APPENDIX D

PROOF OF \((\partial/\partial \tau)h(\tau, \tilde{k}) < (\partial/\partial \tau)g(\tau)\)

We first prove, for the case when \(\tilde{k} = 1\), that \((\partial/\partial \tau)h(\tau, \tilde{k}) < (\partial/\partial \tau)g(\tau)\) in the lower region of \(\tau\), i.e., \(\mathbb{R}_L\). From (21)–(23), when \(\tilde{k} = 1\), we have \(A = 1\), and \(B\) and \(C\) = 0. Therefore, from (20), we have

\[
\frac{\partial}{\partial \tau} h(\tau, 1) = -\frac{1}{\tau} - \frac{2}{T - \tau} \beta - \frac{\beta}{\sqrt{2\tau T}} \left(1 - \sqrt{Q(\alpha + \beta \sqrt{T})}\right) \exp\left(-\frac{(\alpha + \beta \sqrt{T})^2}{2}\right). \tag{24}
\]

The derivative of the function \(g(\tau)\) is given as

\[
(\partial/\partial \tau)g(\tau) = (\beta/\sqrt{T}) \left(\alpha + \beta \sqrt{T}\right).
\]

Therefore, the detailed expression of \((\partial/\partial \tau)h(\tau, 1) < (\partial/\partial \tau)g(\tau)\) is given as

\[
-\frac{1}{\sqrt{T}} - \frac{2}{\sqrt{T}} \beta \left(1 - \sqrt{Q(\alpha + \beta \sqrt{T})}\right) \exp\left(-\frac{(\alpha + \beta \sqrt{T})^2}{2}\right) < \frac{\beta}{\sqrt{T}} (\alpha + \beta \sqrt{T}). \tag{25}
\]

Let \(x = -\alpha - \beta \sqrt{T}\), and since in the lower region of \(\tau\), i.e., \(\mathbb{R}_L\), we have \(\alpha + \beta \sqrt{T} \leq 0\), and therefore, \(x \geq 0\). Substituting \(x\) into (25), we obtain

\[
-\frac{1}{\beta \sqrt{T}} - \frac{2}{\beta (T - \tau)} \frac{1}{\sqrt{2 \tau Q(x)}} \exp\left(-\frac{x^2}{2}\right) < -x \quad \forall x \geq 0. \tag{26}
\]

From (18), it is shown that \((1/\sqrt{2 \tau Q(x)})\exp(-x^2/2) > x\) for all \(x \geq 0\). Therefore, the inequality at (26) is verified, and we have proven \((\partial/\partial \tau)h(\tau, 1) < (\partial/\partial \tau)g(\tau)\) at the lower region of \(\tau\). Next, we will prove \((\partial/\partial \tau)h(\tau, \tilde{k}) < (\partial/\partial \tau)h(\tau, 1)\) for \(\tilde{k} = 2, \ldots, N\). From (20) and since \(\nabla \tilde{P}_f (\tau) < 0\) for \(0 < \tau \leq T\), we have

\[
2(A + B + C) > \frac{2}{\phi(\tau, \tilde{k})} \quad \forall \tilde{k} = 2, \ldots, N \tag{27}
\]

and after some algebraic manipulations, we have \(1 + ((B + C)/A) > 1\) for all \(\tilde{k} = 2, \ldots, N\). Since \(A\), \(B\), and \(C > 0\) for all \(\tilde{k} = 2, \ldots, N\), the inequality \(1 + ((B + C)/A) > 1\) is verified. This completes the proof that \((\partial/\partial \tau)h(\tau, \tilde{k}) < (\partial/\partial \tau)h(\tau, 1) < (\partial/\partial \tau)g(\tau)\) when \(\tau \in \mathbb{R}_L\) and \(\tilde{k} = 2, \ldots, N\).

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Characterizing Intra-Car Wireless Channels

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Abstract—This paper describes the methodology and results of a series of experiments performed to characterize intra-car wireless channels. Specifically, the experiments target parameters such as the coherence time, statistics of channel loss, and fade statistics. Based on previous experiments, flat fading is assumed; the methodology is developed, and the results are interpreted in this context. These efforts are motivated by the need of designing an intra-car wireless sensor network; therefore, some of the implications of results in practical design are discussed. It is found that although the in-vehicle channels exhibit a very large amount of power loss, robust system design can be achieved by utilizing the results of these experiments.

Index Terms—Channel measurements, wireless communications, wireless sensor networks.

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I. INTRODUCTION

As electronic controls in automobiles have grown more advanced and complicated, the need for more sensors to monitor various quantities inside them has also increased. Nowadays, cars have microprocessors that are used to process the data obtained from all these sensors. As seen in Fig. 1, each sensor is wired to an electronic control unit (ECU), which samples and processes the information from that particular sensor. All ECUs in the vehicle are connected through a backbone network. One of the most widely adopted protocols for in-vehicle data networks is known as a controller area network (CAN). The sensors are, in most cases, passive devices with no intelligence on them, whereas the ECUs have some processing capability. The wiring inside the vehicle [1] can be costly, heavy, and, perhaps most importantly, not flexible for the accommodation of modifications to the architecture. Therefore, there is an increasing level of appeal to design a system in which the wired connections to the sensors are replaced with wireless links—in effect, with a wireless network (see Fig. 1). To that end, the feasibility of different technologies [such as radio frequency identification (RFID), Zigbee, etc.] has been investigated [2]–[6].

To address key issues such as reliability and longevity in such an intra-car wireless sensor network while keeping the complexity and cost of the system as low as possible, efforts have been focused on designing a new physical (PHY) layer. Therefore, it has become necessary to obtain an accurate characterization of the behavior of the wireless channels so that the PHY layer can accordingly be designed. Wireless channels are by nature extremely complex and, in most cases, unpredictable systems. There are a number of models and parameters that are used to characterize these channels, many of which are backed up by intuition and physical theory [12]–[16], but no single concrete method of determining the characteristics of a given wireless channel has been established. Therefore, the most accurate method of characterizing a particular wireless channel is experimental measurements, particularly for practical purposes. To the best of the authors’ knowledge, there is no work in the open literature describing extensive experimentation on wireless channels for narrow-band transmission between antennas inside a single car, nor are there theoretical models or simulations specifically focusing on it. Knowledge of the behavior of such channels is critical in designing, simulating, and analyzing any in-car wireless system; gathering experimental data on these channels is, therefore, essential. This is the main contribution of this paper. More specifically, it is shown that the channels generally have a large power loss and large coherence times, but the statistics of the power loss vary from channel to channel and may or may not fit traditional channel models. The general conclusion is that, while the channels are somewhat hostile, given the knowledge obtained from this type of experiment, systems can be designed that function well in such environments.

The rest of this paper is organized as follows. Section II describes the problem at hand and the approach taken. The only work on similar channels in the past has been by the authors themselves; this earlier work is mentioned, and it is shown how these new experiments build upon previous results. Section III gives an overview of the logistics of the experiment. In Section IV, results are presented, and some practical implications are discussed. Section V concludes the paper.

II. PROBLEM STATEMENT

Wireless channels are typically characterized as linear time-variant systems with time-varying impulse responses \( h(t; \tau) \) [17]. Mathematically put, if \( s(t) \) is the complex baseband equivalent of a signal transmitted over a wireless channel and the receiver receives a signal \( r(t) \)

\[
    r(t) = \int_{-\infty}^{+\infty} s(u) h(u; t - u) du
\]

(1)

---

1 It must be noted that, for various reasons, including the cost and complexity restraints on the nodes in such a system, this research has primarily focused on narrow-band communication schemes. There is also ongoing work aiming at developing an intra-car wireless sensor network using wideband communication techniques; see [8]–[11].
then assuming linearity, this model accounts for both the distortion of signals passing through the wireless channel (due to effects such as multipath propagation, which are also known as time dispersion effects), through the $\tau$ parameter, and the time-varying nature of the channel, through the $t$ parameter. In practice, however, it is very difficult to completely and accurately measure the impulse response $h(t; \tau)$ of a wireless channel, which is why measurements usually focus on one of two aspects: either an average characterization of the time dispersion or a statistical characterization of the time variations [12].

Since neither of those two aspects has accurately been modeled for intra-car channels in the past, both must be addressed to obtain a relatively complete picture of the channels. The measurements described in this paper are the third set of experiments in an ongoing project aiming to achieve this. The first set of experiments performed in the in-car environment were measurements of the power delay profile [2], which were generally accurate enough to yield estimates of the first- and second-order delay spread statistics (namely, mean excess delay and RMS delay spread). A major conclusion is that the coherence bandwidth of the channels is larger than the signal bandwidth in most practical scenarios, as will be discussed shortly.

The second round of experiments dealt with the important statistical characterization of received power when transmitting a practical real-world signal [3]; this is related to the time-varying aspect of the impulse response. It is shown that by making studious choices, one can achieve respectable levels of reliability, even over the most hostile in-car channels. Because standard data transmission signals (specifically, binary phase shift keying (BPSK) signals) were used in [3], the results can directly be applied to practical system design. However, they do not directly measure the impulse response $h(t; \tau)$. In addition, those measurements were performed on a small set of channels and at only one carrier frequency (915 MHz). In the new set of experiments, these issues are addressed.

It is desired to obtain a description of the behavior of $h(t; \tau)$ that is as complete as possible, since such information can directly be applied to any design scenario, any choice of signals, etc. Generally, to achieve this, one must make accurate and extensive measurements of both aspects of the channel. However, from [2], it is known that the coherence bandwidth of the in-car channels is larger than the bandwidth needed for the transmissions (a few megahertz for the low-data-rate spread-spectrum applications in mind). This has two major implications [17]:

1) The time dispersion caused by the channel will be much smaller than the symbol rate; in other words, the channel can be modeled as a random time-varying loss and phase shift with additive white Gaussian noise (AWGN).

2) In frequency, the entire spectrum will similarly be affected, which means that any signal will be affected in a similar fashion as a single-tone sinusoid (this is known as flat fading). As a result, the experiments can be carried out on a pure sinusoid, which is far easier to generate and analyze than any data-modulated signal.

Mathematically, the second implication states that if a single carrier wave $A \cos(2\pi f_c t)$ is transmitted and the receiver collects a signal $A \alpha(t) \cos(2\pi f_c t + \theta(t))$, then $\alpha(t)$ and $\theta(t)$ (both random processes) describe the attenuation and phase shift of any narrow-band signal transmitted through that channel.\footnote{The complex baseband equivalent of this process is $A^{\text{channel}} \rightarrow A \alpha(t) e^{j \theta(t)}$.} In other words

$$h(t; \tau) \approx \alpha(t) e^{j \theta(t)} \delta(\tau).$$

In this experiment, a signal generator creates the pure sinusoid. This signal is then fed to one of the antennas placed in the car and a spectrum analyzer is connected to one of the other antennas to collect the output. The spectrum analyzer collects a narrow-band signal around some center frequency and converts it to the complex baseband equivalent. Since the transmitted signal had constant amplitude and phase, this complex baseband equivalent, after normalization by the transmit power, is equal to $\alpha(t) e^{j \theta(t)}$, which is the channel response. The effect of AWGN is made negligible by choosing the bandwidth of the receiver filter to be very small; this is possible only because the transmitted signal is a pure carrier.

Therefore, by measuring the attenuation and phase shift experienced by a pure sinusoid, the attenuation and phase shift of any (narrow-band) signal transmitted over the channel will be obtained.

In summary, comparing these experiments and those carried out in [3], the following should be noted.

1) Both sets of experiments focus on the statistical characterization of channel loss and fading and not on the time dispersion aspect of the channels.

2) Assuming flat fading, based on the coherence bandwidth results from [2], a single set of experiments can, in fact, directly measure the important parameters of $h(t; \tau)$, while also providing ample data to analyze the fading statistics and other metrics. To be more specific, this is achieved by simultaneously obtaining both amplitude and phase data in these measurements, while previous experiments focused only on amplitude.

3) The method used has the added advantages of being simpler and more resilient to noise, which improves accuracy.

4) Whereas previous experiments were performed at 915 MHz, these experiments were performed (primarily) at 2.4 GHz. Both choices were motivated by practical systems under consideration, specifically RFID at 915 MHz [2], [5] and Zigbee at 2.4 GHz [6], [7].

### III. Experimental Setup

In this experiment, a signal generator creates the pure sinusoid. This signal is then fed to one of the antennas placed in the car and a spectrum analyzer is connected to one of the other antennas to collect the output. The spectrum analyzer collects a narrow-band signal around some center frequency and converts it to the complex baseband equivalent. Since the transmitted signal had constant amplitude and phase, this complex baseband equivalent, after normalization by the transmit power, is equal to $\alpha(t) e^{j \theta(t)}$, which is the channel response. The effect of AWGN is made negligible by choosing the bandwidth of the receiver filter to be very small; this is possible only because the transmitted signal is a pure carrier.

As mentioned, the physical setup of the experiment is a number of antennas placed in various locations in a car. In this case, the car is a 1996 Buick LaSabre provided by General Motors (see Fig. 2). The positions of the antennas are as shown in Fig. 2. The antenna under the roof in the passenger compartment (CEN) is an omnidirectional monopole; the others are directional single-feed Yagi–Uda antennas with a maximum gain of 8 dBi. The choice of directional antennas is mainly out of system design considerations; although, for the most part, there is no line-of-sight (LOS) propagation in this environment.
there is a large amount of power loss related to multipath interference. Directional antennas have the property of reducing multipath interference by transmitting (and receiving) in certain directions more than others; therefore, it would seem reasonable for an end system to make use of this property. Furthermore, experiments with variations in the directions and orientations of the antennas (not included in this paper) had been planned to determine whether that was an area where model-specific optimizations could be performed.

The experiments were carried out in two scenarios:

1) parked in a covered parking area with the engine off and no passengers;
2) driving around the suburbs and outskirts of Pittsburgh, PA.

This is almost a best-case–worst-case scenario experiment: in scenario 1, the car is surrounded by completely motionless walls and a roof with other cars and people rarely passing by, while in scenario 2, the car is moving over often rough roads in relatively open areas, uphill and downhill, thereby constantly changing its speed, with passengers and a running engine that cause the channel response to be far from constant. As such, one would expect the results of similar experiments in other scenarios (urban driving, parked with the engine running in an open field, etc.) to be somewhere in the range of the results presented here. Each link (transmitter–receiver pair) in each scenario is tested in both directions, often at various transmission power levels. Each individual experiment is carried out for more than an hour, providing more than 6 000 000 data samples per measurement.

### IV. Analysis and Results

#### A. General Parameters

Table I contains some numerical results obtained for certain links from the experiments. One can make the following observations based on the data shown in Table I.

1) Power loss is very large. The channels chosen for Table I are among the best and worst in terms of average channel loss; even though the transmitter–receiver distance is somewhere between 1 and 3 m for various channels, the average loss is very high and significantly varies from channel to channel.

2) A comparison of the first and second columns shows that the transmission power does not have a major effect on the statistics of the channel response. The result holds for all the channels that were measured with multiple transmission powers. This is a significant result because it verifies the assumption that the intra-car wireless channel responses are indeed *linear*, which is the first step in comparing these channels to well-known wireless channel models [recall (1)]. It is worth noting that this result also holds when studying the coherence time.

#### B. Variations Over Time

The *coherence time* $T_C$ of a wireless channel is defined as the time over which the autocorrelation function of the channel response remains above some certain fraction of the peak. To compute this metric for the various channels, the complex channel-response data (containing amplitude and phase) collected was autocorrelated, and the 50% coherence time was measured. It was observed that the coherence time is very large (hundreds of seconds or more) for all the channels in the parked scenario. However, in the driving scenario, the coherence times drastically drop (10 to 100 times less). One of the more extreme cases is shown in Fig. 3, which shows the autocorrelation computed for the channel from under the engine (UE) to CEN. The important result, however, is that the smallest coherence time measured (in either scenario) was more than 2 s. This is significant: Not only are these channels all *slow-fading channels* in which the channel response can be assumed constant over one symbol duration, but the response is also essentially constant over an entire frame duration for frames of

*TABLE I

Examples of Statistics Retrieved From Raw Data Samples for Some of the Channels*

<table>
<thead>
<tr>
<th>Scenario</th>
<th>Transmitter</th>
<th>Receiver</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Under Engine</td>
<td>In Engine 1</td>
</tr>
<tr>
<td></td>
<td>Under Engine</td>
<td>In Engine 1</td>
</tr>
<tr>
<td></td>
<td>Under Engine</td>
<td>Trunk</td>
</tr>
</tbody>
</table>

| Standard deviation of envelope (normalized) | 0.0129 | 0.0161 | 0.0283 | 0.236 | 0.487 | 0.132 | 0.425 |

Fig. 3. Autocorrelation function of the UE-to-CEN channel.

3) A comparison of the third column with the preceding two (and with the next two pairs of columns) shows that driving the car has a detrimental effect on the performance of the channel. This is quite expected and intuitive, but what is interesting is that the nature and magnitude of the effect differ from channel to channel; for example, the average power loss can increase, stay roughly the same, or even slightly decrease in the driving scenario, while the variations in the power loss always increase.
components, fit the data much better. This can be interpreted as the statistics of the amplitude coefficient $\alpha(t)$, which is the square root of the power gain (or the inverse of the square root of the channel loss). To this end, some popular distributions (namely, Rayleigh, lognormal, Rician, Nakagami, and Weibull [18]) are fitted to the data points using maximum-likelihood parameter estimation, which is similar to the procedure detailed in [3]. The Rayleigh distribution does not work well for channels in the non-LOS regime [2], whereas the lognormal, Rician, Nakagami, and Weibull distributions still fared better. However, it is interesting to observe that the Rayleigh distribution is a special case of the other three distributions (Rician with $k = 0$, Weibull with $\beta = 2$, and Nakagami with $\mu = 1$), any of those distributions will fit any set of data better than Rayleigh. However, in wireless channel modeling, that special case corresponds to the lack of a single dominant signal component; therefore, the difference between the best Rayleigh fit and the best Rician or Weibull fit is a good visual indicator of whether a dominant component exists and, if so, how pronounced its effect is.

To this end, some popular distributions (namely, Rayleigh, lognormal, Rician, Nakagami, and Weibull [18]) are fitted to the data points using maximum-likelihood parameter estimation, which is similar to the procedure detailed in [3]. The Rayleigh distribution does not work well for channels in the non-LOS regime [2], whereas the lognormal, Rician, Nakagami, and Weibull distributions still fared better. However, it is interesting to observe that the Rayleigh distribution is a special case of the other three distributions (Rician with $k = 0$, Weibull with $\beta = 2$, and Nakagami with $\mu = 1$), any of those distributions will fit any set of data better than Rayleigh. However, in wireless channel modeling, that special case corresponds to the lack of a single dominant signal component; therefore, the difference between the best Rayleigh fit and the best Rician or Weibull fit is a good visual indicator of whether a dominant component exists and, if so, how pronounced its effect is.

C. Channel Loss Statistics

What remains to be characterized about the channel response is the statistics of the amplitude coefficient $\alpha(t)$, which is the square root of the power gain (or the inverse of the square root of the channel loss). To this end, some popular distributions (namely, Rayleigh, lognormal, Rician, Nakagami, and Weibull [18]) are fitted to the data points using maximum-likelihood parameter estimation, which is similar to the procedure detailed in [3].

Fig. 4 shows the cumulative distribution functions (cdf’s) of the resulting distributions for two channels. Based on these distributions, one can make the following observations.

The Rayleigh distribution does not work well for channels in the parked state. This is because the channel loss varies slowly and very little (low standard deviation) and is centered at some value; therefore, the received signal almost never drops to zero. Note the limits on the values of $\alpha(t)$ in Fig. 4(a). The Rician and Weibull distributions, which are typically descriptive of channels with dominant LOS propagation components, fit the data much better. This can be interpreted as showing that the channel has a strong almost-constant propagation component similar to LOS. This result can be counterintuitive, because in the car environment, in almost all cases, there is no LOS or even a single dominant propagation path. Therefore, the behavior observed in Fig. 4(a) is probably the result of multiple signal propagation paths that are almost constant, which add up to create a strong almost-constant signal at the receiver, with other (much smaller) time-variant multipath components causing the (very slight) variations. In other words, in these channels, the effect of a single dominant LOS component is mimicked by multiple time-invariant propagation paths (see Fig. 5). The Rayleigh distribution is used to describe channels where there is no dominant signal component, and therefore, it will fail in these cases. Furthermore, in most of the parked scenarios, even the best fits among the distributions were not exceptionally tight; in fact, they all showed variances higher than that of the data they were approximating. This demonstrates that the dominant component in the parked scenario is strong enough to cause deviation from the regular fading models. A similar phenomenon appeared when calculating the coherence time (see Section IV-B), where the coherence time on many of the parked channels was extremely large (virtually constant channel). As a result, one can observe that the parked channels do not behave at all differently from conventional LOS AWGN channels, albeit with high power loss.

For the driving scenarios [such as the case in Fig. 4(b)], the Rayleigh distribution fared much better. However, it is interesting to observe that the Rician, Nakagami, and Weibull distributions still fared better
than Rayleigh for most of the channels. This shows that, interestingly enough, the almost-constant dominant signal component (which, as explained, is probably the aggregate result of multiple propagation paths) exists, even in the driving scenario; however, its effect is much less pronounced. This effect is also visible in the coherence times of the driving scenarios, which are quite large but not nearly as large as that in the parked scenario. Many of the driving scenario channels are comparable with the class of static indoor channels, as they follow similar statistical models (Rician, Weibull, and Nakagami) and have relatively large coherence times (or, equivalently, very small Doppler shifts and spreads) [13]. It is important to understand, however, that there is no direct correspondence to any specific type of indoor channel; rather, analysis and design ideas based on similar indoor models may be applicable to the intra-car channels (particularly in the driving scenario) as well.

These results are qualitatively in line with the results obtained in [3]. This fact, in turn

1) supports the assumption that the effect of the channels (in terms of power loss statistics, coherence time, etc.) on a single-tone carrier is similar to their effect on any narrow-band signal (such as the BPSK signals used in [3]);

2) supports the validity of the results of both sets of experiments, as they were performed using different methods but produced similar results;

3) demonstrates that, at least in terms of loss statistics, the channels do not substantially differently behave at 915 MHz and 2.4 GHz; however, the differences (in terms of average value, variance, and other statistics) are significant enough to warrant separate experiments at each frequency.

D. Fading

Another class of important metrics in the characterization of wireless channels is the fading statistics. In this context, fading is described as the power of the received signal falling below (or, equivalently, the channel loss rising above) some tolerable level. Two important fading metrics are [12] the fade duration and the level-crossing rate (LCR). LCR is defined as the number of times the received power crosses the tolerable level in the positive direction per unit of time. Fade duration is the amount of time the channel remains in a fade (i.e., the received power remains below the level). This, of course, varies from fade to fade; therefore, it is usually averaged over a large number of fades to produce the average fade duration (AFD). From these metrics, the fade proportion \( p \), which is the proportion of time the channel is in a fade or, equivalently, the probability that the channel will be in a fade at some arbitrary point in time, can be approximated as

\[
p = \text{AFD} \times \text{LCR}. \tag{3}
\]

The fading statistics depend on the level that is chosen as tolerable. This, in turn, depends on the level of performance (e.g., the packet error rate) expected from the physical link and the transmission power. Therefore, it is practically useful to store the raw data and use it to compute the fading characteristics for any specific given scenario. Since all these issues and more about fading were discussed in more detail in [3], in this paper, for the sake of brevity, only an example of the application of the data to practical design will be presented:

Suppose a system is designed to use BPSK modulation with a data rate of 250 kbps. The receiver noise figure is 30 dB, and the thermal noise power spectral density (PSD) is the usual assumed value of \(-174 \text{ dBm/Hz}\); therefore, the noise PSD at the receiver will be \( N_0 = -144 \text{ dBm/Hz} \). Suppose the maximum frame length is 1000 Bytes, and there is an error-correction mechanism capable of correcting up to 8 bit errors. This means that if the bit error rate (BER) \( \leq (8/8 \times 1000) = 10^{-3} \), the average frame will correctly be received. To maintain a BER of at least \( 10^{-3} \), based on the theoretical BPSK curve \( \text{BER} = Q(\sqrt{2E_b/N_0}) \)

\[
P_T \geq 4.75 \text{ pW} \approx -83 \text{ dBm}.
\]

Therefore, for this example, any received power level below \(-83 \text{ dBm} \) will be considered to be a fade. Now, the fading statistics can be analyzed for different values of transmission power.

The channel chosen for evaluation is the UE-to-CEN channel in the driving scenario (see the last column in Table I). The average loss of this channel is \( 80.9 \text{ dB} \). Therefore, if the transmission power is set at 0.0 dBm, on the average, the received power will be \(-80.9 \text{ dBm} \), which is \( 2.1 \text{ dB} \) higher than the minimum necessary. However, it can be extracted from the data that, for this received power level, the LCR is 657 times/s, which means that the received power will extremely frequently drop below \(-83 \text{ dBm} \). The AFD in this case is 0.6 ms.

The situation can be improved by increasing the transmission power. Transmitting at 3.0 dBm would result in \( 5.1-\text{dB} \) excess average power at the receiver, but the LCR would drop to 437 times/s, and the AFD would slightly fall to 0.4 ms. A transmitted power of 10.0 dBm would waste 12.1 dB on the average at the receiver, but it would also reduce the LCR to 145 times/s and the AFD to 0.2 ms.

At this point, it is important to once again contrast the behavior of the channels in the parked and driving scenarios. The received signal amplitude in the parked channels only varies over a small range (a few decibels). However, it extremely slowly varies; therefore, if the minimum tolerable level is somewhere within that range, it is likely that the channel will go into a fade and not emerge for a long time (recall the very large coherence times)—which is an unacceptable situation in almost all applications. Therefore, when designing the system, one must either choose a constant transmission power level high enough to avoid this situation altogether or employ an adaptive power control scheme to deal with the slow variations. In the driving scenario, simply setting a higher transmission power is not as effective, as variations are much higher, and fades will occur anyway. However, a fade is not as much of a problem because it will probably be shorter than that in the parked scenario; therefore, even a simple retransmission scheme might be sufficient for reliable data transfer. In any case, the designer must be aware of these different fading characteristics and incorporate them into the system design.

V. Conclusion

This paper has described new results from an ongoing investigation into the characteristics of intra-car wireless channels. Whereas previous measurements have been made focusing on the time dispersion parameters and statistics of the channel loss, the work described herein has attempted to build upon those results with a new set of experiments that targets a more fundamental characterization. The motivation is to develop knowledge of the behavior of these channels that will facilitate the development of an in-car wireless sensor network that will optimally function under these conditions.

It has been observed that although the channels display very high amounts of power loss (considering the distances in question), they are not at all unusable. For example, they tend not to quickly vary (i.e., the coherence times are large). Furthermore, since an extensive set of measurements of the behavior is in hand, the difficulties can be taken into consideration to make intelligent design choices on issues such as modulation schemes, transmission power, adaptive power control, automatic-repeat-request (ARQ) schemes, etc. It is anticipated that these findings will be instrumental in designing a reliable and robust intra-car wireless sensor network.
Pre-DFT Combining for Coded OFDM

S. Nagaraj

Abstract—Multiple receive antennas have been known to improve bit error performance over fading multipath channels by providing spatial diversity. In systems such as orthogonal frequency-division multiplexing (OFDM), this benefit is obtained at the cost of greatly increased system complexity. Pre-discrete Fourier transform (DFT) combining techniques have been proposed by some researchers to obtain multiple antenna gains at a reduced complexity. However, existing works on pre-DFT combining ignore the specific distance properties of the underlying convolutional code, leading to suboptimal solutions. In this paper, we develop a novel technique for pre-DFT combiner design for bit-interleaved coded modulation (BICM)-coded OFDM on Rayleigh block-fading channels. The combiner is designed for the specific code used and outperforms other known pre-DFT combining techniques. Our simulations show about a 2–3-dB improvement in performance over existing techniques.

Index Terms—Bit-interleaved coded modulation (BICM), maximum ratio combining (MRC), orthogonal frequency-division multiplexing (OFDM), single-input multiple-output (SIMO).

I. INTRODUCTION

Orthogonal frequency-division multiplexing (OFDM) is being considered by many researchers as the most suitable modulation format for next-generation wireless communication [1]. OFDM converts a frequency-selective fading channel into several parallel flat-fading subchannels [2], removing the need for complex equalizers. However, the diversity that a frequency-selective channel inherently offers is lost. System performance is determined by the performance of highly faded subcarriers. Bit-interleaved coded modulation (BICM) [3], [4] is a well-known coding technique for achieving diversity in OFDM systems. BICM is used in many systems today (e.g., IEEE 802.11g) to provide coding and diversity gains at high spectral efficiencies.

The performance of coded OFDM on fading channels can further be improved by employing multiple antennas at the transmitter or the receiver. With multiple antennas at the transmitter, space-time block codes [5] or space-time trellis codes [6], [7] may be concatenated to achieve both space and frequency diversity. With multiple antennas at the receiver, the signal at each antenna may be combined using maximum ratio combining (MRC) [8], [9] techniques to achieve both space and frequency diversity. It has been shown that, when space diversity is combined with time or frequency diversity, the diversity order of the system is given by the product of the orders of space and time/frequency diversity.

Traditional MRC has high receiver complexity since one discrete Fourier transform (DFT) operation is necessary for each receive antenna. In [10], which was later extended in [11], the principle of orthogonal design was used to reduce the number of DFTs in half with a loss of only about 3 dB. Okada and Komaki [12] proposed pre-DFT combining so that only one DFT would be sufficient at the receiver.

REFERENCES