Effects of RF Impairments in Communications over Cascaded Fading Channels

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***Abstract*—Direct-conversion architectures can offer highly integrated low-cost hardware solutions to communication transceivers. However, it has been demonstrated that radio frequency (RF) impairments such as amplifier nonlinearities, phase noise and in-phase/quadrature-phase imbalances (IQI) can lead to a severe degradation in the performance of such systems. Motivated by this, the present work is devoted to the quantification and evaluation of the effects of RF IQI on wireless communications in the context of cascaded fading channels for both single-carrier and multi-carrier systems. To this end, closed form expressions are firstly derived for the outage probability (OP) over N*Nakagami$-m$ channels for the cases of ideal transmitter (TX) and receiver (RX), ideal TX and IQI RX, IQI TX and ideal RX, and joint TX/RX IQI. The offered expressions along with several deduced corresponding special cases are subsequently employed in the context of vehicular-to-vehicular (V2V) communications to justify their importance and practical usefulness in the context of emerging communication systems. We demonstrate that considering non-ideal RF front-ends at the TX and/or RX, introduces non-negligible errors in the OP performance that can exceed 20% in several communication scenarios. We further demonstrate that the effects by cascaded multipath fading conditions are particularly detrimental, as they typically result in considerable performance losses of around or over an order of magnitude.**

***Index Terms*—Cascaded fading channels, hardware–constrained communications, I/Q imbalance, multi-carrier communications, N*Nakagami$-m$ fading channels, outage probability, single-carrier communications, vehicle-to-vehicle communications.**

This work was supported by the Academy of Finland under the projects 284694 and 288670.

A limited subset of initial results was preliminary presented at the IEEE 26^{th} Annual International Symposium on Personal, Indoor, and Mobile Radio Communications (IEEE PIMRC ‘15), Hong Kong.

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I. Introduction

The ever-increasing demand for high data rate applications and multimedia services has led to the development of flexible and software-configurable transceivers that are capable of supporting the desired quality of service requirements. In this context, the direct-conversion architecture of such systems provides an attractive front-end solution, as it requires neither external intermediate frequency filters nor image rejection filters [1], [2]. Instead, the essential image rejection is achieved through signal processing methods [3]. Direct-conversion architectures are low cost and can be easily integrated on chip, which render them excellent candidates for modern wireless technologies [4]–[6]. However, direct-conversion transceivers are typically sensitive to front-end related impairments, which are often inevitable due to components mismatch and manufacturing defects [7], [8] . An indicative example is the in-phase and quadrature (I/Q) imbalance (IQI) which corresponds to the amplitude and phase mismatch between the I and Q branches of a transceiver and ultimately leads to imperfect image rejection that incurs considerable performance degradation [3], [9], [10].

It is also widely known that due to the different nature of fading conditions, several statistical models have been proposed for characterizing and modeling fading envelopes under short-term, long-term, and composite fading channels. For example, the Nakagami-m and lognormal distributions have been proposed for accounting for short-term fading, also known as multipath fading, and long-term fading, also known as shadowing, respectively. Likewise, several composite fading models have been proposed for accounting for the simultaneous occurrence of multipath fading and shadowing effects (see [11]–[21] and references therein). In the same context, multiplicative cascaded fading models have been more recently introduced in [22]–[24]. The physical interpretation of these models is justified by considering received signals generated by the product of a large number of rays reflected via N statistically independent scatterers [22]. Based on this, the N*Nakagami$-m$ distribution was introduced in [23] corresponding to the product of N statistically independent, but not necessarily identically distributed Nakagami$-m$ random variables. This model is generic as it includes several special cases of more elementary cascaded fading models. For example, for the specific case of $N = 2$ and $m = 1$, it reduces to the double Rayleigh distribution, which has been shown useful in modeling fading effects in mobile-to-mobile communications [25].

A. Related Work

In spite of the paramount importance of radio frequency (RF) front-ends on the performance of wireless communication systems, the detrimental effects of RF impairments has been overlooked in the vast majority of reported analyses. The effect of RF impairments, modeled as independent and identically distributed (i.i.d) additive Gaussian noise, was investigated in [26]–[32]. In [33], the impact of IQI on the performance of orthogonal frequency division multiplexing (OFDM) systems was investigated. In particular, a signal-to-interference-plus-noise-ratio (SINR) expression was evaluated, considering TX-only IQI, RX-only IQI, and joint TX/RX IQI, with equal levels of IQI at the TX and RX ends. Likewise, the effects of IQI on multi-carrier receivers was analyzed in [34], [35]; specifically, the authors in [34] analyzed the impact of the IQI on the bit error rate (BER) performance of an OFDM based system with $M-$ary quadrature amplitude modulation ($M-$QAM), whereas the case of received OFDM signals subject to an I/Q imbalanced transceiver was derived in [35] along with a respective compensation algorithm.

In the context of cooperative communications, the performance of amplify-and-forward dual hop relay systems under IQI was thoroughly investigated in [36]–[42]. In more details, novel analytical expressions for the symbol error probability (SEP) over Rayleigh fading channels were derived in [36]–[38]. In the former, IQI was assumed only at the destination node, while upper and lower bounds for the respective SEP were reported in [37], [38] for the case of joint TX/RX IQI. Likewise, the authors in [39] derived analytic expression for the outage probability (OP), considering independent and non-identical distribution Nakagami$-m$ fading channels, joint TX/RX IQI at the relay nodes and ideal RF front-ends at the source and destination. Moreover, the case of dual hop opportunistic OFDM in the presence of IQI in all nodes was addressed in [40], where the OP was derived considering statistically independent, frequency-selective channels in all wireless links. Moreover, the effects of IQI in the context of two way amplify-and-forward relaying and multiple-input multiple-output (MIMO) systems were investigated in [41], [42] and [43]–[46], respectively. Specifically, in [41] and [42], the effects of IQI in two-way amplify-and-forward relaying systems for the case of independent, non-identically distributed Nakagami-m fading channels, were studied. Furthermore, the authors in [43] investigated the impact of IQI in single-carrier MIMO systems and proposed a baseband compensation method, whereas the authors in [44] analyzed the performance degradation of both TX and RX IQI in a space division multiplexing based MIMO OFDM system, where the symbol error rate (SER) performance results were derived for the case of Rayleigh multipath fading channels. Finally, in [45], the effects of IQI in maximum ratio transmission beamforming OFDM systems were studied, while, in [46], the authors proposed a low-complexity IQI compensation method for MIMO OFDM systems.

B. Contribution

To the best of the authors' knowledge, the detrimental effects of IQI in digital communications over cascaded fading channels have not been addressed in the open technical literature. Motivated by this, the present work is devoted to the quantification and analysis of IQI on the performance of wireless transmission over cascaded Nakagami-m fading channels. The technical contribution of this paper is outlined below:

- Novel analytic expressions are derived for the OP in single-carrier systems over N*Nakagami$-m$ fading channels for the following three scenarios: i) IQI at the TX only; ii) IQI at the RX only; iii) joint IQI at the TX/RX.
- The above OP analysis is extended to a multi-carrier scenario by additionally taking into account the impact of the presence or the absence of a signal at the mirror frequency channel.
- The offered analytic results along with useful deduced special cases are readily applied in the context of vehicle-to-vehicle (V2V) communication scenarios, which provides explicit justification of their overall importance and practical usefulness.
- A simple and tight lower bound is additionally proposed for the case of RX IQI in the multi-carrier scenario. To this effect, a lower bound for the OP is observed in the corresponding cases of RX IQI and joint TX/RX IQI.

C. Organization & Notations

The remainder of the paper is organized as follows: Section II presents the single-carrier and multi-carrier system models for all possible configurations of ideal/impaired TX/RX. Section III is devoted to the derivation of the novel analytic expressions for the corresponding OP metrics. Section IV demonstrates an application of the offered results in V2V communications over N*Nakagami-m channels. Respective numerical results and discussions are provided in section V, while closing remarks are provided in Section VI.

Notations: Unless otherwise stated, $(\cdot)^*$ denotes conjugation, whereas $\Re\{x\}$ and $\Im\{x\}$ represent the real and imaginary part of x, respectively. Furthermore, the $\mathbb{E}[\cdot]$ and $|\cdot|$ operators denote statistical expectation and absolute value operations, respectively.

II. System and Signal Model

In this section, we revisit the ideal signal model, which is, henceforth, referred to as ideal RF, as well as the realistic IQI signal models in both single-carrier and multi-carrier direct-conversion transmitter (TX) and receiver (RX) scenarios for the case that TX and RX are equipped with a single antenna.

A. Ideal RF front-end

We assume a transmitted signal, s, transmitted over a flat wireless channel, h, with an additive white Guassian noise (AWGN), n. The received RF signal is passed through various processing stages, also known as the RF front-end of the RX. These stages include filtering, amplification, analog I/Q demodulation, down-conversion to baseband and sampling. To this effect, the corresponding baseband equivalent received signal can be expressed as

$$r_{\text{ideal}} = hs + n \tag{1}$$

where h is the channel coefficient and n denotes the circularly symmetric complex additive white Gaussian noise (AWGN). It is assumed that the transmitted signal experiences cascaded fading conditions modeled by a N*Nakagami-m process, which is composed of $N \geq 1$ independent, but not necessarily identical, Nakagami-m random variables. Based on this, the instantaneous signal to noise ratio (SNR) per symbol at the RX input can be given by

$$\gamma_{\text{ideal}} = \frac{E_s}{N_0} |h|^2 \tag{2}$$

where, E_s, denotes the energy per transmitted symbol and N_0 is the single-sided AWGN power spectral density (PSD). Therefore, the corresponding average SNR is

$$\overline{\gamma} = \frac{E_s}{N_0} \prod_{i=1}^{N} \Omega_i, \tag{3}$$

with Ω_i denoting the scaling parameter of the i^{th} Nakagami-m process [23].

In the case of multi-carrier systems, the corresponding baseband equivalent received signal at the k^{th} carrier is represented as follows:

$$r_{\text{id}}(k) = h(k)\, s(k) + n(k) \tag{4}$$

where $s(k)$ is the transmitted signal at the k^{th} carrier, whereas $h(k)$ and $n(k)$ denote the corresponding channel coefficient and the circular symmetric complex AWGN, respectively. Hence, the corresponding instantaneous and average SNRs can be represented as follows:

$$\gamma_{\text{id}}(k) = \frac{E_s}{N_0} |h(k)|^2 \tag{5}$$

and

$$\overline{\gamma}_{\text{id}}(k) = \frac{E_s}{N_0} \prod_{i=1}^{N} \Omega_i(k) \tag{6}$$

respectively.

B. I/Q imbalance Model

The time-domain baseband representation of the IQI impaired signal is given by [5]

$$g_{\text{IQI}} = K_1^{t/r} g_{\text{id}} + K_2^{t/r} g_{\text{id}}^* \tag{7}$$

where g_{id} denotes the baseband IQI$-$free signal and g_{id}^* arises due to the involved IQI effects. Furthermore, the IQI coefficients $K_1^{t/r}$ and $K_2^{t/r}$ are expressed as

$$K_1^{t/r} = \frac{1}{2}\left(1 + \epsilon^{t/r} e^{\pm j\phi^{t/r}}\right) \tag{8}$$

and

$$K_2^{t/r} = \frac{1}{2}\left(1 - \epsilon^{t/r} e^{\mp j\phi^{t/r}}\right) \tag{9}$$

where the positive and negative signs in (8) and the t/r superscripts denote the up and down conversion processes, respectively, whereas the $\epsilon^{t/r}$ and $\phi^{t/r}$ terms account for the TX/RX amplitude and phase mismatch, respectively. It is also noted that the IQI parameters are algebraically linked to each other as

$$K_2^{t/r} = 1 - \left(K_1^{t/r}\right)^*. \tag{10}$$

The $K_1^{t/r}$ and $K_2^{t/r}$ coefficients are associated with the corresponding image rejection ratio (IRR) which determines the amount of attenuation of the image frequency band and is expressed as

$$IRR_{t/r} = \frac{\left|K_1^{t/r}\right|^2}{\left|K_2^{t/r}\right|^2}. \tag{11}$$

It is recalled here that for practical analog RF front-end electronics, the value of IRR is typically in the range of 20dB$-$40dB [4], [31], [47]–[51]. Furthermore, the second term $K_2^{t/r} g_{\text{id}}^*$ is caused by the associated imbalances and in the case of single-carrier transmission it represents the self-interference effect, whereas in multi-carrier transmission it denotes the image *aliasing* effect, which results to crosstalk between the mirror-frequencies in the down-converted signal[1].

C. Single-carrier systems impaired by IQI

In this subsection, we present the signal model for single-carrier transmission in which the TX and/or the RX suffers from IQI.

1) TX impaired by IQI: In this scenario, it is assumed that TX experiences IQI while the RF front-end of the RX is ideal. To this effect, it follows from (7) that the baseband equivalent transmitted signal is expressed as

$$s_{\text{IQI}} = K_1^t s + K_2^t s^* \tag{12}$$

whereas the baseband equivalent received signal is given by

$$r_{\text{IQI}}^t = h s_{\text{IQI}} + n \tag{13}$$
$$= K_1^t h s + K_2^t h s^* + n. \tag{14}$$

Furthermore, the instantaneous SINR per symbol at the input of the RX is expressed as

$$\gamma = \frac{|K_1^t|^2 |h|^2 E_s}{|K_2^t|^2 |h|^2 E_s + N_0} \tag{15}$$

which after basic algebraic manipulations can be re-written as follows:

$$\gamma = \frac{1}{\frac{1}{IRR_t} + \frac{1}{|K_1^t|^2}\frac{1}{\gamma_{\text{ideal}}}}. \tag{16}$$

[1]This is because, in general, complex conjugate in time domain corresponds to complex conjugate and mirroring in the frequency domain. Therefore, the spectrum of the imbalance signal at the k^{th} carrier becomes $G_{IQI}^{t/r}(k) = K_1^{t/r} G(k) + K_2^{t/r} G^*(-k)$, where $G(k)$ and $G(-k)$ denote the spectrum of the IQI free signal at the k and $-k$ carriers, respectively.

In the context of direct-conversion transmitter, the IQI effect can be considered as the so-called self-image problem, which is the case when the baseband equivalent transmitted signal is essentially interfered by its own complex conjugate [7].

2) RX impaired by IQI: In this scenario, it is assumed that the RX experiences IQI while the TX RF front-end is ideal. Based on (7), the baseband equivalent received signal is given by

$$r = K_1^r h s + K_2^r h^* s^* + K_1^r n + K_2^r n^*. \tag{17}$$

The corresponding instantaneous SINR per symbol at the input of the RX is expressed as

$$\gamma = \frac{|K_1^r|^2 |h|^2 E_s}{|K_2^r|^2 |h|^2 E_s + \left(|K_1^r|^2 + |K_2^r|^2\right) N_0} \tag{18}$$

which after basic algebraic manipulations can equivalently be expressed as

$$\gamma = \frac{1}{\frac{1}{IRR_r} + \left(1 + \frac{1}{IRR_r}\right) \frac{1}{\gamma_{\text{ideal}}}}. \tag{19}$$

3) Joint TX/RX impaired by IQI: In this scenario, it is assumed that both TX and RX experience IQI. To this effect and based on (7), it follows that the baseband equivalent received signal can be expressed as

$$r = (\xi_{11} h + \xi_{22} h^*) s + (\xi_{12} h + \xi_{21} h^*) s^* + K_1^r n + K_2^r n^* \tag{20}$$

where

$$\xi_{11} = K_1^r K_1^t, \tag{21}$$

$$\xi_{22} = K_2^r \left(K_2^t\right)^*, \tag{22}$$

$$\xi_{12} = K_1^r K_2^t \tag{23}$$

and

$$\xi_{21} = K_2^r \left(K_1^t\right)^*. \tag{24}$$

Based on this, the instantaneous SINR per symbol at the input of the RX is given by

$$\gamma = \frac{|Z|^2 E_s}{|W|^2 E_s + \left(|K_1^r|^2 + |K_2^r|^2\right) N_0} \tag{25}$$

with

$$|Z|^2 = |\xi_{11} h + \xi_{22} h^*|^2 \tag{26}$$

$$|W|^2 = |\xi_{12} h + \xi_{21} h^*|^2 \tag{27}$$

$$= |\xi_{12}|^2 |h|^2 + |\xi_{21}|^2 |h|^2 + 2\Re\left\{\xi_{12}\xi_{21} h^2\right\} \tag{28}$$

whereas

$$\frac{|\xi_{22}|^2}{|\xi_{11}|^2} = \frac{1}{IRR_r IRR_t} \tag{29}$$

which practically lies in the range of $[-43, -28\text{dB}]$. As a result, it can be accurately assumed that

$$|Z|^2 \approx |\xi_{11}|^2 |h|^2 \tag{30}$$

while due to the inequality

$$2\Re\left\{\xi_{12}\xi_{21} h^2\right\} << |\xi_{12}|^2 |h|^2 + |\xi_{21}|^2 |h|^2 \tag{31}$$

equation (28) can be accurately represented as

$$|W|^2 \approx |\xi_{12}|^2 |h|^2 + |\xi_{21}|^2 |h|^2. \tag{32}$$

To this effect, it follows that (25) can be also re-written as

$$\gamma \approx \frac{|\xi_{11}|^2}{|\xi_{12}|^2 + |\xi_{21}|^2 + \left(|K_1^r|^2 + |K_2^r|^2\right) \frac{1}{\gamma_{\text{ideal}}}}. \tag{33}$$

It is noted here that (33) is particularly accurate since the relative error does not exceed 1%.

D. Multi-carrier systems impaired by IQI

In the case of multi-carrier transmission, we assume that multiple RF carriers are down-converted to the baseband by means of wideband direct-conversion, where the RF spectrum is translated to the baseband in a single down-conversion [4]. For notational convenience, we denote the set of channels as $\mathbf{S}_K = \{-K, \cdots, -1, 1, \cdots, K\}$ and the baseband equivalent IQI$-$free transmitted signal at the k^{th} carrier as $s(k)$. In addition, the $\theta \in \{0, 1\}$ parameter indicates the existence of a signal at channel $-k$.

1) TX impaired by IQI: In this scenario, it is assumed that the RF front-end of the RX is ideal, while the TX experiences IQI. Therefore, with the aid of (7) it follows that the baseband equivalent transmitted signal in the k^{th} carrier is expressed as

$$s_{\text{IQI}}(k) = K_1^t s(k) + \theta K_2^t s^*(-k). \tag{34}$$

Notably, equation (34) exhibits that IQI causes the transmitted baseband equivalent signal at the carrier k, $s(k)$, to be distorted by its image signal at carrier $-k$, $s^*(-k)$. In other words, in a multi-carrier system, the effects of IQI result in crosstalk between the mirror frequencies in the down-converted signal [48]. To this effect, the corresponding baseband received signal becomes

$$r(k) = h(k) s_{\text{IQI}}(k) + n(k) \tag{35}$$

which with the aid of (34) can be equivalently re-written as follows

$$r(k) = K_1^t h(k) s(k) + K_2^t h(k) s^*(-k) + n(k). \tag{36}$$

Moreover, the corresponding instantaneous SINR per symbol at the input of the RX is given by

$$\gamma(k) = \frac{|K_1^t|^2 |h(k)|^2 E_s}{\theta |K_2^t|^2 |h(k)|^2 E_s + N_0} \tag{37}$$

which after carrying out basic algebraic manipulations can be expressed as

$$\gamma(k) = \frac{1}{\theta \frac{1}{IRR_t} + \frac{1}{|K_1^t|^2} \frac{1}{\gamma_{\text{id}}(k)}}. \tag{38}$$

2) RX impaired by IQI: In this scenario, it is assumed that the RF front-end of the TX is ideal, while the RX experiences IQI. Hence, by recalling once more (7), the baseband equivalent received signal in the k^{th} carrier can be represented as follows

$$r(k) = K_1^r h(k) s(k) + \theta K_2^r h^*(-k) s^*(-k) + K_1^r n(k) + K_2^r n^*(-k). \quad (39)$$

With the aid of the above expression, it is shown that IQI is the reason that the received baseband equivalent signal at the k^{th} carrier, $s(k)$, is interfered by the image signal at the carrier $-k$, $s^*(-k)$. The instantaneous SINR per symbol at the input of the RX is expressed as

$$\gamma(k) = \frac{|K_1^r|^2 |h(k)|^2 E_s}{|K_2^r|^2 |h(-k)|^2 E_s + \left(|K_1^r|^2 + |K_2^r|^2\right) N_0} \quad (40)$$

$$= \frac{\gamma_{\text{id}}(k)}{\theta \frac{\gamma_{\text{id}}(-k)}{IRR_r} + \left(1 + \frac{1}{IRR_r}\right)} \quad (41)$$

where $\gamma_{\text{id}}(k)$ is given by (5), and

$$\gamma_{\text{id}}(-k) = \frac{|h(-k)|^2 E_s}{N_0}. \quad (42)$$

3) Joint TX/RX impaired by IQI: Finally, it is assumed that both the TX and RX experience IQI. Thus, with the aid of (7) and after some basic algebraic manipulations, the baseband equivalent received signal can be expressed as follows

$$r(k) = (\xi_{11} h(k) + \xi_{22} h^*(-k)) s(k) + (\xi_{12} h(k) + \xi_{21} h^*(-k)) s^*(-k) + K_1^r n(k) + K_2^r n^*(-k) \quad (43)$$

which indicates that IQI renders the received baseband equivalent signal at the k^{th} carrier, $s(k)$, subject to interference by its image signal at the carrier $-k$, $s^*(-k)$. Therefore, the corresponding instantaneous SINR per symbol at the input of the RX is given by

$$\gamma(k) = \frac{|Z_{MC}(k)|^2 E_s}{|W_{MC}(k)|^2 E_s + \left(|K_1^r|^2 + |K_2^r|^2\right) N_0} \quad (44)$$

where

$$|Z_{MC}(k)|^2 = |\xi_{11} h(k) + \xi_{22} h^*(-k)|^2 \quad (45)$$

and

$$|W_{MC}(k)|^2 = |\xi_{12} h(k) + \xi_{21} h^*(-k)|^2. \quad (46)$$

Likewise, eq. (45) can be accurately expressed as

$$|Z_{MC}(k)|^2 \approx |\xi_{11}|^2 |h(k)|^2. \quad (47)$$

In addition, it is noted that the correlation between the channel responses at the k^{th} carrier and its image is small due to their large spectral separation. To this effect, it is realistic to assume them statistically independent, which satisfies the following expression

$$\mathbb{E}\left[\Re\left\{\xi_{12}\xi_{21}^* h(k) h^*(-k)\right\}\right] \approx 0. \quad (48)$$

Based on this, $|W_{MC}(k)|^2$ can be tightly approximated as

$$|W_{MC}(k)|^2 \approx |\xi_{12}|^2 |h(k)|^2 + |\xi_{21}|^2 |h^*(-k)|^2. \quad (49)$$

Based on the above, it immediately follows that (44) can be expressed as

$$\gamma(k) \approx \frac{|\xi_{11}|^2 \gamma_{\text{id}}(k)}{\theta |\xi_{12}|^2 \gamma_{\text{id}}(k) + \theta |\xi_{21}|^2 \gamma_{\text{id}}(-k) + |K_1^r|^2 + |K_2^r|^2} \quad (50)$$

which is a particularly accurate and simple representation. In general, we wish to emphasize that the nature of IQI is clearly different in the single-carrier and multi-carrier transmission cases. This is primarily because in the multi-carrier case, the image carrier carries an independent data symbol, while in the single-carrier case the image is the complex conjugate of the signal itself. Furthermore, the fading at mirror carriers is generally different while in the single-carrier case the signal and its own conjugate experience the same fading process.

III. Outage Probability over Cascaded Fading Channels

It is recalled that the OP can be defined as the probability that the symbol error rate is greater than a certain quality of service requirement and is computed as the probability that the instantaneous SNR or SINR falls below the corresponding predetermined threshold [52]. In what follows, we derive a novel analytical framework for the OP over N*Nakagami$-m$ fading channels subject to the aforementioned IQI scenarios in both single-carrier and multi-carrier systems. The offered analytic expressions are validated through extensive comparisons with respective results from computer simulations.

A. Ideal RF front-end

In the case of N*Nakagami-m fading channels, the CDF of γ_{ideal} is given by [23, eq. (13)]

$$F_{\gamma_{\text{ideal}}}(\gamma) = \frac{1}{\prod_{i=1}^{N}\Gamma(m_i)} \times G_{1,N+1}^{N,1}\left(\frac{\gamma}{\bar{\gamma}}\prod_{i=1}^{N} m_i \,\middle|\, \begin{matrix} 1 \\ m_1, m_2, \cdots, m_N, 0 \end{matrix}\right) \quad (51)$$

where $\Gamma(\cdot)$ and $G_{s,t}^{v,w}(\cdot)$ denote the gamma function and the Meijer's G-function [53]. Based on this, the corresponding PDF of γ_{ideal} is expressed as [23, eq. (14)], namely

$$f_{\gamma_{\text{ideal}}}(\gamma) = \frac{G_{0,N}^{N,0}\left(\frac{\gamma}{\bar{\gamma}}\prod_{i=1}^{N} m_i \,\middle|\, \begin{matrix} - \\ m_1, m_2, \ldots, m_N \end{matrix}\right)}{\gamma \prod_{i=1}^{N}\Gamma(m_i)}. \quad (52)$$

B. Single-carrier systems impaired by IQI

1) TX impaired by IQI: Using (16) and (51), it follows that the OP can be expressed as

$$P_{\text{out}} = F_{\gamma_{\text{ideal}}}\left(\frac{1}{|K_1^t|^2\left(\frac{1}{\gamma_{th}} - \frac{1}{IRR_t}\right)}\right) \quad (53)$$

with $\gamma_{th} \leq IRR_t$.

2) RX impaired by IQI: With the aid of (19) and (51), the corresponding OP is given by

$$P_{\text{out}} = F_{\gamma_{\text{ideal}}}\left(\frac{1+\frac{1}{IRR_r}}{\frac{1}{\gamma_{th}}-\frac{1}{IRR_r}}\right) \tag{54}$$

with $\gamma_{th} \leq IRR_r$.

3) Joint TX/RX impaired by IQI: Using (33) and (51), the OP in this case is expressed as

$$P_{\text{out}} = F_{\gamma_{\text{ideal}}}\left(\frac{\left|K_1^r\right|^2+\left|K_2^r\right|^2}{\frac{|\xi_{11}|^2}{\gamma_{th}}-\left(|\xi_{12}|^2+|\xi_{21}|^2\right)}\right) \tag{55}$$

with $\gamma_{th} \leq \frac{|\xi_{11}|^2}{|\xi_{12}|^2+|\xi_{21}|^2}$.

C. Multi-carrier systems impaired by IQI

Similar analytic expressions can be derived for the case of multi-carrier transmission.

1) TX impaired by IQI: Based on the signal model in Sec. II-D1 and assuming a known θ, it immediately follows that

$$F_{\gamma}\left(\gamma_{th}\left|\theta\right.\right) = F_{\gamma_{\text{ideal}}}\left(\frac{1}{\left|K_1^t\right|^2\left(\frac{1}{\gamma_{th}}-\theta\frac{1}{IRR_t}\right)}\right) \tag{56}$$

with $\gamma_{th} \leq IRR_t$.

It is also assumed here that θ follows a Bernoulli distribution with CDF

$$P_r\left(\theta\right) = \begin{cases} q, & \theta=1 \\ 1-q, & \theta=0 \end{cases} \tag{57}$$

and $q \in [0,1]$. To this effect, the corresponding unconditional OP can be expressed as follows:

$$P_{\text{out}} = qF_{\gamma_{\text{ideal}}}\left(\frac{1}{\left|K_1^t\right|^2\left(\frac{1}{\gamma_{th}}-\frac{1}{IRR_t}\right)}\right) + \left(1-q\right)F_{\gamma_{\text{ideal}}}\left(\frac{\gamma_{th}}{\left|K_1^t\right|^2}\right). \tag{58}$$

2) RX impaired by IQI: Based on the signal model presented in Sec. II-D2, assuming a given image channel realization, SNR $\gamma_{\text{ideal}}\left(-k\right)$ and a known θ, it follows that

$$F_{\gamma}\left(\gamma_{th}\left|\gamma_{\text{ideal}}\left(-k\right),\theta\right.\right) = F_{\gamma_{\text{ideal}}}\left(\left(\theta\frac{\gamma_{\text{ideal}}\left(-k\right)}{IRR_r}+1+\frac{1}{IRR_r}\right)\gamma_{th}\right). \tag{59}$$

If $\theta=1$, the unconditional CDF can be expressed as follows

$$F_{\gamma}\left(\gamma_{th}\left|\theta=1\right.\right) = \int_0^{\infty}F_{\gamma_{\text{ideal}}}\left(\left(\frac{x}{IRR_r}+1+\frac{1}{IRR_r}\right)\gamma_{th}\right)f_{\gamma_{\text{ideal}}}\left(x\right)\mathrm{d}x. \tag{60}$$

By also taking into consideration (51) and (52), the above expression can be re-written as (61), given at the top of the next page, which after some basic algebraic manipulations can be equivalently expressed as follows:

$$F_{\gamma}\left(\gamma_{th}\left|\theta=1\right.\right) = \frac{1}{\left(\prod_{i=1}^{N}\Gamma\left(m_i\right)\right)^2} \times\int_0^{\infty}\frac{1}{y}G_{1,N+1}^{N,1}\left(y+c\left|\begin{matrix}1\\ m_1,m_2,\ldots,m_N,0\end{matrix}\right.\right) \times G_{0,N}^{N,0}\left(\frac{d}{b}y\left|\begin{matrix}-\\ m_1,m_2,\ldots,m_N,0\end{matrix}\right.\right)\mathrm{d}y \tag{62}$$

where

$$y = bx, \tag{63}$$

$$b = \frac{\gamma_{th}}{\overline{\gamma}}\frac{1}{IRR_t}\prod_{i=1}^{N}m_i, \tag{64}$$

$$c = \frac{\gamma_{th}}{\overline{\gamma}}\left(1+\frac{1}{IRR_t}\right)\prod_{i=1}^{N}m_i \tag{65}$$

and

$$d = \frac{1}{\overline{\gamma}}\prod_{i=1}^{N}m_i. \tag{66}$$

Importantly, equation (62) can be expressed in terms of the Meijer $G-$function in [53] yielding (67), given at the top of the next page. Likewise, when $\theta=0$ the corresponding unconditional CDF can be readily deduced, namely

$$F_{\gamma}\left(\gamma_{th}\left|\theta=0\right.\right) = F_{\gamma_{\text{ideal}}}\left(\left(1+\frac{1}{IRR_t}\right)\gamma_{th}\right). \tag{68}$$

Based on the above, the corresponding unconditional OP is expressed as (69), given at the top of the next page.

3) Joint TX/RX impaired by IQI: By recalling (50) and assuming a given $\gamma_{\text{ideal}}\left(-k\right)$ and a known θ, it immediately follows that

$$F_{\gamma}\left(\gamma_{th}\left|\gamma_{\text{ideal}}\left(-k\right),\theta\right.\right) = F_{\gamma_{\text{ideal}}}\left(\frac{\gamma_{th}\left(\theta\left|\xi_{21}\right|^2\gamma_{ideal}\left(-k\right)+\left|K_1^r\right|^2+\left|K_2^r\right|^2\right)}{\left|\xi_{11}\right|^2-\theta\gamma_{th}\left|\xi_{12}\right|^2}\right). \tag{70}$$

Likewise, in the case of $\theta=1$, the unconditional CDF can be formulated as

$$F_{\gamma}\left(\gamma_{th}\left|\theta=1\right.\right) = \int_0^{\infty}F_{\gamma_{\text{ideal}}}\left(\frac{\gamma_{th}\left(\left|\xi_{21}\right|^2x+\left|K_1^r\right|^2+\left|K_2^r\right|^2\right)}{\left|\xi_{11}\right|^2-\gamma_{th}\left|\xi_{12}\right|^2}\right)f_{\gamma_{\text{ideal}}}\left(x\right)\mathrm{d}x \tag{71}$$

which with the aid of (51)−(52) and carrying out some basic algebraic manipulations can be equivalently expressed as (72), given at the next page. Note that in (72), y, g, l and u stand for

$$y = gx, \tag{73}$$

$$g = \frac{\left|\xi_{21}\right|^2\gamma_{th}}{\overline{\gamma}\left(\left|\xi_{11}\right|^2-\gamma_{th}\left|\xi_{12}\right|^2\right)}\prod_{i=1}^{N}\Gamma\left(m_i\right), \tag{74}$$

$$l = \frac{\left(\left|K_1^r\right|^2+\left|K_2^r\right|^2\right)\gamma_{th}}{\overline{\gamma}\left(\left|\xi_{11}\right|^2-\gamma_{th}\left|\xi_{12}\right|^2\right)}\prod_{i=1}^{N}\Gamma\left(m_i\right) \tag{75}$$

$$F_{\gamma}\left(\gamma_{th}\left|\theta=1\right.\right)=\frac{1}{\left(\prod_{i=1}^{N}\Gamma\left(m_i\right)\right)^2}\int_0^{\infty}x^{-1}G_{0,N}^{N,0}\left(\frac{x}{\overline{\gamma}}\prod_{i=1}^{N}m_i\left|\begin{array}{c}-\\ m_1,m_2,\ldots,m_N\end{array}\right.\right)\times G_{1,N+1}^{N,1}\left(\left(\frac{x}{IRR_t}+1+\frac{1}{IRR_t}\right)\frac{\gamma_{th}}{\overline{\gamma}}\prod_{i=1}^{N}m_i\left|\begin{array}{c}1\\ m_1,m_2,\ldots,m_N,0\end{array}\right.\right)\mathrm{d}x \quad (61)$$

$$F_{\gamma}\left(\gamma_{th}\left|\theta=1\right.\right)\simeq\sum_{k=0}^{p}\frac{(-1)^k\gamma_{th}^k}{\overline{\gamma}^k k!}\left(1+\frac{1}{IRR_t}\right)^k\left(\prod_{i=1}^{N}m_i\right)^k\times G_{N+2,N+2}^{N+1,N+1}\left(\frac{IRR_t}{\gamma_{th}}\left|\begin{array}{c}1,k-m_1+1,k-m_2+1,\ldots,k-m_N+1,k+1\\ m_1,m_2,\ldots,m_N,k,k+1,0\end{array}\right.\right) \quad (67)$$

$$P_{\text{out}}=q\sum_{k=0}^{p}\frac{(-1)^k\gamma_{th}^k}{\overline{\gamma}^k k!}\left(1+\frac{1}{IRR_t}\right)^k\left(\prod_{i=1}^{N}m_i\right)^k G_{N+2,N+2}^{N+1,N+1}\left(\frac{IRR_t}{\gamma_{th}}\left|\begin{array}{c}1,k-m_1+1,k-m_2+1,\ldots,k-m_N+1,k+1\\ m_1,m_2,\ldots,m_N,k,k+1,0\end{array}\right.\right)+(1-q)\frac{1}{\prod_{i=1}^{N}\Gamma\left(m_i\right)}G_{1,N+1}^{N,1}\left(\frac{\left(1+\frac{1}{IRR_t}\right)\gamma_{th}}{\overline{\gamma}}\prod_{i=1}^{N}m_i\left|\begin{array}{c}1\\ m_1,m_2,\cdots,m_N,0\end{array}\right.\right) \quad (69)$$

$$F_{\gamma}\left(\gamma_{th}\left|\theta=1\right.\right)=\frac{1}{\left(\prod_{i=1}^{N}\Gamma\left(m_i\right)\right)^2}\int_0^{\infty}\frac{1}{z}G_{1,N+1}^{N,1}\left(z+l\left|\begin{array}{c}1\\ m_1,m_2,\ldots,m_N,0\end{array}\right.\right)G_{0,N}^{N,0}\left(\frac{u}{g}z\left|\begin{array}{c}-\\ m_1,m_2,\ldots,m_N\end{array}\right.\right)\mathrm{d}z \quad (72)$$

and

$$u=\frac{1}{\overline{\gamma}}\prod_{i=1}^{N}m_i. \quad (76)$$

It is evident that (62) can be also expressed in terms of the Meijer $G-$function [53], namely (77), given at the next page. In the same context, when $\theta=0$, the corresponding unconditional CDF is given by

$$F_{\gamma}\left(\gamma_{th}\left|\theta=0\right.\right)=F_{\gamma_{\text{ideal}}}\left(\frac{\left|K_1^r\right|^2+\left|K_2^r\right|^2}{\left|\xi_{11}\right|^2}\gamma_{th}\right). \quad (78)$$

To this effect, it immediately follows that the unconditional OP can be expressed as (79), given at the next page.

The offered analytic expressions can be readily computed in popular mathematical software packages such as MAPLE, MATHEMATICA and MATLAB. Furthermore, it is noted that to the best of the authors' knowledge, the offered analytic expressions in Sec. III have not been previously reported in the open technical literature.

IV. Applications in Vehicle-to-Vehicle Communications

It is well known that V2V communications constitute a fundamental part of emerging communication systems. This also includes intelligent transportation systems (ITSs), which have been attracting considerable attention due to the large number of applications that they can be deployed for. It is recalled that the transmitted signals in V2V communication systems experience fading effects that typically differ from conventional cellular communications scenarios [54], [55]. This difference arises from the moving nature and the position of the involved TX/RX as well as the presence of reflectors/scatterers in highways and urban environments. As a result, the omnidirectional TX and RX antennas in these systems are located at relatively low elevations and thus, the corresponding wireless channel has been shown to exhibit a non-stationary behavior. As a consequence, the performance of corresponding communication systems is subject to non-negligible deteriorations in terms of throughput and OP, which becomes particularly problematic in certain communications scenarios including safety applications [56]. To this effect, wireless channels in V2V communications should be accurately characterized and modeled in order to evaluate the performance of these systems precisely and incorporate the essential techniques that are capable of ensuring the fulfillment of the corresponding application requirements, resulting to efficient and robust wireless transmission. To this end, we consider a V2V communication system, where the TX and RX are equipped with a single antenna. In this context, we evaluate the corresponding OP over cascaded fading channels for both single-carrier and multi-carrier direct conversion transceivers impaired by IQI.

$$F_\gamma\left(\gamma_{th}\left|\theta=1\right.\right)\simeq\sum_{k=0}^{p}\frac{(-1)^k\gamma_{th}^k\left(\left|K_1^r\right|^2+\left|K_2^r\right|^2\right)^k}{k!\,\overline{\gamma}^k\left(\left|\xi_{11}\right|^2-\gamma_{th}\left|\xi_{12}\right|^2\right)^k}\left(\prod_{i=1}^{N}\Gamma\left(m_i\right)\right)^k \times G_{N+2,N+2}^{N+1,N+1}\left(\frac{\left|\xi_{11}\right|^2-\gamma_{th}\left|\xi_{12}\right|^2}{\left|\xi_{21}\right|^2\gamma_{th}}\left|\begin{matrix}1,k-m_1+1,k-m_2+1,\ldots,k-m_N+1,k+1\\ m_1,m_2,\ldots,m_N,k,k+1,0\end{matrix}\right.\right) \tag{77}$$

$$P_{\text{out}}=q\sum_{k=0}^{p}\frac{(-1)^k\gamma_{th}^k\left(\left|K_1^r\right|^2+\left|K_2^r\right|^2\right)^k}{k!\,\overline{\gamma}^k\left(\left|\xi_{11}\right|^2-\gamma_{th}\left|\xi_{12}\right|^2\right)^k}\left(\prod_{i=1}^{N}\Gamma\left(m_i\right)\right)^k \times G_{N+2,N+2}^{N+1,N+1}\left(\frac{\left|\xi_{11}\right|^2-\gamma_{th}\left|\xi_{12}\right|^2}{\left|\xi_{21}\right|^2\gamma_{th}}\left|\begin{matrix}1,k-m_1+1,k-m_2+1,\ldots,k-m_N+1,k+1\\ m_1,m_2,\ldots,m_N,k,k+1,0\end{matrix}\right.\right) +(1-q)\frac{1}{\prod_{i=1}^{N}\Gamma\left(m_i\right)}G_{1,N+1}^{N,1}\left(\frac{\left|K_1^r\right|^2+\left|K_2^r\right|^2}{\left|\xi_{11}\right|^2\overline{\gamma}}\gamma_{th}\prod_{i=1}^{N}m_i\left|\begin{matrix}1\\ m_1,m_2,\cdots,m_N,0\end{matrix}\right.\right) \tag{79}$$

A. Single-carrier V2V communication system

The complex fading coefficient of the considered wireless communication link is represented by h and is assumed to be the product of statistically independent, but not necessarily identically distributed N*Nakagami$-m$ random variables, namely [54]

$$h=\prod_{i=1}^{N}h_i. \tag{80}$$

It is recalled here that due to the nature of the surrounding environment, the position of the antennas and the mobility of both the transmitter and receiver, the double and triple Nakagami$-m$ distributions have been proven adequate to model fading in basic V2V communications [54]. In what follows, simple analytic expressions are derived for the corresponding measures for these cases which constitute special cases of the more generic N*Nakagami$-m$ fading distribution and are expressed in a simpler algebraic form.

1) Double Nakagami-m channel: It is recalled that the envelope PDF of double Nakagami$-m$ channels i.e. $N=2$, is given in [23, eq. (6)]. Based on this and with the aid of [57, eq. (2.3)], it immediately follows that

$$p_\gamma(\gamma)=\mathcal{A}\gamma^{\frac{m_1+m_2}{2}-1}K_{m_1-m_2}\left(\mathcal{B}\sqrt{\gamma}\right) \tag{81}$$

where $K_n(\cdot)$ denotes the modified Bessel function of the second kind whereas

$$\mathcal{A}=\frac{2}{\overline{\gamma}^{\frac{m_1+m_2}{2}}\prod_{i=1}^{2}\Gamma(m_i)m_i^{-\frac{m_1+m_2}{2}}}, \tag{82}$$

and

$$\mathcal{B}=\frac{2}{\sqrt{\overline{\gamma}}}\prod_{i=1}^{2}m_i. \tag{83}$$

Based on (81), the corresponding CDF is given by

$$F_\gamma(x)=\mathcal{A}\int_0^\gamma x^{\frac{m_1+m_2}{2}-1}K_{m_1-m_2}\left(\mathcal{B}\sqrt{x}\right)\mathrm{d}x \tag{84}$$

which can be expressed in closed-form with the aid of [58, eq. (1.12.2)] yielding

$$\begin{aligned}F_\gamma(\gamma)=&\frac{\mathcal{A}\gamma^{m_1}\Gamma(m_2-m_1)}{m_1 2^{m_1-m_2}\mathcal{B}^{m_2-m_1}}\\&\times\ {}_1F_2\left(m_1;m_1+1,m_1-m_2+1;\frac{\mathcal{B}^2\gamma}{4}\right)\\&+\frac{\mathcal{A}\gamma^{m_2}\Gamma(m_1-m_2)}{m_2 2^{1-m_1+m_2}\mathcal{B}^{m_2-m_1}}\\&\times\ {}_1F_2\left(m_2;m_2+1,m_2-m_1+1;\frac{\mathcal{B}^2\gamma}{4}\right)\end{aligned} \tag{85}$$

where ${}_pF_q(\cdot)$ denotes the generalized hypergeometric function [58]. It is noted here that the above expression is not valid when $m_1=m_2$ as the gamma function is by definition undefined for zero values of its argument[2]. In the same context as above, for the special case that $m_1-m_2\pm\frac{1}{2}\in\mathbb{N}$, the $K_n(\cdot)$ function in (81) can be expressed according to [53, eq. (8.468)]. Based on this and by performing the necessary variable transformation, the SNR PDF of the double Nakagami$-m$ fading model can be alternatively expressed as

$$\begin{aligned}p_\gamma(\gamma)=&\sum_{l=0}^{m_1-m_2-\frac{1}{2}}\sqrt{\frac{\pi}{2\mathcal{B}}}\frac{\mathcal{A}\left(m_1-m_2+l-\frac{1}{2}\right)!}{l!\left(m_1-m_2-l-\frac{1}{2}\right)!(2\mathcal{B})^l}\\&\times\gamma^{\frac{m_1+m_2-l}{2}-\frac{5}{4}}\exp(-\mathcal{B}\sqrt{\gamma})\end{aligned} \tag{86}$$

whereas the corresponding CDF is given by

$$\begin{aligned}F_\gamma(\gamma)=&\sum_{l=0}^{m_1-m_2-\frac{1}{2}}\frac{\mathcal{A}\sqrt{\pi}\Gamma\left(m_1-m_2+l+\frac{1}{2}\right)}{l!\Gamma\left(m_1-m_2-l+\frac{1}{2}\right)(2\mathcal{B})^{l+\frac{1}{2}}}\\&\times\int_0^\gamma x^{\frac{m_1+m_2-l}{2}-\frac{5}{4}}\exp(-\mathcal{B}\sqrt{x})\mathrm{d}x.\end{aligned} \tag{87}$$

The above integral can be expressed in closed-form with the aid of [53, eq. (8.350.1)]. As a result, by performing

[2] In such cases the corresponding results can be obtained with the aid of the generic analytic expressions in Section III. This is also the case in the respective considered scenarios in Section V.

the necessary change of variables and substituting in (87) it follows that

$$F_{\gamma}(\gamma) = \mathcal{A}\sqrt{\pi} \times \sum_{l=0}^{m_1-m_2-\frac{1}{2}} \frac{\Gamma\left(m_1-m_2+l+\frac{1}{2}\right)\gamma\left(m_1-m_2+l+\frac{1}{2},\mathcal{B}\sqrt{\gamma}\right)}{l!2^{l-\frac{1}{2}}\mathcal{B}^{m_1+m_2}\Gamma\left(m_1-m_2-l+\frac{1}{2}\right)} \tag{88}$$

which is also valid for the case that $m_1 - m_2 \pm \frac{1}{2} \in \mathbb{N}$.

To this effect, in the case of single-carrier V2V communication with only TX impaired with IQI, equation (53) can be straightforwardly expressed as follows:

$$P_{\text{out}} = \frac{1}{\Gamma(m_1)\Gamma(m_2)} \times G_{1,3}^{2,1}\left(\frac{1}{|K_1^t|^2\left(\frac{1}{\gamma_{th}}-\frac{1}{IRR_t}\right)\overline{\gamma}}\prod_{i=1}^{2}m_i \middle| \begin{matrix}1\\ m_1,m_2,0\end{matrix}\right) \tag{89}$$

which with the aid of (85) and (87) can be equivalently re-written as (90), given at the top of the next page. Likewise, for the special case that $m_1 - m_2 \pm \frac{1}{2} \in \mathbb{N}$ it can be further simplified to

$$P_{\text{out}} = \sum_{l=0}^{m_1-m_2-\frac{1}{2}} \frac{2^{\frac{1}{2}-l}\Gamma\left(m_1-m_2+l+\frac{1}{2}\right)}{l!\mathcal{B}^{m_1+m_2}\Gamma\left(m_1-m_2-l+\frac{1}{2}\right)} \times \gamma\left(m_1-m_2+l+\frac{1}{2}, \frac{\mathcal{B}}{|K_1^t|\sqrt{\frac{1}{\gamma_{th}}-\frac{1}{IRR_t}}}\right) \tag{91}$$

with $\gamma(a,x)$ denoting the lower incomplete gamma function [58].

In the same context, in the case of only RX impaired with IQI, eq. (54) can be re-written as

$$P_{\text{out}} = \frac{1}{\Gamma(m_1)\Gamma(m_2)} \times G_{1,3}^{2,1}\left(\frac{1+\frac{1}{IRR_r}}{\left(\frac{1}{\gamma_{th}}-\frac{1}{IRR_r}\right)\overline{\gamma}}\prod_{i=1}^{2}m_i \middle| \begin{matrix}1\\ m_1,m_2,0\end{matrix}\right) \tag{92}$$

or alternatively as

$$\begin{aligned} P_{\text{out}} = {} & \frac{\mathcal{A}\Gamma(m_2-m_1)}{m_1 2^{m_1-m_2}\mathcal{B}^{m_2-m_1}}\left(\frac{1+\frac{1}{IRR_r}}{\frac{1}{\gamma_{th}}-\frac{1}{IRR_r}}\right)^{m_1} \\ & \times {}_1F_2\left(m_1; m_1+1, m_1-m_2+1; \frac{\mathcal{B}^2}{4}\frac{1+\frac{1}{IRR_r}}{\frac{1}{\gamma_{th}}-\frac{1}{IRR_r}}\right) \\ & + \frac{\mathcal{A}\Gamma(m_1-m_2)}{m_2 2^{1-m_1+m_2}\mathcal{B}^{m_2-m_1}}\left(\frac{1+\frac{1}{IRR_r}}{\frac{1}{\gamma_{th}}-\frac{1}{IRR_r}}\right)^{m_2} \\ & \times {}_1F_2\left(m_2; m_2+1, m_2-m_1+1; \frac{\mathcal{B}^2}{4}\frac{1+\frac{1}{IRR_r}}{\frac{1}{\gamma_{th}}-\frac{1}{IRR_r}}\right) \end{aligned} \tag{93}$$

which is also valid when $m_1 \neq m_2$. Furthermore, for the special case that $m_1 - m_2 \pm \frac{1}{2} \in \mathbb{N}$, the OP can be expressed as follows:

$$P_{\text{out}} = \sum_{l=0}^{m_1-m_2-\frac{1}{2}} \frac{\mathcal{A}\sqrt{\pi}\Gamma\left(m_1-m_2+l+\frac{1}{2}\right)}{l!2^{l-\frac{1}{2}}\mathcal{B}^{m_1+m_2}\Gamma\left(m_1-m_2-l+\frac{1}{2}\right)} \times \gamma\left(m_1-m_2+l+\frac{1}{2}, \mathcal{B}\sqrt{\frac{1+\frac{1}{IRR_r}}{\frac{1}{\gamma_{th}}-\frac{1}{IRR_r}}}\right). \tag{94}$$

Finally, in the case of joint IQI, equation (55) can be readily expressed as

$$P_{\text{out}} = \frac{1}{\Gamma(m_1)\Gamma(m_2)} \times G_{1,3}^{2,1}\left(\frac{\left(|K_1^r|^2+|K_2^r|^2\right)\gamma_{th}}{\left(\frac{|\xi_{11}|^2}{\gamma_{th}}-\left(|\xi_{12}|^2+|\xi_{21}|^2\right)\right)\overline{\gamma}}\prod_{i=1}^{2}m_i \middle| \begin{matrix}1\\ m_1,m_2,0\end{matrix}\right) \tag{95}$$

which with the aid of (85) and (88) can be alternatively expressed as (96), given at the top of the next page. whereas for the case that $m_1 - m_2 \pm \frac{1}{2} \in \mathbb{N}$, it can be further simplified to (97), given at the top of the next page. It is recalled here that when the distance separating the involved vehicles is larger than 5m, the corresponding line of sight (LOS) component tends to disappear and fading becomes more severe [55]. This also includes double Rayleigh fading conditions [59]–[61], which constitute a special case of the double Nakagami$-m$ for $m_2 = m_1 = 1$.

2) Triple Nakagami-m channel: In case of triple Nakagami$-m$ fading channels, it immediately follows from (51) that the CDF of γ_{ideal} is given by

$$F_{\gamma_{\text{ideal}}}(\gamma) = \frac{1}{\Gamma(m_1)\Gamma(m_2)\Gamma(m_3)} \times G_{1,4}^{3,1}\left(\frac{\gamma}{\overline{\gamma}}m_1m_2m_3 \middle| \begin{matrix}-\\ m_1,m_2,m_3,0\end{matrix}\right). \tag{98}$$

Therefore, in case of TX impaired by IQI, equation (53) can be expressed as

$$P_{\text{out}} = \frac{1}{\Gamma(m_1)\Gamma(m_2)\Gamma(m_3)} \times G_{1,4}^{3,1}\left(\frac{m_1m_2m_3}{|K_1^t|^2\left(\frac{1}{\gamma_{th}}-\frac{1}{IRR_t}\right)\overline{\gamma}} \middle| \begin{matrix}-\\ m_1,m_2,m_3,0\end{matrix}\right) \tag{99}$$

whereas in the case of RX impaired by IQI, equation (54) can be re-written as follows

$$P_{\text{out}} = \frac{1}{\Gamma(m_1)\Gamma(m_2)\Gamma(m_3)} \times G_{1,4}^{3,1}\left(\frac{1+\frac{1}{IRR_r}}{\left(\frac{1}{\gamma_{th}}-\frac{1}{IRR_r}\right)\overline{\gamma}}m_1m_2m_3 \middle| \begin{matrix}-\\ m_1,m_2,m_3,0\end{matrix}\right). \tag{100}$$

Likewise, in the case of joint TX/RX IQI and with the aid of (55), it immediately follows (101), given at the top of the next page.

$$P_{\rm out}=\left(\frac{1}{|K_1^t|^2\left(\frac{1}{\gamma_{th}}-\frac{1}{IRR_t}\right)}\right)^{m_1}\frac{\mathcal{A}\Gamma(m_2-m_1)}{m_1 2^{m_1-m_2}\mathcal{B}^{m_2-m_1}}\,{}_1F_2\left(m_1;m_1+1,m_1-m_2+1;\frac{1}{|K_1^t|^2\left(\frac{\mathcal{B}^2}{4}\frac{1}{\gamma_{th}}-\frac{1}{IRR_t}\right)}\right)$$
$$+\left(\frac{1}{|K_1^t|^2\left(\frac{1}{\gamma_{th}}-\frac{1}{IRR_t}\right)}\right)^{m_2}\frac{\mathcal{A}\Gamma(m_1-m_2)}{m_2 2^{1-m_1+m_2}\mathcal{B}^{m_2-m_1}}\,{}_1F_2\left(m_2;m_2+1,m_2-m_1+1;\frac{\mathcal{B}^2}{4}\frac{1}{|K_1^t|^2\left(\frac{1}{\gamma_{th}}-\frac{1}{IRR_t}\right)}\right) \quad (90)$$

$$P_{\rm out}=\frac{\mathcal{A}\Gamma(m_2-m_1)}{m_1 2^{m_1-m_2}\mathcal{B}^{m_2-m_1}}\left(\frac{|K_1^r|^2+|K_2^r|^2}{\frac{|\xi_{11}|^2}{\gamma_{th}}-\left(|\xi_{12}|^2+|\xi_{21}|^2\right)}\right)^{m_1}{}_1F_2\left(m_1;m_1+1,m_1-m_2+1;\frac{\mathcal{B}^2}{4}\frac{|K_1^r|^2+|K_2^r|^2}{\frac{|\xi_{11}|^2}{\gamma_{th}}-\left(|\xi_{12}|^2+|\xi_{21}|^2\right)}\right)$$
$$+\frac{\mathcal{A}\Gamma(m_1-m_2)}{m_2 2^{1-m_1+m_2}\mathcal{B}^{m_2-m_1}}\left(\frac{|K_1^r|^2+|K_2^r|^2}{\frac{|\xi_{11}|^2}{\gamma_{th}}-\left(|\xi_{12}|^2+|\xi_{21}|^2\right)}\right)^{m_2}{}_1F_2\left(m_2;m_2+1,m_2-m_1+1;\frac{\mathcal{B}^2}{4}\frac{|K_1^r|^2+|K_2^r|^2}{\frac{|\xi_{11}|^2}{\gamma_{th}}-\left(|\xi_{12}|^2+|\xi_{21}|^2\right)}\right) \quad (96)$$

$$P_{\rm out}=\mathcal{A}\sqrt{\pi}\sum_{l=0}^{m_1-m_2-\frac{1}{2}}\frac{\Gamma\left(m_1-m_2+l+\frac{1}{2}\right)\gamma\left(m_1-m_2+l+\frac{1}{2},\mathcal{B}\sqrt{\frac{|K_1^r|^2+|K_2^r|^2}{\frac{|\xi_{11}|^2}{\gamma_{th}}-\left(|\xi_{12}|^2+|\xi_{21}|^2\right)}}\right)}{l!2^{l-\frac{1}{2}}\mathcal{B}^{m_1+m_2}\Gamma\left(m_1-m_2-l+\frac{1}{2}\right)} \quad (97)$$

$$P_{\rm out}=\frac{1}{\prod_{i=1}^{3}\Gamma(m_i)}G_{1,4}^{3,1}\left(\frac{|K_1^r|^2+|K_2^r|^2}{\left(\frac{|\xi_{11}|^2}{\gamma_{th}}-\left(|\xi_{12}|^2+|\xi_{21}|^2\right)\right)\overline{\gamma}}m_1m_2m_3\,\middle|\,\begin{matrix}-\\ m_1,m_2,m_3,0\end{matrix}\right) \quad (101)$$

B. Multi-carrier V2V communication system

In this subsection, we assume that $h(k)$ and $h(-k)$ represent the baseband equivalent wireless communication links complex fading coefficients of the $k^{\rm th}$ and $-k^{\rm th}$ carrier, respectively, whose magnitudes $|h(k)|$ and $|h(-k)|$ follow a cascaded Nakagami$-m$ distribution. This corresponds to the case of the product of statistically independent, but not necessarily identically distributed N*Nakagami$-m$ random variables. To this effect, it immediately follows that

$$|h(k)|=\prod_{i=1}^{N}|h_i(k)| \quad (102)$$

and

$$|h(-k)|=\prod_{i=1}^{N}|h_i(-k)|. \quad (103)$$

It is recalled that due to the nature of the surrounding environment, $|h(k)|$ and $|h(-k)|$ can be also adequately modeled by double or triple Nakagami$-m$ processes.

1) Double Nakagami$-m$ channels: In the case of double Nakagami$-m$ fading channels, the SNR CDF is given by (85). Thus, in the case of TX impaired by IQI, eq. (58) is expressed as (104), given at the top of the next page. Note that (104) is valid when $m_1\neq m_2$. For the special case that $m_1-m_2\pm\frac{1}{2}\in\mathbb{N}$, the OP can be alternatively expressed with the aid of (88), which yields (105), given at the top of the next page.

In the same context, in the case of only RX impaired with IQI, eq. (69) can be re-written as (106), given at the top of the next page, whereas for the case of joint IQI, equation (79) can be expressed as (107), given at the top of the next page. It is again recalled that as the distance between the communicating vehicles becomes larger than 5m, the involved LOS component tends to disappear, which renders the corresponding fading environments more severe, including double Rayleigh fading conditions i.e. ($m_2=m_1=1$).

2) Triple Nakagami-m channels: In case of triple Nakagami-m channels, the OP in the case that the TX is impaired with IQI is given by (58), namely (108), whereas in the case of RX impaired with IQI, equation (69) can be re-written as (109). Finally, in the case of joint TX/RX IQI, eq. (79) can be expressed as (110), given at the top of page 12. To the best of the Authors knowledge, the derived expressions for the above special cases of N*Nakagami$-m$ distribution have not been previously reported in the open technical literature.

V. Numerical Results

In this section, we evaluate and illustrate the effects of IQI on the performance of wireless communications over cascaded Nakagami-m fading channels in terms of the corresponding OP. The notation $m=\{m_1,m_2,\cdots,m_N\}$ denotes up to N*Nakagami-m channels, i.e. $m=\{m_1,m_2,m_3\}$ for $N=3$, with fading $m-$parameters of m_1, m_2, and m_3, respectively.

$$\begin{aligned}P_{\text{out}} =&q\left\{\left(\frac{1}{\left|K_1^t\right|^2\left(\frac{1}{\gamma_{th}}-\frac{1}{IRR_t}\right)}\right)^{m_1}\frac{\mathcal{A}\Gamma(m_2-m_1)}{m_1 2^{m_1-m_2}\mathcal{B}^{m_2-m_1}}\ {}_1F_2\left(m_1;m_1+1,m_1-m_2+1;\frac{\mathcal{B}^2}{4}\frac{1}{\left|K_1^t\right|^2\left(\frac{1}{\gamma_{th}}-\frac{1}{IRR_t}\right)}\right)\right.\\&\left.+\left(\frac{1}{\left|K_1^t\right|^2\left(\frac{1}{\gamma_{th}}-\frac{1}{IRR_t}\right)}\right)^{m_2}\frac{\mathcal{A}\Gamma(m_1-m_2)}{m_2 2^{1-m_1+m_2}\mathcal{B}^{m_2-m_1}}\ {}_1F_2\left(m_2;m_2+1,m_2-m_1+1;\frac{\mathcal{B}^2}{4}\frac{1}{\left|K_1^t\right|^2\left(\frac{1}{\gamma_{th}}-\frac{1}{IRR_t}\right)}\right)\right\}\\&+(1-q)\left\{\left(\frac{\gamma_{th}}{\left|K_1^t\right|^2}\right)^{m_1}\frac{\mathcal{A}\Gamma(m_2-m_1)}{m_1 2^{m_1-m_2}\mathcal{B}^{m_2-m_1}}\ {}_1F_2\left(m_1;m_1+1,m_1-m_2+1;\frac{\mathcal{B}^2}{4}\frac{\gamma_{th}}{\left|K_1^t\right|^2}\right)\right.\\&\left.+\left(\frac{\gamma_{th}}{\left|K_1^t\right|^2}\right)^{m_2}\frac{\mathcal{A}\Gamma(m_1-m_2)}{m_2 2^{1-m_1+m_2}\mathcal{B}^{m_2-m_1}}\ {}_1F_2\left(m_2;m_2+1,m_2-m_1+1;\frac{\mathcal{B}^2}{4}\frac{\gamma_{th}}{\left|K_1^t\right|^2}\right)\right\}\end{aligned}\tag{104}$$

$$\begin{aligned}P_{\text{out}} =& \sum_{l=0}^{m_1-m_2-\frac{1}{2}}\frac{\mathcal{A}q\sqrt{\pi}\Gamma\left(m_1-m_2+l+\frac{1}{2}\right)}{l!2^{l-\frac{1}{2}}\mathcal{B}^{m_1+m_2}\Gamma\left(m_1-m_2-l+\frac{1}{2}\right)}\gamma\left(m_1-m_2+l+\frac{1}{2},\frac{\mathcal{B}}{\left|K_1^t\right|\sqrt{\left(\frac{1}{\gamma_{th}}-\frac{1}{IRR_t}\right)}}\right)\\&-\sum_{l=0}^{m_1-m_2-\frac{1}{2}}\frac{\mathcal{A}(1-q)\sqrt{\pi}\Gamma\left(m_1-m_2+l+\frac{1}{2}\right)}{l!2^{l-\frac{1}{2}}\mathcal{B}^{m_1+m_2}\Gamma\left(m_1-m_2-l+\frac{1}{2}\right)}\gamma\left(m_1-m_2+l+\frac{1}{2},\frac{\mathcal{B}\sqrt{\gamma_{th}}}{\left|K_1^t\right|}\right)\end{aligned}\tag{105}$$

$$\begin{aligned}P_{\text{out}} =& q\sum_{k=0}^{p}\frac{(-1)^k\gamma_{th}^k}{k!\overline{\gamma}^k}\left(1+\frac{1}{IRR_r}\right)^k\left(\prod_{i=1}^{2}m_i\right)^k G_{4,4}^{3,3}\left(\frac{IRR_r}{\gamma_{th}}\left|\begin{matrix}1,k-m_1+1,k-m_2+1,\ldots,k-m_N+1,k+1\\ m_1,m_2,k,k+1,0\end{matrix}\right.\right)\\&+(1-q)\frac{1}{\prod_{i=1}^{2}\Gamma\left(m_i\right)}G_{1,3}^{2,1}\left(\frac{\left(1+\frac{1}{IRR_r}\right)\gamma_{th}}{\overline{\gamma}}\prod_{i=1}^{2}m_i\left|\begin{matrix}1\\ m_1,m_2,0\end{matrix}\right.\right)\end{aligned}\tag{106}$$

$$\begin{aligned}P_{\text{out}} =& q\sum_{k=0}^{p}\frac{(-1)^k\gamma_{th}^k\left(\left|K_1^r\right|^2+\left|K_2^r\right|^2\right)^k}{k!\overline{\gamma}^k\left(\left|\xi_{11}\right|^2-\gamma_{th}\left|\xi_{12}\right|^2\right)^k}\left(\prod_{i=1}^{2}\Gamma\left(m_i\right)\right)^k G_{4,4}^{3,3}\left(\frac{\left|\xi_{11}\right|^2-\gamma_{th}\left|\xi_{12}\right|^2}{\left|\xi_{21}\right|^2\gamma_{th}}\left|\begin{matrix}1,k-m_1+1,k-m_2+1,k+1\\ m_1,m_2,k,k+1,0\end{matrix}\right.\right)\\&+(1-q)\frac{1}{\prod_{i=1}^{2}\Gamma\left(m_i\right)}G_{1,3}^{2,1}\left(\frac{\left|K_1^r\right|^2+\left|K_2^r\right|^2}{\left|\xi_{11}\right|^2\overline{\gamma}}\gamma_{th}\prod_{i=1}^{2}m_i\left|\begin{matrix}1\\ m_1,m_2,0\end{matrix}\right.\right)\end{aligned}\tag{107}$$

$$P_{\text{out}} = \frac{qG_{1,4}^{3,1}\left(\frac{1}{\left|K_1^t\right|^2\left(\frac{1}{\gamma_{th}}-\frac{1}{IRR_t}\right)\overline{\gamma}}\prod_{i=1}^{3}m_i\left|\begin{matrix}1\\ m_1,m_2,m_3,0\end{matrix}\right.\right)}{\prod_{i=1}^{3}\Gamma\left(m_i\right)}+\frac{(1-q)G_{1,4}^{3,1}\left(\frac{\gamma_{th}}{\left|K_1^t\right|^2\overline{\gamma}}\prod_{i=1}^{3}m_i\left|\begin{matrix}1\\ m_1,m_2,m_3,0\end{matrix}\right.\right)}{\prod_{i=1}^{3}\Gamma\left(m_i\right)}\tag{108}$$

$$\begin{aligned}P_{\text{out}} =& q\sum_{k=0}^{p}\frac{(-1)^k\gamma_{th}^k}{k!\overline{\gamma}^k}\left(1+\frac{1}{IRR_r}\right)^k\left(\prod_{i=1}^{3}m_i\right)^k G_{5,5}^{4,4}\left(\frac{IRR_t}{\gamma_{th}}\left|\begin{matrix}1,k-m_1+1,k-m_2+1,k-m_3+1,k+1\\ m_1,m_2,m_3,k,k+1,0\end{matrix}\right.\right)\\&+(1-q)\frac{1}{\prod_{i=1}^{3}\Gamma\left(m_i\right)}G_{1,4}^{3,1}\left(\frac{\left(1+\frac{1}{IRR_r}\right)\gamma_{th}}{\overline{\gamma}}\prod_{i=1}^{3}m_i\left|\begin{matrix}1\\ m_1,m_2,m_3,0\end{matrix}\right.\right)\end{aligned}\tag{109}$$

$$P_{\text{out}} = q\sum_{k=0}^{p}\frac{(-1)^k\gamma_{th}^k\left(|K_1^r|^2+|K_2^r|^2\right)^k}{k!\overline{\gamma}^k\left(|\xi_{11}|^2-\gamma_{th}|\xi_{12}|^2\right)^k}\left(\prod_{i=1}^{3}\Gamma(m_i)\right)^k \times G_{5,5}^{4,4}\left(\frac{|\xi_{11}|^2-\gamma_{th}|\xi_{12}|^2}{|\xi_{21}|^2\gamma_{th}}\,\middle|\,\begin{matrix}1,k-m_1+1,k-m_2+1,k-m_3+1,k+1\\ m_1,m_2,m_3,k,k+1,0\end{matrix}\right) + (1-q)\frac{1}{\prod_{i=1}^{3}\Gamma(m_i)}G_{1,4}^{3,1}\left(\frac{|K_1^r|^2+|K_2^r|^2}{|\xi_{11}|^2\overline{\gamma}}\gamma_{th}\prod_{i=1}^{3}m_i\,\middle|\,\begin{matrix}1\\ m_1,m_2,m_3,0\end{matrix}\right) \tag{110}$$

We also consider that the SNR is normalized with respect to $\gamma_{\rm th}$, which implies that the OP is evaluated as a function of $\gamma/\gamma_{\rm th}$. Furthermore, it is important to note that, unless otherwise is stated, in the following figures, the numerical results are shown with continuous lines, while markers are employed to illustrate the simulation results.

To this end, Fig. 1 illustrates the OP versus the normalized SNR for the different considered TX/RX scenarios for the case of single-carrier communications. Specifically, we compare the OP between the ideal RF front-end, the RX imbalanced, the TX imbalanced and joint TX/RX imbalanced cases when the $IRR = 20$dB, and $\phi = 3^o$. We consider the two cases of $\epsilon < 1$ (continuous lines), and $\epsilon > 1$ (dashed lines). Furthermore, different channels have been considered, where $m = 1$, $m = \{1,1\}$ and $m = \{1,1,1,1\}$ corresponds to the Rayleigh, double Rayleigh and N*Rayleigh with $N = 4$ channels respectively. It is shown that the performance degradation created by the IQI is somewhat less severe compared to the detrimental effects of cascaded fading. For example, for the case of $\gamma/\gamma_{th} = 10$dB, the OP in the case of Rayleigh fading is nearly half the OP value in the case of double Rayleigh fading. In addition, in the case of double Rayleigh fading channels, the assumption of ideal RF front-end results to around 20% error in the corresponding OP. These results highlight the importance of both accurate channel characterization and modeling as well as accounting for RF impairments, in the realistic performance analysis and design of wireless communication systems. It is also interesting to note that, when $\epsilon < 1$, the effects of TX IQI only on the OP degradation are more severe than the corresponding effects of RX IQI only. The underlying reason is that the SINR is higher in the case of RX IQI only than in the case of TX IQI only, since the noise is multiplied by $\left(|K_1^r|^2+|K_2^r|^2\right)$, which for $\epsilon < 1$ does not exceed 1. Moreover, it is worth mentioning that in case of $\epsilon > 1$, the RX IQI effects are the most severe since the noise is multiplied by $\left(|K_1^r|^2+|K_2^r|^2\right)$, which in this case is greater than 1. Interestingly, in case of $\epsilon > 1$, the TX IQI only system outperforms even the corresponding ideal RF front-end system. As can be drawn from (8) and (9), when $\epsilon > 1$, for practical levels of IQI, it follows that $|K_1^t|^2 > 1$, and $|K_2^t|^2 \to 0$. Therefore, eq. (15) can be tightly approximated as

$$\gamma \approx |K_1^t|^2\gamma_{\text{ideal}} \tag{111}$$

which, for $\epsilon > 1$, is greater than γ_{ideal}.

Fig. 2 shows the effects of the IRR on the OP in case of single-carrier communication systems considering double Rayleigh and double Nakagami-m, with $m = \{0.5,0.5\}$, $m = \{2,2\}$ and $m = \{3,3\}$ fading conditions with joint TX/RX IQI. It is evident that the OP decreases as the m values are increased, for a given SNR value, while, as expected, the OP is improved when the IRR is increased. For example, for the case of double Nakagami$-m$ conditions with $\overline{\gamma} = 10$dB, taking an RF front-end with an IRR of 27dB instead of 20dB, decreases the corresponding OP by 30%. Notably, since the IRR of practical RF front-ends lies in the range of $20-40$dB, these results highlight the importance of taking RF impairments such as the IQI into consideration. Likewise, it is also shown that it is of paramount importance to take into account the effects of cascaded fading conditions, as the difference in comparison with non-multiplicative fading is about an order of magnitude in both low and high SNR regimes.

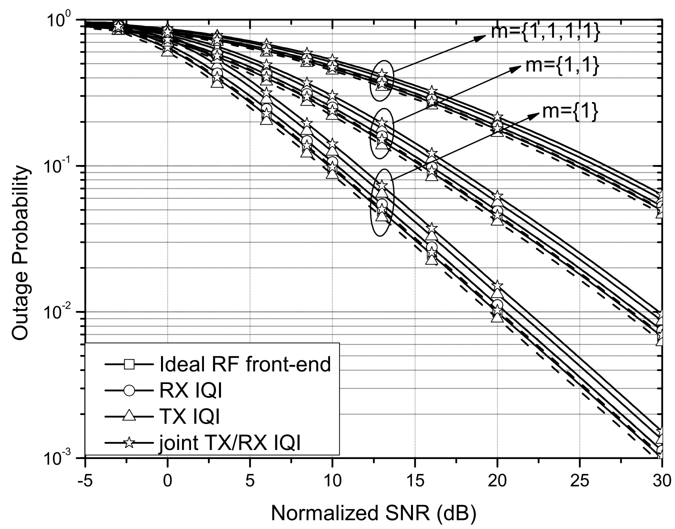

Fig. 1. Single-carrier system P_{out} as a function of the normalized outage SNR, when $IRR = 20$dB, $\phi = 3^o$, $\epsilon \simeq 0.824$ (continuous lines), and $\epsilon \simeq 1.21364$ (dashed lines).

Fig. 3 shows the effects of the IRR on the OP for the different considered TX/RX scenarios assuming double Nakagami-m, with $m = \{2,2\}$ and $m = \{3,3\}$, fading channels. It is evident that the OP is lower for the case of double Nakagami-m fading with $m = \{3,3\}$ compared to the double Nakagami-m fading with $m = \{2,2\}$, while, as expected, the OP is improved when the IRR is increased. For example, for the case of double Nakagami$-m$ with $m = \{3,3\}$ conditions with

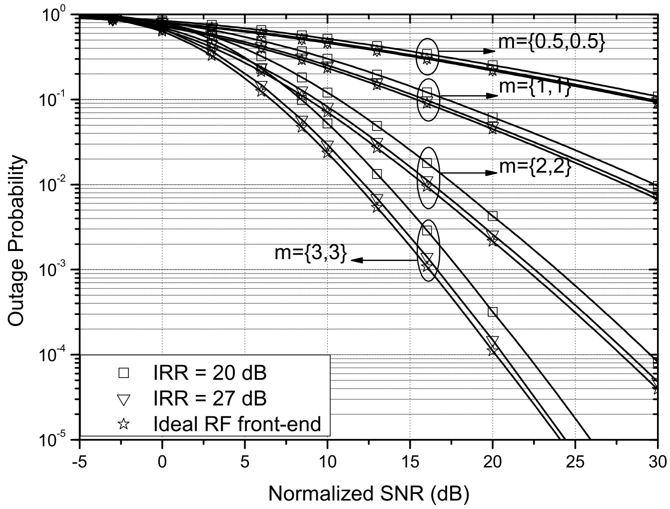

Fig. 2. Single-carrier system P_{out} as a function of the normalized outage SNR, for different values of IRR, when $\epsilon < 1$, and $\phi = 3^o$.

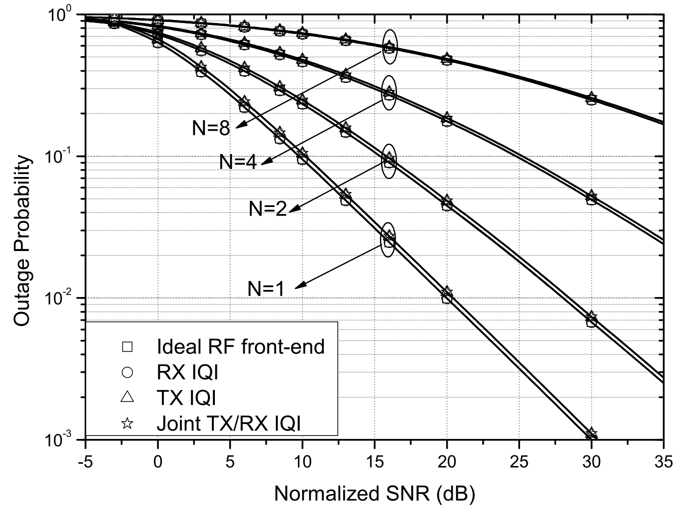

Fig. 4. Multi-carrier system P_{out} as a function of the normalized outage SNR when $q = 0$, $IRR = 25\text{dB}$, $\phi = 3^o$, considering N*Rayleigh channels.

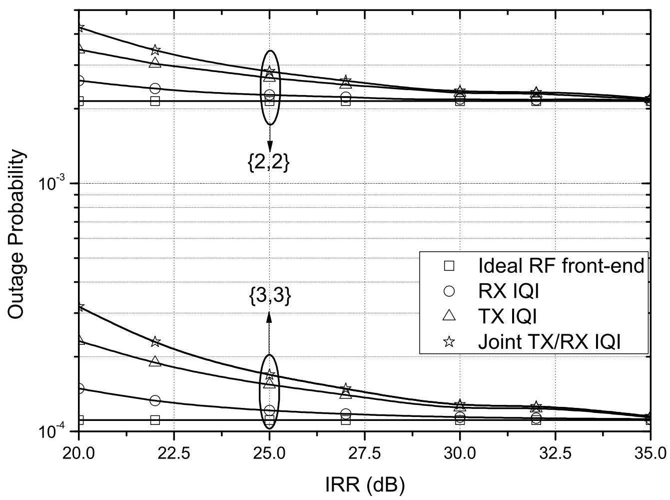

Fig. 3. Single-carrier system P_{out} as a function of the IRR, when $\text{SNR} = 20\text{dB}$ and $\phi = 3^o$.

$\gamma/\gamma_{th} = 20\text{dB}$, taking an RF front-end with an IRR of 25dB instead of 20dB, decreases the corresponding OP by 47%, in case of joint TX/RX IQI. Consequently, for realistic levels of hardware imperfections, these results highlight the detrimental effects of IQI and indicate the importance of taking RF impairments and the statistics of the channel into consideration.

Next, we consider the case of multi-carrier transmission and evaluate the effects of IQI on the OP in the case of signal absence and signal presence at the carrier $-k$, i.e., $q = 0$ and $q = 1$, respectively. We assume mutually uncorrelated channel gains between the carrier k and its image, $-k$. The OP of multi-carrier systems with $q = 0$ over N*Rayleigh channels is demonstrated in Fig. 4, where the impact of cascaded channels on the corresponding performance is clearly observed. Yet, it is noticed that when there is no signal in the image carrier, the signal carried by the carrier k is interference-free. As a result, the performance degradation caused by the RF front-end IQI is much lower than the corresponding degradation in the single-carrier transmission scenario. In fact, when $IRR = 25\text{dB}$, the OP is nearly the same as in the cases that the TX and/or RX are ideal. These results are expected, since the mirror-interference effects of IQI are nullified when there is no signal in the carrier $-k$. Furthermore, we observe that as N increases, the performance degradation become more severe. For instance, for $N = 8$, the OP is higher than 10^{-1}, even for extreme values of normalized SNR. This finding reveals that in case of $q = 0$, the main source of OP degradation is the impact of the cascaded channels and not the hardware imperfections.

Fig. 5 depicts the OP versus the normalized SNR for multi-carrier systems, over both Rayleigh and double Rayleigh channels, for the case of a signal present in the carrier $-k$, i.e. $q = 1$. In this case the IQI causes distortion of the transmitted baseband equivalent signal at the k^{th} carrier, $s(k)$, by its image signal at the carrier $-k$. Furthermore, we observe a lower bound in the case of ideal TX and I/Q imbalanced RX. Intuitively, in case of TX IQI only, mirroring occurs already at TX and the total signal at carrier k, original and mirrored term, both travel through $h(k)$, i.e. fading does not change their ratio. However, in case of RX IQI only, mirroring occurs after wireless transmission over the multiplicative fading channel, so there can be a challenging scenario when $h(k)$ has low value (deep fade) and $h(-k)$ is strong, so the mirrored term can be very strong. Furthermore, as can be drawn from (41), in the high SNR regime, in the presence of signal in the carrier $-k$, the instantaneous SINR per symbol can be accurately approximated by

$$\gamma(k) \approx \frac{\gamma_{\text{ideal}}(k)}{\gamma_{\text{ideal}}(-k)} IRR_r. \tag{112}$$

Furthermore, assuming $\gamma_{\text{ideal}}(k) \approx \gamma_{\text{ideal}}(-k)$, it follows that

$$\gamma(k) \approx IRR_r. \tag{113}$$

In other words, we observe that the maximum achievable SINR is constraint to the IRR levels, because of the effects of IQI.

To this effect, it is shown that the SINR exhibits an upper

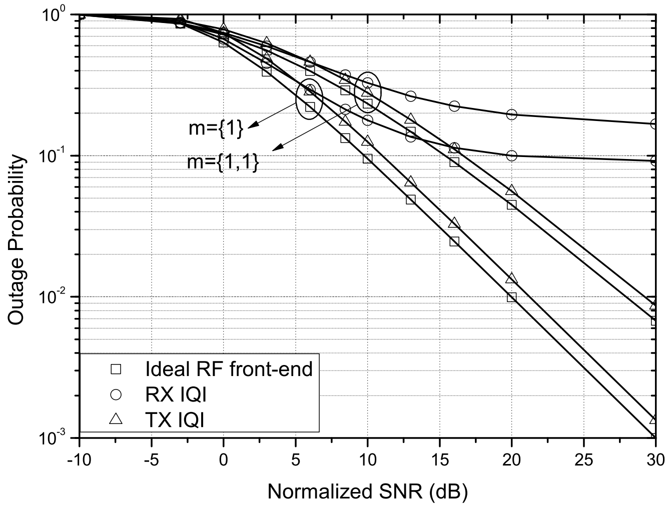

Fig. 5. Multi-carrier system P_{out} as a function of the normalized outage SNR when $q = 1$, $IRR = 20\text{dB}$, $\phi = 3^o$.

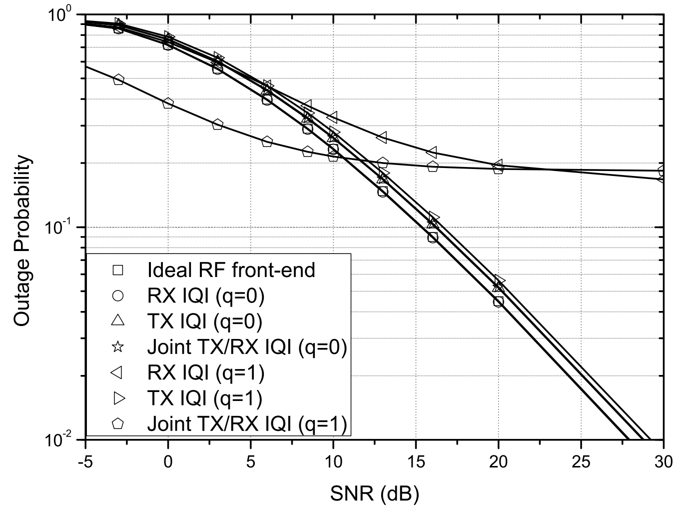

Fig. 6. Multi-carrier system P_{out} as a function of the normalized outage SNR when $m = \{1, 1\}$, $IRR = 20\text{dB}$, and $\phi = 3^o$.

bound which in turn results to a lower bound in the corresponding OP performance. It is noted that the same behavior is experienced in the case of joint TX/RX IQI. Hence, it is evident that this lower bound can create detrimental effects on the performance of communication systems, which stresses the importance of the provided performance analysis and the requirement for efficient compensation techniques.

Finally, Fig. 6 compares the OP of multi-carrier systems over double-Rayleigh channels, when $q = 0$ and $q = 1$. It is observed that the detrimental effects of IQI on the OP performance are significantly increased when a signal is present in the carrier $-k$. Specifically, in the case of $q = 1$ and RX or joint TX/RX IQI, as SNR increases, the mirror-interference increases, resulting to an OP lower bound. In the worst case scenario, where $\text{IRR} = 20$ dB, this bound is in the order of 9×10^{-1}, which may not be acceptable in practice. Meanwhile, the presence of a signal in the carrier $-k$ creates no impact, as expected, on the performance of multi-carrier systems, when the RF front-end is considered ideal.

VI. Conclusions

The present paper investigated the OP performance of single-carrier and multi-carrier systems over cascaded Nakagami-m channels in the presence of IQI at the RF front-end. For the multi-carrier systems, we considered the case that the channel $-k$ is both occupied and unoccupied by an information signal while for each system, we considered three scenarios in our analysis corresponding to ideal TX with I/Q imbalanced RX, I/Q imbalanced TX with ideal RX, and joint I/Q imbalanced TX and RX. The ideal case was also taken into consideration for comparison and the derived analytic results were validated through extensive comparisons with respective results from computer simulations. It was shown that in single-carrier systems the performance degradation caused by IQI in one or both of the RF front-end is more significant than in multi-carrier systems when $q = 0$, while by far the most challenging case is when $q = 1$ in multi-carrier system, in particular in case of RX or joint TX/RX IQI. Furthermore, it was observed in all cases that IQI introduces significant effects that result to non-negligible OP degradations whereas an lower bound on the OP was observed in the high SNR regime. Additionally, it was shown that the effect of cascaded fading conditions on the OP performance are significant as the number of scatterers along with the involved severity of fading can increase or decrease the corresponding performance by about an order of magnitude at both low and high SNR regimes. Finally, the validity and practical usefulness of the offered results were verified by applying them in realistic wireless applications in the context of vehicle-to-vehicle communications over cascaded fading channels providing meaningful insights on the performance of such systems.

Acknowledgments

The authors would like to thank the editor and the anonymous reviewers for their constructive comments and criticism.

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