Intervehicular Communication Systems under Co-Channel Interference and Outdated Channel Estimates

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***Abstract*—In this paper, we study the influence of interference in a inter-vehicular communication system. The interfering links are modeled using the multiple scattering radio channel, which fits quite well with experimental data for mobile-to-mobile communications. For this scenario, we derive the signal-to-interference ratio statistics of a single input single output system. Moreover, the analytical results are generalized by taking into account a diversity reception scenario with two well-known schemes, namely selection diversity and maximal ratio combining. In addition, we analytically study the realistic scenario of outdated (imperfect) channel estimates at the receiver side. The outage probability (OP) is considered as the performance metric for all scenarios considered. Based on the numerical evaluation of the derived analytical expressions for the OP, we depict the impact of interference, diversity schemes and outdated estimates on the system's quality of service. Simulation results are also included to validate the analytical results.**

***Index Terms*- Co-channel interference, multiple scattering, outdated channel estimates, diversity, vehicle-to-vehicle communications.**

I. Introduction

The inter-vehicular communication (IVC) systems that comprise vehicle-to-vehicle (V2V) and vehicle-to-infrastructure (V2I) communications are considered to be crucial components of the intelligent transportation systems (ITS). The ITS integrate telecommunications, electronics and information technologies with transport engineering in order to plan, design, operate and manage transport systems. The characteristics of the V2V communications differ from those of traditional mobile cellular communications. More specifically, in V2V communications the transmitter (Tx) and the receiver (Rx) are in the same height and in similar propagation environments (i.e., peer-to-peer), scattering can occur around both the Tx and the Rx, while the distance over which these communications can take place is comparatively short ($<$ 1km). Furthermore, V2V channels are characterized by high mobility, since both the Tx and Rx as well as many of the important scatterers, are continuously moving. In these systems, diversity techniques may improve the overall performance, e.g., see [1]–[4]. A common observation in the previous works is the assumption of noise limited environment. However, in many practical situations, e.g., due to hidden terminal effect, the performance of these systems can be significantly affected by co-channel interference (CCI).

In this paper, we study the influence of the interfering effects in an IVC scenario. In particular, based on the multiple

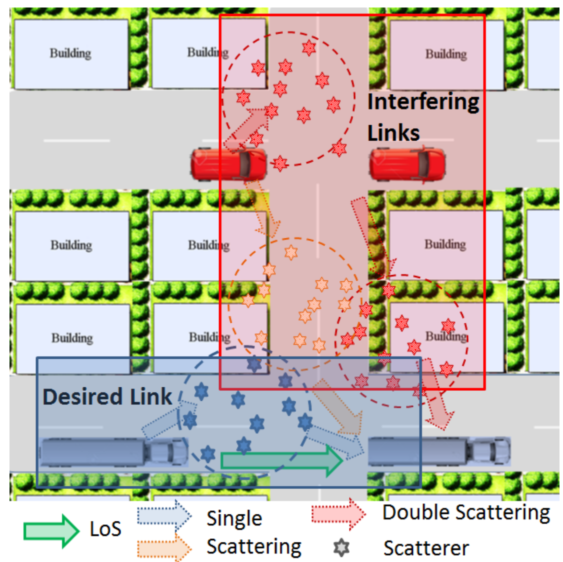

Fig. 1. Example of a LOS in SOS conditions.

scattering channel model, a well established distribution for modeling mobile-to-mobile communication conditions, e.g., [5], the koint probability density function (PDF) of multiple interfering signals has been derived for the first time and used to analyze the performance of the scenario under consideration. Specifically, the quality-of-service (QoS) has been evaluated using the signal-to-interference ratio (SIR) statistics. Moreover, the performance improvement induced by the adoption of diversity techniques has been also analytically investigated. Finally, an analytical framework for quantifying the negative consequences of imperfect channel estimates (that are available) at the receiver has been also developed.

The rest of the paper is organized as follows. Section II contains the system and channel model under investigation. In Section III, we focus on the interference influence to a single antenna communication scenario. In Section IV the performance improvement induced by using diversity reception is studied. In Section IV, the impact of the outdated channel estimates is studied, while in Section V the concluding remarks are provided.

II. System and Channel Model

In general, we consider a communication system with 1 transmitting and L receiving antennas, operating in a vehicular communication environment. In our study, line-of-sight (LoS) conditions exist between the Tx and the Rx, as it is shown in Fig. 1. Moreover, Rx is also subject to interfering signals coming from various mobile sources, which however do not have LoS components, as it is also shown in Fig. 1. Moreover, we also assume that (in general) the level of interference at the receiver is such that the effect of thermal noise on system performance can be ignored (interference limited scenario).

Let us denote the instantaneous signal-to-noise ratio (SNR) of the desired received signal as $\gamma_{d_j} = |h_{d_j}|^2 E_s/N_0$, with h_{d_j} denoting the complex channel gain received at the jth branch, E_s is the average transmitted signal energy and N_0 the additive white Gaussian noise (AWGN) power spectral density. It is assumed that $|h_{d_j}|$ follows the Rice distribution. Thus the PDF of γ_{d_j}, under independent and identically distributed (i.i.d.) fading conditions, is given by [6, eq. (2.16)]

$$\begin{aligned} f_{\gamma_{d_j}}(\gamma) = & \frac{(1+K)\exp(-K)}{\overline{\gamma}_d} \exp\left[-\frac{(1+K)\gamma}{\overline{\gamma}_d}\right] \\ & \times I_0\left[2\sqrt{\frac{K(K+1)\gamma}{\overline{\gamma}_d}}\right] \end{aligned} \tag{1}$$

where $\overline{\gamma}_d = \mathbb{E}\left\langle |h_{d_j}|^2\right\rangle E_s/N_0$ is the average input SNR per branch, with $\mathbb{E} < \cdot >$ denoting expectation, K corresponds to the ratio of the power of the LoS component to the average power of the scattered component and $I_v(\cdot)$ denotes the modified Bessel function of the first kind and order v [7, eq. (8.445)]. Moreover, the corresponding expression for the cumulative distribution function (CDF) of γ_{d_j} is given by

$$F_{\gamma_{d_j}}(\gamma) = 1 - Q_1\left(\sqrt{2K}, \sqrt{\frac{2(1+K)\gamma}{\overline{\gamma}_d}}\right) \tag{2}$$

where $Q_1(\cdot,\cdot)$ denotes the first order Marcum Q-function [6, eq. (4.33)]. Furthermore, let us denote the instantaneous received interference-to-noise ratio (INR) of the ith interfering received signal, with $i \in [1, \ldots, M]$, as $\gamma_{I_i} = |h_i|^2 E_i/N_0 = |h_i|^2 \rho_s$, with corresponding average INR equal to $\overline{\gamma}_{I_i} = \mathbb{E}\left\langle |h_i|^2\right\rangle \rho_s$, where $|h_i|$ denotes the channel gain of the ith interfering signal with energy E_i.

In this paper, we have adopted the multiple-scattering radio channel for modeling the envelopes of the interfering signals. In this context, we focus our attention to the important case of second order scattering (SOS), which has been found to provide a good explanation for the signal envelope for V2V communication conditions. The SOS is characterized by the following impulse response [8]

$$C_{2,i}^* = w_0 e^{j\theta} + w_1 H_{1,i} + w_2 H_{2,i} H_{3,i}. \tag{3}$$

In (3), $C_0 = w_0 e^{j\theta}$ is the LoS component with constant magnitude and uniformly distributed phase over $[0, 2\pi)$, w_0, w_1 and w_2 are non negative real-valued constants that determine the mixture weights of the LoS, single as well as double scattering components. More specifically, by varying w_is, different propagation conditions can be modeled, while (3) includes well-known fading distributions as special cases. For example, assuming $w_2 = 0$, it coincides to Rice distribution, for $w_0 = w_1 = 0$, it coincides to double-Rayleigh distribution and for $w_0 = w_2 = 0$, it coincides to Rayleigh distribution. Here, for the interfering signals, we focus on a scenario of practical interest where only a combination of single and double scattering components exists, i.e., $w_0 = 0$. A potential communication scenario satisfying the assumptions made in this work is given in Fig. 1. In that case the magnitude of second order scattering, $|h_i| = |C_2^*|$, is given by [8, eq. (30)]

$$f_{|h_i|}(r) = 2\exp\left(\frac{w_1^2}{w_2^2}\right)\sum_{m=0}^{\infty}\frac{(-1)^m}{m!}\frac{\Gamma\left(-m, w_1^2/w_2^2\right)}{\left(w_2^2\right)^{m+1}} r^{2m+1} \tag{4}$$

where $\Gamma(\cdot,\cdot)$ is the upper incomplete gamma function [7, eq. (8.350/2)]. Based on the definition of γ_{I_i}, it has been proved that the PDF of γ_{I_i} is given by [1, eq. (38)].

In the system under consideration, the instantaneous output SIR is given by [9]

$$\gamma_{\text{out}} = \frac{\gamma_X}{\gamma_I} \tag{5}$$

where $X \in \{s, \text{mrc}, \text{sd}\}$ when referring to the instantaneous SNR of the single, maximal ratio combiner (MRC) and selection diversity (SD) receivers, respectively, while γ_I denotes the total INR, i.e., $\gamma_I = \sum_{i=1}^{M} \gamma_{I_i}$.

Theorem 1: Let γ_I denote a RV defined as

$$\gamma_I \triangleq \sum_{i=1}^{M} \gamma_{I_i} \tag{6}$$

where the PDF of γ_{I_i} is given in [1, eq. (38)]. The PDF of γ_I can be expressed as

$$f_{\gamma_I}(\gamma) = \frac{1}{2}\exp\left(\frac{w_1^2}{w_2^2}M\right)\sum_{n=0}^{\infty} c_n \frac{\gamma^{M+n-1}}{\Gamma(M+n)} \tag{7}$$

where

$$\begin{aligned} a_0 &= \frac{1}{w_2^2}\Gamma\left(0, \frac{w_1^2}{w_2^2}\frac{1}{\rho_s}\right), \; c_0 = a_0^M \\ a_m &= \frac{(-1)^m}{m! w_2^{2(m+1)}}\Gamma\left(-m, \frac{w_1^2}{w_2^2}\right)\frac{m!}{\rho_s^{m+1}} \\ c_m &= \frac{1}{m a_0}\left(\sum_{t=1}^{m}(tM - m + t) a_t c_{m-t}\right) \end{aligned}$$

with $\Gamma(\cdot)$ being the Gamma function [7, eq. (8.310/1)].

Proof: The moments generating function (MGF) of γ_I is defined as $M_{\gamma_I}(s) = \left[M_{\gamma_{I_i}}(s)\right]^M$. Substituting, the MGF of $M_{\gamma_{I_i}}(s)$, given in [1, eq. (38)], in this definition, making some mathematical manipulations and applying [7, eq. (0.314)], yields the following exact expression for the MGF of γ_I

$$M_{\gamma_I}(s) = \frac{1}{2}\exp\left(\frac{w_1^2}{w_2^2}M\right)\sum_{n=0}^{\infty} c_n \frac{1}{s^{M+n}}. \tag{8}$$

Then, (8) is in appropriate form to apply the inverse Laplace transform, resulting to (7), which also completes the proof. ■

It is noted that the PDF in (7) converges fast, since in most cases a relatively small number of terms is sufficient, i.e., $<$

20, to achieve a high accuracy. Based on (5), the CDF of $\gamma_{\rm out}$ is given by [9]

$$F_{\gamma_{\rm out}}(\gamma) = \int_0^\infty F_{\gamma_X}(\gamma x) f_{\gamma_I}(x) dx. \quad (9)$$

Next, we will examine different system and channel model communication scenarios. More specifically, we will analyze i) the influence of multiple interfering signals to the system performance of a single antenna system, ii) the performance improvement induced by employing diversity in an interference limited scenario and iii) the performance deterioration induced due to outdated channel estimations. It is obvious that in order to better understand the impact of all these contradictory parameters to the system performance, the best approach is to examine them separately.

III. Interference Investigation

In this section, we consider a communication scenario where the Tx communicates with the Rx (with $L = 1$) via a LoS. Moreover, Rx is also subject to interfering effects coming from M sources. In this context, substituting (2) and (7) in (9), where we have assumed $X \equiv s$, using [10, eq. (10)] and after some mathematics, yields the following expression for the CDF of $\gamma_{\rm out}$

$$F_{\gamma_{\rm out}}(\gamma) = 1 - \exp\left(\frac{w_1^2}{w_2^2}M\right) \sum_{n=0}^{\infty} c_n \frac{\overline{\gamma}_d^{M+n}}{[(1+K)\gamma]^{M+n}} \times {}_1F_1(-M-n, 1, -K) \quad (10)$$

where ${}_1F_1(\cdot,\cdot)$ denotes the confluent hypergeometric function [7, eq. (9.210/1)]. Using (10), the outage probability (OP) can be studied. The OP is an important performance indicator for identifying the system's QoS and is defined as the probability that the SIR falls below a predetermined outage threshold $\gamma_{\rm th}$ and is given by $P_{\rm out} = F_{\gamma_{\rm out}}(\gamma_{\rm th})$. In Fig. 2, the OP of a single antenna communication scenario is plotted as a function of the number of interfering sources M. To obtain this figure, we have assumed strong and weak LoS conditions (for the desired signal), that is $K = 11.4$ and $K = 2$, respectively. Moreover, two scenarios regarding the interfering signals propagation conditions have been studied. In the first one, we have assumed $w_1 = 0.2$ (with $w_2 = \sqrt{1-w_1^2}$), which results to dominant double-scattering components for the interfering signals, while in the second one $w_1 = 0.8$, which results to dominant single-scattering components. In Fig. 2, it is shown that the performance improves as K increases, i.e., strong LoS conditions exist for the desired signal. Interesting observations that come out of this figure is that for weak LoS conditions ($K = 2$), the OP for both scenarios of scattering are very close. However, when strong LoS conditions exist, the best performance is when the single scattering components of the interfering signals are dominant. Thus, the single scattering propagation for the interfering signals results in lower INR and thus higher SIR.

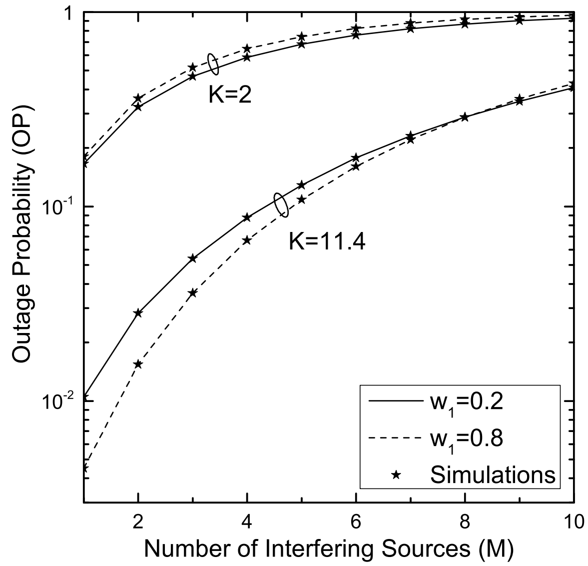

Fig. 2. OP of single antenna scenario vs M.

IV. Multichannel Communications Investigation

In this section, we consider the impact of diversity techniques on the system performance. Two well-known schemes will be analytically investigated, namely MRC and the SD.

A. Maximal Ratio Combining

We consider a communication scenario where the Tx communicates with a Rx (with L antenna branches) supporting MRC reception via a LoS. Moreover, Rx is also subject to interfering signals coming from 1 source. We have assumed that the same interfering signals are present on each diversity branch [11]. In this context, the CDF of the MRC is given by [12, eq. (20)]

$$F_{\gamma_{\rm mrc}}(\gamma) = 1 - Q_L\left(\sqrt{2KL}, \sqrt{\frac{2(1+K)}{\overline{\gamma}_d}\gamma}\right) \quad (11)$$

where $Q_m(\cdot,\cdot)$ denotes the generalized Marcum Q-function [6, eq. (4.59)]. Thus, substituting (11) and (7) in (9), where we have assumed $X \equiv {\rm mrc}$, using again [10, eq. (10)] and after some mathematics, yields the following expression for the CDF of $\gamma_{\rm out}$

$$F_{\gamma_{\rm out}}(\gamma) = 1 - \exp\left(\frac{w_1^2}{w_2^2}\right) \sum_{n=0}^{\infty} \frac{(-1)^n}{(n+1)!} \Gamma\left(-n, \frac{w_1^2}{w_2^2}\right) \times \frac{\Gamma(L+n+1)}{\Gamma(L)} \left(\mathcal{A}_n\right)^{n+1} {}_1F_1(-n-1, L, -KL) \quad (12)$$

where $\mathcal{A}_n = \overline{\gamma}_d / \left(\rho_s w_2^2 (1+K)\gamma\right)$.

B. Selection Diversity

We consider a communication scenario where the Tx communicates with the Rx supporting SD (with $L = 2$ antenna branches) via LoS paths. We have also assumed that the same

$$
\begin{aligned}
F_{\gamma_{\mathrm{out}}}(\gamma) &= 1-\sum_{n=0}^{\infty}\mathcal{B}_n\left(\mathcal{A}_n\right)^{n+1}\Gamma(n+2)\,{}_1F_1(-n-1,1,-K)+\sum_{n=0}^{\infty}\mathcal{B}_n\left(\mathcal{A}_n\right)^{n+1}\exp(-K)\Bigg\{\frac{\Gamma(n+2)}{K}M_{-n-\frac{3}{2},0}(K)M_{-\frac{1}{2},0}(K)\\
&-\left[\sum_{\ell=1}^{n+2}\frac{(n+1)!K^{\ell-1}}{2(\ell-1)!}\binom{n+1}{n+2-\ell}\left[\exp(-K)\sum_{j=1}^{\ell-1}\frac{\sqrt{\pi}\csc\left(\frac{\pi(2j+3)}{4}\right)}{(2K)^{j/2}}\mathcal{G}_{4,5}^{1,3}\left[2K\Big|{}^{0,\frac{1}{2},-\frac{j}{2},\frac{1}{4}}_{\frac{j}{2},-\frac{j}{2},-\frac{j}{2},\frac{j}{2},\frac{1}{4}}\right]\right]\right]-\exp(-K)\sum_{\ell=1}^{n+1}\sum_{j=0}^{n-\ell+1}\frac{(n-j)!}{(\ell-1)!}\\
&\times\binom{n+1}{n+1-\ell-j}\frac{(2K)^{\frac{\ell-1}{2}}\sqrt{\pi}}{2^{\ell+j+1}}\csc\left[\frac{\pi(2(\ell-1)+3)}{4}\right]\mathcal{G}_{4,5}^{1,3}\left[2K\Big|{}^{0,\frac{1}{2},-j-\frac{\ell+1}{2},\frac{1}{4}}_{\frac{\ell-1}{2},\frac{1-\ell}{2},\frac{1-\ell}{2},\frac{\ell-1}{2},\frac{1}{4}}\right]\Bigg\}.
\end{aligned}
\quad (14)
$$

interfering signals are present on each diversity branch. In this context, the CDF of the SD is given by

$$
F_{\gamma_{\mathrm{sd}}}(\gamma)=\left[1-Q_1\left(\sqrt{2K},\sqrt{\frac{2(1+K)}{\overline{\gamma}_d}\gamma}\right)\right]^2. \quad (13)
$$

In the Appendix, it has been proved that the CDF of the output SIR for the scheme under consideration is given by (14) (shown at the top of the next page). In (14), $\mathcal{B}_n=(-1)^n\exp\left(\frac{w_1^2}{w_2^2}\right)\frac{2\Gamma\left(-n,w_1^2/w_2^2\right)}{(n+1)!}$, $\mathcal{G}_{p,q}^{m,n}[\cdot|\cdot]$ denotes the Meijer's G-function [7, eq. (9.301)] and $M_{\mu,\nu}(\cdot)$ is the Whittaker function [7, eq. (9.220/2)]. Thus, by using (12) and (14), the OP of the diversity schemes under consideration can be directly evaluated.

In Fig. 3, the OPs of MRC and SD receivers are plotted as a function of the average input SNR. To obtain this figure we have assumed $w_1=0.7$, which results to intermediate propagation conditions regarding the interfering signals, while different values of ρ_s have been assumed. In this figure, it is shown that the QoS clearly improves, since the OP decreases, when diversity reception is employed, with MRC having always the best performance, as compared to single channel reception. Interesting observations that come out of this figure is that for lower values of ρ_s, the OP improves. This is a reasonable result since an increase on the fading severity in the interfering signals result to a lower INR and thus to a higher SIR. Finally, for the MRC case another plot with $L=3$ branches is also included. In this plot, it is also verified that using MRC, important diversity gain is achieved, despite the negative consequences of the interfering signals.

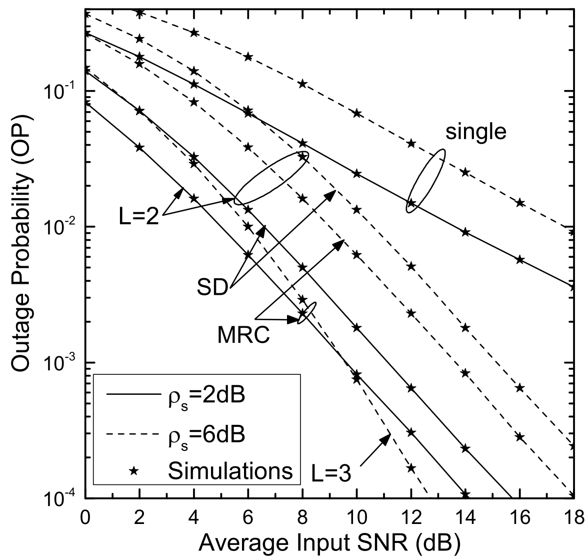

Fig. 3. OP of different diversity schemes vs the average input SNR.

V. Outdated Channel Estimates Investigation

We consider a communication scenario where the Tx communicates with the Rx supporting SD (with $L=2$) via a LoS in a interference free environment. Under ideal conditions, the SD receiver tracks the channel estimate, $\hat{h}_{d_j}$ from the two diversity branches, and selects the branch with the highest SNR value. However, in many practical communication scenarios, the channel estimate $\hat{h}_{d_j}$ and the actual channel gain h_{d_j} are not identical, but a relation between them exists [13]. For example, in time varying communication scenarios, where the Doppler spread may become large, while, at the same time, the wireless medium is fast time varying, the branch which was the best at the selection time instant t may not be the best at the (reception) time instant $t+\tau$ [9], [14]. Mathematically speaking, the joint PDF between h_{d_j} and $\hat{h}_{d_j}$ is given by [15]

$$
\begin{aligned}
f_{h_{d_j},\hat{h}_{d_j}}(x,y) &= \frac{2(1+K^2)xy}{\pi\overline{\gamma}_d^2(1-\rho^2)}\exp\left[-\left(x^2+y^2\right)\beta_1\right]\\
&\times\exp\left(-\frac{2K}{1+\rho}\right)\int_0^{2\pi}\exp\left(2\rho\beta_1 xy\cos(\theta)\right)\\
&\times I_0\left[\sqrt{\frac{4K(x^2+y^2+2xy\cos(\theta))}{\overline{\gamma}_d(1+K)^{-1}(1+\rho)^2}}\right]d\theta
\end{aligned}
\quad (15)
$$

where ρ is the correlation coefficient between h_{d_j} and $\hat{h}_{d_j}$ and $\beta_1=\frac{(1+K)}{(1-\rho^2)\overline{\gamma}_d}$. It is noted that in a time varying scenario ρ is directly related to the Doppler frequency and time delay [16]. Since a PDF in the form of (15) is very difficult, if not impossible, to be used for further analytical purposes, an alternative approach that employs an infinite series representation of this PDF will be adopted. In particular, based on [17], an infinite series expression for the joint PDF of $\gamma_{d_j},\hat{\gamma}_{d_j}$, with $\hat{\gamma}_{d_j}=|\hat{h}_{d_j}|^2E_s/N_0$, is given by

$$
\begin{aligned}
f_{\gamma_{d_j},\hat{\gamma}_{d_j}}(x_1,x_2) &= \sum_{\substack{i,h=0\\ v_1+v_2+v_3=i}}^{\infty}\mathcal{A}\,2^{-i-2}\exp\left[-\beta_1\left(x_1+x_2\right)\right]\\
&\times\left(\mathcal{B}\,x_1^{\beta_2-1}x_2^{\beta_3-1}+\mathcal{C}\,\overline{\gamma}_d^{-1}\,x_1^{\beta_2-1/2}\,x_2^{\beta_3-1/2}\right)
\end{aligned}
\quad (16)
$$

with

$$\mathcal{A} = \frac{2^{v_3+2h-1}(1+K)^{1+\beta_4}\rho^{2h}}{\sqrt{\pi}\,\overline{\gamma}_d^{1+\beta_4}\;(1-\rho^2)^{1+2h}\;v_1!\,v_2!\,v_3!\,i!}$$
$$\left(\frac{K}{(1+\rho)^2}\right)^i \exp\left(-\frac{2K}{1+\rho}\right),$$

$$\mathcal{B} = \frac{\left[1+(-1)^{v_3}\right]\,\Gamma\left[h+(1+v_3)/2\right]}{\Gamma\left[h+1+v_3/2\right]\,\Gamma\left(1+2h\right)},$$

$$\mathcal{C} = \frac{\left[-1+(-1)^{v_3}\right]\,2\rho(1+K)\,\Gamma\left(1+h+v_3/2\right)}{(\rho^2-1)\,\Gamma(2+2h)\Gamma\left[h+(3+v_3)/2\right]},$$

with $\beta_2 = v_1 + \frac{v_3}{2} + h + 1$, $\beta_3 = v_2 + \frac{v_3}{2} + h + 1$ and $\beta_4 = i + 2h + 1$. Thus, the PDF of the actual received instantaneous SNR of the *selected* branch can be expressed as [18]

$$f_{\gamma_{\rm sd}}(\gamma) = \int_0^\infty \frac{f_{\gamma_{d_j},\hat{\gamma}_{d_j}}(\gamma, x)}{f_{\gamma_{d_j}}(x)} f_{\hat{\gamma}_{\rm sd}}(x)dx \tag{17}$$

where the PDF of $\hat{\gamma}_{\rm sd}$ is given by $f_{\hat{\gamma}_{\rm sd}}(\gamma) = 2f_{\hat{\gamma}_{d_j}}(\gamma)F_{\hat{\gamma}_{\rm d_j}}(\gamma)$, with $f_{\hat{\gamma}_{d_j}}(\gamma), F_{\hat{\gamma}_{\rm d_j}}(\gamma)$ given in (1), (2), respectively. Substituting (2) in (17), using the infinite series representation of the Marcum Q-function [6, eq. (4.74)] as well as [7, eq. (3.326/2)], after some mathematics yields the following expression for the PDF of $\gamma_{\rm sd}$

$$\begin{aligned} f_{\gamma_{\rm sd}}(\gamma) &= 2f_{\gamma_{d_j}}(\gamma) - 2 \sum_{\substack{i,h=0 \\ v_1+v_2+v_3=i}}^{\infty} \mathcal{A}\exp\left(-K\right)\gamma^{b_3-1/2} \\ &\times \exp\left(-b_1\gamma\right)\left[\frac{\mathcal{B}}{\gamma^{\frac{1}{2}}}\sum_{n=0}^{\infty}\frac{K^n}{n!}\sum_{k=0}^{n}\left(\frac{K+1}{\overline{\gamma}_d}\right)^k \frac{\Gamma\left(b_2+k\right)/k!}{\left[(1+K)b_5\right]^{b_2+k}}\right. \\ &\left.+\frac{\mathcal{C}}{\overline{\gamma}_d}\sum_{n=0}^{\infty}\frac{K^n}{n!}\sum_{k=0}^{n}\frac{1}{k!}\left(\frac{K+1}{\overline{\gamma}_d}\right)^k\frac{\Gamma\left(b_2+k+1/2\right)}{\left[(1+K)b_5\right]^{b_2+k+\frac{1}{2}}}\right] \end{aligned} \tag{18}$$

where $b_5 = \frac{1}{\overline{\gamma}_d} + \frac{1}{(1-\rho^2)\overline{\gamma}_d}$. The corresponding CDF expression can be derived, using [7, eq. (3.351/1)], as follows

$$\begin{aligned} F_{\gamma_{\rm sd}}(\gamma) &= 2F_{\gamma_{d_j}}(\gamma) - 2 \sum_{\substack{i,h=0 \\ v_1+v_2+v_3=i}}^{\infty} \mathcal{A}\frac{\left(\overline{\gamma}_d(1-\rho^2)\right)^{b_3}}{(1+K)^{b_4+1}} \\ &\times \frac{\exp(-K)}{b_5^{b_2}}\left[\mathcal{B}\sum_{n=0}^{\infty}\frac{K^n}{n!}\sum_{k=0}^{n}\frac{1}{k!}\frac{\Gamma(k+b_2)}{(b_5\overline{\gamma}_d)^k}\gamma(b_3, b_1\gamma) + \mathcal{C}\right. \\ &\left.\times\frac{(1-\rho^2)^{\frac{1}{2}}}{(K+1)}\sum_{n=0}^{\infty}\frac{K^n}{n!}\sum_{k=0}^{n}\frac{\Gamma(k+b_2+1/2)}{k!\,(b_5\overline{\gamma}_d)^{k+\frac{1}{2}}}\gamma\left(b_3+\frac{1}{2}, b_1\gamma\right)\right] \end{aligned} \tag{19}$$

where $\gamma(\cdot,\cdot)$ is the lower incomplete gamma function [7, eq. (8.350/1)]. Based on (19), the OP can be directly evaluated.

In Fig. 4, assuming $K = 4.5$, $\overline{\gamma}_d = 4$dB, the OP is plotted as a function of the correlation coefficient ρ for different values of the outage threshold $\gamma_{\rm th}$. It is depicted that for lower values of ρ, i.e., in a time varying scenario an increased feedback delay exists, the OP is quite high. Moreover, as ρ increases, i.e., in a time varying scenario feedback delay diminishes, the performance improves, which results to an improve to the QoS. The performance also improves, in cases where $\gamma_{\rm th}$ decreases. Moreover, it is interesting to note that for lower values of $\gamma_{\rm th}$ the performance gain due to the increase of the correlation coefficient increases. Monte carlo simulations are also included in all figures to verify the validity of the proposed theoretical approach.

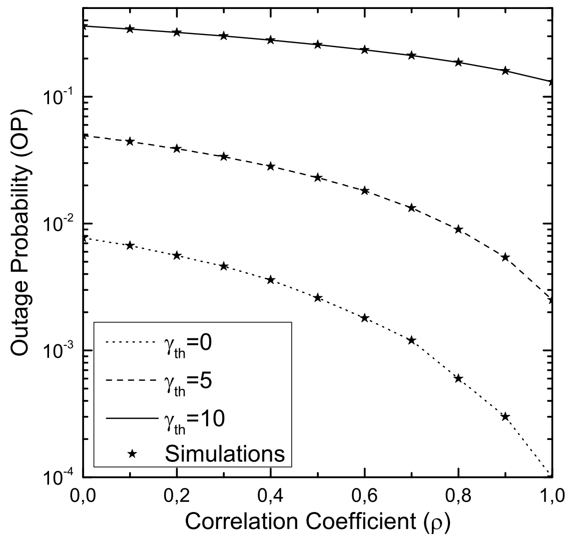

Fig. 4. OP of SD under imperfect channel estimation vs ρ.

VI. Conclusions

In this paper, the influence of the interfering effects on the QoS of a IVC scenario has been analytically evaluated. The multiple scattering distribution has been employed to model the channel gains of the interfering signals. This model has been widely adopted to describe the V2V propagation conditions, since it has a strong physical justification, while it fits very well to empirical data as it is shown in many experimental studies. In this communication environment, the performance improvement achieved using diversity techniques has been also studied. For all cases the CDF expressions of the output SIR have been provided and used to study the OP. In addition, in an interference free scenario the consequences of outdated channel estimates have been also investigated. It is shown that interfering effects degrade seriously the system performance, diversity reception can improve this poor situation, provided that channel estimates are close to the exact ones.

Appendix
Proof for Equation (14)

Substituting (13) and (7) in (9) the following integral appears

$$\begin{aligned} \mathcal{I} &= \int_0^\infty x^n Q_1\left(\sqrt{2K}, \sqrt{\frac{2(1+K)\gamma x}{\overline{\gamma}_d}}\right) \\ &\times Q_1\left(\sqrt{2K}, \sqrt{\frac{2(1+K)\gamma x}{\overline{\gamma}_d}}\right)dx \\ &\overset{(1)}{=} \mathcal{I}_1 + \mathcal{I}_2 \end{aligned} \tag{A-1}$$

where

$$\mathcal{I}_i = \frac{\exp(-2K)}{2^{n+1}(n+1)}\left[\frac{\overline{\gamma}_d}{\gamma(1+K)}\right]^{n+1} \times \int_0^\infty t_{3-i}\exp\left(-\frac{t_{3-i}^2}{2}\right) I_0\left(\sqrt{2K}t_{3-i}\right) \times \underbrace{\int_0^{t_{3-i}} t_i^{2n+3}\exp\left(-\frac{t_i^2}{2}\right) I_0\left(\sqrt{2K}t_i\right) dt_i}_{\mathcal{I}_2}\, dt_{3-i}.$$

In (A-1), (1) holds due to [6, eq. (4.33)]. Moreover, using [7, eq. (6.643/2)], the definition of the Nuttal Q-function, [6, eq. (4.104)], and the corresponding Marcum-Q representation, [7, eq. (4.110)], the following closed-form expression can be derived for $\mathcal{I}_2$

$$\mathcal{I}_2 = \exp(-K)(n+1)!2^{n+1}\left\{\frac{\exp(K/2)}{\sqrt{K}}M_{-n-\frac{3}{2},0}(K) - \sum_{\ell=1}^{n+2}\frac{\binom{n+1}{n+2-\ell}}{(\ell-1)!}K^{\ell-1}Q_\ell\left(\sqrt{2K},t_{3-i}\right)\exp(K) - \exp\left(-\frac{t_{3-i}^2}{2}\right)\sum_{\ell=1}^{n+1}\sum_{j=0}^{n-\ell+1}\frac{(n-j)!}{(\ell-1)!}\frac{\binom{n+1}{n-\ell+1-j}}{(n+1)!2^{j+\frac{\ell+1}{2}}} \times t_{3-i}^{2j+\ell+1}K^{\frac{\ell-1}{2}}I_{\ell-1}(\sqrt{2K}t_{3-i})\right\}. \tag{A-2}$$

Substituting (A-2) in (A-1), the following integrals appear

$$\begin{aligned}
\mathcal{I}_3 &= \int_0^\infty t_{3-i}\exp\left(-\frac{t_{3-i}^2}{2}\right) I_0\left(\sqrt{2K}t_{3-i}\right) dt_{3-i} \\
&\overset{(1)}{=} \frac{\exp(K/2)}{\sqrt{K}}M_{-1/2,0}(K) \\
\mathcal{I}_4 &= \int_0^\infty t_{3-i}\exp\left(-\frac{t_{3-i}^2+2K}{2}\right) I_0\left(\sqrt{2K}t_{3-i}\right) \times Q_\ell\left(\sqrt{2K},t_{3-i}\right) dt_{3-i} \\
&\overset{(2)}{=} \exp(-2K)\sum_{n=1}^{\ell-1}\frac{\sqrt{\pi}}{(2K)^{n/2}}\frac{\csc\left(\pi\left(2n+3\right)/4\right)}{2} \times \mathcal{G}_{4,5}^{1,3}\left(2K\middle|{}_{\frac{n}{2},-\frac{n}{2},-\frac{n}{2},\frac{n}{2},\frac{1}{4}}^{0,\frac{1}{2},-\frac{n}{2},\frac{1}{4}}\right) \\
\mathcal{I}_5 &= \int_0^\infty t_{3-i}^{2j+\ell+2}\exp\left(-t_{3-i}^2\right) I_0\left(\sqrt{2K}t_{3-i}\right) \times I_{\ell-1}\left(\sqrt{2K}t_{3-i}\right) dt_{3-i} \\
&\overset{(3)}{=} \frac{\sqrt{\pi}}{2}\csc\left[\frac{\pi(2(\ell-1)+3)}{4}\right]\mathcal{G}_{4,5}^{1,3}\left[2K\middle|{}_{\frac{\ell-1}{2},\frac{1-\ell}{2},\frac{1-\ell}{2},\frac{\ell-1}{2},\frac{1}{4}}^{0,\frac{1}{2},-j-\frac{\ell+1}{2},\frac{1}{4}}\right].
\end{aligned} \tag{A-3}$$

In (A-3), (1) holds due to [7, eq. (6.614/3)], while for deriving (2), [6, eq. (4.81)] is used (for expressing the generalized Marcum Q-function to Marcum Q-function of the first order), together with [19, eq. (03.02.26.0018.01)] as well as [20, eq. (46)]. Finally, for obtaining (3) [21, eqs. (11 and 21)] have been used. Based on the previous derived analytical expressions and after some mathematics, (14) is finally derived.

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