Self-Interference Cancellation and Beamforming in Repeater-assisted Full-duplex Vehicular Communication

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***Abstract*—Self-driving vehicles will need low-latency and high-capacity vehicular communication for acquiring wider view of their surroundings. Such vehicle-to-vehicle communication can be indirectly supported in some circumstances (e.g., if blocked) through adjacent road side units (RSUs). RSUs will be acting as full-duplex repeaters among the vehicles to ensure low latency and high data rate. However, full-duplex repeaters result in self-interference phenomenon which can degrade the reliability of the communication links. In this work, we aim to enhance the reliability of full-duplex repeaters by canceling out the self-interference impact, and applying a beamforming scheme that is matched to the source-destination composite channel. We show that the proposed self-interference cancellation and beamforming (SICAB) algorithm significantly reduces the error rate for low-isolated repeaters. Finally, we illustrate the impact of the repeater isolation capability on the performance of the proposed SICAB algorithm.**

***Index Terms*—Beamforming, cooperative driving, full-duplex, MIMO, repeaters, self-driving cars, self-interference.**

I. Introduction

The upcoming wave of industrial revolution [1] heavily focuses on *self-driving* vehicles [2]. Enabling autonomous driving depends on two essential and complementary features, namely, *sensing*-based local view and *communication*-based wide view. On one hand, vehicles will be equipped with multiple sensors, cameras and radar equipment, to establish a local view of its adjacent objects. For example, a 3-dimensional mapping can be constructed using LIght Detection and Ranging (LIDAR) technology [3]. On the other hand, vehicles will exploit communications with their neighboring objects to acquire a wider view of their surroundings. Such vehicular communication will include both vehicle-to-vehicle (V2V) and vehicle-to-infrastructure (V2I) communication modes.

Fig. 1 depicts an exemplary scenario, in which there is an urgent need for one self-driving car to provide its local view to the other one in order to save pedestrians' lives. In this scenario, direct V2V may not be possible due to blockage by buildings. Alternatively, employing V2I communication with the neighboring road side unit (RSU) can provide an indirect V2V communication path between the two vehicles. More specifically, the RSU can act as a relay or a repeater broadcasting its received data to the surrounding vehicles. Such scenario falls into the frameworks of *cooperative driving* [4] or *cooperative intelligent transport system* (C-ITS) [5], which are essential for enabling self-driving vehicles.

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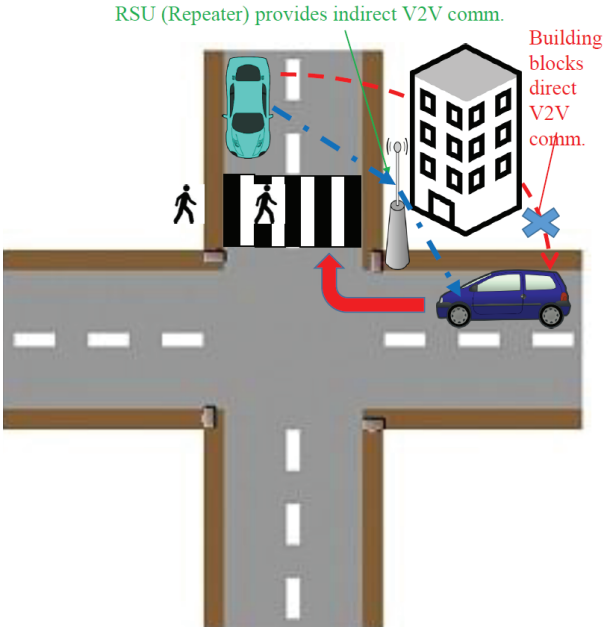

Fig. 1: Cooperative driving can provide indirect vehicle-to-vehicle communication through a road side unit, which acts as a relay (repeater).

RSU-assisted V2V communication, as was exemplified in Fig. 1, needs to provide *high-data-rate* along with *low-latency* communication. *Repeaters*, which are non-regenerative amplify-and-forward relays, are of special importance in vehicular networks as they result in very *low latency* and can operate in *full-duplex* modeEnabling repeaters with full-duplex capability, in which the RSU will be simultaneously transmitting and sending data over the same time-frequency resources, has a great potential to provide low-latency and high-rate communication. Therefore, this work focuses on designing a *repeater-assisted high-rate, low-latency and full-duplex vehicular communication.*

Generally, full-duplex vehicular communication can be utilized to enhance cooperative driving [4]–[6]. For instance, an enhanced carrier senses multiple access with collision avoidance (CSMA/CA) scheme was developed in [4] to enhance

the collision detection using full duplex radios [7]. Despite its advantages, full-duplex mode will result in a *self-interference* phenomenon at the repeater. More specifically, transmitting and receiving at almost the same time causes an oscillatory behavior at the repeater, which dramatically degrades the system performance in terms of its error rate. Therefore, the main *goal* of this paper is to design a *self-interference cancellation* (SIC) scheme that can reduce the error rate in *repeater-assisted full-duplex vehicular communication*.

Generally, self-interference impact can be canceled out to a large dregrre, for example using the null-space of the self-interference channel. However, self-interference cancellation algorithms utilize a portion of the total available degrees of freedom. Hence the number of degrees of freedom, devoted to either providing diversity gain or spatial multiplexing gain, decreases. Such reduction will result in higher error rate. Fortunately, *beamforming* techniques can be utilized to greatly enhance the system reliability as they can achieve both diversity gain and power gain.

Various beamforming techniques have been previously considered in repeater-based networks [8]–[11], in which the channel state information (CSI) can be made available at the source, repeater, or both. The optimum repeater beamforming matrix was obtained in [8] to maximize the channel capacity, and in [9] to maximize the received signal-to-noise-ratio (SNR). Furthermore, joint design of source and repeater beamforming matrices to maximize the channel capacity was presented in [10]. Taking into consideration the source-destination link, a joint design for the source and repeater beamforming matrices was proposed in [11]. We note that the above beamforming works consider maximizing the system performance with *no consideration for the self-interference cancellation*. Beamforming vectors for self-interference cancellation, only at the relay, were proposed in [12], however, there has been no consideration of designing beamforming vectors at the source and destination.

Motivated by the need to make full use of the full-duplex capability of repeaters, while reducing the impact of the repeater self-interference channel, we propose in this work a *self-interference cancellation and beamforming* (SICAB) algorithm. The proposed SICAB algorithm utilizes a limited number of the available degrees of freedom to cancel out the strong contributing directions (eigenvectors) of the self-interference channel. Hence it provides more degrees of freedom, which are utilized to achieve higher diversity and power gains. Moreover, the proposed SICAB algorithm applies transmit and receive beamforming at the source and destination, respectively, which is matched to the composite source-destination channel. In this paper, we show that the proposed SICAB algorithm *reduces the error rate* as it cancels out the effect of repeater self-interference. Such error rate reduction is a major performance metric that indicates having a *reliable* repeater-assisted full-duplex vehicular communication with high data rate and low latency.

The rest of this paper is organized as follows. In the next section, we present the system model of the 3-terminal multiple-input multiple-output (MIMO) communication system. The design of the SICAB algorithm along with the performance analysis are introduced in Section III. In Section IV, we show the simulation results and finally Section V concludes the paper.

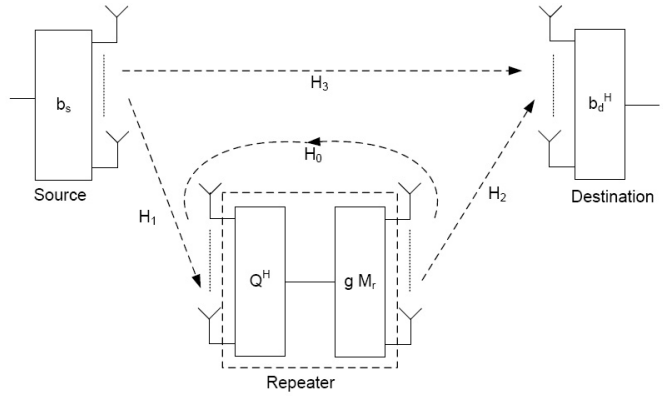

Fig. 2: Repeater-assisted V2V communication system. Source and destination represent two vehicles, while the repeater represents an RSU.

Notations: Matrices and column vectors are denoted by uppercase and lowercase boldface characters, respectively (e.g. $\mathbf{A}$ and $\mathbf{a}$). $\mathbf{H} \in \mathcal{C}^{M\times N}$ denotes that matrix $\mathbf{H}$ is an $M \times N$ matrix of complex elements. An $m \times m$ identity matrix is expressed as $\mathbf{I}_m$. $\mathbf{0}$ denotes an all-zero matrix of appropriate dimensions. The superscript $(.)^H$ denotes matrix conjugate transpose operation. $\mathbf{n} \sim \mathcal{CN}(0, \mathbf{K})$ denotes that vector $\mathbf{n}$ is circular symmetric Gaussian random vector with zero-mean and covariance matrix $\mathbf{K}$. Finally, $E\{.\}$ expresses expectation.

II. System Model

In this section, we introduce the system model of the repeater-based multiple-input multiple-output (MIMO) communication system, which is shown in Fig. 2. It consists of the source, s, the destination, d, and a full-duplex repeater, r. Both the source and destination represent vehicles, while the repeater represent an RSU. In this paper, we consider an orthogonal frequency division multiplexing (OFDM) system with cyclic prefix longer than the delay spread of the effective source-destination channel. Therefore, there is no inter-symbol interference (ISI) resulting from the repeater transmission. In OFDM systems and with no ISI, the system model can be described and analyzed on a subcarrier-by-subcarrier basis. Hence in the following, we introduce the system model based on a single narrow-band subcarrier.

We assume that the source (vehicle) and destination (vehicle) have N_s and M_d omni-directional antennas, respectively. In addition, the repeater (RSU) has M_r and N_r receive and transmit omni-directional antennas, respectively. The repeater self-interference channel matrix is represented by $\mathbf{H}_0 \in \mathcal{C}^{M_r\times N_r}$. In addition, the distance between the transmitting and receiving ends of the repeater is denoted by d_0. The self-interference channel can be modeled as a Line of Sight (LoS)

MIMO channel as in [13]. In particular, the (q,r)-th element of the channel matrix is given by

$$H_0(q,r) = k^{-\frac{1}{2}(q-r)^2}\,, \tag{1}$$

where $k = \exp\left(j\,2\pi\,l^2/(\lambda\,d_0)\right)$, l is the distance between each two antennas in the linear antenna array at the transmitting or receiving end, and λ is the wavelength. Let $\mathbf{H}_1 \in \mathcal{C}^{M_r\times N_s}$, $\mathbf{H}_2 \in \mathcal{C}^{M_d\times N_r}$, and $\mathbf{H}_3 \in \mathcal{C}^{M_d\times N_s}$ represent the source-repeater, repeater-destination, and source-destination channel matrices, respectively, which experience *Rayleigh* fading. Matrices $\mathbf{H}_1$, $\mathbf{H}_2$, and $\mathbf{H}_3$ are mutually independent matrices containing independent and identically distributed (i.i.d.) elements, which are distributed as $\sim \mathcal{CN}(0,1)$. The source-repeater, repeater-destination, and source-destination distances are denoted by d_1, d_2, and d_3, respectively.

Let m be the source symbol with average power of 1, i.e., $E\{|m|^2\} = 1$. The source applies *transmit beamforming* using a unit-norm beamforming vector $\mathbf{b}_s$, which will be determined later in Section III. Thus, the source transmitted signal, $\mathbf{x}_s \in \mathcal{C}^{N_s\times 1}$, can be written as

$$\mathbf{x}_s = \mathbf{b}_s\,m\,. \tag{2}$$

Due to the broadcast nature of the wireless medium, the transmitted signal is received by both the repeater and the destination. The received signal at the repeater, $\mathbf{y}_r \in \mathcal{C}^{M_r\times 1}$, can be modeled as

$$\mathbf{y}_r = \sqrt{P_s\,d_1^{-\alpha}}\,\mathbf{H}_1\,\mathbf{x}_s + \sqrt{d_0^{-\alpha}}\,\mathbf{H}_0\,\mathbf{z}_r + \sqrt{N_0}\,\mathbf{n}_r\,, \tag{3}$$

where P_s is the source power, α is the path-loss exponent, $\mathbf{z}_r$ denotes the repeater transmitted signal, $\mathbf{n}_r \sim \mathcal{CN}(0,\mathbf{I}_{M_r})$ is an additive white Gaussian noise (AWGN), and N_0 is the noise variance.

As shown in Fig. 2, the repeater first cancels out the impact of the self-interference channel using the *unitary* matrix $\mathbf{Q} \in \mathcal{C}^{M_r\times n_q}$, where n_q is the dimension of the matrix $\mathbf{Q}$. Second, the repeater applies a linear mapping operation to map the resulting $n_q \times 1$ signal to the N_r transmit antennas. The mapping is implemented using a mapping matrix $\mathbf{M}_r \in \mathcal{C}^{N_r\times n_q}$, which has a unit-gain per row (corresponding to each receive antenna). Finally, the repeater applies amplification gain per antenna, g. Hence the transmitted signal at the repeater, $\mathbf{x}_r \in \mathcal{C}^{N_r\times 1}$, is equal to

$$\mathbf{x}_r = g\,\mathbf{M}_r\,\mathbf{Q}^H\,\mathbf{y}_r\,. \tag{4}$$

As for the destination, it receives the signal from the source in addition to the amplified signal from the repeater. The received signal at the destination, $\mathbf{y}_d \in \mathcal{C}^{M_d\times 1}$, can be modeled as

$$\mathbf{y}_d = \sqrt{P_s\,d_3^{-\alpha}}\,\mathbf{H}_3\,\mathbf{x}_s + \sqrt{d_2^{-\alpha}}\,\mathbf{H}_2\,\mathbf{x}_r + \sqrt{N_0}\,\mathbf{n}_d\,, \tag{5}$$

where $\mathbf{n}_d \sim \mathcal{CN}(0,\mathbf{I}_{M_d})$ is an AWGN. Substituting (3) and (4) into (5), the received signal at the destination can be rewritten as

$$\mathbf{y}_d = \sqrt{P_s\,d_3^{-\alpha}}\,\mathbf{H}_c\,\mathbf{x}_s + \mathbf{K}_c^{1/2}\,\mathbf{n}_c\,, \tag{6}$$

where $\mathbf{n}_c \sim \mathcal{CN}(0,\mathbf{I}_{M_d})$, $\mathbf{H}_c \in \mathcal{C}^{M_d\times N_s}$ denotes the composite source-destination channel matrix and is given by

$$\mathbf{H}_c = \mathbf{H}_3 + g\,\sqrt{(d_1\,d_2\,/d_3)^{-\alpha}}\,\mathbf{H}_2\,\mathbf{M}_r\,\mathbf{Q}^H\mathbf{H}_1, \tag{7}$$

and $\mathbf{K}_c \in \mathcal{C}^{M_d\times M_d}$ is the covariance matrix of the composite noise, which is equal to

$$\mathbf{K}_c = g^2\,d_2^{-\alpha}\,N_0\,\mathbf{H}_2\,\mathbf{M}_r\,\mathbf{M}_r^H\,\mathbf{H}_2^H + N_0\,\mathbf{I}_{M_d}\,. \tag{8}$$

The self-interference cancellation matrix $\mathbf{Q}$ is a unitary matrix, i.e., $\mathbf{Q}^H\mathbf{Q} = \mathbf{I}_{n_q}$. Based on the system model in (6), we obtain in the next section, the self-interference cancellation matrix, $\mathbf{Q}$, and the transmit beamforming vector, $\mathbf{b}_s$.

III. System Design and Performance Analysis

In this section, we introduce our proposed SICAB algorithm including the derivation of the self-interference cancellation matrix, and transmit and receive beamforming vectors.

A. Self-Interference Cancellation

The singular value decomposition (SVD) of the self-interference channel can be written as

$$\mathbf{H}_0 = \mathbf{U}_0\,\mathbf{D}_0\,\mathbf{V}_0^H\,, \tag{9}$$

where $\mathbf{D}_0 \in \mathcal{C}^{M_r\times N_r}$ is diagonal matrix with ordered non-negative eigenvalues, $\mathbf{U}_0 \in \mathcal{C}^{M_r\times M_r}$, and $\mathbf{V}_0 \in \mathcal{C}^{N_r\times N_r}$ are unitary matrices that consist of the left and right eigenvectors, respectively. Let n_0 denote the number of eigenvalues in $\mathbf{D}_0$, which exceed a certain percentage t_0 (e.g. 10%) of the maximum eigenvalue. We note that if $t_0 = 0$, then n_0 corresponds to the rank of the self-interference channel.

We aim to cancel out the self-interference impact caused by the n_0 strongest directions (eigenvectors). Therefore, the self-interference cancellation matrix $\mathbf{Q}$ includes the eigenvectors that are orthogonal to the n_0 strongest eigenvectors of the self-interference channel. In other words, the matrix $\mathbf{Q}$ consists of the eigenvectors, which lie in the null-space of the matrix $\mathbf{H}_0$ as

$$\mathbf{Q} = \mathbf{U}_0(:,\,n_0+1:M_r)\,, \tag{10}$$

and hence $Q^H\mathbf{H}_0 \simeq \mathbf{0}$. We note that the dimension of the matrix $\mathbf{Q}$ is equal to $n_q = M_r - n_0$. Hence the total degrees of freedom M_r is divided into n_0 degrees, which cancel out the self-interference effect, and n_q degrees, which provide gain (diversity or spatial multiplexing) to the source signal.

B. Beamforming for Composite Channel

In order to obtain the transmit and receive beamforming vectors, we first whiten the colored noise in (6), which has covariance matrix $\mathbf{K}_c$. The whitening step is done by multiplying (6) by $\mathbf{K}_c^{-1/2}$ as

$$\begin{aligned}\mathbf{w}_d &= \mathbf{K}_c^{-1/2}\,\mathbf{y}_d\\ &= \sqrt{P_s\,d_3^{-\alpha}}\,\widetilde{\mathbf{H}}_c\,\mathbf{x}_s + \mathbf{n}_c\,,\end{aligned} \tag{11}$$

where $\widetilde{\mathbf{H}}_c = \mathbf{K}_c^{-1/2}\,\mathbf{H}_c$ is the scaled composite channel and its SVD can be written as

$$\widetilde{\mathbf{H}}_c = \mathbf{U}\mathbf{D}\mathbf{V}^H\,, \tag{12}$$

where $\mathbf{D} \in \mathcal{C}^{M_d \times N_s}$ is a diagonal matrix with ordered non-negative eigenvalues, and $\mathbf{U} \in \mathcal{C}^{M_d \times M_d}$ and $\mathbf{V} \in \mathcal{C}^{N_s \times N_s}$ are unitary matrices. Substituting (12) into (11), we get

$$\mathbf{w}_d = \sqrt{P_s\, d_3^{-\alpha}}\, \mathbf{U}\mathbf{D}\mathbf{V}^H\, \mathbf{b}_s\, m + \mathbf{n}_c\,. \tag{13}$$

Let $\mathbf{u}_1 \in \mathcal{C}^{M_d \times 1}$ and $\mathbf{v}_1 \in \mathcal{C}^{N_s \times 1}$ correspond to the first left and right eigenvectors that correspond to the maximum eigenvalue in $\mathbf{D}$, which is denoted as D_1. The SNR at the destination is maximized when the source and destination are matched to the eigenvectors $\mathbf{v}_1$ and $\mathbf{u}_1$, respectively. Thus, the source and destination beamforming vectors are chosen to be

$$\mathbf{b}_s = \mathbf{v}_1 \;\text{ and }\; \mathbf{b}_d = \mathbf{K}_c^{-1/2}\mathbf{u}_1\,, \tag{14}$$

respectively. We note that the destination receive beamforming vector $\mathbf{b}_d$ includes the whitening step, done in (11). Finally, the processed signal at the destination can be given by

$$\begin{aligned}\hat{m} &= \mathbf{b}_d^H\, \mathbf{y}_d \\ &= \sqrt{P_s\, d_3^{-\alpha}}\; D_1\; m + \hat{n}\,,\end{aligned} \tag{15}$$

where $\hat{n} \sim \mathcal{CN}(0, N_0)$.

C. Performance Analysis

The average probability of error, or symbol error rate (SER), can be computed as [14]

$$p_{e,s} = E_\gamma\{a\, Q(\sqrt{2\, b\, \gamma})\}\,, \tag{16}$$

where γ is the received SNR, $Q(.)$ is the Gaussian Q-function, and a and b are modulation-specific constants. For binary phase-shift keying (BPSK) $a = 1, b = 1$; For binary frequency-shift keying with orthogonal signalling (BFSK) $a = 1, b = 0.5$; For M-PSK with $a = 2, b = \sin^2(\pi/M)$, (16) provides an approximate expression for the SER. From (15), the received SNR can be calculated as [14]

$$\gamma = \frac{P_s\, d_3^{-\alpha}}{N_0}\, D_1^2\,. \tag{17}$$

It is very difficult, if not impossible, to obtain a closed form expression for the distribution of the maximum eigenvalue of the matrix $\widetilde{\mathbf{H}}_c$, which is D_1. Therefore, we calculate the SER in (16) via numerical techniques.

In the following, we comment on the implementation of the proposed SICAB algorithm. First, the repeater needs to estimate the self-interference channel, which is utilized in calculating the self-interference cancellation matrix using (10). In [13] it was shown that the LoS MIMO channel model in (1), which depends on the antenna arrays design, is very adequate to model the self-interference channel matrix. Hence, there is no complexity or overhead in estimating the self-interference channel matrix by the repeater. The repeater also feeds the repeater specifications in (1) to the destination. Second, the destination estimates the source-repeater, repeater-destination, and source-destination channel matrices. Finally, the destination computes the transmit and receive beamforming utilizing (14), and feeds the source beamforming vector to the source.

IV. Simulation Results

In this section, we present the simulation results for the proposed SICAB algorithm. To characterize the impact of the self-interference cancellation, we compare the performance of the proposed SICAB algorithm against that with beamforming but with no self-interference cancellation. For no self-interference cancellation (NSIC), the composite channel is given by

$$\mathbf{H}_{\text{nsic}} = \mathbf{H}_3 + g_{\text{nsic}}\, \sqrt{(d_1\, d_2\, /d_3)^{-\alpha}}\; \mathbf{H}_2\, \mathbf{H}_1\,, \tag{18}$$

and the covariance matrix of the composite noise is equal to

$$\mathbf{K}_{\text{nsic}} = g_{\text{nsic}}^2\, d_2{}^{-\alpha} N_0 \mathbf{H}_2\, \mathbf{M}_{r,\text{nsic}}\, \mathbf{M}_{r,\text{nsic}}^H\, \mathbf{H}_2^H + N_0 \mathbf{I}_{M_d}\,. \tag{19}$$

If $M_r = N_r$ then $\mathbf{M}_{r,\text{nic}} = \mathbf{I}_{N_r}$. Finally, the beamforming vectors are obtained in a similar fashion to (12)-(14) based on the channel matrix $\widetilde{\mathbf{H}}_{\text{nsic}} = \mathbf{K}_{\text{nsic}}^{-1/2}\mathbf{H}_{\text{nsic}}$.

Since the repeater cannot exceed its maximum output power, its amplification gain depends on its input power as $g = P_{\text{out}}/P_{\text{in}}$, where P_{in} and P_{out} are the input and maximum output power, respectively. In the no self-interference case, the repeater input power consists of the power received from the source as well as the self-interference power. On the other hand if the SICAB algorithm is utilized, the input power represents only the power received from the source. Hence, the input power is less for the self-interference cancellation case, which results in higher amplification gain. It can be easily shown that the repeater gain in the no self-interference cancellation case reduces to

$$g_{\text{nsic}} = \frac{g}{1 + 1/\text{SIR}}\,, \tag{20}$$

where SIR expresses the ratio between the power received from the source to the self-interference power.

The simulation parameters can be described as follows. Source, repeater, and destination are equipped with two antennas, i.e., $N_s = M_d = N_r = M_r = 2$ antennas. The source-repeater, repeater-destination, and source-destination distances are equal to $d_1 = 5$m, $d_2 = 5$m, $d_3 = 10$m, respectively. The path-loss exponent is $\alpha = 4$ and QPSK modulation is considered. As for the repeater, its transmit-receive distance is $d_0 = 1$m and its inter-element distance is $l = \lambda/2$. Further, the self-interference channel rank threshold is set to $t_0 = 0.1$ and the repeater gain is set to $g = 80$dB The number of strong interference eigenvectors for the self-interference channel matrix, modeled as in (1), is $n_0 = 1$. Consequently, the rank of the self-interference cancellation matrix is $n_q = 1$. The mapping vector $\mathbf{M}_r$ is the all-ones 2×1 vector, which maps the processed signal to the first and second antennas with unit gain per antenna. Finally, the signal wavelength is $\lambda = 0.15$m.

Fig. 3 depicts the bit error rate (BER) as a function of $P_b/N_0 = P_s/(2\, N_0)$, where P_b is the power per single bit. Multiple values of SIR (0 to 20 dB) are assumed for the ratio between the power received from the source to the self-interference power. For QPSK modulation and based on (16) and (17), the analytical BER is calculated

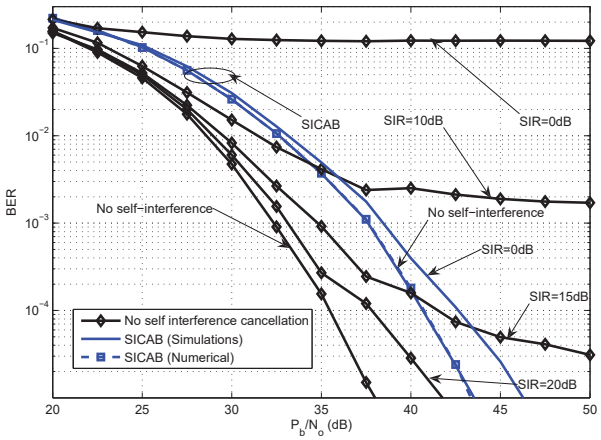

Fig. 3: BER for QPSK modulation and $N_s = M_d = N_r = M_r = 2$ antennas at different SIRs.

as $p_{e,b} = E_{D_1}\{Q(\sqrt{2\,P_b\,d_3^{-\alpha}\,D_1^2/N_0})\}$. First, it is shown that the SIR significantly affects the performance of the no-self-interference cancellation algorithm compared to that of the SICAB algorithm. This is because the SICAB algorithm cancels out the effect of the strongest interference direction, while the NSIC algorithm does not. Second, it is shown that for low SIR, the proposed SICAB algorithm outperforms that of NSIC algorithm and that the NSIC experiences error floor.

For high SIR, on the other hand, the performance of the SICAB algorithm is worse than that of the NSIC algorithm. This dependence of the performance on the SIR can be explained as follows. As indicated previously, the SICAB algorithm reduces the degrees of freedom, devoted to improving the source signal, due to the self-interference cancellation. This partial loss in degrees of freedom is of no use at high SIR. Therefore, at high SIR the NSIC algorithm outperforms the SICAB algorithm as it utilizes more degrees of freedom. Finally, we point out that an advanced repeater will determine whether a self-interference cancellation algorithm is needed or not based on its SIR, which is directly related to its transmit-receive isolation capability.

V. Conclusion

Enabling self-driving vehicles heavily depends on having repeater-assisted high-rate, low-latency, and full-duplex vehicular communication. In this paper, we have proposed a self-interference cancellation and beamforming (SICAB) algorithm, which maximizes the performance of the full-duplex amplify-and-forward repeaters while minimizing the effect of the repeater self-interference phenomenon. The SICAB algorithm 1) cancels out the strongest contributions of the repeater self-interference channel and 2) applies beamforming, which is matched to the source-destination composite channel. It was shown that for low repeater signal-to-interference ratio (SIR) the SICAB significantly improves the performance. On the other hand for high SIR, it was shown that the no-self-interference cancellation (NSIC) algorithm outperforms the SICAB algorithm. Finally, the next steps of this work include 1) the utilization of millimeter-wave frequency band along with investigating the impact of Radio Frequency (RF) propagation characteristics at such frequency band and 2) validating the proposed algorithm using real vehicular testbed.

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