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SIGNAL CORRELATION FOR TIME DIFFERENCE OF ARRIVAL CALCULATION

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ABSTRACT

This paper presents a signal correlation approach for room-precise indoor localization of non-cooperative mobile phones (GSM standard). Instead of widely used Received Signal Strength Indication we use Time Difference of Arrival information for tri-/multilateration. For an aspired spatial resolution of 3 m and less received signal patterns have to be correlated with respect to path delays in the low nanoseconds range. Receiver quality plays an important role in terms of noise figures and other receiver impairments. We describe the major impacts such as local oscillator phase noise and show how this affects correlation results based on both simulated and measured data.

Index Terms— Indoor Localization, Wireless Communication, Trilateration, Time Difference of Arrival (TDoA), GSM, GMSK Modulation, Correlation.

1. INTRODUCTION

Wireless communication advanced to all fields in life, but also implies new risks, especially in security-relevant areas. Therefore, detection, jamming and localization of mobile phones can help to minimize the dangers resulting from unpermitted use. While detection and jamming is easy with existing technology, accurate and reliable indoor localization is still challenging. Our research aims at developing new methods for room-precise blind indoor localization of non-cooperative active wireless communication devices. The paper focuses on signal correlation needed for Time Difference of Arrival computation using the example of GSM standard mobile phones.

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2. LOCALIZATION

2.1. Sensor Network

The detection hardware used is based on the comstop® system [1], which is a distributed sensor network, where each device is connected to a serial bus and controlled by a server. Spatial localization can be achieved through different methods, where triangulation and trilateration are most known. Triangulation uses information about angle of arrival and needs exactly mounted directional antennas, which is not feasible in indoor environments.

Trilateration however is based on distances, which can be derived from either received signal strength or signal propagation delay [2]. Using received signal strength as a distance indication results in good estimation under line-of-sight condition (e.g. free space), but does not perform well in non-line-of-sight scenarios like urban or indoor environments due to path loss in walls [3]. Signal propagation delay is not affected by this, however, signal processing becomes more complicated.

Generally, fading caused by multi-path propagation has strong effects in this scenario. Using both the signal strength and the propagation delay in a localization scenario will most likely give better localization performance than just a single indicator. Since determination of the propagation delay difference between two receiving stations is crucial, in this paper we will present a method to estimate this difference in a highly efficient way with respect to hardware requirements.

Because extensive multipath propagation is likely to happen in indoor environments, the wireless channel, its variation over time and signal bandwidth will have major impacts on the localization strategy. Therefore, in this first approach we concentrate on widely used 2G GSM handsets (without EDGE/8-PSK) and channels with Dominant Direct Path (DDP) profile.

2.2. Trilateration

For spatial localization of blind mobile phones, when no absolute times or distances can be measured on the

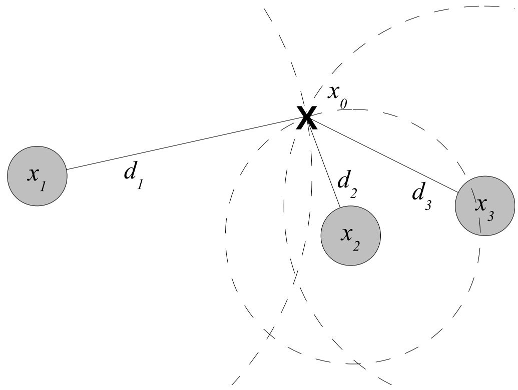

Fig. 1. Trilateration with 3 receivers.

first hand, the Time Differences of Arrival method provides the required information for tri-/multilateration. Assuming n spatially distributed and synchronized detectors (figure 1) with known positions

$$\mathbf{x}_i = [x_i y_i z_i]^T, (i = 1, ..., n), \tag{1}$$

where x_i, y_i, z_i are Cartesian coordinates of the receiver position. The Euclidean distance d_i between detector i with (1) and mobile phone with estimated position vector $\mathbf{x}_0$ is

$$d_i(\mathbf{x}_i, \mathbf{x}_0) = |\mathbf{x}_i - \mathbf{x}_0| \tag{2}$$

$$= \sqrt{(x_i - x_0)^2 + (y_i - y_0)^2 + (z_i - z_0)^2}. \tag{3}$$

With detector $k, (k \in i)$ being the reference, the delay $\tau_i = \frac{d_i - d_k}{c}$ calculated through correlation is the TDoA with respect to position $\mathbf{x}_k$. Since delay and distance are linked by the speed of light $c \approx 3 \cdot 10^8$ m/s, with $d_i - d_k = \tau_i \cdot c$ and (2)/(3),

$$|\mathbf{x}_i - \mathbf{x}_0| = d_k + \tau_i \cdot c. \tag{4}$$

For three-dimensional spatial localization (4) contains four independant unknown variables x_0, y_0, z_0, d_k which have to be estimated, so at least four detectors have to be within range and deliver data to calculate the delays τ_i with respect to the reference detector ($\tau_k = 0$). In reality, it is likely to have more than four receiving devices. Then this system of equations will be overdetermined and an approximate solution can be found using a fitting method like Least Squares [4], [5]. For an intended resolution of 3 m in the spatial domain, time delays as small as $\frac{3\text{ m}}{c} = 10$ ns have to be resolved.

3. DETECTOR

3.1. GSM and GMSK Modulation

Data in GSM networks is transmitted in Gaussian Minimum Shift Keying (GMSK) modulated bursts with a symbol rate of 271 kbauds [6]. Access method for GSM is Time Division Multiple Access (TDMA), where one frame takes 4.6 ms and contains eight timeslots, each with a duration of 577 μs or 156.25 bit times (148 data bits + 8.25 bits guard period).

GMSK is a binary, phase-continuous digital modulation scheme with modulation index $m = 0.5$ [7]. Phase is shifted by $\pm \pi/2$ radians per bit. For higher spectral efficiency baseband quadrature components are filtered using a Gauss filter with a bandwidth-time product $BT = 0.3$ for GSM. Adjacent channels are spaced 200 kHz apart, making a total of 175 channels for a 35 MHz GSM uplink band. Besides TDMA, GSM allows carrier changes on a per-burst basis with a slow hopping frequency of 217 Hz.

Since channel and starting time of a burst transmission are unknown on the detector side, all measurement and calculation needs to be done in relation to a reference, e.g. the detecting device which received the signal first. To determine Time Difference of Arrival, all detectors have to be synchronized (i.e. use a common time base) and received signal patterns must be correlated with the reference pattern. The resulting phase shift yields the propagation delay in relation to the reference.

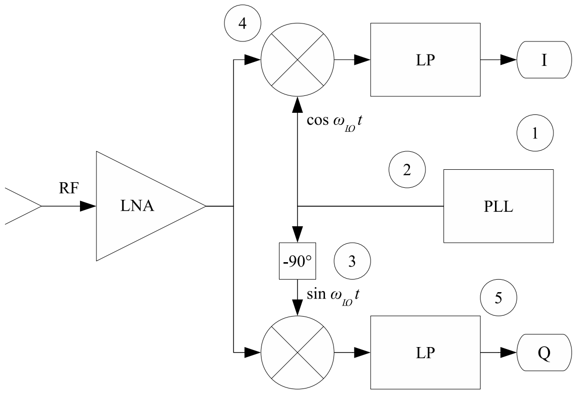

Fig. 2. Receiver architecture.

3.2. Receiver Architecture

Each detector features a phase-locked loop (PLL) controlled direct-conversion receiver which mixes wideband RF (up to 60 MHz) into complex in-phase and quadrature (I/Q) baseband signal. These I/Q components can be used to calculate signal phase $\varphi(t) = \arctan(\frac{Q(t)}{I(t)})$. Since the actual GSM channel (i.e. the exact carrier frequency) is not known a priori and local oscillators (LO) might deviate, baseband signals of the individual detectors cannot be compared directly if high accuracy is necessary. Hence, further digital processing is nessecary which is described in chapter 5.

3.3. Receiver Impairments

Components like PLL, VCO, mixer and filters form an important part of the detector, thus their performance has to be analyzed with respect to signal quality. Tuners can be characterized by mixer impairments and noise figures of oscillators and amplifiers. The following error sources have been identified to influence the results, numbers 1-5 correspond to the positions where the error occurs as depicted in figure 2:

- **(1) Reference frequency offset:**
  Since frequency synthesizers of different tuners are independant of each other, local oscillators might deviate. Continuous phase progresses faster for higher frequencies and has to be compensated in order to correlate signals.
- **(2) VCO/PLL noise:**
  Main sources of local oscillator noise are the various parts of a VCO/PLL as well as their operating parameters. Different noise forms like phase or amplitude noise add to the overall noise power spectral density (PSD) of the oscillator spectrum [8]. During mixing LO noise modulates the input signal and thus directly affects the mixer output. How phase noise relates to cycle-jitter can be seen in [9], and a simple conversion formula is presented in [10].
- **(3) Quadrature skew:**
  Perfect offset of quadrature to in-phase component of the local oscillator is $-\pi/2$ radians. Actual mixers show a skew up to some degrees, the smaller the better.
- **(4) Mixer intermodulation:**
  Main sources of intermodulation are the PLL reference frequency and switching regulators of the power supply. In the spectrum they show as spurious peaks near the carrier at multiples offsets of their own frequency (see figure 3). These effects increase on poorly designed hardware (e.g. lacking separation of digital and analog parts) and add to the overall noise PSD. Another problem is local oscillator leakage which is fed back into the mixer input because of lacking RF/LO isolation. This leads to DC bias and primarily limits the receiver's dynamic range.
- **(5) Baseband I/Q imbalance:**
  Due to hardware tolerances in baseband amplifiers and filters, I/Q outputs can show gain and group delay variation and mismatch as well.

Most static receiver impairments like local oscillator frequency offset can be handled by appropriate post-processing or do not degrade signal correlation accuracy. Particularly complex baseband errors like gain imbalance and quadrature skew can be overcome by sampling only the analog in-phase component and performing a digital demodulation. This does not impose significant additional efforts since digital spectrum analysis and down-conversion is necessary anyway to get the actual carrier frequency.

3.4. LO Noise Measurement

Measurements and simulation results show that local oscillator noise is crucial for accurate correlation results. For this reason, monolithic frequency synthesizers (VCO and PLL combined within one component or even integrated into the mixer) should be preferred over discrete parts.

Most frequency synthesizers specify phase noise numbers as single-sideband (SSB) spectral energy per Hz in relation to the carrier power and frequency offset (e.g. -80 dBc/Hz at 1 kHz). To quantify the effects of receiver impairment on the correlation result, their statistical influence can be determined either on analytical or numerical basis. Since the processing chain contains many non-linear operations, a numerical simulation for both measured and simulated input data is feasible.

For a realistic simulation of this noise source, local oscillator phase noise has been measured for the receiver of our detecing device. For this purpose it can be assumed that any high-quality Spectrum Analyzer's noise figures are good enough to get sufficiently accurate results. For a VCO frequency of 900 MHz measurements have been done with lowest possible RBW of 30 Hz at different offsets from the carrier. Figure 3 shows the power spectral density up to 500 kHz around the carrier.

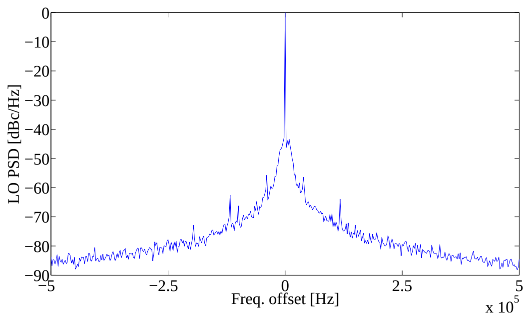

Fig. 3. Power spectral density of LO.

The following numbers have been identified for our detector:

Offset freq. [Hz]	10^3	10^4	10^5	10^6	10^7
PSD [dBc/Hz]	-55	-60	-80	-100	-140

For comparison, the datasheet of a typical monolithic tuner component [11] lists the following phase noise numbers at 1 GHz:

Offset freq. [Hz]	10^3	10^4	10^5	10^6	10^7
PSD [dBc/Hz]	-92	-95	-97	-135	-148

4. SIMULATION

To support our understanding of receiver impairments a MATLAB simulation has been developed. It is controlled by a set of parameters which can be gathered from an actual receiver, either datasheet numbers or measured values (or a combination of both).

Simulation input is a generated ideal RF signal. For completeness the simulation can be enhanced by a preceeding channel model which takes parameters such as transmitter and receiver positions, antenna orientation and environmental constraints.

Most of the error sources are either static or changing slowly (compared to burst duration), e.g. quadrature skew or frequency offsets. Hence, their impact on the correlation result are deterministic. Phase noise on the contrary is highly stochastic and as such it is modelled as spectral sidebands with Additional Gaussian White Noise (AWGN), using the model [12] based on [13]. The simulated effect on signal pattern is shown in figure 4.

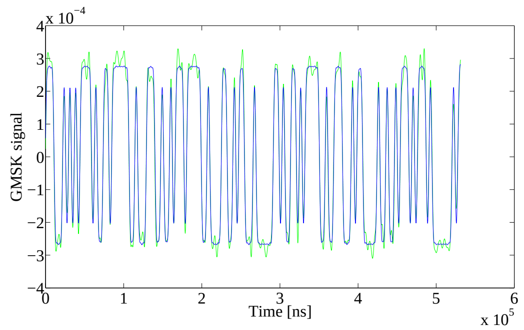

Fig. 4. GMSK pattern (with/without phase noise).

Simulation output consists of a pair of baseband I and Q signals like featured by most current receiver designs. This way measured as well as simulated receiver output is interchangeable and can be used for the following correlation algorithm.

5. CORRELATION

The developed algorithm expects a down-converted signal and calculates its delay against the reference pattern. Since the carrier frequency is unkown, the sampled in-phase component is used for further digital processing (spectrum analysis, complex mixing, phase computation, filtering). At the moment, this algorithm takes a few seconds on a current mid-range personal computer. The following steps are performed, also represented as scheme in figure 5:

- **I Sampling** with $t_s = 1$ ns
- **II Processing**
  - Spectrum analysis
  - Mixing with I/Q demodulation
  - Low-pass filtering
  - Calculation of phase information
  - Filtering
- **III Cross-correlation** with reference

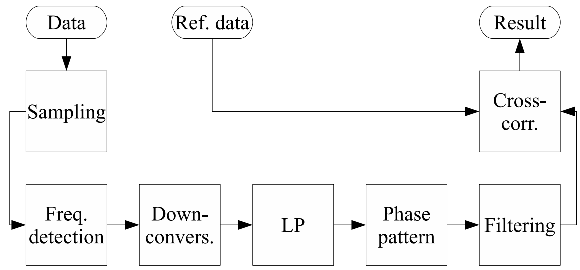

Fig. 5. Correlation scheme.

For a spatial resolution of some meters a sampling time $t_s < 10$ ns is necessary, therefore preprocessing performs data down-sampling of the complete burst pattern with $t_s = 1$ ns. This default sample time helps keeping the simulation simple and the number of filter variants small.

The resulting vector of approx. 500 k samples has to be processed, starting with spectrum analysis to find active channels. At the moment, only a single channel can be handled. The signal is then digitally demodulated using the detected carrier frequency as local oscillator and low-pass filtered with a 60 dB channel filter. From the resulting I/Q signal the phase information is extracted which is continous over π-borders.

Finally, the cross-correlation is calculated at a reasonable number of offsets at steps of t_s. The offset with the highest correlation value then is the estimated delay compared to the reference signal.

6. RESULTS

6.1. Simulated Input Data

Our simulations showed that local oscillator phase noise is the most important error source and has the biggest impact on correlation accuracy. With all other error sources being absent, for a signal delayed by 10 ns (3 m) against the perfect reference signal, three relevant setups have been analyzed:

- **A:** Without phase noise (ideal case)
- **B:** With strong phase noise,
  RMS cycle jitter $\sigma_{rms} = 0.249$ rad, e.g. discrete VCO/PLL/mixer
- **C:** Weak phase noise
  RMS cycle jitter $\sigma_{rms} = 0.043$ rad, e.g. monolithic tuner

Since the first case is perfectly ideal and deterministic, each simulation run returns the true distance of 3 m. For both other cases B (figure 6) and C (figure 7), 100 simulation runs have been performed to get statistically significant results. It can be seen that for both noisy setups the expected value ($\bar{\tau}_{SimB} = 10.08$ ns, $\bar{\tau}_{SimC} = 10.15$ ns) is very close to the actual number of 10 ns (3 m) due to the uncorrelated nature of this phase noise. It is also evident that standard deviation ($\sigma_{\tau,SimB} = 21.76$ ns, $\sigma_{\tau,SimC} = 4.77$ ns) increases significantly for strong noise.

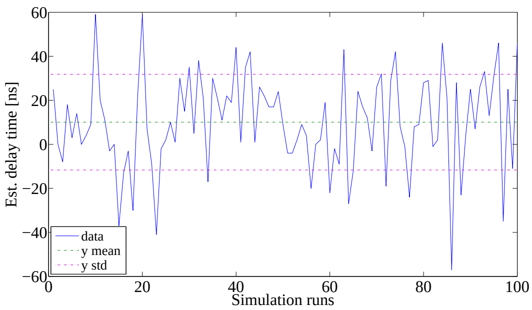

Fig. 6. Corr. results for strong phase noise.

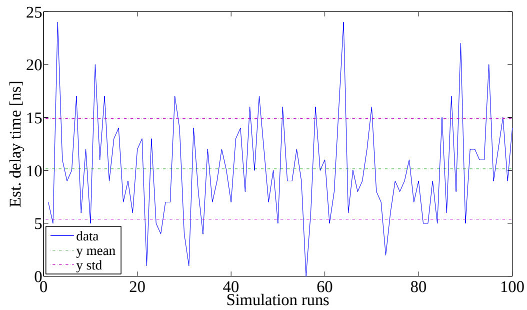

Fig. 7. Corr. results for weak phase noise.

6.2. Measured Input Data

For comparison the simulation setup has been adopted for an actual series of measurements using the existing comstop® receiver hardware.

To minimize dynamic channel behaviour and considering only individual receiver characteristics, the detector positions were choosen to be as close as possible. An artifical signal delay was introduced by extending the measuring line of one detector by a coaxial cable of length $l = 2$ m. Propagation speed within the cable depends on relative permeability $\mu_r \approx 1$ and relative permettivity $\varepsilon_r \approx 2.25$, resulting in

$$c_{coax} = \frac{c}{\sqrt{\varepsilon_r \cdot \mu_r}} \approx 2 \cdot 10^8 \text{ m/s}$$

and an expected delay

$$\tau_{delay} = \frac{l}{c_{coax}} \approx 10 \text{ ns}.$$

61 measurements have been made using a GSM900 cell phone with a distance of 2 m from the transmitter to the detectors. Frequency hopping over eight channels occured, but all measurements were taken into account for higher statistical significance. Data has been sampled by a digital storage oscilloscope and fed into our correlation algorithm. The results of these measurements are shown in figure 8. The estimated value of $\bar{\tau}_{Meas} = 10.05$ ns almost perfectly matches the expected value of 10 ns (3 m), although standard deviation $\sigma_{\tau,Meas} = 30.31$ ns is large.

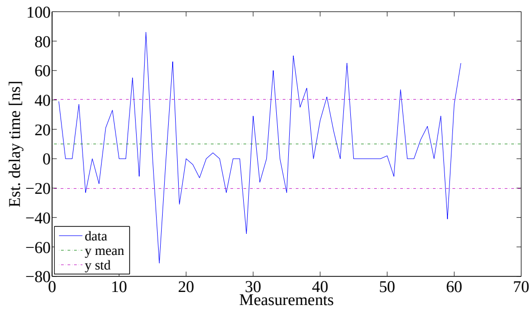

Fig. 8. Corr. results for measuremens.

7. CONCLUSION

In this paper we presented a correlation algorithm for Time Difference of Arrival calculation for GMSK modulated signals. A receiver simulation has been developed in order to identify and understand major receiver impairments. Both simulation and measurement data consistantly show that local oscillator phase noise has a big impact on correlation accuracy and therefore on localization results. Since the expected value of calculated delay is unbiased, better accuracy can be achieved by increasing the number of measurements. Standard deviation of correlation results for measured data is bigger than for simulated data. This could be expected, because impairments of two independant receivers and channel influence add to the total noise figure.

8. OUTLOOK

Further research has to cover the influence of the wireless indoor channel with respect to multipath propagation, fading and signal modulation/bandwidth. For this purpose our simulation will be extended by a priori data (e.g. 3D room description) to serve as a global model unifying all relevant parameters. Additional work also has to be done on detector hardware and highly precise sychronization. Finally the correlation output has to be classified for further processing within a heterogenous sensor data fusion architecture to be developed.

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