Channel Modeling and Performance of Zigbee Radios in an Industrial Environment

Mehrdad Damsaz, Derek Guo, Jeff Peil, Wayne Stark Electrical Engineering and Computer Science University of Michigan Ann Arbor, MI 48109 Nader Moayeri, Richard Candell National Institute of Standards and Technology Gaithersburg, MD 20899

Abstract—In this paper we describe measurements of wireless propagation characteristics to develop path loss models in industrial environments. The models for path loss we develop are two-slope models in which the path loss is a piecewise linear relation with the log distance. That is, the path loss is a inverse power law with two regions, two exponents and a break point, that are optimized to find the best fit to the measured data. Second, the multipath power delay profile is determined. We use a reference measurement and the CLEAN algorithm for processing the measurements in order to determine an estimate for the impulse response of the channel. From this the delay spread of the channel can be determined. Finally we discuss the performance of Zigbee receivers. We compare the performance of different receiver structures for the O-QPSK type of modulation used as one Zigbee physical layer.

I. INTRODUCTION

The pervasive application of wireless communications is well known. One such application is to an indoor factory environment. This environment creates challenges for reliable communications. One potential communication system that could be used to provide wireless communication in this environment is a Zigbee radio. Zigbee radios use of 2 MHz of bandwidth in the 2.4 GHz ISM band. To understand the performance of a Zigbee radio (or any other radio) in this environment the first task is to understand the propagation effects of such an environment. The National Institute for Standards and Technology (NIST) has carried out a measurement campaign at various factory environments, including their own machine shop. We have used these measurements to develop channel models that are suitable for evaluating the performance of wireless communication systems with bandwidth up to about 20 MHz such as a Zigbee radio or a WiFi (802.11) based

system. This paper presents the results of processing the measurements to obtain channel models. From the measurements we determine the propagation loss as a function of distance, the shadowing level, the rms delay spread of the channel. Finally, we present results on the performance of a Zigbee radio when used on the channels considered.

The rest of the paper is organized as follows. In Section II we describe the measurements and the methodology to determine the impulse response for a particular transmitter, receiver location. In Section III the path loss models are described. In section IV the methodology to generate the impulse response from the measurements are discussed. The performance of the Zigbee physical layer is discussed in Section V followed by conclusions.

II. MEASUREMENTS

The channel measurement or sounding campaign was carried out by NIST at various factory or factory-like environments. One location was the NIST machine shop in Gaithersburg, MD. Another location was an automotive assembly plant. NIST researchers from Boulder, CO transmitted a sounding signal and measured the response. The NIST researchers used a cart containing a mobile receiver that moved along a set path defined and measured the received signal from a transmitter located in the shop. The transmitted signal was a 40 MHz wideband signal using a pseudo-noise signal (m-sequence) that was mixed to a carrier frequency (2.4 GHz and 5 GHz). The receiver mixes the received signal to baseband and then samples the signal at an 80MHz rate. The transmitted signal was generated from 8188 (=4x2047) samples from an pseudo-noise sequence (m-sequence) generator. The receiver then sampled the received signal after mixing down to baseband with an IQ demodulator.

Figure 1 shows the layout of the machine shop in Gaithersburg where one set of the measurements were

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Fig. 1. Layout of Room

made. There are a number of industrial machines in the room. The receiver was moved from the "start" location through the room and ended up back at the start (shown as location 11 on map). Various check points with known locations (e.g. locations "Start", 1,...,11) were identified with particular acquisitions of received responses. In between these known locations for certain acquisitions the location was determined by assuming that the receiver moved at a constant speed. By knowing the coordinates of the different check points and the associated measurements, the location of the receiver for other measurements were taken. Various antenna configurations (e.g. polarizations) and two different transmitter heights were used for different runs.

The basic setup of the channel sounding is illustrated in Figure 2. The transmitter and receiver have clocks that were initially synchronized. While this would allow accurate determination of the delay, it was not essential in the measurements channel models we developed. The transmitted signal was a m-sequence of length 2047 sampled four times per chip and then up-converted to a carrier frequency.





The PN code is an m-sequence of length 2047 using



Fig. 3. Output of matched filter for reference system

shift register feedback connection. The signal is generated by first mapping the m-sequence values, 0 and 1, to +1 and -1 respectively and then repeating each chip four times at a sample rate of 80 M samples/second. The duration of the signal is $T = 8188/(80 \times 10^6) = 102.35 \mu s$. Corresponding to each transmission there is a recording of the received signal after mixing down to baseband. The recorded signal is a complex signal corresponding to an IQ demodulator.

In order that the equipment not influence the estimation of the channel characteristic, a measurement was made with only an attenuator inserted between the transmitter and receiver (without the antennas). This reference measurement provides a baseline for determining the effect of the antennas and the channel but not the measuring equipment.

To determine the equipment and channel characteristics (e.g. impulse response) we process the received signal with a filter that is matched to the transmitted signal from the m-sequence generator at the transmitter. The magnitude of the normalized output of the filter matched to the m-sequence is shown in Figure 3 where the normalization is such that the peak output value is 1 (0dB). Figure 3 shows the output due only to the equipment without any channel but with an attenuator between the transmitter and receiver. The sidelobes of the response are roughly 35dB lower than the main lobe (at zero delay). In order to accurately estimate the channel we will "remove" the effect of the sidelobes of the reference signal using a CLEAN-type algorithm.

III. PATH LOSS MODELS

There are several parts of our effort to characterize the channel. The first part is to determine the average received power as a function of distance and to generate an appropriate model. In this part of our characterization it is only the received power that is of importance, as opposed to the actual channel impulse response, which we will calculate later. To determine the path loss we measured the power in the received signal and then compared that to the power in the reference signal (the signal received when the antennas were replaced by an attenuator). By taking into account the attenuation used without the channel and the power of the reference signal we can determine the path loss of the channel (including the antennas) at each distance. The average received power as the receiver moved through various places is shown in Figure 4 for one particular run with one particular type of antenna polarization. The number of measurements for a particular polarization and frequency and transmitter location was 10,500. Each of these is called an acquisition. The average received power as a function of acquisition number and the distance as a function of acquisition number is shown in Figure 4.



Fig. 4. Received Power vs. Acquisition and Distance vs. Acquisition, Cross Polarization, 2.4 GHz, Transmitter Location 1.

Clearly there are various power levels received at a given distance. This is the shadowing of the channel, typically modeled as a lognormal random variable with a certain variance. The path loss models the average received power as a function of distance. We will discuss the shadowing (that adds a variance to the average received power) later. The model for the average received power as a function of distance is typically an inverse power law where the power received is inversely proportional to the distance raised to a power: $P_r = k/d^{\alpha}$ for $d > d_0$. Various estimates for the parameters (k, α ,

and d_0) have been developed for the (average) path loss, PL(d), expressed in dB as a function of distance. As a baseline the free space path loss for f = 2450 MHz (assuming isotropic antennas) is [1]

$$PL(d) = 40.28 + 20 \log_{10}(d)$$

This is a single slope relationship between the distance and path loss since when the path loss in dB is plotted versus distance on a log scale it results in a straight line with a single slope. The generic single slope model for path loss (in dB) is

$$PL(d) = PL(d_0) + 10\alpha \log_{10}(d/d_0), d > d_0.$$
(1)

Wloczysiak [2] has proposed the following received power model for indoor applications, although exactly what type of indoor environment is not specified (industrial versus residential versus office).

$$PL(d) = 50.3 + 40 \log_{10}(d).$$

This model has a slope of 40dB decrease in power per decade of distance, or a received power exponent of 4. In this case $PL(d_0) = 50.3$ and $\alpha = 4$ and d_0 is larger than roughly 10m. Li et al. [3] have proposed models for the received power in a residential environment. The model has additional attenuation for going through walls and for going through floors. These models have a range of slopes between 1.13 and 1.61 for different houses with an overall proposed model with $\alpha = 1.37$.

Monti [4] has proposed a path loss model based on measurements in an office-like environment:

$$PL(d) = 54.5 + 16.4 \log_{10}(d)$$

which has a path loss exponent of $\alpha = 1.6$. Jansen et. al. [5] also proposed models for indoor radio channels in an office/laboratory like environment. A range between 1.86 and 4.46 is given for the path loss exponent. Larger path loss exponents are given for non line-of-sight environments than line-of-sight. Tanghe et. al. [6] proposed path loss models in an industrial-like environment (e.g. manual or automated production line and warehouse). Their path loss models are of the form given in (1)where both α and $PL(d_0)$ are chosen to provide the best fit. They call this the non-fixed intercept model compared to the fixed intercept models described earlier. In the non-fixed intercept models the value of $PL(d_0)$ is chosen to minimize the mean square error of the fit along with the path loss exponent α . The models in [6] have exponents between 1.52 and 2.16 depending on line-of-sight power models for indoor radio channels and $PL(d_0)$ between 67.43 and 80.48dB. For 2.4 GHz frequencies they have the following parameters for the best match choice for $PL(d_0)$. Here we distinguish between different environments between the transmitter and receiver: line of sight (LOS), non line of sight (NLOS) and combined.

Conditions	$PL(d_0)$	α
LOS	67.43	1.72
NLOS (light clutter)	72.71	1.52
NLOS (heavy clutter)	80.48	1.69
All	71.84	2.16

Finally another paper [7] for industrial applications uses the single slope model with a fixed intercept to obtain a path loss exponent between 1.86 and 2.7.

In our model we use a two slope model in which at close in distances the path loss has one slope and at a larger distance the path loss has a second slope. The transition between the two different slopes is optimized to obtain the best overall least squares fit. Our model then is a piecewise linear in that over some initial range of distances there is one value for the slope, α , and then at larger distances there is a second value for α . The path loss in dB then has the form

$$PL(d) = \begin{cases} k_1 + \alpha_1 10 \log_{10}(d), & d < \beta \\ k_2 + \alpha_2 10 \log_{10}(d), & d > \beta \end{cases}$$

with the boundary condition that the path loss is continuous where the slope changes. This model has the advantage that the slope is not influenced by measurements very close to the transmitter. At such distances the path loss is relatively unimportant because the received power will be relatively high (except perhaps to determine amount of interference generated). The model is also simple enough to be used without undue complications. Other approaches, like a second order regression could also be used but would seem to be more complicated. Our model at sufficiently large distances is just an inverse power law model with essentially the minimum distance of applicability determined. The channel model constants $k_1, k_2, \alpha_1, \alpha_2$ and β are to be determined from the measurements. Some of the results for this model are shown below based on the Gaithersburg measurements. In our measurements we have some minimum distance (about 2 meters) and some maximum distance (about 40 meters). We plot the generated model as a solid line between these two limits and a dashed line at smaller distances than the minimum and larger distances than the maximum. Figure 5 shows the attenuation for a 2.4 GHz system with horizontal polarized antennas. Figure 6 shows the attenuation for a 2.4 GHz system with vertical

polarized antennas. Figure 7 shows the attenuation for a 2.4 GHz system with cross polarized antennas. The data for vertical polarized antennas mostly follows a single slope model but for a few distances the attenuation shows an increase in the attenuation. The two slope model finds the best break point between the two slopes and the best slopes such that continuity is maintained. By separating the two regions and finding the optimal α_1, α_2 , and β we can find accurate models for the path loss at distances where the received power level is important. The Matlab Shape Language Modeling toolbox was used to find the best parameters for these models. In Figure 8 we compare the models for different polarizations. As can be seen over a range of distances between 10 and about 30m the path loss exponent is very similar for the different polarizations. Also plotted is the best twoslope piecewise linear model for the aggregate of all polarizations. Here the slope for large distances is about 1.96.



Fig. 5. Attenuation vs. distance, horizontal polarization, 2.4 GHz.

Additional measurements were made at 5 GHz. Figures 9, 10, and 11 show the attenuation for horizontal polarization, vertical polarization and cross polarization at 5 GHz.

The parameters of the model were found based on finding the smallest mean squared error between the model and the measurements. The parameters of the overall model are shown in the Table I. Note that the mean square error of the measurements is also the variance corresponding to a log-normal distribution of the received power. For the above received power versus distance, the inverse power law in the high distance region started at about 12 meters with an exponent of



Fig. 6. Attenuation vs. distance, vertical polarization, 2.4 GHz.



Fig. 7. Attenuation vs. distance, cross polarization, 2.4 GHz.



Fig. 8. Models of attenuation vs. distance, 2.4 GHz.



Fig. 9. Attenuation vs. distance, horizontal polarization, 5 GHz.



Fig. 10. Attenuation vs. distance, vertical polarization, 5 GHz.



Fig. 11. Attenuation vs. distance, cross polarization, 5 GHz.

1.91 and a path loss of about 65 dB at a distance of 12 meters.

 TABLE I

 PARAMETERS FOR CROSS POLARIZATION, 2.4 GHz, T1

Parameter	Value
α_1	0.64
α_2	1.91
β	12.25
k_1	56.52
k_2	42.72
σ^2	12.60

In Figure 12 we compare the received power for these models versus distance. All other models have a simple linear representation of received power in dB versus log distance, we have a piecewise linear model with two different slopes. Overall our model has a power loss exponent close to that of free space at large distances but has a smaller exponent at small distances compared to the other models. The other models mainly were for office spaces as opposed to an industrial setting.



Fig. 12. Various models for overall received power (2.4 GHz) as a function of distance

Shadowing is another factor in determining the performance of a communication system. Shadowing is generally modeled as a log normal random variable. That is, the received power, expressed in dB, is a Gaussian random variable. The mean of the random variable is a function of distance as determined by the path loss model. The variance of the Gaussian random variable measures the effect due to shadowing. Our estimation of the path loss model, by finding the best piecewise linear attenuation model to minimize the mean squared error also results in a mean squared error that is the variance of the Gaussian random variable that models the shadowing. Our results indicate a shadowing variance between 7dB and 14dB. The path loss model used for all polarizations corresponded to a shadowing parameter of about 12dB. This tends to be somewhat larger than other models. So while the average path loss seems to be smaller than other models, the variance tends to be larger.

IV. IMPULSE RESPONSE

The measurement procedure described above allows us to estimate the channel impulse response or the power delay profile of the multipath channel. Consider the output of the reference system after performing a matched filtering and the corresponding output of the measurement system. The output of the reference system where the antennas have been replaced by an attenuator is given by

$$z_r(t) = y_r(t) * h_{MF}(t) = [s_T(t) * h_T(t) * h_R(t) * h_{MF}(t)]A.$$

where $s_T(t)$ is the signal generated by the m-sequence generator, $h_T(t)$ and $h_R(t)$ are the impulse responses for the transmitter and receiver circuitry, $h_{MF}(t)$ is the impulse response of the matched filter and A is the attenuation value. The corresponding output for the measurement system is

$$\begin{aligned} z_m(t) &= y_m(t) * h_{MF}(t) \\ &= [s_T(t) * h_T(t) * h(t) * h_R(t) * h_{MF}(t)] \\ &= [s_T(t) * h_T(t) * h_R(t) * h_{MF}(t)] * h(t) \\ &= \frac{z_r(t) * h(t)}{A}. \end{aligned}$$

Thus the output for the measurement system is the reference response with an additional filtering due to the channel but without the factor due to the attenuator. The output of the matched filter for the reference systems (as seen in Figure 3), $z_r(t)$ is nearly an ideal impulse function but with sidelobes about 35dB lower than the main lobe. As a result, the channel h(t) could be estimated simply as $\hat{h}(t) \approx z_m(t)A$. Here we want to account for the sidelobes of the reference signal to more accurately estimate the impulse response of the channel. Often the multipath aspect of a channel is modeled as a series of impulses of the form

$$h(t) = \sum_{i} \beta_i \delta(t - \tau_i)$$

where β_i is a complex path gain at delay τ_i . For this channel model the result of the measurement would be

$$z_m(t) = \frac{1}{A} z_r(t) * h(t)$$

= $\frac{1}{A} z_r(t) * \sum_i \beta_i \delta(t - \tau_i)$
= $\frac{1}{A} \sum_i \beta_i z_r(t - \tau_i).$

Our goal is to determine the values for β_i and τ_i . Since the function $z_r(t)$ is known, the approximation used above is that h(t) is just a normalized version of $z_m(t)$. However, we can also calculate the β_i by using the known value of $z_r(t)$. In particular we can determine the largest value of β_i by looking at the largest value of the measurement and the associated delay and associating that output value with of β_i . With that determined we can subtract off the effect of the largest β_i , namely $\beta_i z_r (t - \tau_i)$ and continue the process to find the second largest value of β_i and the associated delay. This is generally known as the CLEAN algorithm [8]. In Figure 13 we show in blue the the result of applying the CLEAN algorithm to estimate the impulse response. The error, shown in red, is the left over signal after 250 iterations of the CLEAN algorithm where in each iteration the largest magnitude signal is accounted for by a particular delay and coefficient of the channel. Note that a particular delay could correspond to several iterations that have the largest magnitude residual error signal. In this case it is the (complex) sum of these coefficients that determine the final coefficient at that delay.



Fig. 13. Result of processing measurements with CLEAN algorithm

This approach of estimating the impulse response was applied to 10,500 different acquisitions for a particular run. The magnitude of the impulse responses is shown in Figure 14.



Fig. 14. Impulse response for various acquisitions

With an estimate of the impulse response of the channel various other channel parameters can be calculated. Often the rms delay spread is used to characterize a channel. A large rms delay spread can degrade the performance of certain systems. The rms delay spread can be calculated as follows. Let h(t) be the impulse response of the channel. First we define the power delay profile as

$$P(\tau) = \frac{|h(\tau)|^2}{\int |h(t)|^2 dt}.$$

The power delay profile is a probability density function since it integrates to 1 and is non-negative. The absolute received power level is normalized out in determining the power delay profile. The mean excess delay spread is calculated as

$$\bar{\tau} = \int t P(t) dt.$$

The mean square delay spread is

$$\sigma_{\tau}^2 = \int (t - \bar{\tau})^2 P(t) dt.$$

Then the rms delay spread is σ_{τ} . From the set of impulse responses we can determine the power delay profile for each acquisition and the corresponding rms delay spread. The Gaithersburg machine room indicated rms delay spreads in the range of 90 ns to as much as 400ns. This would indicate a coherence bandwidth of more than 2.5 MHz. Thus, for this delay spread, there should not be significant intersymbol (or interchip) interference in a Zigbee system. However, the measurements in an automotive assembly building indicated delays spreads in the range of 1-5 μ s. This corresponds to a coherence bandwidth as small as 200 kHz and the interchip interference would play a role in determining the performance.

V. ZIGBEE PERFORMANCE

In this section we consider the error probability for a Zigbee communication system. As is known, the 802.15.4 standard specifies how signals are to be transmitted but not how signals are to be received. While there are multiple physical layers defined in the standard, our focus in on signals in the 2.4 GHz band. These signals are designed to transmit data at a maximum rate of 250 kbps but could be smaller. One physical laver defined in the standard is called Offset OPSK. This is a modulation technique that maps groups of four information bits into complex sequences of length 16 chips and then uses offset QPSK with half sine pulse shaping. This is essentially MSK at the chip level. This modulation technique can be demodulated in various ways. A coherent receiver with soft decision demodulation will be the most complex receiver but have the best performance. A noncoherent receiver that does coherent integration over a chip sequence but does not require a coherent phase reference will have worse performance. A receiver that makes a hard decision on each chip using noncoherent demodulation and then finds which of the 16 chips sequences is closest in Hamming distance would have even worse performance.

First we consider a comparison of a purely orthogonal signal set with a perfectly coherent receiver and evaluate the symbol error probability. Note that, in a typical Zigbee application the packet error probability will be of the most interest rather than the bit error probability or the symbol error probability. However, to understand the effects of different modulation techniques and demodulation techniques we evaluate the symbol error probability of a four bits symbol. Figure 15 compares two different modulation techniques and two different receivers. One modulation is an orthogonal signal set. The second modulation is the Zigbee signal set. One receiver is a coherent receiver that requires ideal synchronization and perfect phase estimation. The second receiver is a noncoherent receiver. This noncoherent receiver assumes a constant phase offset for the duration of the time for transmission of the signals (e.g. 16 times the length of a chip). For Zigbee this would be about 16 μ s. As can be seen from the figure, the Zigbee signal set with coherent demodulation requires about 0.6dB more signal-to-noise ratio $(E_b/N_0 \text{ (dB)})$ than an orthogonal signal set at a symbol error probability of 10^{-5} . A noncoherent receiver with an ideal orthogonal signal set at a symbol error rate of 10^{-5} has the same required signal-to-noise ratio (E_h/N_0) as coherent demodulation of the Zigbee signals. However, at higher error rates the coherent receiver for Zigbee signals performs better than the noncoherent receiver for orthogonal signal. The Zigbee signals with noncoherent reception has worse performance at a symbol error rate of 10^{-5} by a little more than 0.6dB than orthogonal signals with noncoherent reception. Note that a symbol error rate of 10^{-5} might correspond to a packet error rate in the range of 10^{-3} with packets on the order of 100 symbols (50 bytes). A receiver making hard decisions on each chip would be expected to be about 2dB worse performance than the receivers shown here.



Fig. 15. Symbol error probability, orthogonal vs. Zigbee, coherent vs. noncoherent reception.

The packet error probability for transmission of information depends first on being able to detect the presence of a transmission and then being able to synchronize to the transmitted signal (timing). After that demodulation of each symbol in a packet is required for the packet to be correct. For Zigbee there is no error control coding technique that could correct symbol errors. The only notion of coding is in the construction of the signal set. In Zigbee a pair of symbols determine a byte of information. A packet in Zigbee can have at most 127 data bytes but could have as few as 9 (ignoring the preamble bytes).

In Figure 16 the packet error probability of an IEEE 802.15.4 system with a coherent receiver and a noncoherent receiver for a packet of length 127 bytes is shown for an additive white Gaussian noise (AWGN) channel. As can be seen the coherent receiver is less than 2dB better than the noncoherent receiver. One reason for such a small gap is that the modulation used in Zigbee is a version of 16-ary orthogonal modulation. As is known, orthogonal modulation has asymptotically (for large number of signals) the same performance for coherent reception and noncoherent reception. Here the signal-to-noise ratio (E_b/N_0) is the average *received* energy per information bit to noise power spectral density.

In indoor and outdoor applications, radio systems need to have a good performance which means a reasonable amount of information loss. As with any other radio system, in order to evaluate Zigbee performance, we started with simulating Zigbee in an additive white Gaussian noise (AWGN) channel as well as Ricean fading channels. In IEEE 802.15.4, at the beginning of PPDU of each packet, there is a 4 bytes-long preamble which consists of 32 zero bits for all packets. We are using these 32 zero bits to find the start of each packet using a matched filter which is matched to each symbol (4 zeros) of the preamble.

The transmitted signal is passed through a complex AWGN channel and the output of the channel is fed into the Zigbee receiver. As the first block of any radiosystem receiver, a synchronization block is designed to find the start of each packet. Since there is a fixed pattern in preamble part of each packet, the receiver uses a matched-filter to locate the separating flag between any two consecutive packets. After finding start of packet, it is possible to pass preamble and demodulate the length of payload-byte of PPDU. Knowing the packet start and the packet length, then the next step is to demodulate the payload which carries the information bits. The demodulation is 16-orthogonal demodulation and is used to detect payload of each packet. The magnitudes only of the 16 demodulator outputs are used to make a decision about the data for the non-coherent receiver. To do coherent demodulation, the real part of outputs of inner products are considered and the maximum is selected. Both coherent and non-coherent receiver have been simulated and their performances in terms of Packet Error Rate (PER) are compared above in Figure 16.

Indoor channel environments are not always well modeled by an AWGN channel. Multipath propagation and obstacle reflections can have a significant impact on system performance. In order to model the multipath propagation, which is a serious factor in indoorcommunication applications, a Rician fading channel has been simulated. Rician fading is a stochastic model for the radio propagation when the signal arrives at the receiver by several different paths. Rician fading can nicely model the environment specially, when one path, which is usually line of sight path, is much stronger than others. This appears to be the case for some of the indoor industrial channel for Zigbee since the bandwidth is relatively small (2 MHz) compared to the bandwidth for WiFi (20 MHz). Our simulation models the amplitude gain using a Rician distribution. Rayleigh fading is used to model the multipath propagation when there is no line of sight. A Rician model with different ratio of direct line-of-sight power versus diffuse power, known as the K factor has been used in our simulation. The packet error rate (PER) for a packet of length 9 and 127 bytes with a coherent receiver is shown in Figure 17 and Figure 18. As expected, the PER converges to AWGN packet error rate curve as the K factor gets large. Notice that there is only a slight performance degradation of the larger packet size relative to the small packet size in these figures, especially for Rayleigh fading. This is due to the fact that the fading is assumed to be a constant for the duration of a packet. If the fade is a bad fade (i.e. destructive interference) then the error probability will be large for symbols and for the packet as a whole. While a good fade (i.e. constructive interference) will result in correct symbols and the packet as a whole. So the packet error probability is dominated by the probability of a good fade versus a bad fade.



Fig. 16. Packet error probability, block length 127 bytes: coherent vs. noncoherent, AWGN

VI. CONCLUSIONS

In this paper we have used measurements to obtain models for indoor industrial environment channels. Our models are piecewise linear relations between the received power (in dB) and the log of the distance. Perhaps the most useful part of the propagation model



Fig. 17. Packet error probability, block length 9 bytes, Rician, Rayleigh channel, coherent reception



Fig. 18. Packet error probability, block length 127 bytes: coherent Rician, Rayleigh, coherent reception

occurs after the breakpoint in the piecewise linear model where the power received becomes small. The received power at short distances is larger than other models while at higher distances the received power is less than most other models. At distances smaller than the break point in the piecewise linear model the received power is going to be quite large and the exact value of the received power is probably not important as the system will have more than adequate power to decode a packet correctly. We have used the CLEAN algorithm to determine the multipath channel characteristics. The multipath delay spread is generally less that 0.5 μ s and is comparable to the inverse bandwidth of a Zigbee system. That is, most of the multipath components will be within a single chip duration of a Zigbee signal. We have used the measurements to evaluate the shadowing

parameter for this environment and our results show a log normal shadowing of between 7 and 12 dB. A Zigbee radio system with different receivers has been simulated and the performance in different channel environments has been determined. While there is a small difference between coherent and noncoherent receivers (e.g. about 2dB), there is a large gap between AWGN performance and Rayleigh faded performance. This is to be expected since the Zigbee signals do not employ error-correcting codes or wide enough bandwidth so that the fading is mitigated.

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