perform any kind of direct pre-equalization for channel shortening (which is usually performed by multiplication or division processes) to reduce the effective channel length so that less CP period can be used. Instead, the scheme generates a superposition signal that is properly designed to be added (not multiplied) on top of the time domain signal in such a way that it cancels the inter-symbol interference caused by the channel while maintaining circularity so that simple one tap frequency domain equalization can still be used just like the one used in standard CP-OFDM design.

IV. PERFORMANCE EVALUATION RESULTS AND DISCUSSION

In this section, we present the performance results of the proposed CP-less OFDM compared to regular CP OFDM. Particularly, we use transmission efficiency (bps/Hz), transmission latency (time samples), and power efficiency as main performance metrics to validate and demonstrate the gain that can be achieved by the proposed CP-less OFDM compared to CP OFDM. Moreover; BER, PAPR, and OOBE metrics are also evaluated to investigate the effect of the proposed design on these performance metrics under various conditions. A comprehensive picture of the performance aspects and measures that can be investigated for providing thorough and comprehensive comparison between CP-less OFDM featured by AS and regular OFDM featured by CP can be seen in Fig. 3. To achieve this inclusive investigation, we perform computer simulations for an OFDM signal having $N = 64$ subcarriers and modulation is determined as QAM with $M = 4$ (unless otherwise stated). The number of OFDM symbols in each frame is assumed to be 20.

The power delay profile of the adopted Rayleigh multipath time dispersive channel is defined to be an exponentially decaying with $(R+1)$ taps and expressed as

$$h(DF, n) = ae^{-DF\times n}; \quad n = 0, 1, \cdots, CDS \times N. \tag{28}$$

where a is the normalization factor, n represents the tap index, and DF is the decaying factor of the channel which is set to have the following values: $DF = 1$, $DF = 2$, and $DF = 3$. Also, CDS is defined as the channel delay spread length of the channel with respect to the OFDM symbol length (N). In this setup, CDS is configured to take the following values[3]: $CDS = 1/8$, $CDS = 1/4$, and $CDS = 1/2$ (unless otherwise stated). The channel is assumed to be changing from one frame to another independently and perfect channel estimation is assumed [14], [35], [36].

A. BIT ERROR RATE (BER)

The bit-error-rate (BER) comparison of a regular CP OFDM signal and proposed CP-less OFDM (or CP-free OFDM) signal is given in Fig. 4. As obviously seen, the proposed CP-less OFDM technique, which totally removes the redundancy of

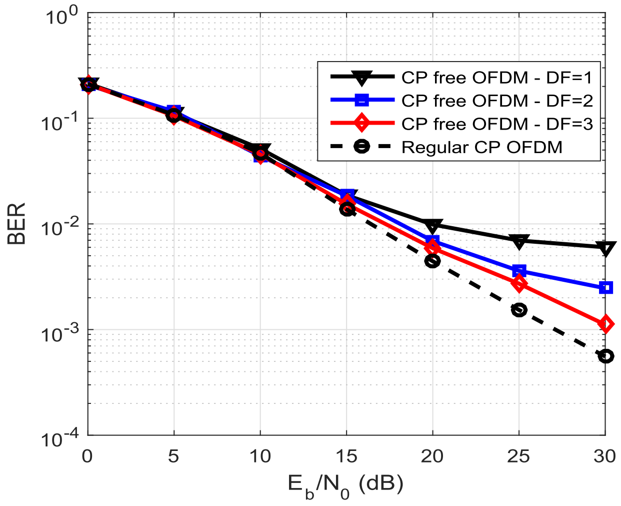

FIGURE 4. BER comparison between CP-less OFDM and CP OFDM with different channel decaying factors when the channel spread length is equal to one forth of the OFDM symbol period. QPSK modulation and $N = 64$ subcarriers are used.

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[3] $CDS = 1/8$ means that the length of the channel spread when $N = 64$ is equal to $1/8 \times 64 = 8$ samples.

CP, exhibits BER performance close to that provided by regular OFDM as the decaying factor of the channel increases, while a slight degradation, resulting in an error floor occurs as the decaying factor of the channel decreases.

After deep investigation and exploration of the reason behind this performance behavior that makes BER affected by the decaying factor of the channel (although the theory provided by the mathematical design in Section 3 shows that we can completely cancel the interference), we found out that it is generally impossible to accurately find the inverse of the term $\left((\mathbf{HP})^{\mathrm{H}}(\mathbf{HP})\right)^{-1}$, which appears in the equation of calculating the alignment signal given as

$$\mathbf{w_i} = \left((\mathbf{HP})^{\mathrm{H}}(\mathbf{HP})\right)^{-1}(\mathbf{HP})^{\mathrm{H}}\,\hat{\mathbf{c}}_i. \tag{29}$$

Computing the accurate precise inverse is impossible in some cases where the channel matrix is overdetermined with low decaying factors (i.e., tall matrix with more rows than columns and a few eigenvalues approaching zero). Consequently, the added alignment signal will not be effective in such cases and will not do its assumed job, which is to cancel the interference and provide circularity. In this case, since the added alignment signal will not be able to perfectly cancel the interference, a certain amount of performance degradation is anticipated. To avoid this degradation, we can measure the length of the interference in terms of time samples and then add a CP of length equal to that of the interference. Another way to avoid this interference is to use decision feedback equalization (DFE) with CP restoration algorithm as proposed in [32]. Besides the aforementioned reason related to matrix inverse, it is also observed that the use of MATLAB pseudo inverse function for overdetermined matrices (which is the case in our design $\mathbf{H}$) results in a noticeable sensitive inaccuracy. This happens as the pseudo-inverse function has a certain level of tolerance that must be used in order to be able to find an approximate pseudo-inverse that is usually calculated using a least square estimate[4] since an accurate result is impossible. This inherent mathematical difficulty results in a slight interference when the channel decaying factor is small, yielding a slight BER degradation as shown in Fig. 4 of the BER performance. To overcome such a difficultly, more accurate and advanced mathematical tools for calculating the inverse with higher accuracy and lower tolerance error are needed to be investigated in future research studies.

B. TRANSMISSION EFFICIENCY

Fig. 5 shows the superiority of the transmission efficiency[5] of the proposed CP-free OFDM design compared to CP OFDM where the CP length is set to be equal to the channel delay spread (i.e., 16 samples, which is the one fourth of the whole OFDM symbol length). It is clear from Fig. 5 that around 0.4 bps/Hz transmission efficiency gain as a result of using the proposed CP-less OFDM compared to regular OFDM. Note that the maximum spectral efficiency that can be achieved at high SNR values can be calculated according to the below formula that takes CP length into account

$$\eta = \frac{\log_2 M \times N}{(N+R)}, \tag{30}$$

where for the case of CP-less OFDM, R is always equal to zero as there is no CP overhead, whereas for the case of CP OFDM, R is equal to the channel delay spread of the channel at minimal.

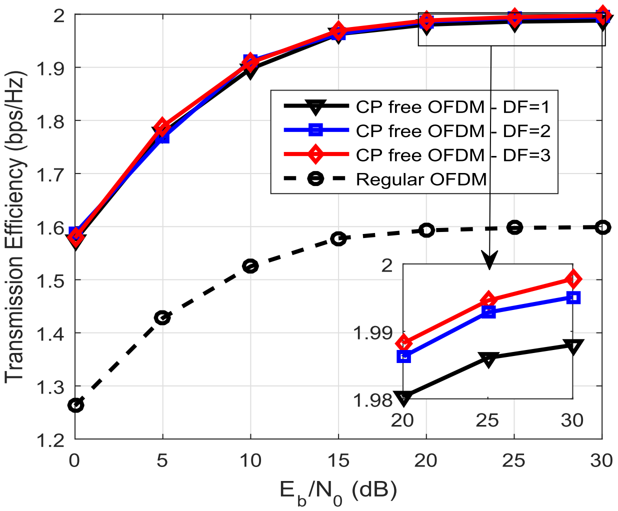

FIGURE 5. Transmission efficiency comparison between CP-less OFDM and CP OFDM with different channel decaying factors when the channel spread length is equal to one forth (1/4) of the OFDM symbol period, QPSK modulation is used, and $N = 64$.

Fig. 6 is plotted to compare the transmission efficiency between CP-less OFDM and regular OFDM when different channel delay spread (*CDS*) values are used. It is shown that the maximum achievable transmission efficiency values of regular OFDM are 1.35, 1.6, and 1.8 for *CDS* values of 1/2, 1/4, and 1/8, respectively. On the other hand, the transmission efficiency of CP-less OFDM is shown to be independent of the *CDS* and always reaches its maximum value at high SNR, which is equal to 2 bps/Hz when QPSK is used. This clearly demonstrates the fact that the spectral efficiency gain difference between CP-less OFDM and CP OFDM increases as the channel delay spread becomes longer. It should be mentioned that the gain in transmission efficiency is achieved at the cost of extra processing at the transmitter alongside

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[4] It should be mentioned that the calculation of approximate pseudo-inverse using least square estimate yields a certain level of complexity at the transmitting BS at the benefit of having minimal equalization complexity at the receiving IoT device. This complexity at BS can be affordable due to having higher processing power capabilities at the BS compared to that available at IoT devices, which have constraints on the power and processing capabilities. For this reason, the proposed CP-less OFDM scheme is best suitable to be implemented in the downlink scenarios (transmission from BS to IoT device). Thus, similar to LTE standard, the frame structure of uplink will use SC-FDMA (SC-FDE), whereas downlink will adopt CP-less OFDM instead of CP-OFDM.

[5] The transmission efficiency is calculated in the same manner as done in the literature of CP-less OFDM design [32], [33].

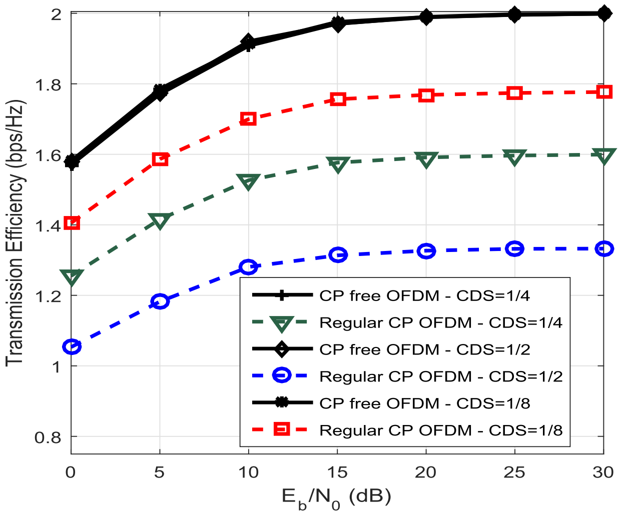

FIGURE 6. **Transmission efficiency comparison between CP-less OFDM and CP OFDM with different channel delay spread lengths, when QPSK modulation is used, *DF* = 1, and *N* = 64.**

the need to use accurate channel estimators with robust calibration techniques [37]. The validity of the received data is directly related to the reliability of the received data, which is measured by average BER metric as shown in Fig. 4.

Furthermore, Fig. 7 sketches the maximum achievable transmission efficiency versus SNR of the proposed scheme compared to regular OFDM under the effect of different modulation orders when the *CDS* is set to be 1/4. It is evident that there are gains of 1.2, 1, 0.8, and 0.4 for modulation types of 64QAM, 32QAM, 16QAM, and 4QAM, respectively. This states that the gain difference enhances as the modulation order increases.

C. TRANSMISSION LATENCY

One of key merits of the proposed CP-less OFDM scheme is that it reduces the latency and delay of data transmission. Thus, it is of importance to quantify the exact amount of latency reduction provided by the proposed CP-less OFDM scheme. Fig. 8 presents the transmission latency caused by regular OFDM with different CP lengths (which are equal to the delay spread of the channel) compared to that of the proposed scheme, where the alignment signal is used instead of CP, under the effect of different channel delay spreads. It is evident from the figure that the delay caused by regular OFDM increases as the *CDS* gets longer since the CP period extends more in time to protect against time dispersion, and thus the whole OFDM transmission time increases. On the other hand, CP-less OFDM scheme shows fixed transmission delay at all *CDS* values (i.e., it is delay spread-independent). This is due to the fact that the alignment signal does not occupy any temporal resources as it is just superimposed and added on top of the OFDM symbol in the power domain. Thus, the transmission latency of CP-less OFDM is always equal to the time period of useful data part of the CP-less OFDM symbol.

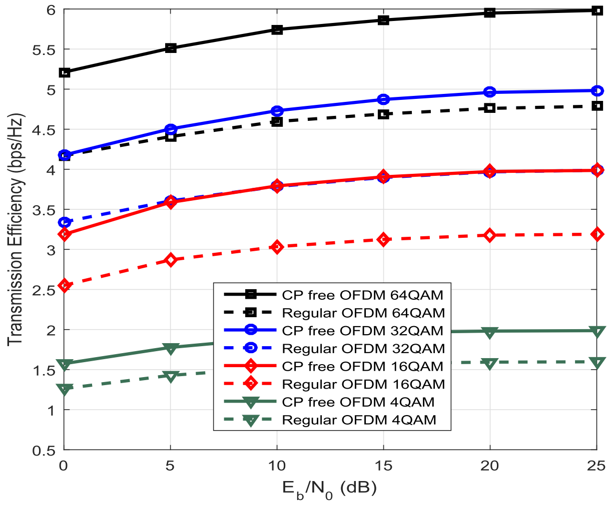

FIGURE 7. **Transmission efficiency comparison between CP-less OFDM and CP OFDM with different modulation orders when the channel spread length is one forth of the OFDM symbol period.**

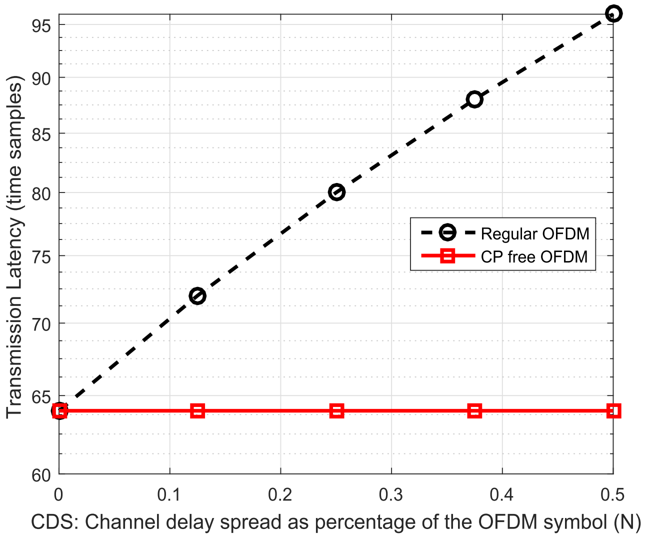

FIGURE 8. **Transmission latency comparison between CP-less OFDM and CP OFDM versus channel delay spread length.**

D. POWER EFFICIENCY

Since the proposed CP-free OFDM scheme totally eliminates the CP guard period, the power consumed by CP in regular OFDM will be saved. However, the proposed scheme on the other hand uses AS which also consumes a certain amount of power to suppress the ISI and provide circularity. Therefore, it is intuitively expected that there will be a trade-off relation between the power saved due to not using CP and the power consumed due to using AS. Accordingly, it would be interesting if we compare the power efficiency of CP-less OFDM with that of regular CP OFDM.

Fig. 9 plots the complementary cumulative distribution function (CCDF) of the sum power consumed by the CP part of regular OFDM, and the sum power distribution of AS (which is function of the channel) used in CP-less OFDM versus different CDS values when $DF = 1$. The figure shows that the power consumed by AS increases slightly as the

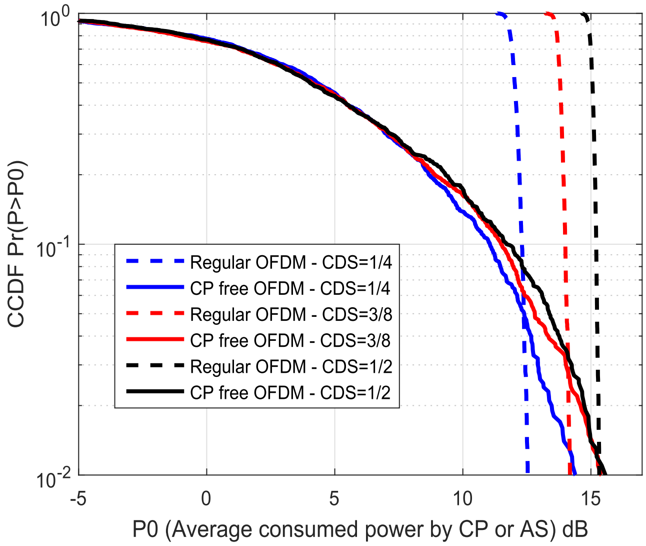

FIGURE 9. **Power efficiency comparison between CP-less OFDM and CP OFDM for different channel delay spread length. Modulation is QPSK, $N = 64$, and $DF = 1$.**

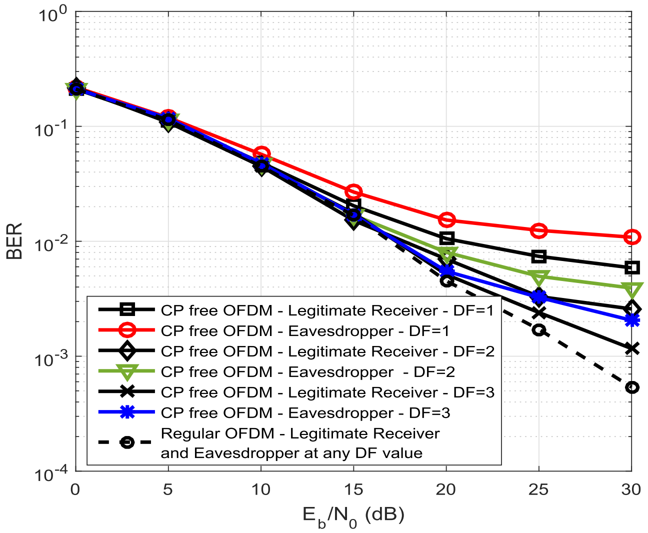

FIGURE 10. **Secrecy performance of the proposed CP-less OFDM using BER secrecy gap between the legitimate receiver and eavesdropper versus SNR at different channel decaying factors (DF). Modulation is QPSK, $N = 64$, and $CDS = 3/8$.**

CDS increases, whereas the power consumed by CP increases significantly as CDS goes higher. As shown from the figure, the total average power consumed by AS is just 0.05 percent of the time greater than that consumed by CP when $CDS = 1/4$ and it is about 0.01 percent of the time when $CDS = 1/2$. In other words, the probability that the power consumed by AS greater than that of the CP is 0.01 when $CDS = 1/2$. Based on these results, one can conclude that the power efficiency of the proposed CP-less OFDM is much better that that of regular CP OFDM scheme. This specific merit makes the proposed design suitable for green type communication, which is very crucial requirement for future environmentally friendly and health-friendly communication systems.

E. SECRECY PERFORMANCE

Besides the aforementioned verified and validated advantages of the proposed CP-less OFDM scheme in terms of low equalization complexity at the Rx, enhanced spectral efficiency, reduced latency, and better power efficiency; there is another extra advantage related to providing physical layer security against eavesdropping [16], [43], [44]. This additional merit related to security comes for free as a result of designing alignment signals dependent on the channel of the legitimate receiver, which is naturally different from that of the eavesdropper whose location is normally assumed distant by at least half wavelength from that of the legitimate receiver [14], [16]. Consequently, a secrecy gain will occur as the added AS (which is function of the legitimate receiver's channel) at the Tx will not be aligned or canceled at the eavesdropper side, but rather causing extra interference and resulting in a tangible degradation in the performance.

To investigate the secrecy performance provided by the proposed scheme, we use BER-based secrecy gap between the legitimate receiver and eavesdropper as a security performance measure [14], [16], [45]. Fig. 10 presents the secrecy performance of the proposed CP-less OFDM using BER secrecy gap versus SNR at different channel decaying factors (DF). It is clear from the figure that the eavesdropper performance is worse than that of the legitimate receiver. For instance, when $DF = 1$, there is a gap of around 10 dB between the two receivers (i.e., the legitimate receiver and eavesdropper) at BER $= 10^{-2}$. Such a large gap in performance can allow providing quality of service (QoS)-based security services [46], where the BER of the legitimate receiver is made less than a target BER (e.g., $BER = 10^{-2}$) to satisfy the required quality of a certain service (e.g., voice service), while eavesdropper's BER is made above this target BER to prevent eavesdropper from satisfying the QoS required for that specific service. It should be noted that this performance gap, which is obtained by considering uncoded system, may get smaller when advanced channel coding is used. Thus, it would be interesting to consider studying and investigating the effect of different coding schemes on the performance gap between Bob and Eve. This is left for future research studies on this topic as it is beyond the scope of this paper.

F. ROBUSTNESS AGAINST CHANNEL ESTIMATION ERRORS

To evaluate the robustness of the proposed CP-less OFDM scheme against imperfect channel estimation, we add intentional errors denoted by ($\Delta\mathbf{h}$) to the true channel ($\mathbf{h}$) in order to obtain new erroneous channels given by $\hat{\mathbf{h}} = \mathbf{h} + \Delta\mathbf{h}$ [14]. $\Delta\mathbf{h}$ is modeled as an independent complex Gaussian noise with zero mean and variance $\sigma^2 = mse \times 10^{\frac{-SNR_{dB}}{10}}$, where mse is a variable related to the mean square error of the estimator's quality. Then, the alignment signal is designed based on this erroneously estimated channel $\hat{\mathbf{h}}$, which is different from the

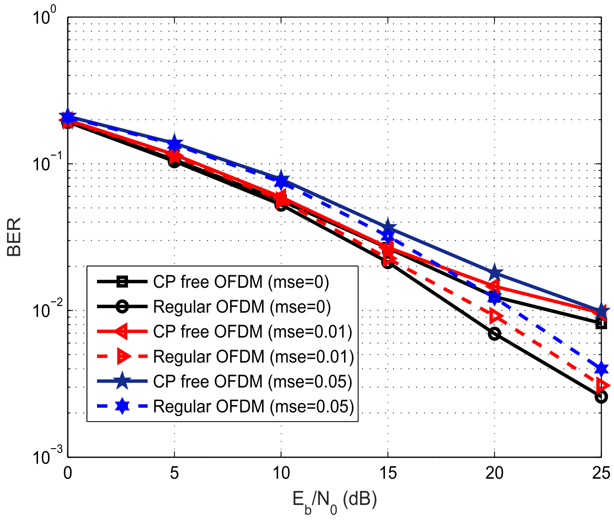

FIGURE 11. BER comparison between CP-less OFDM and CP OFDM with imperfect channel estimation factors when the channel delay spread (*CDS*) length is equal to one forth of the OFDM symbol period and decaying factor (*DF*) is 1. Modulation is 16-QAM and the number of subcarriers is $N = 64$.

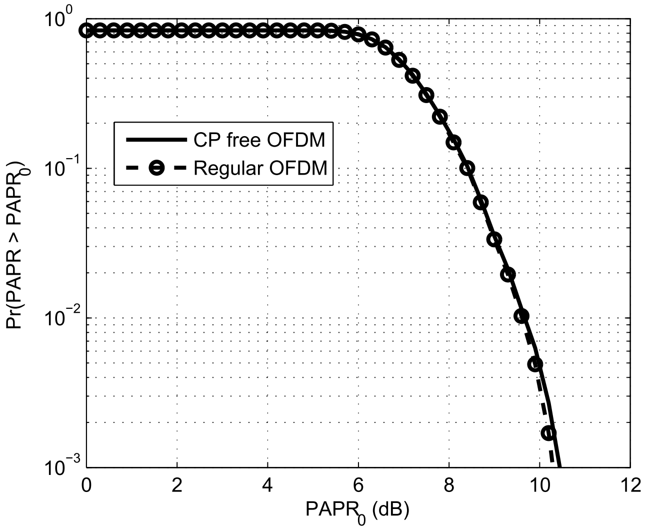

FIGURE 12. PAPR comparison between CP-less OFDM and CP OFDM. Modulation is 16QAM, $N = 64$, $CDS = 1/8$, and $DF = 2$.

true channel $\Delta\mathbf{h}$. Fig. 11 presents the performance under different estimation qualities with $mse = 0$ (perfect estimation), $mse = 0.01$ and $mse = 0.05$. It is shown that imperfect channel estimation leads to a small degradation due to having error mismatch between the alignment signal generated from the estimated channel and its actual value coming from the true channel. This results in a small performance degradation in the BER as the added alignment signal in this case will not perfectly cancel the actual interference coming from the true channel. However, this degradation can be further mitigated by increasing the length or power of the training sequence.

G. PAPR AND OOBE

In order to check how AS affects the other critical waveform characteristics in terms of out-of-band emission (OOBE) and peak-to-average-power ratio (PAPR), we also plot the CCDF of PAPR and power spectral density (PSD) in Fig. 12 and Fig. 13, respectively. Due to the usage of an additional signal, a very small increase in PAPR and OOBE (< 0.2 dB) is observed which is not considerably effective in system performance.

H. COMPLEXITY

It should be stated that the proposed CP-less OFDM scheme costs extra number of multiplications compared to conventional CP-OFDM. The extra processing cost at the transmitter is resulted from the need to calculate: 1) the aligning matrix $\mathbf{P}$ that consists of the null space vectors of $\mathbf{BH_u}$ with complexity of $O(N^2(N-R))$, and 2) the added signal $\mathbf{w}$, which is function of the inverse term given by $(\mathbf{HP})^{-1}$ with complexity of $O(N(N+R)R)$. However, due to the sparsity nature of the channel Toeplitz matrix $\mathbf{H}$ where there are many zero elements, the computational complexity of the proposed scheme can be significantly reduced by using advanced computing algorithms that take sparsity into account [47]. In particular, a sparse matrix can be solved with efficient algorithms that are capable of reducing complexity from $O(N^2)$ to $O(N)$, which represents linear complexity with sparse representation. Thus, the net complexity of the proposed scheme can at least be reduced to $O(\sqrt{(N-R)N^2}) + O(\sqrt{N(N+R)R})$.

I. COMPATIBILITY WITH MIMO

Since multiple-input multiple-output (MIMO) is an essential technology that has been adopted in 4G and 5G systems to enhance their performances in terms of data rates and reliability. The applicability of the proposed scheme for MIMO is highly desirable. It should be stated that although the structure of the proposed scheme is based on OFDM waveform, which is known for its easy integration with MIMO [2] due to its orthogonality where simple equalization can be performed for each subcarrier and antenna separately; CP-less OFDM scheme, which is proposed for SISO systems in this work, may require some modifications to make it fully compatible with MIMO due to having multiple channels interaction that may impact the orthogonality among subcarriers when CP is not used. Therefore, the compatibility of the proposed scheme with different modes of MIMO (such as transmit diversity, spatial multiplexing, and spatial modulation) will be a subject of future research studies as this is beyond the scope of this paper.

J. CHANNEL CODING

It should be mentioned that for the sake of concentration, simplicity and avoiding distraction in the presentation, we have considered an uncoded system as is the case in many research studies available in the literature, especially the recent ones related to CP-free/CP-short OFDM design as in [19]–[33]. In addition, this is assumed as so due to the fact that the

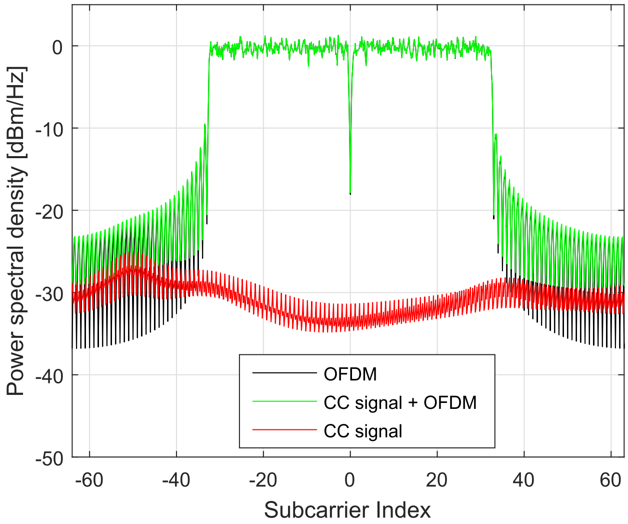

FIGURE 13. OOBE comparison between CP-less OFDM (with CP Canceling (CC) signal) and CP OFDM. Modulation is 16QAM, $N = 64$, $CDS = 1/8$, and $DF = 2$.

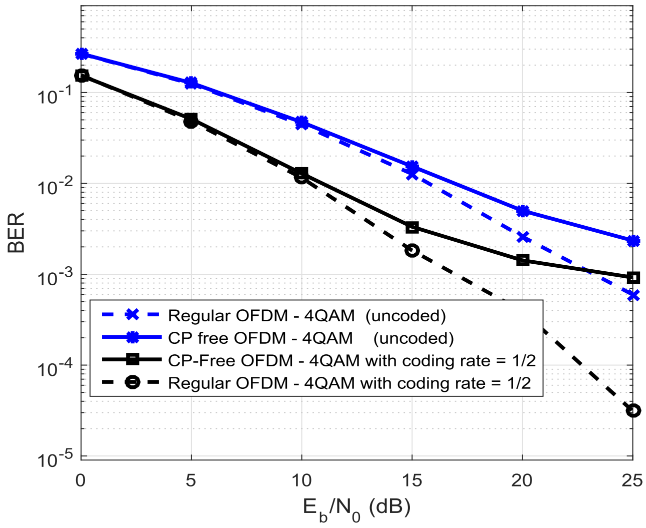

FIGURE 14. BER comparison between CP-less OFDM and CP OFDM with convolutional channel coding of rate equals to 0.5. The channel spread length is equal to one forth of the OFDM symbol period and decaying factor (DF) is 1. Modulation is 4QAM and the number of subcarriers is $N = 64$.

proposed design does not rely on coding to compensate the effect of inter-symbol interference (ISI) caused by not using a cyclic prefix in a time dispersive channel, but rather it utilizes a novel, special alignment signal design to remove ISI from one side and to achieve circularity from another side. However, when considering a coded system, which is typically the case in practical systems, then further improvement in the BER performance is expected to occur according to the coding rate and decoding algorithm adopted. To evaluate the impact of channel coding on the performance of the proposed scheme, we have performed additional computer simulations with channel coding. Fig. 14 shows the BER of the proposed technique under the effect of channel coding. Particularly, a convolutional encoding scheme and Viterbi decoding algorithm are adopted at the transmitter and receiver sides, respectively. The convolutional encoder has rate = 1/2, constraint length = 3, and generator polynomials given in binary domain as $[111; 101]_2$. It is clear from Fig. 14 that channel coding results in further improvement in the BER performance as expected.

K. SYNCHRONIZATION

Synchronization is an important aspect for the successful operation of the proposed design. It is very essential to identify the beginning of the frame so that exact alignment of the superimposed signal and thus zero interference can be guaranteed. In general, synchronization in OFDM is achieved by either transmitting a known preamble (training sequence) or by utilizing the CP structure [35]. Preamble-based synchronization techniques can be conveniently used with our proposed design, where the alignment signal is not generated during the synchronization phase. However, to substitute the absence of the alignment signal during the synchronization phase, suitable guard period has to be used to avoid interference.

CP-based synchronization is based on the fact that the CP samples are equal to the data samples situated at the end of the OFDM symbol. These similar samples in the CP and the data portion of the OFDM symbol are spaced by N samples apart. Using a sliding window correlator, this information can be used to determine the start of the OFDM symbol. However, after applying the proposed scheme with alignment signal, there will be no CP. As such, the CP may no longer be utilized for synchronization purposes. To overcome this issue, the alignment signal can be designed so that it leaves part of the CP and the corresponding samples in the data duration of the OFDM symbol unaffected. Accordingly, part of the CP samples are used by the alignment signal for interference suppression while the rest are used for synchronization as well as guard protection. This partial CP usage is only used during the synchronization phase; once synchronization is achieved, the full CP length can be utilized by the proposed design.

V. CONCLUSION AND FUTURE WORK

This work has proposed a CP-free OFDM design that eliminates the need of using excessive CPs between OFDM symbols. The design is shown to increase the spectral efficiency, enhance power efficiency, reduce latency, and improve physical layer security while maintaining low receiver complexity, making it a strong candidate for meeting the requirements of future 5G and beyond services and applications. Particularly, a novel power domain-based method that removes the requirement of CP while keeping the whole detection process the same at the receiver side is achieved by using a special design of alignment signals. These signals are added in the power domain of the transmitted OFDM symbols in order to achieve two goals simultaneously: 1) removing the inter-symbol-interference between symbols and 2) making the signal circular at the receiver side. Simulation results showed that spectral and power efficiency got enhanced, latency got reduced, and secrecy got improved while maintaining low complexity equalization at the Rx side.

Future work: Most of the performance aspects shown in Fig. 3 have been investigated in this paper. However, still there are other important aspects related to CP-less OFDM with AS which need to be studied and deeply investigated such as the effect of timing and carrier synchronization issues and how it should be performed in the proposed CP-free scheme. Moreover, channel estimation and how to handle error mismatch are also important issues to deal with. Besides, quantifying and checking the suitability of using the proposed scheme as a physical layer security technique for providing secrecy is as significant as the other performance measures. In addition, measurements and practical test bed implementations are also needed in order to validate the advantages of the proposed scheme and compare simulations with real conditions. All these factors and aspects are promising future research directions to be conducted in upcoming studies so that a complete picture of the proposed design including its merits and demerits can be carefully drawn and fully understood. As a final challenge, it is very substantial to come up with mathematical tools and techniques that can enable finding an accurate (or very close to accurate) inverse of the channel tall overdetermined matrix that appears in the calculation of the AS signal. This will help guarantee having precise, accurate AS that ensures causing no interference and thus having intact BER performance.

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