Design of PAM-8 VLC Transceiver System Employing Neural Network-Based FFE and Post-Equalization

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Abstract: Wireless communication technology adopting electromagnetic waves in an unlicensed spectrum, such as visible light, for communication has attracted wide research efforts. Visible light communication (VLC) utilizes visible light as a communication medium to transmit signals but faces a limited communication bandwidth and low data rate, which is caused by the intrinsic characteristics of LEDs. This paper first studies a mathematical model of limited bandwidth and its effect on transmitted signals and then analyzes the free space and underwater channel loss. With the theoretical analysis, a VLC transceiver system is presented for solving bandwidth limitation by utilizing a pulse-amplitude modulation-8 (PAM-8) scheme and a hybrid equalization method. The proposed hybrid equalization combined a passive equalizer, a neural network (NN)-based feed-forward equalization (FFE), and a radial basis function neural network (RBF-NN). The feasibility of this VLC system was verified through a co-simulation platform with both free-space and underwater channels. Compared with a VLC system adopting a deep neural network (DNN)-based post-equalization method, the proposed VLC system could achieve a data rate of 3.6 Gbps with a bit error rate (BER) of $3.8 \times 10^{-3}$ over a 3 m free-space channel. The RBF-NN achieved a reduced training time of 10 min, which was 86.7% lower than the conventional DNN-based post-equalization method.

Keywords: visible light communication (VLC); neural networks (NNs); feed-forward equalization (FFE); PAM-8 modulation; optical wireless transceiver circuit

1. Introduction

With the rapid growth in network traffic caused by the increasing number of communication devices, high-quality media transmission, and real-time contact requirements, radio spectrum congestion has become a severe problem. Radio frequency (RF) signals only cover the 20 kHz to 300 GHz bandwidth in the electromagnetic spectrum, while visible light covers the 400 THz to 800 THz spectrum bandwidth [1]. This wide, unlicensed spectrum bandwidth provides a solution to the radio spectrum congestion problem, and it has led to research focused on visible light communication (VLC).

VLC technology utilizes visible light as a communication medium to transmit programmed data by changing the intensity of the light source. Then, the modulated light is received by a photodiode (PD) and transferred to a photoelectric current, which is processed in a receiver to recover the transmitted signal [2]. Beyond providing sufficient bandwidth to relieve radio spectrum congestion, other unique characteristics of visible light, such as its line-of-sight (LOS) property and lack of electromagnetic interference (EMI), guarantee the superiority of VLC compared with traditional RF-based technologies [3]. Therefore, VLC technology has been considered as a potential indoor communication method with high security and low cost. Meanwhile, visible light owns a much lower attenuation in water than RF signals. Thus, VLC technology is suitable for underwater communication.

However, VLC technology suffers from intrinsic problems, such as limited communication bandwidth and limited signal-to-noise ratio (SNR) for high-order modulation.
limited communication bandwidth impedes the transmitted signal data rate in VLC systems since commercial white LEDs only provide from several MHz to 10 MHz of communication bandwidth \[4,5\]. To increase the communication data rate, there are three solutions. The first one is changing the device structure to downscale the size of LEDs. To increase the communication data rate, the most straightforward method is increasing the symbol rate through extending the modulation bandwidth. The main limitation of LED bandwidth is caused by large intrinsic capacitance, which can be reduced by downscaling the size of LEDs. Therefore, micro-LEDs (µ-LEDs), the size of which is a small percent of normal LEDs, have hundreds of MHz of bandwidth \[6,7\]. Due to the limitation on bandwidth caused by the intrinsic characteristics of LEDs, some works have sought µ-LEDs with larger bandwidths for implementing high-speed VLC systems \[8–10\]. The second solution is applying high-order modulation schemes, such as pulse-amplitude modulation (PAM), QAM, carrier-less amplitude and phase modulation (CAP), and discrete multi-tone modulation (DMT) \[11–14\], to increase the efficiency of modulation bandwidth. This method increases the data rate without changing the symbol rate by improving the data number carried by each symbol. The third one is utilizing equalization schemes, such as passive and active equalizers, feed-forward equalization (FFE), decision feedback equalization (DFE), and neural network (NN)-based equalization \[12,13,15–18\], to extend the modulation bandwidth. These solutions have been adopted synthetically to further extend the modulation bandwidth. In \[8\], by adopting a µ-LED that owned the 906 MHz bandwidth, a 4 Gbps underwater VLC with a PAM-4 modulation scheme was achieved over a distance of 2 m. A 655 MHz µ-LED was utilized for implementing a data transmission rate of 11.95 Gbps for a VLC with an orthogonal frequency-division multiplexing (OFDM) modulation scheme \[9\]. With a T-bridge pre-equalizer, the −3 dB bandwidth of a VLC system based on HV-LED was extended to more than 700 MHz, which was used to achieve a 3 Gbps underwater VLC transmission over 1.2 m with a 16-quadrature amplitude modulation (QAM) \[10\]. In \[12\], PAM-4 and FFE were combined to increase the communication data rate to 2 Gbps. In \[13\], a 64-QAM was adopted for a higher data rate with an adaptive deep-learning equalizer. A CAP modulation scheme with a hybrid post-equalizer was utilized in \[13\], achieving a data rate of 8 Gbps. DMT modulation with an adaptive bit- and power-loading algorithm was adopted in \[14\] to achieve a data rate of 1 Gbps. Ref. \[15\] combined DCO-OFDM and long short-term memory (LSTM), while \[16\] presented 64-QAM signal transmission with a joint time–frequency neural network to realize an underwater VLC system with a data rate of 2.85 Gbps. In \[17\], PAM-4 and PAM-8 schemes were adopted with an LSTM equalizer to reach a communication distance of 1.2 m at 1.15 Gbps, and an underwater VLC system with a data rate of 1.22 Gbps was implemented in \[18\] with PAM-7 modulations and a density-based spatial clustering of applications with noise (DBSCAN) algorithm.

VLC systems implemented with discrete components can lead to large parasitic components, which introduce extra noise to the transmitted signals, increase the signal loss, and limit the bandwidth. The aforementioned results cause stability and installation problems. High-bandwidth µ-LEDs are expensive and not achievable in practical implementation. From the perspective of practical application, a highly integrated VLC system is superior to a discrete system. The typical architecture of an integrated VLC transceiver system is presented in Figure 1, which consists of a transmitter, a light source, a light signal detector, and a receiver. On the transmitter side, the source data are encoded and modulated in the baseband. Then, the modulated signal is sent to the LED driver. This signal controls the driving current of the light source to generate a corresponding optical signal. On the receiver side, the transmitted optical signal is converted into an electrical signal by a PD. A transimpedance amplifier (TIA) converts the photoelectronic current into voltage with amplification for further signal processing. A clock and data recovery (CDR) module then extracts the clock and retimes the data for subsequent demodulation and decoding processes in the baseband.
There are several works related to the design of integrated VLC systems. In [19], a fully integrated VLC transmitter was designed and implemented in a 0.35 um process with a 266 kbps data rate. Ref. [20] proposed a micro-LED display driver with a VLC transmitter implemented in a 0.5 um process to achieve a data rate of 1.25 Mbps. Ref. [21] presented a 24 Mbps VLC receiver with an ambient light rejection function at a 1.6 m distance. In [22], a fully integrated receiver in a 0.13 um process for visible laser communication was presented with a maximum data rate of 500 Mbps. However, it can be noticed that these works have only focused on the design of either the transmitter or the receiver without systematically evaluating the integrated transceiver system. In addition, these works have performed with rather low data rates compared with discrete VLC transceiver systems [13–18] due to a lack of suitable equalization schemes. In this paper, we propose a PAM-8 VLC transceiver system for implementing Gbps communication with a μ-LED with a 50 MHz bandwidth. It consists of an integrated analog front-end (AFE) and an offline digital baseband, which aims to overcome limited bandwidth by the combination of a high-order modulation scheme and a hybrid equalization method. The AFE is composed of a current-mode logic (CML) driver, passive equalizers, and a TIA. A 50 MHz μ-LED modelled in both circuit and Verilog-A code is used as the light source, which is fitted from the experimental data. To extend the bandwidth of this LED, passive equalizers applied after the driver and an NN-based FFE algorithm in the baseband are adopted on the transmitter side, while a radial basis function neural network (RBFNN)-enabled post-equalization is employed in the off-line baseband of the receiver. PAM-8 modulation is chosen for further increasing the data rate. The communication channel is either a 3 m free-space channel or a 1.2 m underwater channel. The main contributions of this work are summarized as follows:

1. A high-speed PAM-8 VLC transceiver system is proposed and simulated through the cooperation of MATLAB and Cadence. The AFEs of the transmitter and receiver, LED model, channel models, and PD model are all designed and implemented in Cadence, while the digital baseband, binary-weighted DAC in the transmitter, and flash ADC in the receiver are implemented in MATLAB.

2. NN-based equalization schemes are proposed and adopted as both the pre- and post-equalizations in the PAM-8 VLC system. In the transmitter, an FFE is implemented through a traditional NN in the baseband with the benefit of higher accuracy of the tap weights. In the receiver, an RBF-NN is proposed and adopted as the post-equalization, which is utilized to compensate for channel loss and solve the problems of modulation bandwidth limitation. With the combination of an NN-based FFE with passive equalizers serving as pre-equalization and the RBF-NN as post-equalization, the highest achievable data rate with a 3 m free-space channel is 3.6 Gbps.

3. Different communication channels are taken into consideration and compared. The proposed NN-enabled VLC transceiver system is also applied to realize underwater VLC. An underwater channel model is designed and analyzed for simulating the performance of underwater VLC with the same transceiver. This underwater VLC system is compared with experimental results and verifies the feasibility of the proposed hybrid equalization scheme in an underwater channel.

The remainder of this paper is organized as follows. Section 2 discusses the adopted models of the devices, as well as the channels, and their main limitations caused in VLC transceiver systems. Section 3 provides solutions to relieve these limitations for improving
communication quality. In Section 4, the proposed VLC transceiver system, including both digital basebands and AFEs, is presented and analyzed in detail. The simulation results and the corresponding analysis are demonstrated in Section 5. Lastly, Section 6 concludes this paper.

2. Analysis of Models and Limitations in VLC Systems

In a VLC system, the data rates of transmitted signals are mainly limited by the physical restrictions caused by the intrinsic characteristics of optical devices. In addition, the channel loss, nonlinearity of devices and noises can lead to inter-symbol interference (ISI) and the distortion of original signals. To achieve a transmission with a higher data rate, eliminating these unwanted effects is significant. Moreover, realistic models of the optical devices and channels are necessary for better elimination of limitations since they can provide systematic analyses with mathematical calculations of how the physical restrictions relate to the interference in the signal.

2.1. µ-LED Model

The LED device utilized to generate light signals in this VLC system was a red-green-blue (RGB) µ-LED with a pixel size of 200 µm × 100 µm and whole-package dimensions of 0.79 mm × 0.79 mm. As shown in Figure 2a, the compact equivalent circuit model was composed of a PCB test fixture model, a bonding wire model, intrinsic device model I, and a Verilog-A code model [22–24]. L1, R1, C1, L2, R2, and R3 were the passive components representing the PCB fixture model, while Lb, Rb, and Cb were the bonding wires. The intrinsic device model of the LED consisted of contact resistor Rg, active region capacitor Cj, and junction resistor Rs, which were all current-dependent. Since the parasitic components were negatable compared with the components of the intrinsic device in the low-frequency domain, the parasitic model mainly affected signals operating in the frequency domain beyond 500 MHz with second-order distortion. Therefore, when considering the lower-frequency parts of the LED frequency response curve, the dominant part was determined by the intrinsic circuit model, which was a first-order RC filter. The code model represented the opto-electric conversion, and the output optical power with a current through Rj followed the relationship of the first-order low-pass filter function. The values of these components in the compact model were extracted from the measured S-parameters of a fabricated µ-LED with the fitting process in ADS.

![Figure 2](image.png)

**Figure 2.** (a) Compact equivalent circuit model of µ-LED; (b) simulated and measured bandwidth curves of µ-LED.

The simulated results of bandwidth with this proposed compact equivalent model of µ-LED under different bias currents corresponded with the measured bandwidth curves and are presented in Figure 2b. It can be noticed that the −3 dB bandwidth of this µ-LED was 50 MHz when the bias current was 70 mA. This limited bandwidth was the main obstruction to increasing the communication data rate.
2.2. PD Model

The equivalent circuit model of the PD is presented in Figure 3, and it consisted of an intrinsic bandwidth limitation model and a diode dark current noise model. The intrinsic model comprised a Verilog code model representing carrier lifetime and photoelectric response and intrinsic parasitic capacitor $C_T$ with parasitic resistor $R_f$ for the $-3$ dB bandwidth of the PD. For the dark current noise model, $C_T$ was the terminal capacitance, which included the capacitances of the PD, pad, and bonding wire. To determine the values of these components for an accurate PD model, physical manufacturing principles and photoelectric conversion were adopted to build the Verilog model, while test results and data from the datasheet were used for parameter fitting.

![Figure 3. Compact equivalent circuit model of PD.](image)

2.3. Free-Space Channel Model and Underwater Channel Model

In the VLC communication system, the most common channel is free space, and the loss of this channel has the aforementioned effects on waveform transmission. For visible light transmission, due to the rectilinear propagation characteristic, this free-space channel can be treated as an LOS channel in which the main restation is signal loss. The gain in the LOS channel can be expressed as below [25]:

$$PL_{LOS} = 10 \log_{10} \left( \frac{A_{PD}(m + 1)}{2 \pi D^2} \cos^m(\alpha) \cos^2(\beta) G_{ct} G_{cr} \right),$$

where $A_{PD}$ is the PD area; $D$ is the distance between the light source and PD; $\alpha$ and $\beta$ are the emission and receiving angles, respectively, with $\cos(\varphi) = \cos(\psi) = h/D$ in which $h$ is equal to $D$ when the PD aligns to the LED; and $m$ is the Lambert emission order. $G_{ct}$ is the gain in the lens at the transmitter that is utilized to increase the intensity by narrowing the half-power angle, while $G_{cr}$ is the gain in the non-imaging concentrator that is applied before PD [25].

Another typical channel for a VLC system is water due to the low loss of visible light in water. For an underwater VLC, path loss consists of attenuation loss and geometrical loss, which should be considered for diffused light sources, such as LEDs [25]. The geometrical loss for LED can be represented by the LOS channel loss since it is caused by spreading during transmission. The attenuation loss is determined by [26]:

$$PL_{AT} = 10 \log_{10} \left( e^{-c(\lambda)D} \right),$$

which is the Beer–Lambert law. $c(\lambda) = a(\lambda) + b(\lambda) + a_T(\lambda)$ [27], which is the extinction coefficient representing the summation of water absorption coefficient $a(\lambda)$, water scattering coefficient $b(\lambda)$, and tank attenuation coefficient $a_T(\lambda)$. The final overall channel loss of the underwater channel is:

$$PL_{Water} = PL_{AT} + PL_{LOS}.$$  

With the path loss equations of both the free-space channel and the underwater channel, the path loss variance with distance of these two types of channels could be simulated. The parameters of the VLC link model that applied to the simulation are summarized in Table 1, and the corresponding curves are presented in Figure 4. For a channel length equal to 3 m, the path loss of the free-space channel was 57.4 dB, while that of the underwater channel was 62.5 dB.
Table 1. Simulation parameters for the VLC link model.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Typical Value</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Transmitter</strong></td>
<td></td>
</tr>
<tr>
<td>Total Power</td>
<td>20 W</td>
</tr>
<tr>
<td>Half-power Angle</td>
<td>50°</td>
</tr>
<tr>
<td>-3 dB Bandwidth</td>
<td>~50 MHz</td>
</tr>
<tr>
<td><strong>Receiver</strong></td>
<td></td>
</tr>
<tr>
<td>PD Area</td>
<td>0.12 mm²</td>
</tr>
<tr>
<td>Photosensitivity</td>
<td>0.51 A/W</td>
</tr>
<tr>
<td>Dark Current</td>
<td>100 pA</td>
</tr>
<tr>
<td>Terminal Capacitance</td>
<td>1.6 pF</td>
</tr>
<tr>
<td>Field-of-view Angle</td>
<td>30°</td>
</tr>
<tr>
<td>-3 dB Bandwidth</td>
<td>1 GHz</td>
</tr>
<tr>
<td><strong>Free-Space Channel</strong></td>
<td></td>
</tr>
<tr>
<td>Link Range</td>
<td>0.1–5 m</td>
</tr>
<tr>
<td>Emission Angle α</td>
<td>0°</td>
</tr>
<tr>
<td>Receiving Angle β</td>
<td>0°</td>
</tr>
<tr>
<td><strong>Underwater Channel</strong></td>
<td></td>
</tr>
<tr>
<td>Link Range</td>
<td>0.1–5 m</td>
</tr>
<tr>
<td>Water Type</td>
<td>Clear sea water</td>
</tr>
<tr>
<td>Absorption Coefficient a(μ) [28]</td>
<td>0.114 m⁻¹</td>
</tr>
<tr>
<td>Scattering Coefficient b(λ) [28]</td>
<td>0.037 m⁻¹</td>
</tr>
<tr>
<td>Tank Attenuation Coefficient aT(λ) [27]</td>
<td>0.265 m⁻¹</td>
</tr>
</tbody>
</table>

Figure 4. Simulated path loss versus distance in free-space channel and water channel.

2.4. Bandwidth Limitation and Nonlinearity

The LED model mentioned in Section 2.1 was utilized for analyzing the effect on the modulation bandwidth in the VLC system. As mentioned in Section 2.1, it could be treated as a first-order low-pass filter in the low-frequency region, as the intrinsic model was the dominant item. Considering the analysis of the impulse response, the expression of this low-pass filter in the s-domain was calculated as below:

\[ c(s) = \frac{A}{1 + \frac{s}{\omega_0}}, \]  

where A is the amplitude of the signal, and \( \omega_0 \) is the characteristic frequency of the low-pass filter. After applying the inverse Laplace transform to Equation (5), the limited bandwidth expression in the time domain was shown as the equation below:

\[ c(t) = \omega_0 A \times e^{-\omega_0 t} \times u(t), \]  

(5)
where $u(t)$ is the unit step function, and $e^{-\omega_0 tf}$ is the time-dependent item complying with exponential attenuation. Therefore, $c(t)$ caused undesired tails leading to ISI in the time domain by adding the time-dependent exponential attenuation to the input signals. This effect of bandwidth limitation on the original input signal is shown in Figure 5, which presents the generated tails spread into adjacent symbols with overlapping [29].

![Figure 5. Simulated path loss versus distance in free-space channel and water channel.](image)

To evaluate the intrinsic nonlinearity of the LED device, the I–V relationship of the LED at forward bias needed to be calculated and was [30]:

$$I_{LED} = (I_n + I_p) \left( e^{\frac{qV}{kT}} - 1 \right),$$  \hspace{1cm} (6)

where $(I_n + I_p)$ is the electrons and holes current, and $e^{\frac{qV}{kT}}$ is a temperature-dependent item that has an exponential relationship with the bias voltage. It can be noticed that the driving current through the LED was nonlinear with a bias voltage of the LED that caused amplitude asymmetry with the DC bias point.

3. Analysis of Equalization Methods

As mentioned in the last section, the main restriction in the transmission data rate in the integrated VLC system was caused by the limited bandwidth. To extend the bandwidth, a single equalization method was not enough, so various equalization methods were combined to corporately contribute to the signal transmission. In the proposed VLC system, both passive equalizer and NN-based FFE algorithms were adopted at the transmitter, while an RBFNN-enabled equalization algorithm was utilized in the receiver.

3.1. Passive Equalizer

The schematic of the adopted passive equalizer is presented in Figure 6, and it was a basic high-pass filter for simplifying the structure of the discrete circuit. When the signal frequency was low, the capacitive reactance was large since $X_C = 1/(2\pi fC)$, while the inductive reactance was small since $X_L = 2\pi fL$. For the high-frequency signal, these two component values changed reversely. Thus, the low-frequency signals were bypassed to the ground through the inductor, while the high-frequency signals were sent to the output of this filter. Therefore, this high-pass filter could decrease the low-frequency gain and increase the high-frequency gain of a frequency response, so that contributed to the modulation bandwidth extension. By properly setting the zeros and poles of this high-pass filter, the high-frequency gain loss could be compensated for at the target frequency. Moreover, the values of $L$ and $C$ determining the zeros and poles of this filter could be calculated with the following:

$$L = \frac{L_0 \times K}{2\pi \times f_{target}},$$  \hspace{1cm} (7)
where $L_0$ is a unique inductor with the value of 1 H, $C_0$ is a unique capacitor equal to 1 F, and $K$ is the impedance ratio of the designed filter to the standard filter, which is normally equal to 50 due to the 50 Ω impedance matching. $f_{\text{target}}$ is the target frequency of compensation through the passive equalizer, which can be treated as the −3 dB bandwidth of the LED without equalizers.

\[
C = \frac{C_0}{K \times 2\pi \times f_{\text{target}}},
\]  

(8)

Figure 6. Schematic of the adopted passive equalizer.

### 3.2. FFE

An FFE is an equalization technology that is commonly adopted in high-speed optical communication transmitter systems since it is easy to facilitate through digital or analog methods and owns a superior performance at eliminating ISI [31]. To generate the FFE signal, the first step was delaying the original signal with 1 or 0.5 UI to generate the required number of taps. Secondly, these delayed signals were inverted and then amplified with the corresponding tap weights to form the pre- or post-tap signals. Finally, all of these taps were added together with the main tap signal, which was the original one. After combination, the desired FFE signal was generated. The expression of the FFE signal was:

\[
y(n) = \sum_{k=0}^{L} \omega_k x(n + k),
\]  

(9)

where $\omega_k$ is the weight of each tap in the FFE, and $x(n + k)$ is the signal for each tap. The FFE contributed to removing the ISI and extending the limited bandwidth since the combined waveforms reshaped the rising and falling edges of the main tap signal and attenuated the low-frequency components in the signal, thus both decreasing the low-frequency gain and increasing the high-frequency gain of the signal frequency response.

The traditional methods of realizing an FFE are either generating modulated signals with an FFE in a digital LUT or combining different taps together through analog circuits. However, these two methods face the problem of low weight precision due to the resolution of DAC or analog circuit mismatch in fabrication. Therefore, a NN-based FFE implementation method was proposed in this work. As shown in Figure 7, this NN-based FFE consisted of two layers of neurons. Input signals with calculated delay times were sent to the input layer towards the corresponding neurons. To train this network, the generated FFE signal was utilized as a training signal, and the weights in this network were adjusted to be the same as the weights of each FFE tap. After the training process, the number of neurons in the input layer with corresponding weights was fixed. There was only one neuron in the output layer to linearly combine the previous data from the input layer and provide a signal modulated with FFE, which was the same as the expression in Equation (9). With this trained network, the test signal could be modulated into an FFE equalization signal. In the proposed VLC transceiver system, a three-tap FFE with one-third UI delay was adopted.
After the training process, the weights in this three-layer DNN were determined, respectively; and $W_m$ was the weight of the output layer. The typical activation function in a DNN is the sigmoid function, and its training target was the originally transmitted signal. After the training process, the weights in this three-layer DNN were determined, and this trained DNN could be utilized as a post-equalizer to compensate for the newly received signals.
The radial basis function utilized the distance between the input and fixed center point to determine the value of the output, and the common expression of this function is:

\[ \phi(x) = \phi(||x - c||), \]

(11)

where the function \( \phi(\cdot) \) is the radial function, the distance is called Euclidean distance, and \( c \) is the radial kernel center. With this definition, the RBF-NN is a neural network that utilizes RBF as the activation function [32]. A conventional RBF-NN only consists of three neuron layers: the input layer, the hidden layer, and the output layer, as shown in Figure 9. The input layer sends input data to the hidden layer, where the activation function is the radial basis function. In this hidden layer, the Euclidean distance between the input data and the kernel center is calculated and utilized to determine the output of the hidden layer. There are various choices for the activation function, such as the Gaussian kernel function, multiquadric function, and inverse quadric function. The output layer linearly combines the output from the hidden layer and provides the final output of the RBF-NN. The weights of this linear layer are also changed during the training process.

![Figure 9. Structure of RBF-NN.](image)

The RBF-NN adopted in the proposed VLC system is presented in Figure 9. The input signals were received from the TIA after down-sampling and synchronization, and then the sampled data were delayed, forming a matrix with a scale equal to the window length of \((L + 1)\) in which the window length represented the delay time of the original signal and was related to the size of the training dataset. Therefore, it could impact the training performance, and a moderate value was required for the tradeoff of training performance and training time. This input matrix was the input of the radial basis function, and the Gaussian kernel function was adopted as the activation function in the hidden layer considering its approximation to the limited bandwidth effect in the time domain, as analyzed in Section 2. The output layer was a linear combination of outputs from the hidden layer. The training process of this RBF-NN utilized the original PRBS-9 signal as the training target, and then adjusted the weights in the hidden layer and the output layer. The final output of the RBF-NN was [32]:

\[ y(n) = f(\sum_{i=0}^{1} \omega_i \phi_i(x_i) + b), \]

(12)

where \( b \) is the extra bias item, and \( \phi(x) \) is the Gaussian kernel function with the expression:

\[ \phi(x) = e^{-\frac{||x-c||^2}{2\sigma^2}}. \]

(13)

The RBF-NN had several advantages compared with a traditional DNN for its unique structure and training process, as summarized in Table 2. The first advantage was that the activation function of the RBF-NN was more suitable to approximate the bandwidth limitation and nonlinear effects. Due to the flexible choice of various activation functions,
the RBF-NN could compensate for channel models with nonlinear characteristics, especially when using the Gaussian kernel function as its radial basis function. The second advantage was that the RBF-NN was a local approximation neural network due to the impact of one neuron on another one being related to their Euclidean distance, and it only required changing part of the weights during the training process instead of changing all the weights for each iteration. Traditional neural networks, such as DNNs and CNNs, utilize the latter method for the training process and consume many more resources and take a longer time. The third advantage was that the RBF-NN owned a simpler neural network structure since it only contained three layers compared with other NNs with several hidden layers. This simple structure also contributed to decreasing the training time.

Table 2. Comparison of RBF-NN and DNN.

<table>
<thead>
<tr>
<th>Type of NN</th>
<th>Activation Function</th>
<th>Approximation Approach</th>
<th>Neuron Layers</th>
</tr>
</thead>
<tbody>
<tr>
<td>RBF-NN</td>
<td>Gaussian Kernel Function</td>
<td>Local Approximation</td>
<td>3</td>
</tr>
<tr>
<td>DNN</td>
<td>Sigmoid Function</td>
<td>Universal Approximation</td>
<td>&gt;3</td>
</tr>
</tbody>
</table>

4. Architecture of VLC Transceiver Circuits

The architecture of the proposed VLC transceiver system is presented in Figure 10, and it includes a digital baseband, which was implemented in MATLAB, and an AFE, which was designed in Cadence. The digital baseband in the transmitter aimed to generate a transmitted signal and to encode and modulate this signal by equipping the NN-based FFE, while the baseband in the receiver decoded and demodulated the received signal with the RBF-NN-based equalizer. The AFE in Cadence included a LED driver circuit, a passive equalizer, and a TIA. In addition, the models of the LED, PD, and channels that were built in Section 2 were implemented in Cadence as the VLC link for analyzing the system performance. To verify the design of the proposed system, a co-simulation platform comprising MATLAB and Cadence was utilized.

![Figure 10. Architecture of proposed VLC system with NN-based equalizations.](image-url)

4.1. Transmitter in VLC System

The digital baseband in the proposed VLC transmitter consisted of a pseudo-random binary sequence (PRBS) waveform generator, a PAM-8 signal modulator, and an NN-based pre-equalizer using a FFE algorithm, and a DAC generated the analog PAM-8 signal with the FFE. A PRBS-9 signal was generated and encoded, and then it was sent to the PAM-8 modulator. An NN-based equalizer following the modulator added an FFE algorithm to compensate for the limited bandwidth. After this equalizer, the PAM-8 FFE signal was upsampled and transmitted to the AFE in the form of .dat-formatted data.
The core circuit of the AFE on the transmitter side presented in Figure 10 was the integrated LED driver, and its structure is presented in Figure 11. It consisted of four transistors: M1 and M2 were the differential pair, while M3 was the tail current source and cooperated with M4 to form a current mirror [33]. The differential pair worked in the saturation region for driving the LED with the required output swing. The differential input voltage signals generated from the digital baseband in the transmitter were sent to the gate nodes of M1 and M2, and then the differential outputs were elicited from the drain nodes. The gain in $V_{\text{out}}$ to $V_{\text{in}}$ is equal to $g_m(R_1/r_o)$, where $g_m$ is the transconductance of the transistor, and $r_o$ is the output resistance of M1 or M2. The current mirror composed of M3 and M4 was utilized to provide a tail current to the differential pair. Since referred current $I_{\text{ref}}$ from outside flowing through M4 was smaller than the required tail current, a current mirror that could provide a current with the ratio of the transistor size was necessary. The value of the tail current through M3 was $I_{\text{ref}}(W_3L_3/W_4L_4)$, where $W_3L_3/W_4L_4$ is the transistor size ratio of M3 and M4. The whole CML circuit owned several advantages that made it suitable for a high-speed communication circuit. It owned a simple structure, and the differential pair guaranteed low common mode noise. In addition, the low switching noise decreased the rising and falling times, which improved the switching speeds of M1 and M2. The large output impedance contributed to the low power consumption.

![Schematic of CML LED driver.](image)

The passive equalizers shown in Figure 6 were placed after the CML LED driver to extend the modulation bandwidth by serving as a high-pass filter. The model of LED used in this system owned a limited bandwidth of 50 MHz, and it was based on the compact equivalent model of the $\mu$-LED introduced in Section 2. This $\mu$-LED model was also implemented in Cadence with fixed RLC components and a Verilog-A-code-based current-dependent model. For the free-space and underwater channels, their models are illustrated in Section 2, and the channel loss at the alignment occasion of the LED and PD was adopted in this VLC system as the total loss.

4.2. Receiver in VLC System

On the receiver side, the optical signal was received by the PD and converted into the photo-electronic current, and then this current was transformed into voltage and amplified by the TIA. The following ADC sampled the TIA output and saved it as .csv-formatted data. In the MATLAB offline baseband, this received signal was synchronized first, and then an NN-based post-equalizer compensated for this signal, followed by PAM-8 demodulation.

For the AFE in the VLC receiver shown in Figure 10, the TIA was the core circuit, and its schematic is presented in Figure 12. The TIA unit converted the photoelectronic current generated by the PD to a voltage output through feedback resistor R1. The output current of the PD was as small, from several to dozens of uA, to guarantee the subsequent signal process, and it required a conversion of the signal from current to voltage with large
amplification, which could be satisfied by the TIA unit. Since the signal conversion in the TIA was processed by the feedback resistor following equation $V_{\text{out}} = R_1 \times I_{\text{in}}$, the gain of the TIA was determined by the value of the feedback resistor when the stability of the core amplifier was ensured. Due to the requirement of SNR in receiving the signal, the noise of the TIA should be as low as possible to avoid bringing large noise into the receiver [34]. The noise in the TIA unit consisted of the thermal noise of the TIA and the equivalent input thermal noise of the MOSFET. These two types of noises were inversely proportional to the transimpedance of the TIA. Taking the requirement of both high gain and small noise into consideration, the realization of the high-sensitivity receiver required a large feedback resistor. Meanwhile, the first-stage amplifier should own a large transimpedance to further increase the performance of the TIA. Moreover, the three-stage structure adopted in this TIA provided a stable high gain due to the alleviation of bandwidth limitation brought by the low noise requirement.

Figure 12. Schematic of TIA circuit.

5. Simulation Results and Discussion

To evaluate the performance of the proposed PAM-8 VLC system, a co-simulation platform of MATLAB and Cadence was adopted. This platform was based on the VLC system structure presented in Figure 10, and it combined the digital baseband in MATLAB with the AFE in Cadence to incorporate the advantages of these two software design tools. The assessment of the proposed VLC system consisted of a transceiver system performance evaluation, a data transmission evaluation, and a training performance evaluation. In addition, to prove the superiority of the RBF-NN, a comparison with a DNN as a post-equalization method was also performed.

5.1. Pre-equalization Performance Evaluation

The bandwidth of the AFE in the transmitter was analyzed to evaluate the performance of the passive equalizer. Figure 13a shows the frequency response of the designed passive equalizer. This passive equalizer served as a high-pass filter to extend the limited modulation bandwidth caused by the LED. As presented in Figure 13b, the $-3$ dB bandwidth of the transmitter including limitation from the LED was 52.42 MHz, and it could be extended to 313.39 MHz after adopting the passive equalizer with the sacrifice of low-frequency component loss. The corresponding eye diagrams of 2.7 Gbps and 3.6 Gbps transmissions without and with the NN-based FFE were detected at the output of the LED to verify the feasibility of the FFE. The first pre-tap weight and first post-tap weight for the 2.7 Gbps data transmissions were 0.05 and 0.35, respectively, while the values for 3.6 Gbps were 0.07 and 0.5. Figure 14a,c are the eye diagrams acquired only with the passive equalizer under the data rates of 2.7 Gbps and 3.6 Gbps separately. Meanwhile, Figure 14b,d are the eye diagrams acquired with both the passive equalizer and FFE under the data rates of 2.7 Gbps and 3.6 Gbps. When comparing Figure 14a to Figure 14b, it can be
noticed that the eye quality increased dramatically with the equalization of the FFE, which proved the feasibility of the proposed NN-based FFE. However, when the data rate was increased from 2.7 Gbps to 3.6 Gbps, the top and bottom eyes diagrams of Figure 14d were almost closed. Nevertheless, the case of the data rate equal to 3.6 Gbps also suffered from nonlinearity caused by the LED. After transmission through the channels, the quality of the eye diagrams was further decreased. Therefore, these problems of bandwidth limitations and path loss remained to be solved by the RBF-NN-enabled post-equalization.

Figure 13. (a) Frequency response of passive equalizer; (b) −3 dB bandwidth extension of transmitter using passive equalizer.

Figure 14. Eye diagrams of PAM-8 signal transmissions with passive equalizers at LED output: (a) 2.7 Gbps data rate without FFE; (b) 2.7 Gbps data rate with FFE; (c) 3.6 Gbps data rate without FFE; (d) 3.6 Gbps data rate with FFE.
5.2. Post-Equalization Performance Evaluation

To evaluate the post-equalization performance, the bandwidth of the AFE on the receiver side was analyzed first. The $-3$ dB bandwidth of the AFE in the receiver, including the PD model and the corresponding total gain, were simulated as shown in Figure 15. It can be noticed that the total bandwidth was 906.8 MHz, with a TIA gain of 73.9 dB.

![Figure 15. $-3$ dB bandwidth and gain of receiver AFE including PD.](image)

The BER is one of the most important evaluation criteria of data transmission quality, and the relationship of the BER with the data rate was studied to find the maximum data rate that the VLC system could support with the BER criteria of $3.8 \times 10^{-3}$. The data transmission quality of using the RBF-NN as the post-equalizer, using a DNN, and using a traditional equalization scheme of FFE were compared under a 3 m free-space channel, and the results are presented in Figure 16a. The highest data rate of using the RBF-NN as a post-equalizer was 3.6 Gbps, while the highest data rates that could be achieved with the DNN equalizer or FFE were 3.3 Gbps and 2.7 Gbps, respectively. By utilizing the RBF-NN as the equalization scheme, the communication data rate could be improved by 20% when compared with the DNN due to the more suitable activation function. In addition, the highest data rate that this system could support with different types of channels was also studied. As shown in Figure 16b, the highest achievable data rate for a 1.2 m water channel was 4.2 Gbps, which was higher than the 3 m free-space channel due to less channel loss with the RBF-NN.

![Figure 16. Simulated results of BER vs. data rate: (a) using RBF-NN, DNN, or FFE as a post-equalizer with 3 m free-space channel; (b) 3 m free-space and 1.2 m water channel with RBF-NN.](image)
The simulated received eye diagrams using the RBF-NN as a post-equalization scheme at two different data rates with two types of channels are presented in Figure 17. Figure 17a,b shows the eye diagrams acquired with a 3 m free-space channel model utilized as the VLC channel under the data rates of 2.7 Gbps and 3.6 Gbps, respectively. Comparing Figure 17a to Figure 17b, it can be noticed that the eye quality decreased with the increase in data rate. Figure 17a shows an eye diagram acquired with a BER of $10^{-5}$, while Figure 17b is the case a BER equal to $3.8 \times 10^{-3}$. These two figures also present the case of error-free and the case of maximum BER tolerated for signal transmission. Meanwhile, Figure 17c,d displays the eye diagrams acquired with a 1.2 m underwater channel model under the data rates of 3.3 Gbps and 4.2 Gbps. According to Figure 4, the path loss of the 3 m free-space channel was around 5 dB larger than the 1.2 m underwater channel. Therefore, by comparing the upper two figures with the lower two figures, it can be concluded that, under the same BER criteria, the communication data rate increased when the path loss decreased. In addition, when taking Figure 14b,d into comparison, the feasibility of the RBF-NN enabled post-equalization could be proved. Since Figure 14b,d represent the eye diagrams acquired with the FFE and passive equalizers on the transmitter side, the upper two eye diagrams are the results of transmission through the 3 m free-space channel and processing by the receiver. It can be noticed that the eye quality was hugely increased when comparing Figure 14b,d.

**Figure 17.** Simulated eye diagrams of PAM-8 signal with RBF-NN as post-equalizer: (a) communication data rate of 2.7 Gbps with 3 m free-space channel; (b) communication data rate of 3.6 Gbps with 3 m free-space channel; (c) communication data rate of 3.3 Gbps with 1.2 m underwater channel; (d) communication data rate of 4.2 Gbps with 1.2 m underwater channel.
5.3. Training Performance Evaluation

For the RBF-NN adopted in the proposed VLC transceiver system, evaluating its training performance was essential to prove its feasibility and superiority compared with traditional DNNs. In addition, the training time of the RBF-NN is critical for implementing it through a hardware design for real-time signal processing. Long training time causes extra delay time for signal transmission and even leads to the failure of compensating for signal loss and distortions. Therefore, decreasing the training time is important for building a real-time signal-processing baseband.

The convergence curves of training the RBF-NN with the 2.7 Gbps and 3.6 Gbps datasets are presented in Figure 18. It can be noticed that the trained model fixed well with the expectation and owned an error less than $10^{-8}$ after 1000 epochs, which proved the feasibility of choosing the RBF-NN as the post-equalizer. The training times of different types of NNs vary since the complexity of their structures and training methods determine the time cost. Taking into consideration the implementation of NNs to systems-on-chip as a real-time equalization method, the training time should be as short as possible. The training times of both the RBF-NN and a traditional DNN at different communication data rates were simulated, and the simulated results are presented in Figure 19. It is obvious that, with an increased data rate, the training time for the NN also increased. This proportional relationship reflects the fact that the training resources were dramatically consumed when the transmitted signal deviated hugely from the original. Moreover, compared with the traditional DNN, the RBF-NN consumed much less training time due to its superior characteristic of local approximation. With the same data rate achieved, the training time of the equalization scheme utilizing the RBF-NN could be reduced by 86.7%. Since the RBF-NN was a local approximate network, only part of the weights of all the neurons needed to be trained during each iteration, which could speed up the training process of the RBF-NN. The DNN required a much longer training time since it was a global approximation network, which meant that the weights of all the neurons should be adjusted during each iteration. Therefore, utilizing the RBF-NN contributed to saving computing resources.

![Performance is 1.99474e-09, Goal is 0](image1)
![Performance is 3.36365e-09, Goal is 0](image2)

**Figure 18.** Training convergence curves of RBF-NN: (a) 2.7 Gbps data transmission; (b) 3.6 Gbps data transmission.
Specifications of VLC System 

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<th>Specfications of VLC System</th>
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Figure 19. Simulated results of training time vs. data rate with RBFNN or DNN as equalizer.

5.4. Comparison of Other VLC Systems with Neural Network

To evaluate the performance of the proposed PAM-8 transceiver system with neural-network-enabled pre- and post-equalization schemes, the system architecture, training performances, and data transmission of the proposed VLC system were compared with those of previous works, as well as summarizing the characteristics of these works in Table 3. It can be noticed that our proposed VLC transceiver system could achieve the highest data rate with the longest channel length since a hybrid equalization scheme consisting of the passive equalizer, NN-based FFE, and RBF-NN-enabled post-equalization was adopted. In addition, the highest modulation bandwidth in the receiver also contributed to achieving this data rate. The AFEs of our proposed VLC transceiver system were implemented with integrated circuits, which further improved the performance of the system. Our proposed system could also be adopted to both free-space and underwater channels for communication.

Table 3. Comparison of neural-network-enabled VLC systems.
6. Conclusions

A PAM-8 VLC transceiver system composed of digital basebands, AFEs, an LED model, a PD model, and channel models was proposed and implemented. The limitations in VLC systems caused by the intrinsic characteristics of LEDs and channels were systemically studied and could lead to bandwidth limitation, such as ISI, and path loss. Based on the analysis of ISI and signal loss, a VLC transceiver system that combined an NN-based FFE, passive equalizers, and an RBF-NN-enabled post-equalizer to extend the limited bandwidth was proposed. The whole VLC system was implemented in a co-simulation platform consisting of MATLAB and Cadence. MATLAB was utilized to build these digital basebands, the DAC, and the ADC. The designs of the AFEs with the LED and PD device models were implemented in Cadence. In addition, both free-space and underwater channel models were established and utilized for the simulation to prove that the proposed system could work with different channel models. The data transmission performances of using the free-space model and the underwater channel model were also analyzed and compared. To verify the feasibility and evaluate the performance of the VLC system, simulations on the data rates of signal transmission versus the BER with the RBF-NN or a DNN as post-equalization schemes were conducted, and the training time was also studied. The simulation results indicated that the RBF-NN-enabled VLC transceiver system could achieve a data rate of 3.6 Gbps while taking less than 10 min of training time.

Author Contributions: Conceptualization, C.P.Y. and B.X.; methodology, B.X.; validation, B.X. and T.M.; writing—original draft preparation, B.X.; writing—review and editing, C.P.Y. All authors have read and agreed to the published version of the manuscript.

Funding: This work was funded by the Foshan-HKUST Project under the Government of Foshan Municipal City (grant number FSUST20-SHCIRI05C); the Project of Hetao Shenzhen–Hong Kong Science and Technology Innovation Cooperation Zone (grant number HZQ8-KCZYB-2020083); the Hong Kong Research Grants Council under the General Research Fund (GRF) project (grant number 16215620); and the HKUST-Qualcomm Joint Innovation and Research Laboratory.

Data Availability Statement: Not applicable.

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

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