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Zabih Ghassemlooy, Luis Nero Alves, Stanislav Zvánovec, Mohammad-Ali Khalighi

Channel Modeling

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Channel Modeling

Zabih Ghassemlooy, Mohammad-Ali Khalighi, and Dehao Wu

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3.1 Introduction

An important step in the design of a visible light communications (VLC) system is to comprehend the limitations arising from the optical wireless channel. Accurate channel characterization is an important prerequisite to set the system parameters appropriately in order to establish a high-quality link since it permits better exploitation of the available energy and spectral resources in view of optimizing the system design. An accurate channel model is also necessary to precisely predict the performance of VLC systems.
In this chapter, we address channel modeling for VLC systems mainly focusing on indoor systems. We introduce different sources of impairment in VLC systems arising from beam propagation or transmitter (Tx)/receiver (Rx) devices. Indeed, the latter could be attributed to the “global channel” comprising the blocks between the signal Tx and signal detection at the Rx.

After this introduction, we give an overview of different propagation modes in Section 3.2. In addition, we explain methods for numerical channel simulation. Analytical channel modeling for the cases of single- and multiple-source systems is presented in Section 3.3. Then, in Section 3.4, we outline the limitations arising from the aggregate channel while focusing on the problem of intersymbol interference (ISI) and how it affects the link performance particularly in the absence of a line of sight (LOS). This could be due to multipath-induced channel time dispersion, that is, multiple reflections from people and objects within an indoor environment, the lack of sufficient bandwidth of the transmitting device, most commonly a light-emitting diode (LED), the photodetector (PD), and the cabling used for lighting installations. Limitations due to the LED nonlinear characteristics could also be considered as an impairment of the global channel, which is the subject of Section 3.5 where channel distortion modeling is investigated. Finally, channel modeling for multiple-input multiple-output (MIMO) VLC systems will be presented next in Section 3.6.

3.2 Signal Propagation

3.2.1 Propagation Modes

For indoor links, six different configurations have been defined in [1], basically classified depending on the existence/nonexistence of the LOS path between the Tx and Rx. Here we consider some of these configurations that apply to the case of indoor VLCs. For LOS configuration, which is the most basic, the emitter beam angle and the receiver field-of-view (FOV) will specify the transmission channel. For the case of directive links, the Tx and Rx have a small divergence angle and FOV, respectively (see Figure 3.1a), thus requiring very accurate alignment and suffering from blocking due to the movement of people or presence of objects within the room. For the so-called hybrid links, Tx and Rx have different degrees of directionality [1]. In non-directive links, Tx and Rx both have a wide angle—see Figure 3.1b. In the case of the diffuse configuration, see Figure 3.1c (this may be one of the most popular and widely used schemes); the source position plays an important role in the power levels at various points within a room. In this configuration, the Tx pointing up toward the ceiling has a wide beam angle and the Rx has a wide FOV, which collects reflected diffused light from the ceiling, floor, walls, and objects in the room [2]. In general, to establish high data rate links,
the availability of an LOS path is essential since nondirected LOS or diffused configurations will limit the achievable data rate [3]. Indeed, the LOS helps in having a much higher received light intensity (i.e., higher signal-to-noise ratio (SNR) that can be traded to increase the data rate or the link span) and also reduces the risk of ISI, as will be discussed later in Section 3.4, but at the cost of limited mobility or possibility of shadowing/beam blockage.

When considering mobility within an indoor environment, it is essential that a tracking capability is adopted for alignment of the Tx and Rx [4]. Assuming that the Tx is fixed and placed on the ceiling, there are three possible configurations: full-tracking (FT), half-tracking (HT), or nontracking (NT). For the FT link, shown in Figure 3.2a, tracking mechanisms at the Tx and Rx permit the alignment of the Tx and Rx of small apertures, and hence, provide a relatively high SNR at the Rx. For the HT configuration, see Figure 3.2b; only the Tx (or the Rx) tracks the Rx (Tx), thus allowing the use of a lower complexity tracking mechanism, which is more applicable for multiuser systems. For the NT system, see Figure 3.2c; the orientations of the Tx and Rx are fixed and vertical to each other, which permits the implementation of a low-cost system.
In what follows, we will not consider the mobility but focus on the case of nondirected LOS (Figure 3.1b) as the default configuration since it corresponds to a typical indoor VLC system. The non-LOS configuration overcomes the blocking problem by using multiple diffuse reflections from walls and ceiling. In this configuration, the Rx will receive signals from a number of paths, thus ensuring 100% link availability at all times but at the cost of reduced data due to multipath induced ISI. However, it offers much flexibility and ease in the link setup, and hence, it is suitable for point-to-multipoint applications. Note that in general for optical wireless communication (OWC) systems, ISI depends on the data rate and FOV of the Tx and Rx. In VLC systems, the Tx typically has a wide angle of irradiance for the function of lighting in most applications, except the multispot lighting where a Tx with a narrow irradiance angle is normally used.

3.2.2 Channel Simulation

The ideal LOS channel impulse response (CIR) is essentially a time delayed and scaled delta function representing amplitude degradation of the transmitted signal. Therefore, the link attenuation becomes an important parameter that can be derived from the photometric parameters mostly adopted for characterization of LED illumination capability. For the purposes of performance evaluation of VLC systems, one may resort to experimental measurements [5–7] or to numerical simulation of the propagation channel. The second approach is faster and less costly but it can also be helpful prior to the experimental verifications. The propagation channel is fully characterized by the CIR. Whatever the link configuration, the CIR needs to be determined at several points within the indoor environment. Then, the necessity of an accurate and computationally efficient simulation method is obvious. The most critical point in the simulation is the reflections from walls and other objects within the room, which will take time depending of course on the number of reflections taken into consideration.

The reflection characteristics of the surface within an indoor environment depend on a number of factors including the material, operating wavelength of the light source, and the angle of irradiance. The smoothness or roughness of the surface relative to the wavelength will also affect the shape of the reflected patterns. A smooth surface like a mirror or a shiny object reflects the incident beam only in one well-defined direction (i.e., specular reflection), whereas a rough surface reflects the beam in random directions (i.e., diffuse reflection). In practice, most reflections are typically diffuse in nature where the Lambertian model can appropriately be adopted [2,8]. Several measurement campaigns have validated the basic diffuse reflection model, illustrating the importance of the orientation of Tx and Rx as well as the importance of shadowing. There have been several proposed approaches to simulate the diffuse light components. In [2], it was proposed to decompose the room surface into a number of reflecting elements, which scatter the light according to the
Lambertian model, and to sum the reflected light from all these elements at the Rx. However, only single reflections were considered in this method. An extension of the approach of [2] to account for higher-order reflections was proposed in [9] using a recursive algorithm. This method has been widely used in the literature; however, it suffers from high computational complexity. In fact, in order to improve the reliability of the calculated CIR, the number of reflections taken into account should be increased, but this will result in an exponential increase in the computing time [10]. On the other hand, if the number of considered reflections is not sufficient, then the path loss and channel bandwidth will be overestimated [10].

A statistical Monte Carlo approach based on the ray-tracing scheme was proposed in [11] where the ray directions are randomly generated according to the emitter radiation pattern. The trajectories of emitted photons from the Tx are then calculated until they are lost or received in the PD active area. Using a large number of generated photons, the CIR is then estimated. In a more recent work, a lower complexity ray-tracing method was proposed [12]. An iterative method was also presented in [13] that is much faster than the recursive approach of [9]. Using this method, the one-order reflection of tiny surfaces is first calculated, and then used to determine the higher-order reflections in an iterative manner. A so-called integrating sphere method was also proposed in [10] to obtain an analytical channel model that is used to approximate the CIR.

However, most of these works consider the infrared (IR) transmission rather than VLC. The main difference is that for IR communications, a narrowband near-monochromatic IR light source is used and a constant reflectance is considered for the reflectors, independent of the wavelength [14]. It is worth mentioning that most VLC systems use phosphor-based white LEDs with emission in the visible spectrum (from 380 nm to 780 nm) for the reasons of reduced fabrication complexity and cost, compared to red, green, and blue (RGB)-based white LEDs, where all three colors are transmitted simultaneously. However, the bandwidth of these LEDs is very limited (typically several MHz), which is due to the slow time constant of the phosphor. An alternative approach is to use a narrowband blue filter at the receiver in order to filter out the slow yellow component to improve the modulation bandwidth of the LED [15–17]. In this case, most of the results of the work done in the IR band can be exploited as we effectively work in the narrowband. Meanwhile, it should be noticed that the reflectivity of the IR band is higher than that of the visible band [18]. For instance, VLC channel characterization based on the recursive algorithm of [8] is studied in [15,19] where constant reflectance was considered. If the white light is directly used for signal transmission and detection, then we need to modify the channel model to take into account the wavelength-dependent nature of reflectors. This is investigated in [14] where the wideband nature and power spectral distribution of the visible light source are taken into consideration. Indeed, if the number of reflections considered in the simulations is not too high, as a good approximation, the average reflectivity over the entire visible spectrum can be used [14,20].
3.3 Channel Model

3.3.1 Illuminance of LEDs

The illuminance intensity is normally adopted to define the brightness of an LED or an illuminated surface, assuming that the source has a Lambertian radiation pattern, which is given in terms of the spatial angle $\Omega$ and the luminous flux $\Phi$ as:

$$I = d\Phi/d\Omega.$$  \hfill (3.1)

Consequently the transmitted (emitted) power $P_E$ is defined as:

$$P_E = \int_{\Lambda_{\text{min}}}^{\Lambda_{\text{max}}} \int_{0}^{2\pi} \Phi_e d\phi d\lambda,$$  \hfill (3.2)

where $\Lambda_{\text{min}}$ and $\Lambda_{\text{min}}$ are defined by the sensitivity plots of PD, $\lambda$ is the wavelength, $\phi$ is the incident angle, and $\Phi_e$ is the energy flux.

The luminous intensity when illuminating a surface defined in terms of the angle of irradiance $\phi$ is given as [1]:

$$I(\phi) = \frac{m+1}{2\pi} I(0) \cos^m(\phi), \quad \phi \in \left[-\frac{\pi}{2}, \frac{\pi}{2}\right],$$  \hfill (3.3)

where $I(0)$ represents the center luminous intensity of the LED Tx and $m$ indicates its Lambertian order, given by [21]:

$$m = \frac{-\ln(2)}{\ln(\cos(\Phi_{1/2}))},$$  \hfill (3.4)

where $\Phi_{1/2}$ is the semiangle at half illuminance of the Tx. $I(\phi)$ can also be written in terms of the incident power:

$$I(\phi) = \rho \frac{m+1}{2\pi} P_E \cos^m(\phi),$$  \hfill (3.5)

where $\rho$ is the surface reflection coefficient.

3.3.2 General Transmission Link Model

Like in most OWC systems, intensity modulation with direct detection (IM/DD) is used in most VLC systems [8] for the reasons of reduced cost and implementation complexity. This way, the intensity of the LED, $x(t)$, is modulated by the input signal. Denoting the photocurrent generated
by the PD at the receiver by \( y(t) \), the baseband equivalent of the optical link is described by (see Figure 3.3):

\[
y(t) = Rx(t) \otimes h(t) + n(t),
\]

where \( R \) is the PD responsivity, \( h(t) \) is the baseband CIR, \( \otimes \) denotes convolution, and \( n(t) \) is the additive white Gaussian noise. Note that as it represents optical intensity, \( x(t) \) is nonnegative.

The receiver noise \( n(t) \) is mainly due to the ambient light and in the form of shot noise. The main sources of ambient noise are sunlight and artificial light such as that of incandescent and fluorescent lamps [22,23]. The power spectral density of different ambient light sources and that of the corresponding electrical signals can be found in [1,24]. During daytime, sunlight through windows is typically stronger than the other two sources. In addition, if LEDs are exclusively used for indoor lighting, we are only concerned by the sunlight. Otherwise, to reduce the interference from fluorescent lighting, for example, discrete multitone techniques (DMT) can be used [15] (see Chapter 4). The produced shot noise due to ambient light can degrade the performance of the VLC system. Note that in the case where the blue light is used at the Rx for signal detection by narrow spectral filtering, the influence of ambient light is considerably reduced. If the ambient light is negligible, then the dominant noise source is the Rx preamplifier thermal noise.

3.3.3 Channel Model for Single Source Case

Let us focus on the CIR \( h(t) \). If for the sake of simplicity we neglect the diffuse propagation component, that is, consider only the LOS path, the received intensity will depend on the emitter radiation pattern, the receiver optics, and the PD active area. Denoting the emitted optical intensity by \( P_E \), the received optical power \( P_R \) is given by:

\[
P_R = H(0)P_E,
\]

FIGURE 3.3
Baseband-equivalent model of the optical link with IM/DD.
where $H(0)$ is the channel DC gain given by:

$$H(0) = \int_{-\infty}^{\infty} h(t) dt.$$  \hfill (3.8)

If we model the emitter by a generalized Lambertian pattern, we have [25]:

$$H(0) = \begin{cases} \frac{(m + 1) \text{APD}}{2 \pi d^2} \cos^m(\phi) T_s(\phi) g(\phi) \cos(\phi), & 0 \leq \phi \leq \phi_c, \\ 0, & 0 \geq \phi_c \end{cases},$$  \hfill (3.9)

where $\text{APD}$ is the PD surface area, $\phi_c$ is the Rx FOV (semiangle), and $d$ is the distance from LEDs to the Rx point. Also, $T_s(\phi)$ is the optical filter gain, and the optical concentrator gain $g(\phi)$ is defined as [8]:

$$g(\phi) = \begin{cases} \frac{n^2}{\sin^2 \phi_c}, & 0 \leq \phi \leq \phi_c, \\ 0, & 0 \geq \phi_c \end{cases}.$$  \hfill (3.10)

where $n$ is the concentrator refractive index.

Consider Figure 3.4 that shows the geometry of the optical Tx, Rx, and surface reflectors for a typical indoor VLC system. The illuminance at a given point on the receiving plane is given by $I(\phi) \cos(\phi)/d^2$ [26].

FIGURE 3.4
Geometry of optical Tx, Rx, and reflectors.
Considering power due to the non-LOS paths, the DC channel gain of the reflected path is given by [2]:

\[
d_{H_{ref}}(0) = \frac{(m+1)A_{PD}}{2\pi d_{1}^2 d_{2}^2} \rho d_{wall} \cos^{m}(\phi) \cos(\alpha) \cos(\beta) T_{s}(\phi) g(\phi) \cos(\varphi), \quad 0 \leq \varphi \leq \varphi_{c},
\]

\[
0, \quad \varphi \geq \varphi_{c}
\]

(3.11)

where \(\beta\) represents the angle of irradiance from the reflective area of the wall, \(\alpha\) is the angle of irradiance to the wall, \(d_{1}\) and \(d_{2}\) are the distances between the Tx and the wall, and the wall and a point on the receiving surface, respectively, and \(d_{wall}\) is the size of the reflective area.

Now if we consider both multipath propagation and the LOS component, for the more general case, the total received power \(P_{R}\) is given by [27]:

\[
P_{R} = P_{E} H_{LOS}(0) + \int_{walls} P_{E} d_{H_{ref}}(0).
\]

(3.12)

Note that the electrical SNR, which expresses the quality of transmission, can be defined in terms of \(P_{R}\) as:

\[
\text{SNR}_{ele} = \left( \frac{RH(0) P_{R}}{\sigma_{T}^{2}} \right)^{2},
\]

(3.13)

where \(\sigma_{T}^{2}\) is the total noise variance.

The transmitted intensity of a single light beam undergoes a number of \(k\) reflections (bounces) prior to collection at the Rx, which is described by the CIR as [28]:

\[
h(t; T_{j}, R_{i}) = \sum_{k=0}^{\infty} h_{S}^{(k)}(t; T_{j}, R_{i}),
\]

(3.14)

where \(h_{S}^{(k)}(t; T_{j}, R_{i})\) is the impulse response corresponding to the \(k\)th reflection. The LOS contribution to CIR is given in terms of the delayed Dirac delta function as:

\[
h_{S}^{0}(t; T_{j}, R_{i}) = V I(\phi_{ij}) \left( \frac{A_{R_{i}} g(\phi)}{d_{ij}^{2}} \right) \times \delta(t - d_{ij}/c),
\]

(3.15)

where \(V\) is the visibility factor \(0 < V \leq 1\), with \(V = 1\) representing unobstructed LOS path, \(c\) is the speed of light, \(d_{ij}\) is the distance between the Tx and the Rx, and \(A_{R_{i}}\) is the optical collection area. Also, \(g(\phi)\) is the Rx optical gain function, defined as follows:

\[
g(\phi) = \begin{cases} 
\cos(\phi) & \text{if } 0 \leq \phi \leq \pi/2 \\
0 & \text{otherwise}
\end{cases}.
\]

(3.16)
Similarly, the $k$-bounce response can be calculated using the $(k - 1)$-bounce response, which is given by Carruthers et al. in [29]:

$$h^{(k)}_S(t; T_j, R_i) = \int_S \rho \, d\varepsilon' \cdot h^{(k-1)}_S(t; T_j, d\varepsilon') \otimes h^0_S(t; d\varepsilon', R_i),$$  \hspace{1cm} (3.17)

where the integral is over the surfaces in $S$, $d\varepsilon'$ and $d\varepsilon'$ represent a differential surface of area $dr^2$, where the first one acts as Rx with respect to $T_j$ and then as a source with respect to $R_i$.

Note as $k \to \infty$, $\|h^{(k)}_S(t; T_j, d\varepsilon')\| \to 0$ since $\rho < 1$ everywhere; then we can estimate the overall CIR for a number of $N$-bounce as:

$$h_S(t; T_j, R_i) \approx \sum_{k=0}^{N} h^{(k)}_S(t; T_j, R_i).$$  \hspace{1cm} (3.18)

A good approximation can be achieved for $3 < N < 10$, as outlined in [29].

### 3.3.4 Channel Model for Multiple Sources

Let us consider a general VLC channel with $M$ light sources (or $M$ small elements per each facet) and multiple propagation with $N$ non-LOS paths between a Tx and an Rx. The general link geometry is shown in Figure 3.5. The Rx $R_j$ receives radiation emitted from multiple sources including $T_i$.
via the LOS, as well as from k-number of reflections from walls, ceiling, and floor within the room.

For the indoor VLC configuration as shown in Figure 3.5, assuming that \( T_i \) emits a unit impulse at \( t = 0 \) and normalizing \( P_E \) to 1, the LOS (\( k = 0 \)) CIR for a particular source \( T_i \) and a detector \( R_j \), is given by [30]:

\[
h_S^0(\tau; T_i, R_j) = \frac{I(\phi_{ij})A_{R_j}}{d_{ij}^2} T_s(\varphi_{ij}) g(\varphi_{ij}) \cos(\varphi_{ij}) \text{rect} \left( \frac{\varphi_{ij}}{\varphi_c} \right) \delta \left( t - \frac{d_{ij}}{c} \right),
\]

(3.19)

where \( d_{ij} \) is the distance between \( T_i \) and \( R_j \), and \( \delta(\cdot) \) is the Dirac delta function. Also, \( \text{rect}(x) \) stands for a rectangular function defined as:

\[
\text{rect}(x) = \begin{cases} 
1 & \text{for } |x| \leq 1, \\
0 & \text{for } |x| > 1.
\end{cases}
\]

(3.20)

The CIR for \( k \)-bounce (\( k \leq 1 \)) is given by [29]:

\[
h_S^k(\tau; T_i, R_j) = \sum_{n=1}^{M} \rho \left( d_{en}^t \right) \cdot h_S^{k-1}(\tau; T_i, d_{en}^t) \otimes h_S^0(\tau; d_{en}^t, R_j).
\]

(3.21)

Following the methodology described in [29], all reflective surfaces are represented by a number of small-area elements \( e_n \), that is, \( M \). In Equation 3.21 and Figure 3.5, both \( d_{en}^t \) and \( d_{en}^r \) represent small reflecting areas that are acting as an Rx with respect to the light source \( T_i \), and then as a source to \( R_j \). The overall CIR taking into account multiple transmitters and multiple reflections can be written as [29]:

\[
h(\tau; T_i, R_j) = \sum_{i=1}^{M} \sum_{k=0}^{\infty} h_S^k(\tau; T_i, R_j).
\]

(3.22)

For the case of high data rate indoor VLC systems that we consider here, because of relatively slow movement of people and fixed objects within a room, the channel can effectively be considered as time invariant. To determine \( h_S^k(\tau; T_i, R_j) \), we carry out the following: (i) calculate the \( M \) impulses responses \( h_S^0(t, d_{en}^t, R_j) \); (ii) continue computing \( h_S^1(t, d_{en}^t, R_j) \) until we have \( h_S^{k-1}(t, d_{en}^t, R_j) \); and (iii) use (3.21) to calculate \( h_S^k(t, d_{en}^t, R_j) \) for each receiver. Note that the computation time \( t_{\text{com}} \) really depends on three key parameters of \( N \) and \( M \) and the number of Rxs. For a single Rx, \( t_{\text{com-1Rx}} = (M^2 \cdot N^2) \) [29]. In a typical room of size 4 \( \times \) 4 m\(^2\) with \( M \) of 2024 facets, 100 sec \(< t_{\text{com}} < 0.03 \) sec for a number of Rx elements from 1 to 1000, respectively.

From the CIR, two important items are deducted: channel gain and the root mean square (RMS) delay spread \( \tau \). It has been shown that \( \tau \) and the channel gains are more than sufficient to model diffuse configurations.
The delay spread $\tau$ provides a good estimate to how susceptible the channel is to ISI, and can be computed from the CIR using:

$$\tau = \left( \frac{\int (t-\mu)^2 h^2(t) dt}{\int h^2(t) dt} \right)^{\frac{1}{2}},$$

(3.23)

where $t$ is the propagation time and $\mu$ is the mean excess delay given by:

$$\mu = \frac{\int t h^2(t) dt}{\int h^2(t) dt}.$$  

(3.24)

Obviously, different room and Tx–Rx configurations can significantly affect $\tau$, and smaller values of $\tau$ indicate a higher system transmission bandwidth [13,31]. In OWC systems the most important feature is the channel gain (defined as the ratio between $P_R$ and $P_E$), which determines the achievable SNR for a given $P_E$ [29].

### 3.4 Channel Limitations and ISI

#### 3.4.1 Multipath Dispersion

In optical communications, the information carrier signal has a frequency of about $10^{14}$ Hz. Typically, the PD active area in VLC receivers is about millions of square wavelengths. Since the total generated photocurrent is proportional to the integral of the optical power over the entire PD surface area, this will provide an inherent spatial diversity. Therefore, indoor VLC systems are effectively not subject to multipath fading [8,25]. For the same reason, we are concerned with negligible Doppler spreads and the channel can mostly be considered as time invariant (except when shadowing or beam blockage occurs) [32]. However, multipath propagation of emitted signals in these systems leads to time dispersion and ISI, which will limit the transmission rate [24].

The signal intensity received on the PD surface includes contributions from the LOS (with respect to the transmitters), as well as from reflections of walls or objects within the room [15]. For the LOS contribution, the channel response is modeled by Dirac pulses, whereas for the diffuse part, it is represented by an integrating sphere model [10]. The diffuse component is almost constant and depends on the room properties and the Rx aperture size.

To perform a more detailed analysis, a case study is presented. Consider a room of dimension $5 \times 5 \times 3 \text{ m}^3$ with a single Tx in the middle of the ceiling. The Rx is placed at a height of 0.5 m that corresponds to a typical desktop. Using equations (3.4, 3.7, 3.9, and 3.10), and for a transmit optical power of 2 W, $\Phi_{1/2}$ of 60°, $I(0)$ of 200 Lux at 700 mA of current, four LEDs, Rx FOV of 60°, PD surface area of 16 mm$^2$, $\rho$ of 0.8, a unity gain optical filter, and a lens at PD with a refractive index of 1.5, the received optical power distributions
corresponding to the LOS path and multipaths (only the first-order reflections) are shown in Figure 3.6a and b, respectively.

We notice that the impact of multipath reflections is most significant at room corners. They have a much lower impact when the Rx is located beneath the Tx, however, which is quite logical as the LOS has the main contribution in the received signal.

In most practical cases, however, the influence of the diffuse component is masked by the strong LOS component. It is shown in [15] that it has no significant influence on the overall channel bandwidth. Indeed, for typical room dimensions, the channel time dispersion corresponding to the LOS component is negligible [15]. For example, considering a room of dimension $5 \times 5 \times 3$ m$^3$ with Tx s configuration as shown in Figure 3.7 the maximum delay between two LOS paths is around 5.5 ns only.

**FIGURE 3.6**
Received optical power distribution corresponding to (a) LOS and (b) first-order reflection multipath. $5 \times 5 \times 3$ m$^3$ room, Rx height 0.5 m.

**FIGURE 3.7**
Example of ceiling lighting design using four LEDs of $1 \times 1$ m$^2$ spaced 7 cm apart with a total number of 900 chips. (Adapted from Grubor, J., et al., *J. Lightwave Technol.*, 26, 3883–3892, 2008.)
In general, the limitation of ISI depends on the transmission scenario and the room properties, and can be quantified by evaluating the channel cut-off frequency. Consider an LED modulation bandwidth of 20 MHz (which is the case when blue filtering is performed, as explained next in Subsection 3.4.2), the Nyquist symbol period is limited to 25 ns, and ISI will occur if transmitted data symbols experience delays larger than 12.5 ns, assuming that LEDs are synchronously driven. Simulation results provided in [15] showed that channel bandwidth limitation corresponds to the minimum bandwidth of 90 MHz at the worst location at the desktop surface, which is significantly above the 20 MHz limitation of the LED itself. As a result, the channel can effectively be considered as frequency nonselective (flat) over the bandwidth of interest.

Note that, alternatively to the cut-off frequency, the transmission bit rate $R_b$ can be used given the RMS delay spread $\tau$ as [27]:

$$R_b \leq \frac{1}{10\tau}. \quad (3.25)$$

It should be noticed that in large rooms such as conference halls where we have a noticeable difference between the optical path delays, we may be concerned with ISI for higher transmission data rates. This is specially the case when the Rx is placed in the corner of the room, where the diffuse propagating component becomes predominant [33]. In such cases, advanced modulation schemes should be used.

Note that, in addition to the room dimensions, channel dispersion also depends on the receiver FOV and the distance between the Tx and the Rx. When smaller beam divergence angles are used at the Tx or Rxs of smaller FOV are used, the dispersion is due to multipath scattering and reflections, and hence the ISI is reduced. This is the case for tracked directed links. Such systems can potentially support data transmission speeds of more than 100 Mbps [10,25] but require sophisticated tracking mechanisms to ensure link connectivity. On the other hand, the Tx radiation pattern can be optimized in order to improve link properties. More specifically, we can optimize the Tx Lambertian order through the use of a beam diffuser to maximize the LOS path gain and to minimize the ISI. This is particularly interesting in multicell scenarios. We consider three different room sizes and cell configurations as specified in Table 3.1.

We have shown in Figure 3.8 the normalized CIR for the three cases of Rooms A, B, and C in Table 3.1 where the typical first-order Lambertian order and optimized Lambertian order LEDs are used. We notice that for the Room A in Figure 3.8a, the amplitude of CIR corresponding to LOS increases from 43.5% to 72% by using the Optimum Lambertian order (OLO). Also, the LOS component increases from 35.6% to 81.6% and from 25.6% to 80.3% by using OLO for the two other cases of Rooms B and C, shown in Figure 3.8b and c, respectively. At the same time, the contribution of the reflected paths, and hence the ISI, decreases significantly.
We have also shown in Figure 3.9 the profile of the RMS delay spread for different Rx positions for the nonoptimized and optimized source patterns for the case if a four-cell scenario and $5 \times 5 \times 3$ m room size. The average RMS delay spread by using OLO decreases from ~1.5 ns to ~0.4 ns and the peak RMS delay spread which corresponds to the room corners decreases from ~2.3 ns to ~0.5 ns [34].

In the case of multiple emitting sources, the main factor that impacts the channel frequency selectivity is the asymmetry between the multiple LOS paths rather than multipath reflections [20,35]. As a matter of fact, the RMS delay spread is a useful metric for comparing the degree of frequency selectivity of the different link configurations. However, its absolute value cannot be used to determine the limitation on the transmission rate [20]. One may resort to the channel 3-dB cut-off frequency to determine the degree of channel frequency selectivity. However, this metric is of limited interest in practice except for the case of a purely diffuse channel, that is, blocked LOS. Otherwise, the oscillating behavior of the frequency response due to the contribution of the LOS component makes the 3 dB bandwidth meaningless [20,35]. It is shown in [20,35] that a

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**TABLE 3.1**

Specification of Studied Indoor VLC Systems

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>LED wavelength ($\lambda$)</td>
<td>(500–1000) nm</td>
</tr>
<tr>
<td>LED power</td>
<td>200 mW</td>
</tr>
<tr>
<td>Half angle FOV of receiver</td>
<td>60 (deg.)</td>
</tr>
<tr>
<td>Active area of photodiode</td>
<td>16 mm$^2$</td>
</tr>
<tr>
<td>Gain of optical filter</td>
<td>1.0</td>
</tr>
<tr>
<td>Refractive index of a lens at a photodiode</td>
<td>1.5</td>
</tr>
<tr>
<td>Reflection coefficient (wall, ceiling, floor)</td>
<td>(0.8, 0.8, 0.3)</td>
</tr>
<tr>
<td>Room A (width, length, height)</td>
<td>$5 \times 5 \times 3$ m</td>
</tr>
<tr>
<td>Number of cells</td>
<td>4</td>
</tr>
<tr>
<td>Cell radius ($r$)</td>
<td>1.77 m</td>
</tr>
<tr>
<td>OLO</td>
<td>5.7</td>
</tr>
<tr>
<td>Room B (width, length, height)</td>
<td>$4 \times 6 \times 3$ m</td>
</tr>
<tr>
<td>Number of cells</td>
<td>6</td>
</tr>
<tr>
<td>Cell radius ($r$)</td>
<td>1.41 m</td>
</tr>
<tr>
<td>OLO</td>
<td>9</td>
</tr>
<tr>
<td>Room C (width, length, height)</td>
<td>$5 \times 5 \times 3$ m</td>
</tr>
<tr>
<td>Number of cells</td>
<td>9</td>
</tr>
<tr>
<td>Cell radius ($r$)</td>
<td>1.17 m</td>
</tr>
<tr>
<td>OLO</td>
<td>13</td>
</tr>
</tbody>
</table>

*Note: FOV, field of view; OLO, optimum Lambertian order.*
FIGURE 3.8
The normalized CIR when using first-order and optimized Lambertian order light sources in:
(a) Room A, (b) Room B, and (c) Room C configurations.
useful metric in quantifying the amount of ISI is signal-to-ISI ratio (SIR), defined as the ratio of the received powers corresponding to the “desired” signal and ISI, respectively. Defining the ISI on the sampled signal at the Rx, the interest of this metric was demonstrated in [20]. Furthermore, taking into account the effect of the Rx filter, it was shown that a simple Bessel low-pass filter is preferable to the matched filter (assuming perfect time synchronization at the Rx) since it provides a higher SIR at relatively high data rates.

3.4.2 LED Bandwidth Limitation

As a matter of fact, the main limitation on high data rate transmission arises from the limited bandwidth of the LED. This bandwidth limitation is due to the power-bandwidth trade-off and also the parasitic elements in the packaging of the LED [32]. Whereas the trichromatic LEDs have bandwidths of several hundred megahertz and more [17,25], for the reasons of fabrication cost and also color-shift over time, white light LEDs (blue LED plus phosphorous layer) are envisioned for indoor lighting. Typically, the bandwidth of white LEDs is limited to several MHz mainly due to the slow time constant of phosphor. If blue filtering is performed at the Rx to suppress the phosphorescent portion of the optical spectrum, this bandwidth can be extended to about 20 MHz [15].

To increase the transmission rate over this limit, a number of solutions have been proposed so far. One solution is the pre-equalization of the driving circuitry, by which the bandwidth can be increased up to 50 MHz [36,37] but at the price of high SNR penalty (around 20 dB). Another solution is to use multilevel modulation together with DMT modulations, which require complex driving circuitry at the Tx [15,38–40], or to perform frequency-domain equalization (FDE) at the price of increased complexity of the Rx [41–43]. Another alternative is to use MIMO architectures to benefit from spatial
multiplexing at the Tx [44–46]. We will consider channel modeling for MIMO systems in Section 3.6.

Note that bandwidth limitation is much more important in the case of using organic devices. The typical bandwidth for organic PDs and organic LEDs (OLEDs) is in the order of 30 KHz and 90 KHz, respectively [47], which is much lower than those for the inorganic counterparts. The use of advanced transmission techniques such as equalization, multilevel modulation, DMT, or MIMO is particularly promising in this case for increasing the data rate beyond the limitations of the components [48–50].

Lastly, although it is not a fundamental issue, cabling can impact the transmission speed in indoor VLC systems. In fact, the difference between the electrical signal paths driving the LEDs on the ceiling can lead to ISI. This obviously depends on the distribution of LED lamps on the ceiling. It is shown in [15] that the critical cabling difference is about 1.6 m considering the Nyquist symbol rate for a bandwidth of 20 MHz. This can be managed easily in installations in medium-size rooms.

3.5 Signal Distortion

3.5.1 Nonlinear LED Characteristics

Another impairment that can affect the performance of a VLC system is the nonlinearity of the LED transfer function, regarding both voltage–current and current-emitted optical power relationships [51–53]. In addition, the signal is also limited due to the limited dynamic range of the LED.

On the other hand, as linear power amplifiers cannot be used for reasons of their high power consumption, we should admit even more nonlinearity [32]. The effect of LED nonlinearity is especially important when high-order constellations are used. It is also problematic when quadrature amplitude modulation (QAM) with DMT is employed in order to increase the data rate or to reduce the impact of the ambient noise from artificial light sources such as fluorescent lighting (see Section 3.3.2). In such a case, the LED nonlinear transfer function causes cross-talk between the subcarriers [53]. It is hence important to consider appropriate modeling of LED nonlinearity in order to evaluate/predict the effective system performance.

Note that, although this impairment does not concern the physical propagation channel, we can consider it as a part of the global channel model, incorporating the imperfect effects of the Tx and Rx devices.

3.5.2 Distortion Modeling

The most common approach to account for LED nonlinearity is to consider memory or memoryless models mostly based on a static model (i.e., neglecting
the change in the characteristics of the LED over time) [51]. Considering modulation frequencies well below the LED 3-dB bandwidth, the classical approach is to consider a memoryless model and to use a polynomial fit to the nonlinear transfer function. However, to obtain a realistic model, the polynomial order should be more than five, though a second-order polynomial can provide a fair description of the transfer function.

This way, the output power $P_{out}$ is described as a function of the input current $I_{in}$ as follows:

$$P_{out} = b_0 + b_1(I_{in} - I_{DC}) + b_2(I_{in} - I_{DC})^2,$$

where the coefficients $b_0$, $b_1$, and $b_2$ are the polynomial coefficients and $I_{DC}$ is the DC bias current.

For a more accurate model that can be used for higher modulation frequencies (i.e., large signal bandwidths) the memoryless model is not adequate. Indeed, the frequency-dependency of the current–voltage characteristics of the LED necessitates taking the memory effects of the nonlinearity into consideration [51]. One solution is to use the active region carrier density rate equation [54]. The Volterra series representation of the nonlinearity is the most accurate method but the practical interest of this model is limited due to the high computational complexity for calculating the model parameters that makes it inappropriate for real-time applications [55,56]. As an alternative to this model, a memory polynomial model can be used, as suggested in [57]. Another simplification of this model is to consider two blocks of a linear time-invariant (LTI) system and a memoryless nonlinear system. The order of these two blocks results in Wiener (LTI followed by memoryless nonlinear) [58] or Hammerstein [51] models (otherwise).

### 3.6 MIMO VLC Systems

#### 3.6.1 Interest of MIMO Structures

In most VLC systems, we have a relatively high SNR available. In order to achieve high data rates despite the limited bandwidth of LEDs, one solution is to perform spatial multiplexing by using the MIMO technique. MIMO systems offer a higher data throughput as well as increased link range without the requirement for additional power or bandwidth. MIMO systems have been widely proposed for optical interconnects between source and detector arrays in order to simplify the source-detector alignment [59]. In VLC systems, however, MIMO systems have attracted attention for their ability of increasing channel capacity [16,17,44–46,60].
In contrast to radio frequency [61] or free-space optical communication [62], where MIMO systems are used to achieve increased link reliability by providing spatial diversity gain, in VLC systems we are concerned with a deterministic channel, and hence the only interest of MIMO architectures is for increasing the data throughput.

### 3.6.2 Channel Modeling for MIMO VLC Systems

For MIMO VLC systems, there are two approaches of nonimaging and imaging receivers. Let us start with the simpler one, that is, the nonimaging system where at the Rx side, nonimaging lenses are used for collecting the transmitted intensity. Figure 3.10 shows an example of such a system where the LEDs and receivers are arranged in a $2 \times 2$ array. Each PD, through nonimaging concentrators, collects the light from the LEDs with different intensities.

Let us consider the general case of an MIMO architecture with $N_T$ transmitters and $N_R$ receivers. As explained in Section 3.4.1, the LOS propagation component dominates the diffuse one in most practical cases. Furthermore, we can practically neglect the difference between the propagation delays of the different LOS paths [44]. So, the MIMO channel can be described by a matrix $H$ of dimension $N_R \times N_T$, whose entries are the DC channel gains between a pair of Tx–Rx.
\[ H = \begin{bmatrix} h_{11} & \cdots & h_{1N_T} \\ \vdots & \ddots & \vdots \\ h_{N_k1} & \cdots & h_{N_kN_T} \end{bmatrix}. \] (3.27)

For instance, \( h_{ij} \) represents the channel gain between the \( j \)th PD and \( i \)th LED. The channel matrix includes the LOS component as well as the diffuse component arising from multipath reflections. A general form of \( H \) is proposed in [63] as follows:

\[ H = G_r(D + F_s \cdot \Pi), \] (3.28)

where \( D \) represents the contribution of the LOS components, \( F_s \) and \( G_r \) denote the Tx and Rx profiles, respectively, and \( \Pi \) is the environment matrix representing the contribution of surface reflectors.

If we denote the vectors of transmitted and received signals at a given time reference by \( X = [x_1, \ldots, x_{N_T}]^T \) and \( Y = [y_1, \ldots, y_{N_R}]^T \), respectively, we can write:

\[ Y = R P_{\text{LED}} H X + n, \] (3.29)

where \( R \) is the detector responsivity, \( n \) is the noise, and \( P_{\text{LED}} \) is the average transmitted power. Then, the transmitted data are obtained, for instance, based on channel inversion:

\[ \hat{X} = H^{-1} Y, \] (3.30)

where it is assumed that \( H \) is known at the Rx, which can be realized through the transmission of some pilot signals. Inverse filtering is justified by the relatively high SNR available at the Rx in VLC systems. However, in order to estimate the transmitted signals from Equation 3.30, the channel matrix must obviously be of full rank. For the configuration shown in Figure 3.10, this is not the case when the Rx is situated in the center of the room or along its axes. As a result, by nonimaging MIMO, the channel bandwidth is position dependent; depending on the LED configuration geometry and Rx position, the channel matrix can be ill-conditioned, and in the worst case, rank-deficient. Note that inverting an ill-conditioned \( H \) results in a significant noise amplification and consequently a considerable bit error rate (BER) increase. (The reader is referred to [44], Figure 3, which shows the dependence of the BER to the Rx position for a configuration similar to Figure 3.10.)

To circumvent the problem of rank-deficient channel matrix, an imaging lens system [64] can be used at the Rx. Figure 3.11 illustrates the imaging MIMO structure, where the LED arrays are “imaged” to the Rx plane via the imaging lens [44]. This requires a large enough Rx area so that the images of the LED arrays fall on the detectors for all possible Rx positions inside the room [65]. By paraxial approximation, we neglect image distortion due to the dependence of magnification on the angle of incidence of the rays.
This technique is quite efficient but the Rx imaging lens is bulky and introduces additional expense and complexity. Another approach is to use a standard camera technology [66], but the problem is the limited FOV as such cameras are designed to produce focused images that match the human eye. Lastly, the use of a hemispherical imaging lens has been recently proposed in [45], which has the advantage of providing a very wide FOV with low correlation between the underlying subchannels (in the case of using several PDs at the Rx).

References


