

Integrating FTN-GFDMA With A-RoF: A Proof-of-Concept for eRAC Applications

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Abstract—This paper investigates the integration of Faster-Than-Nyquist Generalized Frequency Division Multiple Access (FTN-GFDMA) into an Analog Radio-over-Fiber (A-RoF) system for enhanced Radio Access Connectivity (eRAC). A baseband-equivalent model accounting for Mach-Zehnder Modulator (MZM) nonlinearities is developed, and a machine-learning-based digital predistortion (DPD) technique is applied to mitigate electro-optical distortions. Results show that the proposed DPD significantly improves bit error rate (BER) performance, enabling the FTN-GFDMA A-RoF system to operate close to the theoretical additive white Gaussian noise (AWGN) limit.

Keywords—6G, A-RoF, eRAC, FTN-GFDMA, multiple access.

I. INTRODUCTION

The evolution of sixth-generation of mobile networks (6G) requires communication systems capable of providing affordable and ubiquitous connectivity, including remote and rural regions that remain underserved by current infrastructures [1], [2]. Despite advances in mobile communications, a significant portion of the population still lacks reliable network access due to limited coverage and high deployment costs [3]. These challenges become more critical in enhanced remote area communications (eRAC) scenarios, where spectrum scarcity, long transmission distances, and infrastructure constraints demand highly spectrally efficient and cost-effective solutions [4].

In this context, analog radio over fiber (A-RoF) and TV white space (TVWS) emerge as promising technologies for enabling long-range wireless coverage with reduced deployment complexity. However, limited bandwidth availability and nonlinear impairments introduced by optical fronthaul components remain major challenges [5], [6].

Faster than Nyquist - generalized frequency division multiple access (FTN-GFDMA) is a spectrally efficient waveform that combines FTN signaling [7], [8] with generalized frequency division multiplexing (GFDMA) [9], enabling controlled time-frequency overlap to improve capacity and reduce out-of-band emission (OOBE). These features suit A-RoF architectures, though their use in optical-wireless systems remains largely unexplored.

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This paper presents a proof-of-concept implementation of FTN-GFDMA in an A-RoF system designed for eRAC applications. The work examines the benefits and implementation challenges of integrating FTN-GFDMA waveforms with optical fronthaul architectures. A baseband-equivalent model incorporating the nonlinear behavior of the Mach-Zehnder modulator (MZM) is adopted to evaluate the impact of electro-optical distortions on the FTN-GFDMA waveform [10]. To mitigate these impairments, a multi layer perceptron (MLP)-based digital pre-distortion (DPD) learns the inverse transmitter nonlinearity [11], [12], reducing distortion and bringing the FTN-GFDMA A-RoF bit error rate (BER) close to the theoretical additive white Gaussian noise (AWGN) bound.

II. SYSTEM MODEL FOR INTEGRATING FTN-GFDMA INTO A-RoF FOR eRAC APPLICATIONS

In FTN-GFDMA, time-frequency resources are allocated to users by defining elements of the data symbol matrix and setting the remaining elements to zero [13].

Each resource element represents the transmission of a quadrature amplitude modulation (QAM) symbol $s_{k,m} \in \mathbb{C}^{1 \times 1}$ using a time and frequency-shifted version of the prototype filter $\mathbf{g}_{k,m} \in \mathbb{C}^{N \times 1}$, where $s_{k,m}$ is the data symbol at the k th subcarrier and m th subsymbol, and $\mathbf{g}_{k,m}$ is the prototype filter shifted to the same positions. Each resource element is allocated to a single user and is not shared with others.

The prototype filter $\mathbf{g}$ is represented by $\mathcal{P}$ periods with $\mathcal{S}$ samples per period, where $N = \mathcal{P}\mathcal{S}$ is the total number of filter samples. Thus, the v th sample of the prototype filter shifted to the k th subcarrier and m th subsymbol is given by

$$g_{k,m}[v] = g \left[\left(v - m\tau\mathcal{S} \right)_{\mathcal{N}} \right] \exp \left(j2\pi k \frac{\phi}{\mathcal{S}} v \right), \quad (1)$$

where $k \in \{0, \dots, K-1\}$ is the subcarrier index, with K denoting the subcarriers number; $m \in \{0, \dots, M-1\}$ is the subsymbol index, with M representing the subsymbol number; $v \in \{0, \dots, N-1\}$ is the time index; and τ and ϕ represent the time and frequency overlap factors, respectively, given by

$$\tau = \frac{\Delta_K}{\mathcal{S}} \quad \text{and} \quad \phi = \frac{\Delta_M}{\mathcal{P}}. \quad (2)$$

Here, Δ_K is the distance between the subsymbols in the time domain, and Δ_M is the distance between the subcarriers in the frequency domain. When $\Delta_K < \mathcal{S}$, subsymbols overlap in time, increasing spectral efficiency but causing intersymbol interference (ISI). Similarly, when $\Delta_M < \mathcal{P}$, subcarriers overlap in frequency, improving spectral efficiency while introducing intercarrier interference (ICI). Thus, the numbers of

subsymbols M and subcarriers K in the FTN-GFDM block are given by

$$M = \frac{\mathcal{PS}}{\Delta_K} = \frac{\mathcal{P}}{\tau} = \left\lfloor \frac{N}{\Delta_K} \right\rfloor, \quad K = \frac{\mathcal{PS}}{\Delta_M} = \frac{S}{\phi} = \left\lfloor \frac{N}{\Delta_M} \right\rfloor. \quad (3)$$

The FTN-GFDM transmission matrix $\mathbf{A} \in \mathbb{C}^{N \times N}$ consists of the $N = KM$ versions of the prototype filter, circularly shifted in time and frequency, and can be defined as

$$\mathbf{A} = \begin{bmatrix} \mathbf{g}_{0,0} & \mathbf{g}_{1,0} & \cdots & \mathbf{g}_{K-1,0} & \mathbf{g}_{0,1} & \cdots & \mathbf{g}_{K-1,M-1} \end{bmatrix}. \quad (4)$$

Thus, the $(k + mK)$ th column of $\mathbf{A}$ corresponds to $\mathbf{g}_{k,m}$.

Assuming a FTN-GFDMA scenario where a base station (BS) communicates with U users, the allocated resource elements to user u are represented by $\mathcal{K}_{\text{on}}^{(u)} \times \mathcal{M}_{\text{on}}^{(u)} \subset \mathcal{K}_{\text{on}} \times \mathcal{M}_{\text{on}}$, where $u \in \{1, \dots, U\}$, and $\mathcal{K}_{\text{on}}$ and $\mathcal{M}_{\text{on}}$ are the sets of available subcarriers and subsymbols, respectively. The u th user data symbol matrix $\mathbf{S} \in \mathbb{C}^{K \times M}$ can be represented by

$$\mathbf{S}_{k,m}^{(u)} = \begin{cases} s_{k,m} & \text{if } k \in \mathcal{K}_{\text{on}}^{(u)} \text{ and } m \in \mathcal{M}_{\text{on}}^{(u)} \\ 0 & \text{otherwise} \end{cases}. \quad (5)$$

In the downlink FTN-GFDMA scenario, the data of all users are combined into the data symbol matrix, i.e.,

$$\mathbf{S} = \sum_{u=1}^U \mathbf{S}^{(u)}. \quad (6)$$

The signal transmitted by the BS is given by

$$\mathbf{x} = \mathbf{A}\mathbf{s}, \quad (7)$$

where $\mathbf{s} = \text{vec}(\mathbf{S}) \in \mathbb{C}^{N \times 1}$ is a vector containing the QAM symbols of all users. Thus, the v th sample of the signal inserted in the MZM is given by

$$x[v] = \sqrt{\frac{N}{N}} \sum_{k=0}^{K-1} \sum_{m=0}^{M-1} s_{k,m} g[\langle v - m\tau S \rangle_N] \exp\left(j2\pi k \frac{\phi}{S} v\right). \quad (8)$$

Before transmission, a cyclic prefix (CP) is added to the signal, leading to a faster than Nyquist (FTN)-GFDM spectral efficiency given by

$$\eta = \frac{R_c M K \log_2(J)}{\tilde{N}}, \quad (9)$$

where J is the modulation order, R_c is the forward error correction (FEC) coding rate, and $\tilde{N} = N + N_{\text{CP}}$ and N_{CP} is the CP length.

In the access network considered in this work, a long-reach optical fronthaul is used to transport signals between central office (CO) and BS. Extending the fronthaul link requires increasing both radiofrequency (RF) and optical power to compensate for the additional attenuation. This increased power leads the system to operate within a nonlinear region, as the MZM exhibits nonlinear behavior. A low or controlled peak to average power ratio (PAPR) waveform is desirable [14].

An equivalent baseband model can be employed to represent the nonlinear behavior of the access network. The memoryless nonlinear response of the MZM can be modeled as [10]

$$b[v] = \sum_{\iota=0}^{I-1} \zeta_{\iota} |x[v]|^{\iota} x[v], \quad (10)$$

where ζ_{ι} denote the polynomial model coefficients, $x[v]$ represents the transmitted GFDM signal (or the FTN-GFDMA waveform in the multiuser case), and I denotes the nonlinear order of the model. For a block of v samples, (10) can be equivalently expressed in matrix form as

$$\mathbf{b} = \mathbf{X}\boldsymbol{\zeta}, \quad (11)$$

where $\mathbf{b} = [b[0], b[1], \dots, b[v-1]]^T$ is the output signal vector, $\boldsymbol{\zeta} = [\zeta_0, \zeta_1, \dots, \zeta_{I-1}]^T$ is the coefficient vector, and $\mathbf{X} \in \mathbb{C}^{v \times I}$ is the nonlinear regression matrix defined as

$$\mathbf{X} = \begin{bmatrix} x[0] & |x[0]|x[0] & \cdots & |x[0]|^{I-1}x[0] \\ x[1] & |x[1]|x[1] & \cdots & |x[1]|^{I-1}x[1] \\ \vdots & \vdots & \ddots & \vdots \\ x[v-1] & |x[v-1]|x[v-1] & \cdots & |x[v-1]|^{I-1}x[v-1] \end{bmatrix}. \quad (12)$$

Assuming that the photodetector (PD) does not introduce nonlinear distortions due to the low incident optical power, the signal at PD output can be described by

$$\bar{\mathbf{x}} = \mathbf{b} + \boldsymbol{\omega}, \quad (13)$$

where $\boldsymbol{\omega}$ is the noise inserted by PD.

Assuming perfect synchronization between the network and the BS clock, and the use of a sufficiently long CP, the received signal at the u th user terminal can be expressed as

$$\mathbf{y}^{(u)} = \mathbf{H}^{(u)} \bar{\mathbf{x}} + \mathbf{w}^{(u)}, \quad (14)$$

where $\mathbf{H}^{(u)} \in \mathbb{C}^{N \times N}$ represents the circulant matrix of the wireless channel impulse response between the BS and the u th user terminal, and $\mathbf{w}^{(u)} \in \mathbb{C}^{N \times 1}$ denotes the AWGN vector.

At the user terminal u , after the removal of the CP and under the assumption of perfect channel state information (CSI), the received signal is equalized as

$$\bar{\mathbf{y}}^{(u)} = \left(\mathbf{H}^{(u)}\right)^{-1} \mathbf{H}^{(u)} \bar{\mathbf{x}} + \left(\mathbf{H}^{(u)}\right)^{-1} \mathbf{w}^{(u)}. \quad (15)$$

The equalized signal is then processed by a matched filter (MF), whose output can be expressed as

$$\mathbf{r}^{(u)} = \mathbf{A}^H \bar{\mathbf{y}}^{(u)} = \mathbf{A}^H \left(\mathbf{H}^{(u)}\right)^{-1} \mathbf{H}^{(u)} \bar{\mathbf{x}} + \mathbf{A}^H \left(\mathbf{H}^{(u)}\right)^{-1} \mathbf{w}^{(u)}. \quad (16)$$

A nonlinear detector is used to detect the symbol sequence and cancel the inter-user interference (IUI). Then, the users perform resource demapping to extract their data and discard the rest. This process is performed at all user terminals.

A. Artificial Neural Network-Based Digital Pre-distortion

To mitigate nonlinear distortions introduced by the electro-optical modulation process in the A-RoF transmitter chain, a supervised machine-learning-based DPD technique was adopted. The baseband DPD scheme learns an inverse nonlinear mapping to linearize the MZM response and reduce distortion prior to transmission. Machine-learning-based DPD

schemes provide robust performance by exploiting their generalization capability to adapt to varying operating conditions while maintaining manageable computational complexity [15].

The proposed DPD employs a MLP trained to map distorted baseband samples to undistorted ones. The complex signal is represented by its real and imaginary components, and the output provides the predistorted I/Q samples. The network has three hidden layers with 288, 544, and 224 neurons using rectified linear unit (ReLU), linear, and scaled exponential linear unit (SELU) activations, followed by a two-neuron linear output layer. The weights are initialized with a zero-mean Gaussian distribution with standard deviation 0.05.

Training is supervised using experimental baseband data from our previous A-RoF transmission chain [12]. Post-MZM distorted samples are used as inputs and the original undistorted samples as targets. Both signals are power-normalized before the data are split into training and validation subsets. The network is optimized with Adam using a learning rate of 3.98×10^{-5} , and the loss is the mean squared error between predicted and target I/Q components. A batch size of 4096 samples is used during the 500-epoch training, with early stopping to avoid overfitting. After training, the artificial neural network (ANN)-based DPD is inserted into the transmitter chain, as shown in Fig. 1. The predistorted signal $\mathbf{x}'$ is obtained by applying the trained ANN to the input signal $\mathbf{x}$.

III. INTEGRATING FTN-GFDMA WITH A-ROF SYSTEMS

Integrating FTN-GFDMA with A-RoF systems offers several benefits, including improved spectral efficiency, waveform flexibility, and centralized processing. Nevertheless, it also introduces challenges such as increased interference, nonlinear distortions, and higher receiver complexity. The following subsections discuss the main benefits and challenges of this integration.

A. Advantages

The opportunistic use of TVWS channels is essential to extend connectivity to remote and rural areas. Since the available spectrum may be limited, maximizing channel utilization while minimizing adjacent-channel interference is crucial.

Orthogonal frequency division multiplexing (OFDM) generates high OOB due to abrupt transitions between time-domain blocks, which may interfere with adjacent communications. In contrast, FTN-GFDMA reduces OOB through circular filtering, facilitating coexistence between primary and secondary users. This characteristic is fundamental for the technical feasibility and regulatory approval of TVWS deployments, especially in fragmented spectrum scenarios.

A key advantage of FTN-GFDMA over orthogonal frequency division multiple access (OFDMA) is its flexibility in pulse shaping and parameter configuration, including the number of subcarriers, subsymbols, and overlap factors in time and frequency [16]. This flexibility enables more precise spectral control through the selection of the prototype filter, $\mathbf{g}$ [17]. Furthermore, unlike OFDMA, which relies on orthogonal subcarriers and rectangular filtering, FTN-GFDMA allows adjustable subcarrier spacing [18].

FTN-GFDMA further improves spectral efficiency by allowing temporal and spectral pulse overlap, increasing throughput and system capacity [17]. This flexibility enables resource allocation according to user demands and application requirements. Thus, the system can either support more users without degrading quality of service (QoS) or increase the data rate per user, depending on the network objective.

B. Challenges

In FTN-GFDMA, independently modulated subcarriers are combined to form the transmitted signal. When these subcarriers align in phase, high power peaks may occur, increasing fluctuations in the instantaneous signal power and, consequently, the PAPR. This effect becomes more pronounced as the number of subcarriers increases. Although prototype filters improve spectral behavior, they may also increase signal amplitude variations. Furthermore, violating the Nyquist criterion in FTN signaling introduces inherent ISI and ICI, which further contribute to higher PAPR.

Fig. 2 shows the complementary cumulative distribution function (CCDF) of PAPR for FTN-GFDMA and GFDMA systems using a root raised cosine (RRC) filter with roll-off factor α , $S = 5$, $P = 4$, and the parameters listed in Table I. In FTN-GFDMA, higher interference levels caused by overlapping subcarriers and subsymbols increase PAPR, while prototype filters with nonzero roll-off factors help mitigate this effect by reducing sidelobes in the time domain.

TABLE I: Parameters. BPSK, RRC filter, $P = 4$, and $S = 5$.

Parameters	FTN-GFDMA-I	FTN-GFDMA-II	GFDMA
Time spacing between subsymbols (Δ_K)	4	4	5
Frequency spacing between subcarriers (Δ_M)	4	3	4
Time overlapping factor (τ)	0.8	0.8	1
Frequency overlapping factor (ϕ)	1	0.75	1
Subcarriers per block (K)	5	6	5
Subsymbols per block (M)	5	5	4
Density of data symbols per signal sample (N/N)	1.25	1.5	1

The high PAPR of FTN-GFDMA exacerbates MZM nonlinearities, particularly under dynamic power allocation in A-RoF systems, leading to signal distortion and quality degradation. However, suitable mitigation techniques can reduce these effects while preserving the flexibility, low OOB, and spectral efficiency of FTN-GFDMA.

In addition to nonlinear distortions, integrating FTN-GFDMA into A-RoF systems increases the impact of inherent ICI and ISI, particularly in dispersive and time-varying wireless channels. High Doppler frequencies can significantly degrade BER performance, especially for long transmission blocks, where the assumption of channel invariance becomes inaccurate. Reliable CSI estimation is therefore essential, since estimation errors lead to suboptimal equalization and reduced system reliability.

Another challenge is the reduction in spectral efficiency caused by longer CP requirements. In long optical fronthaul

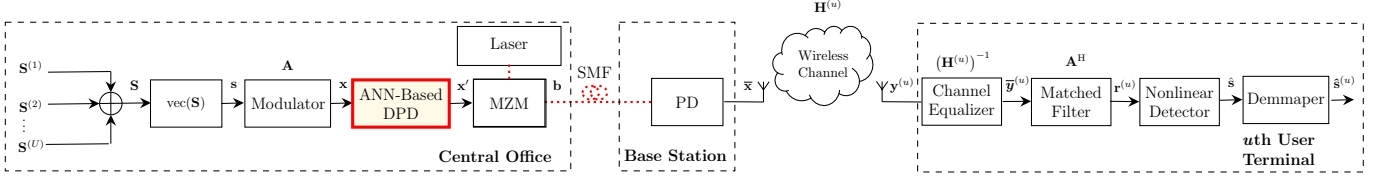


Fig. 1: Block diagram of the FTN-GFDMA downlink transceiver integrated with the A-RoF link, including the ANN-based DPD block at the transmitter/central office, for the u th user terminal. CP insertion and removal are omitted for clarity.

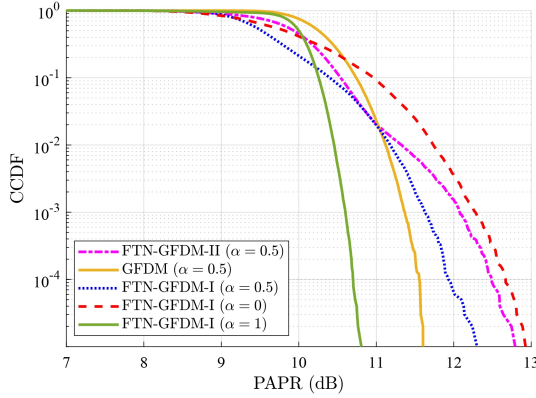


Fig. 2: PAPR of FTN-GFDM and GFDM systems, where the complementary cumulative distribution function was empirically constructed from 10^5 frames, each containing 40 FTN-GFDM blocks, using an oversampling factor of 4.

links, chromatic dispersion may become significant depending on the signal bandwidth. To avoid inter-block interference, the CP must exceed the combined impulse response of the optical and wireless channels, increasing transmission overhead and reducing the time available for payload transmission.

To address these impairments, linearization techniques capable of compensating MZM nonlinearities are required. PAPR reduction methods are also important for mitigating nonlinear distortion. To reduce infrastructure cost and complexity, these techniques should preferably be implemented at the CO. In this context, DPD and artificial intelligence (AI)-based solutions are promising approaches for reducing nonlinear effects and implementation complexity [11], [19].

Finally, receiver complexity remains a major challenge. Simultaneously mitigating ICI, ISI, nonlinear distortions, and channel impairments requires advanced signal processing algorithms with high computational cost. Efficient low-complexity detection and equalization techniques are therefore essential for the practical implementation of FTN-GFDMA systems in challenging communication environments.

IV. RESULTS AND DISCUSSIONS

This section evaluates the performance of FTN-GFDMA integrated into an A-RoF system. Chromatic dispersion and PAPR reduction were not considered, since the optical fronthaul is relatively short and the narrow TVWS bandwidth is unlikely to cause significant dispersion effects.

A. Baseline System Performance

Fig. 3 shows the uncoded BER performance of the faster than Nyquist - generalized frequency division multiplexing (FTN-GFDM) system for different E_b/N_0 values using the parameters in Table I. This analysis considers only the FTN-GFDMA signal and serves as a baseline for subsequent evaluations including fiber transmission effects. Compared to conventional GFDM, the spectral efficiency increases by 25% for FTN-GFDM-I and 50% for FTN-GFDM-II.

The adopted channel models include an AWGN channel with impulse response $h = 1$ and a time-variant frequency-selective (TVFS) channel given by $\mathbf{h} = [e^0 h_0 \ e^{-3} h_1 \ e^{-6} h_2 \ e^{-9} h_3]^T$, where $h_n \sim \mathcal{CN}(0, 1)$. The four-tap TVFS channel provides flat fading per subcarrier and frequency selectivity over the full signal bandwidth. Perfect CSI is assumed at the receiver, and nonlinear sphere decoder (SD) is used for symbol recovery. The simulated uncoded BER agrees with the theoretical curves¹, indicating that FTN-GFDM improves spectral efficiency and data rate with limited BER degradation. Similar behavior is expected for the joint detection adopted in the FTN-GFDMA downlink, where the entire received data sequence is detected before discarding the symbols allocated to other users. Alternatively, single-user detection can be employed as proposed in [13], where each user detects only the symbols allocated to its resources, reducing computational complexity and improving network security while proper resource allocation mitigates IUI without significant BER degradation.

¹Analytical BER expressions for binary phase shift keying (BPSK) modulation under AWGN and TVFS channels are available in [9].

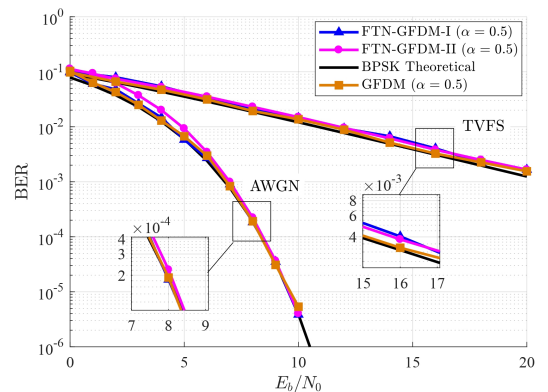


Fig. 3: BER performance of FTN-GFDM systems using BPSK and an RRC filter with roll-off factor $\alpha = 0.5$.

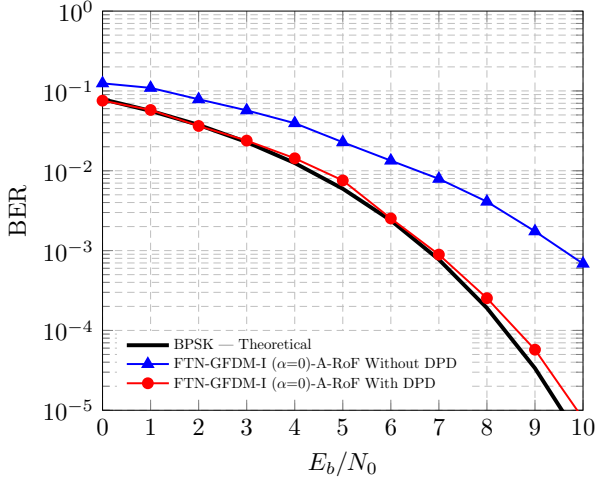


Fig. 4: BER performance of FTN-GFDM systems employing BPSK modulation and an RRC filter with roll-off factor $\alpha = 0$.

B. DPD Performance Evaluation

The DPD evaluation follows the block diagram in Fig. 1. Assuming an AWGN channel, the channel matrix $\mathbf{H}$ was set to the identity matrix. Fig. 4 presents the BER performance of the FTN-GFDM-I signal transmitted over the A-RoF link with and without DPD, alongside the theoretical BPSK bound.

Without DPD, the system exhibits significant performance degradation due to nonlinear distortions introduced by the MZM. Applying DPD substantially improves performance, with the BER approaching the theoretical BPSK curve, especially at low and moderate E_b/N_0 values. These results indicate that the ANN-based DPD effectively compensates for the dominant nonlinear impairments by learning the inverse characteristics of the nonlinear transmission chain. Residual gaps at high E_b/N_0 values are mainly associated with unmodeled higher-order and memory nonlinearities.

V. CONCLUSIONS

This paper presented an initial proof-of-concept of the integration of FTN-GFDMA into A-RoF systems for eRAC scenarios in future 6G networks. The study discussed the main benefits and challenges of combining the spectral efficiency and flexibility of FTN-GFDMA with the centralized and cost-effective characteristics of A-RoF architectures, targeting coverage expansion in rural and underserved areas.

The analysis showed that the intentional time-frequency overlap of FTN-GFDMA improves spectral efficiency but also increases ISI, ICI, PAPR, and nonlinear distortions in the A-RoF fronthaul. To mitigate these impairments, a machine-learning-based DPD scheme was applied at the central office to linearize the transmitter chain without increasing remote-unit complexity. The results demonstrated that the proposed DPD significantly reduces nonlinear distortions, enabling the FTN-GFDMA A-RoF system to achieve BER performance close to the theoretical AWGN bound.

Overall, the results demonstrate the potential of FTN-GFDMA-based A-RoF systems to provide spectrally efficient wireless connectivity in remote and underserved

regions. Although challenges such as PAPR, interference management, chromatic dispersion, and receiver complexity remain, the proposed framework provides a basis for future investigations. Future work will investigate different time-frequency overlap factors, higher-order modulation schemes, nonlinear operating conditions, and alternative detector and DPD configurations to provide a deeper assessment of the proposed framework beyond the initial proof-of-concept presented in this work.

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