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Blind Beamforming for Multiple-IRS Assisted Wireless Transmission

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***Abstract*—Conventional beamforming methods for intelligent reflecting surfaces (IRSs) typically entail the full channel state information (CSI). However, channel acquisition is costly when IRSs are extensively deployed in the network. To overcome this difficulty, this paper proposes a novel strategy called blind beamforming that coordinates multiple IRSs by means of the received signal statistics without knowing CSI. Blind beamforming just requires measuring the received signal power at the user side with respect to a sequence of randomly generated phase shifts across all the IRSs. Its main idea is to extract the essential statistical information for beamforming by exploring only a small portion of the whole solution space of phase shifts. We show that blind beamforming guarantees a signal-to-noise ratio (SNR) boost of $\Theta(N^{2L})$ under certain conditions, where L is the number of IRSs and N is the number of reflecting elements per IRS. The above result significantly improves upon the previous studies (including those with CSI) on multiple-IRS-assisted communication. Furthermore, we demonstrate the effectiveness of the proposed blind beamforming method through field tests at the commercial spectrum band of 2.6 GHz.**

***Index Terms*—Intelligent reflecting surface (IRS), reconfigurable intelligent surface (RIS), multi-IRS/RIS systems, blind beamforming without channel state information (CSI).**

I. Introduction

Intelligent reflecting surface (IRS) is an emerging wireless network device that aims to improve the wireless environment by manipulating signal reflections [2], [3]. As compared to small base-station and relay, IRS enjoys the advantages of much lower price and much lower energy consumption, thus providing a practical solution to the throughput, coverage, and reliability requirements of future networks. While the early studies about IRS are focused on a single IRS, many recent works take multiple IRSs into account [4], [5]. Because the existing methods for multi-IRS coordination typically require full channel state information (CSI), their implementations become intractable when IRSs are extensively deployed in the network. To overcome this technical difficulty, we propose a novel strategy called blind beamforming that optimizes phase shifts across multiple IRSs in the absence of CSI.

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Our approach stems from two recent works [6], [7] that demonstrate the potential of optimizing phase shifts blindly for a single IRS without CSI. Their proposed schemes try out a small subset $\mathcal{S}$ of the whole phase shift solution space Ω at random, from which a statistical quantity (e.g., the conditional sample mean) of the received signal power is obtained and then is used to decide phase shifts. The resulting solution is not limited to the test set $\mathcal{S}$. This work seeks a generalized blind beamforming strategy for multiple IRSs.

Due to the fact that the number of channels grows exponentially with the number of IRSs, channel estimation is practical only for some simple cases, e.g., when there are two IRSs [8] or when the multi-hop reflected channels are all neglected [9]. Aside from its high computational complexity, channel estimation is also difficult to implement in practice because the current network protocol does not provide sufficient support. Actually, the existing prototype realizations of IRS rarely involve channel estimation [6], [10], [11].

Even with the perfect CSI available, optimizing phase shifts for multiple IRSs is still challenging, because every multi-hop reflected channel is incident to more than one reflecting element and hence their phase shifts must be coordinated judiciously. To bypass this difficulty, many existing works [12]–[16] approximate the beamforming problem by ignoring all the multi-hop channels. But this approximation can undermine the capability of multi-IRS coordination. Without multi-hop channels, multiple IRSs are conceptually the same as one single IRS, and hence the signal-to-noise ratio (SNR) boost is limited by $\Theta(L^2N^2)$ from above according to [7], where L is the number of IRSs and N is the number of reflective elements (REs) of each IRS. In contrast, this work shows that the highest possible SNR boost of $\Theta(N^{2L})$ can be achieved by blind beamforming which harnesses the multi-hop reflections properly rather than ignoring them. This highest possible SNR boost is also established in [17] but under a much stronger assumption—only the longest reflection is nonzero. This work is based on a more general setting without any zero approximation.

Throughout the paper, $f(n)=O(g(n))$ if there exists some $c>0$ such that $|f(n)|\le cg(n)$ for n sufficiently large; $f(n)=\Omega(g(n))$ if there exists some $c>0$ such that $f(n)\ge cg(n)$

for n sufficiently large; $f(n) = \Theta(g(n))$ if $f(n) = O(g(n))$ and $f(n) = \Omega(g(n))$ both hold. Moreover, we denote by $\angle x$ the phase of a complex number $x \in \mathbb{C}$, and denote by $[a : b]$ the discrete set $\{a, a+1, \ldots, b-1, b\}$ for two integers $a < b$.

II. System Model

Consider a point-to-point wireless transmission in aid of $L \geq 2$ IRSs. Assume that each IRS $\ell \in [1 : L]$ has N REs. We use $n_\ell \in [1 : N]$ to index the REs of IRS ℓ. Let $\theta_{n_\ell} \in [0, 2\pi)$ be the phase shift induced by RE n_ℓ into its associated reflected channels. The practical implementation of IRS requires that each θ_{n_ℓ} can only take on values from the following uniform discrete set

$$\Phi_K = \{\omega, 2\omega, \ldots, K\omega\} \;\text{ where }\; \omega = \frac{2\pi}{K}$$

given a positive integer $K \geq 2$. We denote by $h_{n_1,\ldots,n_L}$ the cascaded reflected channel associated with the REs $(n_1, n_2, \ldots, n_L)$. If a reflected channel $h_{n_1,\ldots,n_L}$ is not related to any RE of IRS ℓ, then we set $n_\ell = 0$. For instance, considering a double-IRS system with $L = 2$, $h_{3,0}$ is the single-hop reflected channel only associated with RE 3 of IRS 1, while $h_{3,6}$ is the two-hop reflected channel associated with RE 3 of IRS 1 and also with RE 6 of IRS 2. In particular, the direct channel from the transmitter to the receiver is written as $h_{0,\ldots,0}$ since it is not related to any IRS.

The received signal $Y \in \mathbb{C}$ is given by

$$Y = \sum_{(n_1,\ldots,n_L)\in[0:N]^L} h_{n_1,\ldots,n_L} e^{j\sum_{\ell=1}^{L}\theta_{n_\ell}} X + Z, \quad (1)$$

where the transmit signal $X \in \mathbb{C}$ satisfies the power constraint $\mathbb{E}[|X|^2] = P$ and the additive noise $Z \in \mathbb{C}$ has the complex Gaussian distribution $\mathcal{CN}(0, \sigma^2)$. In (1), we let $\theta_{n_\ell} = 0$ whenever $n_\ell = 0$. The received SNR can be computed as

$$\mathrm{SNR} = \left| \sum_{(n_1,\ldots,n_L)\in[0:N]^L} h_{n_1,\ldots,n_L} e^{j\sum_{\ell=1}^{L}\theta_{n_\ell}} \right|^2 \frac{P}{\sigma^2}. \quad (2)$$

This paper aims to evaluate the performance gain reaped from the multi-IRS coordination. We take the SNR without the aid of IRSs as the baseline:

$$\mathrm{SNR}_0 = |h_{0,\ldots,0}|^2 \frac{P}{\sigma^2}. \quad (3)$$

We seek the optimal tuple of phase shifts that maximize the gain over the baseline SNR, i.e.,

$$\underset{\{\theta_{n_\ell}\}}{\text{maximize}} \quad \frac{\mathrm{SNR}}{\mathrm{SNR}_0} \quad (4a)$$

$$\text{subject to} \quad \theta_{n_\ell} \in \Phi_K,\ \forall n_\ell. \quad (4b)$$

We are faced with two challenges. First, the optimization variables are discrete. Second, none of the channels $\{h_{n_1,\ldots,n_L}\}$ is available.

III. Blind Beamforming

We begin by illustrating the motivation behind the proposed blind beamforming strategy. The traditional method for IRS beamforming consists of two stages: it first estimates channels and then optimizes phase shifts. However, channel acquisition does not scale well with problem size due to the fact that the number of channels is exponential in the number of IRSs. Instead of estimating channels separately, one may propose to estimate the channel matrix between every pair of IRSs and then obtain the cascaded channels $\{h_{n_1,\ldots,n_L}\}$ by multiplying the corresponding between-IRS channel matrices together. As a result, the number of channels to estimate now decreases to only $2NL + \binom{L}{2}N^2 = O(N^2L^2)$. However, this method is costly in practice because it requires deploying a signal sensor at each RE for the pilot signal measurement purpose. To address this issue, we suggest getting rid of channel acquisition completely and optimizing phase shifts without any channel information.

A. Preliminary: Single-IRS Case

Before proceeding to the main result, we first look at the single-IRS system and review the so-called *conditional sample mean (CSM)* method in [7]. For ease of notation, we drop ℓ and simply use n to index each RE of the IRS.

In the ideal case, if the phase shift of each RE is continuous, i.e., as $K \to \infty$, then we can readily show that the optimal solution to (4) is given by $\theta_n^\star = \angle h_0 - \angle h_n$. When K is finite, a natural idea is to round the relaxed solution to the closest point in the discrete set Φ_K, so

$$\theta_n^{\text{CPP}} = \arg\min_{\theta\in\Phi_K} \left|\theta - \theta_n^\star\right| = \arg\min_{\theta\in\Phi_K} \left|\theta + \angle h_n - \angle h_0\right|. \quad (5)$$

namely the *closest point projection (CPP)* method. Clearly, the above method requires CSI.

Nevertheless, as shown in [7], we can somehow recover the solution of CPP by means of statistics without knowing specific channels. This blind beamforming approach works as follows. First, generate a total of T random samples of phase shifts, each denoted by $\boldsymbol{\theta}^{(t)} \triangleq \{\theta_n^{(t)} \text{ for all } n \in [1 : N]\}$, i.e., each $\theta_n^{(t)}$ is generated independently. Next, measure the received signal power $|Y^{(t)}|^2$ at the user terminal with respect to every random sample $\boldsymbol{\theta}^{(t)}$. Denoting by $\mathcal{G}_{n,k} \subseteq [1 : T]$ the set of random samples satisfying $\theta_n^{(t)} = k\omega$, i.e.,

$$\mathcal{G}_{n,k} \triangleq \Big\{ t \in [1 : T] \Big| \theta_n^{(t)} = k\omega \Big\}, \quad (6)$$

we compute the conditional sample mean of $|Y^{(t)}|^2$ within each $\mathcal{G}_{n,k}$ as

$$\widehat{\mathbb{E}}[|Y|^2|\theta_n = k\omega] = \frac{1}{|\mathcal{G}_{n,k}|} \sum_{t\in\mathcal{G}_{n,k}} |Y^{(t)}|^2. \quad (7)$$

Finally, choose each phase shift to maximize the corresponding conditional sample mean of received signal power, i.e.,

$$\theta_n' = \arg\max_{\varphi\in\Phi_K} \widehat{\mathbb{E}}[|Y|^2|\theta_n = \varphi]. \quad (8)$$

The resulting solution $\{\theta_n'\}$ is referred to as the *conditional sample mean (CSM)* method.

To quantify the performance of CSM, we first introduce the

Algorithm 1 Blind Beamforming for L-IRS System

Initialize all the θ_{n_ℓ}'s to zero.
for $\ell = 1, \ldots, L$ **do**
  Generate T random samples $\{\boldsymbol{\theta}^{(t)} | t \in [1:T]\}$ where each $\boldsymbol{\theta}^{(t)} = \{\theta^{(t)}_{n_\ell} | n_\ell = 1, \ldots, N\}$.
  for $t = 1, \ldots, T$ **do**
    Measure the received signal power $|Y^{(t)}|^2$.
  end for
  for $n_\ell = 1, \ldots, N$ **do**
    for $k = 1, \ldots, K$ **do**
      Compute the conditional sample mean in (7).
    end for
    Decide each θ_{n_ℓ} for IRS ℓ according to (8).
  end for
end for

notion of the average-reflection-to-direct-signal ratio as

$$\rho \triangleq \frac{\sum_{n=1}^{N} |h_n|^2}{N \cdot |h_0|^2}. \tag{9}$$

The performance analysis of CSM is stated in the following proposition.

Proposition 1 (Theorem 2 in [7]): The CSM method is equivalent to the CPP method in (5) and yields a quadratic SNR boost in the number of REs in expectation, i.e.,

$$\mathbb{E}\left[\frac{\text{SNR}}{\text{SNR}_0}\right] = \rho \cdot \Theta(N^2), \tag{10}$$

provided that $K \geq 3$ and $T = \Omega(N^2(\log N)^3)$.

B. General L-IRS Case

We now consider L IRSs. The extension of CSM to this multi-IRS case is fairly straightforward: apply CSM to only one IRS at a time while holding the phase shifts of the rest IRSs fixed. The details of this sequential CSM method are summarized in Algorithm 1.

Simple as the above extended CSM method looks, it is by no means trivial to analyze its performance. There is a possible misconception: since CSM can enable an SNR boost of $\Theta(N^2)$ for every IRS as shown in Proposition 1, the accumulated boost would be up to $\Theta(N^{2L})$. However, this argument is problematic because the boost factor ρ in (10) of the previous IRS can be changed dramatically by the configuration of the current IRSs. We illustrate this point through a double-IRS example. Suppose that the channels between the two IRSs are all zeros so that only $h_{0,0}, h_{n_1,0}, h_{0,n_2}$ survive, then the two IRSs can be recognized as one whole IRS, and then the highest possible boost is $\Theta(N^2)$. The reason is that each $h_{0,n_2}e^{j\theta_{n_2}}$ is viewed as the direct channel when optimizing the phase shifts for IRS 1, but later on it can be altered significantly by the phase shift optimization for IRS 2. Thus, the key question is how to preserve the SNR boost of the previous IRSs while optimally configuring the current IRS? The following theorem gives a sufficient answer to the above question.

Theorem 1: If an L-IRS system satisfies the following three conditions:

C1. there exists a set of complex values $\{u^{(\ell)}_{n_l} \in \mathbb{C} \,|\, n \in [1:N],\ \ell \in [1:L]\}$ such that every L-hop channel (which is related to every IRS) can be decomposed as

$$h_{n_1,\ldots,n_L} = \prod_{\ell=1}^{L} u^{(\ell)}_{n_\ell}; \tag{11}$$

C2. the number of phase shift choices $K \geq 2L-1$;

C3. there exists a constant $\gamma \in [0, \frac{\pi}{2(L-1)} - \frac{\pi}{K})$ such that

$$\frac{\sum_{(n_1,\ldots,n_L)\in\mathcal{A}^{(\ell)}_m} |h_{n_1,\ldots,n_L}|}{\prod_{i>\ell} \left|\sum_{n_i=1}^{N} u^{(i)}_{n_i}\right| \cdot \prod_{i<\ell} \left[\sum_{n_i=1}^{N} |u^{(i)}_{n_i}| \cos(\gamma + \frac{\pi}{K})\right]} \leq |u^{(\ell)}_m| \cdot \sin\gamma \tag{12}$$

for $\ell \in [1:L-1]$ and $m \in [1:N]$, where

$$\mathcal{A}^{(\ell)}_m \triangleq \left\{(n_1,\ldots,n_L) \,\middle|\, n_\ell = m, \prod_{\ell=1}^{L} n_\ell = 0\right\}, \tag{13}$$

then Algorithm 1 yields the following SNR boost:

$$\mathbb{E}\left[\frac{\text{SNR}}{\text{SNR}_0}\right] = \frac{\prod_{\ell=1}^{L} \delta_\ell^2}{|h_{0,\ldots,0}|^2} \cdot \Theta(N^{2L}), \tag{14}$$

where

$$\delta_\ell \triangleq \frac{1}{N} \sum_{n_\ell=1}^{N} |u^{(\ell)}_{n_\ell}|, \quad \forall \ell \in [1:L]. \tag{15}$$

It is highly nontrivial to establish the result in the above theorem. We only sketch the proof here due to the page limit; the complete proof can be found in [1].

Proof Sketch of Theorem 1: Our goal is to show that the previous $L-1$ IRSs can still jointly achieve a $\Theta(N^{2(L-1)})$ SNR boost after the last IRS L has been optimized, i.e.,

$$\left|\sum_{(n_1,\ldots,n_{L-1})\in[0:N]^{L-1}} h_{n_1,\ldots,n_L} e^{j\sum_{i=1}^{L-1}\theta'_{n_i}}\right| = \Theta(N^{L-1}) \tag{16}$$

for each $n_L \in [1:N]$ assuming that θ'_{n_i} is obtained from Algorithm 1. We use Q as the shorthand for the left-hand side of (16), which can be approximately bounded from below as

$$\begin{aligned} Q &\approx \left|\sum_{(n_1,\ldots,n_L)\in\mathcal{E}^{(L)}_{n_L}} h_{n_1,\ldots,n_L} e^{j\sum_{i=1}^{L-1}\theta'_{n_i}}\right| \\ &\overset{(a)}{\geq} |u^{(L)}_{n_L}| \sum_{(n_1,\ldots,n_L)\in\mathcal{E}^{(L)}_{n_L}} \cos\left[(L-1)\left(\gamma + \frac{\pi}{K}\right)\right] \prod_{\ell=1}^{L-1} |u^{(\ell)}_{n_\ell}| \\ &\overset{(b)}{=} |u^{(L)}_{n_L}| \cdot \cos\left[(L-1)\left(\gamma + \frac{\pi}{K}\right)\right] \cdot N^{L-1} \cdot \prod_{i=1}^{L-1} \delta_i \end{aligned} \tag{17}$$

where $\mathcal{E}^{(L)}_{n_L}$ denotes the set of channels related to RE n_L and

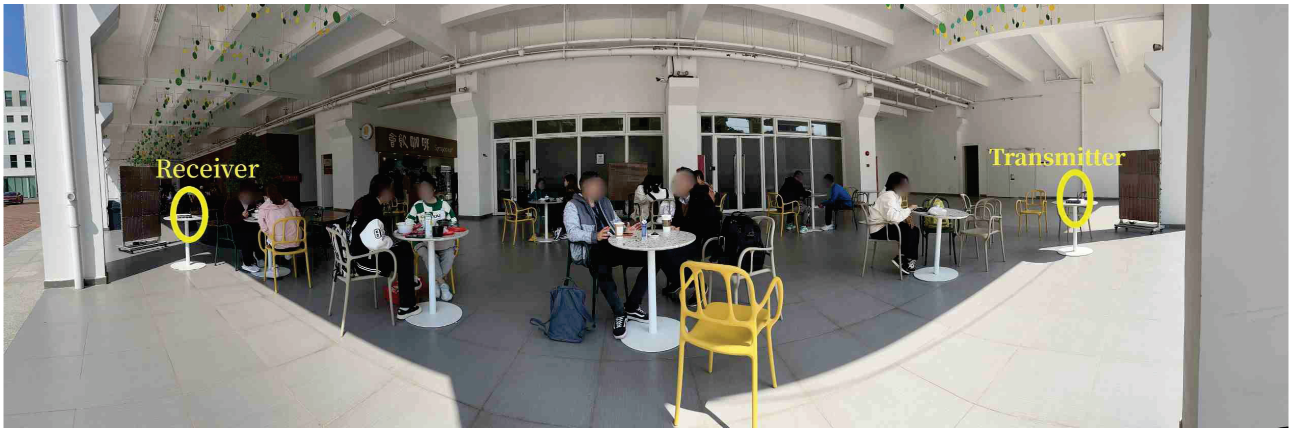

Fig. 1. Field test with three IRSs deployed alongside an open café.

all the rest $L-1$ IRSs, i.e.,

$$\mathcal{E}_{n_L}^{(L)} \triangleq \left\{(m_1,\ldots,m_{L-1},m_L)\middle| m_L = n_L, \prod_{i\in[1:L-1]} m_i \neq 0\right\}. \tag{18}$$

In deriving the above lower bound, step (a) follows by the three conditions in Theorem 1, and step (b) follows by (15). We can then show that $Q = O(N^{L-1})$ and hence verify (16).

Remark 1: Theorem 1 implies that only one round of configuration (i.e., every IRS is optimized one time regardless of L) suffices to attain an SNR boost of $\Theta(N^{2L})$. This is of practical significance when the IRSs are extensively deployed in the network.

C. Comparison to State of the Art: Double-IRS Case

Although we have developed blind beamforming for the fully general case with L IRSs in the former subsection, it is still worthwhile to specialize the result to the double-IRS case with $L=2$ in order to compare our new result with the existing result in [17].

When $L = 2$, it can be shown after some algebra that Theorem 1 reduces to the following proposition.

Proposition 2: If a double-IRS system satisfies the following three conditions:

D1. the channels between the two IRSs are line-of-sight (LoS) so that the two-hop channel matrix has rank one and can be factorized as

$$\begin{bmatrix} h_{1,1} & \cdots & h_{1,N} \\ \vdots & & \vdots \\ h_{N,1} & \cdots & h_{N,N} \end{bmatrix} = \begin{bmatrix} u_1^{(1)} \\ \vdots \\ u_N^{(1)} \end{bmatrix} \begin{bmatrix} u_1^{(2)} & \cdots & u_N^{(2)} \end{bmatrix}; \tag{19}$$

D2. the number of phase shift choices $K \geq 3$;

D3. there exists a constant $\gamma \in [0, \frac{\pi}{2} - \frac{\pi}{K})$ such that

$$|h_{n_1,0}| \leq \sin\gamma \cdot \left|\sum_{n_2=1}^{N} h_{n_1,n_2}\right|, \quad \forall n_1 \in [1:N], \tag{20}$$

then Algorithm 1 yields a quartic SNR boost as

$$\mathbb{E}\left[\frac{\mathrm{SNR}}{\mathrm{SNR}_0}\right] = \frac{\delta_1^2\delta_2^2}{|h_{0,0}|^2}\cdot\Theta(N^4), \tag{21}$$

where

$$\delta_1 \triangleq \frac{1}{N}\sum_{n_1=1}^{N}|u_{n_1}^{(1)}| \quad \text{and} \quad \delta_2 \triangleq \frac{1}{N}\sum_{n_2=1}^{N}|u_{n_2}^{(2)}|. \tag{22}$$

Remark 2: Actually, the state-of-the-art work [17] already establishes the quartic boost for a double-IRS system, but under much stronger assumptions than those in Proposition 2. Specifically, the algorithm in [17] guarantees the quartic boost only when the following conditions all hold:

D1'. the channels between the two IRSs are LoS, same as D1' in Proposition 2;

D2'. $K\to\infty$, namely the continuous beamforming;

D3'. the direct channel and the one-hop reflected channels are all zeros, i.e., $h_{0,0} = h_{n_1,0} = h_{0,n_2} = 0, \forall (n_1,n_2)$.

It is easy to see that D2' and D3' are the special cases of D2 and D3, respectively. Thus, the sufficient condition for achieving the quartic boost in the double-IRS network as stated in Proposition 2 is much more general than that in the existing work [17].

IV. Field Test

This section provides the field test to verify the proposed blind beamforming strategy. We assume that the transmit power is fixed at -5 dBm and the carrier frequency is 2.6 GHz. The following three IRSs are used:

- IRS 1 with 294 REs and 2 phase shift choices $\{0,\pi\}$ for each RE, i.e., $N=294$ and $K=2$;
- IRS 2 also with $N=294$ and $K=2$;
- IRS 3 with $N=64$ and $K=4$.

Notice that our field test does not assume the same values of N and K for IRSs as in the theoretical model in Section II. We consider an outdoor environment where the three IRSs are deployed alongside an open café as shown in Fig. 1. The

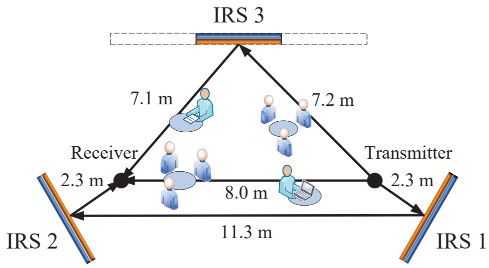

Fig. 2. Layout drawing of the field test. In particular, for Physical Single-IRS, we move IRS 1 and IRS 2 to the positions indicated by the dashed lines.

testbed layout is specified in Fig. 2. The transmission is occasionally blocked by the crowd and also suffers interference which is treated as noise.

The following six methods are compared:

- *Without IRS:* IRS is not used.
- *Zero Phase Shifts:* Fix all phase shifts to be zero.
- *Random Beamforming:* Try out $L \times 1000$ random samples of phase shift vectors and choose the best.
- *Virtual Single-IRS:* Ignore the multi-hop channels and treat multiple IRSs as a single one; optimize phase shifts by the method in [7] with $L \times 1000$ random samples.
- *Physical Single-IRS:* Put multiple IRSs together at the same position to form a single larger IRS; optimize phase shifts by the method in [7] with $L \times 1000$ random samples.
- *Proposed Blind Beamforming:* Coordinate multiple IRSs by Algorithm 1 that uses 1000 random samples per IRS.

TABLE I summarizes the SNR boost performance of these methods, by taking "without IRS" as a baseline. The result of Zero Phase Shifts shows that placing IRSs in the environment already increases SNR by nearly 3 dB even without any optimization. Then a simple heuristic optimization method such as Random Beamforming can reap a higher SNR gain. Observe also that Virtual Single-IRS achieves the highest SNR boost among the four benchmark schemes.

In contrast, the proposed Blind Beamforming further improves SNR by more than 3 dB, as compared to Virtual Single-IRS. This further gain is due to the capability of blind beamforming to take those multi-hop reflections into account. This reason also explains why Physical Single-IRS is worse than most of the other methods. Although its phase shifts have been carefully optimized by the method in [7], its performance is still limited by the deficiency of multi-hop reflections.

V. Conclusion

This work aims at an extension of the blind beamforming strategy to $L \geq 2$ IRSs, thereby achieving a remarkable SNR boost of $\Theta(N^{2L})$ without channel estimation. Field test shows that the proposed method can be efficiently implemented in the real world. Furthermore, we examine the theoretical aspect of this new method and obtain a more general optimality condition than the existing result in the literature.

TABLE I
SNR Boosts Achieved by the Different Methods

Method	SNR Boost (dB)
Zero Phase Shifts	2.91
Random Beamforming	8.48
Virtual Single-IRS	10.80
Physical Single-IRS	7.06
Blind Beamforming	14.09

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