APPLICATION OF ALGORITHMS FOR DOA ESTIMATION AND BEAMFORMING TO INFRARED PHOTODIODE ARRAYS

K. Iversen   *M. Wolf*   *D. Mämpel*   *H. Schubert*   *C. Schmidt*[1]

Heinrich-Hertz-Institut für Nachrichtentechnik Berlin GmbH, Einsteinufer 37, 10587 Berlin, Germany
[1] Technical University of Ilmenau, Department of Communication and Measurement
P.O.Box 100565, 98684 Ilmenau, Germany

ABSTRACT

In this paper we present a novel infrared communication system for large cells and low user mobility. The proposed transmitter technique allows nearly perfect direction-of-arrival (DoA) estimation in time division multiple access (TDMA) environments only with three photodiodes. Using a photodiode array we show that known algorithms for DoA-estimation can be applied in spatial division multiple access (SDMA) systems.

1. INTRODUCTION

Wireless data transmission based on infrared (IR) radiation is known as a serious alternative to radio especially for high speed indoor applications [2, 11, 12]. For example, advantages are the huge utilizable bandwidth — there are over 200 THz in the 700-1500 nm range alone — which is not regulated. The use of optical standard components such as plastic lenses, photo- and laserdiodes or LEDs allows cost efficient systems by principle. Here it must be noted that the photodetector generates the square of the envelope of the optical bandpass signal and hence inherently provides the bandpass to baseband conversion. Furthermore, due to the resistance against electromagnetic fields, IR is applicable in sensitive environments like intensive-care units.

On the other hand IR has some drawbacks. The most important is that the mobility is limited. As a result of its very small wavelength IR will not be diffracted by man-made barriers and hence cannot pass through (opaque) obstacles. For this reason, diffuse applications seem to be the only way to offer good mobility since they provide redundant paths between transmitter and receiver leading to resistance against shadowing [6]. But unfortunately these multiple paths simultaneously generate multipath dispersion [3, 8]. Furthermore, the range of those diffuse systems is strongly limited. Noise interference due to ambient light is one reason, another is the square-law nature of a direct-detection receiver which doubles (in dB) the path loss compared to a linear receiver. So, if power restrictions due to eye safety are considered, diffuse IR-systems are limited to small rooms.

However, a direct-beam line-of-sight (LOS) IR-link provides excellent channel properties and power budgets even over large distances [5, 11]. In this case an indoor network consists of a base station which acts as an active repeater and provides the connection to the backbone, see Fig. 1. Both transmitters, one of the base station and one of the terminal, are characterized by small beam widths (spots) and must be aligned for this reason. The concept causes the requirement of an unobstructed path between transmitter and receiver limiting applicability to configurations where this condition can be met easily. For example, possible applications are computer IR-interconnections in booths at a fair, large area offices and manufacture halls compatible to HIPERLAN IV. Here employment of wireless transmission reduces installation effort enormously. While pointing of the terminal transceiver is done manually, the base station should be able to detect the DoA of the terminal to redirect its transmitter automatically, see Fig. 2. One possibility for DoA-estimation is the use of a two-dimensional charge coupled detector (CCD) in conjunction with an focusing lens as proposed by Wisely [11]. The problem of this configuration is that the solution is discrete with resolution depending on the size of the array and therefore not practicable for large rooms.

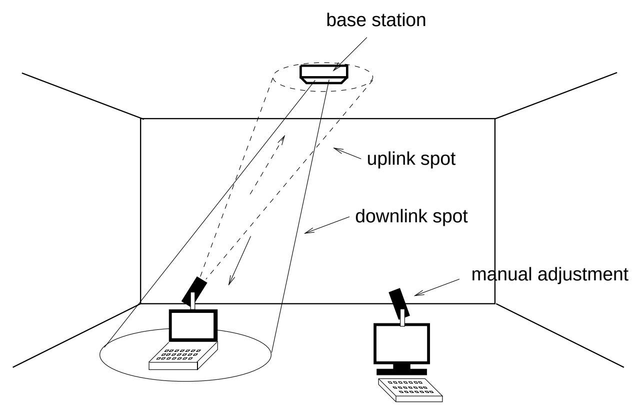

Figure 1. Proposed direct-beam system for large rooms providing placement flexibility.

In this paper we propose a new feasibility of DoA-estimation which is independent of cell size. Our approach is based on electrical subcarriers. As a result we can achieve nearly perfect resolution. In the following we will show the operation in TDMA and SDMA environments.

2. DOA ESTIMATION WITH THREE PHOTODIODES FOR TDMA

2.1. System description

First we propose an efficient method to estimate DoA of terminals based on system shown in Fig. 1. The transceivers of the quasi-stationary mobiles are aligned manually. Beside a data transceiver each terminal has a separate transmitter only for DoA estimation which consists of an optical microwave generator (OMG), see Fig. 3. This OMG is active only during the manually alignment process of the terminal, hence collisions at the base station are prevented. The

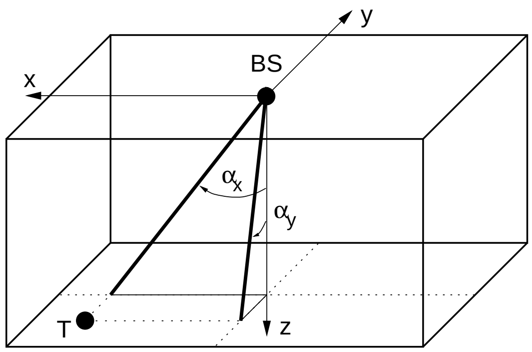

Figure 2. The DoA-estimation of terminals (T) contains the angels in the x-z-plane and the y-z-plane of an orthogonal x-y-z-reference system with base station (BS) in the coordinate-center.

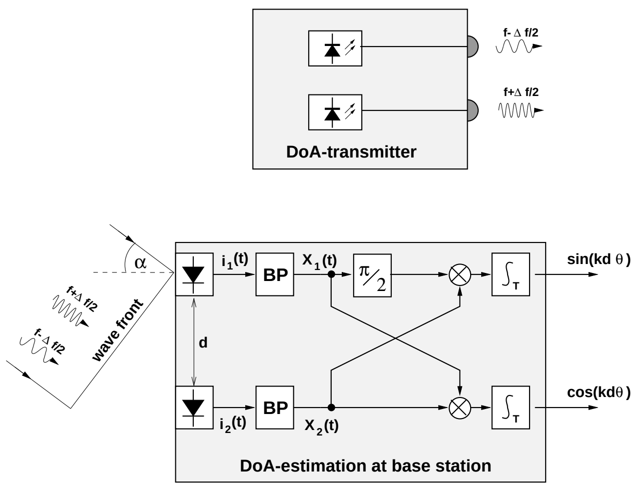

Figure 3. Transmitter (terminal) and receiver (base station) for DoA-estimation ($k = 2\pi/\lambda$, $\theta = \sin(\alpha)$).

basestation consists of a large field-of-view data receiver, a steerable data transmitter and a second receiver for DoA estimation, only. For each angle α_x and α_y to be estimated it needs two photodiodes aligned to the orthogonal reference system, where one diode can be used for both reducing the number of diodes to three. If the basestation detects the DoA information of one terminal a Time Division Multiplex (TDM) downlink and a TDMA uplink, synchronized by the downlink, can be established.

2.2. Principle of operation

If the first photodiode of the receiver is excited by two infrared signals $p_1(t)$ and $p_2(t)$ with equal polarization and unique optical power P_{opt}

$$p_1(t) = \sqrt{P_{opt}} \cdot e^{j(2\pi(f+\Delta f/2)t+\phi_p)}\,, \qquad (1)$$

$$q_1(t) = \sqrt{P_{opt}} \cdot e^{j(2\pi(f-\Delta f/2)t+\phi_q)}, \qquad (2)$$

where Δf is the frequency difference, and ϕ_p and ϕ_q are random phase, the output photocurrent is given by

$$i_1(t) = |p_1(t) + q_1(t)|^2 = 2RP_{opt}(1 + \cos(2\pi\Delta f t + \phi))\,. \quad (3)$$

R denotes the responsivity of the diode and the term $\phi = \phi_p - \phi_q$ represents the phase difference. After bandpass filtering (see Fig. 3) the complex envelope of the intermediate frequency signal of frequency Δf under consideration of receiver noise yields

$$x_1(t) = 2RP_{opt}e^{j\phi} + n_1(t) = s(t) + n_1(t), \qquad (4)$$

where $s(t)$ denotes the signal component (which is unmodulated in this approach) and $n_1(t)$ is additive white Gaussian noise.

If we assume plane waves the time delay of the second photodiode can be expressed as $\tau = d\sin(\alpha_i)/c$ where $\alpha_i,\ i \in [x, y]$ denotes the arriving angle and d the distance between both diodes. If the coherence length of the electrical carrier is large compared to d, i. e. $s(t)$ is constant while traveling over both diodes, the narrow band assumption

$$s(t-\tau)e^{j2\pi\Delta f(t-\tau)} = s(t)e^{-jkd\theta}e^{j2\pi\Delta f t} \qquad (5)$$

holds. With (5) the complex envelope at output of the second bandpass filter can be written as

$$x_2(t) = s(t)e^{-jkd\theta} + n_2(t) = s(t)a(\theta) + n_2(t). \qquad (6)$$

The term $k = 2\pi\Delta f/c$ is the wave number of the subcarrier and $a(\theta) = e^{-jkd\theta}$ is called the steering component containing the direction information $\theta = \sin(\alpha_i),\ i \in [x, y]$ which is of special interest.

2.3. DoA Estimation

To estimate DoA, spatially information must be considered and hence it seems reasonable to crosscorrelate both signals $x_1(t)$ and $x_2(t)$. The real and imaginary part of the complex crosscorrelation term $r_{1,2} = \mathrm{E}[x_1(t)x_2^*(t)] = 2RP_{opt}a^*(\theta)$ are given by

$$\mathrm{Re}\{r_{1,2}\} = 2RP_{opt}\cos(kd\theta) \quad \text{and} \qquad (7)$$

$$\mathrm{Im}\{r_{1,2}\} = 2RP_{opt}\sin(kd\theta), \qquad (8)$$

respectively. Eqn. (7) and (8) contain all information to estimate the direction θ in the range $[-\pi/(k \cdot d), \pi/(k \cdot d)]$ if normalization by RP_{opt} can be achieved. This can be done easily by normalization with respect to the absolute value $|r_{1,2}| = 2RP_{opt}\sqrt{\sin^2(kd\theta) + \cos^2(kd\theta)}$. Another possibility is to autocorrelate a bandpass output with an appropriate delay so that the noise can be neglected.

Our method for DoA-estimation doesn't need search algorithms. Hence, crosscorrelation and further normalization can be seen as a simple strategy to estimate DoA of one terminal at a time. This condition is ensured by the manually alignment. As it is shown in Fig. 3 both parts of $r_{1,2}$ can obtained by analog devices. The imaginary part of $r_{1,2}$ is obtained by an additional phase shift of $\pi/2$. If crosscorrelation is done digitally the sampling frequency can be lower than Δf (e. g. like in sampling oscilloscopes).

a) Required frequency difference

The angels α_x and α_y are limited to be smaller then $\pi/2$ depending on room size. Hence the period of $a(\theta)$ changes with the maximum allowed arriving angle $|\alpha_{max}|$. If the entire period of $a(\theta)$ is used, i. e. $2\pi = 2kd\sin(|\alpha_{max}|)$, the relationship between Δf and d is given by

$$d = \frac{\lambda}{2} \cdot \frac{1}{\sin(|\alpha_{max}|)} = \frac{c}{2\Delta f}\frac{1}{\sin(|\alpha_{max}|)}. \qquad (9)$$

Fig. 4 shows the required frequency difference as a function of d. It is can be seen that Δf can be adopt to available

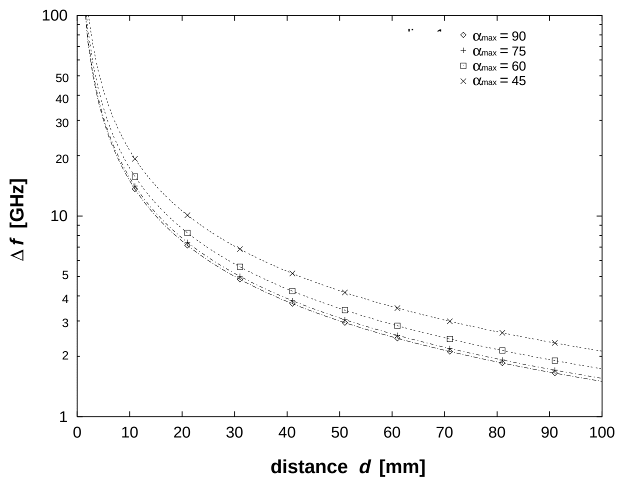

Figure 4. Required frequency difference as a function of d

OMG-modules and high speed photodiodes. Off course, the subcarriers can be also generated at terminals by an electrical oscillator. However, the OMG can be realized as a Photonic Integrated Circuit (PIC) [4] which allows cost advantages.

b) Influence of photodetector area on estimation accuracy

In wireless IR-links the electrical signal-to-noise power ratio (SNR) depends on the photodetector size since shot noise power (which is induced mostly by background light) increases linearly with detector area while signal power increases with the square of the area. In the previous system the detector areas of the DoA-estimator can be small since the subcarriers can be filtered with a very small bandwidth. But if the system needs to detect all signals parallel — in this case the OMG must be modulated with the data streams — the photodetector area must be larger to reach the required SNR [1].

For this reason the influence of the photodetector size on estimation accuracy is discussed now. The photodiode integrates the incoming signal over the entire surface. If we assume a diode with a rectangular detecting area with dimension l in the direction of interest the subcarrier signal $\sqrt{P}\cos(2\pi\Delta f t)$ of an ideal point-like receiver changes to

$$\frac{\sqrt{P}}{l}\int_0^l \cos(2\pi\Delta f t - \gamma k\theta)\mathrm{d}\gamma = 2\frac{\sqrt{P}}{lk\theta}\sin(\frac{lk\theta}{2})\cos(2\pi\Delta f t - \frac{lk\theta}{2}) \quad (10)$$

As expected the increase of detector dimension causes an attenuation of the signal depending on θ as well as an phase shift. The phase shift is of no interest if all diodes have the same dimensions. The attenuation reaches its maximum at $\alpha = \alpha_{max}$ and is smaller then $4dB$ if $l = d$.

3. PHOTODIODE ARRAYS FOR SDMA

If the OMG of the j-th of M terminals is modulated with the data information the signal component behind the bandpasses in Fig. 3 can be simply expanded to

$$s_j(t) = m_j(t)2RP_{opt}e^{j\phi_j}. \quad (11)$$

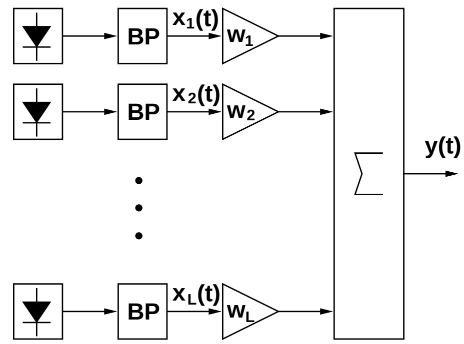

Figure 5. Beamformer with L elements

where $m_j(t),\ j \in [1\cdots M]$ denotes the modulation function in the complex baseband representation. If the outputs $x_1(t)$ and $x_2(t)$ of the bandpasses are combined (added) $s_j(t)$ can be still detected but received power depends on the direction θ. If the arrangement of two photodiodes for each angle component α_x and α_y is replaced by an one dimensional array of L elements whose outputs are weighted, see Fig. 5, it is possible to minimize power in multiple directions, i. e. directions of undesired users. This process supposes that the directions of this users are known.

3.1. Definitions

The weighting coefficients are expressed as the vector

$$\mathbf{w} = [w_1\ w_2\ \cdots\ w_L]^T. \quad (12)$$

For convenience all other signals proposed in the last section are defined as vectors now. The number of active users is denoted by M, where the condition $M < L$ must be met.

symbol	structure	dimension
$\mathbf{x}(t)$	$[x_1(t), x_2(t), \cdots, x_L(t)]^T$	$L \times 1$
$\mathbf{a}(\theta_j)$	$[a_1(\theta_j), a_2(\theta_j), \cdots, a_L(\theta_j)]^T$	$L \times 1$
$\mathbf{A}(\Theta)$	$[\mathbf{a}(\theta_1), \mathbf{a}(\theta_2), \cdots, \mathbf{a}(\theta_M)]^T$	$L \times M$
$\mathbf{s}(t)$	$[s_1(t), s_2(t), \cdots, x_M(t)]^T$	$M \times 1$
$\mathbf{n}(t)$	$[n_1(t), n_2(t), \cdots, n_M(t)]^T$	$M \times 1$

The vector $\mathbf{x}(t)$ is referred as the data vector, $\mathbf{a}(\theta_j)$ as the steering vector, $\mathbf{A}(\Theta)$ as the steering matrix, $\mathbf{s}(t)$ as the signal vector and $\mathbf{n}(t)$ as the noise vector. In this vector-based notation the output signal $y(t)$ can be simply be written as

$$y(t) = \mathbf{w}^T\mathbf{x}(t) \text{ with } \mathbf{x}(t) = \mathbf{A}(\Theta)\mathbf{s}(t) + \mathbf{n}(t). \quad (13)$$

To exploit the spatial information between the elements now the covariance matrix can be used

$$\mathbf{R_{xx}} = E[\mathbf{x}(t)\mathbf{x}^H(t)]\ . \quad (14)$$

3.2. Calculation of the weight coefficients

One method to suppress all $(M-1)$ undesired users is to steer zeros in these directions while keeping the desired signal j. This is equal to find a vector $\mathbf{w}$ which is orthogonal to all steering vectors $\mathbf{a}(\theta_i)$, $i = 1\cdots M, i \neq j$ but not to $\mathbf{a}(\theta_j)$

$$\mathbf{w}^T\mathbf{x}(t) = s_j(t)\mathbf{w}^T\mathbf{a}(\theta_j)\ . \quad (15)$$

The vector $\mathbf{w}$ can be found by projection. If all directions are known the projection matrix $\mathbf{P}$ can be calculated by [7]

$$\mathbf{P} = \mathbf{I} - \widetilde{\mathbf{P}} \quad \text{where} \quad \widetilde{\mathbf{P}} = \mathbf{Q}^*(\mathbf{Q}^t\mathbf{Q}^*)^{-1}\mathbf{Q}^t \quad (16)$$

The matrix $\mathbf{Q}$ is the steering matrix reduced by the desired vector $\mathbf{a}(\theta_j)$. The weighting vector can be obtained by

$$\mathbf{w} = \mathbf{P}\mathbf{a}^*(\theta_j) \ . \tag{17}$$

Eqn. (15) is fulfilled if $\mathbf{a}(\theta_j)$ is linear independent on all other steering vectors which is satisfied if $\mathbf{a}(\theta_j)$ lies in another direction.

Another approach is to maximize the signal-to-noise plus interference power ratio. In case the optimal weighting vector is $\mathbf{w}_{opt} = \mathbf{C}^{-1}\mathbf{a}^*(\theta_j)$, where $\mathbf{C}$ is noise-plus-interference covariance matrix. In the case of minimizing the mean square error the optimal weight vector is $\mathbf{w}_{opt} = \mathbf{R_{xx}}^{-1}\mathbf{a}^*(\theta_j)$.

3.3. DoA estimation of multiple signals

The possibility of beamforming by arrays of size $L > M$ which is shortly described in the last section requires the parallel DoA-estimation of multiple signals. For this reason DoA estimation proposed in the first section cannot be employed. But there are many algorithms known from literature [9] dealing with this problem. These algorithms can be applied since the circumstances at the bandpasses outputs in Fig. 5 are the same as in scenarios for which they have been originally designed. For optical systems the optical signals from multiple OMGs must not coherently interfere. For this reason the coherence times of the optical signals must be very short against the integration time of the photodiode while the electrical subcarriers must satisfy the narrow band condition. Our approach has the advantage that only reflections of orders higher than one appear which have neglectable influence to the desired signals due to the high path loss. Hence there are no problems with distortion due to coherent signals. For that reason searching methods like MUSIC [10], which is an subspace based method similar to the described null steering, becomes possible. Because the terminals are quasi stationary the calculation time of signal-processors is not critically. Hence complex parametric methods like deterministic maximum likelihood become possible.

ACKNOWLEDGMENTS

This work was supported by the German Federal Ministry of Education, Science, Research, and Technology and the country of Berlin.

REFERENCES

[1] M. D. Audeh and J. M. Kahn. Performance Evaluation of Baseband OOK for Wireless Indoor Infrared LAN's Operating at 100 Mb/s. *IEEE Transactions on Communications*, 43(6):2085–2094, 1995.

[2] J. R. Barry. *Wireless Infrared Communications.* Kluwer Academic Publishers, Boston Dordrecht London, 1994.

[3] J. R. Barry, J. M. Kahn, W. J. Krause, E. A. Lee, and D. G. Messerschmitt. Simulation of multipath impulse response for indoor wireless optical channels. *IEEE Journal on Selected Areas in Communications*, 11(3):367–379, 1993.

[4] R.P. Braun and G. Grosskopf et. al. Optical microwave generation and transmission experiments in the 12 and 60 GHz-region for wireless communication. *IEEE MTT-S, International Microwave Symposium, San Francisco*, pages paper WE 1D–5, June 1996.

[5] T. S. Chu and M. J. Gans. High Speed Infrared Local Wireless Communication. *IEEE Communications Magazine*, 25(8):4–10, August 1987.

[6] Fritz R. Gfeller and Urs Bapst. Wireless In-House Data Communication via Diffuse Infrared Radiation. In *Proceedings of the IEEE*, volume 67, No. 11, pages 1474–1486, 1979.

[7] J.E. Hudson. *Wireless Infrared Communications.* Peter Peregrinus Ltd.,London,United Kingdom, 1991.

[8] J. M. Kahn, W. J. Krause, and J. B. Carruthers. Experimental Characterization of Non-Directed Indoor Infrared Channels. *IEEE Transactions on Communications*, 43(2/3/4):1613–1623, February/March/April 1995.

[9] H. Krim and M. Viberg. Two Decades of Array Signal Processing Research. *IEEE Signal Processing Magazine*, pages 67–94, July 1996.

[10] R.O. Schmidt. *A Signal Subspace Approach to Multiple Emitter Location and Spectral Estimation.* Ph.D. thesis,Stanford University,Stanford, CA, 1981.

[11] D. R. Wisely. A 1 Gbit/s Optical Wireless Tracked Architecture for ATM Delivery. In *IEE Coll. on Optical Free Space Comm. Links,London,UK*, pages 14/1–7, Febr. 1996.

[12] G. Yun and M. Kahvehrad. Spot-diffusing and fly-eye receivers for indoor infrared wireless communications. In *IEEE International Conference on Selected Topics in Wireless Communications; Vancouver, Canada*, pages 262–265, 1992.