A medium with advantages over radio...and some drawbacks

Non-Directed Infrared Links for High-Capacity Wireless LANs

Joseph M. Kahn, John R. Barry, Malik D. Audeh, Jeffrey B. Carruthers, William J. Krause, and Gene W. Marsh

The emergence of portable information terminals in future work and living environments is expected to accelerate the introduction of wireless LANs. Such portable terminals should have access to all of the services that will be available on wired networks. Unlike their wired counterparts, portable devices are subject to severe limitations on power consumption, size, and weight. The desire for inexpensive, high-speed links satisfying these requirements has motivated recent interest in infrared wireless communication [1-5].

As a medium for short-range, indoor communication, infrared offers several significant advantages over radio, including a virtually unlimited spectral region that is unregulated worldwide. Near-infrared and visible light are close together in wavelength, and they exhibit qualitatively similar behavior. Both are absorbed by dark objects, diffusely reflected by light-colored objects, and directionally reflected from shiny surfaces. Both types of light penetrate through glass, but not walls or other opaque barriers. As a result, infrared communications can readily be secured against eavesdropping. Moreover, it is possible to operate at least one infrared link in every room of a building without interference, so that the potential capacity of an infrared-based network is extremely high. When an infrared link employs intensity modulation with direct detection (IM/DD), the short carrier wavelength and large, square-law detector lead to efficient spatial diversity that prevents multipath fading. By contrast, radio links are typically subject to large fluctuations in received signal magnitude and phase.

The infrared medium is not without drawbacks, however. In many indoor environments there exists an intense infrared ambient, arising from sunlight, incandescent lighting, and fluorescent lighting, which induces noise in an infrared receiver. In virtually all short-range, indoor applications, IM/DD is the only practical transmission technique. The signal-to-noise ratio of a DD receiver is proportional to the square of the received optical power, implying that IM/DD links can tolerate only a comparatively limited path loss. Often, infrared links must employ relatively high transmit power levels and operate over a relatively limited range. While the transmitter power level can usually be increased without fear of interfering with other users, transmitter power may be limited by concerns of power consumption and eye safety, particularly in portable transceivers. Some of the characteristics of infrared and radio indoor wireless links are compared in Table 1.

Using directional infrared transmitters and receivers, it is possible to achieve high bit rates and long link ranges using relatively modest transmitter power [6]. In most applications of wireless LANs, however, it is desirable to form links using omnidirectional transmitters and receivers, alleviating the need for careful alignment between them. This article will focus on such non-directed links. As illustrated in Fig. 1, non-directed infrared links may be classified into two categories: line-of-sight (LOS) and diffuse. LOS links depend upon the existence of an unobstructed path between transmitter and receiver. Diffuse links alleviate the need for a direct LOS path by relying on light scattered from a large diffuse reflector, such as a ceiling. Because it is difficult to block all of the light reflected from such a large surface, diffuse links are more robust than LOS links, and may be preferable for many applications.

Fig. 2 illustrates two different paradigms for creating wireless infrared LANs serving portable information terminals. When two or more terminals are located in the same room, they may communicate directly with each other on a peer-to-peer basis, forming an ad hoc network. Portable transceivers designed for such ad hoc interconnection should consume little power and be relatively inexpensive. Alternatively, infrared links may also be used to connect portable devices to base stations that are interconnected by a wired backbone network. Such an installed network would permit terminals to communicate with multimedia and compute servers, or with portable devices located in other rooms. In this scenario, the portable terminals should be inexpensive and low-power, but it might be permissible for the base stations to be more complex and to consume greater power. In some future high-performance multimedia wireless computing environments, the portable terminals may serve mainly as a human interface, accepting pen and keyboard input, but displaying full-motion video. The very high-capacity downlinks (tens of Mb/s per base station) and moderate-capacity uplinks (several Mb/s per base station) required of such a system would be particularly well-matched to the capabilities of infrared communication. Smaller rooms could be served by a single base station, while rooms larger than about 10 m x 10 m may require more than one base station. Techniques for accommodation of multiple base stations in one room will be touched upon below.

Despite a relative scarcity of research publications on wireless infrared communications, the technology has found wide commercial application. Directed infrared beams are commonly used in remote-control devices, as well as in serial links for
computer peripherals operating at bit rates up to 112 kbs [7]. Diffuse infrared has been employed for several years in commercial audio transmission systems. During the past year, there have been several new products using diffuse infrared transmission to permit interconnection of portable computers. Both IBM and Photonics are marketing modems[8,9] that permit ad hoc, peer-to-peer interconnection of notebook computers at a bit rate of 1 Mbps, achieving coverage of a 10 m x 10 m room. These modems use light-emitting diode (LED) sources with 16-pulse-position modulation (PPM), and support a CSMA/CA protocol. Spectrix is offering portable terminals [10] that use 2-PPM of LED sources to achieve a 4-Mbps transmission speed over a 15-m range. These links permit wireless communication with base stations connected to a backbone network, making use of a deterministic reservation/polling SNMP protocol. While there are currently no standard transmission formats or protocols for wireless infrared networks, a subgroup of the IEEE 802.11 committee is expected to draft the first standards for wireless infrared networks in early 1994 [11].

<table>
<thead>
<tr>
<th>Channel Property</th>
<th>Non-Directed Infrared</th>
<th>Radio</th>
</tr>
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<tbody>
<tr>
<td>Path loss</td>
<td>High</td>
<td>High</td>
</tr>
<tr>
<td>Multipath fading</td>
<td>No</td>
<td>Yes</td>
</tr>
<tr>
<td>Multipath distortion</td>
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<td>Yes</td>
</tr>
<tr>
<td>Dominant noise</td>
<td>Shot noise from</td>
<td>Interference from</td>
</tr>
<tr>
<td></td>
<td>background light</td>
<td>other users</td>
</tr>
<tr>
<td>Input X(t) represents</td>
<td>Power</td>
<td>Amplitude</td>
</tr>
<tr>
<td>SNR proportional to X(t)</td>
<td></td>
<td>1</td>
</tr>
<tr>
<td>Average power proportional to X(t)</td>
<td></td>
<td>1</td>
</tr>
<tr>
<td>Bandwidth limitation</td>
<td>Photodiode capacitance and transit time</td>
<td>Regulatory</td>
</tr>
</tbody>
</table>

Table 1. Comparison of infrared and radio for indoor communications.

Using technology employed in current commercial systems, it should be relatively straightforward to extend the bit rate of non-directed infrared communication links to approximately

Figure 1. Configurations of non-directed indoor infrared links: (a) line of sight, (b) diffuse. T and R denote transmitter and receiver, respectively.

Figure 2. Network of portable multimedia terminals using wireless infrared access to wired backbone. A peer-to-peer interconnection of two portables is also shown (from [22]).
Nature of Non-Directed Optical Channels

After non-directed propagation, an infrared beam typically consists of an unknown superposition of modes, making it very difficult to implement efficiently heterodyne or homodyne optical detection. Such techniques are also too complex for this cost-sensitive application, so that IM/DD is the only practical alternative. The modeling of non-directed infrared channels with IM/DD is illustrated in Fig. 3. The transmitted waveform \( X(t) \) is the instantaneous optical power of the infrared emitter. The received waveform \( Y(t) \) is the instantaneous current in the receiving photodetector, which is proportional to the integral over the photodetector surface of the total instantaneous optical power at each location.\(^1\) As shown in Fig. 3(a), the received electric field generally displays spatial variation of magnitude and phase,\(^2\) so that “multipath fading” would be experienced if the detector were smaller than a wavelength. Fortunately, typical detector dimensions are thousands of wavelengths, leading to efficient spatial diversity that prevents multipath fading. As the transmitted optical power \( X(t) \) travels along various paths of different lengths, non-directed infrared channels are still subject to multipath-induced distortion. The channel can be modeled as a baseband linear system, with input power \( X(t) \), output current \( Y(t) \), and an impulse response \( h(t) \), which is fixed for a given arrangement of transmitter, receiver and intervening reflectors. A mathematical derivation of this channel model can be found in [12], where it is shown that the linear relationship between \( X(t) \) and \( Y(t) \) is a consequence of incoherent propagation. By contrast, we note that when IM/DD is used with narrow-linewidth sources in dispersive single-mode optical fibers, the relationship between \( X(t) \) and \( Y(t) \) is nonlinear [13].

In many applications, non-directed infrared links are operated in the presence of intense infrared and visible background light. As we will see in the following section, it is possible to reduce this

10 Mb/s. Higher transmission speeds will be desirable in future wireless computing environments. In this article, we describe the physical obstacles to achieving higher bit rates, and we discuss the technical means to achieve bit rates as high as 100 Mb/s.

![Figure 4](image_url) Equivalent electrical magnitude response (a) and impulse response (b) of non-directed infrared channels. Measurements were performed in an empty 5.5 m x 7.5 conference room having a ceiling height of 3.5 m (from [12]).
received ambient light by optical bandpass or longpass filtering, but this background still adds a white, nearly Gaussian shot noise \(n(t)\) that is the limiting factor in the signal-to-noise ratio (SNR) of a well-designed receiver. Our channel model is summarized by

\[
Y(t) = X(t) \otimes h(t) + n(t),
\]

(1)

where the \(\otimes\) symbol denotes convolution. As in linear electrical or radio channels with additive noise, the SNR is proportional to \(|X(t)|^2\). However, infrared channels differ from conventional channels because the channel input \(X(t)\) represents power. Thus, \(X(t)\) cannot be negative, and the average power is proportional to a time integral of \(X(t)\),

\[
P_{\text{avg}} = \lim_{T \to \infty} \frac{1}{2T} \int_{-T}^{T} X(t) \, dt,
\]

(2)

rather than the usual \(|X(t)|^2\), which is appropriate when \(X(t)\) represents amplitude.

While non-directed propagation alleviates the need for physical alignment between transmitter and receiver, a major drawback of this approach is the signal distortion caused by reflections from ceilings, walls, and other objects. To evaluate the possible impact of multipath distortion on high-speed infrared links, we have performed experimental characterization of about 90 channels in five offices and conference rooms [12]. We used a vector network analyzer to perform swept-modulation frequency characterization [14] of the channel frequency response \(H(f)\). The 832-nm transmitter emitted a Lambertian radiation pattern, i.e., with power-per-unit solid angle proportional to the cosine of the angle with respect to transmitter surface normal. Our receiver exhibited a 3-dB cutoff frequency of 150 MHz. During all measurements, the receiver was placed at desk height and pointed upwards. To form LOS configurations, the transmitter was placed near the ceiling and pointed straight down, while for diffuse configurations, it was placed at desk height and pointed straight up. Shadowed LOS and diffuse configurations were formed by having a person stand next to the receiver so as to block the dominant reception path. The channel frequency response \(H(f)\) is scaled so that \(H(0)\) represents the ratio of the optical power received by a 1-cm\(^2\) detector to the total transmitted optical power. We obtain the channel impulse response \(h(t)\) by inverse Fourier transformation of \(H(f)\).

Fig. 4 presents the magnitude and impulse responses of four different non-directed infrared link configurations, measured in an empty conference room. While the details of channel responses depend on the link geometry, responses measured at all positions in all rooms exhibit qualitative similarity to Fig. 4. Unshadowed LOS impulse responses are dominated by a short initial pulse, and the strongest distinct reflections typically arrive 15 to 20 ns after the initial pulse. Dominance of the short initial pulse leads to magnitude responses that are flat at high frequencies. Unshadowed diffuse impulse responses exhibit a significantly wider initial pulse, which has a width of about 12 ns at 10 percent height, corresponding to the existence of a continuum of different path lengths between illuminated portions of the ceiling and the receiver.

**Figure 5.** Optical path loss vs. horizontal separation between transmitter and receiver for diffuse, unshadowed channels. Measurements were performed in a 8.5 m x 9 m office having a ceiling height of 2.4 m, with transmitter and receiver arranged as in Fig. 1(b). Solid line is result of a model taking account only of the first reflection from the ceiling, which has a diffuse reflectivity of 80 percent. At large separation, results of that model approach a \(d^4\) law (from [12]).

This continuous distribution of path delays leads to a steady decrease in the channel magnitude response at high frequencies. For all channels, the impulse response may contain significant energy as long as 70 ns after its initial zero-crossing. The d.c. gain of all channels is enhanced over that at high frequencies, because it includes the contribution due to the entire duration of the impulse response.

Shaded channels exhibit characteristics that are slightly less predictable than the unshadowed channels. The shadowed LOS impulse response typically resembles the corresponding unshadowed response with the dominant initial pulse removed, since only indirect propagation paths remain. We observe that in LOS configurations, shadowing significantly degrades the channel frequency and impulse responses. Diffuse configurations are far less vulnerable to shadowing than their LOS counterparts, because in diffuse configurations there exist many possible propagation paths between the illuminated ceiling area and the receiver. In diffuse configurations, shadowing produces a slight broadening of the impulse response, and a slight increase in the rate of falloff of the magnitude response with increasing frequency.

As the frequency response \(H(f)\) is relatively flat at low frequencies, the equivalent d.c. electrical gain \(H(0)\) is probably the single most important quantity characterizing a non-directed infrared channel. The same information is conveyed by the optical path loss, which we define as the ratio of the average transmitted power to the average power received by a detector of 1-cm\(^2\) area, i.e., the reciprocal of \(H(0)\). The path losses of diffuse channels measured in one office are shown in Fig. 5. These measured path losses can be fitted with reasonable accuracy by a model [15] that considers only the light undergoing one diffuse reflection from the ceiling en route from source to receiver. We note that for large horizontal separation between trans-

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1 The detector is equivalent to a two-dimensional array of many antennas whose receptions are squared, lowpass filtered, and summed.

2 Under very unusual circumstances — with point or plane-wave source, LOS propagation, and carefully aligned detector — this spatial variation would exhibit a regular pattern, but in typical cases of LOS or diffuse propagation, it appears to be random.
minter and receiver, this model predicts that diffuse path losses should increase as the fourth power of horizontal separation. Considering measurements performed in five rooms, for a given horizontal separation, the unshadowed LOS configuration generally yields the lowest path loss, and the path loss of the corresponding unshadowed diffuse configuration is typically 1 to 3 dB higher. In the presence of shadowing, however, LOS configurations exhibit typically a 7 to 10 dB increase in path loss, while in the diffuse case, the corresponding increase is about 2 to 5 dB. As a result, shadowed diffuse configurations typically yield path losses that are 2 to 5 dB smaller than the corresponding shadowed LOS configurations, indicating the robustness of an extended, diffuse source.

In comparing the multipath dispersion of different channels, we will consider the channel r.m.s. delay spread. The delay spread is calculated from the impulse response according to:

\[
\text{r.m.s. delay spread of } h(t) = \left[ \frac{\int (t - \mu)^2 h^2(t) \, dt}{\int h^2(t) \, dt} \right]^{1/2}
\]

where the mean delay \( \mu \) is given by

\[
\mu = \frac{\int t h^2(t) \, dt}{\int h^2(t) \, dt},
\]

and the limits of integration in (3) and (4) extend over all time. We emphasize that since \( h(t) \) is fixed for a given configuration, so is the delay spread. Fig. 6 presents the r.m.s. delay spread versus horizontal separation for our measured channels. In the absence of shadowing, LOS channels, whose impulse response is dominated by a short initial pulse, generally yield the smallest delay spreads, ranging from a measurement-limited 1.3 ns to about 12 ns. Unshadowed diffuse channels exhibit delay spreads that lie in the same range, but which are systematically slightly larger, due to the finite temporal spread of the dominant reflection from the ceiling. Shadowing increases the delay spread of both LOS and diffuse channels but, as might be expected, the increase is relatively modest for the latter. The LOS channels consistently exhibit the largest delay spreads, typically between 7 and 13 ns. In the section on performance of modulation techniques that follows, we will see that the delay spread is a reasonably accurate predictor of the multipath power penalty incurred in links using baseband on-off keying (OOK).

We have performed detailed simulations of non-directed infrared propagation [16], using a recursive technique that can account for an arbitrary number of diffuse reflections from room surfaces. For both LOS and diffuse channels, simulations are in good quantitative agreement with measured channel responses, leading us to believe that multipath infrared propagation can be well-described by simple models.

Having discussed the characteristics of non-directed infrared channels, we will now examine how to design reliable, high-speed links. We will first describe strategies for achieving a high SNR, as they are applicable to all links, independent of bit rate and modulation format. Then, we will proceed to discuss various modulation techniques, addressing their power efficiency and, for high bit rates, their robustness against multipath distortion.

Achieving a High Signal-to-Noise Ratio

The wavelength band between about 780 and 950 nm is presently the best choice for non-directed links, due to the availability of low-cost light-emitting diodes (LEDs) and laser diodes (LDs), and because it coincides with the peak responsivity of inexpensive, low-capacitance silicon photodiodes. LEDs are currently used in all commercial systems, due to their extremely low cost and because most LEDs emit light from a sufficiently large surface area that they are generally considered eye-safe. Potential drawbacks of LEDs include:

- Typically poor electro-optic power conversion efficiency of 10 to 20 percent (though new devices have efficiencies as high as 40 percent).
- Modulation bandwidth that is typically limited to tens of MHz.
- Broad-spectral width (typically 25 to 100 nm), which requires the use of a wide receiver optical passband, leading to poor rejection of ambient light.
- The fact that wide modulation bandwidth is usually obtained at the expense of reduced electro-optic conversion efficiency.

LDs are much more expensive than LEDs, but offer many nearly ideal characteristics:

- Electro-optic conversion efficiencies of 30 to 70 percent.
- Wide modulation bandwidths, which range from hundreds of MHz to more than 10 GHz.
- Very narrow spectrum (spectral widths ranging from several nm to well below 1 nm are available).

To achieve eye safety with a LD requires that one pass the laser output through some element that destroys its spatial coherence and spreads the radiation over an extended emitting aperture. New eye-safety regulations are likely to restrict diffuse power densities at wavelengths near 850 nm to levels of about 370 W/m² for continuous viewing [17, 18]. With an LD, eye safety can be achieved using a transmissive diffuser, such as a thin plate of translucent plastic. While such diffusers can achieve efficiencies of about 70 percent, they typically yield a Lambertian radiation pattern, offering the designer little freedom to tailor the source radiation.
pattern. Computer-generated holograms (CGHs) offer a means to generate arbitrary radiation patterns with efficiencies approaching 100 percent, but must be fabricated with extreme care to ensure that any residual image of the LD emission aperture is tolerably weak [17].

Infrared links are subject to two different classes of ambient light. Sunlight, skylight, and incandescent lights represent broadband visible and infrared sources that vary slowly with time. These sources induce a white, nearly Gaussian shot noise whose electrical power spectral density (PSD) is proportional to the total detected optical power. This shot noise can thus be limited by use of an optical bandpass filter or longpass filter that rejects most of the background light, but passes the desired signal. The emission of fluorescent lights, on the other hand, is intensity-modulated, either at the power-line frequency or, with new compact-ballast fixtures, at frequencies of tens of kHz. This near-d.c. interference includes harmonics that are particularly strong for fixtures that can be detected at frequencies as high as 1 MHz. While the optical emission from fluorescent lights is strongest in the visible and near-ultraviolet region, significant emission also occurs in the near-infrared, especially at 1017 nm, a spectral line of mercury.

Interference due to fluorescent lighting may be controlled by a judicious combination of optical and/or electrical filtering. Longpass optical filtering reduces fluorescent-light interference, while narrow bandpass optical filtering at a carefully chosen near-infrared wavelength provides an even greater reduction of interference. Nonetheless, because of residual interference, it is usually necessary for the link to employ a passband modulation scheme that avoids the near-d.c. interference, such as PPM or multiple-subcarrier modulation (MSM). Alternatively, the link can utilize a baseband modulation technique, such as on-off keying (OOK), in conjunction with a highpass electrical filter that removes the interference. In this case, line coding or active baseline restoration is needed to prevent baseline wander. These modulation techniques will be discussed in the following section.

We will now discuss how to design the receiver optics in order to achieve a high SNR in the face of steady background illumination. A well-designed receiver will be shot-noise-limited under conditions of bright illumination, so we will assume that background-induced shot noise dominates. Under these conditions, it is appropriate to utilize a simple positive-intrinsic-negative (p-i-n) photodiode [19] rather than the avalanche photodiodes that are used in IM/DD receivers when background illumination is weak [20]. It is desirable to use a large photodetector area; we will see that the shot-noise-limited SNR is proportional to the detector area A. However, large-area detectors have high capacitance, which can limit receiver bandwidth and greatly increase receiver thermal noise [20]. Accordingly, it is desirable to reduce the required physical detector area by use of an optical concentrator, which accepts light from a large collection area and concentrates it to the somewhat smaller detector area. One type of concentrator, the dielectric compound parabolic concentrator (CPC), has been used in prototype free-space infrared links [21, 17]. The CPC is an angle-transforming device, which can concentrate light from a large input area A_{in} down to a smaller detector area A_{det}, yielding an optical gain of A_{in}/A_{det}.

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In doing so, however, it reduces the receiver solid angle of acceptance by the ratio Ω_{in}/Ω_{det} = n^2 A_{det}/A_{in}, where n is the CPC refractive index. Receivers using a single CPC-based element may be excessively directional for many applications of non-directed infrared links. An array of differently oriented CPC elements may be used to construct an angle-diversity receiver [15, 21], which can use selection or combining techniques to improve SNR or reduce the impact of multipath distortion. A second type of optical concentrator is the dielectric truncated spherical concentrator, of which a special case is the dielectric hemisphere [15, 22]. A dielectric hemisphere of index n can provide an optical gain of n^2 that is nearly omnidirectional, as long as the hemisphere radius R is at least n^2 times the detector radius r, and the two are suitably indixed. For many applications, it is desirable to build a simple receiver containing a single, omnidirectional element, and we will restrict our attention to receivers that utilize a hemispherical concentrator.

The best rejection of ambient light can be achieved using a spectrally narrow source, such as an LD, in conjunction with a narrow optical bandpass filter. Efficient, narrowband optical interference filters can be constructed using multiple layers of thin dielectric films [23]. Unfortunately, the phase shift through the dielectric layers changes with angle of incidence, so that for non-normal incidence, the filter passband shifts to shorter wavelengths. Fig. 7 illustrates the angle-dependent passband of a typical 29-layer filter having a 30-nm half-power bandwidth at normal incidence. It is seen that the passbands for θ = 0° and θ = 30° barely overlap.

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<table>
<thead>
<tr>
<th>Parameter</th>
<th>Hemispherical filter</th>
<th>Planar filter</th>
</tr>
</thead>
<tbody>
<tr>
<td>Filter bandwidth $\Delta \lambda$</td>
<td>11.6 nm</td>
<td>71.7 nm</td>
</tr>
<tr>
<td>Required average transmitter power</td>
<td>224 mW</td>
<td>492 mW</td>
</tr>
</tbody>
</table>

Table 2. Optimized filter bandwidth and required average transmitter power in 100 Mbps OOK links. LOS source is optimized to cover a 5 m x 5 m room using minimum total power. A non-distorting channel, bright skylight $P_{sky} = 5.8 \mu W/(cm^{-2}-nm)$ and BER $= 10^{-5}$ (SNR$_B = 144$) are assumed. Other parameters include: detector area $A = 1 cm^2$, concentrator gain $G = 4.6$ dB, wavelength $\lambda = 810$ nm, and responsivity $r = 0.53$ A/W (from [22]).

Fig. 8 displays two different ways that a thin-film interference filter can be combined with a hemispherical concentrator and large-area detector. As shown in Fig. 8(a), it is straightforward to place the filter between the hemisphere and detector. It has been determined through ray tracing that in this geometry, the angle-dependent filter passband makes it possible to simultaneously achieve the desired properties of narrow optical bandwidth and wide FOV [22]. Alternatively, an optical bandpass filter can be deposited or bonded onto the outer surface of the hemispherical concentrator, as shown in Fig. 8(b). Ray tracing calculations [22] show that regardless of the angle $\psi$ from which the signal is received, rays that reach the detector are incident upon the filter at angles $\theta$ between $0^\circ$ and $\theta_{max}$, where

$$\theta_{max} = \sin^{-1}(n/R)$$

Typically, $\theta_{max}$ is less than $30^\circ$. Thus, with a hemispherical filter, it is possible to simultaneously obtain a narrow bandwidth and wide FOV.

With the aim of optimizing the receiver optical design, we will calculate the SNR of a shot-noise-limited receiver, explicitly considering the effects of the concentrator gain, bandpass filter properties, and detector area. We consider a link using OOK modulation over a channel that is free of multi-path distortion. In the following section, the performance of OOK on the ideal channel will be used as a reference for evaluating the power efficacy of OOK and other modulation techniques on multipath channels. We assume that the OOK transmission, at bit rate $1/T$, consists of rectangular pulses of duration $T$ having a peak power of either 0 or $P_p$. Assuming that ones and zeros are equally probable, the average power is $P_{ave} = P_p/2$. The channel, being ideal, has a gain $H(\theta)$ at all frequencies. The receiver employs a concentrator with optical gain $G$ and an optical filter that transmits the signal with efficiency $\eta_{sig}$, which represents the average over rays striking the filter at different angles with respect to its surface normal. We emphasize that $\eta_{sig}$ depends on the signal angle of incidence $\psi$. The optical signal is detected by a p-i-n photodiode of area $A$ and responsivity $r$. The detected current is passed through a filter having rectangular impulse response of duration $T$ and peak value $1/T$, and the filter output is sampled. The resulting samples have values of $0$ or $P_p H(0) \eta_{sig} (\psi) Ar T$. Incident upon the receiver is a background flux of density $p_{bg}$ which has units of (W/cm$^2$-nm), and is assumed to be constant over the filter passband. The filter has peak transmission $\eta_{sig}$ and optical noise bandwidth $\Delta \lambda$. Including the concentrator gain $G$, a background irradiance $p_{bg} \eta_{sig} \Delta \lambda$ is incident on the detector. This induces a shot-noise current of two-sided PSD $p_{bg} \eta_{sig} \Delta \lambda$ $Ar$ where $e$ is the elementary charge, resulting in a sampled filter output with variance equal to this PSD. The peak receiver SNR is thus:

$$SNR_p = \frac{P_p H^2(0) \eta_{sig} (\psi) Ar T}{q \rho \eta_{sig} \Delta \lambda},$$

and the bit-error rate (BER) is given by

$$BER = Q\left\{SNR_p \right\},$$

where $Q(\cdot)$ is the Gaussian Q function [24]. For example, the peak SNR should have a value of 144 (21.6 dB) to achieve a BER of $10^{-9}$.

Examining (5), we see that the SNR is improved by utilization of a detector having a large area $A$ and

---

**Figure 8.** Cross-sectional view of receiver structures that attempt to achieve wide field of view and narrow passband. The hemispherical concentrator of refractive index $n$ provides a nearly omnidirectional optical gain of approximately $n^2$, as long as its radius exceeds $n^2$ times the detector radius. Structure (a) uses a flat optical bandpass filter; rays arriving at angle $\psi$ strike the filter at angles $\theta$ that are close to $\psi$. Structure (b) uses a hemispherical filter; independent of the direction $\psi$, from which light is received, all rays reaching the photodiode strike the filter at angles $\theta$ that are near normal incidence — typically between $0^\circ$ and $30^\circ$ (from [22]).
the highest available responsivity \( r \) in conjunction with a concentrator having high optical gain \( G \). The optical filter should be designed so as to maximize the figure of merit \( \eta_{opt} / \Delta \lambda \). To first order, this implies that one should choose the narrowest possible filter bandwidth \( \Delta \lambda \) that still achieves high signal efficiency \( \eta_{opt} \). Assuming a hemispherical filter, as in Fig. 8(b), this can be achieved approximately as follows. One should choose the filter center wavelength so that for \( \theta = 0 \), the short-wavelength edge of the filter passband lies at the signal wavelength. The filter bandwidth should then be chosen so that for \( \theta = \theta_{max} \), the long-wavelength edge of the passband is at the signal wavelength. For example, the filter of Fig. 7 has been designed using this procedure for a signal wavelength of 805 nm and for \( \theta_{max} = 28^\circ \). In fact, the shot-noise-limited SNR is maximized by choice of an even narrower filter bandwidth, which reduces the power that but rejects more ambient light. We have simultaneously optimized the filter characteristics and transmitter radiation pattern of a LOS link to minimize the total transmitted power while yielding the required SNR everywhere within the transmitter coverage area [22]. Table 2 presents results for a 100 Mbps OOK link operating over a 5 m x 5 m area. The optimized hemispherical-filter bandwidth is 11.6 nm, and the required average transmitter power is 224 mW. When a flat filter is utilized instead, the optimum filter bandwidth is increased to 71.7 nm, requiring 492 mW of transmitter power, a 3.4-dB increase.

Receiver preamplifier design is a crucial consideration for high-speed infrared links, and is discussed in detail in [22]. Among candidate preamplifier designs [20], the transimpedance amplifier probably is the best choice for most applications, due to its superior dynamic range. The preamplifier should be designed so that, under conditions of the brightest ambient illumination to be encountered, the receiver achieves sufficient SNR and is shot-noise limited, or nearly so. One should choose a p-i-n photodiode having low capacitance per unit area, so minimize a strong component of receiver noise whose PSD is proportional to the square of receiver input capacitance and to the square of frequency. Unfortunately, low capacitance per unit area requires that the photodiode be relatively thick, and this may lead to a transit-time limitation of receiver bandwidth [19], especially for devices illuminated through the \( n \) contact. It is also important to choose a photodiode having low series resistance. Analysis of FET-based preamplifiers [22] shows that the front-end FET should be chosen to have high transconductance \( g_{m} \); the FET cutoff frequency \( f_{c} \) is important mainly in that the FET gate capacitance should be smaller than the photodiode capacitance.

### Performance of Modulation Techniques

Non-directed infrared channels can be modeled as fixed, linear systems with additive, white, Gaussian noise, as summarized in (1). As it represents instantaneous optical power, the channel input \( X(t) \) must be nonnegative, and it is related to the average transmitted optical power by (2). The channel, described by impulse response \( h(t) \), exhibits multipath distortion that can induce intersymbol interference (ISI) in high-bit-rate links. When the ISI is relatively mild, it leads to an optical power penalty, but if it is severe, it may lead to a BER floor.

In evaluating candidate modulation techniques, the most important criterion is the optical average-power requirement, as it generally corresponds to transmitter electrical power consumption and optical hazard. At high bit rates, one must consider the effect of multipath ISI on this power requirement, as well as any reduction of the ISI that can be achieved through techniques such as adaptive equalization. The second most important attribute is the receiver electrical bandwidth requirement, as it can be difficult to achieve flat frequency response and low noise over a wide bandwidth using large-area photodiodes. Other important criteria for comparison of modulation techniques are the complexity and power consumption of a portable receiver.

Fig. 9 presents a comparison of the average power and bandwidth requirements [22] of a number of modulation schemes, assuming a distortionless channel. The power and bandwidth requirements are normalized to those of baseband OOK. These modulation techniques include single-carrier schemes, such as L-PPM and \( L \)-pulse-amplitude modulation (L-PAM), of which an important special case is 2-PAM (OOK). They also include MSM schemes that employ \( N \) subcarriers modulated using binary or quadrature phase shift keying (BPSK or OQPSK), and 16- or 64-quadrature-amplitude modulation (QAM). In the following section, we discuss the performance of several of these modulation techniques, emphasizing their power requirement in the face of multipath ISI, which we have evaluated using a large set of measured channel impulse responses. Our reference point for comparison will be the power requirement of OOK on an ideal channel, which was calculated in the previous section.

It is evident in Fig. 9 that OOK represents a

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3 We note that the electrical bandwidth requirement of a modulation technique has little bearing on the electrical bandwidth occupied by an \( I \) signal. This optical bandwidth is dominated by the large spectral spread of practical infrared sources. For example, a 1-mm width corresponds to 469 GHz, assuming a wavelength of 800 nm.
good compromise between power requirement and bandwidth requirement, at least on the ideal, distortionless channel. When considering the impact of multipath ISI, the power requirement of OOK links can be calculated using the technique described in [16]. The impulse responses of non-directed infrared channels we have characterized [12] contain significant energy only within the first 100 ns after their first zero-crossing. It is thus not surprising that a bit rate of 10 Mb/s, where the bit duration is 100 ns, ISI induces little impairment. For example, normalized power requirements of 10-Mb/s OOK links on shadowed, diffuse channels do not exceed 1.0 dB, as indicated by the symbols in Fig. 10. This figure also displays the power requirement of links using several different types of L-PPM, which will be discussed below. Clearly, unequalized OOK transmission is feasible at 10 Mb/s. At higher bit rates, OOK links incur more significant degradation from multipath ISI [12]. For example, at 30 Mb/s, normalized power requirements on shadowed, diffuse channels do not exceed 4.3 dB. While relatively high, these power requirements may be acceptable for some applications, particularly when transmitting from a base station. At a bit rate of 100 Mb/s, however, very large power requirements or even BER floors are incurred on all four types of channels (LOS and diffuse, with and without shadowing), suggesting that unequalized OOK is not practical in this case. Fig. 11 displays the normalized power requirements of unequalized OOK links transmitting on all four channel types at bit rates of 30 and 100 Mb/s, denoted by o and + symbols, respectively. It is seen that there is a systematic relationship between the normalized power requirement and the normalized channel delay spread, which is the channel delay spread (τ) divided by the bit duration T. Despite the fact that the various impulse responses differ in their detailed features, for purpose of calculating the impact of ISI on OOK links, channels of all four types can be roughly described by a single-parameter model.

A practical, though sub-optimal, means to reduce the multipath ISI penalty is by using a decision-feedback equalizer (DFE) that can adapt automatically to the channel impulse response [24]. Considering measured channel responses, the power requirements of OOK links using DFEs have been evaluated in [12]. At 30 Mb/s, normalized power requirements for all channels are less than 2.3 dB. Thus, OOK with a DFE is a feasible technique for transmission at 30 Mb/s. At 100 Mb/s, no BER floors are observed with a DFE, in contrast to the unequalized case. With a DFE, power requirements for unshadowed LOS and diffuse channels are less than 6.7 and 5.7 dB, respectively, while those for shadowed LOS and diffuse channels do not exceed 9.1 and 7.1 dB, respectively. These power requirements, while high, may be small enough to make OOK with a DFE practical at 100 Mb/s, particularly for transmission from a base station. Fig. 11 presents the normalized power requirements of OOK links with a DFE operating at 30 and 100 Mb/s, denoted by x and symbols, respectively, transmitting over all four channel types. Again, we observe a systematic relationship between the normalized power requirement and normalized delay spread.

While we have considered zero-forcing DFEs to simplify calculation of the power requirements, their performance at low BER is virtually identical to that
of DFEs adapted according to the minimum-mean-square error criterion [24, 25]. In practice, a DFE typically would be implemented using digital or discrete-time analog signal-processing techniques. Numerical simulation [25] of 100 Mbits/s OOK with constrained-complexity DFEs has shown that little degradation is observed when the forward and reverse filters are three T/2-spaced taps and five T-spaced taps, respectively, or by employing four-bit quantization of tap weights. Adaptation to a training sequence using the least-square algorithm is found to occur within about 200 bits, i.e., about 2 μs. Most indoor applications of non-directed infrared links will not involve rapidly moving receivers, so that channel impulse responses will change significantly only on time scales of tens to hundreds of μs. Thus, we expect that once it is initially adapted to the channel response, a DFE should be able to track easily any changes in that response.

As mentioned above, fluorescent lighting may induce near-d.c. interference in wireless infrared receivers. The impact of this interference on a baseband OOK receiver may be reduced by electrical highpass filtering of the preamplifier output. To avoid excessive baseline wander in a baseband OOK link, however, the highpass filter cut-on frequency cannot be higher than about 10^3 of the bit rate [26], unless line coding or active baseline restoration is utilized [25].

Considering L-PAM in general, as, for example, L is increased from 2 (recall that OOK is equivalent to 2-PAM), there is a monotonic decrease in bandwidth requirement, at the expense of a monotonic increase in power requirement on an ideal channel (Fig. 9). For a given bit rate, as L increases and the symbol duration increases, the decrease in noise admitted by the corresponding receiver filter is not sufficient to overcome the reduction in noise immunity associated with closer spacing of signal levels, as in conventional systems [24]. Considering multipath channels, one might hope that as L increases and, for a given bit rate, the symbol duration increases relative to the channel delay spread, a reduction in ISI penalty might overcome the poor inherent power efficiency of higher-level L-PAM, making it more efficient than OOK. Unfortunately, this is not the case. For example, at 30 Mbits/s, unqualified 4-PAM has a normalized power requirement of 5.3 dB, compared to 2.7 dB for unqualified 2-PAM (OOK), averaged over 17 diffuse, shadowed channels. These two power requirements are compared in Fig. 12, where they are indicated by * symbols. This figure also presents the power requirements of several other schemes, which will be discussed below. We conclude that OOK is probably the best L-PAM technique for most applications.

L-PAM is a transmission technique [22, 26] that offers an improvement in average-power efficiency over OOK, at the expense of an increased bandwidth requirement (Fig. 9). This technique utilizes symbols consisting of L time slots, which we will refer to as chips. A constant power L_Pmean is transmitted during one of these chips (P_{avg} is the average transmitted optical power), and zero power is transmitted during the remaining (L-1) chips, thereby encoding log L bits in the position of the non-zero chip. For a given bit rate, L-PAM requires more bandwidth than OOK by a factor L/log_2L, e.g., 16-PAM requires four times more bandwidth than OOK. On distortionless channels, L-PAM yields a decrease in average-power requirement that improves steadily with increasing L; the increased noise associated with a (L/log_2L)-fold wider receiver noise bandwidth is outweighed by the improved noise immunity arising from a L-fold increase in peak power. The excellent average-power efficiency of L-PAM can result in a significant decrease in transmitter power consumption, making the technique especially useful for portable transmitters. Because the power spectra of L-PAM vanishes at d.c. for all values of L, it is possible to pass a received L-PAM waveform through a highpass filter having a cut-on frequency as high as 0.05 to 0.10 times the bit rate with little baseline wander [26]. This makes L-PAM an excellent choice in the presence of near-d.c. interference from fluorescent lighting. Two drawbacks of L-PAM as compared to OOK should be noted, i.e., an increased transmitter peak-power requirement, and the need for more precise synchronization.

When L-PAM is transmitted over multipath channels, the non-zero transmitted chips can induce interference in chips both within the same symbol (intrasyMBOL interference) and in adjacent transmitted symbols (intersymbol interference); we will refer to these effects collectively as ISI. For transmission at a fixed bit rate on a given channel, the impact of ISI should generally increase steadily with increasing L, due to the decrease in duration of the L chips within each symbol. As L is increased, the ISI penalty will eventually overcome the inherent average-power efficiency associated with

![Figure 12. Optical average-power and electrical-bandwidth requirements of several multiple-subcarrier (MSC) and pulse-amplitude modulation (PAM) schemes.](image-url)
large $L$. We have analyzed the impact of ISI on this transmission technique, and have evaluated the power requirement of unequaled $L$-PPM links using measured channel impulse responses [26]. The average-power requirement of OOK and several $L$-PPM orders are compared in Fig. 10, for transmission at 10 Mbit/s over diffuse, shadowed channels. At this bit rate, while $L$-PPM links incur ISI penalties that increase with increasing $L$, 16-PPM still yields superior average-power efficiency on all four types of channels (LOS and diffuse, with and without shadowing). At the higher bit rate of 30 Mbit/s, the same general trends are evident. However, the optical power requirements are higher, and the power requirement does not decrease steadily with increasing $L$ for all channels. For the shadowed LOS configurations, higher PPM orders $L \geq 4$ incur BER floors on several channels, demonstrating once again that for applications where shadowing is probable, a diffuse system is more robust. At a bit rate of 100 Mbit/s, excessive or infinite power penalties are incurred on all channels except shadowed LOS channels, where 8-PPM is consistently found to yield the best power efficiency.

Given its inherent power efficiency, it would be desirable to develop techniques to mitigate the impact of multipath ISI on $L$-PPM, in the hope that this might permit reliable transmission at very high bit rates. Recently, Barry has discussed maximum-likelihood sequence detection of $L$-PPM in the presence of ISI [27], and has shown that it can be implemented by generalization of the Viterbi algorithm to the case of a vector channel. He has also discussed several sub-optimal adaptive equalization techniques, which include linear and decision-feedback equalizers operating at either the chip or symbol rates, as well as hybrid DFEs that make use of tentative chip decisions to cancel intrasymbol interference, but use more reliable symbol decisions to cancel intersymbol interference. We are currently using measured channel characteristics to quantify the performance of these detection and equalization schemes.

MSM is a technique that promises the flexibility of frequency-division multiplexing and multi-level modulation, while maintaining the simplicity of IM/DD [22, 28]. In MSM, several independent bit streams are modulated onto carriers at several frequencies (say, of the order of 1 to 100 MHz). The modulated subcarriers are summed together to form a frequency-division multiplexed signal, and this signal is used to modulate the intensity of an optical transmitter. After transmission and optical-to-electrical conversion, the individual bit streams can be recovered using multiple bandpass demodulators. Through simultaneous transmission of several narrow-band subcarriers, MSM may make possible very high aggregate bit rates without requiring adaptive equalization to overcome ISI, and may allow individual receivers to process only a subset of the total transmission. While MSM is less power-efficient than OOK or $L$-PPM, it may be well-suited for transmission of multiplexed bit streams from a base station to a collection of several portable receivers.

The optical average-power and bandwidth requirements of several MSM schemes on ideal channels have been derived in [22], and are summarized in Fig. 9. A single BPSK or QPSK subcarrier requires 1.5 dB more optical power than OOK; BPSK and OOK transmissions of average power $P_{av}$ are equivalent to binary antipodal signals plus a d.c. bias $P_{dc}$ carrying no information, but the BPSK waveform uses sinusoidal pulses having 3-dB less electrical power, thus requiring 1.5 dB more optical power for achievement of the same receiver SNR. Single 16- and 64-QAM subcarrier schemes are less power-efficient than BPSK and
QPSK for conventional reasons [24]. An N-subcarrier transmission requires more average power than the corresponding single-subcarrier scheme by a ratio that increases steadily with increasing N, because the amplitude of each subcarrier must not exceed $P_{\text{avg}}/N$, to assure that the transmitted optical-power waveform $X(t)$ is non-negative.

When MSM is transmitted over a multipath channel, several effects further degrade its power efficiency, as compared to transmission of OOK over an ideal channel [28]. As multipath channels are generally lowpass in nature, subcarriers are subject to an attenuation that generally increases with increasing subcarrier frequency. In addition, subcarriers may overlap and interfere both in phase and quadrature phases of one subcarrier, and interference between adjacent subcarriers that may overlap partially in frequency. To reduce these three interferences, it is desirable to use a large number of subcarriers, but this leads to an excessive penalty for large N, as described above. We have evaluated the performance of a large number of different MSM schemes, for transmission at total bit rates of 30 and 100 b/s over measured multipath channels [28]. The amplitudes of all subcarriers were maintained equal, and the average optical power was scaled as necessary to achieve the required BER. We found that the best performance was achieved typically by two and four-subcarrier formats. For example, Fig. 12 presents the power and bandwidth requirements of the MSM schemes that yielded the best BER-floor-free, average-power performance at 30 Mb/s. Averages over 17 diffuse, shadowed channels, the best normalized power requirement of 6.5 dB was achieved by two QPSK subcarriers using 100 percent-excess bandwidth, root-raised-cosine pulses. While this is 5.2 dB more than the power required using OOK with a DFE, the increased power may be accommodated by a base-station transmitter. At a bit rate of 100 Mb/s, the best MSM format, 2-QPSK, yielded a normalized power requirement of 11.1 dB, averaged over the same 17 channels. Improvement in MSM performance, at the expense of a variable transmission bit rate, can be achieved through dynamic carrier selection, i.e., by not transmitting at a subcarrier frequency at which the channel attenuation and/or dispersion is excessive.

**Experimental 50-Mb/s Diffuse Infrared Link**

Over the past 18 months, we have been designing and constructing an experimental link to test the performance limits of high-speed non-direct infrared communication. In its initial form, this prototype uses baseband OOK transmission at 50 Mb/s [29]. Our transmitter uses a cluster of eight LDs whose output is passed through a translucency plate, a diffuse reflector, to create an approximately Lambertian radiation pattern having 475-mW average power at a wavelength of 805 nm. In typical operation, the transmitter emission is directed upward toward the ceiling, creating a diffuse link configuration. A block diagram of the receiver is shown in Fig. 13. The receiver employs an optical "antenna" of the design shown in Fig. 8(b). A 1-cm² silicon-photodiode detector is index-matched to a hemispherical concentrator of 2-cm radius, having a refractive index of 1.76. An optical bandpass filter designed to have the wavelength-dependent transmission characteristics shown in Fig. 7 is bonded to the hemisphere's outer surface. This filter-concentrator combination achieves a bandwidth of 30 nm, a net gain of 1.5 dB, and a FOV of 65°. The photodiode capacitance of 35 pF, in conjunction with the preamplifier load resistance of 10 kΩ, leads to a 455-kHz pole that is compensated by a passive R-C circuit. This receiver achieves a 3-dB cutoff frequency of 25 MHz, which is limited by the transit time of holes across the depletion region of the photodiode, which is illuminated through the n contact. Our receiver has an equivalent input noise current density (one-sided) of 7.8 pA/√Hz, averaged over the bandwidth of the 25-MHz Bessel lowpass filter. Residual interference from fluorescent lighting is removed using 1.6-MHz, single-pole highpass filter, and quantized feedback through a 1.6-MHz, single-pole lowpass filter is used to prevent baseline wander. In order to reduce the impact of multipath ISI, our receiver employs a DFE, with the forward and reverse filters realized using cable delays and manually adjustable, variable-gain amplifiers. Both filters have four taps; those of the forward filter are half-band-spaced, while those of the reverse filter are banded-spaced.

Fig. 14 presents measurements of the BER achieved by a diffuse link transmitting over a 3-m horizontal range, with no ambient light present. It is seen that the DFE is extremely effective in mitigating multipath ISI, reducing the optical power penalty by 2.5 dB (equivalent to an SNR improvement of 5 dB). In the absence of illumination, the receiver sensitivity is -32.5 dBm/cm² at 10⁻⁷ BER. Fluorescent lighting induces an optical power penalty of only 0.1 dB, while bright sunlight leads to a penalty up to 1.5 dB. In the presence of bright sunlight, the link achieves a horizontal range of 2.9 m at 10⁻⁷ BER.
Interference in Multi-User Systems

As mentioned earlier, a key advantage of infrared LANs is that interference can arise only between users not separated from each other by opaque boundaries. In this section, we discuss briefly how to accommodate multiple infrared links within a single room. Further discussion on this topic can be found in [4, 22].

Ad hoc LANs, formed by direct peer-to-peer communication between two or more portable devices in the same room, typically operate using a single shared channel, e.g., one transmission wavelength accommodating packetized, L-PPM transmission [8]. Such a LAN can be operated very efficiently using a well-known protocol, such as CSMA-CA, though protocol modification is sometimes necessary to provide for the possibility of “hidden nodes,” i.e., situations in which some of the colocated portables cannot communicate with each other.

LANs that involve duplex communication between portables and base stations can employ a single transmission wavelength for uplink and downlink communications. Single-carrier modulation schemes, such as OOK and LPPM, will provide a single channel that must be shared by uplinks and downlinks using an appropriate protocol. Multiple-carrier modulation schemes, such as MSM or hybrid baseband/MSM schemes, can provide multiple, non-interfering channels for separate use by uplinks and downlinks. The single-wavelength approach has the advantage that peer-to-peer communication between portables is also possible, making use of the same portable receiver used for downlink reception. Use of separate wavelengths for uplink and downlink transmission can increase the capacity of LANs involving uplink and downlink communications. When such wavelength multiplexing is used, however, portables must be able to receive at two wavelengths if peer-to-peer communication between portables is also desired.

In LANs that utilize base stations, rooms up to 10 m x 10 m in size, or even larger, may be served by a single base station, depending on uplink and downlink transmission range. In this case, downlink transmissions will not be subject to interference from other downlinks, and will be essentially free of interference if the LAN employs wavelength multiplexing of uplink and downlink transmissions. Downlinks may interfere with each other in larger rooms that require multiple base stations. In this case, the greatest aggregate downlink capacity can be achieved by partitioning the room into non-overlapping regions covered by different base stations, taking advantage of the very rapid falloff of interference with distance (see below). Unfortunately, this approach will introduce “dead zones” between base-station coverage regions. In rooms that can be covered by two or three base stations, it is possible to achieve full coverage by having all base stations transmit in unison, but this may lead to an unacceptable increase of multipath distortion. The best general means of achieving full coverage of large areas is probably to adopt a cellular approach, partitioning the downlink into time intervals (with single-carrier modulation schemes) or frequency bands (with MSM schemes) that are not used in adjacent cells, but that may be reused in distant cells.

It is worthwhile to discuss briefly the modeling of interference in non-directed infrared systems. Consider M simultaneous IM transmissions X(t), j = 1, ..., M, which are incident upon a DD receiver. Let h_j(t) denote the impulse response of the channel between transmitter j and the receiver. Then the total received photocurrent Y(t) is given by

\[ Y(t) = \sum_{j=1}^{M} X_j(t) \otimes h_j(t) + n(t). \]  

We emphasize that Y(t) is linear in each of the IM envelopes X(t), and that there is no need to consider the relative phases of the underlying optical carriers. The derivation of expression (7) is a simple generalization of the derivation of (1), which is provided in [12].

As an example of a multuser system with interference, we will consider the case of OOK transmission. We assume that the bit rate is low enough that channel dispersion can be neglected. The desired signal, of peak power P_d, transmitted over a channel having d.c. gain H(0). There are K-independent, synchronized interferers, for which the corresponding quantities are P_i and H_i(0), j = 1, ..., K. We assume that ones and zeros are equiprobable in all transmissions. Using (7), it is straightforward to derive the peak SIR:

\[ SIR = \frac{4P_d^2H^2(0)}{\sum_{j=1}^{K} P_i^2H_j^2(0)}. \]

The numerator and denominator of (8) are proportional to equivalent electric powers, as one would expect. We note, however, that the equivalent d.c. electrical power gains H^2(0) and H^2(0) are equal to the inverse squares of the respective optical path losses. For example, suppose that the path loss increases as the fourth power of distance (Fig. 5). In this case, the SIR increases proportional to the eighth power of the interferer separation, in contrast with the fourth-power increase that would be seen in a system using linear detection techniques. While this advantage of wireless infrared links using IM/DD is not yet widely appreciated, it may lead to significant enhancement of infrared LAN capacity, as compared to LANs employing radio links.

Summary and Conclusions

In this article we have reviewed the advantages and drawbacks of non-directed infrared radiation as a communication medium for indoor wireless LANs. The physical characteristics of IM/DD infrared channels, including path loss and multipath distortion, were described. We discussed techniques
for achieving a high receiver SNR in the face of high path loss and intense ambient infrared radiation. Several promising modulation formats were described, including OOK, JPPM, and MSM, and we quantified the performance of these techniques on real multipath channels. We have described preliminary tests of a 50-Mb/s diffuse infrared link, showing that OOK transmission with a DFE is feasible at this bit rate. Finally, we discussed interference in multi-user infrared systems.

At this writing, non-directed infrared LANs operating at bit rates up to 4 Mb are commercially available. The prospects appear good that much higher transmission speeds can be achieved, and that infrared will play a significant role in future high-capacity indoor wireless LANs. In order for infrared LANs to reach their full potential, much R & D work remains to be performed. Ferile research topics include optimization of transmitter and receiver optics, equalization techniques for JPPM, diversity reception techniques, techniques to mitigate interference in multi-user systems, and low-power implementation of relevant analog and digital circuitry.

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References

[6] For example, a bit rate of 125 Mbps over a 30-m range, using transmitter and receiver fields of view of 1° x 1° and 6° x 6°, respectively, is achieved by the IL-889-15 transmitter manufactured by Jolta, Ltd., Israel.
[9] Photronics Corporation San Jose, CA.

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