

Distance Estimation Based on Molecular Absorption at THz Frequencies

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Abstract—One of the main approaches for distance estimation is the received signal strength (RSS) based techniques. Their drawbacks include the requirement of accurate knowledge of the transmit power and antenna gains. Also, the traditional RSS-based techniques do not take into account the molecular absorption that occurs at terahertz frequencies. In this paper, we propose a distance estimation method for the line-of-sight case that actually takes advantage of the molecular absorption. The proposed method measures the RSS at two frequencies and does not require known transmit powers and antenna gains, which are assumed to be almost the same at the two frequencies. Therefore, the two frequencies have to be relatively close to each other. The proposed distance estimation method calculates the difference of the received powers (in the dB domain) at the two frequencies and finds the link distance based on it and the humidity level.

Index Terms—Distance estimation, THz communications, molecular absorption, positioning.

I. INTRODUCTION

The exponential growth of data traffic volume during the last decade is expected to continue [1]. Such growth is a result of disruptive technologies such as virtual reality [2] and machine-to-machine communications in the Internet of Things [3]. As a result, it is expected that current wireless communication infrastructures are unable to cope with such traffic demands [4]. Terahertz (THz) Band (0.1–10 THz) is considered one of the key enabling technologies to satisfy such high data rate demands for 6G and Beyond wireless systems [5]. One drawback of the THz link is that it can be highly attenuated by atmospheric water vapour [6]. Also, the free space path loss can be large. Therefore, especially higher THz frequencies are more suitable for short-range wireless communications.

The link distance estimation between two nodes is important for multiple applications such as wireless sensor networks (WSNs). In addition, it is desired to be able to estimate the nodes' separation without installing specialized hardware for distance estimation in low-cost WSNs. In the literature, various techniques have been proposed for localization and distance estimation [7]. The received signal strength (RSS) based approach has gained much attention due to its simplicity and suitability for estimation of short distances [8]–[10]. Other techniques include utilizing the time of arrival (TOA) which measures the propagation time of a radio signal traveling between a sender and a receiver to estimate their distance.

In practice, TOA-based ranging methods require accurate knowledge of the time of the transmission for calculating the time of fly [11]–[13].

In this paper, we measure the RSS at two close-by frequencies and use the difference (in the dB-domain) of the received powers and channel propagation characteristics, specifically, humidity level for the link distance estimation. The basic idea of the proposed RSS-based link distance estimation is simple. Let us assume that we have two frequencies with molecular absorption values of, say, 5 dB/m and 8 dB/m for a given humidity level. The difference in frequency leads to different loss in molecular absorption. Assume that the transmit powers and the antenna gains are the same at these two frequencies, which thereby typically need to be close to each other. If the received power differs 6 dB between these two frequencies, we can infer the distance to be 2 meters. We do not need to estimate or measure the absolute power levels, just their relative difference. The reason this works is that the molecular absorption can be specified in terms of dBs per meter unlike for example free space path loss.

TOA-based ranging method requires accurate knowledge of the time of the transmission, while the proposed method requires more sensors (humidity and temperature sensors). The proposed method does not need to know a lot of information, which make this method suitable for example for IoT devices on THz frequencies.

In practice, measuring the distance based just on the power difference is not straightforward, since we have to take into account for example path loss variation as a function of frequency, the impact of phase noise etc. However, if the frequency gap is not too large, it should be doable.

The distance estimation performance improves as the difference in molecular absorption among the chosen two frequencies increases. However, there are practical problems with measuring very low signal powers so frequencies with very large molecular absorption should be avoided. Therefore, the approach is most promising for the THz band. Also, if all frequencies have the same level of molecular absorption (flat part of the absorption curve), the link distance estimation cannot be done using the approach proposed in this paper. In the lower frequencies, the molecular absorption due to the oxygen at around 60 GHz might differ enough compared to the molecular absorption due to the humidity at the higher frequencies.

For example, devices based on the IEEE 802.11ay standard are operating at around 60 GHz. However, the level of the molecular absorption at 60 GHz is less than the values typical at the THz frequencies, while the proposed method could still be applied with longer link distances. At the THz frequencies, for example, device-to-device (D2D) communication systems may prefer operation at frequencies with significant molecular absorption with the so called molecular absorption peaks due to the reduced interference between systems [14]. In fact, the D2D systems may choose a suitable frequency based on the typically unknown link distance [14].

II. SYSTEM MODEL

The system model for distance estimation is shown in Fig. 1. We assume a transmitter and a receiver with a humidity sensor, a temperature sensor, and possibly a pressure sensor in the system. Our aim is to estimate the distance between the transmitter and the receiver. The distance estimation is performed at the receiver based on the received signals. Specifically, an energy detector is used for distance estimation [15].

The proposed distance estimation approach can be used with normal wireless communication signals, especially with wideband single-carrier modulation so that the same modulation symbol is affecting both frequencies. Distance can be estimated at the same time as the wireless communication is taking place. It is also possible to transmit a dedicated signal for distance estimation.

In the channel propagation of the THz band, the molecular absorption is a unique feature of the THz frequencies compared to lower frequencies. The molecular absorption loss depends on not only the frequency but also distance, humidity level, and temperature level. The proposed distance estimation exploits the feature of the molecular absorption loss.

The received power model for the assumed line-of-sight case including the free space path loss and molecular absorption loss is shown in Section IV. Friis's free space path loss is used but the fading effect caused by multipath components is not considered so performance evaluations assume non-rich multipath environment where the effect of multipath component is relatively lower. The proposed technique does not apply a radar-type technique using the reflected signal from the receiver with some radar cross section.

III. DISTRIBUTION OF RECEIVED SIGNAL STRENGTH

Let us show the model of the received signal strength. We assume that the received power is estimated with a standard energy detector. Assume that the received signal power is P_R and let us denote the noise power with N_{pow} . Ideally, the noise power is $N_{pow} = kT_K B$, where k is the Boltzmann's constant, T_K is the temperature in Kelvin, and B is the bandwidth of measurement. We further assume that there are $2K$ degrees of freedom in the output, where K can be the number of complex time-domain samples the energy detector output is based on K complex baseband inphase-and-quadrature samples. In practice, we may not need access to the complex baseband

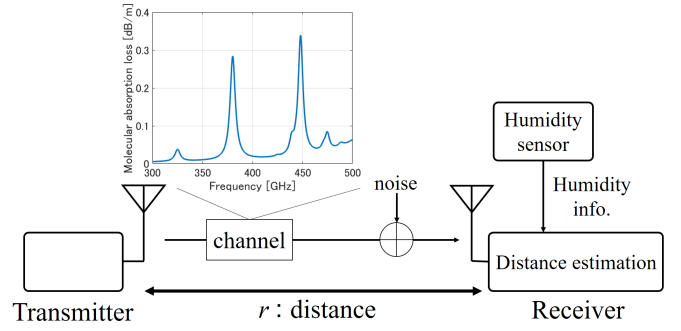


Fig. 1. System model.

samples if the energy detector is equivalently implemented with analog components.

Usually, the normalized energy detector output is assumed to follow the non-central chi-squared distribution with $2K$ degrees of freedom with noncentrality parameter being in our case $\lambda = 2KP_R/N_{pow}$. We divide the energy detector output with K to get power per sample. Now, based on the theory of the non-central chi-squared distribution, we get that Gaussian approximation for the measured received power X has mean $\mu = N_{pow} + P_R$, so that the average measured power is the sum of the noise power and the received signal power. We notice that noise power limits how low received signal powers we can measure. In theory, it would be possible to subtract the noise power from the received power but this would require accurate knowledge of the noise power. The uncertainty of noise power is a well-known limitation of energy detectors [16]. The variance of the Gaussian approximation of X is obtained as $\sigma^2 = N_{pow}(N_{pow} + 2P_R)/K$. We notice that as K approaches infinity, the variance of the measured power approaches zero. However, it is of course not practical to measure for very long times.

We approximate also the distribution of received power in the dB domain as Gaussian. In this case, based on first terms from the Taylor expansion of $10 \log_{10}(X)$, see https://en.wikipedia.org/w/index.php?title=Taylor_expansions_for_the_moments_of_functions_of_random_variables&oldid=1063034706 and [17], we get as approximation that the mean of the $10 \log_{10}(X)$ is

$$\mu_{LOG} = 10 \log_{10}(P_R + N_{pow}) \quad (1)$$

and variance is

$$\sigma_{LOG}^2 = \frac{100 N_{pow}(N_{pow} + 2P_R)}{\ln(10)^2 (N_{pow} + P_R)^2 K}, \quad (2)$$

where \ln means the natural logarithm.

Fig. 2 shows an example of the Gaussian approximation in the dB-domain. It can be observed that approximation is good.

Let us assume that we measure received power at two center frequencies, f_1 and f_2 . Let us denote the measured (noisy) signal powers X_1 and X_2 . Also denote the received signal powers with $P_{R,1}$ and $P_{R,2}$. Now $X_{dB,1} = 10 \log_{10}(X_1)$ and

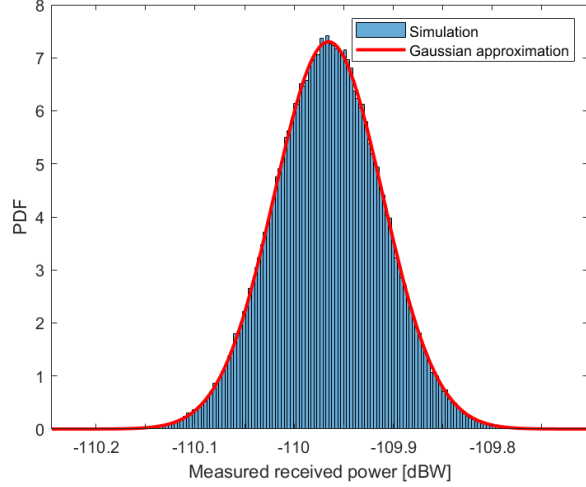


Fig. 2. Gaussian fit to received power when $P_R = 10^{-11}$ W (-110 dBW / -80 dBm), $K = 100$, $N_{pow} = 8 * 10^{-14}$.

$X_{dB,2} = 10 \log_{10}(X_2)$ both have the Gaussian approximation with mean given by (1) and variance given by (2).

IV. RECEIVED POWER MODEL

The well-known Friis's equation for free space propagation tells that received signal power P_R is

$$P_R = P_T \frac{G_T G_R c^2}{(4\pi r f)^2}, \quad (3)$$

where P_T is the transmitted power, G_T is the gain of the transmitting antenna, G_R is the gain of the receiving antenna, c is the speed of light, r is the link distance, and f is the frequency.

Based on the Friis's equation we can estimate the link distance with

$$r_{\text{Friis}} = \sqrt{\frac{P_T G_T G_R c^2}{P_R (4\pi f)^2}}. \quad (4)$$

A drawback with the approach above is that receiver needs to be able to measure the absolute power levels, it must know the transmit powers, as well as the transmitter and the receiver antenna gains. What is more, it will not work well in the THz frequencies, where molecular absorption is present, as molecular absorption does not follow the square-law behavior in terms of distance.

In the THz frequencies, we need also to include the molecular absorption when considering the received power. The molecular absorption factor is well known to be [18]

$$e^{-k(f)r}, \quad (5)$$

where $k(f)$ is a frequency-dependent absorption coefficient that can be obtained numerically using for example the am software [19] or the ITU model ITU-R P.676-8, where the ITU model is up to 1000 GHz [20]. We [21] have proposed

a simplified model for calculating the molecular absorption coefficient. Also the humidity level needs to be known since it is the main contributing factor to $k(f)$. This can be obtained with a humidity sensor in a node. A temperature sensor is typically also needed.

Combining the Friis's equation and molecular absorption loss, we get the received signal power as

$$P_R = P_T \frac{G_T G_R c^2}{(4\pi r f)^2} e^{-k(f)r}. \quad (6)$$

The equation above can also be solved for r provided that P_T , G_T , and G_R are known. We assume to store $k(f)$ into a local memory as a function of the humidity. The result is

$$r_{\text{comb}} = \frac{2}{k(f)} W_0 \left(\frac{k(f) \sqrt{\frac{G_T G_T P_T c^2}{16 P_R f^2 \pi^2}}}{2} \right), \quad (7)$$

where W_0 is the principal branch of the Lambert W function [22].

We can write (6) in the dB domain as

$$P_R[\text{dB}] = \alpha - 20 \log_{10}(r) - ar, \quad (8)$$

where $\alpha = 10 \log_{10} \left(P_T \frac{G_T G_R c^2}{(4\pi f)^2} \right)$ and $a = \frac{10k(f)}{\ln(10)}$. In [23, Eq. (9)], an iterative solution has been presented. For free space path loss exponent 2, it converges to the same value as our closed-form solution (7). It should be noted that [23] is about attenuation due to rain, not gaseous attenuation. However, the received power can be written in the same form (8) in both cases. In fact, we can interpret $k(f)$ to contain both gaseous attenuation and rain attenuation. However, in this case, a rain sensor or accurate information from a weather database is needed in addition to the humidity sensor to get the value of $k(f)$.

V. PROPOSED DISTANCE ESTIMATION TECHNIQUE

In the case that we do not know P_T , G_T , G_R , let us assume that we measure the received power at two frequencies f_1 and f_2 . We further assume that we know the absorption coefficients at both frequencies with the similar humidity information as discussed above. Assume that the frequencies are sufficiently close to each other so that P_T , G_T , and G_R are the same. The ratio between the received power at f_1 and the received power at f_2 is denoted as γ . We get that

$$\gamma = \frac{f_2^2 e^{-r(k_{f_1} - k_{f_2})}}{f_1^2}. \quad (9)$$

By multiplying with $\frac{f_1^2}{f_2^2}$ and taking the logarithm, we get that the link distance is

$$r_{\text{prop}} = -\frac{\ln \left(\gamma \frac{f_1^2}{f_2^2} \right)}{k_{f_1} - k_{f_2}}, \quad (10)$$

where k_{f_1} and k_{f_2} are the known absorption coefficients. Different to the basic set-up with one frequency measurement above, we do not need to know neither the transmit power nor the antenna gains as typical with received signal strength based

methods. They are just assumed to be equal. We can rewrite (10) in terms of measured noisy received powers in dB ($X_{dB,1}$ and $X_{dB,2}$) as

$$r_{\text{prop}} = \frac{X_{dB,2} - X_{dB,1}}{10 \log_{10}(e)(k_{f_1} - k_{f_2})} - \frac{2 \ln(f_1/f_2)}{k_{f_1} - k_{f_2}}. \quad (11)$$

It can be noted that the distance estimation only depends on the difference of the received powers in the dB domain. The absolute received power levels are not needed avoiding the necessity of accurate calibration schemes.

VI. ANALYSIS OF THE PROPOSED TECHNIQUE

As the received powers in dB can be approximated as independent Gaussians, since the thermal noise is independent in two different frequency channels, (11) has subtraction of one Gaussian from another Gaussian. Then we get the Gaussian approximation for the estimated distance with the proposed algorithm to have mean

$$\mu_{\text{prop}} = \frac{E[X_{dB,2}] - E[X_{dB,1}]}{10 \log_{10}(e)(k_{f_1} - k_{f_2})} - \frac{2 \ln(f_1/f_2)}{k_{f_1} - k_{f_2}}, \quad (12)$$

where for example $E[X_{dB,2}] = 10 \log_{10}(P_{R,2} + N_{\text{pow}})$ as given by equation (1). The variance of the estimated distance

$$\sigma_{\text{prop}}^2 = \frac{\text{Var}[X_{dB,2}] + \text{Var}[X_{dB,1}]}{(10 \log_{10}(e)(k_{f_1} - k_{f_2}))^2}, \quad (13)$$

where for example $\text{Var}[X_{dB,2}]$ is (as given by (2))

$$\text{Var}[X_{dB,2}] = \frac{100 N_{\text{pow}}(N_{\text{pow}} + 2P_{R,2})}{\ln(10)^2 (N_{\text{pow}} + P_{R,2})^2 K}. \quad (14)$$

Now we have a fully theoretical model for the distance estimation accuracy of the proposed technique.

VII. NUMERICAL RESULTS

Fig. 3 shows a comparison between simulated histogram and the theoretical probability density function (PDF) by the Gaussian approximation for the estimated distance. The used parameters are $f_1 = 377$ GHz, $f_2 = 378$ GHz, transmitter antenna gain 30 dB, receiver antenna gain 20 dB, $K = 100$, $N_{\text{pow}} = 8 * 10^{-14}$. The proposed method works also at other frequencies given that there is sufficient difference in absorption coefficient.

The water vapour density, which comes from the relative humidity level ¹, is 7.5 g/m^3 . The transmitted power inside each measured band is 1 mW. True distance $r = 10$ m. It can be observed that the proposed Gaussian approximation is very accurate. It can also be seen that the estimator is unbiased for these input parameters.

For results shown in Fig. 3, the mean square error of the proposed estimator is 0.0089. The mean square error with the traditional Friis-based approach using (4) was much larger, more than 3. This shows that molecular absorption may have to be taken into account or performance can be greatly reduced. When taking into account molecular absorption with (7), performance with the Friis-based approach greatly improved

and the mean square error was ≤ 0.0005 . As expected, performance is better than with the proposed approach since more information is assumed to be known.

Fig. 4 shows the mean square error of the distance estimation with error in the humidity sensor. It can be seen that the proposed method is sensitive to the error in the humidity. We see that the positive humidity error is better than negative humidity error. The shown theoretical values are obtained by calculating the sum of theoretical bias squared and the theoretical variance of the estimated distance. It can be seen that when there is only a small error in the humidity value, increasing the value of K significantly improves performance. However, when there is a significant error in the humidity values, increasing the number of samples K does not much improve performance. The reason is that there is a bias component that cannot be reduced by increasing the amount of averaging. Increasing the number of samples will reduce the variance but not bias.

Fig. 5 shows the mean square error of the distance estimation as a function of the true distance. As expected, the mean square error increases with the true distance. We can see that increase of the mean square error is not linear as a function of the distance but more like exponential. However, the approach is promising for reasonable distance range for indoor use cases.

Fig. 6 shows the mean square error with two different noise figures (NFs), 0 dB and 20 dB. It can be seen that performance with a large noise figure is significantly reduced due to the increased power of the noise at the receiver. However, even with large noise figure, when the number of samples (K) is increasing, performance is improving. This means that bias component is not dominating since increasing K mainly affects the variance. At very larger number of samples performance improvement starts gets less, the reason is the bias component.

Please note that the MSE difference appears very small (less than 0.01) in Fig. 4 for $K = 100$ and $K = 1E5$, compared to MSE difference in Fig. 6 for $K = 100$ and $K = 1E5$ with both NF = 0 dB and 20 dB due to use of linear scale vs logarithmic scale.

THz radiation power can be quite limited so it is important to evaluate performance with low transmit power. Fig. 7 shows the mean square error as a function of the power inside each measured band. In case where normal wireless communication signals are used for distance estimation, the power inside each measured subband comes from the fraction of the total transmit power of the communication signal within the two measured sub-bands centered at f_1 and f_2 . When a dedicated signal is transmitted for distance estimation, power can be focused on the sub-bands leading to potentially larger power values. We can see that when power is reduced performance gets much worse as expected. If the power inside each measured band is small, performance can be significantly improved by increasing the number of samples K .

From the above discussion we note that proposed approach is sensitive the noise figure and the power within the two sub-

¹see for example <https://www.cactus2000.de/js/calchum.pdf>

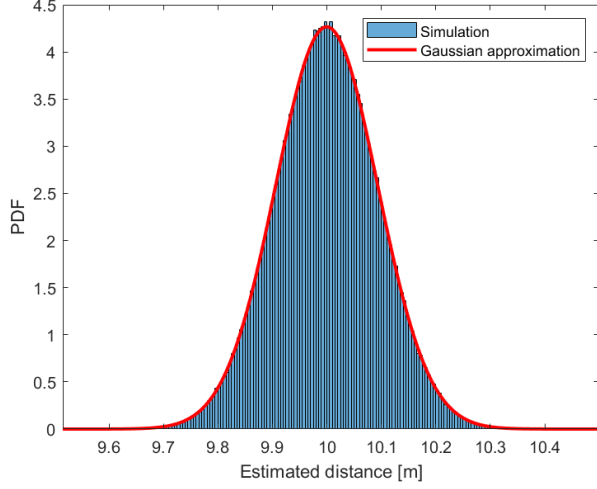


Fig. 3. Probability density function (PDF) of the estimated distance, $f_1 = 377$ GHz, $f_2 = 378$ GHz, transmitter antenna gain 30 dB, receiver antenna gain 20 dB, $K = 100$, $N_{pow} = 8 \times 10^{-14}$. Water vapour density is 7.5 g/m^3 (no error assumed in the humidity sensor). Transmitted power inside each measured band 1 mW. True distance $r = 10$ m.

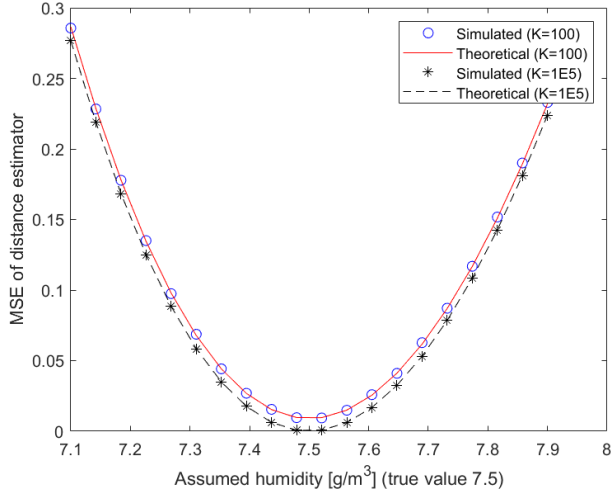


Fig. 4. Mean square error of estimated distance with error in the humidity sensor, $f_1 = 377$ GHz, $f_2 = 378$ GHz, transmitter antenna gain 30 dB, receiver antenna gain 20 dB, $N_{pow} = 8 \times 10^{-14}$. True water vapour density is 7.5 g/m^3 . Transmitted power inside each measured band 1 mW. True distance $r = 10$ m.

bands. However, as the results show, problems related to these can often be addressed by increasing the number of samples K . The proposed approach is also sensitive to error in the assumed humidity value. This problem cannot be solved by increasing the averaging but instead, in case of problems, more accurate humidity sensor is needed. In case of THz, there may be limitation in transmit power and significant spreading loss. Then, the noise power at the receiver may determine the potential of distance estimation in THz.

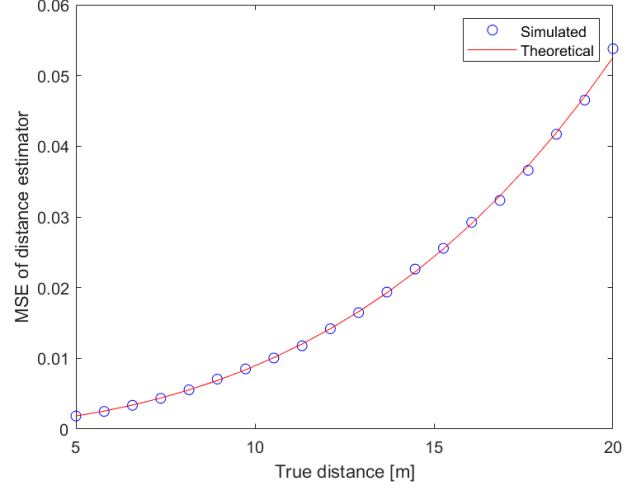


Fig. 5. Mean square error of estimated distance as a function of the true distance, $f_1 = 377$ GHz, $f_2 = 378$ GHz, transmitter antenna gain 30 dB, receiver antenna gain 20 dB, $K = 100$, $N_{pow} = 8 \times 10^{-14}$. True water vapour density is 7.5 g/m^3 (no error assumed in the humidity sensor). Transmitted power inside each measured band 1 mW.

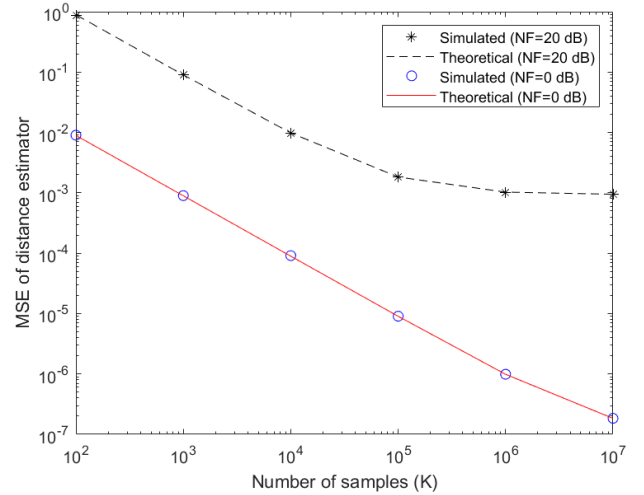


Fig. 6. Mean square error of estimated distance, $f_1 = 377$ GHz, $f_2 = 378$ GHz, transmitter antenna gain 30 dB, receiver antenna gain 20 dB, ideal noise level (noise figure NF = 0 dB) $N_{pow} = 8 \times 10^{-14}$. Water vapour density is 7.5 g/m^3 (no error assumed in the humidity sensor). Transmitted power inside each measured band 1 mW. True distance $r = 10$ m.

VIII. CONCLUSION

In this paper, a simple distance estimation method was proposed for THz links. It is based on assuming that the receiver is equipped with a humidity sensor so that it can calculate the molecular absorption coefficients using a known model at different frequencies. The receiver measures the received signal strength at two frequencies and calculates the received power ratio in the linear domain or the difference in the dB-domain. Based on the measured difference and the molecular absorption coefficient, the proposed method can find the distance between the transmitter and the receiver. The

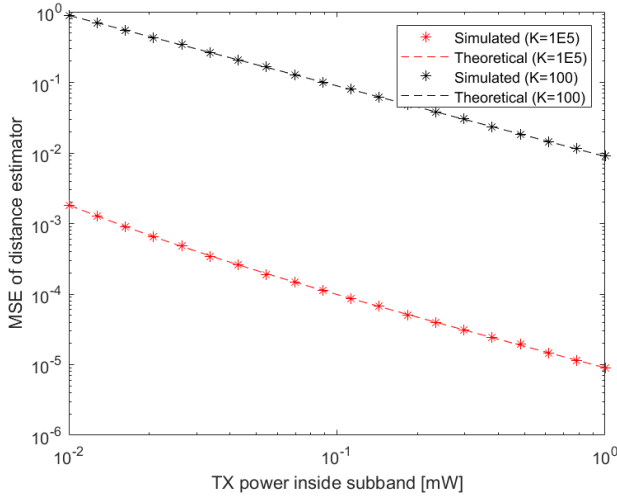


Fig. 7. Mean square error of estimated distance, $f_1 = 377$ GHz, $f_2 = 378$ GHz, transmitter antenna gain 30 dB, receiver antenna gain 20 dB, noise level $N_{pow} = 8 \times 10^{-14}$. Water vapour density is 7.5 g/m^3 (no error assumed in the humidity sensor).

results modeled two sources of randomness: the thermal noise and the error in the humidity sensor output value. The present work is mostly suited for single carrier systems where the same modulation system is affecting both measured frequencies.

In the future work, a third source of randomness due to different modulation symbols at different frequencies could be studied. It is particularly relevant for multi-carrier systems with large peak-to-average power ratios of the transmit signals. Also, effect of multipath could be considered. Since all practical aspects are difficult to model in a simulation study, we also plan to verify the performance with practical measurements in the future work.

ACKNOWLEDGMENT

This work was supported by the Academy of Finland 6Genesis Flagship (grant no. 318927).

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