High Capacity 64-Quadrature Amplitude Modulation Based Optical Coherent Transceiver for 60 GHz Radio over Fiber System

Devendra Chack · Sunil Narayan Thool

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Abstract
The high-capacity data transmission over an optical communication network is the bottleneck to meeting future communication needs due to the substantial growth in the data transmission requirements, which increased yearly at an average rate of more than 25%. The new generation 5G, beyond 5G and 6G proven to be the most promising technology to meet such high-capacity data transmission requirements. Also, 60 GHz (mm-wave) would be the most promising candidate for wireless high-capacity data transmission between multiple radio terminals. The optical domain techniques, such as coherent detection and optical heterodyning for generating 60 GHz (mm-wave) signals, are adequate, mature, and cost-effective solutions for high-capacity data transmission. We propose a 60 GHz radio over fibre (RoF) system design and its performance analysis which established a noteworthy 168 Gbits/s high-capacity data transmission using 64-Quadrature Amplitude Modulation, optical heterodyning, coherent detection, and advanced digital signal processing (ADSP) chain. The optimized version of the optical heterodyne technique generates a 60 GHz radio frequency signal for connecting various radio nodes over the wireless channel. The decisive and crucial part of the designed system is the ADSP chain which is effectively used to mitigate the impairments generated, and it also improves the signal spectral efficiency and results in long-distance data transmission. The proposed RoF system design and its performance analysis prove that the 28 Gbaud high-capacity data transmission is achieved for more than 172 km distance of Standard Single Mode Fiber.

Keywords Optical communication · Radio over fiber · Optical heterodyne coherent detection · Advanced DSP · 5G · 6G
1 Introduction

We are in the era of 5G, forthcoming beyond 5G and 6G, Internet of things (IoT), Distributed Data Centre etc. and these technologies are progressively used to provide the high-capacity data transmission and facilitate the end user’s data communication needs such as high definition (HD) video transmission which would be highly useful to enhance the interconnectivity across the world [1]. In order to address these requirements and to accommodate the substantial growth of massive bandwidth utilization generated by new networking standards, it is essential to upgrade the existing data transmission scheme presently used in optical communication and wireless data transmission. IEEE 802.11ad is a 60 GHz wireless networking standard that provides multigigabit data transmission [2].

60 GHz (mm-wave) can be used to transmit the high-capacity data, such as uncompressed videos, over wireless channels for shorter-distance communication. Therefore, the 60 GHz wireless channel is now becoming a more popular and an alternate solution for the conventional Wireless Fidelity (Wi-Fi) system and the same to be used in 5G and beyond technologies [3, 4]. As indicated above, 60 GHz is an imminent and most appropriate option for rapidly increasing huge bandwidth utilization requirements, especially in broadband wireless communication. 60 GHz has its inherent limitation of transmission loss and results in the degradation of system performance. However, recent studies use a 60 GHz wireless channel, which is the most suitable solution for indoor communication. Furthermore, 60 GHz is the unlicensed spectrum, and it is effectively used for high-capacity data transmission [5–9]. The recent studies and development in the area of optical fibre-based RoF cater for limited bandwidth utilization, and mitigation techniques of wireless channel impairments and electromagnetic interference are becoming the most critical bottleneck of the system design [10]. Any RoF system mainly consists of a Central Station (CS), Base Station (BS) and Radio User Terminals (RUT). Preliminarily, baseband signal processing, such as signal conditioning and modulation and conversion of modulated signal into the optical signal, is carried out in the CS, followed by optical signal transmission over a longer distance standard single mode fibre (SSMF). The BS intercepts the optical signal over the SSMF and is converted into 60 GHz radio frequency (RF) for connecting various RUTs over wireless channels. Any wireless communication system is highly needed to transmit signals, which provide high stability and high spectral efficiency. Also, it allows full utilization of the available spectrum bandwidth. With the recent development in the higher order modulation format and IEEE wireless standard, M-ary QAM (m-QAM) represents higher spectral efficiency as compared to the other modulations formats such as binary phase-shift keying (BPSK), amplitude shift keying (ASK) and quadrature phase-shift keying (QPSK) [11–13]. Therefore, M-ary QAM is widely used in mm-wave wireless access systems. It is also important to note that in the case of QAM-based modulation format, as the order of m-ary increases with lower symbol rates shows better spectral signal efficiency, and when such signal is transmitted over optical fibre, the effect of dispersion reduces effectively and hence, m-QAM becomes more potential modulation format for high-capacity data transmission over long haul RoF link [14–18].

Numerous pieces of literature are available on RoF, and it is reported that the mm-wave RoF wireless communication systems realised several tens of gigabyte signal transmission. However, it is limited to a significantly shorter distance SSMF and uses a higher RF band for wireless transmission. As mentioned in the references [11–22], these studies and their outcomes encountered many technical challenges. Among all major challenges, a significant and crucial challenge is to combat the fibre’s impairments due to its dispersive nature.
and polarisation mode dispersion effect. Hence, a single sideband (SSB) based modulation format and 90° hybrid couplers are used in optical modulation in CS to reduce fibre dispersion. Also, highly precise and optimised biasing is applied to the optical modulator. Sharp roll-off-based optical filters are used to increase the spectral efficiency of the optical signal to transmit over SSMF. In the BS unit where optical heterodyne and coherent detection are carried out, a transmitter continuous wave (CW) laser and receiver CW lasers are exhibited a frequency offset which can be in the order of some few megahertz to gigahertz. This may lead to a complete loss of phase with respect to the reference baseband modulation constellation symbols. Therefore, frequency offset correction is carried out using the fourth power algorithm [23–25]. The linewidth dependencies for the phase noise contribution by free-running CW laser would increase the design complexity in the receiver DSP chain for carrier phase noise estimation and correction [26–29].

In view of the above, considering all complexity and the system requirements, we propose an RoF system design that establishes the high-capacity data transmission at 168 Gb/s input data rate as represented in (Fig. 1). The higher order m-ary 64-QAM is used to modulate the input data. A higher linewidth-based CW laser is used as an optical source, and its phase noise impact is analysed on the baseband 64-QAM signal. The optical modulation is designed using an orthogonal polarised Mach-Zender Modulator (MZM). MZM is the modulating device that mainly works on the principle to generate the amplitude and phase modulation in the form of interference of the signal, which is propagated over SSMF. The baseband In-Phase (I) and Quadrature (Q) components of the 64-QAM signal are optically modulated using two orthogonal arms of the LiNbO3 MZM modulator (LiNbO3-MZM). The transmitter CW laser is tuned at the operating frequency of 193.4144 THz at CS. The optically modulated 64-QAM signal is transmitted over SSMF with chromatic dispersion of 16.75 ps/nm/km. Similarly, at the receiver side in the BS, orthogonal polarised optical heterodyne and coherent detection are carried out, which generates a 60 GHz mm-wave RF signal. Optical heterodyne coherent detection is designed using 90° mm-wave RF using 3-dB couplers and one phase shifter. The 90° optical hybrid can be used for coherent signal demodulation for either homodyne or heterodyne detection [30, 31]. At BS, the receiver CW laser is tuned at a frequency of 193.3544 THz with the same linewidth value.

[Fig. 1] Block diagram of proposed 60 GHz (mm-wave) RoF system
as the transmitter CW laser. After mixing the transmitted optical carrier from CS with the receiver CW laser at BS, the offset frequency difference of 60 GHz mm-wave RF signal is generated. The other generated frequency components due to the mixing process are discarded using the appropriate filter. This generated 60 GHz RF signal is transmitted over a wireless link to connect several RUTs.

At the RUT, the generation of the orthogonal arm of the baseband signal and demodulation is carried out using an IQ demodulator. The IQ demodulator mainly consists of a 60 GHz local oscillator (LO) signal generator, mixer and low pass filter (LPF). The channel noise of wireless transmission highly impairs the received 60 GHz signal. The received noisy RF signal is mixed with LO and passed through the LPF with the cut-off frequency of 0.2xdata rate, which essentially reduces the intermodulation products generated by the mixing process. The baseband 64-QAM received signal is then passed through ADSP. The ADSP chain is designed with different sections consisting of filtering, resampling, IQ imbalance compensation, chromatic dispersion (CD), polarisation mode dispersion (PMD) and fibre non-linearity (NL) effect compensation. It also consists of timing recovery to compensate for the delay, adaptive equalisation to mitigate PMDs, desired symbol trajectory recovery, and a down-converter to convert signal at the desired sample per symbol rate to match with the symbol rate of baseband signal generation at the transmitter. Carrier Phase estimation (CPE) and Frequency Offset estimation (FOE) are also used to mitigate the phase and frequency deviation encountered while the optical demodulation modulation process. The deviation in the phase and frequency is generated due to the optical device’s characteristics, such as laser linewidth and inherent frequency offset between two lasers. After processing the signal through the ADSP chain, the received symbols are remapped using a 64-QAM demodulator and generate the received data stream. The received data and transmitted data are analysed, and obtained results are discussed in the form of Bit Error Rate (BER), Error Vector Magnitude (EVM) and received signal constellation diagrams.

This paper is organized as follows: In Section 2, we present an overview of the theoretical foundations of the proposed 60 GHz RoF system design. More precisely, we emphasized the ADSP structure and the overall ADSP chain implementation and its framework impact in the high-capacity RoF system. In Section 3, we elaborated the complete system design aspects with all requisite practical parameters, and, most important, subsystem ADSP and its impact on the overall system design, followed by the system performance outcome results and analysis, are very well compared with other proposed systems. For the designed 60 GHz RoF system, we conclude that the system’s performance in Section 4 and elaborate on the proposed system’s importance and appropriateness fulfil the requirement of future communication needs.

2 System Design and Advanced Digital Signal Processing Chain

The proposed RoF system is mainly categorized into three parts viz i. CS ii. BS and iii. RUT. The detailed system design is shown in (Fig. 2). The CS mainly consists of high data rate input bits stream generation followed by the modulation using 64-QAM and conversion of the electrical signal into an optical signal with an orthogonal arm-based LiNbO3-MZM optical modulator. The transmitter CW laser is tuned at a fundamental frequency \( f_0 = 2\pi/\omega_0 \), which is emitted the light waves of \( E_0(t) = E_0\exp(j\omega_0t) \). The modulated 64-QAM baseband I and Q signals are optically modulated with light waves generated by the CW laser source using two orthogonal polarized LiNbO3-MZM arms.
The biasing voltages of LiNbO3-MZM are adjusted and optimized to modulate I and Q signals with lower optical signal loss. The biasing arrangement of the LiNbO3-MZM is generated SSB modulated optical signal tuned at CW laser frequency. The baseband I and Q signals of 64-QAM are represented as,

$$S(t) = I(t)\sin\omega_{RF}t + Q(t)\cos\omega_{RF}t$$

The SSB modulation generation using MZM modulator is configured with bias voltages of $v_\pi/2$ and phase difference of $\pi/2$. The modulation index is chosen as $\alpha = \pi I(t)/V_\pi$ and $\beta = \pi Q(t)/V_\pi$. The generated SSB optical modulated signal is represented as,

$$E_{eq}(t) = \gamma_1\gamma_2E_0I_2(m_h)\left\{ \exp(j(\omega_0 + 2\omega_{RF})t + \frac{\pi}{V_\pi}[I(t) + jQ(t)]\exp(j(\omega_0 + 2\omega_{RF} + \omega_{IF})t) \right\}$$

where $\gamma_1, \gamma_2$ are the insertion loss of MZM. This $E_{eq}(t)$ optical signal generated at CS is transmitted through SSMF with chromatic dispersion of 16.75 ps/nm/km and attenuation of 0.2 dB/km. Nonlinearity parameters and PMDs are also considered while carrying out the system analysis.

60 GHz mm-wave signal generation is done by intercepting the transmitted optical signal through the SSMF at BS. The BS is designed using the optical heterodyning coherent detection technique. The coherent detection is realised by using a high-speed photodiode (PD). PDs are optoelectronic devices that generate an electric current. $I_p(t)$ which is proportional to the squared modulus of the optical input field $E_{in}(t)$. The
generated mm-wave signal at the coherent detector is expressed by Eq. (3). This consists of modulating signal and carrier signal.

\[
I(t) = \mu r_1^2 \gamma E_0^2 J_0(m_0) \\
\left\{ 1 + \frac{1}{\gamma_1^2} + \frac{\pi}{\gamma_2^2} \left[ I^2(t) + Q^2(t) \right] + 2 \gamma_2 \cos 4\omega_{RF} \\
+ 2 \gamma_2 \pi \left[ I(t) \cos (4\omega_{RF} + \omega_{IF}) t + Q(t) \sin (4\omega_{RF} + \omega_{IF}) t \right] \\
+ 2 \gamma_2^2 \left[ I(t) \cos \omega_{IF}(t) + Q(t) \sin \omega_{IF}(t) \right] \right\} 
\]

\( \mu \) is the sensitivity of photo diode.

In the BS, optical heterodyne coherent detection is designed using a 2×2 balance detector, 90° optical hybrid and 3 dB couplers for converting the optical signal into the orthogonal electrical I and Q signals [21, 22]. This orthogonal electrical I and Q signal are combined using a power combiner and converted to an RF signal centred at the frequency of 60 GHz. To generate 60 GHz RF using an optical heterodyne process, the receiver CW laser source is tuned at a 60 GHz frequency offset with reference to the transmitter CW laser operating frequency.

60 GHz RF signal is transmitted over the air to connect the different RUTs. The RUTs are designed with IQ demodulator, ADSP chain and 64-QAM de mapper. The received noisy signal is initially converted into the baseband by mixing the input signal with 60 GHz LO, and the intermodulation product generated by this mixing process is filtered out using sharp roll-off LPF. This baseband signal is then passed through the ADSP chain comprised of different signal processing blocks starting with preprocessing, dc blocking, chromatic dispersion estimation and correction, adaptive equalization, CPE and correction, FOE and correction, and finally ending with signal recovery. The sequence of blocks adapted in the ADSP system is depicted in (Fig. 3). At the preprocessing stage, signal conditioning is carried out by passing the signal through Bessel filters, resampling and IQ compensations. The implementation design details are described as follows,
2.1 Bessel Filter, Resampling and IQ Compensation

The received RF signal at the input of RUT is highly affected by the transmission channel bandwidth noise. Due to the optical modulation and demodulation process, there is an extra DC shift in the signal caused by voltages applied to the optical devices. A third-order Bessel Filter is used to compensate for this DC shift, which has a better shaping factor, flatter phase delay, and group delay. The output signal at Bessel Filter is then passed through the resampling process, where the signal is resampled at 2 samples/symbol followed by IQ compensation and normalization. The cubic interpolation method is designed to resample the I and Q signals in this preprocessing implementation. The IQ signal’s amplitude and phase are also imbalanced due to fibre impairments. To mitigate this, an IQ compensation, an orthogonalization block, is used. The orthogonalization is accomplished by the Gram-Schmidt orthogonalization procedure (GSOP). In addition to the orthogonalization, it normalized the IQ components with unitary power. The operation involved in the GSOP has been elaborated in the ref. [32].

\[
\begin{align*}
    r_{I}^{ort}[n] &= \frac{r_{I}[n]}{\sqrt{E\{r_{I}^{2}[n]\}}} \\
    r_{Q}^{int}[n] &= r_{Q}[n] - \frac{E\{r_{I}[n]r_{Q}[n]\}r_{I}[n]}{E\{r_{I}^{2}[n]\}} \\
    r_{Q}^{ort}[n] &= \frac{r_{Q}^{int}[n]}{\sqrt{E\{(r_{Q}^{int}[n])^{2}\}}} \\
\end{align*}
\]

\(r_{I}[n]\) and \(r_{Q}[n]\) are the I and Q discrete time signal and \(E[.\] is the expectation operator. Initially, \(r_{I}[n]\) is normalised and generated the \(r_{I}^{ort}[n]\) followed by projection of \(r_{Q}[n]\) with reference to \(r_{I}[n]\) in orthogonal direction and generated \(r_{Q}^{int}[n]\) and finally with normalization process generates \(r_{Q}^{ort}[n]\).

2.2 Chromatic Dispersion, Non-Linear Compensation and Timing Recovery

In the proposed implementation, digital filtering is used to compensate chromatic dispersion of the optical fibre. A frequency domain and time domain filtering are designed and implemented as per ref. [33]. The frequency domain transfer function is given by,

\[
G(z, \omega) = \exp\left(-j.D.\frac{\lambda^2}{4.\pi.C}.\omega^2\right)
\]

where \(z\), \(\omega\), \(\lambda\), \(C\), \(S\), and \(\lambda_{0}\) are the transmission distance, angular frequency, wavelength, speed of light, dispersion slope and reference wavelength respectively. The dispersion coefficients of the fibre are represented as \(D = D_{0} + S \times (\lambda - \lambda_{0})\). Similarly, the time domain implementation uses \(N\) taps finite impulse response (FIR) filter. The tap weights of the FIR are calculated as,

\[
a_{k} = \sqrt{\frac{j.\pi.C.T^2}{D.\lambda^2 \times z}} \cdot \exp\left(-j.\pi.C.T^2 \cdot \frac{k^2}{D.\lambda^2 \cdot z}\right) - \left\lfloor \frac{N}{2} \right\rfloor \leq k \leq \left\lfloor \frac{N}{2} \right\rfloor
\]

\(T = \pi/\omega_{n}\), \(\omega_{n}\) is the Nyquest frequency.
A digital back propagation (BP) algorithm compensates for the nonlinear fibre impairments [33]. The BP mainly designed by considering the inverse nonlinear Schrödinger equation (NLSE) to solve the optical link parameter. The NLSE is represented by,

$$\frac{\partial E}{\partial (\pm z)} = (D + N)E$$  \hspace{1cm} (7)

$D$ and $N$ are the differential operator and nonlinear operator respectively, and $E$ is the signal received. $D$ and $N$ can be estimated as,

$$D = \frac{j}{2} \beta_2 \frac{\partial^2}{\partial t^2} a$$  \hspace{1cm} (8)

$$N = j\gamma |E|^2$$  \hspace{1cm} (9)

$\beta_2$ is dispersion parameter group velocity, $a$ is the attenuation factor and $\gamma$ is the nonlinearity parameter. The Split-Step Fourier method (SSFM) is used to solve the above Eqs. (7 and 8), [34]-[35]. The phase shifts between each sample are calculated by,

$$\theta_{NL}(t) = k\gamma L_{eff} |E|^2$$  \hspace{1cm} (10)

where $k$ is optimised compensation factor, and effective length of each step is represented by $L_{eff}$. In case of more than one fiber spans, it is then compensated by each BP step and accordingly $L_{eff}$ is represented as,

$$L_{eff} = s \frac{1 - \exp(-\alpha L_{span})}{\alpha}$$  \hspace{1cm} (11)

where $s$ and $L_{span}$ are the number of fiber spans and length of each span respectively.

The timing recovery algorithm adaptively determines the sample timing instances of the respective symbols. The sampling frequency and sampling phase for the received symbols can be deviated due to oscillator frequency drift stated for specific symbol rates. Similarly, the filtering process also introduces a substantial timing delay. To mitigate these issues, the timing recovery algorithm is implemented by the digital square and filter algorithm [36]. The received signal is given by,

$$r(t) = \sum_{k=-\infty}^{\infty} a_k g_T[t - k.T - \epsilon(t).T] + n(t)$$  \hspace{1cm} (12)

where $a_k$ are the transmitted symbols, $g_T(t)$ is the transmission signal with symbol duration of $T$, and $n(t)$ is that channel noise. $\epsilon$ is a time delay which varies slowly. The received signal is passed through the filter with the impulse response of $g_R(t)$ and sampled the signal at a rate of $4/T$ and resulting in the following samples,

$$\tilde{r}(t) = r(t) \otimes g_R(t)$$  \hspace{1cm} (13)

$$\tilde{r}_k = \tilde{r} \cdot \frac{KT}{4}$$  \hspace{1cm} (14)

The sampled filtered and square input signal is represented given below at $1/T$. 

\begin{figure}[h]
\centering
\includegraphics[width=\textwidth]{ Springer}
\end{figure}
\[
x_k = \left| \sum_{m=-\infty}^{\infty} a_m \cdot g \left( \frac{KT}{4} - m.T - \varepsilon.T \right) + \tilde{n} \left( \frac{KT}{4} \right) \right|^2
\]  

(15)

\[
g(t) = g_{t1}(T) \otimes g_{R}(T)
\]  

(16)

Equation (16) represents the filtered samples of the squared input signal. This spectral component is calculated and determined for every section of the length from \(4L\) signal samples.

At last complex Fourier coefficients at the desired symbol rate is computed and represented below,

\[
X_n = \sum_{k=4nL}^{4(n+1)LN-1} x_k e^{-j2\pi k/n}
\]  

(17)

The normalized phase is calculated by,

\[
\hat{\varepsilon} = -\frac{1}{2\pi} \arg(x_n)
\]  

(18)

### 2.3 Adaptive Equalizer

An adaptive equalizer (AE) is used to reduce the residual chromatic dispersion, PMD and inter-symbol interference. The AE is performed by a finite-impulse response filter applied to the orthogonal polarization signal. The AE is implemented by butterfly structure architecture with a two-stage constant modulus algorithm and radius-directed equalization (CMA-RD) [37]-[38]. The generalized cost function of the CMA is given by,

\[
J(k) = E \left[ |y(k)|^2 - R_p \right]^2
\]  

(19)

The AE scheme is designed using following cost function,

\[
J(k) = E \left[ |y(k)|^2 - R_p \right]^2 - E \left[ |y(k)|^2 \right]^2
\]  

(20)

The second term forces the equalizer to be non-circularly symmetric [39]. From the above Eqs. (19 and 20), the \(y(k)\) is the equalizer output and \(E[\ldots]\) is the statistical expectation. \(R_p\) is depending on the input data symbol \(a(k)\) with dispersion order \(p\) and it is represented as,

\[
R_p = \frac{E[|a(k)|^{2p}]}{E[|a(k)|^p]}
\]  

(21)

The equalizer output \(y(k)\) is obtained from the following,

\[
y(k) = W^H \cdot X(k)
\]  

(22)

\[
W = \left[ w_0(k), w_1(k) \ldots w_{N-1}(k) \right]^T
\]  

(23)
where $W$ is the weight vector of the equalizer, $X(k)$ is the input data vector and $N$ is the length of the equalizer tap weight. The tap weights vector is calculated using stochastic gradient algorithm with $\mu$ step size as per the following,

$$W_{(k+1)} = W(k) + \mu X(k) e^*(k)$$  \hspace{1cm} (25)

And error signal $e(k)$ is estimated by,

$$e(k) = y(k).\left( R_p - |y(k)|^2 \right)$$  \hspace{1cm} (26)

For a 64-QAM signal, $e(k)$ cannot reach zero; therefore, fine tuning equalisation is carried out by CMA as a first-order convergence and followed by the RD algorithm. The output of CMA $e(k)$ signal is optimised to the nearest constellation using second stage RD algorithm. RD determines the criteria for the detection of $e(k)$ as given below,

$$e(k) = \hat{R}_k^p - |y(k)|^p$$  \hspace{1cm} (27)

where, $R_k$ is the radius of the nearest constellation symbol for each equalizer output. The tap weights for the RD are further updated as follows,

$$W(k+1) = W(k) + \mu X(k) e^*(k)$$  \hspace{1cm} (28)

$$e(k) = \hat{R}_k^p - |y(k)|^p$$  \hspace{1cm} (29)

The AE implementation in the proposed system design is done with a large data sample and carries multiple iterations to get the desired result.

2.4 Frequency Offset and Carrier Phase Estimation

In coherent heterodyne detection, the transmitter and receiver laser are not operated at the same frequency and result in the frequency offset, which affects the complete loss of the symbol’s phases during demodulation. Assuming that after perfect equalization and demultiplexing of orthogonal polarization, the received signal is impaired by frequency offset, phase noise and additive noise, and such received signal can be expressed by,

$$S(k) = C(k).e^{(2\pi \Delta f T + \varphi_k)} + n(k)$$  \hspace{1cm} (30)

where $C(k)$ are the input data symbols, $\Delta f$ is the frequency offset and $\varphi_k$ is the carrier phase which varies due to $\Delta f$. The frequency recovery algorithm is used for the spectral estimation of received signal raised to the fourth power [26]. For 64-QAM signal with frequency offset, when it raised to the fourth power, the offset frequency peak in the spectrum can be easily estimated. The frequency offset $\hat{\Delta f}$ is estimated by,

$$\hat{\Delta f} = \frac{1}{4} \max_f \left| \text{FFT}\{ (y[k])^4 \} \right|$$  \hspace{1cm} (31)

where $\text{FFT}\{ . \}$ is the Fast Fourier Transform and $f$ is maximal spectral amplitude. After estimation of the $\hat{\Delta f}$, then carrier frequency offset is compensated by,
\[ z(k) = S(k)e^{-jk2\pi\Delta fT_s} \]  

(32)

The blind phase search algorithm recovers the phase mismatch between the receiver CW laser and the signal [28]. This is implemented to apply test rotations of \( \theta_b \) to the received signal \( z(k) \). The range of rotations \( p \) is equal to the angles of constellation symbols of the 64-QAM. For 64-QAM signal \( p = \frac{\pi}{2} \) is considered. Rotation angle \( \theta_b \) is estimated as,

\[ \varphi_k = \frac{b\pi}{B}, b \in \left\{ -\frac{B}{2}, \ldots, 0, 1, \ldots, \frac{B}{2} \right\} \]  

(33)

All rotated symbols \( z_b(k) \) are expressed by,

\[ z_b(k) = z(k)e^{\theta_b} \]  

(34)

Post rotation, the symbols are passed through the decision unit, which consists of minimum distance operator and quadratic distance calculation between symbols before and after. The squared distance between the symbols \( |d_{k,b}|^2 \) are calculated by,

\[ |d_{k,b}|^2 = |z_b(k) - [z_b[k]]_D|^2 \]  

(35)

where \([\cdot]_D\) is minimum distance operation.

An averaging method is used to mitigate the effect of noise. It is considered as the \( N \) symbols block, which has same phase noise. The summation of such \( N \) consecutive test symbols by the same carrier phase angle \( \varphi_k \) is represented by,

\[ S_{k,b} = \sum_{n=1}^{N} |d_{k-n,b}|^2 \]  

(36)

The CW laser line-width and symbol duration \( T \) product affect the determination of optimum value of \( N \). The choice of \( N \) is made based on the calculation of noise power and SNR. Finally, the estimated phase \( \hat{\theta}[k] \) is made based on \( \theta_b[k] \) from the minimized distance sum of the decoded output symbols which is selected from \([z_b[k]]_D\). This proposed design uses unwrapping after calculating phase noise to remove the pole ambiguity of 64-QAM constellation demapping.

3 System Analysis, Results and Discussion

The proposed system design of a high-capacity 60 GHz mm-wave RoF link is implemented in the optisystem. The performance analysis and validation of the outcome results are carried out in Matlab. The transmitter CW laser in CS is tuned at the operating wavelength of 1550 nm (193.4144 THz) with a large linewidth of 10 MHz. The larger linewidth characterises the phase noise impact on the received 64-QAM symbols. The laser is acted as a light source of LiNbO3-MZM orthogonal optical modulator at CS. The biasing voltages of LiNbO3-MZM are adjusted and optimised to the value of \( V\pi/2 \). The 64-QAM modulator is used to generate the baseband I and Q signal for the input data stream of the random bit generator at the data rate of 168 Gb/s. The symbol rate of the baseband signal is 28 Gbaud. Further, this baseband signal is passed through the optical modulator to convert electrical into an optical signal and subsequently transmitted to the BS through SSMF.
The generated optical spectrum of the 64-QAM signal is shown in (Fig. 4). The maximum amplitude of the 64-QAM optical signal is observed as -50 dBm. The lower power generation is due to losses incurred in the modulation process. In order to compensate for these losses, an optical booster amplifier is used with a gain of 20 dB. The SSMF, which has a dispersion of 16.75 ps/nm/km and attenuation of 0.2 dB/km, is used to transmit the optical signal from CS to BS at a maximum distance of 172 km. The loss of the SSMF is compensated by the optical pre-amplifier so that desired amplitude of the received signal can be reproduced at BS. The amount of optical power interception at the BS mainly depends upon the modulation scheme type and its associated symbol rate. The wide band signal with a 28 Gbaud symbol rate is passed through the SSMF, and essential amplification of 20 dB is provided to the signal before and after optical fibre. During the system design and analysis, the best possible achievable symbol rate of 28 Gbaud using a 64-QAM modulation scheme is identified with several iterations and implementation to transmit the data for such a long distance. The amount of optical signal generated with a -50 dBm signal is sufficient to transmit the signal through SSMF with additional pre and post-booster amplifiers.

At the BS, the optical signal is detected and converted into the RF signal using the optical heterodyne coherent detection technique. The Gaussian filter with a bandwidth of 2\times symbol rate is designed to filter out the unwanted noise generated due to channel characteristics before the optical receiver. The coherent detector is implemented using a 90-degree optical hybrid consisting of a high-speed photodiode. In the optical heterodyne receiver, a 90-degree hybrid is generated four intermediate frequency (IF) signals. These signals are generated by coupling and phase shift operation between the receive optical signal and the receiver CW laser signal, its orthogonal counterpart. These four-arm signals are combined and generate the IF signal. The receiver CW laser is tuned at the operating wavelength of 1550.48 nm (193.3544 THz). The operating wavelength of the receiver CW laser is tuned at 60 GHz offset with reference to the transmitter CW laser and results in the electrical signal generation centred at 60 GHz. The 60 GHz RF signal consisting of a 64-QAM baseband envelope is transmitted over the wireless link to connect various RUTs.

The generated RF spectrum of the 60 GHz signal at BS is shown in (Fig. 5). The maximum amplitude of the detected RF signal is observed as -30 dBm at a distance of 172 km.
of SSMF. It can be elucidated from (Fig. 5) the 60 GHz centre carrier of the RF spectrum is occupied with $2 \times$ Symbol rate bandwidth, and such huge bandwidth data are propagated over the air to connect the RUTs.

Table 1 shows the parameters used to analyse the proposed system design, demonstrating the satisfied values of received optical power, EVM (%) and BER for the maximum distance of 172 km of SSMF. The system parameters in Table 1 are very close to the commercially available optical devices. Especially SSMF parameters such as operating
temperature, dispersion, dispersion slope and PMD coefficient are vital in the system design and analysis. Similarly, the other parameters, such as MZM bias voltage and extinction ratio, are major deciding parameters that generate the spectrally efficient optical spectrum over which the high-capacity data transmission is carried out. At the optical heterodyne coherent detection, the 90-degree optical hybrid network is designed with a high-speed photodiode and a responsivity factor of 1 A/W.

The wireless channel is considered an Additive White Gaussian Noise (AWGN) channel in this system design. The RUTs are designed with the chain of IQ demodulator, LPF, ADSP

![Fig. 6](image)

**Fig. 6**  
(a) Constellation diagram of 64-QAM signal prior to ADSP at a distance of 172 km of SSMF,  
(b) Constellation Diagram of 64-QAM signal after processing through ADSP at a distance of 172 km of SSMF
and 64-QAM demodulator. The IQ demodulator is implemented with a 60 GHz local oscillator (LO) with considerably lower phase noise and a mixer. The LO signal is mixed with the received 60 GHz RF from BS and generates the 64-QAM baseband signal, which is substantially affected by both optical and wireless channel impact. Due to the mixing process, it is also generated the intermodulation product. It is passed through the LPF with a cut-off frequency of 0.2×data rate to eliminate such unwanted signals. The constellation diagram of the 64-QAM baseband signal at the IQ demodulator output is shown in (Fig. 6a). This constellation diagram represents the signal detected at 172 km of SSMF. The constellation diagram is indicated that the phases of the symbol’s trajectories are completely mismatched due to channel impairments, as discussed in section-2. In order to mitigate these impacts, ADSP blocks are designed, and the received baseband signal is processed through these blocks. The ADSP module is designed with a chain of DSP algorithms, as discussed in section-2, paragraphs (2.1–2.4).

As depicted in the ADSP chain, the received signal is processed through various pre-processing followed by the recovery stage. At the processing stage, the noise sampled signal is added with the received sampled signal and passed through the DC blocking to correct any offset generated due to the voltages applied to the modulator, followed by normalization with reference to the 64-QAM format. The third-order Bessel filter with a bandwidth of 0.75×symbol rate filters out the spectrum band noise. The signal is resampled at the rate of 2 samples/symbols using the cubic interpolation method. Furthermore, the signal is passed through the GSOP algorithms to compensate for the IQ imbalances in phase and amplitude of the I-Q signals. Post IQ compensation, chromatic dispersion compensation is applied to the signal using a time domain finite impulse response filter. NLSE is applied to the filtered signal to mitigate the nonlinear characteristics of the optical fibre by calculating phase shifts of each sample which is compensated and optimized to the corresponding effective length of each step of the optical fibre. Subsequently, the signal is further passed through the digital square and filter timing recovery algorithm to synchronize the symbols, and the normalized phase is calculated by computing the Fourier coefficient of the symbol rate.

Two-stage CMA-RD algorithm compensates the inter-symbol interference, PMDs and residual chromatic dispersion. Initially, the CMA algorithm minimizes the error power, but it is not obtained the perfect equalization with error signal zero; hence the signal is further passed through the RD algorithm to get fine-tuned equalization. In order to get fine-tune the equalization, we used multiple iterations with extensive sample data followed by signal down sampled by 1 sample/symbol. The signal is further processed through the FOE using a signal raised to the fourth power to estimate the frequency offset and corrected accordingly.

The most important part of the ADSP is CPE using the BPE algorithm. As depicted in Table 1, the CW laser linewidth of 10 MHz is used, which is more than the commercially available CW laser. However, to analyse the phase noise impact with reference to considered linewidth is compensated by effectively implementing the BPE algorithm followed by the averaging method to reduce the noise. This effective implementation of the ADSP chain recovered the baseband I-Q signal at a distance of 172 km, as shown in (Fig. 6b). The outcome of the ADSP chain significantly improved the signal representation of the noisy 64-QAM baseband signal. From the constellation diagram, it can be easily seen that the symbol trajectory of the processed signal is matched with reference to the 64-QAM constellation, as depicted in (Fig. 6b). Similarly, to get a more precise picture, the pre- and post-ADSP constellation diagram of the 64-QAM signal at a distance of 150 km of SSMF is shown in (Fig. 7a) and (b) respectively.
The received signal after ADSP is passed through the 64-QAM demodulator and converted into the required digital bits stream. EVM (%) and BER are calculated on the transmitted and received symbols and bits stream. EVM (%) of the received baseband signal is calculated by,

\[
EVM(\%) = \sqrt{\frac{\sum n (\hat{x}_n - a_{\text{opt}} x_n)^2 + (\hat{y}_n - a_{\text{opt}} y_n)^2}{\sum n (a_{\text{opt}} x_n)^2 + (a_{\text{opt}} y_n)^2}}
\]  \hspace{1cm} (37)

Fig. 7  a Constallation diagram of 64-QAM signal prior to ADSP at a distance of 150 km of SSMF, b Constallation diagram of 64-QAM signal after processing through ADSP at a distance of 150 km of SSMF
This equation represents $n$th symbol of the received I-Q signal $\hat{x}_n$ and $\hat{y}_n$ respectively. Similarly, $x_n$ and $y_n$ are the input symbols and $a_{x}^{opt}$ and $a_{xy}^{opt}$ are the optimized attenuation coefficient of I-Q signal level and calculated as, $a_{x}^{opt} = \langle \hat{x}_n \rangle$ and $a_{xy}^{opt} = \langle \hat{y}_n \rangle$ and finally mean of variation is calculated for transmitted and received symbols in percentage. The EVM is calculated for 7.015% against the received optical power of 7.886 dBm at 172 km of SSMF. Similarly, at 150 km of SSMF, the EVM is calculated as 1.006% against the received power of 8.35 dBm. It can be noted that there is an eventual reduction in the EVM with the reduction in the lengths of the SSMF varying from 172 to 150 km. The reduction in the EVM can also be seen with increased optical power at a lower distance. The EVM (%) vs received power plot at the longer distance of SSMF is shown in (Fig. 8). The optical power generation at the SSMF distance varies from 150 to 172 km is shown in (Fig. 9). It is also observed that the reduction into the optical power from 8.35 dBm to 7.886 dBm with the distance variation of 150 km to 172 km.

In order to get system quality of service, BER is estimated between the transmitted and received bits stream of 64-QAM by,
where, $E_b$ is the energy of the bit and $N_0$ denoted the amount noise injected in the signal. According to the BER curve, as shown in the (Fig. 10), it is observed that BER value $2.7 \times 10^{-3}$ is achieved at SSMF distance of 172 km, which is reasonably good and would be most suitable for long haul communications. It can be clearly seen from the (Fig. 10), that the variation of BER value from $10^{-5}$ to $2.7 \times 10^{-3}$ is function of length of the optical fiber from 150 to 172 km. Practically, for the reproduction of any transmitted data over the channel at the receiver end, the best possible BER requirement would be $10^{-3}$ with the desired SNR, and it is also depending upon the type of modulation scheme and signal occupied bandwidth. This system analysis shows that 30 dB SNR is required to reproduce the signal and achieve the BER of $2.7 \times 10^{-3}$ at 172 km for 28 Gbaud symbol rate transmission using the 64-QAM modulation process.

Table 2 is highlighted the comparison studies of the similar designs based on the results of the system’s performance of a 60 GHz RoF system using 64-QAM. As per the comparison studies shown in Table 2, it is observed that the proposed system design is supported for long-distance communication by using the ADSP technique upto the 172 km with an achievable min BER of $2.7 \times 10^{-3}$. Apart from this, it has increased a significant amount of bandwidth utilization for 28 Gbaud high payload data transmission over longer distance SSMF and maintained the EVM well below its limit of $< 7.015\%$.

4 Conclusion

The proposed system design of 60 GHz mm-wave RoF at the data rate of 168 Gb/s modulated using 64-QAM and 28 Gbaud high-capacity symbol rate transmission of the optical signal generated by orthogonal arm MZM modulator with optimum value of CW laser biasing generated spectrally efficient signal at CS. The transmitted optical signal at a maximum distance of 172 km of SSMF is intercepted by the high-speed photodiode-based
optical heterodyne coherent detection technique at the BS. It would become a cost-effective solution for generating a 60 GHz RF signal. On the air, transmitted 60 GHz RF is received at RUTs and processed through the ADSP chain. The significant improvement and achievement of the system performance are analysed by the effective implementation of ADSP at such a long-distance transmission at a 28 Gbaud symbol rate. The ADSP chain is responsible for reconstructing the baseband 64-QAM signal trajectory for the 172 km distance of SSMF. The system performance is also evaluated and analysed through a received signal constellation diagram, EVM calculation and BER curve. It is also observed that the ADSP module’s signal processing maintained the EVM of less than 7.015% and BER of 2.7 × 10⁻³ at a 172 km distance. Given the above, the proposed RoF system is the most suitable and reliable solution for forthcoming beyond 5G and 6G technologies, which requires high-capacity data transmission, maximum bandwidth utilisation and long-distance communication.

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**Declarations**

**Conflicts of interest** The authors declare that they have no conflict of interest.

**Human Participants and/or Animal** This article does not contain any studies with human participants or animals performed by any authors.

**Informed Consent** All referred study is highlighted in the Literature Review.
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Devendra Chack is an Associate professor at the Department of Electronics Engineering, Indian Institute of Technology (ISM) Dhanbad. He received his Bachelor of Engineering degree in Electronics Engineering from Madhav Institute of Technology and Science (MITS) Gwalior, India. He received an M. Tech. and a PhD degree from the Indian Institute of Technology (ISM), Dhanbad, India. His research interests include On-chip silicon Photonics, Advanced optical Communication. He is a potential reviewer of many journals and conferences, such as the IEEE, SPIE, and Elsevier publications. He is currently handling many research projects in the capacity of a Principal Investigator and Co-Principal Investigator sponsored by different funding agencies like the Science and Engineering Research Board SERB/DST, BRNS/DAE and TEQIP-III. He is a senior member of IEEE, Optica and SPIE. He is also a faculty advisor of Optica (formally known as OSA) student chapter IIT (ISM) Dhanbad.

Sunil Narayan Thool received the B.E and M. Tech. Degree in Electronics & Telecommunication Engineering and VLSI from Nagpur University, Nagpur, Maharashtra, India, in 2004 and 2010 respectively. He is pursuing a PhD in Electronics Engineering at the Indian Institute of Technology (ISM), Dhanbad, Jharkhand. His research interests include Optical signal processing, Radio Over Fiber using 60 GHz mm-wave and Probabilistic Constellation Shaping technique optimization for high data rate optical communication systems, and Digital Signal Processing.