Blind Beamforming for Multiple Intelligent Reflecting Surfaces

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***Abstract*—Channel acquisition is a major challenge faced by the conventional beamforming methods when dealing with multiple intelligent reflecting surfaces (IRSs), because the number of unknown channels grows exponentially with the number of IRSs. This work proposes to sidestep channel estimation and to configure the IRSs blindly based on the statistical information which is extracted from a set of random samples of the received signal power. The proposed blind beamforming method has provable performance in terms of the signal-to-noise ratio (SNR) boost. For instance, it yields a quartic SNR boost of $\Theta(N^4)$ for a double-IRS system under certain condition, where N is the number of reflected elements of each IRS. We remark that the above $\Theta(N^4)$ result is more sophisticated than the existing ones about the double-IRS system in the literature. Furthermore, we numerically demonstrate the advantage of the proposed blind beamforming method through prototype tests with multiple IRSs.**

***Index Terms*—Practical discrete beamforming without channel state information, multiple intelligent reflecting surface (IRSs), quartic signal-to-noise ratio (SNR) boost, prototype tests.**

I. Introduction

Intelligent reflecting surface (IRS) is a low-cost and energy-efficient device (as compared to base-station and relay) that harnesses the signal reflection to improve the data rate, coverage, connectivity, and reliability of wireless networks. In order to capture as many impinging signals as possible from the environment, there has been considerable interest in deploying multiple IRSs [1], [2]. This work advocates a blind beamforming policy for coordinating multiple IRSs in the absence of channel information.

The proposed method stems from a recent discovery [3] which seems at first surprising: the reflected channels can be aligned with the direct channel without knowing any channel state information. While [3] has demonstrated the essential power of the above result in enhancing a single-IRS system, this work goes further to take multiple IRSs into account. We show that blind beamforming enables a quartic signal-to-noise ratio (SNR) boost of $\Theta(N^4)$ for a double-IRS system, where N is the number of reflected elements (REs). The previous work [1] shows the $\Theta(N^4)$ boost under a much stronger assumption; it can be thought of as a special case of the result of this paper. Moreover, we validate the performance gain of the proposed algorithm called sequential conditional sample mean (SCSM) in field tests.

Blind beamforming does not require any channel knowledge. By contrast, the previous studies on the multi-IRS systems typically follow the traditional two-stage paradigm of first estimating channels and then configuring IRSs. However, the complexity of channel acquisition grows exponentially with the number of IRSs. Actually, the state-of-the-art channel estimation methods [4]–[7] are limited to merely two IRSs. As a result, many existing works focus on the beamforming problem for a double-IRS system. Assuming that the phase shifts can be tuned continuously and that there exist only two-hop reflected channels from the transmitter to the receiver, [1] shows that it is optimal to direct the beams of the first IRS toward the other IRS, and consequently an SNR boost of $\Theta(N^4)$ can be attained. Our method SCSM is capable of achieving the same SNR boost under a more general double-IRS setting, yet without requiring channel information. The authors of [8] assume that only the reflected channels (either one-hop or two-hop) exist in a double-IRS system with multiple receivers; an upper bound on the max-min objective is devised to facilitate the alternating optimization of the two IRSs. The above problem is also considered in [9] from a fractional programming perspective. Moreover, [10] utilizes the double-IRS system to prevent eavesdropping by means of majorization-minimization.

When a large number of IRSs are deployed in the system, not only does the exponential growth of the number of channels impede channel acquisition, but it can also render the IRS coordination a challenging task. To find a tractable approximation of the multi-IRS problem, it is common practice in the literature to assume that only a small portion of channels are sufficiently strong while the rest can all be ignored. For instance, [11] assumes that only the direct channel and the longest reflected channel exist, and [12] assumes that only the direct channel, one-hop channels, and two-hop channels exist. The resulting simplified beamforming problems can be addressed by the reinforcement learning [11] and the weighted minimum mean squared error (WMMSE) algorithm [12].

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Another line of studies [13], [14] aim to find the optimal cascaded path out of the survived channels, namely beam routing. In particular, [13] proposes a routing scheme that does not entail channel information, but it requires each IRS to constantly measure the received signal power and report it to base-station. In comparison, the blind beamforming strategy proposed in this paper aims at a fully general multi-IRS setup without ignoring any channels, which does not require each IRS to measure its received signals and also does not require base-station to collaborate.

II. System Model

Consider a point-to-point wireless transmission in aid of $L \geq 2$ IRSs. The transmitter and receiver are equipped with one antenna each. Assume that every IRS consists of N reflective elements. We use $n_\ell = 1, \ldots, N$ to index each RE of the IRS ℓ. Let $\theta_{n_\ell} \in [0, 2\pi)$ be the phase shift induced by the RE n_ℓ of the IRS ℓ into its associated reflected channels. From a practical stand, assume that each θ_{n_ℓ} can only take on values from a discrete set

$$\Phi_K = \{\omega, 2\omega, \ldots, K\omega\} \quad \text{where } \omega = \frac{2\pi}{K} \tag{1}$$

for some prescribed integer $K \geq 2$, namely *passive discrete beamforming*. We use $h_{n_1,\ldots,n_L}$ to denote the cascaded channel associated with the RE n_ℓ of IRS ℓ, for $\ell = 1, \ldots, L$; in particular, we let $n_\ell = 0$ if the cascaded channel is not related to the ℓth IRS, so $h_{0,\ldots,0}$ represents the direct channel from the transmitter to the receiver. Thus, with the transmit signal $X \in \mathbb{C}$ and the additive complex Gaussian noise $Z \sim \mathcal{CN}(0, \sigma^2)$, the received signal $Y \in \mathbb{C}$ is given by

$$Y = \sum_{(n_1,\ldots,n_L)\in\{0,\ldots,N\}^L} h_{n_1,\ldots,n_L} e^{j\sum_{\ell=1}^L \theta_{n_\ell}} X + Z \tag{2}$$

with a dummy variable $\theta_{n_\ell} = 0$ whenever $n_\ell = 0$. In particular, when $L = 2$, we have

$$Y = h_{0,0}X + \sum_{n_1=1}^N h_{n_1,0} e^{j\theta_{n_1}} X + \sum_{n_2=1}^N h_{0,n_2} e^{j\theta_{n_2}} X + \sum_{n_1=1}^N \sum_{n_2=1}^N h_{n_1,n_2} e^{j(\theta_{n_1}+\theta_{n_2})} X + Z, \tag{3}$$

where the first term is the direct signal, the second term is the signal reflected by the first IRS alone, the third term is the signal reflected by the second IRS alone, and the fourth term is the signal sequentially reflected by the two IRSs, as illustrated in Fig. 1.

With the transmit power $\mathbb{E}[|X|^2]$ denoted by P, the received SNR with IRSs can be computed as

$$\mathrm{SNR} = \left| \sum_{(n_1,\ldots,n_L)\in\{0,\ldots,N\}^L} h_{n_1,\ldots,n_L} e^{j\sum_{\ell=1}^L \theta_{n_\ell}} \right|^2 \frac{P}{\sigma^2}. \tag{4}$$

We wish to quantify the benefit brought by IRSs. Toward this end, define the baseline SNR without IRS as

$$\mathrm{SNR}_0 = |h_0|^2 \frac{P}{\sigma^2}. \tag{5}$$

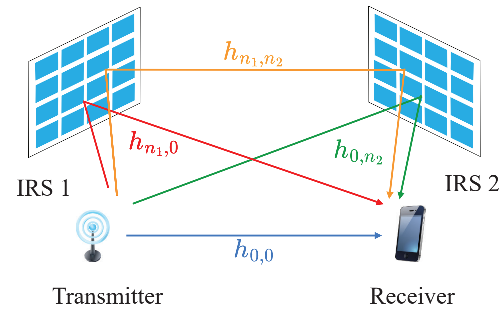

Fig. 1. A double-IRS system with $L = 2$.

This work aims to maximize the *SNR boost*, i.e., the ratio of the improved SNR to the baseline SNR by choosing the phase shifts $\{\theta_{n_\ell}\}$ properly under the discrete constraint Φ_K. The above task can be formally stated as

$$\underset{\{\theta_{n_\ell}:\forall(\ell,n)\}}{\text{maximize}} \quad \frac{\mathrm{SNR}}{\mathrm{SNR}_0} \tag{6a}$$

$$\text{subject to} \quad \theta_{n_\ell} \in \Phi_K, \ \forall(\ell, n). \tag{6b}$$

We remark that the channels $\{h_{n_1,\ldots,n_L}\}$ in the above problem are unknown *a priori*.

III. Blind Beamforming for Multiple IRSs

We first motivate the blind beamforming approach by showing how difficult it is to estimate channels for massive IRSs. The existing works mostly estimate each cascaded channel $h_{n_1,\ldots,n_L}$ separately. It can be seen from (2) that the number of cascaded channels grows exponentially with the number of IRSs. Alternatively, one may suggest estimating the channel between each pair of IRSs and further obtaining $h_{n_1,\ldots,n_\ell}$ via concatenation, thereby reducing the number of estimated channels to $2NL + \binom{L}{2}N^2 = O(N^2L^2)$. The main issue with the above approach is that the cost and the complexity of IRS would then increase significantly because a receive antenna must be deployed at each RE for channel acquisition. Hence, we wish to get rid of channel estimation altogether and learn how to configure the IRSs without channel information.

A. Conditional Sample Mean (CSM)

Before proceeding to a fully general algorithm for multiple IRSs, let us first consider how to perform blind beamforming for a single-IRS system, i.e., the special case with $L = 1$. Since there is only one IRS, we drop the IRS index ℓ throughout this subsection, e.g., θ_{n_1} is written as θ_n.

If the channels are already known, then a natural idea is to align each reflected channel with the direct channel so as to maximize their superposition; if the perfect alignment cannot be achieved due to the discrete constraint Φ_K, we could choose each θ_n, $n = 1, \ldots, L$, to rotate h_n to the closest possible

Algorithm 1 Sequential Conditional Sample Mean (SCSM)

1: Initialize all the θ_{n_ℓ}'s to feasible values.
2: **for** $\ell = 1, \ldots, L$ **do**
3: Generate a total of T random samples.
4: Compute the conditional sample mean as in (8).
5: Decide θ_{n_ℓ} according to (9) for IRS ℓ.
6: **end for**

position to h_0, namely *closest point projection (CPP)*. As a result, the SNR boost ρ is quadratic in N in most cases.

It turns out that the CPP method can be carried out implicitly in the absence of channel knowledge. We first generate a total of T random samples of $\{\theta_n^{(t)} : n = 1, \ldots, N\}$, where $t = 1, \ldots, T$ is the sample index and each entry θ_n is drawn i.i.d. uniformly from Φ_K. Let $\mathcal{G}_{n,k} \subseteq \{1, \ldots, T\}$ be a subset of indices of those random samples with $\theta_n^{(t)}$ randomly set to $k\omega$, i.e.,

$$\mathcal{G}_{n,k} = \left\{ t \in [1:T] \text{ such that } \theta_n^{(t)} = k\omega \right\}. \tag{7}$$

Measure the corresponding received signal power $|Y^{(t)}|^2$ for each random sample t, and further compute the conditional sample mean of $|Y^{(t)}|^2$ within each $\mathcal{G}_{n,k}$:

$$\widehat{\mathbb{E}}[|Y|^2|\theta_n = k\omega] = \frac{1}{|\mathcal{G}_{n,k}|} \sum_{t \in \mathcal{G}_{n,k}} |Y^{(t)}|^2. \tag{8}$$

The CSM method is to decide each θ_n such that the related conditional sample mean is maximized, denoted as

$$\theta'_n = \arg\max_{\varphi \in \Phi_K} \widehat{\mathbb{E}}[|Y|^2|\theta_n = \varphi]. \tag{9}$$

Proposition 1 (Theorem 2 in [3]): The CSM method in (9) yields an SNR boost that is quadratic in the number of REs, i.e.,

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

where the factor ρ is defined to be

$$\rho = \frac{\sum_{n=1}^N |h_n|^2}{N|h_0|^2}. \tag{11}$$

B. Sequential CSM for Multiple IRSs

We propose extending the above CSM method to multiple IRSs in a sequential fashion. Specifically, for each iteration, we run CSM on one particular IRS while holding the phase shifts of the rest IRSs fixed.

Simple as the above extended CSM seems, it is guaranteed to be yield equally good performance as the existing method in [1] based on continuous beamforming and perfect channel information. The following proposition provides more details.

Proposition 2: Consider a double-IRS system with the number of phase shift choices $K > 2$. Assume line-of-sight (LoS) propagation between the two IRSs so that the two-hop channel matrix is rank one [1] and can be decomposed as

$$\begin{bmatrix} h_{1,1} & \cdots & h_{1,N} \\ \vdots & & \vdots \\ h_{N,1} & \cdots & h_{N,N} \end{bmatrix} = \begin{bmatrix} u_1 \\ \vdots \\ u_N \end{bmatrix} \begin{bmatrix} v_1 & \cdots & v_N \end{bmatrix}. \tag{12}$$

If there exists some $\gamma \in [0, \frac{\pi}{2} - \frac{\pi}{K})$ such that

$$|h_{n_1,0}| \le \sin\gamma \cdot \left| \sum_{n_2=1}^N h_{n_1,n_2} \right|, \quad \text{for all } n_1 = 1, \ldots, N, \tag{13}$$

then the SCSM method yields a quartic SNR boost as

$$\mathbb{E}\left(\frac{\text{SNR}}{\text{SNR}_0}\right) = \frac{\delta_1^2\delta_2^2}{|h_0|^2} \cdot \Theta(N^4), \tag{14}$$

where

$$\delta_1 = \frac{1}{N}\sum_{n_1=1}^N |u_{n_1}| \quad \text{and} \quad \delta_2 = \frac{1}{N}\sum_{n_2=1}^N |v_{n_2}|, \tag{15}$$

after only one iteration, i.e., when the two IRSs have been optimized one time each.

Proof: Since $|h_0|^2$ and P are fixed, it suffices to show that $\mathbb{E}[|g|^2] = \delta_1^2\delta_2^2 \cdot \Theta(N^4)$, where g represents the superposition of all the channels with the IRS phase shifts θ_{n_1} and θ_{n_2}, i.e.,

$$g(\theta_{n_1}, \theta_{n_2}) = h_{0,0} + \sum_{n_1=1}^N h_{n_1,0}e^{j\theta_{n_1}} + \sum_{n_2=1}^N h_{0,n_2}e^{j\theta_{n_2}} + \sum_{n_1=1}^N\sum_{n_2=1}^N h_{n_1,n_2}e^{j(\theta_{n_1}+\theta_{n_2})}. \tag{16}$$

To establish $\mathbb{E}[|g|^2] = \delta_1^2\delta_2^2 \cdot \Theta(N^4)$, we need to verify the converse $\mathbb{E}[|g|^2] = \delta_1^2\delta_2^2 \cdot O(N^4)$ and the achievability $\mathbb{E}[|g|^2] = \delta_1^2\delta_2^2 \cdot \Omega(N^4)$.

The converse is evident since

$$|g|^2 \le \left| |h_{0,0}| + \sum_{n_1} |h_{n_1,0}| + \sum_{n_2} |h_{0,n_2}| + \sum_{n_1,n_2} |h_{n_1,n_2}| \right|^2 = \delta_1^2\delta_2^2 \cdot O(N^4).$$

The rest of the proof focuses on the achievability.

Without loss of generality, assume that both θ_{n_1} and θ_{n_2} are initialized to zero at the beginning of the proposed algorithm; otherwise the initial phase shifts can be incorporated into the channels. According to the algorithm, we first configure IRS 1 with IRS 2 held fixed, by treating all the channels related to IRS 1 as the reflected channel and the rest as the direct channel. Thus, the continuous solution of θ_{n_1} is to align the reflected channel with the direct channel exactly, i.e.,

$$\theta^\star_{n_1} = \angle\underbrace{\left(h_{0,0} + \sum_{n_2} h_{0,n_2}\right)}_{\text{direct channel}} - \angle\underbrace{\left(h_{n_1,0} + \sum_{n_2} h_{n_1,n_2}\right)}_{\text{reflected channel}}. \tag{17}$$

According to Proposition 1, configuring IRS 1 by conditional sample mean is equivalent to rotating the reflected channel to the closest position to the direct channel, i.e.,

$$\theta'_{n_1} = \arg\min_{\theta\in\Phi_K} \left|\theta - \theta^\star_{n_1}\right|. \quad (18)$$

Clearly, we have

$$\left|\theta'_{n_1} - \theta^\star_{n_1}\right| \le \frac{\pi}{K}. \quad (19)$$

Moreover, we approximate the continuous solution $\theta^\star_{n_1}$ by removing the single-hop reflect channels, which is

$$\hat{\theta}^\star_{n_1} = \angle\left(h_{0,0} + \sum_{n_2} h_{0,n_2}\right) - \angle\left(\sum_{n_2} h_{n_1,n_2}\right). \quad (20)$$

Because of (13), the error of the above approximation can be bounded above as

$$\left|\hat{\theta}^\star_{n_1} - \theta^\star_{n_1}\right| \le \gamma. \quad (21)$$

Combining (19) and (21) gives

$$\left|\hat{\theta}^\star_{n_1} - \theta'_{n_1}\right| \le \frac{\pi}{K} + \gamma. \quad (22)$$

Subsequently, IRS 2 is configured with each phase shift of IRS 1 fixed at θ'_{n_1}. We now treat all the channels related to IRS 2 as reflected channel and treat the rest as the direct channel. Thus, the continuous solution of θ_{n_2} is given by

$$\theta^\star_{n_2} = \angle\underbrace{\left(h_{0,0} + \sum_{n_1} h_{n_1,0}e^{j\theta'_{n_1}}\right)}_{\text{direct channel}} - \angle\underbrace{\left(h_{0,n_2} + \sum_{n_1} h_{n_1,n_2}e^{j\theta'_{n_1}}\right)}_{\text{reflected channel}}. \quad (23)$$

Again, by Proposition 1, it can be shown that

$$\theta'_{n_2} = \arg\min_{\theta\in\Phi_K} \left|\theta - \theta^\star_{n_2}\right| \quad (24)$$

and

$$\left|\theta'_{n_2} - \theta^\star_{n_2}\right| \le \frac{\pi}{K}. \quad (25)$$

For ease of notation, we write

$$\xi_{n_2} = h_{0,n_2} + \sum_{n_1} h_{n_1,n_2}e^{j\theta'_{n_1}}. \quad (26)$$

It can be shown that

$$|g(\theta'_{n_1},\theta'_{n_2})|^2 = \left|h_{0,0} + \sum_{n_1} h_{n_1,0}e^{j\theta'_{n_1}} + \sum_{n_2} e^{j\theta'_{n_2}}\xi_{n_2}\right|^2 \overset{(a)}{\ge} \left(\cos\frac{\pi}{K}\cdot\sum_{n_2}|\xi_{n_2}|\right)^2, \quad (27)$$

where the lower bound in (a) is obtained by projecting each $e^{j\theta'_{n_2}}\xi_{n_2}$ onto $h_{0,0}+\sum_{n_1} h_{n_1,0}e^{j\theta'_{n_1}}$ and by using the fact that $\left|\angle(e^{j\theta'_{n_2}}\xi_{n_2}) - \angle(h_{0,0} + \sum_{n_1} h_{n_1,0}e^{j\theta'_{n_1}})\right| = |\theta'_{n_2} - \theta^\star_{n_2}| \le \pi/K$ according to (25). We further bound the value of $|\xi_{n_2}|$:

$$\begin{aligned}
|\xi_{n_2}| &= \left|h_{0,n_2} + \sum\nolimits_{n_1} h_{n_1,n_2}e^{j\theta'_{n_1}}\right| \\
&= \left|\sum\nolimits_{n_1} h_{n_1,n_2}e^{j\theta'_{n_1}}\right| + o(N) \\
&= \left|\sum\nolimits_{n_1} h_{n_1,n_2}e^{j\hat{\theta}^\star_{n_1}}e^{j(\theta'_{n_1}-\hat{\theta}^\star_{n_1})}\right| + o(N) \\
&\overset{(b)}{=} \left|\sum\nolimits_{n_1} u_{n_1}v_{n_2}e^{j(\eta-\angle u_{n_1})}e^{j(\theta'_{n_1}-\hat{\theta}^\star_{n_1})}\right| + o(N) \\
&= |v_{n_2}|\cdot\left|\sum\nolimits_{n_1} |u_{n_1}|e^{j(\theta'_{n_1}-\hat{\theta}^\star_{n_1})}\right| + o(N) \\
&\overset{(c)}{\ge} |v_{n_2}|\cdot\cos\left(\gamma+\frac{\pi}{K}\right)\cdot\sum\nolimits_{n_1}|u_{n_1}| + o(N),
\end{aligned} \quad (28)$$

where a new variable $\eta = \angle\big(h_{0,0}+\sum_{n_2} h_{0,n_2}\big)-\angle\big(\sum_{n_2} v_{n_2}\big)$ is introduced in step (b) and can be ignored in the rest steps since it is independent of n_1. In the above sequence of inequalities, step (b) follows by the rank-one assumption in (12), and step (c) follows by the bound between θ'_{n_1} and $\hat{\theta}^\star_{n_1}$ as stated in (22). Finally, combining (27) and (28) gives

$$\begin{aligned}
|g(\theta'_{n_1},\theta'_{n_2})|^2 &= \Omega\left(\left(\sum\nolimits_{n_2}\sum\nolimits_{n_1}|v_{n_2}||u_{n_1}|\right)^2\right) \\
&= \delta_1^2\delta_2^2\Omega(N^4).
\end{aligned} \quad (29)$$

The proof is then completed. ■

C. Some Comments on Proposition 2

We first provide intuition on the condition assumed in Proposition 2. The channel matrix factorization in (12) plays a key role in bounding below the term $|\xi_{n_2}|$ in (28). Intuitively, we aim to make each $|\xi_{n_2}|$ be in order of N. Because $\xi_{n_2} \approx \sum_{n_1} h_{n_1,n_2}e^{j\theta'_{n_1}}$, it requires at least a constant portion of $\{h_{n_1,n_2}e^{j\theta'_{n_1}}\}$ to be aligned. But we have only N phase shift variables θ'_{n_1} to manipulate; the degrees of freedom are far insufficient to achieve the above alignment for every $|\xi_{n_2}|$, $n_2 = 1,\ldots,N$. Nevertheless, with the channel matrix $[h_{n_1,n_2}]$ factorized as in (12), we gain the great convenience that once $\{h_{n_1,n_2}e^{j\theta'_{n_1}}\}$ are aligned for a particular ξ_{n_2} then the alignments are automatically attained for the rest ξ_{n_2}'s. This desirable property enables us to reach linear growth for every $|\xi_{n_2}|$ by choosing θ'_{n_2} properly. This result suggests deploying the IRSs at the LoS locations.

It is also worth noting that Proposition 2 generalizes the existing result in [1] in two respects. First, [1] assumes continuous beamforming while we assume discrete beamforming. Second, [1] assumes that $h_{0,0} = h_{n_1,0} = h_{0,n_2} = 0$ for any n_1, n_2 while we only assume that they are sufficiently small (and not necessarily approaching zero). The following remark summarizes the above comparison.

Remark 1: The previous work [1] also shows that an SNR boost of $\Theta(N^4)$ can be achieved for a double-IRS system under certain condition. However, the condition assumed in [1] is much stricter than that assumed in Proposition 2. First, [1] only considers continuous beamforming by letting $K \to \infty$ whereas Proposition 2 allows finite K. Second, [1] requires that both the direct channel $h_{0,0}$ and the single-hop reflected channels $\{h_{n_1,0}, h_{0,n_2}\}$ be zero whereas Proposition 2 only requires them to be sufficiently small. Most importantly, [1] assumes perfect channel information available whereas this work pursues blind beamforming.

Furthermore, as an extension of Proposition 2, it can be shown that an SNR boost of $\Theta(N^{2L})$ can be reaped for an

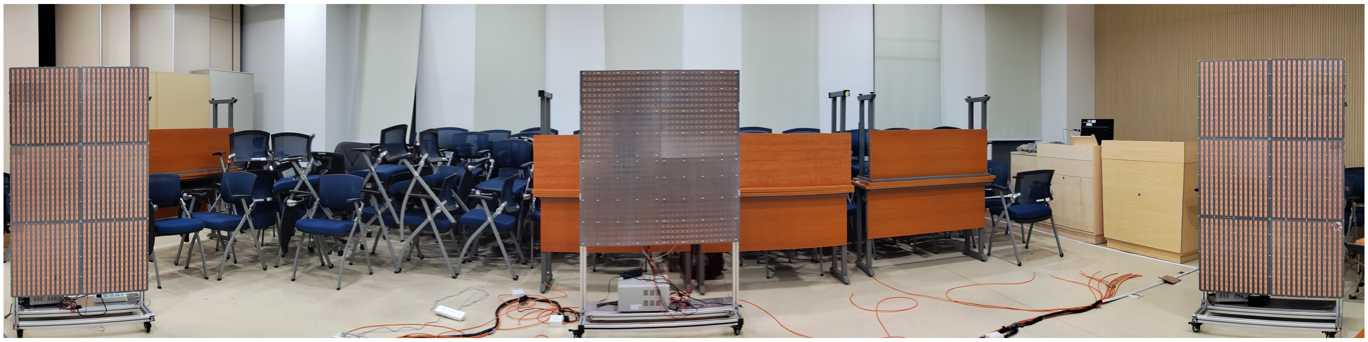

Fig. 2. The triple-IRS system in our prototype tests. The IRS in the middle comprises 64 REs and provides 4 phase shift options $\{0, \pi/2, \pi, 3\pi/2\}$, while the other two IRSs comprise 294 REs each and provide 2 phase shift options $\{0, \pi\}$. We reduce it to a double-IRS system by removing the middle IRS.

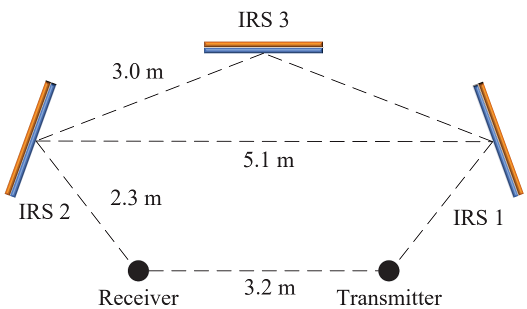

Fig. 3. Layout drawing of the triple-IRS system in our field tests. It reduces to the double-IRS system when IRS 3 (located in the middle) is removed.

L-IRS system when $L \geq 2$. Lack of space prevents our discussing this nontrivial extension. We refer the reader to the journal version of this work for the full generalization. Nevertheless, the experiments in this paper involve more than two IRSs, as shown in the following section.

IV. Experiments

A. Field Tests

Our field tests are carried out in an indoor environment as displayed in Fig. 2; the detailed locations of all the devices can be found in Fig. 3. We use the 2.6 GHz spectrum band for the transmission. Three IRSs are used:

- IRS 1 comprises 294 REs each and provides 2 phase shift options $\{0, \pi\}$, i.e., $N_1 = 294$ and $K = 2$.
- IRS 2 is identical to IRS 1, i.e., $N_2 = 294$ and $K = 2$.
- IRS 3 comprises 64 REs and provides 4 phase shift options $\{0, \pi/2, \pi, 3\pi/2\}$, i.e., $N_3 = 64$ and $K = 4$.

Aside from the proposed Algorithm 1, the following methods are included in field tests as benchmarks:

- *Single-IRS:* Further remove IRS 2 from the double-IRS system, and configure IRS 1 by the CSM method [3].
- *Double-IRS non-LoS:* Put IRS 1 and IRS 2 together at the location of IRS 1 to form a larger IRS (with 588 REs); CSM [3] is used to configure this combined IRS.

TABLE I
Performance of the Different Methods in Field Tests

Algorithm	SNR (dB)	SNR Boost (dB)
Without IRS	10.07	0.00
Single-IRS	16.96	6.89
Double-IRS non-LoS	18.09	8.02
Double-IRS uncoordinated	18.94	8.87
Double-IRS	20.13	10.06
Triple-IRS	21.89	11.82

- *Double-IRS uncoordinated:* IRS 1 and IRS 2 are located as in Fig. 3. Configure them using CSM [3] by treating the two IRSs as a single IRS.

Each of the above three benchmarks uses 2000 random samples. The proposed SCSM method uses 1000 random samples for each IRS.

The test results are summarized in Table I. It shows that a remarkable SNR boost of around 7 dB can already be achieved with only one IRS deployed. When two IRSs are available, distributing them as in Fig. 3 is more beneficial than putting them together (i.e., double-IRS non-LoS); this result agrees with our conclusion from Proposition 2. By comparing “double-IRS uncoordinated” and “double-IRS”, it can be observed that the proposed algorithm outperforms the baseline of treating two IRSs as a single IRS. Observe also that adding one more IRS to the system can further raise the SNR boost by about 2 dB, even through the added IRS consists of merely 64 REs.

B. Simulation Tests

We further consider some other tests in simulations; these tests are difficult to do on the prototype machines because of the hardware issue. Consider a double-IRS system. The transmitter, receiver, IRS 1, and IRS 2 are located at $(1, 0, 2)$, $(1, 50, 0)$, $(0, -5, 1)$, and $(0, 55, 1)$, respectively, all in meters. The transmit power is 20 dBm and the background noise power is -80 dBm. The direct pathloss model is $32.6 + 36.7 \log_{10} d$ and the reflected pathloss model is $30 + 22 \log_{10} d$, where d

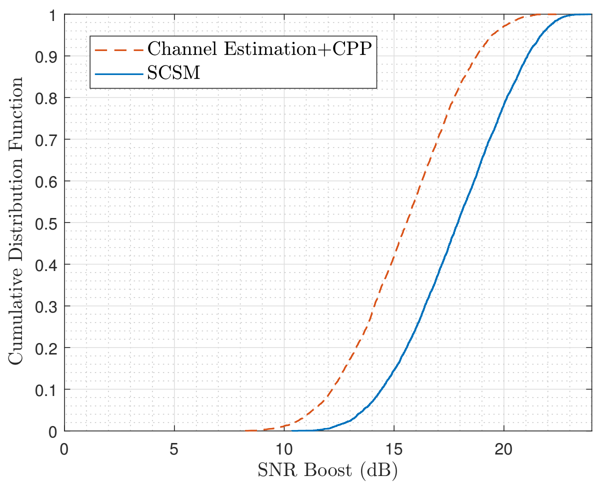

Fig. 4. SCSM vs. the conventional channel estimation based method.

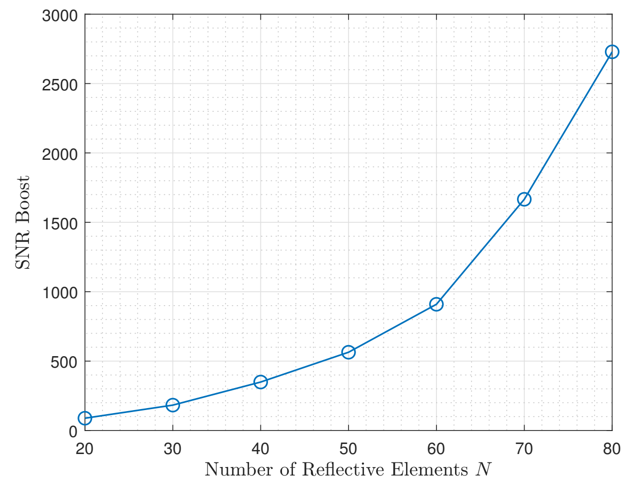

Fig. 5. SNR Boost achieved by SCSM vs. the number of REs per IRS.

is the distance in meters. Rayleigh fading is assumed. The set of phase shift choices is $\{0, \pi/2, \pi, 3\pi/2\}$.

Fig. 4 shows the cumulative distribution of the SNR boost for the proposed blind beamforming method versus the channel estimation based method. The channels are estimated by the DFT method [15]. After obtaining the channels, the two IRSs are optimized in an alternating fashion such that each reflected channel is aligned with the direct channel, and ultimately the continuous solution is rounded to the discrete set, namely the *closest point projection (CPP)*. As shown in Fig. 4, the proposed method SCSM outperforms the channel estimation plus CPP significantly. At the 50th percentile, the proposed method achieves approximately 2 dB higher in terms of the SNR boost than the channel estimation based method.

Moreover, we plot how the SNR boost of the proposed method scales with the number of REs N in Fig. 5. It shows that the proposed method already yields a near-quartic growth of the SNR boost even for a modest rise in N.

V. Conclusion

This work aims to coordinate multiple IRSs in the absence of channel state information to avoid the complexity of cascaded channel estimation. The main idea is to broaden the scope of the blind beamforming strategy in [3] from the single-IRS case to the multi-IRS case. The extended blind beamforming algorithm guarantees a quartic SNR boost for a double-IRS system under a more general setting than the previous study in [1]. As shown in the real-world experiments, applying the proposed algorithm to a triple-IRS system can raise SNR by almost 12 dB.

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