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Abstract. We consider indoors communications networks using modulated LEDs to transmit the information packets. A generic indoor channel equalization formulation is proposed assuming the existence of both line of sight and diffuse emitters. The proposed approach is of relevance to emergent indoors distributed sensing modalities for which various lighting based network communications protocols are considered.

1. Introduction

With the wide proliferation of RFIDs, smart-phones and other portable communication devices, as well as distributed sensing modalities proposed by sensors manufacturers for pervasive sensing in medical/healthcare applications (e.g, hospitals), in domestic electronics e.g. for smart homes as well as in environmental sensing e.g. food/agriculture applications within controlled environments (greenhouses, plant growth rooms super market shelves etc) it is realised that current radio communication links, have limited bandwidth and are unlikely to be able to cope with the large datasets generated in the long run. Although terahertz links have been suggested for indoors communications [1,2], a simpler solution for setting up multi-hop local indoor communications networks is to use the existing lighting infrastructure in a building and modulate indoor lighting appropriately to enable non-directed communications in an ad-hoc manner. LEDs are well suited to such applications as they are cheap, robust, more efficient from an energy consumption perspective and can be very quickly modulated in the time domain to transmit data packets at high data rates. They are also rapidly becoming a preferred lighting solution by many architect firms. LED-based indoor communications can be conveniently achieved assuming direct line-of-sight or diffuse links.

Optical wireless systems use simple baseband modulation schemes such as on-off keying (OOK), or pulse position modulation (PPM) [3, 4]. Alternatives that are superior from a power efficiency perspective include digital pulse-interval modulation (DPIM) [5], dual header pulse-interval modulation (DHPIM) [6] and differential pulse-position modulation (DPPM) [7]. Differential amplitude pulse-position modulation (DAPPM) is a new hybrid modulation technique that has attracted interest recently [8] as it incorporates aspects from both pulse-amplitude modulation (PAM) as well as DPPM. Demodulation of the encoded signals is commonly performed using hard-threshold decision (HTD), maximum-likelihood sequence detection (MLSD) or zero-forcing decision-feedback

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equalization (ZF-DFE) modalities assuming intensity demodulation is performed using a direct detection technique [8].

Line-of-sight links can assume a simple exponentially decaying channel model whereas diffuse links require a ceiling-bounce channel model [9, 10]. In the time domain, the output $y(t)$ from the communications channel is given by:

$$y(t) = Rh(t) * x(t) + n(t) \tag{1}$$

where $h(t)$ is the channel impulse response, R is the detector responsivity, $n(t)$ is white Gaussian noise due to the lighting in the room and $*$ denotes the convolution operator. A ceiling-bounce model such as the one assumed by Carruthers and Kahn [11] can be used for the channel model:

$$h(t,a) = L_0 \frac{6a^6}{(t+a)^7} u(t) \tag{2}$$

where $u(t)$ is the unit step function and a depends on the room size, and the transmitter and receiver position. The parameter L_0 is given from:

$$L_0 = \frac{rA_r}{3\pi L^2} \tag{3}$$

where L is the height of the ceiling, r is the reflectivity of the ceiling above the transmitter and receiver and c is the speed of light. The parameter $a = 2L/c$ is related to the rms delay spread D_{rms} through the following expression [8]:

$$D_{rms} = \frac{a}{12}\sqrt{\frac{13}{11}} \tag{4}$$

In the following sections, a generic channel equalization scheme with amplitude, slew rate and power constrains is developed for the above channel.

2. Optimization of the input waveform

The theory of matched filters [12] could be adopted to customize the excitation waveform in order to either maximize the efficiency of energy transfer from the transmitter to the receiver and improve the signal-to-noise ratio of the data in frequency regions where the channel identification used for the equalization process is ill-conditioned. The problem of maximizing the power transfer can be stated as follows. Given the Laplace transform $H(\omega)$ of $h(t)$, find the Laplace transform $U(\omega)$ of $u(t)$ that maximizes the following function:

$$J = \int_0^\infty |H(\omega)U(\omega)|^2 \, d\omega \tag{5}$$

subject to the following restrictions related to the capabilities of the emitter:

$$|U(\omega)| \le M(\omega), \forall \omega > 0 \tag{6}$$

where $M(\omega)$ is an upper bound for the spectral profile of the input pulse at each frequency and the restriction on the overall power of the generator is given from:

$$\int_0^{\infty} |U(\omega)|^2 \, d\omega \leq P_{\max} \tag{7}$$

where $M(\omega)$ is the maximum rms value that the generator is able to deliver at frequency ω and P_{max} is the maximum overall power output of the emitter. This problem can be re-stated by introducing a variable $Z(\omega) = |U(\omega)|^2$ so that the objective function becomes:

$$J = \int_0^{\infty} |H(\omega)|^2 \, Z(\omega) d\omega \tag{8}$$

subject to:

$$Z(\omega) \leq M^2(\omega), \forall \omega > 0 \tag{9a}$$

$$\int_0^{\infty} Z(\omega) d\omega \leq P_{\max} \tag{9b}$$

Under the hypothesis that H is known, Eq. (8) is a linear functional of $Z(\omega)$, which must be minimized subject to the linear constraints expressed in (9a) and (9b). This optimization problem could be solved by using variational calculus techniques. Alternatively, by discretizing the frequency axis and considering discrete frequencies up to a certain maximum ω_{max}, the problem becomes a finite-dimensional Linear Program, which can be solved in an efficient manner by techniques such as the Simplex method.

A different problem would be to shape the pulse in order to obtain a specified pulse shape at the output of the receiver. This problem can be formulated as follows. Let $Y_d(\omega)$ be the frequency profile of the desired output pulse. Given $H(\omega)$, find $U(\omega)$ that minimizes the following function:

$$J' = \int_0^{\infty} |Y_d(\omega) - H(\omega) U(\omega)|^2 \, d\omega \tag{10}$$

subject to the generator restrictions:

$$|U(\omega)| \leq M(\omega), \forall \omega > 0 \tag{11a}$$

$$\int_0^{\infty} |U(\omega)|^2 \, d\omega \leq P_{\max} \tag{11b}$$

This problem is more complex than the previous one, because the cost function J' cannot be transformed in a linear function by a change of variables, as it was done previously. In this case, the problem must be solved by numerical nonlinear programming techniques. The formulation can be regarded as a pre-equalization process. In fact, instead of transmitting the desired pulse shape, a different pulse shape is realized in order to account for the distortion caused by the transmission through the devices and channel under consideration. It is worth noticing that, if no restrictions are placed on the generator capabilities, then the solution to this problem is:

$$U(\omega) = \frac{Y_d(\omega)}{H(\omega)} \tag{12}$$

which may not be a feasible solution in practice.

A different approach consists of formulating the pre-equalization as an optimal control problem in the time domain. Assuming that $y_d[k]$ is the desired pulse shape at the output of the device under consideration, the cost function can be written as:

$$J' = \sum_{k=0}^{K} (y_d[k] - y[k])^2 \tag{13}$$

subject to the dynamics of the device:

$$\begin{aligned} x[k+1] &= Ax[k] + Bu[k] \\ y[k] &= Cx[k] + Du[k] \end{aligned} \tag{14}$$

and also the time-domain restrictions of the generator:

$$-u_{\max} \le u(k) \le u_{\max},\ k = 0,1,\ldots,K \qquad \text{(amplitude)} \tag{15a}$$

$$-\Delta u_{\max} \le \Delta u(k) \le \Delta u_{\max},\ k = 0,1,\ldots,K \qquad \text{(slew rate)} \tag{15b}$$

$$\sum_{k=0}^{K} (u[k])^2 \le P_{\max} \qquad \text{(power)} \tag{15c}$$

where $\Delta u[k] = u[k] - u[k-1]$. Equations (13), (14), (15a), (15b) and (15c) define a linear-quadratic control problem with input constraints. This problem can be solved numerically. Moreover, if the power restriction (15c) is found to be of small importance and is eliminated from the optimization problem, then efficient quadratic programming methods can be used to find the solution, because the remaining input restrictions (15a and 15b) are both linear. An alternative consists of transforming the hard input constraints into soft input constraints by adding input-related terms to the cost function (13):

$$J'' = \sum_{k=0}^{K} (y_d[k] - y[k])^2 + \rho(u[k])^2 + \eta(\Delta u[k])^2 \tag{16}$$

where ρ and η are design weights. An exact solution to this problem can be obtained by the use of a Ricatti equation. If it is found to violate the input restrictions, then weight ρ or η can be increased in order to force a reduction in the input amplitude and rate of change respectively. This process can be iterated until a satisfactory solution is found. Examples from the application of the above formulation to indoors DAPPM encoded communications channels will be discussed at the conference.

References

[1] Piesiewicz R, Jemai J, Koch M, and Kurner T, THz channel characterization for future wireless gigabit indoor communication systems, in *Proc. SPIE—Terahertz and Gigahertz Electronics and Photonics IV*, 2005, vol. 5727, pp. 166–176.

[2] Galvão R K H, Hadjiloucas S, Izhac A, Becerra V M and Bowen J W, Wiener system subspace identification for mobile wireless mm-wave networks, *IEEE Transactions Vehicular Technology* vol. 56, No 4, 1935-1948, 2007.

[3] Gfeller F R and Bapst U H, Wireless in-house data communication via diffuse infrared radiation, *Proc. IEEE*, vol. 67, no. 11, pp. 1474–1486, 1979.

[4] Barry J. R., *Wireless Infrared Communications*, KluwerAcademic Publishers, Norwell, Mass, USA, 1994.

[5] Z. Ghassemlooy, A. R. Hayes, N. L. Seed, and E. D. Kaluarachchi, Digital pulse interval modulation for optical communications, *IEEE Commun. Mag.*, vol. 36, no. 12, pp. 95–99, 1998.

[6] Aldibbiat N M, Ghassemlooy Z, and McLaughlin R, Dual header pulse interval modulation for dispersive indoor optical wireless communication systems, *IEE Proceedings Circuits, Devices and Systems*, vol. 149, no. 3, pp. 187–192, 2002.

[7] Shiu D -S and Kahn J M, Differential pulse-position modulation for power-efficient optical communication, *IEEE Trans. Commun.*, vol. 47, no. 8, pp. 1201–1210, 1999.

[8] Sethakaset U, and Gulliver T A, Differential Amplitude Pulse-Position Modulation for Indoor Wireless Optical Communications *EURASIP Journal on Wireless Communications and Networking* vol 1, pp. 3–11, 2005.

[9] Barry J R, Kahn J M, Krause W J, Lee E A, and Messerschmitt D G, Simulation of multipath impulse response for indoor wireless optical channels, *IEEE J. Select. Areas Commun.*, vol. 11, no. 3, pp. 367–379, 1993.

[10] Fernando X.N., and Balendran B., Adaptive Denoising and Equalization of InfraredWireless CDMA System, *EURASIP Journal on Wireless Communications and Networking* vol 1, pp. 20–29, 2005.

[11] Carruthers J B and Kahn, J M, Modeling of nondirected wireless infrared channels, *IEEE Trans. Commun.*, vol. 45, no. 10, pp. 1260–1268, 1997.

[12] Taub H, and Schilling D L, *Principles of Communication Systems*, 2nd ed. McGraw-Hill, New York, 1986.