Bachelor Term Project

Phase 3

Configuring Passive Intelligent Surfaces for maximizing data-rates in wireless communication

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Abstract

Wireless Communication is ever improving with the 5G based wireless communication being now being deployed in a lot many geographies. With the advent of beamforming technology and MIMO based communication, concerns over power consumption, security and logistical issues are bound to be raised. Many solutions have come forward for what is labelled as “5G and beyond” stack, one such solution is the use of reflecting surfaces which have reflecting elements which can inflict a phase and magnitude shift on the incoming wave and reflect it without the use of any active components.
Passive Intelligent Surfaces (PIS) or Intelligent Reflecting Surfaces (IRS) as they are sometimes called are an up and coming advancement to the ever improving field of wireless communication. PIS have a wide variety of applications including but not limited to unclogging dead zones, enhancing physical layer security and enhancing the power of the received signal. There are a wide variety of problems related to the development, deployment and operation of such a surface but this thesis concerns with the specific class of problems related to setting the reflection coefficients of the IRS elements such that the data rates for information users are maximised in linear time.

Contents

1 Introduction

5G and beyond networks come with more demands than their predecessor, like improving energy efficiency, decreasing monetary costs, higher reliability and domineeringly lower latency. However, emerging solutions to 5G services (e.g. ultra-reliable and low latency communication (URLLC)) include an increasing number of active nodes, packing more antennas and migrating to higher frequencies involves increasing energy, hardware and cost requirements. Hence this need for cheaper energy-efficient smart solutions has led to the conceptualization of Passive Intelligent Surfaces(PISs) (also known as Intelligent Reflecting Surfaces (IRSs). These are surfaces (not necessarily flat) which reflect signals from base station to the user and vice-versa. In addition to reflecting they are also capable of changing the phase and magnitude of the signal and are configurable by a controller.

Formally, PIS is a re-configurable environment for Beyond 5G wireless communication systems constituting passive elements which reflect incoming signals and additionally inflict a controllable phase and magnitude change to them [1].

PIS reflection surface is realizable using the existing programmable meta surfaces [2]. These reflection elements are tuned by the help of a controller connected to the base station, and Micro Electro Mechanical Switches (MEMS) are used to control the reflection coefficients. This thesis does not delve into the physical implementation any more than this but details can be found in [3].

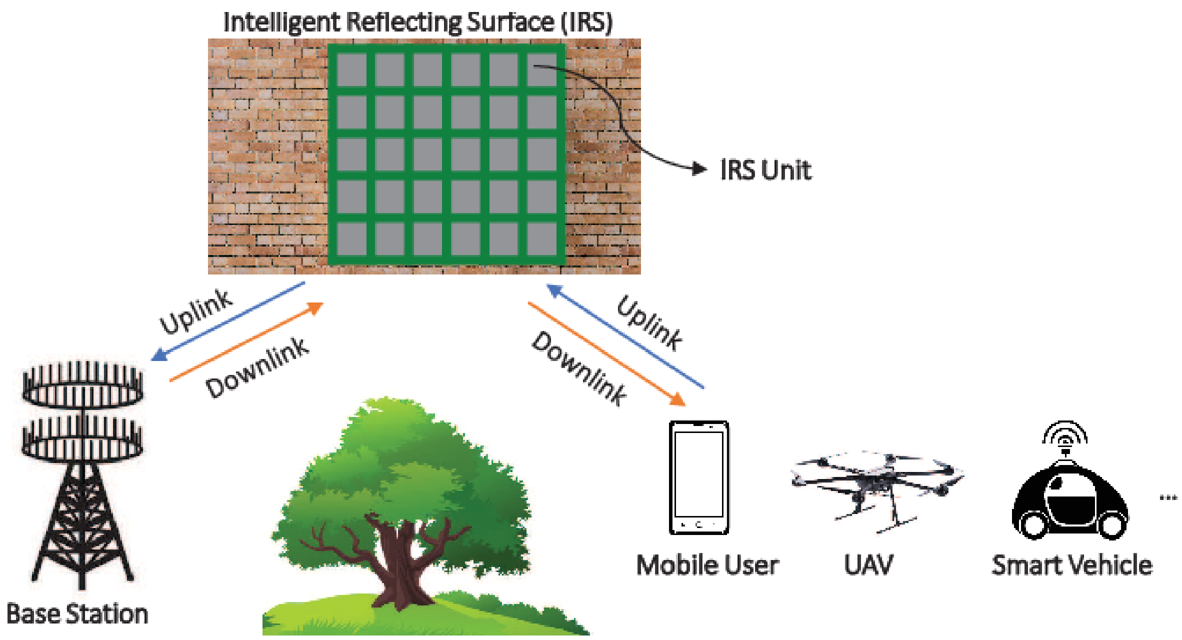

Figure 1: Basic setup of how Passive Intelligent Surfaces or Intelligent Reflecting Surfaces are deployed and work. Image Credits: Figure 1 of [4]

2 Literature Review

PIS has a variety of different use cases been explored in the areas of Wireless Communication and Energy Transfer as surveyed in [5]. Configuring a PIS to maximize energy recieved or data-rates requires information about the channel coefficients [6]. Estimation can be done in both the frequency and time domain, and using different sequences of PIS elements is required, as was studied in the previous phase of this project and also in [7].

Once estimated, the reflection coefficients need to be tuned to meet the optimization objective. This could be done as a joint beamforming optimization where both active beamforming at the Base Station and passive beamforming by setting coefficients of PIS, similar to [8]. The phase shifts of the elements need not be discrete and could be based in a practical setting as studied in [9] and [10]. [10] suggested iterative alternating optimization algorithms to optimize the SINR of the recieved signal.

There has been a lot of research on channel estimation, passive beamforming and placement of IRS in a network in recent years as is summarized in [11]. This current work concerns with passive beamforming at the IRS. Passive beamforming is tuning the reflection coefficients of the IRS elements to obtain a particular objective. The objective can be to optimize datarates [12], to cancel out signals to eavesdroppers [13] or to optimally transfer energy [6] etc. For this paper, we are concerned with optimizing the achievable datarates. There has been extensive research on passive beamforming for optimizing datarates on a single subcarrier, on both continous [14] and discrete [15] phase shifts of IRS elements. Different methods have been suggested for passive beamforming in order to achieve optimal datarates. Research [16] also deal with jointly doing active and passive beamforming using an Semidefinite Relaxation(SDR) based approach to optimize an approximate version of the datarate expression. There has also been research [13] on jointly optimizing the IRS Coefficients and the power allocation to different users. More recently Gradient Projection and Cross Entropy based algorithms like the ones in [17], [18] and [19] have been suggested as more efficient solutions to directly using conventional optimization methods. What has been a relatively nascent research area is optimizing phase shifts in an OFDM based setup. While some techniques have been looked at in a continous phase shift based IRS [16] [20], a few recent papers have looked at discrete phase shifts [21], [22] as well. But these techniques suffer on two counts, the continous phase shifts don't capture the practicalities of deployment and for methods on discrete phase shifts, the methods are either based on only binary reflection elements or are computationally complex for very large number of IRS [23].

State of the Art and Novel Contributions:

Practical models involving PIS aided MIMO-OFDM based communication have been explored in [22] recently, which go on to show that solving the optimization problem of maximizing the data-rates has non-convex constraints, which makes it tough to solve. Some techniques like Successive Convex Approximation [24] and Block Coordinate Descent [22] have been used to find an approximate solutions but such approaches have non linear time complexity.

These methods while do help achieve the maximum have a high computational penalty, something which not be desirable since IRS are modelled to be deployed at the edge of the network where computing power might not be abundant. It is also desirable that the time taken to beamform to be as low as possible since the total time for training and beamforming might be limited and this helps gain more time for training. Having low time is also desirable for situations where the user is mobile.

In this paper, we propose a heuristic algorithm to obtain the optimal IRS phase-shifts for every sub-carrier in an OFDM system with the objective of maximizing the data rate under discrete phase-shift constraint at the IRS. We show that the proposed algorithm gives near-optimal data rates at lower computational complexity of $\mathrm{O}(NK\log K)$ as opposed to $\mathrm{O}(N^3K^2)$ with SDR/cross entropy based state-of-the-art algorithms for N IRS elements and K subcarriers. We also show that given just 3 reflection bits, our discrete algorithm performs almost equal to the continous counterpart.

Our model is novel in the following aspects: Firstly no work has looked at optimizing the data rates in an OFDM system with respect to passive beamforming in an OFDM system when the phase-shifts at the IRS are constrained to take values from a discrete set. Secondly, this work takes a comprehensive view of the problem at hand by elucidating the performance-complexity trade-off of the proposed algorithm against the state-of-the-art algorithms proposed for IRS with continuous phase-shifts.

Notations: In this thesis vectors are denoted by bold small letters and matrices are given by capital ones. $\mathbb{C}^{M\times N}$ denotes space of $M \times N$ complex valued matrices. For complex vectors v, $||v||$, v^T, v^H and $diag(v)$ each denote the l_2-norm, transpose, conjugate transpose, and diagonal matrix with each diagonal element being the corresponding element in v respectively. Scalars s are denoted by normal italic letters and s^* denotes conjugate. $\odot$ denotes element wise matrix multiplication. $\text{RoundOff}(a, \mathbb{A})$ is a function which approximates a to the nearest value from set $\mathbb{A}$.

3 System Model

As shown in figure [figure number], there are N reflection elements in the IRS setup, with M discrete phase shifts on each IRS element. That is, the i^{th} element can take a phase shift θ_i from the set $\mathbb{A} = \{0, 2\pi\frac{1}{M}, 2\pi\frac{1}{M}, \dots, 2\pi\frac{M-1}{M}\}$. The reflection gains of the IRS elements are taken to be 1. There are K subcarriers and T taps. At the source (AP), the information symbol s_{ik} with $E[|s_{ik}|^2] = 1$ is transmitted with power $p_k = \frac{P}{K}$ to create transmit symbol x_k for the k^{th} subcarrier, where $x_k = p_k s_i$ and P is the total power available at the source. The received symbol y_k at the user on the k^{th} sub-carrier can be written as,

$$y_k = (h_k^{si^H}\ \Phi\ h_k^{iu} + h_k^{su*})\ x_k + z_k \tag{1}$$

where $h_k^{si^H} \in \mathbb{C}^{1\times N}$, $h_k^{iu} \in \mathbb{C}^{N\times 1}$ and $h_k^{su*} \in \mathbb{C}$ denote the channel coefficients for the Source-IRS Channel, the IRS-User Channel and the direct channel for the k^{th} subcarrier. $\Phi = \text{diag}([e^{j\theta_1}, e^{j\theta_2}, \dots, e^{j\theta_N}])$ denotes the IRS reflection coefficients where $\theta_i \in \mathbb{A}$ is the phase shift. The term ${h_k^{si}}^H \ \Phi \ h_k^{iu} + h_k^{su*}$ is denoted by $h_\Theta[k]$ for the remaining paper and is the channel coefficient for the cascaded plus the direct channel. z_k is CWGN with power spectral density N_0. The received symbols at different sub-carriers can be vectorised to $y^T = [y_1, y_2, \dots, y_K]$ using the vector transmit symbol x, where both y and x belong to $\mathbb{C}^{K\times 1}$. The updated equation for y,

$$y = ({H^{si}}^H \odot {H^{iu}}^H \Phi + h^{su*})x + z \tag{2}$$

where ${H^{si}}^H \in \mathbb{C}^{K\times N}$,${H^{iu}}^H \in \mathbb{C}^{K\times N}$ and $h^{su*} \in \mathbb{C}^{K\times 1}$ denote the channel coefficients for the BS-IRS Channel, the IRS-User Channel and the direct channel. $z \in \mathbb{C}^{K\times 1}$ is CWGN with power spectral density N_0. . For the purpose of discussion and problem formulation, Θ is a $N \times 1$ vector whose elements are the reflection angles of the IRS elements and belong to $\mathbb{A}$, i.e. $\Theta^T = [\theta_1, \theta_2 \dots \theta_N]$ $\theta_i \in \mathbb{A}$ $\forall i$

4 Problem Description

The main objective that we are working on is the optimizing the datarates by passive beamforming of a suitable IRS configuration (i.e $\theta = [\theta_1, \theta_2, ...\theta_N]$ for all N IRS elements). The data rate for a particular IRS configuration is computed as

$$R = \frac{B}{K+T-1} \sum_{k=0}^{K-1} \log_2 \left(1 + \frac{p_k |h_\Theta[k]|^2}{BN_o}\right) \tag{3}$$

where B is the Bandwidth(symbol rate), $h_\theta[k] \in \mathbb{C}^{1\times(N+1)}$ is a vector of effective channel coefficients of all the channels (direct channel and the N channels through IRS) for k^{th} sub-carrier and N_o is the Noise Power Spectral Density.

Mathematically, the optimization problem can be stated as

$$\begin{aligned} \max_{\Theta} \quad & R \\ \text{s.t.} \quad & \theta_i \in \mathbb{A} \ \forall \ \theta_i \in \Theta \\ & |\theta_i| = 1 \ \forall i \end{aligned} \tag{4}$$

An exhausitive search over the whole space of Θ would require 2^N computations and would prove out to be inefficient when N is in thousands. Therefore there is a need to reduce the amount of computations while almost achieving the optimal datarates. These algorithms would preferably should be of linear time complexity [1]

[1] It can be shown that the minimum possible time complexity can be linear since the channel via each element of the PIS would have to be evaluated unless grouping the PIS elements is an option.

5 Reflection Coefficient Optimization to maximize achievable rates

5.1 Ideal Optimization

Optimizing without the constraint of having discrete phase shift has already been done in [24] and [25]. The result was is that the PIS configuration that attain the optimal data-rates is one where the phase shift induced by the PIS is equal to the difference in phase shift of the direct channel and that of the cascaded channel. In other words, the PIS would be expected to compensate for the delays induced by the rest of the channel. But since deploying continuous phase shifts practically is not possible, PIS elements often have discrete phase shifts. It makes sense to then try and compensate as much as possible for the phase shifts induced. But this approach becomes redundant when the number of possible phase shifts is very low, for example 2. The continuous phase shifts churned out from these algorithms could also be discretized, but these still suffer from complexity issues. The algorithms developed ahead still follow the same underlying philosophy but also exploiting the discrete nature of the phases.

5.2 Optimizing across Multiple Frequencies: OFDM

If we consider the frequency response for the channel to be flat across the bandwidth the Rate becomes,

$$R = \frac{BK}{K+L-1} log_2 (1 + \frac{P|G_\nu \Theta|^2}{BN_o})$$

and maximizing this is equivalent to maximizing just the Channel Coefficient $L2$ Norm. Therefore our equation 4 becomes,

$$\max_{\Theta} |G_\nu \Theta|^2 \tag{5}$$

$$\text{s.t. } \theta_i \in \{0, \pi\} \ \forall i \in \mathbb{N} \tag{6}$$

The algorithms developed here are under this assumption and hence the vector sum of the channel coeffiecients are maximized by optimaly and efficiently selecting the reflection phase shifts. But if the channel is not flat across the bandwidth then one of the approaches as taken by [12], would be to estimate the channel coefficients in the time domain, select some T strongest taps and then optimally selecting the delays of PIS elements using analogous algorithms. Although in [12], the algorithms developed optimally in $O\left(N^3\right)$, we try an devise a log linear algorithm.

5.3 Constraints and Considerations while devising the algorithm

While designing such an algorithm to figure out the optimal phase shift angles, certain considerations are needed to be considered. First of them, as already discussed would be to keep the time complexity to a minimum, preferably linear time, i.e. $O(N)$. The second would be restricting to only discrete phase shift values.

6 Descriptions of Proposed Algorithms

6.0.1 Vector Heuristic

The complex coefficients can be considered as two dimensional vectors. Defining $h_{ref} = max(|h^{siu}|)$ i.e. vector with highest magnitude or the strongest channel. Further, let $\Delta_i = \text{Phase}(\frac{h^{siu}}{h_{ref}})$. For all i, θ_i is set to $\text{RoundOff}(-\Delta_i, \mathbb{A})$. The intuition behind this comes from the passive beam-forming rule of continuous phase shifts where the IRS element exactly compensates the phase difference between direct and cascaded channel. Simply put, all IRS elements would be aligned along the strongest channel (among the IRS elements).

The time complexity of this algorithm is $O(NK)$. Although this does give us performance gains, experimentally as will be shown in results this can be further improved by considering the strongest tap.

Algorithm 1 Vector Heurestic Algorithm for Passive Beamforming

Input: $h_k^{siu}, h_k^{su}\ \forall\ i, k$
Output: Θ
$h^{siu} \leftarrow \frac{1}{K}\sum_{k=0}^{K} h_k^{siu}\ \forall i$
$h_{ref} \leftarrow max(|h^{siu}|)$ among all i and direct channel
$\Theta \leftarrow [0, 0 \ldots 0]_{1\times N}$
for all i in $1, 2 \ldots N$ **do**
  $\Delta_i \leftarrow \text{Phase}(\frac{h^{siu}}{h_{ref}})$
  $\Theta[i] \leftarrow \text{RoundOff}(-\Delta_i, \mathbb{A})$
end for
return Θ

6.0.2 Strongest-Tap Selection

To further improve performance we could consider only the strongest tap of all the N paths via the IRS elements and the direct channel. A similar optimization is performed in [20] and [26] optimization along the strongest taps, but with continous phase shifts. The IDFT for the K channel coefficients for the $N+1$ sub channels is taken and trimmed to first L entries. Therefore for each subchannel, we define:

$$g^{siu} = \text{IDFT}(h^{siu})[1:L]\ \forall\ i \tag{7}$$

$$g^{su} = \text{IDFT}(h^{su})[1:L] \tag{8}$$

Algorithm 2 Strongest Tap Enabled Vector Heurestic Algorithm for Passive Beamforming

Input: $h_k^{siu}, h_k^{su}\ \forall\ i, k$
Output: Θ
$h^{siu} \leftarrow [h_1^{siu}, h_2^{siu} \dots h_k^{siu} \dots h_K^{siu}]\ \forall i$
$h^{su} \leftarrow [h_1^{su}, h_2^{su} \dots h_k^{su} \dots h_K^{su}]$
$H^{siu} \leftarrow IDFT([h^{siu})[1:L]\forall i$
$H^{su} \leftarrow IDFT([h^{su})[1:L]\forall i$
$H_{\text{strong}}^{siu} \leftarrow max(H^{siu})\forall i$
$H_{\text{strong}}^{su} \leftarrow max(H^{su})\forall i$
$H_{ref} \leftarrow max(|H_{\text{strong}}^{siu}|)$ among all i and direct channel
$\Theta \leftarrow [0, 0 \dots 0]_{1\times N}$
for all i in $1, 2 \dots N$ **do**
  $\Delta_i \leftarrow \text{Phase}(\frac{H_{\text{strong}}^{siu}}{H_{ref}})$
  $\Theta[i] \leftarrow \text{RoundOff}(-\Delta_i, \mathbb{A})$
end for
return Θ

The strongest entry out of these L is considered for each N IRS elements and these are used and processed to optimize for the passive beamforming.

$$g_{\text{strong}}^{siu} = \max(g^{siu})\ \forall\ i \tag{9}$$

$$g_{\text{strong}}^{su} = \max(g^{su}) \tag{10}$$

A DFT is then taken of these coefficients, to obtain the frequency domain equivalents.

$$h_{\text{strong}}^{siu} = \text{DFT}(g_{\text{strong}}^{siu})\ \forall\ i \tag{11}$$

$$h_{\text{strong}}^{su} = \text{DFT}(g_{\text{strong}}^{su}) \tag{12}$$

We use these $N+1$ coefficients, i.e. $h_{\text{strong}}^{siu}\ \forall\ i$ and h_{strong}^{su} as input for Algorithm 1 to obtain the optimal N IRS coefficients. The time complexity of the resulting algorithm is $O(Nlog(N) + NK)$.

7 Numerical and Simulation Results

7.1 Setup

In this section, the algorithms proposed are bench-marked against the existing state of the art solutions both in terms of rate maximization optimality and computation time performance or complexity. As highlighted above this is a single antenna base station and single user setup with an IRS stationed between the BS and user. The IRS is stationed at the origin (0,0,0) and the base station is at (0,-10,0), making the distance between IRS and BS to be fixed at 10m. The user is kept at (0,y_u,-2). This is portrayed in Figure 1. For our simulations we are considering a system with a default number of sub-carriers as

500, number of IRS elements as 64, number of taps between different links as 20, and the number of IRS bits as 1. These parameters are varied from their base values to see the impact on the performance metrics. The direct link between BS and User, the BS-IRS link and IRS-User links are modelled as Rician Channels with different parameters. The Path loss coefficient for the direct link is taken to be 3.6 and for IRS channel is taken to 2.2 [2] The carrier frequency is taken to be 3.5 Ghz, hence the wavelength λ is 0.0857 m. The separation between two IRS elements is taken to be $\lambda/4$. In [Rhui Zhang Paper Ref], an successive convex approximation (SCA) based algorithm has been proposed with initialization using Semi Definite Programming. The number of iterations required for the SCA to converge depends on the values that it is initialized with. This algorithm although suboptimal has been used in different studies, we use this as the benchmark (without any discretization).

Results are derived for three cases, changing the number of IRS Elements, the number of sub-carriers and the number of discrete IRS bits. Figure [2] displays the effect of changing the number of IRS elements on the data rates for the different distance scenarios and time taken for the algorithm to run. Figure [3] shows the effect of the number of sub-carriers with similar axes for data rates and time taken. Figure [4] and [5] are plotted while varying both the user locations and the number of discrete bits on the time taken and capacity achieved.

7.2 Discussion

Figure [2] shows an expected trend that increasing the number of IRS elements increase the performance of the data rate. A more interesting trend is how close a 2 bit IRS can get to the SCA based setup. Looking at the corresponding time performance we see a stark improvement in the number of complex computations in comparison to the benchmark. Our algorithm increase with the order of $\mathrm{O}(N)$ with increase in IRS elements and SCA based algorithm increases with $\mathrm{O}(K^{4.5}N^{3.5})$. Figure [3] shows a similar trend as well where increasing the number of sub-carrier increases the performance, and the 2 bit vector based approach is similar to the continuous SCA based approach, and the time performance is much better for the vector based approach. Finally, reviewing Figure [5] we see that as we increase the number of bits the performance of algorithm increases, reaches the SCA based and even surpasses it to reach what may be considered the optimal. This also shows that 2 to 3 bit IRS elements should be enough to give as good a data rate performance as the continuous SCA based approach.

[2]Results are derived for 1000 iterations on an Intel i5 8th Gen Processor with 16 GB memory.

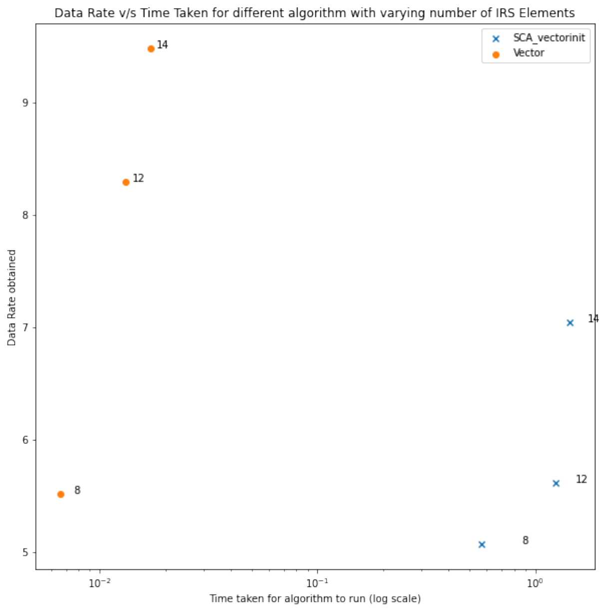

Figure 2: How our algorithm performs in comparision to the SCA algorithm with changing number of IRS elements

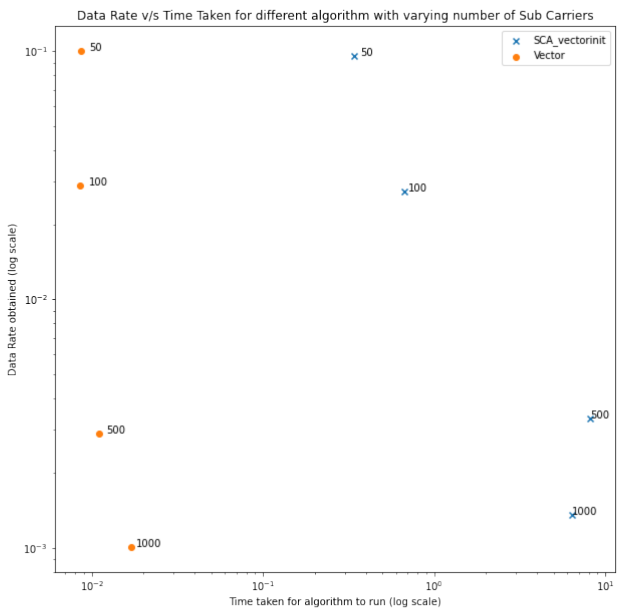

Figure 3: How our algorithm performs in comparision to the SCA algorithm with changing number of Subcarriers

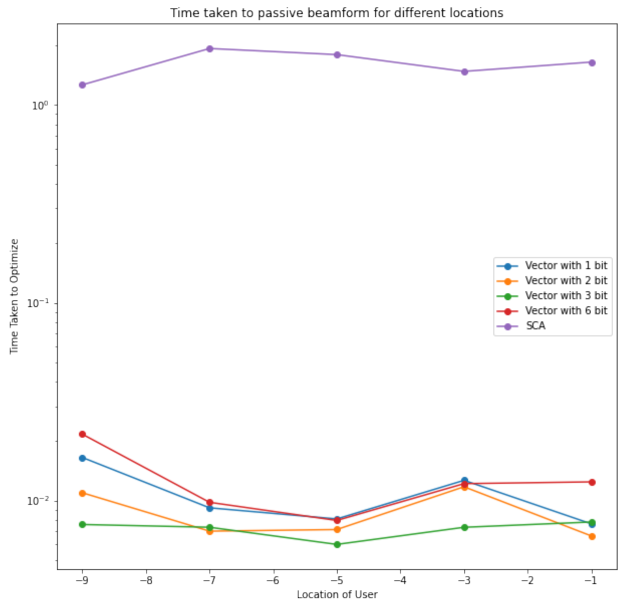

Figure 4: Time taken for SCA and our method for different user locations

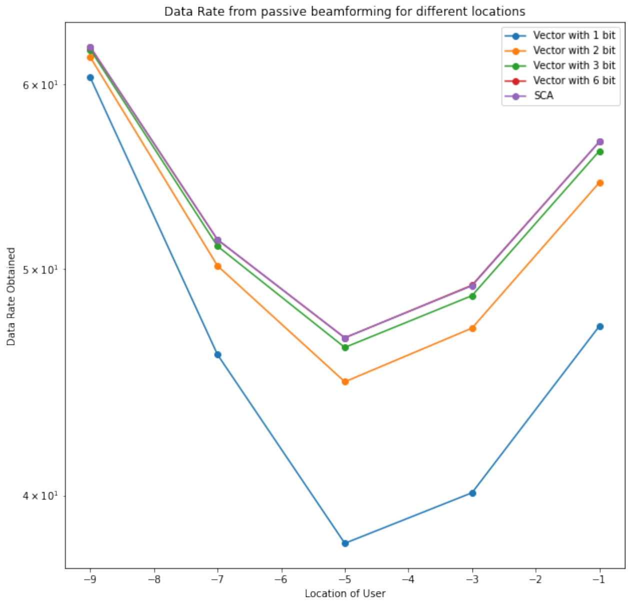

Figure 5: Datarates for SCA and our method for different user locations

8 Discussions and Future Work

Results show that our simpler algorithm for passive beamforming using discrete phase shifts in an OFDM based IRS aided wireless setup is comparable in data rate performance and has a significant advantage in terms of the time performance. This gain in time performance makes it a better choice, when more time for training is needed or when the user is mobile and optimal IRS phase shifts need to be determined again and again. This particular setting with discrete IRS elements of different number of bits has not been studied earlier and is of practical consequence, since while deployment IRS elements will end up having discrete phase shifts. Our method is also not dependent on any initialization unlike many methods like the SCA, whose actual number of iterations depend on the different initialization. Extension to a joint optimization where power allocation across different subcarriers is fairly straightforward as highlighted in previous research using alternating optimization. Although these results give an optimistic outlook, further work is needed to make this applicable, including looking into a mutli-user based setup and MIMO based transmission. These additions would definitely complicate the optimization equation but the strongest tap heuristic technique still might prove useful and is up for future research.

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