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A Novel Flip-Filtered Orthogonal Frequency Division Multiplexing-Based Visible Light Communication System: Peak-to-Average-Power Ratio Assessment and System Performance Improvement

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Abstract: Filtered orthogonal frequency division multiplexing (F-OFDM), employed in visible light communication (VLC) systems, has been considered a promising technique for overcoming OFDM's large out-of-band emissions and thus reducing bandwidth efficiency. However, due to Hermitian symmetry (HS) imposition, a challenge in VLC involves increasing power consumption and doubling inverse fast Fourier transform IFFT/FFT length. This paper introduces the non-Hermitian symmetry (NHS) Flip-F-OFDM technique to enhance bandwidth efficiency, reduce the peak-average-power ratio (PAPR), and lower system complexity. Compared to the traditional HS-based Flip-F-OFDM method, the proposed method achieves around 50% reduced system complexity and prevents the PAPR from increasing. Therefore, the proposed method offers more resource-saving and power efficiency than traditional Flip-F-OFDM. Then, the proposed scheme is assessed with HS-free Flip-OFDM, asymmetrically clipped optical (ACO)-OFDM, and direct-current bias optical (DCO)-OFDM. Concerning bandwidth efficiency, the proposed method shows better spectral efficiency than HS-free Flip-OFDM, ACO-OFDM, and DCO-OFDM.

Keywords: complexity reduction; Flip-F-OFDM; PAPR; spectral efficiency; visible light communication



Received: 17 December 2024

Revised: 10 January 2025

Accepted: 12 January 2025

Published: 15 January 2025

Citation: Hujjo, H.S.R.; Ilyas, M. A Novel Flip-Filtered Orthogonal Frequency Division Multiplexing-Based Visible Light Communication System: Peak-to-Average-Power Ratio Assessment and System Performance Improvement. *Photonics* **2025**, *12*, 69. <https://doi.org/10.3390/photonics12010069>

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1. Introduction

Recently, attention has been directed towards visible light communication (VLC) systems, which have been indicated as having the potential for enhancing wireless communications using radio frequencies (RFs). The claim of high-speed wireless transmission has risen nowadays due to the limitations of radio frequencies. These limitations are due to spectrum determinations, unrestricted frequency bands, and open signal propagation [1]. Furthermore, the growing number of users led to considerable spectrum congestion in RF-based wireless telecommunication.

In VLC, high-level frequency bands are used in visible light that belongs to the visible light spectrum, offering a new transmission medium compared to RF communication [1,2]. VLC, which utilizes the visible light spectrum, can offer substantial impedance to electromagnetic interference and high transmission rates because of its bandwidth advantages.

As a result, it has rapidly emerged as a prominent research topic in wireless communication [2]. Orthogonal frequency division multiplexing (OFDM) has been widely observed to be a groundbreaking source of knowledge that uses a number of orthogonal subcarriers to transfer data. OFDM presents several merits, including the use of the high-frequency spectrum and its impedance of multipath interference [3]. Despite its various advantages, OFDM has a bandwidth limitation due to its considerable out-of-band emissions, which cause spectral inefficiency [4]. Hence, a filtered OFDM scheme can handle this problem [5].

Recently, the layout and execution of employing OFDM in direct detection and intensity modulation (DD/IM) technology have been scrutinized in optical communications. The issue is that the signal power must be present in nonnegative and noncomplex amounts, but the OFDM baseband signal is present in nonreal and negative amounts [6,7]. Regarding real signal production for intensity modulation in OFDM systems, two main approaches are typically used [8]. Firstly, the complex signal of the inverse fast Fourier transform (IFFT) output is passed through other converting processes to create a non-negative-valued signal. Despite the requirements of this technique and the need for additional hardware, like a digital-to-analog converter (DAC) and an in-phase and quadrature (IQ) mixer, the compensation of the carrier frequency equalizer may also be required for effective application [9]. The other technique for generating a real-valued signal in OFDM systems is harnessing the Hermitian symmetry (HS) limitation on the input datum at the IFFT. This technique permits the direct generation of a real-valued signal without the need for the additional hardware mentioned previously [9]. The HS constraint, often used in Discrete Multi-Tone (DMT) techniques, produces a real-valued signal that can be transmitted immediately via the optical channel [10,11]. The HS algorithm allows a noncomplex OFDM signal to be generated by constraining the frequency symbols such that the input of the IFFT satisfies the HS constraint [12,13]. The FFT and the reverse operation, the IFFT, together form the essentials of an OFDM system, and they play an important role in orthogonal subcarrier data modulation and demodulation [14].

Indeed, the desired length of the IFFT/FFT directly influences the OFDM transceiver power consumption, peak-average power ratio (PAPR), and chip size. While expanding the length of the IFFT/FFT, the complex multiplication and addition processes increase, leading to a higher power consumption. Additionally, the increased number of processing elements required for larger IFFT/FFT sizes may result in a larger occupied chip area, which can also increase the overall cost of the OFDM transceiver [14]. Furthermore, the HS constraint in optical OFDM schemes leads to doubled FFT/IFFT blocks. Specifically, using a $2N$ -point subcarrier IFFT in the modulator results in an N -point subcarrier FFT at the demodulator. This increases the energy usage and the employed chip size of the transceiver and imposes computational complexity and hardware cost constraints [15]. Moreover, increasing the subcarrier number leads to a cumulative effect of the PAPR [16].

The related literature has studied several unipolar OFDM schemes. These schemes were designed according to various frequency domain constraints and further expanded by specific time domain processes. Some popular, substandard examples of these schemes are DC-bias optical OFDM (DCO-OFDM) and asymmetrically clipped optical OFDM (ACO-OFDM) [17–19]. DCO-OFDM employs DC biasing to append a positive amount of DC to the signal, thereby elevating the nonpositive polarity of the dipolar signal [19]. DCO-OFDM is extensively used in VLC [20,21]. However, it experiences depressed power efficiency because it relies on DC bias. A significant DCO-OFDM defect is represented by the high-level peak-average power ratio, which escalates with an increase in the input data. A considerable DC bias must be increased in the signal to counteract the high negative peak amplitudes. This extensive DC bias results in reduced power efficiency and aggravates fiber nonlinearity [22]. Many approaches to reducing the PAPR have been discussed in the

literature to decrease this defect. The pilot-assisted (PA) technique, introduced in [23], is regarded as an effective method for reducing the PAPR in DCO-OFDM-based systems.

Conversely, cutting the nonpositive signal polarity in ACO-OFDM converts it into a unipolar form, introducing clipping noise. Although this method decreases spectral efficiency by half, the signal can still be identified similarly to any OFDM signal by eliminating the non-odd carriers that bear the cutting noise [24]. On the other hand, Flip-OFDM is a nondipolar signaling system that saves the nonpositive signal values by negating the polarity. This method considers a technique to offset the loss in spectrum efficiency compared to ACO-OFDM [17,22,24]. Flip-OFDM divides the two components of the signal (negative; positive) from a real OFDM bipolar symbol to keep the Hermitian symmetry of the data symbols to be transmitted. The negative components' polarity is squared to transmit the positive and negative components, one OFDM symbol after another. The fact that the signal transmitted is always positive allows Flip-OFDM to be classified as a unipolar OFDM system applicable to unipolar communications [22,25]. However, Flip-OFDM and ACO-OFDM experience diminished spectral capacity because they utilize a sole half of the subcarriers compared to DCO-OFDM. To enhance OFDM, ADO-OFDM was introduced, combining the strengths of DCO-OFDM and ACO-OFDM [26]. Nevertheless, a common drawback of these OFDM tactics is the out-of-band (OOB) emanations, which can lead to significant interference with adjacent frequency bands. A hybrid DCO-OFDM-PTM scheme was introduced in [27], which integrates DCO-OFDM-PWM and DCO-OFDM-PPM under varying channel conditions, demonstrating reduced PAPR and enhanced BER performance compared to ACO-OFDM and ACO-OFDM-PTM. Moreover, in [28], two OFDM-based QG-MIMO schemes, TD-QGSM and TD-QGSMP, are introduced for a VLC system to enhance spectral efficiency and bit error rate performance. The simulation results indicate that performing IQ decomposition of constellation signals before OFDM modulation significantly improves the system's spectral efficiency.

Many flipped OFDM formats based on HS have been suggested in the literature to boost the efficiency of the system. Nonetheless, because of the HS restriction, just one-half of the subcarriers will be able to transmit information, as mentioned previously, leading to a doubled IFFT size in the transmitter. In [29], the efficacy of Flip-OFDM and ACO-OFDM was compared, and Flip-OFDM was considered to have a 50% reduction in receiver equipment intricacy in comparison with ACO-OFDM. At the same time, the performance of both methods was equivalent. The researchers in [30] mentioned that Flip-OFDM constructions are time-compression types of classical Flip-OFDM. The model outcomes indicated that this time compression significantly enhances both efficiency bandwidth and DC electrical power. Moreover, in [31], the authors proposed a Flip-OFDM system employing channel coding to ameliorate the signal–noise ratio, while in [32], the researchers enhanced the system performance using an iterative receiver for the Flip-OFDM scheme. On the other hand, the authors in [33] examined the Flip-Wavelet Packet Modulation (Flip-WPM) as a substitutional method for Flip-OFDM to produce a unipolar signal. The study showed that Flip-WPM required less computational difficulty and was evenly power-efficient.

Recent studies [34–36] have presented the conception of filtered OFDM (F-OFDM). F-OFDM divides the bands of the OFDM signal into several sub-bands, each configured with a specific configuration, and applies filtering to these sub-bands [35]. This approach offers advantages over traditional OFDM, including better OOB suppression, support for asynchronous transmission, PAPR reduction, enhanced efficiency spectral, and simplified designing [34]. This study aims to boost spectral efficiency, degrade system intricacy, and explore the effect of the PAPR by implementing a new scheme based on Flip-Filter-OFDM. This approach generates real signals corresponding with IM/DD systems without requir-

ing the usage of Hermitian symmetry, thereby reducing the intricacy and consumption of power in the system while improving spectral efficiency.

The rest of this article is structured as follows: Section 2 comprehensively describes the proposed Flip-Filtered OFDM scheme and its mathematical formulation, along with a comparison to the conventional technique. Section 3 presents and analyses the simulation results, focusing on bandwidth efficiency, signal-to-noise ratio (SNR), bit error rate (BER), computational complexity, and PAPR effect. Lastly, Section 4 reviews the main outcomes and suggests concluding remarks.

2. System Model

In this part, we present both the conventional and the suggested system model of Flip-Filtered-OFDM for IM/DD systems, covering the transmitter, receiver, and optical channel model components.

2.1. The Conventional Flip-F-OFDM

In the conventional Flip-F-OFDM transmitter, depicted in Figure 1, the input data flow is digitally modulated and converted into frequency samples. To guarantee the creation of real F-OFDM signals, the input frequency samples to the IFFT bulk are subjected to the HS restriction, which is defined as follows [15,29]:

$$X_{HS} = (X_0, X_1, X_2, \dots, X_{N-1}, X_{N-1}^*, \dots, X_2^*, X_1^*) \quad (1)$$

The real signal, denoted as X_{HS} , is obtained from complex frequency samples after applying HS. The HS constraint confirms that the conjugate of X is appended to the original frequency samples, while setting $X_0 = X_N = 0$ prevents the presence of remaining complex components in the time domain signal [37]. The complex signals after quadrature amplitude modulation (QAM), represented by X_1 to $X_{(N-1)}$, contribute to generating the final real signal.

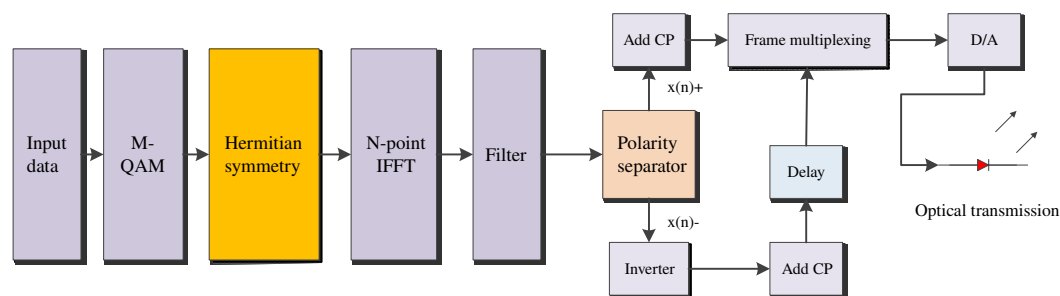


Figure 1. Conventional Flip–F–OFDM transmitter.

As a result, the signal obtained adheres to the requirement of being a real signal. It is to be distinguished that the second half of the input carriers carry the complex conjugate value of the first half. This symmetrical arrangement ensures the generation of a real signal at the IFFT output. However, it is fundamental to note that a sole half of the carriers carry useful data for obtaining the noncomplex signal, resulting in reduced spectral efficiency [38]. To obtain the discrete signal $x(n)$, the symbol signals $X(\kappa)$ are applied to the IFFT and are mathematically extracted as follows:

$$x(n) = \frac{1}{N} \sum_{\kappa=0}^{N-1} X(\kappa) \exp \frac{2\pi j \kappa n}{N}, \quad n = 0, 1, \dots, N-1 \quad (2)$$

Here, N signifies the FFT/IFFT length, $X(\kappa)$ is the subcarrier symbols, and $j = \sqrt{-1}$. However, due to the imposition of the Hermitian symmetry on the IFFT input signals, the IFFT size turns into $2N$ points, and hence, (2) can be re-arranged as follows:

$$x(n) = \frac{1}{2N} \sum_{\kappa=0}^{2N-1} X_{Hs} \exp \frac{j\pi \kappa n}{N}, \quad n = 0, 1, \dots, 2N-1 \quad (3)$$

where X_{Hs} represents the real signal. Applying the HS constraint ensures that the time domain signal imaginary part is zero at the IFFT output. This guarantees that the F-OFDM signal is noncomplexly valued, though it is not yet unipolar [38]. To obtain the unipolar signal, a polarity separator is employed to decompose the signal as follows:

$$x(n) = {}^+x(n) + {}^-x(n) \quad (4)$$

where the positivity and negativity components are presented as:

$${}^+x(n) = \begin{cases} x(n), & \implies x(n) \geq 0 \\ 0, & \implies x(n) < 0 \end{cases} \quad (5)$$

$${}^-x(n) = \begin{cases} x(n), & \implies x(n) < 0 \\ 0, & \implies x(n) \geq 0 \end{cases} \quad (6)$$

In order to change the dipolar signal, $x(n)$, into a nondipolar signal, the nonnegative portion is conveyed in the first subframe of F-OFDM, while the inverted nonpositive portion is conveyed in the following subframe of F-OFDM, as shown below:

$$x_{2N}(n) = \begin{cases} {}^+x(n), & n = 0, 1, \dots, N-1 \\ -{}^-x(n-N), & n = N, \dots, 2N-1 \end{cases} \quad (7)$$

The real nonunipolar F-OFDM signal, on the receiver side, $y(n)$, is reconstructed as shown below:

$$y(n) = {}^+y(n) - {}^-y(n) \quad (8)$$

In this context, ${}^+y(n)$ and ${}^-y(n)$ exemplify the two F-OFDM signal subframes received, respectively. The transmitted signal samples are then retrieved by applying a standard OFDM demodulation process.

2.2. The Proposed Flip-Filtered OFDM

2.2.1. Transmitter

To address the Hermitian symmetry (HS) constraints in real signal generation, we introduce a novel scheme, non-Hermitian symmetry (NHS), that reduces system hardware complexity by eliminating the necessity for frequency symbols to adhere to HS constraints. The schema of the suggested Flip-F-OFDM is illustrated in Figure 2. Indeed, the proposed Flip-F-OFDM technique eliminates the need for HS constraints by directly applying the IFFT to the frequency symbols. This method enables the creation of a nonreal F-OFDM signal at the output of IFFT. Because quadrature amplitude modulation (QAM) can lower noise interference and enhance system performance, thereby enabling high-speed data transmission, it is used in this research [15,37]. The nonreal signal from the IFFT output is rehabilitated into a noncomplex signal via a signal processing method known as the juxtaposition process.

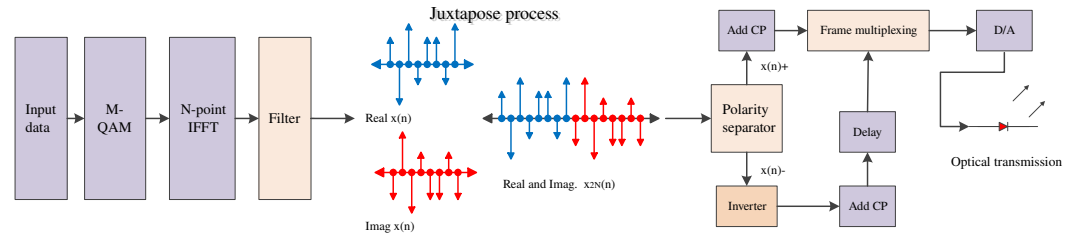


Figure 2. The transmitter of the suggested Flip-Filter-OFDM.

The input data stream of the transmitter is depicted in Figure 2, which is converted into QAM symbols and then directly fed into the N -point IFFT. Since there is no HS constraint applied, the output stream of the IFFT will be complex and is stated as follows [37]:

$$x(n) = \sum_{\kappa=0}^{N-1} [X_{\Re}(\kappa) + jX_{\Im}(\kappa)] \exp \frac{2\pi j \kappa n}{N} \quad (9)$$

Here, $X_{\Re}(\kappa)$ and $X_{\Im}(\kappa)$ are denoted as the two components of $X(\kappa)$. Then, the output IFFT signal is processed using a window function and a carefully constructed sub-band filter to create the F-OFDM signal [38,39]. To produce a double N -index noncomplex F-OFDM signal, $x_{2N}(n)$, the portions of the nonreal F-OFDM signal, real and imaginary as indicated in Figure 3, are combined in the time domain, resulting in the below juxtaposition:

$$x_{2N}(n) = \begin{cases} x_{\Re}(n), & \Rightarrow n = 0, 1, \dots, N-1 \\ x_{\Im}(n-N), & \Rightarrow n = N, \dots, 2N-1 \end{cases} \quad (10)$$

The subsequent $2N$ -point bipolar signal is then forwarded to a polarity separator. This separator treats the bipolar signal as a combination of a positive value, $^+x_{2N}(n)$, and a negative value, $^-x_{2N}(n)$, which is stated as follows:

$$^+x_{2N}(n) = \begin{cases} x_{2N}(n), & \Rightarrow x(n) \geq 0 \\ 0, & \Rightarrow x(n) < 0 \end{cases} \quad (11)$$

$$^-x_{2N}(n) = \begin{cases} x_{2N}(n), & \Rightarrow x(n) < 0 \\ 0, & \Rightarrow x(n) \geq 0 \end{cases} \quad (12)$$

The generated unipolar signal is then represented as:

$$x_{4N}(n) = \begin{cases} ^+x_{2N}(n), & n = 0, 1, \dots, 2N-1 \\ ^-x_{2N}(n-N), & n = 2N, \dots, 4N-1 \end{cases} \quad (13)$$

For the Flip-Filtered OFDM, the first part of the signal, $^+x_{2N}(n)$, is sent out in the first subframe, whilst the subsequent subframe is dedicated to the transmission of the flipped signal, $^-x_{2N}(n)$, as depicted in Figure 4. Given that the communication takes place over a dispersing optical channel, cyclic prefixes are inserted for each Filtered OFDM subframe, delaying the second subframe by $(N + CP)$, where N and CP represent the IFFT size and cyclic prefix, respectively [29]. The generated unipolar signal facilitates intensity modulation and is suitable for transmitting through the optical channel.

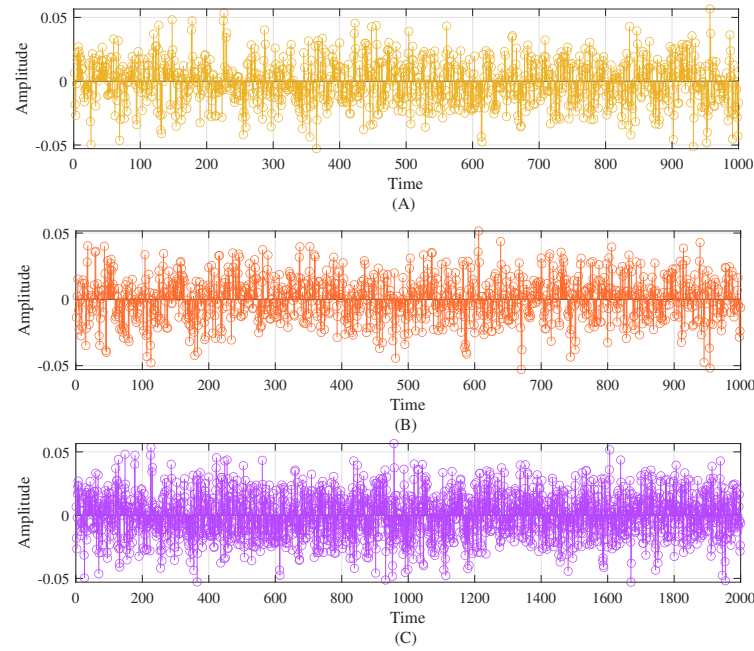


Figure 3. (A) Real signal, (B) imaginary signal, and (C) juxtaposed signal.

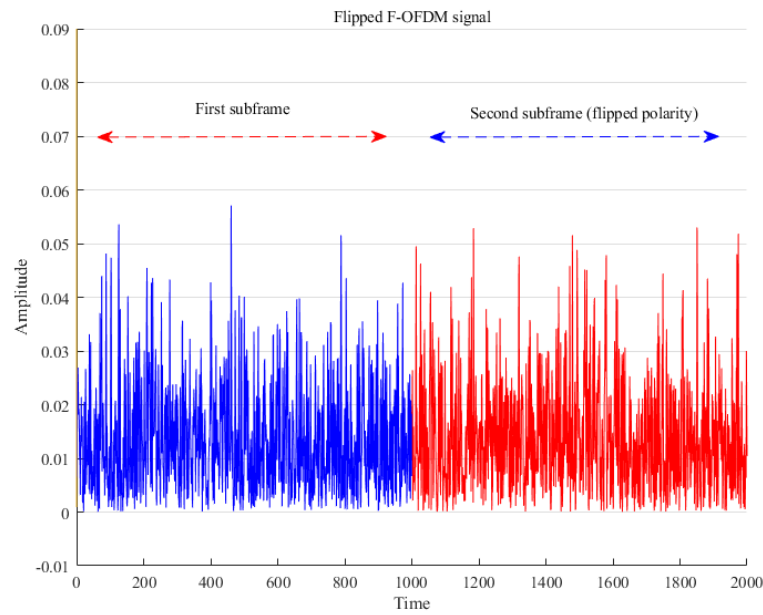


Figure 4. Unipolar (flipped) Filtered-OFDM signal.

2.2.2. Receiver

The Flip-F-OFDM receiver is exemplified in Figure 5. In this receiver, we assume a white Gaussian noise with a plane channel and idealistic equalization, resulting in the received signal $y(t)$. Subsequently, the photodetector changes the acquired signal into an electrical signal. Next, the received signal undergoes analog-to-digital conversion, and the cyclic prefix allied with each F-OFDM subframe is eliminated. To acquire the bipolar signal, $y_{2N}(n)$, the produced signal, $y_{4N}(n)$, is reconstructed according to the following expression:

$$y_{2N}(n) = {}^+ y_{4N}(n) - {}^- y_{4N}(2N + n) \quad (14)$$

Consequently, the noncomplex signal, $y_{2N}(n)$, is reconstructed to the complex signal, as expressed below:

$$\begin{cases} y_{\Re}(n) = y_{2N}(n), & \Rightarrow n = 0, 1, \dots, N-1 \\ y_{\Im}(n) = y_{2N}(N+n), & \Rightarrow n = 0, 1, \dots, N-1 \end{cases} \quad (15)$$

where $y_{\Re}(n)$ and $y_{\Im}(n)$ represent two portions of the complex signal, $y(n)$, correspondingly. Afterward, the generated complex signal is passed through the subband filter to the demodulator, following the standard procedure of complex F-OFDM systems. The retrieved frequency signals $Y(\kappa)$ are then expressed as follows:

$$Y(\kappa) = \sum_{n=0}^{N-1} y_n \exp^{-\frac{2\pi j \kappa n}{N}} \quad (16)$$

$$Y(\kappa) = \sum_{n=0}^{N-1} [y_{\Re}(\kappa) + jy_{\Im}(\kappa)] \exp^{-\frac{2\pi j \kappa n}{N}} = Y_{\Re}(\kappa) + jY_{\Im}(\kappa) \quad (17)$$

Here, $y_{\Re}(\kappa)$ and $y_{\Im}(\kappa)$ signify the values of the two parts of the received frequency symbols.

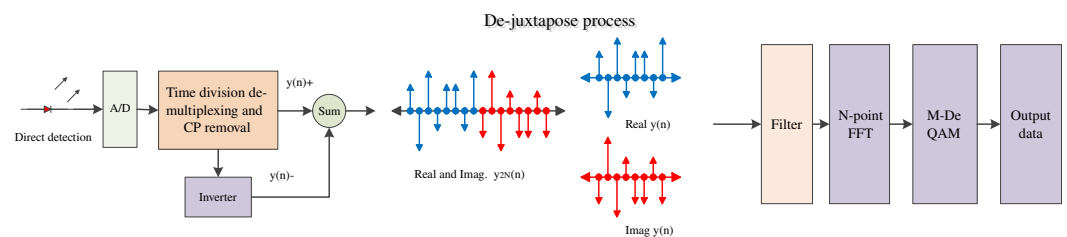


Figure 5. The receiver of the suggested Flip-Filter-OFDM.

2.2.3. Optical Channel Model

In the optical wireless communication systems of Figure 6, the two predominant sources of noise are photons and thermals in the receiving circuitry [40]. Photon noise is created from the discrete nature of photon arrivals, primarily caused by background light sources. Although optical filtering can minimize the received background light, shot noise is still generated even in a well-designed photodetector [41]. Since background light power is typically more substantial than the sent signal, the impact of the conveyed optical signal on this shot noise is insignificant.

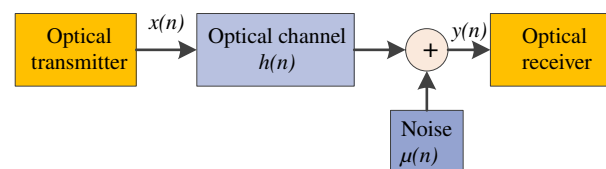


Figure 6. Optical communication model.

Consequently, it is typically assumed that shot noise is autonomous of the signal and can be quantified as white Gaussian noise. Similarly, in the absence of background light, the thermal (physical noise) is the dominant noise from the receiver preamplifier, which would also be signal-independent and follow a Gaussian distribution. In both instances, the whole noise source can be effectively constituted as Gaussian noise and signal-independent [40,41].

In this research, the transmitted signal, $x(n)$, passes out of an optical impulse response channel represented by $h(n)$, as shown in Figure 6. Then, an AWGN channel is employed,

and the received signal, $y(n)$, can be given as follows, taking into account the addition of noise with a variance of σ_μ^2 and mean of zero [37]:

$$y(n) = x(n) * h(n) + \mu(n) \quad (18)$$

where $*$ is the convolution, $x(n)$ denotes the signal transmitted, and $\mu(n)$ signifies the noise component. Then, the frequency symbol $Y(\kappa)$ is obtained by convolving the frequency response $H(\kappa)$ with the transmitted symbols $X(\kappa)$ and adding the noise term $M(\kappa)$, which in turn is extracted by:

$$Y(\kappa) = X(\kappa)H(\kappa) + M(\kappa) \quad (19)$$

3. Result Analysis and Discussion

In this part, the suggested system will be assessed, scrutinized, and juxtaposed against conventional Flip-F-OFDM and established methods such as HSF Flip-OFDM [42], ACO-OFDM [19,37,43], JM-DCO-F-OFDM [38], and DCO-OFDM [14,19,43] considering bandwidth efficiency, signal-to-noise ratio (SNR), bit error rate (BER), peak-average power ratio (PAPR), and computational complexity. This evaluation will be conducted through simulation verification and mathematical analysis.

3.1. Spectral Efficiency and Data Rate Analysis

Filter-OFDM offers substantial benefits in enhancing the system's spectral efficiency to face the demands of high transmission rates. Based on [38,39], the filter design used in F-OFDM effectively minimizes out-of-band emissions, enabling a very small gap between sub-bands and thus promoting efficient spectrum utilization. In the proposed Flip-F-OFDM technique, the absence of Hermitian symmetry (HS) allows all carriers (N) to be used for data transmission, resulting in an efficient data rate [12]. Conversely, in conventional Flip-F-OFDM, the imposition of HS reduces the data transmission rate to $(N/2)$, as only half of the available carriers are utilized for data transfer [19].

Otherwise, in DCO-OFDM, the data transmission is reduced to $(N/2)$ due to the IFFT Hermitian symmetry. In ACO-OFDM, the data transmission is reduced to $(N/4)$ because it employs HS and only uses odd subcarriers. Furthermore, although Flip-OFDM employs even and odd subcarriers, it requires two subframes to recreate the dipolar signal and decode the data. As a result, Flip-OFDM consumes double the number of samples as ACO-OFDM to convey double the data symbols. Thus, the transmission rate for both ACO-OFDM and Flip-OFDM is congruent [40]. Unlike Flip-OFDM and ACO-OFDM, HSF Flip-OFDM preserves a similar data rate as the proposed method since it does not require the imposition of HS [42,44]. Table 1 summarizes the transmitted data rate, transmitted data rate ratio, and spectral efficiency for the previously mentioned techniques.

Table 1. The transmitted data rate, transmitted data rate ratio, and spectral efficiency for the proposed Flip-F-OFDM, Conventional Flip-F-OFDM, HSF Flip-OFDM, Flip-OFDM, ACO-OFDM, and DCO-OFDM.

Technique	Transmitted Rate	Transmitted Rate Ratio	Spectral Efficiency
Proposed Flip F-OFDM	N	100%	$\frac{\log_2 M}{1+(2/N)}$
Conventional Flip F-OFDM	N/2	50%	$\frac{(\log_2 M)/2}{1+(2/N)}$
HSF Flip-OFDM [44]	N	100%	$\frac{\log_2 M}{1+(2/N)}$
Flip-OFDM [19,43]	N/4	25%	$\frac{(\log_2 M)/2}{1+(2/N)}$
ACO-OFDM [19,37,43]	N/4	25%	$\frac{(\log_2 M)/2}{1+(2/N)}$
DCO-OFDM [14,19,43]	N/2	50%	$\frac{\log_2 M}{1+(2/N)}$

3.2. Signal-to-Noise Ratio (SNR)

The SNR is a metric that quantifies the ratio of signal power to noise power. In Flip-Filtered OFDM, the unipolar signal is produced by transmitting the positive part in the first subframe of the F-OFDM signal, and the reversed negative component is in the second subframe. Let us assume $^+\mu(n)$ and $^-\mu(n)$ are the noise elements of the channel inserted into the two F-OFDM subframes, respectively. Subsequently, the two F-OFDM-receiving subframes are expressed as:

$$^+y(n) = ^+x(n) + ^+\mu(n) \quad (20)$$

$$^-y(n) = -^+x(n) + ^-\mu(n) \quad (21)$$

Then, the transmitted signal power $\sigma_{x,Flip}^2$ is defined as:

$$\sigma_{(x,Flip)}^2 = \frac{1}{2N} \sum_{n=0}^{N-1} |^+x(n)|^2 + \frac{1}{2N} \sum_{n=N}^{2N-1} |^-x(n)|^2 \quad (22)$$

In a similar procedure, the power noise at the receiver, $\sigma_{\mu,Flip}^2$, can be given by:

$$\sigma_{(\mu,Flip)}^2 = \frac{1}{2N} \sum_{n=0}^{N-1} |^+\mu(n)|^2 + \frac{1}{2N} \sum_{n=N}^{2N-1} |^-\mu(n)|^2 \quad (23)$$

In conventional Flip-F-OFDM, the dipolar signal is retrieved at the receiver by combining the two parts of the transmitted signal. Thus, the power of the bipolar signal for the N -points at the receiver is defined as:

$$P_{(\mathbb{S},Flip)} = \frac{1}{N} \sum_{n=0}^{N-1} |^+x(n)|^2 + \frac{1}{N} \sum_{n=0}^{N-1} |^-x(n)|^2 \quad (24)$$

$$P_{(\mathbb{S},Flip)} = 2\sigma_{(\mathbb{S},Flip)}^2 \quad (25)$$

Similarly, the noise power can be expressed as:

$$P_{(\mu,Flip)} = \frac{1}{N} \sum_{n=0}^{N-1} |^+\mu(n)|^2 + \frac{1}{N} \sum_{n=0}^{N-1} |^-\mu(n)|^2 \quad (26)$$

$$P_{(\mu,Flip)} = 2\sigma_{(\mu,Flip)}^2 \quad (27)$$

Accordingly, the SNR of the conventional Flip-F-OFDM at the receiver is presented as:

$$\text{SNR}_{\text{Flip-FOFDM}} = \frac{\sigma_{(\mathbb{S},Flip)}^2}{\sigma_{(\mu,Flip)}^2} = \frac{2\sigma_{\mathbb{S}}^2}{2\sigma_{\mu}^2} = \frac{\sigma_{\mathbb{S}}^2}{\sigma_{\mu}^2} \quad (28)$$

Combining real and imaginary components in the proposed Flip-F-OFDM further doubles both the signal power and noise power to restore the complex transmitted signal. As a result, the following expression can be used to show the SNR at the receiver of the proposed Flip-F-OFDM:

$$\text{SNR}_{\text{Proposed Flip-FOFDM}} = \frac{4\sigma_{\mathbb{S}}^2}{4\sigma_{\mu}^2} = \text{SNR}_{\text{Flip-FOFDM}} \quad (29)$$

3.3. Bit Error Rate Analysis

This subsection introduces a simulation examination of the proposed Flip-F-OFDM error rate compared to HSF Flip-OFDM, ACO-OFDM, DCO-OFDM, JM-DCO-F-OFDM,

and conventional Flip-F-OFDM. This study considers AWGN to be the communication channel as in [43], and the simulation was conducted in a MATLAB R2023a environment utilizing a 1024-point IFFT for various M-QAM schemes [38,39,43]. Filter design is vital in the F-OFDM technique to minimize out-of-band emissions. The spectrum can be more efficiently utilized by keeping the guard interval between sub-bands minimal [38,39]. Table 2 provides the detailed simulation parameters employed in this study. Figure 7 assesses the BER efficacy of the aimed Flip-F-OFDM, HSF Flip-OFDM, ACO-OFDM, JM-DCO-F-OFDM, and DCO-OFDM for 4QAM, 16QAM, 64QAM, and 256QAM, respectively.

The imitation results indicate that the suggested technique surpasses the previously mentioned techniques in all utilized QAM orders. This improvement is attributed to the employed filter's windowing technique, which ensures satisfactory temporal localization within the truncated filter's response. Thus, the F-OFDM signal maintains an acceptable inter-symbol interference (ISI) limit. Similarly, regarding JM-DCO-F-OFDM, the proposed scheme provides a better result because no DC bias effect is inserted into the signal.

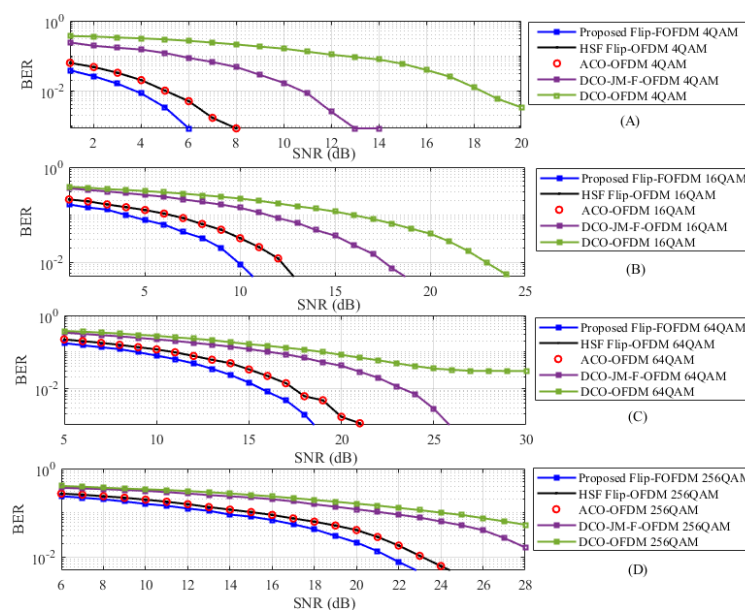


Figure 7. Evaluation of BER among the proposed Flip-F-OFDM, HSF-Flip-OFDM, ACO-OFDM, JM-DCO-F-OFDM, and DCO-OFDM: (A) 4QAM, (B) 16QAM, (C) 64QAM, and (D) 256QAM.

Table 2. Simulation parameters.

Parameter	Values
IFFT/FFT size	1024
No. of F-OFDM symbols	1000
No. of subcarriers	600
Filter type	Raised cosine
Modulation technique	QAM
Cyclic prefix	72
Constellation order	4, 16, 64, 256
Channel model	AWGN

Additionally, per the mathematical analysis in [37,40], ACO-OFDM exhibits the same bit error rate performance as HSF Flip-OFDM for the different QAM constellations. As illustrated in Figure 7A–D, the simulation outcomes fully agree with the analytical findings for the given QAM constellations. Figure 8, on the other hand, illustrates the bit error rate comparison between the suggested Flip F-OFDM and classical Flip F-OFDM. According to

the simulation results, the proposed and conventional F-OFDM techniques exhibit similar BER performance, consistent with the mathematical conclusions.

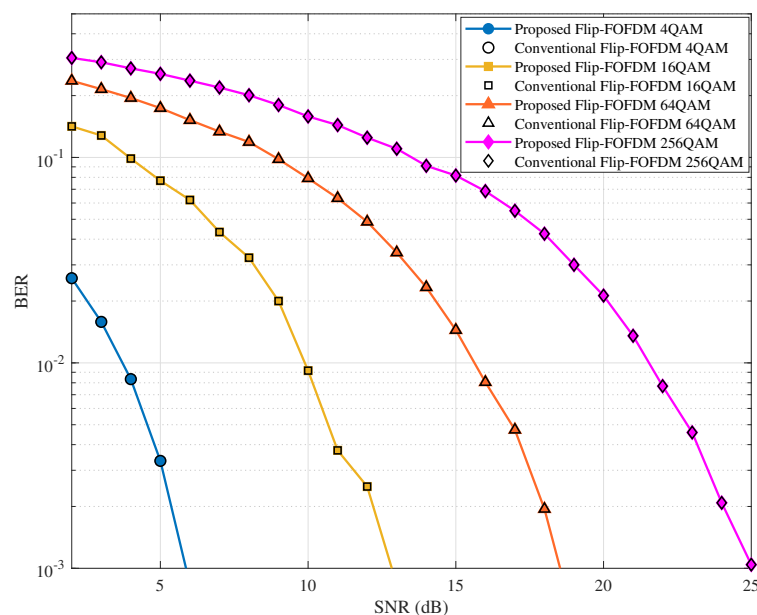


Figure 8. Evaluation of BER between the proposed Flip-F-OFDM and conventional Flip-F-OFDM for 4QAM, 16QAM, 64QAM, and 256QAM.

3.4. System Complexity

This metric provides an estimate of the computational resources required for the system. The impact of increasing FFT/IFFT sizes on computational complexity is detailed in [45]. According to [45], increasing the FFT/IFFT size leads to a rise in the subcarriers' number, which in turn expands the dynamical breadth of the OFDM signal. This expansion involves greater accuracy in arithmetic processes. The findings presented in [45] clearly illustrate the influence of various FFT/IFFT scopes on the error vector magnitude (EVM), with sizes ranging from 32 to 1024. The analysis indicates that reducing the FFT/IFFT sizes leads to a noticeable decrease in EVM for a fixed FFT/IFFT bit accuracy. Additionally, the study shows that both power usage and chip area utilization increase as the FFT/IFFT sizes grow.

In particular, this increase is significantly more noticeable for larger transform sizes, particularly when $N \geq 512$. According to the comparison outlined in [45], reducing the FFT size from $N = 1024$ to $N = 512$ led to a decrease of nearly 40% in chip area and 38% in power consumption. The foremost goal of the proposed Flip-F-OFDM procedure is to simplify the transmission and reception components. This is achieved by producing a noncomplex F-OFDM signal using N -length IFFT/FFT and creating a $4N$ -length unipolar F-OFDM signal. According to [45], this approach effectively reduces the computational complication of the system while retaining the desired functionality.

In the suggested Flip-F-OFDM technique, computational complexity is greater than that of HSF Flip-OFDM in the time domain because of the additional system complexity imposed by the filter at the transmitter side. The computational complication of F-OFDM is determined by increasing the OFDM computational complexity by the size of the filter. Consequently, the addition processes are similar for both F-OFDM and OFDM systems, while the multiplication processes increase according to the filter length for F-OFDM [46]. Compared to the standard Flip-F-OFDM method, the proposed Flip F-OFDM technique provides significant reductions in hardware complexity, achieving an improvement of $\approx 50\%$. Utilizing the radix-2 algorithm of FFT, the required number of additions and

multiplications for computing an $N/2$ -point real Flip F-OFDM signal using HS can be extracted as follows [47,48]:

$$\mathbb{O}_{\text{Add FFT}} = \frac{3N}{2} \log_2 \left(\frac{N}{2} \right) - \frac{3N}{2} + 4 \quad (30)$$

$$\mathbb{O}_{\text{Multip. FFT}} = \frac{N}{2} \log_2 \left(\frac{N}{2} \right) - \frac{3N}{2} + 4 + \frac{N}{2} (L - 1) \quad (31)$$

Here, N represents the FFT length, and L signifies filter size. Alternatively, the number of the operations (additions and multiplications) demanded for computing the FFT transformation in the proposed Flip F-OFDM receiver using $N/4$ -point IFFT is provided by [47,48]:

$$\mathbb{O}_{\text{Add FFT}} = \frac{3N}{4} \log_2 \left(\frac{N}{4} \right) - \frac{3N}{4} + 4 \quad (32)$$

$$\mathbb{O}_{\text{Multip. FFT}} = \frac{N}{4} \log_2 \left(\frac{N}{4} \right) - \frac{3N}{4} + 4 + \frac{N}{4} (L - 1) \quad (33)$$

Through mathematical analysis, Figures 9 and 10 offer a comparison of computational complexity in terms of multiplications and additions between the proposed Flip F-OFDM and traditional Flip F-OFDM methods for generating real signals. As noted in [45], these figures illustrate that with an increase in the size of the FFT/IFFT, both the occupied chip area and power consumption also increased.

Through the utilization of an optimized IFFT algorithm [46,47], the proposed Flip F-OFDM can achieve further reductions in the computational intricacy of the transmitter. This is a crucial aspect to consider in practical applications. Conversely, the conventional Flip F-OFDM method offers no complexity improvements on the transmitter side. The transmitter complexity in conventional Flip F-OFDM remains high due to the large number of subcarriers and the necessity to conduct a full IFFT on each subcarrier.

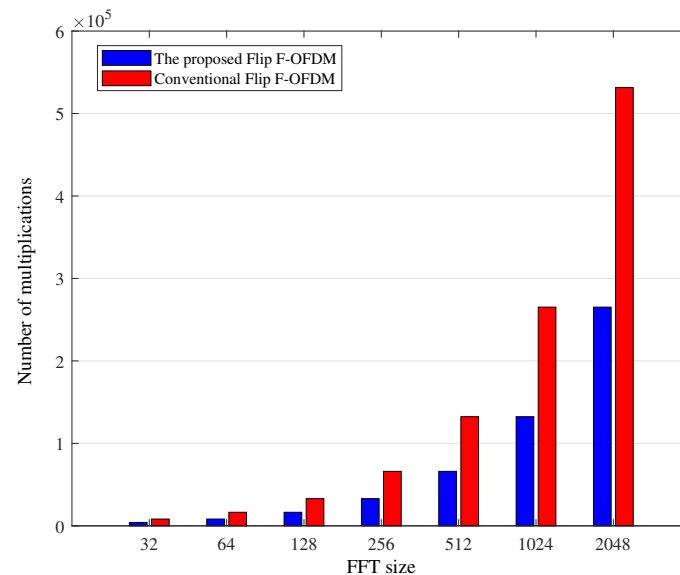


Figure 9. Flip–F–OFDM multiplication complexity.

Table 3 provides a comparative example of the computational complications of the proposed Flip-F-OFDM, conventional Flip-F-OFDM, HSF Flip-OFDM, Flip-OFDM, ACO-OFDM, JM-DCO-F-OFDM, and DCO-OFDM methods, based on a transmitted signal length of 128. According to Table 3, the proposed Flip-F-OFDM technique attains a significant reduction of $\approx 66.5\%$ in multiplication processes and around 57% in addition processes on

both the transmission and reception sides, compared to the classical Flip-F-OFDM and JM-DCO-F-OFDM methods. This illustrates the substantial computational complexity savings provided by the proposed Flip-F-OFDM approach. However, the proposed Flip-F-OFDM procedure has more multiplications than the HSF Flip-OFDM, Flip-OFDM, ACO-OFDM, and DCO-OFDM procedures. This increase in multiplications is attributed to the additional complexity introduced by the filter. Consequently, the computational complexity of the proposed Flip-F-OFDM relative to these techniques depends on the filter size used.

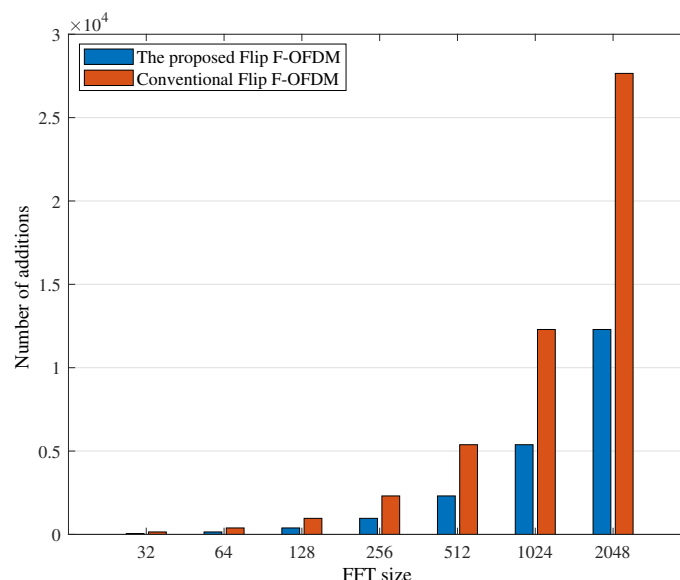


Figure 10. Flip-F-OFDM addition complexity.

Table 3. Computational complexity assessment among the proposed Flip-F-OFDM, conventional Flip-F-OFDM, JM-DCO-F-OFDM, HSF Flip-OFDM, Flip-OFDM, ACO-OFDM, and DCO-OFDM.

Schemes	Additions	Multiplications
Proposed Flip-F-OFDM	388	16,452
Conventional Flip-F-OFDM	964	32,964
JM-DCO-F-OFDM [38]	964	32,964
HSF Flip-OFDM [44]	964	196
Flip-OFDM [19,43]	2308	516
ACO-OFDM [19,37,43]	2308	516
DCO-OFDM [14,19,43]	2308	516

3.5. Peak-to-Average-Power Ratio (PAPR)

The PAPR is a crucial parameter in optical OFDM systems, significantly influencing the power efficiency of optical transmission. A large PAPR indicates a substantial disparity between the average and maximum power levels, which can lead to enlarged nonlinearity and the introduction of undesirable distortions. Conversely, a low PAPR indicates a more regular energy distribution [49,50]. In the Flip-F-OFDM technique, the dipolar signal is split into two unipolar values: the positive and negative halves, which include all negative values that are inverted and transmitted as positive. These components are conveyed in separated time slots, and the receiver subsequently recombines them to reconstruct the original F-OFDM signal. The maximum power for Flip-F-OFDM will occur in either the positive side $^+x(n)$ or the flipped negative side $^-x(n)$. Thus, the peak power for Flip-F-OFDM is specified by:

$$\max\left\{\left|x_{\text{Flip}}(n)\right|^2\right\} = \max\left\{\max\left|x^+(n)\right|^2, \max\left|x^-(n)\right|^2\right\} \quad (34)$$

But since both halves are derived from the same signal $x(n)$, then (34) can be rewritten as:

$$\max\left\{\left|x_{\text{Flip}}(n)\right|^2\right\} = \max\left\{\left|x(n)\right|^2\right\} \quad (35)$$

Furthermore, the overall average power in Flip-F-OFDM is the cumulative total of the average powers of the two components (the flipped negative and the positive values) that are transmitted independently. Each segment utilizes an entire time slot; thus, the total average power of the Flip-OFDM signal is articulated as follows:

$$E\left\{\left|x_{\text{Flip}}(n)\right|^2\right\} = E\left\{\left|x^+(n)\right|^2\right\} + E\left\{\left|x^-(n)\right|^2\right\} \quad (36)$$

Since $x^+(n)$ and $x^-(n)$ are derived from the same signal, the average power is split evenly between the two halves as follows:

$$E\left\{\left|x_{\text{Flip}}(n)\right|^2\right\} = 2 \times \frac{1}{2} E\left\{\left|x(n)\right|^2\right\} = E\left\{\left|x(n)\right|^2\right\} \quad (37)$$

Using (35) and (37) for peak power and average power, the PAPR for Flip-OFDM is given as:

$$\text{PAPR}_{\text{Flip-F-OFDM}} = \frac{\max\left\{\left|x_{\text{Flip}}(n)\right|^2\right\}}{E\left\{\left|x_{\text{Flip}}(n)\right|^2\right\}} \quad (38)$$

For the purported Flip-F-OFDM scheme, the peak power stays congruous with that of the conventional Flip-F-OFDM, as the elimination of Hermitian symmetry impacts the spectral structure without altering the peak amplitude of the time domain signal. In addition, since the real and imaginary parts contribute to the overall signal power, the average power is maintained at the same level as in conventional Flip-F-OFDM. Consequently, the PAPR of the proposed Flip-F-OFDM is articulated as follows:

$$\text{PAPR}_{\text{NHS Flip-F-OFDM}} = \frac{\max\left\{\left|x_{\text{Flip}}(n)\right|^2\right\}}{E\left\{\left|x_{\text{Flip}}(n)\right|^2\right\}} = \text{PAPR}_{\text{Flip-F-OFDM}} \quad (39)$$

The mathematical analysis indicates that the NHS-Flip-F-OFDM method effectively prevents any rise in the PAPR [43]. As a result, the NHS-F-OFDM approach facilitates the generation of unipolar F-OFDM signals that maintain the same PAPR levels as those produced by conventional unipolar Flip-F-OFDM techniques. Figure 11 illustrates the CCDF curves corresponding to the F-OFDM signals generated by the proposed NHS-Flip-F-OFDM, traditional Flip-F-OFDM, DCO-F-OFDM, ACO-F-OFDM, DCO-OFDM, and ACO-OFDM. These curves are derived from a 16-QAM constellation with a signal length of $N = 512$.

As shown in Figure 11, the CCDF diagrams for the proposed and standard F-OFDM are closely aligned, exhibiting a lower PAPR than ACO-F-OFDM and DCO-F-OFDM, respectively. However, compared with DCO-OFDM, due to the DC bias used, the proposed Flip-F-OFDM introduced a lower PAPR. In contrast, ACO-OFDM illustrates a lower PAPR than Filp-Filtered OFDM due to differences in signal structures and clipping mechanisms. ACO-OFDM inherently reduces the PAPR by clipping the signal at zero, ensuring only the

positive part of the signal is transmitted [37]. This clipping process removes prominent peaks, which significantly lowers the PAPR. F-OFDM filters each sub-band to confine the signal within a specified bandwidth. While this improves spectral efficiency and reduces out-of-band emissions, filtering can introduce additional peaks in the time domain signal due to the convolution process.

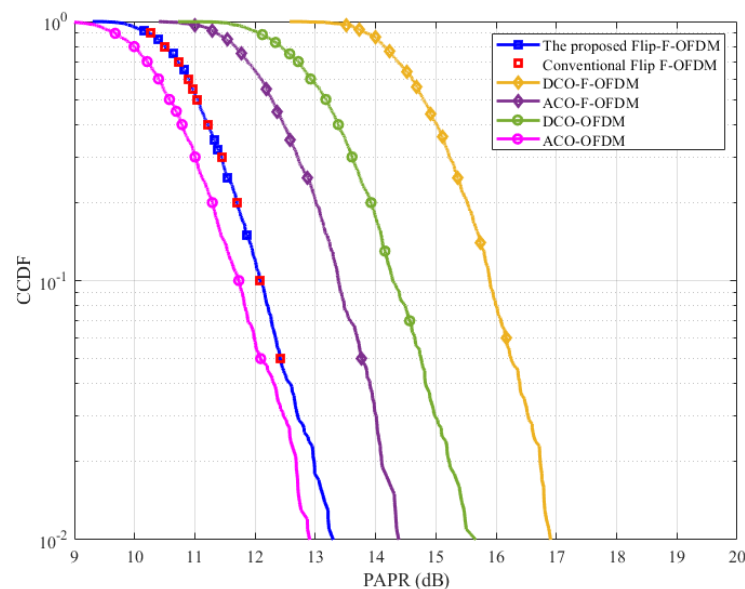


Figure 11. The CCDF comparison of the proposed NHS-Flip-F-OFDM, conventional Flip-F-OFDM, DCO-F-OFDM, ACO-F-OFDM, DCO-OFDM, and ACO-OFDM using a signal length $N = 512$.

4. Conclusions

The Flip-F-OFDM approach in this study introduces a new way to obtain a real and nondipolar signal applicable to IM/DD optical communication systems. Unlike traditional methods that require Hermitian symmetry (HS), the proposed method removes the dependence on HS yet maintains full compatibility with IM/DD systems. By laying the two parts of the complex F-OFDM signal (real and imaginary) together, the resulting real signal is then used to produce a non-bipolar signal. Moreover, this method only requires $N/4$ -size IFFT to obtain N -size FFT, thus simplifying the system design and reducing the computational complexity. As a result, this complexity reduction leads to lower power consumption and a smaller occupied chip area for the overall system.

In addition, the proposed NHS-Flip-F-OFDM approach enhances spectral efficiency, the efficiency of spectral control, and bit error rate performance compared to the HSF Flip-OFDM, JM-DCO-F-OFDM, ACO-OFDM, and DCO-OFDM methods. We also noted that although the proposed approach provides better efficiency in terms of hardware complexity, the proposed solution does not compromise on error rate performance. Therefore, the BER of the NHS-Flip-F-OFDM method is compatible with that of the conventional Flip-F-OFDM approach, in spite of using QAM constellations of different forms. On the other hand, the suggested technique prevents any increase in PAPR compared with the conventional Flip-F-OFDM, which provides an improvement compared with ACO-F-OFDM and DCO-F-OFDM. Further studies can investigate how techniques like pre-distortion, peak cancellation, or tone reservation in F-OFDM systems mitigate PAPR effects.

Author Contributions: Conceptualization, H.S.R.H. and M.I.; methodology, H.S.R.H.; software, H.S.R.H.; validation, H.S.R.H. and M.I.; formal analysis, H.S.R.H.; investigation, H.S.R.H. and M.I.; resources, H.S.R.H. and M.I.; data curation, H.S.R.H.; writing—original draft preparation,

H.S.R.H.; writing—review and editing, H.S.R.H.; visualization, H.S.R.H.; supervision, M.I.; project administration, M.I.; funding acquisition, H.S.R.H. and M.I. All authors have read and agreed to the published version of the manuscript.

Funding: This research received no external funding.

Institutional Review Board Statement: Not applicable.

Informed Consent Statement: Not applicable.

Data Availability Statement: The data underlying the results presented in this paper are not publicly available but may be obtained from the authors upon reasonable request.

Conflicts of Interest: The authors declare no conflicts of interest.

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