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Performance analysis of a low cost wireless indoor positioning system with distributed antennas

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Abstract. In this paper an approach for indoor localization based on already installed communication transceiver infrastructure is presented. Only marginal additional hardware is required to implement the localization functionality. The idea is to use a power level detection evaluation of the received signals in communication receivers to localize participants. In the planned configuration the power transmitted from mobile terminals is received by a leaky wave cable. Its length and coupling properties are dimensioned according to the dynamic range of two access points connected to the two cable heads. In order to adapt the problem to circuit-based analysis techniques the leaky wave cable is represented by a distributed antenna system. Thereby the considered design consists of a cascade of coupling structures with broadband Vivaldi antennas connected to the coupling ports. An experimental system has been built and also tested using an automated frequency domain measurement setup. Measurement results relating the system bandwidth to positioning accuracy show good agreement with theoretical investigations.

1 Introduction

In recent years indoor positioning systems (IPS) have gained importance in research as commercially available satellite based positioning systems become inaccurate or even drop out completely inside buildings (Gu et al., 2009). Ultra-wideband (UWB) or active radio frequency identification (RFID) systems are a state-of-the-art replacement (Bill et al., 2004). They can provide accuracy down to the centimeter range.

However, these systems are very costly due to complex digital signal processing. A very cost efficient approach is to deploy the already installed network infrastructure. Thus it can be trusted that only little additional hardware is needed for the purpose of localization. One possible approach is to combine the observed power levels at the access points of the system with a-priori knowledge about the properties of the antenna system (Saha et al., 2003) into an estimator for the positions of the client terminals.

A serious disadvantage of this positioning approach is the fact that amplitude detection is rather sensitive with respect to the inherent amplitude fading effects. Hence a larger number of access points (observations) is necessary to reach the required accuracy. A refinement of the positioning resolution can be reached by placing distributed antenna systems between the access points (Saleh et al., 1987).

A novel approach is to achieve localization via power level detection by using leaky wave cables (Engelbrecht et al., 2010). Furthermore communication systems for tunnels and indoor environments based on leaky wave cables are well investigated (Delogne and Deryck, 1980; Motley and Palmer, 1983). However, the dimensioning of a leaky wave cable is very challenging as it is difficult to allocate a radiation pattern to a radiating slot (Walter, 1965). Furthermore any modification of the cable performance involves a redesign of the entire cable structure. To derive design rules a distributed antenna system is perfectly suited as it is possible to change single components with well known properties individually.

In this paper the evaluation of the localization performance of a distributed antenna system related to the properties of a leaky wave cable is reported. Section 2 comprises the basic principles of the localization system followed by a description of the developed antenna architecture in Sect. 3. The key features of the implemented simulation tool is described in Sect. 4. In Sect. 5 the performance of the wireless indoor positioning system is analyzed by simulations and measurements before the paper is concluded in Sect. 6.

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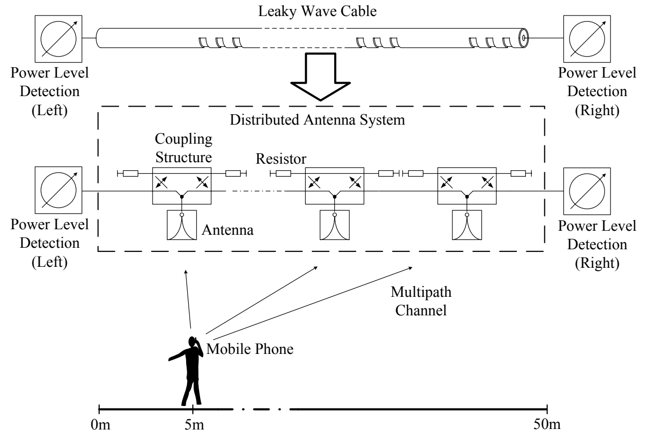

Fig. 1. Schematical diagram of an indoor wireless coverage system and the positioning approach under study. The leaky wave cable is represented by an equivalent distributed antenna system with appropriate coupling factors.

2 Functional principle

The functional principle of the wireless indoor positioning system is depicted in Fig. 1. A leaky wave cable is used to realize the radio link of a participant with a mobile. The position of the mobile should be estimated by means of a comparison of the output power levels at the cable heads. The output power level difference

$$\frac{\Delta P}{\mathrm{dB}} = \frac{P_{\mathrm{left}}}{\mathrm{dBm}} - \frac{P_{\mathrm{right}}}{\mathrm{dBm}} \tag{1}$$

is detected by subtracting the logarithmic power levels P_{left} and P_{right} at the cable heads from each other. By comparing the result with an a-priori known reference characteristic the power levels are mapped to an estimated position in the room (Weber et al., 2009). This so-called RF fingerprinting technique has the advantage that the estimate is based on the level difference ΔP rather than absolute levels. Thus the system is very robust against motions orthogonal to the cable as well as polarization loss. The system requirements have been chosen to achieve a linear curve progression with respect to position and level difference. Thus the location error of simulation and measurement is referred to the deviation from a least mean square error (LMSE) algorithm (Hagan et al., 1996).

The slots of the leaky wave cable have been replaced by a cascade of coupling structures and antenna elements. Consequently the design complexity reduces significantly as coupling between the radiating elements can be neglected. Furthermore the leaky wave cable has to be considered as an array of single radiating elements. The radiation behavior of one slot is not that easy to determine and to control whereas the radiation pattern of a single antenna element is well known.

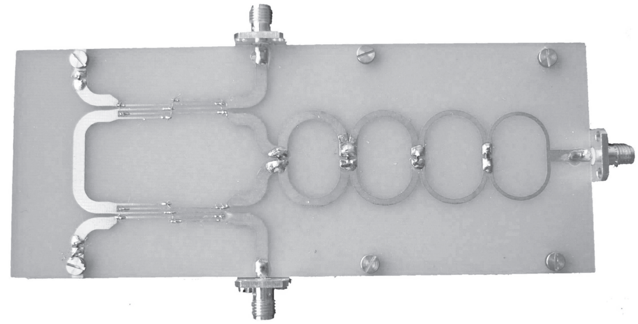

Fig. 2. Photo of the developed coupling structure prototype.

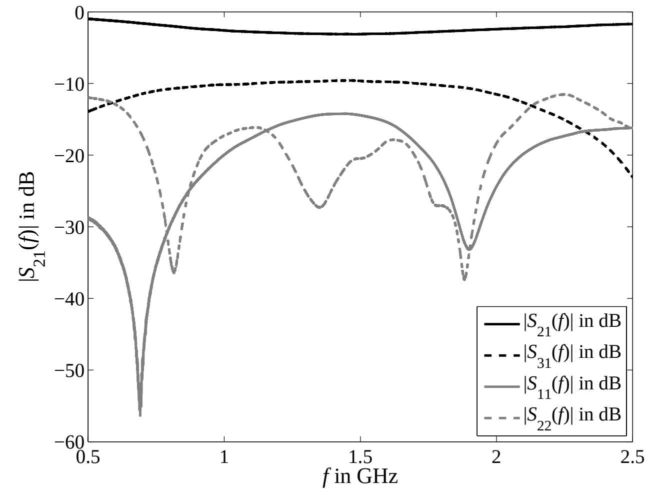

Fig. 3. Scattering parameters of the developed coupling structure (measurement).

3 Antenna prototype

The experimental antenna system has been built-up as a chain of identical individual parts including a coupling structure and an antenna (see Fig. 1). This elementary cell is supposed to represent a radiating slot (or a group of slots) on the shielding of the leaky wave cable. Thus the coupling structure of the elementary cell has been dimensioned in such way that only a small portion of the guided energy from the backbone cable is radiated by each antenna element. A prototype of a single coupling structure is depicted in Fig. 2. The structure consists of two mirrored Lange couplers (Lange, 1969) and a four stage broadband Wilkinson divider (Wilkinson, 1960). Similar to a single radiating slot it implements a weak coupling with no directivity. A drawback of this concept is that almost the complete received energy of the antennas is dissipated in the load elements. The scattering parameters of the coupling structures are illustrated in Fig. 3. The signal from the antenna is attenuated by 9 dB and the signal which passes the coupling structure is attenuated by 3 dB.

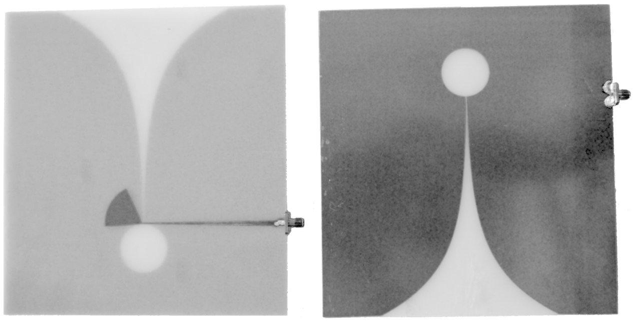

Fig. 4. Photograph of the developed Vivaldi antenna prototypes.

Based on the simulation results we found that Vivaldi antennas (Gibson, 1979) are perfectly suited as receiving antennas. Two antenna prototypes with rear and front view are depicted in Fig. 4. By feeding the antennas with a balun a symmetrical radiation pattern can be achieved. Furthermore the antenna is low directive within the frequency range from 800 MHz to 2 GHz and thus field distributions from different angle of arrivals are equally weighted.

4 System simulation tool

A detailed comparison between various scenarios within different environments provides a deep insight into fundamental limitations of positioning accuracy and reliability. Especially inside large buildings like concert halls or train stations prototypes are costly as many elementary cells are needed and the timbering can be difficult. Hence a simulation tool which enables to plan the entire system has been developed. An overview of the system simulation approach is given in Fig. 5. According to the individual system requirements the number of antenna elements M and the spacing between the antenna elements can be variably chosen. The entire system of cascaded couplers is modeled by a scattering matrix which enables to compute the power waves at the output ports $M+1$ and $M+2$ from the input waves at the M antenna ports. They are computed from the incoming rays of a ray tracing process weighted with the antenna properties.

To estimate the maximum fidelity which can be achieved by the antenna system we identified the impact of reflections in the environment and on the antenna system itself as essential effects for signal distortions. Thus a ray tracer based on a shooting and bouncing ray (SBR) algorithm similar to (Buddendick and Eibert, 2010) has been applied. As ray finding in complex scenarios can be very time intensive SBR algorithms are a good choice to characterize complex scenarios (Ling et al., 1989). Thereby the electrical field at the receiver is computed from the arriving rays on a sphere around the nominal receiver position. Furthermore Computer Aided Design (CAD) data can be applied and simulated.

To obtain accurate and reliable simulation results the receiving antenna elements were characterized by angle and frequency dependent transfer functions similar to Kunisch (2007), Farr and Baum (1998). Consequently this approach also enables to test the system with different antenna properties. The antenna transfer functions have been chosen in a way that the electrical field in free space is related with the power waves at the output ports of the antennas. Given the wave $a(\omega)$ incident at the antenna port and the normalized vector valued antenna transfer function $\boldsymbol{H}_{\mathrm{TX,N}}(\omega,\varphi,\vartheta)$ the electrical field launched by the TX antenna (mobile phone) can be expressed by

$$\boldsymbol{E}_{\mathrm{TX}}(\omega,\varphi,\vartheta)=\frac{\mathrm{j}\omega Z_{\mathrm{F0}}}{2\pi c_0}\boldsymbol{H}_{\mathrm{TX,N}}(\omega,\varphi,\vartheta) \cdot\frac{\sqrt{Z_{\mathrm{L}}}}{Z_{\mathrm{A}}(\omega)+Z_{\mathrm{L}}}a(\omega) \quad (2)$$

with c_0 the speed of light, $Z_{\mathrm{A}}(\omega)$ the antenna terminal impedance, Z_{F0} the intrinsic impedance of free space and Z_{L} the reference impedance of the transmission lines, respectively. Thereby the normalized antenna transfer function is related to the antenna transfer function in the receiving case. Assuming that the transmit antenna and receive antenna m are separated by a distance of R_m in free space, the field incident at antenna m is given by

$$\boldsymbol{E}_{\mathrm{RX},m}(\omega,\varphi,\vartheta)=\boldsymbol{E}_{\mathrm{TX}}(\omega,\varphi,\vartheta)\frac{\mathrm{e}^{-\mathrm{j}\omega\tau_m}}{R_m} \quad (3)$$

with the path delay $\tau_m=R_m/c_0$. In the receiving case K different rays $\boldsymbol{E}_{\mathrm{RX},m}(\omega,\varphi_k,\vartheta_k)$ arrive from arbitrary directions (φ_k,ϑ_k) at the surface of the antenna. Thus the outgoing wave at the antenna port m is related to the sum of all incoming electrical field contributions and the antenna transfer function in the receiving case by

$$b_{1,m}(\omega)=\frac{\sqrt{Z_{\mathrm{L}}}}{Z_{\mathrm{A},m}(\omega)+Z_{\mathrm{L}}} \cdot\sum_{k=1}^{K}\boldsymbol{H}_{\mathrm{RX,N},m}(\omega,\varphi_k,\vartheta_k)\boldsymbol{E}_{\mathrm{RX},m}(\omega,\varphi_k,\vartheta_k). \quad (4)$$

According to the reciprocity theorem of antennas the afore stated ansatz leads to the equivalence of the normalized transmitting and receiving antenna transfer functions

$$\boldsymbol{H}_{\mathrm{TX,N}}(\omega,\varphi,\vartheta)=\boldsymbol{H}_{\mathrm{RX,N}}(\omega,\varphi,\vartheta). \quad (5)$$

Furthermore both transmitting and receiving transfer functions can be related to the classical properties of antennas. Thus the magnitude of the transfer function in the transmitting case is related to the antenna gain $G_{\mathrm{TX}}(\omega,\varphi,\vartheta)$ according to

$$|\boldsymbol{H}_{\mathrm{TX,N}}(\omega,\varphi,\vartheta)|^2=\frac{4\pi c_0^2 G_{\mathrm{TX}}(\omega,\varphi,\vartheta)\mathrm{Re}\{Z_{\mathrm{A}}(\omega)\}}{\omega^2 Z_{\mathrm{F0}}}. \quad (6)$$

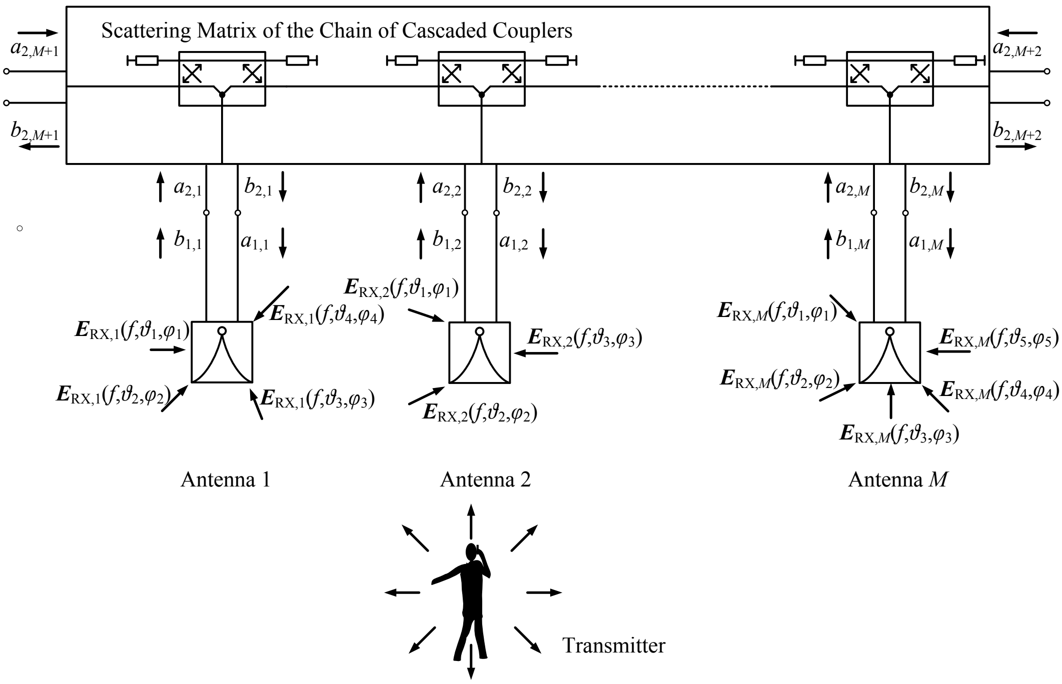

Fig. 5. Architecture of the system simulation.

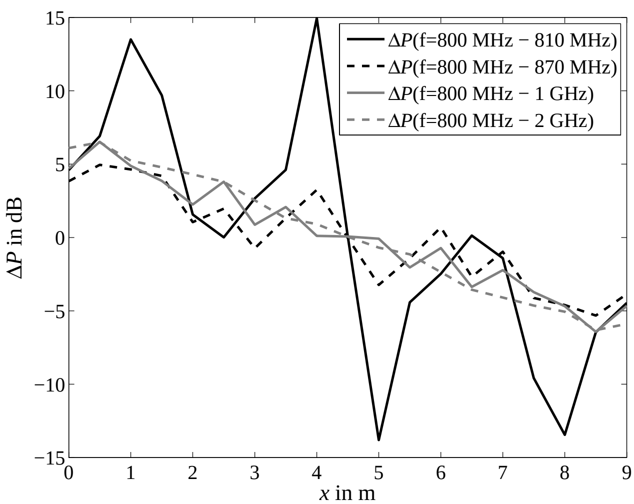

Fig. 6. Simulated level difference at the output ports of a 9 m long distributed antenna system with four equally spaced antenna elements.

Similarly, the receiving transfer function

$$|\boldsymbol{H}_{\mathrm{RX,N}}(\omega,\varphi,\vartheta)|^2 = \frac{4A_{\mathrm{e}}(\omega,\varphi,\vartheta)\,\mathrm{Re}\{Z_{\mathrm{A}}(\omega)\}}{Z_{\mathrm{F0}}} \tag{7}$$

is related to the effective area $A_{\mathrm{e}}(\omega,\varphi,\vartheta)$. By considering a transmission between two antennas in free space it is possible to extract the antenna properties from the transmission scattering parameter $S_{21}(\omega)$. Thus the antenna transfer function can either be deduced from simulations or from measurements in an anechoic chamber.

Furthermore the waves $b_{1,m}(\omega)$ are used as input waves into the coupling structures. To obtain accurate results all coupling structures and transmission lines are cascaded and mathematically described within one scattering matrix. To obtain the system matrix an efficient algorithm according to Simpson (1981) is applied. To accelerate the computation time the entire scattering matrix of the system is computed in a preprocessing step before the simulation is started. The power level difference of the system is determined according to

$$\frac{\Delta P}{\mathrm{dB}} = 10\cdot\log_{10}\left[\int_{f_{\mathrm{c}}-B/2}^{f_{\mathrm{c}}+B/2}\frac{\left|b_{2,M+1}(2\pi f)\right|^2}{2B}\mathrm{d}f\right] - 10\cdot\log_{10}\left[\int_{f_{\mathrm{c}}-B/2}^{f_{\mathrm{c}}+B/2}\frac{\left|b_{2,M+2}(2\pi f)\right|^2}{2B}\mathrm{d}f\right]. \tag{8}$$

Thereby the influence of the signal baseband bandwidth B is considered by computing the mean value of the output waves at the center frequency f_{c}.

5 Performance analysis

A prototype with four antenna elements was tested in a 6 m high building with a length of 12 m and a width of 10 m. The spacings of the single antenna elements was chosen to 3 m. In this paper exemplary results of stationary measurements at

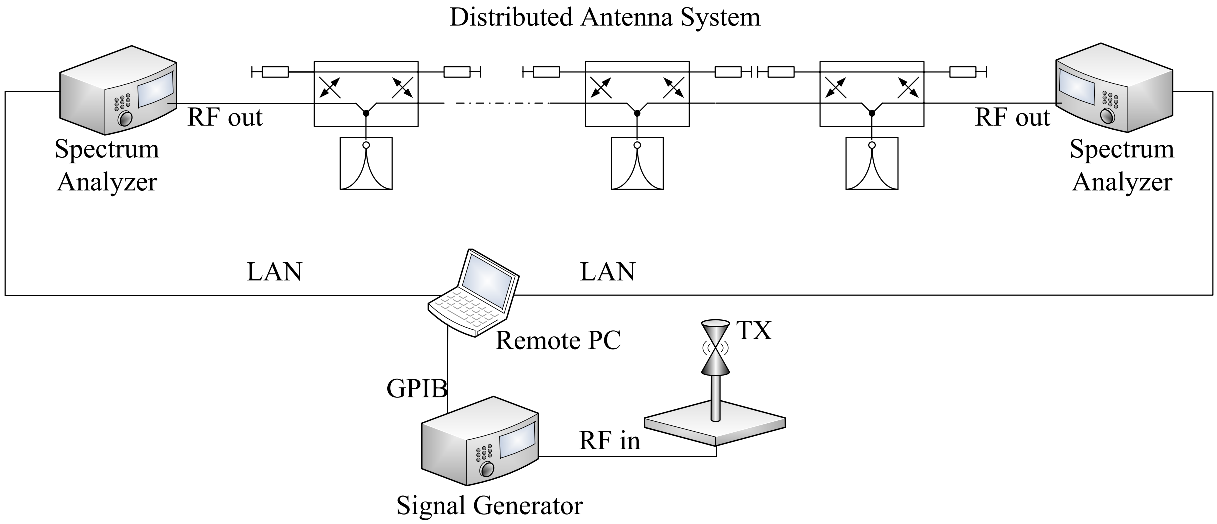

Fig. 7. Illustration of the automated frequency domain measurement setup.

different transmitter positions are presented. Starting under the first antenna element of the system measurements ranging from 0 m to 9 m with an incremental step of 0.5 m were carried out. To identify the potential for commercial use the impact of the system bandwidth was analyzed. Therefore continuous wave (CW) sweeps over the entire frequency range from 800 MHz to 2 GHz at equidistant frequency steps were performed. Averaging the received level differences at different frequency ranges yields an estimation of the localization accuracy. To achieve comparable results equivalent conditions were considered in the simulations.

5.1 Simulations

In the simulations the ray tracing process was started for discrete frequencies with a step width of 1 MHz over the entire frequency range at each transmitter position. In Fig. 6 simulated level differences at varying system bandwidths are presented. It can be observed that with increasing bandwidth the curves become smoother with less variations and therefore less ambiguities in the estimation of position x from power difference ΔP. An almost linear behavior is reached when the signal baseband bandwidth exceeds 70 MHz.

5.2 Measurements

Measurements were carried out using an automated measurement setup including two spectrum analyzers and a signal generator. An important advantage of frequency domain analysis is that the measured data can be used flexibly for evaluation of different system bandwidths. A schematical drawing of all components of the measurement setup is depicted in Fig. 7. The transmitting part consists of a signal generator connected to an omnidirectional low dispersive biconical antenna (Ott and Eibert, 2010). This setup enables to test the system under ideal conditions and makes the measurement results comparable to the simulations with an isotropic transmitting antenna. Instead of root mean square (RMS) level detection two spectrum analyzers were connected to the output ports of the distributed antenna system. A remote personal computer is used to provide communication between the devices over Local Area Network (LAN) and General Purpose Interface Bus (GPIB) connections. This setup enables to characterize the entire frequency range of the signal by applying a CW sweep. Thus predictions within the entire system bandwidth are possible without filtering the signal. Furthermore a larger dynamic range can be investigated compared to measurements with network analyzers.

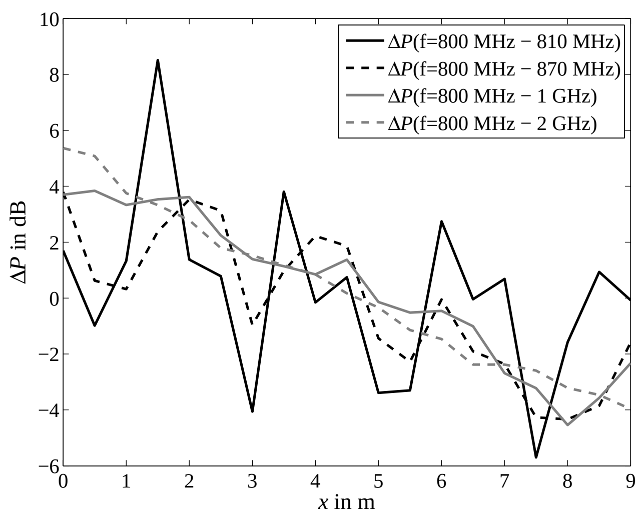

Fig. 8. Measured level difference at the output ports of a 9 m long distributed antenna system with four equally spaced antenna elements.

Figure 8 shows the measurement results carried out in a hall at the authors institute. As expected from the simulations location accuracy is increased significantly with arising bandwidth. Within this measurement campaign the required localization fidelity of 3 m is achieved at a system bandwidth of 70 MHz. Comparing measurement and simulation results stronger frequency selective fading effects occurred in the simulations at a signal bandwidth of 10 MHz. A reasonable explanation is that within the simulations much stronger multipath effects occurs due to walls with metallic surfaces.

5.3 Appraisal of results

Measurement and simulation results showed that localization with a distributed antenna system based on simple level detection is possible provided that the signal bandwidth is sufficiently large. On the other hand RMS level detection averages the available signal power over a certain period of time. Thus the system also offers the possibility localize narrowband signal sources with Frequency Hopping Spread Spectrum (FHSS). Further improvement can be expected by applying statistical tracking filters for parameter estimation (Kalman, 1960). It is intended to evaluate the performance of the system also with time domain signals in the future.

6 Conclusions

A localization concept for a wireless indoor positioning system has been presented. The algorithm is based on information from simple power measurements of received signals. Thus positioning with communication transceivers is possible without the need for too much additional hardware. This makes the implementation of the system very cost efficient. Both numerical system simulations and practical measurements in an exemplary indoor environment show that location accuracy of 3 m can be reached for signals with a bandwidth of at least 70 MHz.

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