

The Expected Achievable Distortion of Two-User Decentralized Interference Channels

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Abstract- This paper concerns the transmission of two independent Gaussian sources over a two-user decentralized interference channel, assuming that the transmitters are unaware of the instantaneous CSIs. The availability of the channel state information at receivers (CSIR) is considered in two scenarios of perfect and imperfect CSIR. In the imperfect CSIR case, we consider a more practical assumption of having an MMSE estimation of the channel gain at the receivers. In this case, minimizing the expected achievable distortion associated with each link is considered. Due to the absence of CSI at the transmitters, the Gaussian sources are encoded in a successively refinable manner and the resulting code words are transmitted over the channel using a multi-layer coding technique. Accordingly, the optimal power assignment between code layers leading to the least expected achievable distortion, under a mean-square error criterion is derived for both, the perfect and imperfect CSIR scenarios. Finally, some numerical examples are provided and it is demonstrated that the proposed method results in better performance as compared with the conventional single-layer approach, termed as outage approach.

Index Terms- Calculus of Variations, Multi-layer Coding, Successive Refinement.

I. INTRODUCTION

Interference is one of the important barriers in front of improving the quality of service in the wireless communication systems, and two-user Gaussian interference channel is a simple model for describing two transmitter-receiver pairs interfering with each other. Addressing the capacity of Gaussian interference channel has gained a considerable attention in many papers and it is derived in some special cases like [1]-[3]. Furthermore, the aforementioned channel as well as some of its variations are the subject of newer studies in this era [4], [5].

On the other hand, multi-layer coding is deemed to be an advantageous mechanism for maximizing the average achievable rate in point-to-point communication channels, when the transmitter is

oblivious about the instantaneous CSI and the channel is under a block-fading characteristics [6]-[12]. This is because the communications system using the multi-layer coding approach is equipped with a variable-rate strategy that matches the fading characteristics of the channel. Dealing with a single-hop point to point channel, the authors of [6] have come up with an optimal expression for power allocation function to the code layers, which maximizes the average achievable rate in a Single-Input Single-Output (SISO) block-fading channel. However, their attempt to extend the approach to a Multi-Input Multi-Output (MIMO) case did not result in an optimal solution. In [7]-[10] the notion of multi-layer coding is extended to two-hop channels and in [11] it is extended to multi-hop networks, where maximizing the average achievable rate is studied in various cases and the optimal power allocation functions are addressed. [12] is an example of considering the multi-layer coding approach in wiretap channels, where the physical layer secrecy rate is considered as the main objective and the optimal average secrecy rate in various scenarios are derived. In [13], multi-layer coding is applied to a decentralized interference channel, where it is shown that the resulting expected achievable rate associated with multi-layer coding surpasses the conventional outage approach.

Minimizing the expected mean-square error (MSE) distortion of a Gaussian source, transmitted over a communication channel has been considered in various cases. There are many works like [14]-[19] that consider the joint source-channel scheme in minimizing the average achievable distortion. Among them, some papers like [14] and [15] tackle the subject by incorporating the rate-distortion function [20] and some focus on studying the achievable distortion in high SNR regime, where minimizing the average distortion changes to maximizing the distortion exponent as the performance measure [16]-[19]. On the other hand, there are a wide range of papers that look at the subject of minimizing the achievable distortion through the source-channel separation perspective and consider the hierarchically source coding [21], coupled with multi-layer channel coding to derive the average achievable distortion in a communication channel [22]-[25]. In [22] and [23] the average achievable distortion of Gaussian source in a point to point communication channel is minimized and the concept is extended into relay-assisted channels in [24] and [25].

We first formalize a generalized notion for the distortion for a one-hop communications channel and then consider minimizing the expected achievable distortion of two Gaussian sources, transmitted over a decentralized interference channel. To this end, we consider two scenarios of having perfect channel state information at the receivers regarding the direct channel gains, namely full-CSIR scenario, and a more practical assumption of having an MMSE estimate of the direct channel gains, namely imperfect-CSIR scenario. In both cases, it is assumed that the communication is occurring in a block-fading environment, where the transmitters are only aware of probability density functions, associated with their corresponding channel gains.

The organization of the rest of the paper is as follows. In Section II, the background information about hierarchical source coding and the notion of multi-layer channel coding are provided.

Furthermore a generalized formulation of the expected distortion at the destination of a single hop channel is presented. Section III is dedicated to investigating the minimization of the expected distortion of two independent Gaussian sources transmitted over an interference channel, where two scenarios of full-CSIR and imperfect CSIR are addressed under subsections A and B, respectively. Numerical results and conclusion of the paper are provided in Section IV and Section V, respectively.

II. HIERARCHICAL SOURCE AND MULTI-LEVEL CHANNEL CODING

Let's consider a transmitter, that is set out to send a complex Gaussian source over a point-to-point block fading single-input single-output channel in the presence of a circularly symmetric zero-mean complex Gaussian noise of unit variance, i.e., $\mathbf{CN}(0,1)$. The receiver uses the CSI to decode its received signal, while the transmitter is oblivious about it. In this case, it is widely recognized that employing a multi-layer code and assigning the available power to the code layers in the optimal way, maximizes the average achievable rate at the destination. Incorporating an infinite layer code at the transmitter, it is shown in [6] that the achievable rate associated with an instantaneous strength of the channel realization, say s , is

$$R(s) = \int_0^s \frac{u\pi(u)du}{1+u\Pi(u)}, \quad (1)$$

where $\pi(s)ds$ and $\Pi(s) = \int_s^\infty \pi(u)du$ denote respectively, the fractional power assigned to layer s , and the amount of power assigned to the undecodable layers.

Considering a block-fading environment, the Gaussian source symbols of the length l_s are mapped to the channel symbols of the length l_c in a linear way, where the source/channel mismatch factor is defined as the ratio between them, i.e., $b = l_c / l_s$. It is assumed that the transmitted blocks are shorter than the dynamics of the channel, however, they are long enough to approach the rate-distortion limit.

On the other hand, transmitting a Gaussian source with the rate $R(s)$ results in the distortion level of $D(s) = \exp(-bR(s))$ at the destination [20]. The assistant function $T(s) = D(s)^{-1/b}$ is recommended in [22] to formulate the average achievable distortion problem under a fixed transmission power constraint P_t . Incorporating a positive continuous weighting function like $w(\cdot)$ which is constant in the regions that the power allocation does not take place, a generalized notion of expected distortion can be formulated as follows,

$$\bar{D} = \min_{T(\cdot)} \int_0^{\infty} w(s) \frac{f_s(s)}{T(s)^b} ds$$

$$s.t. \begin{cases} \int_0^{\infty} \frac{T(s)}{s^2} ds \leq P_t \\ \frac{d}{ds} T(s) > 0 \end{cases} \quad (2)$$

where $f(s)$ is the probability density function (pdf) associated with the layer s , and the last constraint ensures that the solution provides a non-negative power allocation for each layer. Tackling the above problem, one can write the following Lagrangian form and use the method of calculus of variations [26],

$$\mathcal{L}(T) = \int_0^{\infty} \left(w(s) \frac{f_s(s)}{T(s)^b} + \lambda \frac{T(s)}{s^2} - \gamma(s) T'(s) \right) ds \quad (3)$$

In the above equation, λ is the Lagrange multiplier associated with the power constraint and $\gamma(\cdot)$ is an arbitrary nonnegative function that is used to ensure the positivity of $T'(\cdot)$.

Taking the variational notation of $A(s, T, T') = \frac{f_s(s)}{T(s)^b} + \lambda \frac{T(s)}{s^2} - \gamma(s) T'(s)$, and noting

$A_T + \frac{d}{ds} A_T' = 0$ for the optimal solution [26], one can arrive at,

$$\frac{-bw(s)f(s)}{T(s)^{b+1}} + \frac{\lambda}{s^2} + \gamma'(s) = 0, \quad (4)$$

where $\lambda \left(\int_0^{\infty} \frac{T(s)}{s^2} ds - P_t \right) = 0$ and $\gamma(s) T'(s) = 0$ are slackness conditions. Obviously, the

optimal solution takes place in some intervals that $T'(s) > 0$ which results in $\gamma(s) = 0$. Considering a single interval of power allocation like $[s_1, s_2]$ and noting that $T(s) \Big|_{s=s_1} = 1$, the following optimal assistant function becomes,

$$T(s) = \left(\frac{s^2 w(s) f_s(s)}{s_1^2 w(s_1) f_s(s_1)} \right)^{\frac{1}{b+1}}. \quad (5)$$

Re-writing (3) with single interval of power allocation assumption, after some mathematics one can arrive at the equation $1 - F_s(s_2) = s_2 f_s(s_2)$ to determine the optimum endpoint of power allocation.

Additionally, the positive power allocation constraint should be discussed. As noted earlier, the optimal power allocation takes place on intervals which $T'(s)$ is greater than zero. Thus, the

corresponding slack function, namely $\gamma(s)$, must be equal to zero. In this case (4) changes to,

$$\frac{-bw(s)f_s(s)}{T(s)^{b+1}} + \frac{\lambda}{s^2} = 0, \quad (6)$$

which leads to the following condition for the positive power allocation,

$$\frac{d}{ds}(s^2w(s)f_s(s)) > 0. \quad (7)$$

Now, it should be proved that once the start and end points of a power allocation interval is derived, there is no discontinuity in it. To prove this, suppose otherwise; i.e., consider M intervals of power allocation, where, the inequality (7) holds and there is a discontinuity between the start and end points of power allocation in the i^{th} interval. To be more clear, assume that the power allocation interval is $[s_{il}, s_{iu}]$ and we have the discontinuity in $[d_l, d_u] \in [s_{il}, s_{iu}]$. Since the derived solution is optimal, the following corner conditions [26] must be satisfied,

$$\begin{aligned} \mathbf{L}'_T \Big|_{s=s_c^-} &= \mathbf{L}'_T \Big|_{s=s_c^+}, \\ (\mathbf{L} - T' \mathbf{L}'_T) \Big|_{s=s_c^-} &= (\mathbf{L} - T' \mathbf{L}'_T) \Big|_{s=s_c^+}. \end{aligned} \quad (8)$$

Plugging \mathbf{L} from (3) to (8) and evaluating the relations at corner points of d_l and d_u , one would arrive at $\gamma(d_l) = \gamma(d_u) = 0$. Also for $s \in (d_l, d_u)$, where no power allocation exists and $T(s)$ is equal to $T(d_l)$, the following holds,

$$\gamma'(s) = \frac{bw(s)f(s)}{T(d_l)^{b+1}} - \frac{\lambda}{s^2}. \quad (9)$$

On the other hand, according to (6) for the point $s = d_l$ we have $\frac{bw(s)f(s)}{T(d_l)^{b+1}} - \frac{\lambda}{s^2} = 0$. However, since $s^2w(s)f_s(s)$ has a strictly positive derivative, it can be concluded that $\gamma'(s)$ described in (8) is positive in $s \in (d_l, d_u)$. This statement contradicts $\gamma(d_l) = \gamma(d_u) = 0$, therefore, the assumption of having discontents power allocation intervals is not true.

For instance, in the Rayleigh fading case we have $M = 1$ interval of power allocation. So, deriving the values of s_1 and s_2 , one can arrive at the optimal power allocation function that minimizes the average achievable distortion in a Rayleigh point to point block-fading channel.

III. THE PROPOSED APPROACH

In this section, we concentrate on $w(x) = 1$ and formalize the problem of minimizing the average achievable distortion of two complex Gaussian sources, transmitted over a two-user decentralized

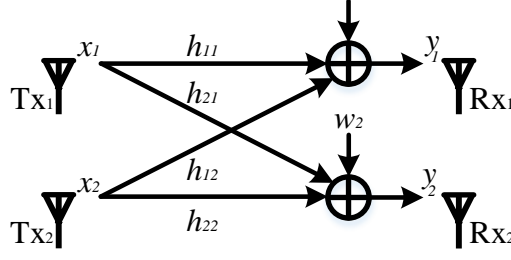


Fig. 1. Two-user interference channel, considered in the current study.

interference channel, when the transmitters are unaware of the channel state information associated with their forward links (*Fig. 1*). The received signal at each receiving node can be represented as,

$$y_j = \mathbf{h}_j^T \mathbf{x} + w_j, \quad (10)$$

where $j \in \{1, 2\}$ determines the intended receiver, $\mathbf{h}_j = [h_{j1}, h_{j2}]^T$ represents the channel coefficients vector associated with either of the transmitters to the j^{th} receiver with power $s_{ji} = |h_{ji}|^2$, $\mathbf{x} = [x_1, x_2]^T$ denotes the transmitted Gaussian codebooks associated with transmitters and finally w_j is a zero-mean circularly symmetric Gaussian noise with the variance of unity. Moreover, the first and the second transmitter are subject to the peak power constraint P_1 and P_2 , respectively. According to the symmetry of the problem, without loss of generality, one can consider the minimization of the first link's achievable distortion as the main goal. We discuss the problem in two scenarios of having and having not access to the perfect channel state information at the receivers.

A. The full-CSIR scenario

In the case of having full-CSIR, we can formulate the first receiver's received signal as follows,

$$y_1 = h_{11}x_1 + h_{12}x_2 + n_1. \quad (11)$$

In this case, the mutual information associated with the first transmitter and receiver, regarding the full-CSIR condition would be,

$$\begin{aligned} I(x_1; y_1 | h_{11}) &= h(y_1 | h_{11}) - h(y_1 | x_1, h_{11}) \\ &= 2\pi e \cdot \log(s_{11}P_1 + s_{12}P_2 + 1) - 2\pi e \cdot \log(s_{12}P_2 + 1) \\ &= \log\left(1 + \frac{s_{11}}{s_{12}P_2 + 1} P_1\right). \end{aligned} \quad (12)$$

Thus, the received signal strength would be $s = \frac{s_{11}}{s_{12}P_2 + 1}$, and the equivalent CDF of the first link

can be calculated from the following equation,

$$F_S(s) = \int_0^{\infty} \int_0^{s(s_2 P_2 + 1)} f_{s_{11}}(s_{11}) f_{s_{12}}(s_{12}) ds_{11} ds_{12}, \quad (13)$$

For instance, incorporating the exponential distribution for s_{11} and s_{12} , one can derive the CDF and the pdf of the equivalent channel as,

$$F_S(s) = 1 - \frac{\exp(-s)}{sP_2 + 1},$$

$$f_S(s) = \frac{\exp(-s)}{(sP_2 + 1)^2} (sP_2 + P_2 + 1). \quad (14)$$

Now, considering the problem formulation of (2), the optimal assistant function that minimizes the average achievable distortion in a decentralized interference channel is derived as follows,

$$T^{opt}(s) = \left[\exp(s_1 - s) \cdot \frac{(s_1 P_2 + 1)^2}{(s P_2 + 1)^2} \cdot \frac{s P_2 + P_2 + 1}{s_1 P_2 + P_2 + 1} \right]^{\frac{1}{b+1}}. \quad (15)$$

Moreover, plugging (14) into the equation $1 - F_S(s_2) = s_2 f_S(s_2)$ to determine the endpoint of the power allocation interval, leads to a second order equation which results in,

$$s_2 = \frac{-1 + \sqrt{1 + 4P_2}}{2P_2}. \quad (16)$$

It is worth noting that tending P_2 to zero, results in approaching s_2 to one, which is derived in [22]. The start point of power allocation interval, namely s_1 , can be determined by inserting (15) and (16) to the power constraint of (2). Finally, the mean achievable distortion of the network can be derived using the objective function of (2).

In what follows, we are going to state that the single interval of power allocation leads to the optimal assistant function. Considering (14) the $\frac{d}{ds}(s^2 f_S(s)) > 0$ becomes,

$$\left(-P_2^2 s^3 - 2P_2 s^2 + (2P_2 - 1)s + 2P_2 + 2 \right) \cdot \frac{s \exp(-s)}{(sP_2 + 1)^3} > 0. \quad (17)$$

Considering the above representation as $P(s) \cdot \frac{s \exp(-s)}{(sP_2 + 1)^3}$, it is argued in [13] according to the

"Descartes' Rule of Signs" in [27], that the polynomial $P(s)$ has only one positive root, due to having just a single variation in the consecutive signs of the coefficients. Moreover, taking the negative sign of the highest degree's coefficient into account, it is obvious that for large values of s , $P(s)$ is negative. On the other hand, inserting s_2 into the $P(s)$ we have,

$$P(s)\Big|_{s=s_2} = \frac{1+4P_2+\sqrt{1+4P_2}}{2}, \quad (18)$$

which is a positive value. Therefore, one can conclude that the positive root of $P(s)$, namely s_r , is greater than s_2 . Thus $\frac{d}{ds}(s^2 f_s(s))$ results in the single interval of $[0, s_r]$ which subsumes the derived power allocation between s_1 and s_2 and the resulting solution is optimal.

B. The imperfect-CSIR scenario

For the more practical case of having the imperfect CSI at the receiver, one can consider the channel state information is estimated at the receiver by the use of MMSE approach. The MMSE estimation of the channel gain for the current channel is [28],

$$\hat{h}_{11} = \frac{P_1}{P_1 + s_{12}P_2 + 1} h_{11}. \quad (19)$$

Where $\hat{n}_{11} = \frac{s_{12}P_2 + 1}{P_1 + s_{12}P_2 + 1} h_{11}$ is the estimation error and we have $h_{11} = \hat{h}_{11} + \hat{n}_{11}$. In this case, the received signal can be represented as follows,

$$y_1 = \frac{P_1}{P_1 + s_{12}P_2 + 1} h_{11}x_1 + \frac{s_{12}P_2 + 1}{P_1 + s_{12}P_2 + 1} h_{11}x_1 + h_{12}x_2 + n_1. \quad (20)$$

The first term in the left hand side of the above equation, represents the decoded signal component at the destination. The second term is the estimation noise which is added with the interference and noise, i.e., third and fourth terms. The mutual information associated with the first link can be formulated as follows,

$$\begin{aligned} I(x_1, y_1 | \hat{h}_{11}) &= h(y_1 | \hat{h}_{11}) - h(y_1 | x_1, \hat{h}_{11}) \\ &= 2\pi e \cdot \log \left(\left(\frac{P_1}{P_1 + s_{12}P_2 + 1} \right)^2 s_{11}P_1 + \left(\frac{s_{12}P_2 + 1}{P_1 + s_{12}P_2 + 1} \right)^2 s_{11}P_1 + s_{12}P_2 + 1 \right) \\ &\quad - 2\pi e \cdot \log \left(\left(\frac{s_{12}P_2 + 1}{P_1 + s_{12}P_2 + 1} \right)^2 s_{11}P_1 + s_{12}P_2 + 1 \right) \\ &= \log \left(1 + \frac{\left(\frac{P_1}{P_1 + s_{12}P_2 + 1} \right)^2 s_{11} P_1}{\left(\frac{s_{12}P_2 + 1}{P_1 + s_{12}P_2 + 1} \right)^2 s_{11}P_1 + s_{12}P_2 + 1} \right) \end{aligned} \quad (21)$$

In the above formula, noting that the error part is orthogonal with respect to the estimated part of the signal is necessary. Defining the equivalent channel gain as,

$$s = \frac{\left(\frac{P_1}{P_1 + s_{12}P_2 + 1}\right)^2 s_{11}}{\left(\left(\frac{s_{12}P_2 + 1}{P_1 + s_{12}P_2 + 1}\right)^2 s_{11}P_1 + s_{12}P_2 + 1\right)}, \quad (22)$$

the CDF of the equivalent channel gain can be derived from the following,

$$F_S(s) = \Pr(S \leq s) = \Pr\left(S \leq \frac{\left(\frac{P_1}{P_1 + s_{12}P_2 + 1}\right)^2 s_{11}}{\left(\left(\frac{s_{12}P_2 + 1}{P_1 + s_{12}P_2 + 1}\right)^2 s_{11}P_1 + s_{12}P_2 + 1\right)}\right) \quad (23)$$

Considering $\left(\frac{P_1}{P_1 + s_{12}P_2 + 1}\right)^2 < \left(\frac{s_{12}P_2 + 1}{P_1 + s_{12}P_2 + 1}\right)^2 sP_1$, the relation can be expressed as follows,

$$F_S(s) = \Pr\left(s_{11} \leq \frac{s(s_{12}P_2 + 1)}{\left(\frac{P_1}{P_1 + s_{12}P_2 + 1}\right)^2 - \left(\frac{s_{12}P_2 + 1}{P_1 + s_{12}P_2 + 1}\right)^2 sP_1}\right) \quad (24)$$

Absolutely, there are some states for the parameters P_1 , P_2 , s_{11} and s_{12} that result in

$$\left(\frac{P_1}{P_1 + s_{12}P_2 + 1}\right)^2 < \left(\frac{s_{12}P_2 + 1}{P_1 + s_{12}P_2 + 1}\right)^2 sP_1. \text{ In this case the direction of the inequality as well as the}$$

sign of the right side of the equality changes and we are facing the case that $F_S(s) = \Pr(s_{11} \geq \alpha)$,

where α is negative value and the probability of this event is simply equal to unity. Therefore,

denoting \mathbf{R} as the region of s_{12} that satisfies the inequality

$$\left(\frac{P_1}{P_1 + s_{12}P_2 + 1}\right)^2 \geq \left(\frac{s_{12}P_2 + 1}{P_1 + s_{12}P_2 + 1}\right)^2 sP_1 \text{ and } \mathbf{R}' = [0, \infty) - \mathbf{R} \text{ as the region of satisfying the}$$

inequality $\left(\frac{P_1}{P_1 + s_{12}P_2 + 1}\right)^2 < \left(\frac{s_{12}P_2 + 1}{P_1 + s_{12}P_2 + 1}\right)^2 sP_1$, one can derive the equivalent channel CDF as

follows:

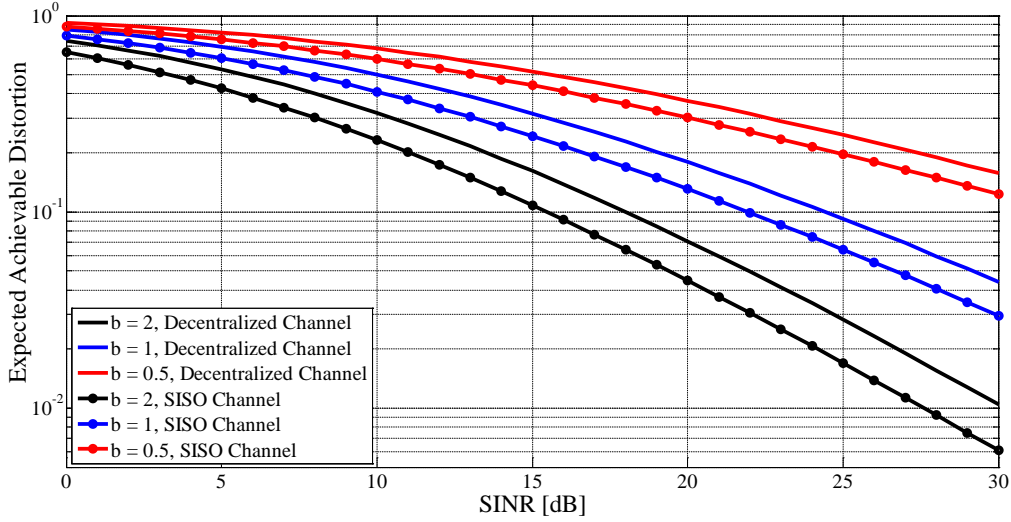


Fig. 2. Comparing the expected achievable distortion of the decentralized channel with that of SISO channel.

$$\begin{aligned}
 F_S(s) &= \int_{\mathbf{R}} \int_0^{\frac{s(s_{12}P_2+1)}{\left(\frac{P_1}{P_1+s_{12}P_2+1}\right)^2 - \left(\frac{s_{12}P_2+1}{P_1+s_{12}P_2+1}\right)^2} sP_1} f_{s_{11}}(s_{11}) f_{s_{12}}(s_{12}) ds_{11} ds_{12} + \int_{\mathbf{R}'} f_{s_{12}}(s_{12}) ds_{12} \\
 &= \int_{\mathbf{R}} e^{-s_{12}} \times \left(1 - e^{\frac{-s(s_{12}P_2+1)}{\left(\frac{P_1}{P_1+s_{12}P_2+1}\right)^2 - \left(\frac{s_{12}P_2+1}{P_1+s_{12}P_2+1}\right)^2} sP_1} \right) ds_{12} + \int_{\mathbf{R}'} e^{-s_{12}} ds_{12} \\
 &= \int_{\mathbf{R}} e^{-s_{12}} \times \left(1 - e^{\frac{-s(s_{12}P_2+1)(P_1+s_{12}P_2+1)^2}{P_1^2 - sP_1(s_{12}P_2+1)^2}} \right) ds_{12} + \int_{\mathbf{R}'} e^{-s_{12}} ds_{12}
 \end{aligned} \tag{25}$$

Using the above equation, the CDF of the channel gain can be derived numerically and the corresponding channel pdf can be derived, afterwards. Furthermore, the optimal assistant function is derived numerically and after finding the start and the end point of the power allocation function, the expected achievable distortion is calculated.

IV. NUMERICAL RESULTS

In this section, numerical examples are provided to demonstrate the performance of the proposed method in a Rayleigh block-fading environment.

Fig. 2, compares the expected distortion of a SISO network in the Rayleigh fading environment with the same performance metric in the considered decentralized channel at the same SINR. As it can be seen, there is a constant SNR loss which was predictable, because the equivalent channel gain in the decentralized format deviates from the Gaussian form.

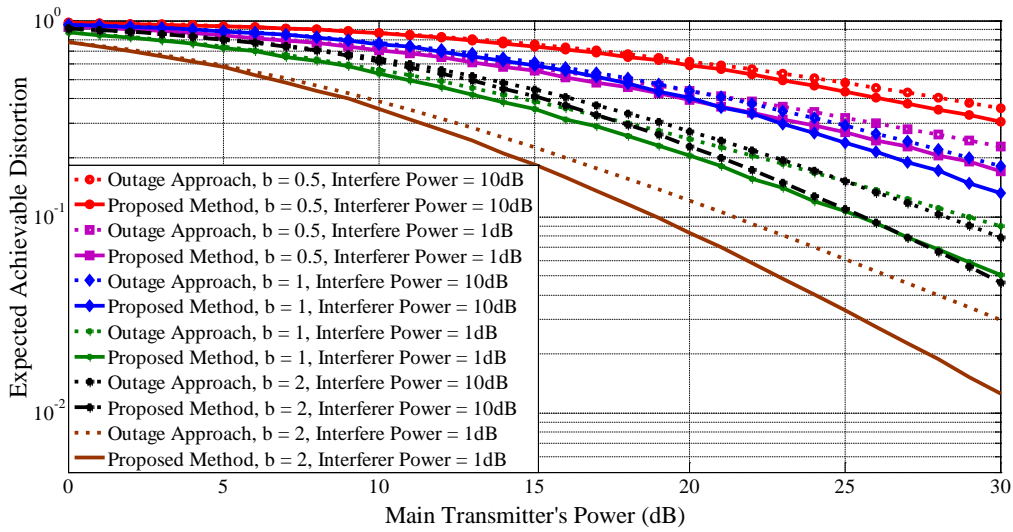


Fig. 3. Expected distortion minimization using the proposed method and the conventional outage approach. The interferer's power is fixed and the intended transmitter's power is swept.

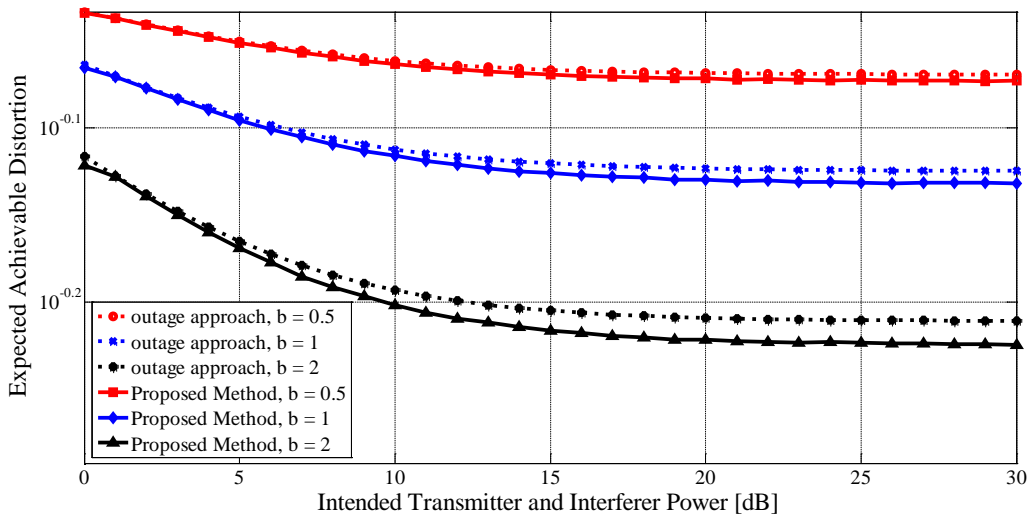


Fig. 4. Expected distortion minimization using the proposed method and the conventional outage approach. The interferer's power as well as the intended transmitter's power are swept.

In Fig.3 and Fig.4 the comparison of the proposed method with the outage approach is provided. In the outage approach the transmitter uses a single level code, e.g., $R_o = \log(1 + s_{th}P)$ to send the information to the destination and if the channel gain falls below s_{th} , nothing can be decoded. In this case, the average achievable distortion, which obviously is a function of s_{th} , can be formulated as,

$$D_{ave}(s_{th}) = F_S(s_{th}) \times 1 + (1 - F_S(s_{th})) \times (1 + s_{th}P)^{-b} \tag{26}$$

The best performance of the outage approach comes from tuning the s_{th} into its optimum value, so that $D_{ave}(s_{th})$ sticks to its minimum.

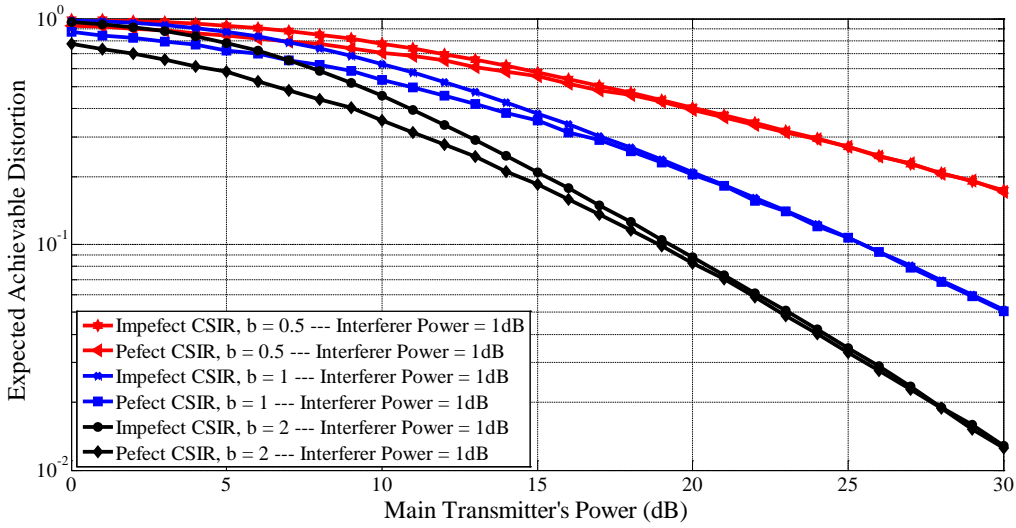


Fig. 5. Comparing the Imperfect channel estimation with the perfect case in deriving the minimum expected distortion of a decentralized channel.

Fig. 3 compares the proposed multi-layer approach to the conventional outage approach. The numerical results are provided for sweeping the intended transmitter's power from 0dB to 30dB and the interferer's power equal to 1dB or 10dB. It is shown that the proposed method outperforms the outage approach for all mismatch factors and interferer's powers, however, it is more advantageous in greater mismatch factors and transmit powers, compared to the interferer. As an example, an SNR gain of around 5 dBs is gained for $b = 2$, when reaching to the average distortion level of 2×10^{-2} is required.

The performance of the proposed method is compared with the outage approach in an interference limited case in Fig. 4, by sweeping the intended transmitter's power as well as the interferer's power. It can be seen that the advantage of using multi-layer coding approach increases for greater mismatch factors, however, increasing the intended transmitter's power does not lead to a viable decrease in expected distortion in the interference limited case.

The Fig. 5 compares the perfect CSIR scenario with the imperfect CSIR case. As it can be seen, the MMSE channel estimation, used in the imperfect CSIR case, degrades the performance of the proposed method at low SNRs, however, the performance degradation due to the imperfect channel estimation vanishes when coming to high SNRs.

V. CONCLUSION

We considered a decentralized interference channel, where two transmitter-receiver pairs attempt to communicate each other in a block-fading environment. It is assumed that the CSI associated with the direct and cross links are not available in the transmitters. Accordingly, we proposed a multi-layer coding approach that can be coupled with hierarchical representation of a Gaussian source, in order to

minimize the expected achievable distortion of the network at the destination, under two scenarios of having and having not access to the perfect channel state information at the receiver. In this case, the optimal assistant function for power allocation as well as the start and end point of power allocation interval are addressed. Moreover, it is proved that the provided solution in this paper is optimal in the Rayleigh block-fading case for the perfect CSIR scenario, however, deriving the relations in the imperfect CSIR scenario is not possible. Finally, simulation results approved that the proposed method outperforms the conventional outage approach in terms of minimizing the expected distortion.

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