Radar Systems

Auditorium J6

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Use of Homodyne Methods of Microwave Phase Measurements in a Task of Precision Indoor Positioning

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1. Summary

Microwave indoor propagation offers a good opportunity for object positioning. The use of the pulse radar method for measuring distances and angles are quite unsuitable for indoor applications. The resolution of this method is too low and there is a minimal distance requirement of the pulse radar measurement that is usually higher than the room size. In this paper the phase method of multipoint distance measurements is presented. The microwave phase progression was used for these measurements. The resolution of the phase method of distance measurements is determined by the microwave length. Depending on the wavelength one can reach an accuracy of 10 mm and better [1] and [2]. No doubts, the phase method causes an ambiguity because the phase measurements can only have values in an interval between 0 - 2π. In this paper the way of bypassing this problem is shown.

2. Approach to a problem

Realizing the homodyne method of microwave phase measurements and consequently distances determination, we offer to place the radio beacons in the corners of a room or in the middle of an extended wall. The transponder is placed on the object that is to be located. The positioning of object is characterized by the distances \( d_i \) from the object to each of beacons.

The beacon radiates the microwave signal that can be described as

\[ U_1(t) = U_0 \sin[\omega_0 t + \varphi_0] \]

The microwave, propagated along the distance \( d_i \), obtains the attenuation \( A_i \) and phase progression \( kd_i \),

\[ U_2(t) = A_i U_0 \sin[\omega_0 t + kd_i + \varphi_0] , \]

where \( k = \frac{2\pi}{\lambda} \) is propagation constant, \( \lambda \) is the wavelength.

The transponder receives this signal, shifts the frequency and phase of the received microwave signal on values \( \Omega \) and \( \Phi_{LF} \) by means of a controlled phase shifter

\[ U_3(t) = A_i U_0 \sin[(\omega_0 + \Omega)t + kd_i + \varphi_0 + \Phi_{LF}] , \]

and reradiates this frequency-phase transformed microwave signal back in direction of the beacon. The secondary microwave signal that is received in the beacon will be

\[ U_4(t) = A_i^2 U_0 \sin[(\omega_0 + \Omega)t + kd_i + k^1 d_i + \varphi_0 + \Phi_{LF}] , \]

where \( k^1 \) takes into account the frequency shift \( \omega_0 + \Omega \). If the frequency shift \( \Omega \) is much lower than the initial frequency \( \omega_0 \), then \( k^1 \approx k \). This secondary received signal is mixed with the original microwave signal and at the mixer’s output the low-frequency signal of difference is selected and amplified up to a certain limit. This low-frequency signal will be

\[ U_5(t) = U_0 \sin[\Omega t + 2kd_i + \Phi_{LF}] . \]
The initial frequency and initial phase of origin microwave signal both are subtracted in a mixer. The only double phase progression of the microwave signal is of interest for the distance definition.

A low-frequency signal is obtained on the output of each mixer of each beacon, but the phase shift will be unique for each beacon and will be determined by the each distance \( d_i \). These signals are delivered to the signal processing unit and the phases of these signals are compared with the phase of low-frequency signal reference with the same frequency \( \Omega \).

Generally it is possible to measure a phase difference between 0 and \( 2\pi \). The phase progression \( kd \) will be represented as \( 2\pi n + k\Delta d \), where \( n \) is integer. In order to avoid this problem we serially change the operating frequency of each beacon and we measure the phase differences. After that we calculate the distance as

\[
\Delta d_i = \frac{(\Delta \phi_1 - \Delta \phi_2)}{2(\Omega_1 - \Omega_2)} c.
\]

Certainly, these calculations yield the rough results of distance determination. These calculations let us obtain the number of phase cycles \( n \) and the possibility to determine the distance in terms of integer numbers of wavelengths. The exact value of distance \( d \) can be obtained by measuring the phase difference \( k\Delta d \). Taking into consideration the accuracy of phase measurements in 1.4° (8 digits) and possible wavelength in 0.2 m, the resolution in distance determination will be about 1 mm. We must understand the measured distance will be conditional distance, taking into account antennas' phase centers and all feeders' lengths. Further, as each beacon operates as stand alone unit, there is a possibility to measure the phase difference between beacons' mixers' output signals. Mentioned opportunity let us improve the accuracy of coordinate's determination, as it was pointed out in [3] and [4]. In this paper the design of the equipment and the algorithm are discussed. The problem of phase synchronization of low-frequency oscillators is discussed as well.

### 3. Conclusion

Having defined distances to the object from several beacons (not less than two) and knowing the exact coordinates of each beacon, by means of the simple software we will calculate exact coordinates of the object in a room. The time of the object's coordinates determination will be derived from the time of the phase difference measurement at consecutive iterations. One iteration can last tens or hundreds of milliseconds. The time of the PC computation cannot be taken into account. We can reduce the measurements and the calculations accordingly to a minimum by using a tracking mode.

### 4. References


Power Level Surveillance for an FMCW-based Local Positioning System

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1 Introduction

In the frequency modulated continuous wave (FMCW) based local positioning system LPM the time of flight distance is determined by evaluating the frequency difference between the down converted chirp signal of a reference transponder (RT) and the measurement transponder (MT) inside every base station (BS). In a typical LPM setup several BSs surround the measurement field whereas the RT is located around the center of the setup. Both, the BSs and RT are located at known positions, while the MT is movable. A schematic of an exemplary LPM setup is depicted in Figure 1. For a more detailed explanation of the LPM working principle the reader is referred to [1]. The LPM system also provides the power level (PL) value of the received signals which can be used for monitoring the quality of the current measurement. In this contribution, a signal model for the received signal powers is derived with respect to the antenna characteristics, the hardware parameters of the LPM system and the Friis equation for the line of sight (LOS) path. This information can be used as a weighting factor for the pseudo-ranges in the position estimation process which is shown on real measured data.

\[ P_{m,n} = \eta_{SP,HW} \gamma_{CL} \eta_{BA[\theta]}(\theta[m,n], \phi[m,n]) \xi[m,n] \kappa_{TA[\theta]}(\theta[m,n], \phi[m,n]) \xi_{CL} P_T, \]

Figure 1: Schematic and photograph (point of view around BS 2) of an LPM setup with 12 BSs (circles) for a 3D measurement scenario at the LPM test site in Regau, Austria. The RT (square) is at fixed position located approximately in the middle of the setup, whereas the MT (diamond) is mobile. The lines mark the LOS path from the RT (dashed), and MT (dotted) respectively, towards the BSs.

2 Method and signal model for PL estimation

The basic model for the received PL can be defined as
where $P_r[m, n]$ denotes the measured PL from the $m$th transponder (TP) in the $n$th BS, where $P_t$ is the transmitting power of each TP. Since position estimation in the LPM system is strictly restricted to LOS paths, the path loss is modeled with the well-known Friis equation [2] in $\zeta[m, n]$. Moreover, hardware parameters such as the antenna gains $\eta_{BA[n]}$ and $\kappa_{TA[m]}$, the cable attenuations $\xi_{CL}$ and $\gamma_{CL}$ as well as the signal amplification of the BSs $\eta_{SP,HW}$ have a major effect on the PL. The PL of the MT ($P_r[1, n]$) is highly influenced by the BS antenna gain towards its position, which is dependent of the azimuth and elevation angles of the incident signal. Since the antenna characteristics provided by the manufacturer are often limited to a measurement of one azimuth/elevation plane, this paper will present a method to interpolate this information over the whole unit sphere.

3 Simulation and measurement results

As depicted in the left plot of Figure 2, an arbitrary movement with a remote-controlled vehicle (RCV) equipped with an MT was performed. For the estimated positions, the PLs of an arbitrarily chosen BS (here BS 12) were simulated. The right plot of Figure 2 shows that the simulated data fits the measured values, except in the regions where the RCV leaves the field of view of BS 12. Note that the displayed PLs are scaled by an internal factor $P_0$. For a schematic of the setup of the BS and RT, see Figure 1.

![Figure 2: (Left) Test movement of an MT mounted on the RCV. (Right) Simulated (dashed line) and measured (solid line) PLs of BS 12.](image)

4 Conclusion

In this contribution the feasibility of precise PL estimation for the LPM measurement system is shown, which can be used to monitor the measurement quality of the BSs. This additional measurement information can then be used in an Extended Kalman filter to weigh the pseudo-ranges depending on the deviation from the predicted PL.

5 References


Non-Stochastic Multipath Simulations for an Indoor Local Positioning System

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1 Summary
The measurements of an active pulsed reflector in an indoor local positioning system revealed that multipath propagation has a large influence on the positioning accuracy. In this study, a channel model is developed, which calculates the multipath components based on base station and reflector positions in a room. This model allows identifying correlations between the reflector position and the distance error. The distance error is correlated to the channel damping due to multipath propagation with a correlation factor of 0.59. By comparing the distance errors of one reflector measured by 17 base stations, the reflectors with the largest standard deviation are close to a wall. The standard deviation of 3400 simulations is 17.19 cm. This is comparable to the measurement results, which resulted in a standard deviation of 26.6 cm.

2 Introduction
An active pulsed reflector can be used as a backscatter in a FMCW radar based indoor positioning system [1]. The base stations measure the round-trip time-of-flight by detecting the frequency difference in the spectrum (Figure 1a). The measurement results revealed that multipath propagation has a large influence on the positioning accuracy. This paper presents multipath simulations for the active reflector system and evaluates the influence of the position of the reflector to the 1D distance error.

3 Channel Model
In wide-band channel models, the propagation channel is modelled with stochastic parameters. For the simulation of the accuracy of 3D indoor positioning system, these channel models are not suited, because the propagation channels from different base stations are correlated. Our channel model assumes a 10x10x3 m³ large room. With the known position of the base station and the reflector, the multipath and line-of-sight (LOS) components are calculated. In this paper, we present results for an empty room. This approximation is valid e.g. for interactive guiding in a museum. With non-empty rooms, the channel model can still be used, but the calculation of the multipath components will be more elaborate.

4 Simulation Results
The multipath simulations were conducted with 17 base stations and 200 random reflector positions. This results in a total of 3400 different propagation channels. For each of these propagation channels, one distance measurement was simulated. This includes the multipath propagation, the start-up behaviour of the reflector, and the down-conversion in the base station with a subsequent zoom-FFT. Figure 1a illustrates the base band spectrum with and
without multipath propagation. The detected distance error has a standard deviation of 17.19 cm. The distance error is correlated to the channel attenuation due to constructive and destructive interference from multipath components. The correlation factor is 0.5946. The largest distance errors occur close to the two opposite walls. However, the error is not correlated to the distance to the wall, only the standard deviation of the distance errors measured from several base stations is. Figure 1b illustrates the 60 reflector positions with the highest standard deviations. All of these reflectors are less than 2 m away from the closest wall.

Figure 1.a: Base-band spectrum without and with multipath propagation. The frequency difference between the peaks is proportional to the distance $d$ from the base station to the reflector. b: This plot illustrates the 60 reflector positions of 200 with the largest standard deviation of the distance error.

5 Comparison with Measurement Results
The measurement results for the active reflector in a strong multipath environment were presented in [1]. The measured standard deviation of the 1D distance error is 26.6 cm. This value is comparable to the simulated standard deviation of 17.19 cm. The simulation underestimates the error, because it uses a simple multipath model, neglecting multiple reflections and reflections at objects inside the room. Moreover, the laboratory room was smaller and the doors and part of the walls were metallic. Nevertheless, the measured and the simulated spectra are very similar. Thus, the simple channel model enables us to predict the distance error due to multipath propagation and allows testing of the detection and positioning algorithms.

6 Conclusions and Outlook
The presented channel model allows estimating the influence of multipath propagation for indoor positioning systems. The multipath propagation has the highest influence on reflectors close to a wall due to the smallest multipath delay time. The results can be used to find the optimal position of the base stations.

References
Input Amplifier for Sensitivity Improvement in an M-Sequence Radar Front-End

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1 Summary

In this paper, a new input amplifier is presented which is intended to improve the dynamic range of an M-sequence radar front-end. The latter is part of a sensor array and at the current state its dynamic range is confined by spurious tones caused by cross-talk. Owing to the advantageous characteristics of the M-sequence device it is possible to predict the receive signal by digital signal processing and thus to compensate for spurious tones. For this purpose an additional input is provided by the amplifier which is used for signal subtraction.

2 Extension of the basic M-sequence principle

Figure 30 depicts a principal block level schematic of the M-sequence system (solid lines) and the intended extension for sensitivity improvement (dashed lines). Due to the steep clock slopes in this system driven by a square pulse source, jitter is kept low and sampling is absolutely equidistant so that sophisticated digital signal processing is feasible.

As the scene under test varies slow compared to the clock rate and wave propagation, samples of the same chip in consecutive sequences can be averaged to generate a prediction signal. Also, the data rate transferred to the FPGA can be reduced by such a high speed processing without significant loss in accuracy and thus be adapted to the capabilities of the FPGA. The digital prediction signal is then converted to an analogue representation and fed into the second input of the designed input amplifier. This way it is subtracted from the input signal and only the difference signal, usually of small amplitude, is sampled and converted by the ADC. In section 3, the design of such an input amplifier is examined in more detail and first measurement results of the fabricated chip are shown.
3 Input Amplifier with Integrated Signal Subtraction

Focussing the schematic in Figure 31 (a), the principle of operation of this input amplifier can be explained. The signal from the main branch $V_{in+}$ is amplified by a cascoded common emitter amplifier while input matching is achieved by resistive feedback. The second input signal $I_{DAC+}$ is provided by a current steering DAC which implies signal subtraction in the current domain. As the amplifier is driven by a differential signal, subtraction can be performed by two current mirrors $Q_5-Q_6$ and $Q_7-Q_8$ which share a common output node. To ensure output matching, the biasing resistor is split in two portions, $R_{o1}$ and $R_{o2}$, presenting the required impedance to the load.

![Figure 31: Half circuit schematic (a) and chip photo (b) of the amplifier with signal subtraction](image)

Figure 31 (b) shows a chip photograph of the amplifier produced in IHPs 250 nm SiGe:C BiCMOS technology. Measurement results for this chip are presented in Figure 32 in which 1 and 2 are considered to be the differential input and output of the main branch, respectively.

![Figure 32: Transfer characteristics and noise figure for the amplifier with signal subtraction](image)

4 Conclusion and Outlook

In this abstract a new input amplifier for dynamic range improvement of an M-sequence radar front-end is presented. Its main branch is characterized by first measurement results. They show the desired gain curve progression and encourage further tests within the target system.