Visualization tool of the urban microcell radio propagation paths

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Abstract: This paper presents a tool for the visual inspection of the small interacting objects (IOs) for the microwave propagation in urban environment, by utilizing 3D point cloud data of townscape and double directional channel sounding results. It is demonstrated that the tool can effectively visualize the location of the IOs.

Keywords: radio channel measurement, microwave, microcell, interacting object

Classification: Antennas and Propagation

References

1 Introduction

According to the raise of the carrier frequency, the design of the cell site tends to be more specific, since various objects in the environment become electrically large. The ray tracing prediction of the propagation channel has been widely utilized, in the conventional ultra high frequency (UHF) band, by considering the simplified building structure. According to the increase of the carrier frequency, more detailed structures of the environment such as surface irregularity and small objects are not negligible in terms of radiowave scattering. The behavior of the propagation channel tends to be more site-specific, and the prediction of the channel property is getting more important in the site planning. Therefore more complex techniques are needed to model these small objects. Various extensions of prediction methods have been proposed such as physical optics, effective roughness model, and hybrid T-matrix method. To explore more details about the scattering mechanisms, it is crucial to observe the measured channel properties.

The authors have conducted the 24×24 MIMO channel sounding at 11 GHz with bandwidth of 400 MHz in indoor and outdoor environments [1, 2]. A parametric model in double-directional and delay domains is introduced and a high resolution maximum likelihood parameter estimation is applied to extract the angles of departure, arrival, and delay time of individual path. In the indoor environment, a measurement-based ray tracing (MBRT) combined with the geometry based clustering has been implemented to identify the interacting object (IO) and interaction loss (IL) [3]. However, it is more complicated to conduct MBRT in the outdoor scenario due to difficulty of modeling the larger environment with more detail of the surface of the objects. This study aims at the empirical identification of the IOs in the urban microcell environment. 3D townscape is captured by using the mobile mapping system through laser scanner. Angles of arrival and departure of paths are utilized to visualize the specific IOs.

2 Radio channel measurement

2.1 Description of the measurement site

The outdoor measurements were conducted in an urban microcell in Ishigaki city, Okinawa, Japan at 11GHz band. The map of the environment where the measurements were carried out is depicted in Fig. 1 [4]. The transmitter (Tx) was positioned at different locations represented by the red line of the measurement route 2, while the receiver (Rx) is fixed. Both Tx and Rx antennas are mounted on the cars at the height of 3 meters.
2.2 Channel sounding and parameter estimation

24×24 MIMO channel sounder with 400 MHz bandwidth was used to measure double-directional and delay properties of the propagation channel [1]. The base station (BS) and the mobile station (MS), which are Rx and Tx respectively, both use the 12-elements uniform circular antenna array. This antenna has both vertical and horizontal polarized elements spaced by 11.9 mm (0.44λ). The transmit signal is an unmodulated multi-tone signal having 2048 tones. Maximum likelihood parameters estimation was applied to the measurement results to extract the discrete path parameters, i.e. angles of arrival and departure, delay time and polarimetric gain [5].

The double directional channel model for single frequency bin is represented by [5].

\[
H(f) = \sum_{l=1}^{L} a_R(\theta^R_l, \phi^R_l, f) \Gamma_l a_T(\theta^T_l, \phi^T_l, f) e^{-j2\pi f \tau_l}
\]

where \(0^T\) and \(q^T\) are the zenith and the azimuth of departure respectively at Tx side, \(0^R\) and \(q^R\) are the zenith and the azimuth of arrival respectively at Rx side as shown in Fig. 2(b). \(H\) is the channel matrix, \(L\) is the total number of propagation path, \(a_R\) and \(a_T\) are the Rx and Tx polarimetric antenna responses and \(\Gamma_l\) is the path weight matrix which is expressed as a matrix of all four polarization pairs.

\[
\Gamma_l = \begin{bmatrix}
\gamma_{VV,l} & \gamma_{VH,l} \\
\gamma_{HV,l} & \gamma_{HH,l}
\end{bmatrix}
\]

3 Visualization tool

The prediction of channel properties is more important for the higher frequency with smaller coverage. Although ray tracing simulation has been widely used for the purpose, there are several modeling challenges such as surface irregularity comparable with the wavelength and objects smaller than first Fresnel zone. Measurement based approach to identify the significant IOs is considered as the diagnosis of the
discrepancy in ray tracing simulation. In order to acquire the details of the environment, the introduction of the point cloud data has been found useful [6]. The advantage of using the point cloud is proven in the prediction for both specular reflections and diffraction as well as the total channel [7]. During the measurement campaign, a mobile mapping system (Topcon IP-S2 Lite) captured the 3D facade of the townscape. It provided the point cloud data of the objects surrounding the MS by a laser scanner, together with the Global Positioning System (GPS) coordinates, and the panoramic photos.

The tool is implemented for the visualization of the small IOs. OpenGL library is used to read and visualize point cloud data of the townscape of the measurement site, and the extracted angular information at each snapshot to reproduce the measurement scenario.

MS, which is Tx, is depicted by a cone with the bottom on the ground and the vertex at the location of the antenna.

For any point \( M = (x_M, y_M, z_M) \), the line equation through \( T = (x_T, y_T, z_T) \) and \( M \) is given by:

\[
M = T + \rho u
\]  

(3)

where \( \rho \) the distance from the Tx to the point \( M \) and the unit vector pointing from the Tx to the point \( M \) is represented as \( u = (\sin \theta_T \cos \phi_T, \sin \theta_T \sin \phi_T, \cos \theta_T) \). Here, \( \theta_T \) and \( \phi_T \) represent zenith and azimuth of departure of a propagation path obtained through the measurement.

![Fig. 2. Illustration of a path and IO](image)

The determination of the IO for each path is performed through three steps:

1. The selection of the candidate scattering points (CSPs) representing IOs:

   A point \( P \) in the cloud is a candidate point if the distance between \( P \) and the
propagation path (Eq. (3)) is within a threshold \(d_{th}\).

\[
d^2 = \|P - T\|^2 - ((P - T) \cdot u)^2 \leq d_{th}^2
\]  

(4)

2. The determination of the scattering points (SP):
   To determine the SP representing IO, the distance between \(T\) and the projection of \(P\) onto the propagation path

\[
\rho = (P - T) \cdot u
\]  

(5)

is evaluated for each SIP, and select the smallest \(\rho\). Finally, substituting this \(\rho\) into Eq. (3), \(M\) represents the SP.

3. Visual classification of the IO:
   The propagation path is then visualized as the line segment between Tx and SP, and SP is displayed as a sphere. Classification of the IO is done visually, by moving the viewpoint within the visualization tool as shown in the attached video, and with the help of panoramic photo if necessary.

Fig. 3. Modeling of the equipment tests
4 Results: visualized objects

Considering the density of the point cloud, $d_{th}$ is fixed to 0.9 m.

Different IOs are visualized at different MS locations. In this section some of the visualized IOs are presented as the examples of sources of scattering. The first example of the IO is the building facade. It is represented by the green circles in the output of visualization tool Fig. 3(a) and the spherical photo Fig. 3(e). Building facade is the most collided surface in urban area. The second type of IOs most present are the lampposts. Figure 3(b) depicts a lamppost as the spot of the ray with the red circle. They approximately measure less than 50 cm in width and more than 4 m in height and are located in sidewalk. Similarly Fig. 3(c) allows to identify a parked car which corresponds to yellow circle. The orange circle in Fig. 3(d) corresponds to the signboard. These objects are depicted by the circles with the same color in the spherical photo Fig. 3(e).

5 Conclusion

This paper has presented a visualization method of the urban mobile propagation paths identified from the measurement data. Mobile mapping system has been used to capture the townscape of the measurement site. Point cloud data obtained from the scanning has been utilized to implement the 3D reconstruction of the measurement site. At every snapshot, the extracted angular information of the propagation path is traced and the interacting objects are identified. This tool has been developed to identify small IOs via visual inspection to facilitate detailed analysis of the urban propagation for microwave frequencies that have been omitted in many ray tracing simulators.
An efficient method of sub-signal channel modulation on hitless redundancy switching systems

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Abstract: In the hitless redundancy switching system, the selector of the redundancy switching device is controlled on the basis of the control information such as the connection status of the path to transmit frames, end-to-end delay of the path, and jitter. However, the control signal transmitted superimposed on the main signal channel in the hitless redundancy switching system is not highly confidential because all the information is transmitted superimposed on the two paths. In our previous study, we modulated the transmission timing of frames in the main signal channel by assigning “0” and “1” codes to the two paths in the hitless redundancy switching system. In this paper, we propose a method to efficiently modulate sub-signal channels in a hitless redundancy switching system to improve the communication bandwidth, and evaluate its effectiveness in a numerical simulation.

Keywords: hitless, redundant path network, Ethernet

Classification: Network System

References

1 Introduction

Wide-area Ethernet has become a useful option for low-cost communication between multiple locations. In wide-area Ethernet services, networks are used to increase availability. Hitless redundancy switching systems (HRSS) have been discussed that transfer frames to multiple paths for communication without interruption [1, 2, 3].

In HRSS, the selector of the redundant switching device is controlled on the basis of the control information such as the connection status of the path to transmit frames, end-to-end (e2e) delay of the path, and jitter. In general, the control signals to control the device itself are sent on a separate channel from the main signal channel where user data is sent [4, 5], and the control signals to control the transfer of user data between devices are often sent superimposed on the main signal channel. The control signals exchanged between devices are also sent by superimposing them on the main signal channel as they are, so it is difficult to say that the system has excellent confidentiality. However, if the control signals can be deciphered only by combining the information of both control signals transmitted over two paths, high-security transmission can be achieved by taking advantage of the characteristics of the HRSS. However, to the best of our knowledge, there have been few reports on the communication of such control signals.

We previously proposed a method to transmit binary code information separately from data in user frames by modulating the transmission timing of frames in the main signal channel with a code of “0” and “1” for each of the two paths in a HRSS and having the receiver first arrive at the main signal channel frame of the intended path by a sequence number [6].

In this paper, we propose a method to efficiently modulate sub-signal channels in HRSS to improve the communication bandwidth and evaluate its effectiveness by numerical simulation.

2 Sub-signal channel multiplex on HRSS

2.1 Conventional HRSS

A conventional HRSS with a two-route redundant configuration consists of a pair of redundancy switching devices as shown in Fig. 1. Each device has one port that connects to the user NW and two ports that connect to the relay NW. Let us consider sending from user #1 to user #2 using the HRSS. Redundancy switching device #1 assigns a sequence number to the frame received from user #1, duplicates it, and sends it to NW #1 and NW #2. Redundancy switching device #2 checks the frames received from the respective relay NWs and confirms the order of arrival of frames with the same sequence number. The first arriving frame is sent to the user NW, and the second arriving frame is discarded. With the above operation, the HRSS...
provides uninterrupted communication between users #1 and #2 even if one of the routes becomes disconnected.

2.2 Proposed sub-signal channel modulation method

The proposed sub-signal channel modulation method is shown in Fig. 1(b). The main signal channel and the sub-signal channel are superimposed at layer 1.

The main signal channel duplicates the user frame at redundancy switching device #1 and forwards it to redundancy switching device #2 via two relay NWs, which is the basic function of the HRSS. The sub-signal channel transmits control signals. Figure 1(b) shows, as an example, a case where a control signal with the code sequence “1010101” is transmitted.

The sub-signal channel does not prepare the bits to be transmitted as in the user frame of the main signal channel. Instead, each bit of the control signal is composed of two frames that are duplicated to transfer the redundant path in the main signaling channel, and the transmission timing is adjusted so that the intended arrival order is achieved upon reception. In Fig. 1(b), the code conversion is performed so that the transmission timing at which NW #1 arrives first is “1,” and the transmission timing at which NW #2 arrives first is “0.” To ensure that the signal of the intended path arrives first, redundancy switching device #1 delays the frame transmission time of the path to be arrived at later by $T_{\text{DelayModulation}}$. For example, the first code “1” of the sub-signal channel in Fig. 1(b) delays the frame transmission time of sequence number 1 of NW #2 by $T_{\text{DelayModulation}}$ to control the frame of sequence number 1 of the main signal channel so that the frame via NW #1 arrives first. Redundancy switching device #2 decodes the information of the sub-signal channel by judging the port number (routes) to which the frame with the same sequence number arrived.
first as a binary code. We note that the main signal transmits continuously in order for the sub signal to transmit.

For transmission using the above method, the modulation speed \( v \) of the sub-signal channel is expressed by the following equation

\[
v = \frac{\alpha}{(\beta + \alpha \times T_{\text{DelayModulation}})},
\]

where \( \alpha \) is the wire rate of the main signal channel, \( \beta \) is the bit length of the user frame, and \( T_{\text{DelayModulation}} \) is the delay time interval. From eqn. (1), we can see that the modulation rate of the sub-signal channel can be increased by keeping \( T_{\text{DelayModulation}} \) small.

\( T_{\text{DelayModulation}} \) can be set to a sufficiently small value when the difference between the average path delays of NW #1 and NW #2, \( |T_{\text{DelayNW #1}} - T_{\text{DelayNW #2}}| \), is zero and the jitter can be regarded as zero, because the intended reception timing will not be shifted.

On the other hand, if \( |T_{\text{DelayNW #1}} - T_{\text{DelayNW #2}}| > 0 \), then \( T_{\text{DelayModulation}} > |T_{\text{DelayNW #1}} - T_{\text{DelayNW #2}}| \) must be satisfied to prevent the intended reception timing from shifting due to the difference in propagation delay caused by the different routes.

Figure 2 shows the timing when redundancy switching device #1 sends frames to each path and the timing when redundancy switching device #2 receives frames from each path when NW #1 has a larger average path delay than NW #2. Here, the code sequence transmitted by the sub-signal channel is assumed to be the “1010101” signal.

Figure 2(a) shows an existing timing adjustment method in which frames to be post-delivered are transmitted with a \( T_{\text{DelayModulation}} \) delay. The reception timing is the intended reception timing in each code. However, improvement is needed considering that the difference between the two path lengths for forwarding the frames is large enough to reduce \( v \).

![Fig. 2. Proposed reception timing correction method.](image-url)
Figure 2(b) shows the proposed reception timing correction method. In this method, the reception timing of a frame received from a path with a short average path delay is delayed by the average path delay difference from a path with a long average path delay. As a result, since the average path delay difference can be excluded from the value of $T_{\text{DelayModulation}}$, which delays the frame to be post-delivered, the modulation speed $v$ of the sub-signal channel can be increased by setting $T_{\text{DelayModulation}}$ to a smaller value than in Fig. 2(a).

### 3 Numerical analysis

To evaluate the effectiveness of the proposed method, we calculate the existing modulation rate $v$ and the proposed modulation rate $v'$ of the sub-signal channel in a HRSS. Since the average path delay difference is 100 msec, the delay time interval $T_{\text{DelayModulation}}$ of $v$ is 100 msec. On the other hand, $v'$ does not need to take the average path delay difference into account. Then, $T_{\text{DelayModulation}}$ of $v'$ is 1 msec taking only the jitter into account. The code sequence of the sub-signal channel used for the evaluation is a code sequence of alternating “1” and “0” such as “1010...10”. Figure 3(a) shows the results of evaluating the values of transmission speeds $v$ and $v'$ of the sub-signal channel when the value of $\beta$ is changed from 64 to 1522 bytes when the wire rate of the main signal channel is $\alpha = 1$ Gbps with values of $T_{\text{DelayModulation}}$. $v$ is about 10 b/s because the value of the average path delay difference (100 msec) is included in $T_{\text{DelayModulation}}$. On the other hand, $v'$ is about 1 kb/s in the frame length range when considering only the jitter (1 msec) because the value of the average path delay difference is not included in $T_{\text{DelayModulation}}$. This shows that the proposed method increases the modulation speed by about 100 times compared with the existing method.

The physical limit of the proposed method can be specified as the speed at $T_{\text{DelayModulation}} = 0$, which is shown by the black line. The two factors that increase $T_{\text{DelayModulation}}$ are the average path delay difference and jitter. Then, we can see that the magnitude of jitter is the factor that determines the speed of the proposed method. In this evaluation, the jitter is assumed to be 1 msec, but the speed can be improved by using a path with even smaller jitter.

Next, to evaluate the impact of the proposed method on the main signal, we calculated the transmission rate of the main signal with the code sequence where the

![Fig. 3. Numerical results of transmission rate of (a) sub-signal channel and (b) main-signal channel.](image-url)
same code sequence occurs with probability $1/2$. Figure 3(b) shows the transmission rate of the main signal with values of $T_{\text{DelayModulation}}$. As shown in Fig. 3(b), the parameters that affect the speed of the main signal are $T_{\text{DelayModulation}}$ and frame length. In other words, in a system where the frame length of the main signal is long enough and the jitter is kept small, the impact of the proposed method on the main signal is minimal.

Finally, we discuss the bit error of the proposed method. In the proposed method, the bit error is caused by the reversal of the arrival order of frames with the same sequence number received from each path, so that one frame order reversal is a 1-bit error. The bit rate of the sub-signal channel is equal to the frequency of frames that receive a delay greater than the value of $T_{\text{DelayModulation}}$. To suppress bit errors, we can take two approaches. One is to use the proposed method in a delay-guaranteed network where the maximum delay is guaranteed. Another is to optimize value of $T_{\text{DelayModulation}}$ according to jitter of networks.

4 Conclusion

We proposed a method to efficiently modulate the sub-signal channel in a HRSS to improve the modulation rate by reflecting the processing related to the average path delay difference in the timing. Numerical simulations show that the proposed method significantly increases the modulation rate compared with the existing methods.
Attribute-based low-complexity network access control policy with optimal grouping algorithm

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Abstract: Zero Trust (ZT) Model [1] is a security approach to prevent malicious activities assuming the presence of an attacker in the environment. The fine-grained access control should be executed on ZT in accordance with various information, which requires a large complexity of access control policy due to the large patterns of attributes [2]. Our focus is the low-complexity of policy management. We propose a method to reduce and evaluate the complexity of policies for network access control. This letter discloses the optimal grouping algorithm to reduce the complexity, and shows the higher performance in comparison with the existing methods.

Keywords: network access control, policy, complexity, clustering, attribute-based access control

Classification: Network Management/Operation

References


1 Introduction

In an enterprise network, a variety of devices, users, and resources are connected at anytime from anywhere. There is always a possibility of presence of attackers in the internal network. To deal with it, Zero Trust (ZT) model [1] evaluates all access requests from devices, and executes access control with fine-granularity. An important functionality on ZT is the policy definition. Network operators have to configure the rule sets to define how access control should be performed. Attribute-based access control (ABAC) policy is a common choice to implement the fine-grained access control, such as ZT. A detailed ABAC policy definition is very costly because the privilege space becomes too large to deal with as the attributes’ pattern increases. A solution to reduce the policy definition cost is to systematically generate the policy by rule-mining with audit logs [2]. Logic programming or natural language processing-based rule extractions are also known.

However, such methods require log data, declaration of constraints without excess or deficiency, enough to execute the expected access controls. Insufficient assumption will lead to undesirable policy decisions. The decrease of initial policy definition cost will increase the management cost to verify or modify the generated policy to apply it to the network.

We propose a method to reduce the policy management cost by reducing the complexity of the policy, targeting the network access control (NAC) policy. We assume that a fine-grained policy is given by systematic policy generation methods, one of which we introduce in this letter to improve fineness without detailed assumptions. Our main proposal reduces the complexity without affecting the fineness of the given generated policy. We disclose the grouping-based optimization algorithm and the evaluation result.

2 Proposed method

2.1 Overview of access control model

In this section, we briefly explain the overview of our access control model, the policy generation method to reduce policy definition cost, and the complexity reduction method to reduce the management cost of the generated policy.

Network access control model Our target is the network access control (NAC) policy because of its feasibility. It’s defined with IP address, port number, and any additional variables on network or application layer. We can easily deploy it with policy enforcement points (PEPs) as gateways. Our model has the policy generator (PG) which performs the systematic policy generation. For feasibility, the PG also updates the NAC policy in response to changes of information sources such as asset database and activity logs [3]. Our method reduces the complexity of the policy generated by PG at each time by grouping the nodes, e.g. IP address, as shown in Fig. 1. The figure also shows Access control list (ACL) policy and the grouped policy.

Policy generation method We briefly describe the simple PG mechanism by modeling the network operators’ intents for partially delegating the policy definition process to the system. Our model expresses the positive, negative effects and the trade-off between them. All values of attributes should be represented as scores
to indicate the positive effects to the policy decision. The score of location of user = “home” will be lower than “office”. Combinations of attributes have scores to express access rights. The policy is the result of fine evaluation of trade-off, and is updated dynamically. It’s difficult to manage, verifying and modifying due to the fineness.

**Complexity reduction method** We provide an abstract of our complexity reduction method. The previous study [4] had proposed the ABAC policy extraction from ACL tuple using k-means clustering algorithm, achieving both low complexity and fine-granularity i.e. reproduction rate of the original policy. They quantified the complexity with the number of rules, and grouped the similar rules based on attributes. We can group nodes or tuples based on their attributes as well. However, this is not the optimal in NAC. It has the difficulty of parameter tuning and additional pruning of the same rules. These are caused by the conflicts of two independent objectives, the squared error on attributes, and the policy quality metric.

Our contribution is twofold. First, we disclose an algorithm to achieve a NAC policy with the lowest complexity and the finest granularity. We propose a distance metric instead of the squared error. Second, we quantify the entropy-based complexity to evaluate the policy management cost.

### 2.2 Complexity reduction algorithm based on optimal grouping

In this section, we disclose an algorithm to minimize the complexity of NAC policy, while keeping the fineness of the policy, using the following definitions.

NAC policy consists of a set of node $X$, content $E$ such as port number, and
action $A = \{ \text{"Accept", "Deny"} \}$. We assume that a generated policy $P_G$ which defines actions as scores is given, i.e. $P_G : X \times X \times E \to [0, 1] \subset \mathbb{R}$. $P_G$ is generated based on the relevance between node $x$ and attributes, e.g. location of the users assigned to that node. The two nodes $X \times X$ correspond to source and destination, respectively. The meaning of the score is as follows. The access flow containing $x, x' \in X$ and $e \in E$ should be absolutely denied if $P_G(x, x', e) = 0$, whereas accepted if $P_G(x, x', e) = 1$. When $P_G(x, x', e) = 0.4$, the flow should be denied with low confidence. In our proposal, we define a set of group $C$ and a grouping function $\zeta : X \to C$ to reduce the complexity. We describe the set of nodes in $c$ as $X_c = \{ x \in X \mid \zeta(x) = c \}$. The cardinalities of $X, E, C$ are $n, m, k$, respectively. We describe the policy with fixed nodes $R_{x \to x'}$ so that $R_{x \to x'}(c) = P_G(x, x', c)$. The policy can be represented by source and destination groups $c, c' \in C$, such as $P_{c \to c'} : E \to [0, 1]$. The actual actions are obtained as $A_{c \to c'} : E \to A$ discretizing $P_{c \to c'}(e)$.

Our algorithm converts a given policy $R_{x \to x'}$ into a less complex policy $P_{\zeta(x) \to \zeta(x')}$: It solves the optimization problem to minimize $k$, while limiting the dissimilarity between $R_{x \to x'}$ and $P_{\zeta(x) \to \zeta(x')}$ lower than a threshold $\epsilon_{\max}$ in two steps. First, we optimally group the nodes $c = \zeta(x)$. Second, we select the policies $P_{c \to c'}$ for each pair of groups $c, c' \in C$.

Optimal grouping At first, the algorithm calculates $X_c$ in group $c$ with $k$-medoid clustering. To achieve the optimal solution, the algorithm defines a distance matrix based on the policy $R_{x \to x'}$ instead of using the relevant attributes as a feature vector. We define the distance between nodes $d(x, x')$ to describe how much the whole policy changes if $x$ and $x'$ are exchanged.

$$d(x, x') = \sum_{\zeta \in X \setminus \{ x, x' \}} (D(R_{x \to x'}, R_{x' \to \zeta}) + D(R_{\zeta \to x}, R_{\zeta \to x'})) + D(R_{x \to x'}, R_{x' \to x}) \quad (1)$$

where $D(R_{x \to y}, R_{x' \to y'}) = (\sum_{e \in E} (R_{x \to y}(e) - R_{x' \to y'}(e))^p)^{\frac{1}{p}}$ describes the dissimilarity between policies measured by $L^p$-norm. We used $p = 1$. The algorithm evaluates $L_{\text{global}} = \max_{c \in C} [\max_{x, x' \in X_c} [d(x, x')]]$, the maximum intra-class loss. It finds the smallest $k$ satisfying $L_{\text{global}} \leq \epsilon_{\text{max}}$.

Policy selection In the rest of Algorithm 1, the policies for each pair of groups $P_{c \to c'}(e)$ are chosen as representative policies. The algorithm averages the original policies defined between groups. $|X_c|$ is the cardinality of $X_c$. The policies are discretized, and grouped NAC policy $A_{c \to c'}$ is achieved.

### 2.3 Quantification of policy management costs with entropy-based complexity

In this section, we propose a quantification of management cost of NAC policy with entropy in a similar way with [5]. Shannon entropy corresponds to the complexity of optimally designed management interface. The entropy of an original policy generated by PG is $H_{\text{original}} = n(n-1)$, i.e. $O(n^2)$.

We describe two formulae of complexity of grouped $P_G$. The first one is the complexity when the nodes are grouped. The algorithm in the previous section is one example of this kind. The grouping function is $\zeta : X \to C$. The shape of the policy is “$C \times C \times E \to A$.” The first term indicates the number of grouped policies.
Algorithm 1 Optimal complexity reduction algorithm

Require: Original policy $R_{x \rightarrow x'}$ for all $x, x' \in X$ and $\epsilon_{\text{max}}$ are given.

Ensure: Optimize groups $X_c$ and grouped policies $P_{c \rightarrow c'}$.

for $i = 1, 2, \ldots, n$ do
    for $j = 1, 2, \ldots, n$ do
        $d_{i,j} \leftarrow d(x_i, x_j)$
    end for
end for

$k \leftarrow 0$

$L_{\text{global}} \leftarrow \epsilon_{\text{max}} + 1$

while $L_{\text{global}} > \epsilon_{\text{max}}$ do
    $k \leftarrow k + 1$
    $L_{\text{global}} \leftarrow 0$
    for $c = 1, 2, \ldots, k$ do
        $X_c \leftarrow$ Group $c$ obtained by $k$-medoids with the distance matrix $\{d_{i,j}\}$.
        $L_{\text{global}} \leftarrow \max (L_{\text{global}}, \max_{x_i, x_j \in X_c} \{d_{i,j}\})$
    end for
end while

for $c = 1, 2, \ldots, k$ do
    for $c' = 1, 2, \ldots, k$ do
        $P_{c \rightarrow c'} \leftarrow \frac{1}{|X_c| |X_{c'}|} \sum_{x_i \in X_c} \sum_{x_j \in X_{c'}} \{R(x_i, x_j)\}$
    end for
end for

The second term is the complexity to manage the grouping. The symbol $\{\cdot\}$ means the Stirling number of the second kind.

$$H_{\text{node}} = k^2 + \frac{1}{m} \log_2 \left\{ \binom{n}{k} \right\}$$

(2)

The second one is grouping tuples of nodes as $\zeta : X \times X \rightarrow C$ instead of grouping node itself. The shape of the policy is “$C \times E \rightarrow A$.”

$$H_{\text{tuple}} = k + \frac{1}{m} \log_2 \left\{ \binom{n(n-1)}{k} \right\}$$

(3)

For large $n$, the orders are $O(nm^{-1} \log_2 k)$, $O(n^2m^{-1} \log_2 k)$, respectively. $H_{\text{node}}$ is lower than $H_{\text{tuple}}$ for the same number of grouped policies $k^2, k$.

3 Evaluation

3.1 Simulation conditions

To evaluate the performance of the complexity reduction, we set the number of IP address to $n = 30$, the number of port to $m = 10$. We simulated the attributes of users, devices, IP addresses, port numbers, user roles, resources.

In the evaluation scenario, we updated one of the device trust from high value to low. Trust in the whole network was re-evaluated and policies were updated by the PG. We used this as the original policy. We compared our proposed method
with the performance of the non-optimal node clustering method like [4] (Attribute-based node shrinkage), and with the other optimal strategy grouping tuples (Policy shrinkage).

3.2 Results

Figure 2 shows the result. We plotted the coarseness i.e. reproduction error rate $\frac{1}{mn^2} \sum_{x, x' \in X} D(R_x \rightarrow x', P_{\xi}(x) \rightarrow \xi(x'))$ while varying the complexity. The complexity is $H_{\text{node}}$ (Eq. (2)) for Attribute-based node shrinkage and the proposed method, and is $H_{\text{tuple}}$ (Eq. (3)) for policy shrinkage. Original policy granularity is the coarseness that only a single rule is different to the original policy ($\epsilon_{\text{max}} = 1$). Static RBAC (Role-based access control) is for reference which have two pre-defined user roles (groups). This is the coarsest policy.

Our method achieved the same policy as that of the original one (less than a single error) at the complexity 72, corresponding to eight number of groups. It is 89 percent lower than that of the attribute-based node shrinkage, and 81 percent lower than that of policy shrinkage. The proposed method is plotted in the lower-left side of the figure, indicating the best performance.

![Fig. 2. Complexity vs coarseness of grouped policies.](image-url)

4 Conclusions

This letter proposed the complexity reduction method of the systematically generated fine-grained policy for NAC. Our proposed method groups nodes whereas the objective describes the whole policy. The reason of the best performance is that the policy definition process is implicitly affected by two kinds of information, node specific attributes and other than that, e.g. conditional properties and trade-offs. The practicality of the method depends upon the policy generation, which will also be evaluated as the whole system.
Adaptive spreading factor assignment for VLC-OFDM-IDMA with PIC

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Abstract: This letter proposes a signal-to-Noise Ratio (SNR) based spreading factor adaptation for orthogonal frequency division multiple interleave division multiple access (OFDM-IDMA) incorporated with parallel interference cancellation (PIC) on visible light communication (VLC) systems. Orthogonal variable spread factor (OVSF) code is an effective solution to eliminate multiple access interference (MAI) in exchange for spectral resource consumption. Channel ranking based spreading factor optimization can maximize the system throughput performance. The major contribution of this letter is to achieve its adaptive allocation according to the estimated value of SNR. The computer simulation results clarify its effectiveness.

Keywords: VLC, OFDM, OVSF codes, channel ranking

Classification: Wireless Communication Technologies

References

1 Introduction

Visible light communication (VLC) has received significant attention in recent years as an alternative to wireless communication technology that operates in the RF band [1]. In VLC, optical signals are spatially multiplexed at the receiver side. Orthogonal frequency division multiplexing (OFDM) and interleave division multiple access (IDMA) have already been proposed [2]. To further improve the multiuser detection capability of OFDM-IDMA, an orthogonal variable spread factor (OVSF) codes [3, 4, 5] and parallel interference cancellation (PIC) [6] are supportive. OVSF codes spread the mapped symbol to some subcarriers with a specified sequence orthogonal to all users [7, 8, 9]. We have previously studied optimal combinations of spread factors based on channel gains for each user in order to maximize overall throughput performance [6]. For more practical use, this letter provides an adaptive spread factor allocation algorithm based on the estimated signal-to-noise power ratio (SNR). The combinations of spread factors are determined with reference to the predefined throughput–SNR characteristics. Simulative evaluation makes it possible to verify that the proposed system works and maximizes throughput performance.
2 System model: VLC-OFDM-IDMA with PIC

In VLC-OFDM, the means of transmission between the transmitter and receiver is visible light, so the transceiver needs to generate signals only in the amplitude range [2]. To support VLC modulation, the frequency-domain symbol sequence would have to be rearranged such that its time-domain transformation becomes a real-valued signal [10, 11]. Let $N_c$ and $X_u(k)$ $(k \in \{0, 1, \cdots, N_c - 1\})$ denote the number of subcarriers and the mapped symbol of the $u$-th user whose information data is interleaved with the specified pattern, OFDM symbol for intensity modulation (IM) can be generated as,

$$X_u(2N_c - k) = X_u^*(k),$$
$$X_u(0) = X_u(N_c) = 0,$$

where $(.)^*$ stands for the complex conjugate. The output signal after inverse discrete Fourier transform (IDFT) is expressed by

$$s_u(t) = \sum_{k=0}^{2N_c-1} X_u(k)e^{j2\pi kt/2N_c},$$

$s_u(t)$ is a time-domain OFDM having real values suitable for IM and visible light transmission. The photodetector receives the signal but it contains interference from other transmitters. The received symbol $r(k)$ at the $k$-th subcarrier is given as,

$$r(k) = \sum_{l=1}^{U} H_l(k)X_l(k) + n(k),$$

where $U$ is the number of users, $H_l(k)$ is the channel coefficient at the $k$-th subcarrier of the $l$-th user, and $n(k)$ is an additive white Gaussian noise (AWGN). Rewrite (4) for the $m$-th user as,

$$r(k) = H_m(k)X_m(k) + \sum_{l=1,l\neq m}^{U} H_l(k)X_l(k) + n(k),$$

$$= H_m(k)X_m(k) + \gamma(k).$$

Here, multiple access interference (MAI) is replaced with $\gamma$ and it can be removed by PIC [6]. As the number of users increases, the impact of MAI also becomes significant. Forward error correction (FEC) and OVSF code are generally utilized to solve this problem. In addition, adaptive optimization is the key technology to achieve high-quality communications [12]. This letter focused on the OVSF code and its adaptive optimization to maximize throughput performance.

3 Proposed framework

3.1 Orthogonal variable spreading code and channel ranking

In OVSF codes, orthogonality is guaranteed between the codes even with the different spreading factors. Doubling the spreading factor halves the information bits of data to be transmitted. Appropriate use of OVSF codes facilitates multiuser detection [13]. There is a trade-off between the MAI suppression capability and
throughput performance. We have already introduced channel ranking to assist the MAI suppression function to maximize the throughput performance [6]. When the influence of MAI is significant, a large spreading factor should be assigned and vice versa. Based on the estimated CSI, the path gain of each user is calculated and sorted in descending order. Optimal spreading factors are then assigned to each user.

3.2 Adaptive spreading factor assignment based on SNR estimation

Our previous study investigated that the possibility of determining the optimal spreading factors based on average SNR in users [6]. The key proposal of this work is to adaptively assign OVSF codes based on the predefined relationship between the spreading factors and SNR in order to maximize the system throughput performance. As shown in Fig. 1(a), the receiver determines the combination of spreading factor based on the estimated SNR and instantaneous channel gain of each user. The transmitters (users) then use the spread factor fed by the receiver.

Figure 1(b) shows the combination of spreading factors that can achieve the maximum throughput performance in each SNR region. From (4), estimate values of average SNR at the receiver side, $\Gamma_{\text{est}}$ can be calculated as,

$$\Gamma_{\text{est}} = \frac{1}{U} \sum_{k=0}^{N_c-1} |r(k)|^2.$$

(6)

In the proposed scheme, an appropriate spreading factor is assigned to each user according to instantaneous estimated SNR in (6) and predefined in Fig. 1(b).

4 Computer simulation

4.1 Simulation parameters

Table I lists the simulation parameters. The proposed scheme assigns spreading factors as 2, 4, 8, and 16 for each user and optimizes them based on estimated instantaneous SNR. The terminal bit of the convolution code is 64. The propagation environment is assumed to be the line-of-sight (LOS), and the channel coefficient, $H$, is defined as [1],

$$H = \frac{(e + 1)A}{2\pi d^2} \cos^e(\phi)T(\theta)G(\theta)\cos(\theta).$$

(7)
Table I. Simulation parameters.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Transmission scheme</td>
<td>VLC-OFDM-IDMA</td>
</tr>
<tr>
<td>Modulation</td>
<td>QPSK</td>
</tr>
<tr>
<td>Transmission bandwidth</td>
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<tr>
<td>Number of subcarriers</td>
<td>128</td>
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<tr>
<td>FFT size</td>
<td>128</td>
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<tr>
<td>Guard Interval</td>
<td>16</td>
</tr>
<tr>
<td>Number of pilot symbols</td>
<td>2</td>
</tr>
<tr>
<td>Number of data symbols</td>
<td>64</td>
</tr>
<tr>
<td>Channel model</td>
<td>Line of sight</td>
</tr>
<tr>
<td>Number of users</td>
<td>4</td>
</tr>
<tr>
<td>Number of PIC iterations</td>
<td>2</td>
</tr>
<tr>
<td>Error correction code</td>
<td>Convolutional code, rate 1/2</td>
</tr>
<tr>
<td>Spreading factor</td>
<td>2, 4, 8, 16</td>
</tr>
<tr>
<td>Input data bits</td>
<td>1888, 880, 376, 124</td>
</tr>
<tr>
<td>Interleaver type</td>
<td>Random</td>
</tr>
</tbody>
</table>

where $\phi$ is the angle of irradiance, $\theta$ is the angle of incidence, $T(\theta)$ and $G(\theta)$ denote the gains of an optical filter and concentrator, respectively. $e$ is the order of Lambertian emission. This evaluation assumes $T(\theta)$, $G(\theta)$, and $e$ are 1.0. Users are located in the distance $d$ from 1 to 4 m range, which provides the difference of channel gain among them. Angles $\theta$ and $\phi$ are with randomly determined in $0 \sim \pi/3$ rad. The evaluation metric is the throughput performance; one transmission frame is composed of 64 OFDM symbols.

4.2 Simulation results

First, Fig. 2(a) plots throughput performances versus transmit SNR with various spreading factor patterns, creating the spreading factor assignment table in Fig. 1(b). It shows the sum-rate performance of all four users. It demonstrates that the optimal

![Fig. 2. Simulation results.](image-url)
combination of spreading factors can be determined by the average SNR. The maximum-value and throughput performance in the low SNR region differ. The optimum spreading factor is assigned according to the SNR value.

Figure 2(b) then presents throughput performance obtained by the proposed scheme. The dashed line shows the envelope of the throughput values of Fig. 2(a) as a reference. This result verifies our proposed adaptive spreading factor assignment can maximize the throughput performance at given SNR conditions. Actual SNR varies depending on the users’ location. To maximize the throughput performance, it is necessary to assign the spreading factor according to the instantaneous SNR. The reference (dashed line) determines the spreading factor assignment fixedly as regard average SNR. On the other hand, the proposed method can refer to the instantaneous SNR. Therefore, it shows superior performance compare to the reference method in all SNR regions.

5 Discussion

Adaptive control methods based on average SNR are superior in terms of computational cost and are commonly used. Since this method cannot account for instantaneous SNR variations, there is still a potential for improvement in terms of maximum throughput. In particular, this letter deals with adaptive control in multiple users; thus, there is an instantaneous SNR variation for each user. In the proposed method, the improvement by considering the instantaneous SNR can be obtained for each user, and the throughput can be significantly improved.

The estimation error of SNR is an important factor that affects the communication performance. When adaptive control is performed using average SNR, it is strongly affected by this error because of the error enhancement that occurs during averaging. However, when the control is performed using instantaneous SNR, the error can be rounded off, and this effect can be minimized because the control is performed successively. In this letter, SNR used for the control is idealized, but from the above, the proposed method is effective even in an environment where estimation error exists.

6 Conclusion

This letter proposed the adaptive spreading factor assignment by the instantaneous SNR estimation and channel ranking to further improve the throughput performance of the VLC-OFDM-IDMA system. Simulated evaluation justified its effectiveness in the scenario of four users sharing the visible light communication channel. We can conclude that our proposed method is one of the most effective solutions for high system functionality.
A study on coherently combining sparse-multiband processing

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Abstract: The authors proposed Coherently Combining Sparse-Multiband Processing (CCSMP) [1], which coherently combines signals of the separated multi-band obtained by the stepped multiple frequency method [2]. CCSMP implements the Recursive Signal Subtraction Frequency Estimation Method [3], which iteratively estimates the range of each target while separating the signals one by one from the observed signals from multiple targets obtained in each band. Hence, CCSMP can improve the range estimation performance in multiple-target situations where the main lobes of the targets overlap and interfere with each other. In this letter, we derive the Cramér-Rao Lower Bound (CRB) for the target range estimation by combining the separated multi-band signals, and the performance of CCSMP is discussed by comparing the derived CRB values with statistical evaluations by simulation.

Keywords: sparse frequency bands, stepped multiple frequency method, coherent processing

Classification: Sensing

References

1 Introduction

Achieving an improvement of detection performance is an important research issue in radar technology. The authors have proposed Stepped Multiple Frequency CPC (complementary phase code) modulation [4], which combines the synthetic bandwidth method and pulse compression method, and have experimentally demonstrated that it is possible to achieve a high resolution with a low processing load while realizing long range detection with a narrow receiver bandwidth [5]. Furthermore, we proposed the Coherently Combining Sparse-Multiband Processing (CCSMP) method [1], which improves performance of range estimation in the presence of multiple targets by combining the signal outputs from the stepped multiple frequency CPC radars in separated non-coherent multi-band.

CCSMP employs the Recursive Signal Subtraction Frequency Estimation Method [3], in which the target range is estimated while subtracting unwanted signals other than the signals from targets. In consequence it can improve the performance for range estimation even when there is interference between multiple targets. In this letter, we derive the Cramér-Rao Lower Bound (CRB) of the target range estimation performance by combining the signals of separated multi-band. Additionally, the performance of CCSMP is evaluated by comparing the derived CRB values with statistical evaluations by simulation, especially when the spacing between multiple targets is narrow and the signals from these targets are interfering.

2 Coherently combining sparse-multiband processing

Assume that \( K \) target signals are received by the radar at \( N \) frequency steps from a range of \( R = [R_0, \cdots, R_{K-1}]^T \) and similar reception is performed in \( FN \) bands of different receiving frequency. \( z_{iF}(t) \) is the signal received at time \( t = 0, \cdots, L - 1 \) in the \( iF \)th band. \( z_{iF}(t) \) of all acquired bands is concatenated to form a signal \( z(t) = [z_0(t), \cdots, z_{FN-1}(t)]^T \) with vector length \( M = FN \cdot N \). This \( z(t) \) is given by

\[
z(t) = A(R)s(t) + e(t)
\]

where

\[
a(R_k) = \exp \left(-j4\pi \frac{n\Delta f}{c} R_k \right)
\]

where \( \Delta f \) is the step frequency interval, \( c \) is the speed of light, and \( n \) is the number of the step frequency for \( 0 \leq n \leq N - 1 \). Let \( F_{iF} \) be the frequency at step number

---

$n = 0$ in the $iF$th band. Also, let the complex gain/phase of the receiver at this frequency be $\beta_{iF}$. The complex amplitude $\alpha_{iF,k}$ of the received signal $z$ is given by

$$\alpha_{iF,k} = \beta_{iF} \cdot \exp\left(-j4\pi \frac{F_{iF}}{c} R_k \right)$$  \hspace{1cm} (3)$$

Let $s_k(t)$ be the waveform of the $k$th target signal at time $t$ with $s(t) = [s_0(t), \cdots, s_{K-1}(t)]^T$. Let $e(t) = [e_0(t), \cdots, e_{M-1}(t)]^T$ be an additive complex white Gaussian noise with mean zero and variance $\sigma^2$. We also assume that the noise is uncorrelated and independent of the signal. The received signal in the $iF$th band from the $k$th target is given by

$$y_{iF,k} = \alpha_{iF,k} \cdot a(R_k)s(t)$$  \hspace{1cm} (4)$$

and the signal $z(t)$ in (1) can be given as follows,

$$z(t) = \left[ I_{FN} \otimes a(R_0)s_0(t) \right] \cdots \left[ I_{FN} \otimes a(R_{K-1})s_{K-1}(t) \right] + e(t)$$

$$= H(R, t)\alpha + e(t)$$  \hspace{1cm} (5)$$

where “$I$” is an identity matrix and “$\otimes$” is the Kronecker product.

The target range $R_k$ can be estimated by estimating $\alpha_{iF,k}$ so that the phase in each $iF$th band is aligned.

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Fig. 1. Coherently combining sparse-multiband processing. (a) Signal reception by separate bands. (b) Complex amplitude of the received signal in each band.
3 Cramér-Rao lower bounds

The probability density function of \(z(0), \ldots, z(L-1)\) derived from (1) is as follows

\[
p(z(0), \ldots, z(L-1)|\mathbf{R}, \alpha, \{s(t)\}_{t=0}^{L-1}, \sigma^2)
= \frac{1}{\sqrt{(2\pi)^L \sigma^2 I_M^L}} \exp\left\{-\frac{1}{\sigma^2} \sum_{t=0}^{L-1} \|z(t) - A(R)s(t)\|_2^2\right\}
\]
(6)

After ignoring the terms independent of the unknown variables, the log-likelihood function by (6) is given as follows,

\[
\mathcal{L}(\mathbf{R}, \alpha, \{s(t)\}_{t=0}^{L-1}, \sigma^2) = ML \ln \sigma^2 - \frac{1}{\sigma^2} \sum_{t=0}^{L-1} \|z(t) - A(R)s(t)\|_2^2
= ML \ln \sigma^2 - \frac{1}{\sigma^2} \sum_{t=0}^{L-1} \|z(t) - \mathbf{H}(R,t)\alpha\|_2^2
\]
(7)

where \(\mathbf{R}\) and \(\sigma^2\) are real variables and \(\alpha\) and \(s(t)\) for different \(t\) are complex ones. The complex variables of \(\alpha\) and \(s(t)\) can be decomposed into real and imaginary parts, i.e., \(s(t) = s_R(t) + js_I(t)\) and \(\alpha = \alpha_R + j\alpha_I\), and the real variable set determining the measurements is \(\Omega = [R^T, \alpha_R^T, \alpha_I^T, (s_R^T(t), s_I^T(t))_{t=0}^{L-1}]^T\). The expression for CRB is obtained from the inversion of the Fisher information matrix (FIM) [6], which contains information about all the unknown parameters in \(\Omega\). Based on the log-likelihood function given in (7) and the FIM, straightforward mathematical derivations (detailed in the literature [6]) can be carried out to obtain the CRB for the target range estimation as follows,

\[
CRB_{R}^{-1} = \frac{2}{\sigma^2} \text{Re}\left( F_{RR} - F_{Ra} F_{aa}^{-1} F_{Ra}^H \right)
\]
(8)

where

\[
F_{RR} = \sum_{t=0}^{L-1} \left( X^H(t)D^H(R)\Pi_A^H D(R)X(t) \right)
\]
(9)

\[
F_{Ra} = \sum_{t=0}^{L-1} X^H(t)D^H(R)\Pi_A^H \mathbf{H}(R,t)
\]
(10)

\[
F_{aa} = \sum_{t=0}^{L-1} \mathbf{H}(R,t)\Pi_A^H \mathbf{H}(R,t)
\]
(11)

\[
X(t) = \text{diag}(\{s_0(t), \ldots, s_{K-1}(t)\}^T)
\]
(12)

\[
D(R) = \text{diag}\left( \frac{\partial a(R_0)}{\partial R}, \ldots, \frac{\partial a(R_{K-1})}{\partial R} \right)^T
\]
(13)

\[
\Pi_A^H = I - A(R)(A^H(R)A(R))^{-1} A^H(R)
\]
(14)

4 Numerical results

In order to evaluate the effectiveness of CCSMP, numerical simulations and performance comparison with the derived CRB values are performed. The simulation is based on the Recursive Signal Subtraction Frequency Estimation Method (detailed
in literature [1, 3]) and is performed as follows

$$\hat{\alpha}^{(p)} = \left( \sum_{t=0}^{L-1} H^H(\hat{R}^{(p-1)}, t) H(\hat{R}^{(p-1)}, t) \right)^{-1} \sum_{t=0}^{L-1} H^H(\hat{R}^{(p-1)}, t) z(t)$$  \hspace{1cm} (15)

$$\hat{s}(t)^{(p)} = \left( A^H(\hat{\alpha}^{(p)}, \hat{R}^{(p-1)}) A(\hat{\alpha}^{(p)}, \hat{R}^{(p-1)}) \right)^{-1} A^H(\hat{\alpha}^{(p)}, \hat{R}^{(p-1)}) z(t)$$  \hspace{1cm} (16)

$$\hat{y}_{iF,k}(t) = z_{iF}(t) - \sum_{i=0}^{K-1} a_i \hat{R}_{iF}^{(p-1)} \cdot \hat{\alpha}_{iF,i}^{(p)}$$  \hspace{1cm} (17)

$$\hat{x}_{iF,k}(t) = \hat{y}_{iF,k}(t)/\hat{\alpha}_{iF,i}^{(p)}$$  \hspace{1cm} (18)

$$\hat{R}_k^{(p)} = \arg \max_r \left[ \sum_{t=0}^{L-1} a(r)^H \hat{x}_{iF,k}(t) \hat{x}_{iF,k}^H(t) a(r) \right]/\left[ a(r)^H a(r) \right]$$  \hspace{1cm} (19)

where the subscript $^{(p)}$ on the right shoulder denote the value of the $p$th iteration.

The average root-mean-square error (RMSE) of the target range estimation of the three target signals ($K = 3$) is used for statistical range estimation performance evaluation, which is defined as

$$\text{RMSE} = \sqrt{\frac{1}{K} \sum_{k=0}^{K-1} \left( \frac{1}{N_{tr_y}} \sum_{tr_y=1}^{N_{tr_y}} \left( \hat{R}_k^{(tr_y)} - R_k \right)^2 \right)}$$  \hspace{1cm} (20)

where $N_{tr_y}$ denotes the number of simulations and we set $N_{tr_y} = 100$, and $R_k$ and $\hat{R}_k^{(tr_y)}$ are the true and estimated range of the $k$th target, respectively, and the amplitude and phase of the noise are varied in different trials. The simulations are performed assuming three targets set at $R = [12.5 \text{ m}, 12587 \text{ m}, 12.674 \text{ m}]^T$. And the number of frequency steps $N = 32$, the frequency step interval $\Delta f = 13.438 \text{ MHz}$, the band frequencies $F_{iF} = 79 \text{ GHz} + 500 \text{ MHz} \cdot iF$, and the number of snapshots $L = 1$ are assumed. The target spacing is 0.087 m, which is 1/4 of the resolution (0.349 m) of one bandwidth ($N \cdot \Delta f$) in which the three targets interfere with each other. Figure 2(a) shows the rerationship between SNR and RMSE in different number of combined bands, and comparison of RMSE by simulations and calculated CRB values. Figure 2(b) shows the rerationship between number of combined bands and
RMSE in different SNR, and comparison of simulated RMSE and calculated CRB values. In Fig. 2, the simulated RMSE and CRB values are in good agreement. Figure 2(a) shows that as the SNR increases, the RMSE decreases approximately in proportion to the SNR to the 1/2 power. Figure 2(b) shows that RMSE decreases as the number of bands combined by CCSMP increases. In particular, this RMSE has a large decrease in the number of combined bands from 1 to 2.

As another simulation, we compared the case where the total number of received signals is $TN$ and CCSMP. It is performed with $TN$ multi-band signals with $L = 1$ snapshots in each band, and the case where CCSMP is performed with $L = TN$ snapshots in one band. The SNR of each is 30 dB, respectively. Figure 3 shows that CCSMP can reduce the RMSE by using multiple bands, and CCSMP can improve target range estimation performance because the complex amplitude $\alpha_{iF,k}$ are estimated and combined so that the received signal $z_{iF}$ of each band can be coherently combined.

![Fig. 3. Comparison of RMSE and CRB by the number of received signals. The red line is from multiple snapshots by 1 band ($FN = 1, L = TN$). The blue line is from multiple bands by 1 snapshot ($FN = TN, L = 1$). The thick line is RMSE. The thin line is CRB.](image)

### 5 Conclusion

In this letter, we derive the Cramér-Rao Lower Bound (CRB) for the target range estimation by combining the separated multi-band signals, and the performance of CCSMP is showed by comparing the derived CRB values with statistical evaluations by simulation. The evaluation results show that CCSMP can improve the range estimation performance in multiple-target situations where the main lobes of the targets overlap and interfere with each other, and the simulated range estimation RMSE by CCSMP and the derived CRB values show good agreement.

It is expected that the range estimation RMSE can be reduced more by reducing the number of complex amplitude parameters of the received signal which do not directly affect the target range estimation, and this is a future research topic.
920MHz band propagation characteristics near metal ceiling for secure IoT communication

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Abstract: In this paper, we investigate the path loss characteristics near metal ceiling in 920 MHz band for secure wireless communication of indoor-use equipment such as lighting systems. In order to clarify the propagation mechanism, experiment and electromagnetic simulation are conducted using two models as use cases: a simple model with only a metal ceiling and an office with a metal ceiling. The results show that the path loss characteristics for vertical polarization near the metal ceiling has a smaller than the free space loss. The path loss characteristic for horizontal polarization, however, changes significantly depending on the floor condition.

Keywords: metal ceiling, indoor propagation, path loss, ray tracing

Classification: Antennas and Propagation

References

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1 Introduction

In recent years, with the growth of the IoT (Internet of things) market, the wireless systemization of various equipment such as lighting and ventilation system have been accelerating. The applications are, for example, remote control, monitoring, and data collection into the cloud and its analysis. The equipment is often used 920 MHz band in Japan since it has good propagation characteristics and less interference than the 2.4 GHz band which is also used for IoT devices.

In order to realize secure communication, it is important to predict the path loss characteristics in the environment where the equipment is actually used. In particular, the indoor-use equipment is generally installed near the ceiling as well as the access points connected to the equipment. Therefore, it is necessary to clarify the path loss characteristics including the influence of the ceiling material on the equipment.

In this paper, we investigate the path loss characteristics when both antennas are close to near metal ceiling. Metal ceiling is mainly used in non-residential buildings such as offices, hospitals, shopping malls, and airports, for its advantage that it can eliminate the bias of air conditioning with radiation effect generated by cooling or heating the ceiling chamber.

Various studies for the prediction of path loss within the same building floor have been conducted [1–3]. However, they do not clarify the specific propagation characteristics near metal ceiling. In [4], an empirical indoor propagation prediction model is standardized. However, it is difficult to predict the propagation characteristics accurately since parameters such as construction materials and structures are not taken into account. Furthermore, the propagation characteristics with respect to the polarization characteristics of the antenna have not been clarified.
In this study, the following two models are considered in order to clarify the propagation mechanism near metal ceiling in the 920 MHz band. Model 1 denotes the case only with a metal ceiling. This model assumes that the reflection from the metal ceiling is predominant propagation mechanism and the reflection from the floor can be negligibly small. The propagation characteristics of both vertical and horizontal polarizations (denoted as V-pol. and H-pol. respectively) for Model 1 are obtained by experiment and theoretical analysis based on image method. Model 2 denotes the case with an actual metal ceiling office with a floor. To clarify the propagation characteristics, an experiment was conducted in an actual office and numerical simulation using an equivalent simplified model. The simulation method used in the propagation prediction is the ray-tracing method, which is commonly used in site-specific propagation prediction [3, 5, 6].

2 Experimental site and simulation model

Figures 1(a) and 1(b) show the experimental site and corresponding simplified simulation model for Model 1.

We use a five-sided anechoic chamber for EMC (Electromagnetic Compatibility) test site showed Fig. 1(a) as an experimental site to eliminate the influence from other components of the metal ceiling. The metal floor of the anechoic chamber models a metal ceiling of the upside-down office. For the simulation model shown in Fig. 1(b), the ceiling is simply assumed to be a plane PEC (Perfect Electrical Conductor).
antennas at the Tx and Rx are half-wavelength dipole antennas, respectively. The distance from the ceiling to the antenna feed point is set to $h = 0.1 \text{ m} \ (\approx 0.3\lambda)$, which is the same for Tx and Rx, where $\lambda$ is the wavelength.

Figures 1(c) and 1(d) show the actual metal ceiling office and the corresponding simulation model for Model 2. In Fig. 1(d), the materials of the ceiling and the floor are assumed to be PEC, the distance between the ceiling and the floor is 2.5 m in order to match the conditions with the experimental site shown in Fig. 1(c). In the simulation model, concrete walls of the non-residential buildings are not considered because the experiment was conducted in the large room with low ceiling, 20 m (W) $\times$ 100 m (D) $\times$ 2.5 m (H). The conditions of Tx and Rx antennas are the same as in the case of Model 1. In both Model 1 and Model 2, the path loss characteristics are obtained by experiments and simulations with the Tx and Rx antennas placed in parallel to each other in both cases of V-pol. and H-pol. and $d$ is the distance between Tx and Rx. $d$ is set within the range 1 m to 20 m, which represents the typical communication range of IoT devices.

### 3 Experimental and simulation results

#### 3.1 Model 1: Metal ceiling only

Figure 2(a) shows the path loss characteristics of experiments and geometrical optics theory for Model 1. The experimental plot is the power average of 461 points measured within a radius of about 2 wavelengths at each $d$. The theoretical values are calculated using Eqs. (2), (3), and (4) described below. As shown in Fig. 2(a), the experimental and theoretical results are almost same. It is also found that the results of the V-pol. and H-pol. are quite different. The interpretation of the results is explained as follows. The model of Fig. 1(b) is the well-known two-path model. In this case, the path gain $G$ defined as the ratio of the transmitted power $P_T$ to the received power $P_R$ is given by the following equation [8]:

$$G = \frac{P_R}{P_T} = \left( \frac{\lambda}{4\pi d} \right)^2 G_T \cdot G_R |1 + R \cdot \exp( jk\Delta l)|^2 ,$$

where $G_T$ is Tx antenna gain, $G_R$ is Rx antenna gain, $R$ is the reflection coefficient of the ceiling, $k = 2\pi/\lambda$ is the wave number, $\Delta l$ is the path difference of direct wave
and reflection wave. $\Delta l$ can be expressed as the following equation:

$$\Delta l = \sqrt{(h_T + h_R)^2 + d^2} - \sqrt{(h_T - h_R)^2 + d^2} \approx \frac{2h_T h_R}{d},$$

where $h_T$ and $h_R$ are the distances between the ceiling and Tx antenna, and between the ceiling and Rx antenna, respectively. In this case, $h_T = h_R = h$. According to the image theory [9], reflection coefficient $R$ of V-pol. of the plane PEC is 1, and that of H-pol. is $\frac{1}{2}$. This is the reason of the difference for both polarizations. By considering $R$ for V and H-pol. and taking the reciprocal of Eq. (1) and substituting Eq. (2) into Eq. (1), the path losses for V-pol. and H-pol. in the condition of near metal ceiling can be approximated by the following equation:

$$L_V = \left(\frac{2\pi d}{\lambda}\right)^2 \cdot \frac{1}{\cos\left(\frac{k\Delta l}{2}\right)} \approx \left(\frac{2\pi d}{\lambda}\right)^2,$$

$$L_H = \left(\frac{2\pi d}{\lambda}\right)^2 \cdot \frac{1}{\sin\left(\frac{k\Delta l}{2}\right)} \approx \left(\frac{d}{\lambda}\right)^4.$$

In the applicable range of Eq. (3) and Eq. (4), the breakpoint of the two-path model is defined by following equation [8]:

$$d_{bp} = \sqrt{2kh_T h_R}.$$

Only range beyond breakpoint is discussed in this study because $h$ is sufficiently small. The breakpoint is 0.27 m when it is calculated by Eq. (5) in case of $h = 0.1$ m. Therefore, in the range of Fig. 2(a), the path loss exponent is $\alpha = 2$ for V-pol and $\alpha = 4$ for H-pol, respectively. Note that the path loss component is $\alpha = 2$ in the free space.

### 3.2 Model 2: Metal ceiling office

Figure 2(b) shows the path loss characteristics of the experimental and the simulation results of Model 2 for V-pol. and H-pol., respectively.

The experimental plot is the power average of 461 points measured within a radius of about 2 wavelengths at each $d$. The measurement in the office of Model 2 was performed in the distance range $d$ of up to 10 m due to the constraints of the experimental site. The simulation result is obtained by using the ray tracing [10]. The simulation model and conditions are shown in Fig. 1(d) and Table I, respectively. In order to investigate the effect of the metal floor, the ray tracing simulation is conducted with considering only the metal ceiling and floor. To be consistent with the measurement data, the simulation plots are the moving averages within the range of $\pm 2$ wavelengths along the observation line.

As shown in Fig. 2(b), the path loss for V-pol. is less than the free space loss as in the case of Model 1. It can be also seen that the experimental and simulation results are in good agreement. That is, the path loss is more dominated by the metal ceiling rather than the floor.

In contrast, the path loss for H-pol. is very different from that of Model 1 and is also smaller than the free space loss as in case of V-pol. It is considered that the wave...
reflected from the metal floor in Model 2 results in the small path loss of H-pol., as the guiding effect in the parallel flat plate consisting of the metal ceiling and floor. In case of H-pol. path loss measurement, the dipole antennas were horizontally set, as shown in Fig. 1(d). From this, radio waves with the same power were radiated to the floor and ceiling, due to omni-directional radiation pattern of the dipole antenna. Consequently, the guiding effect can be observed.

On the other hand, it was found from Fig. 2(b) that the waveguide effect of V-pol. is smaller than that of H-pol., since the wave reflected from the floor was small because of the radiation pattern of the vertically set dipole antenna.

In order to properly calculate the guiding effect in ray tracing simulation, the maximum number of reflections was set to be 100 shown in Table I [11].

### 4 Conclusion

This paper presented the path loss characteristics near metal ceiling in 920 MHz band. It was found from the experimental and theoretical results of the geometric optics for Model 1 that the path loss for V-pol. has different behavior from two-path model, and it is smaller than the free space loss.

It was also found from the experimental and simulation results for Model 2 that the path loss for V-pol. is the same as in Model 1. On the other hand, the path loss for H-pol. is very different from that of Model 1 and is also smaller than the free space loss as the guiding effect in the parallel flat plate consisting of the metal ceiling and floor. As a result, the path losses of both V and H-pol. are smaller than the free space loss.
Microwave 3D imaging of near-field scatterers mutually coupled with an antenna array

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Abstract: In this study, we propose a three-dimensional (3D) imaging method for near-field scatterers that are mutually coupled with an antenna array, such as the support structure or fixtures of the antenna, based on bistatic microwave synthetic aperture imaging. This method can be used to evaluate the effectiveness of the antenna design with irremovable near-field scatterers. Numerical simulation and measurement results show the validity of the proposed imaging scheme.

Keywords: antenna array, mutual coupling, 3D imaging

Classification: Sensing

References


1 Introduction

Mutual coupling refers to the electromagnetic interaction between antenna array elements. It is known to degrade the accuracy of high-resolution direction-of-arrival estimation algorithms. Furthermore, near-field scatterers around the antenna array, such as the support structure or fixtures, exhibit a similar undesired mutual coupling effect.

The scattering parameter (S-parameter) between two antenna elements is a commonly used metric to characterize the effect of mutual coupling [1]. A raster-scanning electric field probe [2] can be used to identify locations of intense radiation for near-
field scatterers. To overcome the requirement of the raster-scanning, [3] proposed a method based on two-dimensional (2D) synthetic aperture imaging using circular scanning. However, because realistic antennas have three-dimensional (3D) structures, considering an extension of the previous algorithm to a 3D case is important. Therefore, this paper proposes the following extensions: First, we employ cylindrical scanning instead of conventional circular scanning to obtain an image resolution in the height dimension. Second, a 3D spatial image is generated instead of the conventional 2D image.

We present the numerical simulation and measurement results to show the validity of the proposed scheme.

2 Problem formulation

2.1 System and signal model

Figure 1 shows the system model considered in this study. The receiving elements of an antenna array are distributed at known 3D spatial locations $r_m$, and the discrete near-field scatterers are located at unknown spatial locations $s_n$. Here, $m \in \{1, 2, \ldots\}$ and $n \in \{1, 2, \ldots\}$ are the indices representing the individual array element and scatterers, respectively.

As shown in Fig. 1, the transmitting antenna cylindrically moves around the antenna array, where the radius of the cylinder, the azimuth angle of the transmitting antenna location, and antenna altitude are $R_c$, $\phi$, and $z_c$, respectively. At each azimuth angle and height, the transmitting antenna transmits an electromagnetic wave, and the array elements receive the transmitted and scattered waves. This refers to a multistatic scattering experiment using a single transmitting antenna and multiple receiving antennas.

We define the distance between the transmitting antenna at $(\phi, z_c)$ and a spatial point $p = (x, y, z)$ and the distance between two arbitrary points, $p$ and $q$, as follows:

$$d(\phi, z_c; p) = \sqrt{(R_c \cos \phi - x_p)^2 + (R_c \sin \phi - y_p)^2 + (z_c - z_p)^2} \quad (1a)$$

$$d(p, q) = \sqrt{(x_p - x_q)^2 + (y_p - y_q)^2 + (z_p - z_q)^2} \quad (1b)$$

$$p = (x_p, y_p, z_p), \quad q = (x_q, y_q, z_q). \quad (1c)$$

When the transmitting antenna at $(\phi, z_c)$ transmits an electromagnetic wave of an angular frequency $\omega$, we denote the signal received by the $m$th array element as $S_m(\omega; \phi, z_c)$. The received signal $S_m(\omega; \phi, z_c)$ comprises the direct wave from the transmitting antenna to the $m$th array element, denoted by $S_m^d(\omega; \phi, z_c)$, and

![Fig. 1. System model of a 3D antenna array with near-field scatterers.](image-url)
the scattered wave from the scatterer at $s$, denoted by $S_m^s(\omega; \phi, z_c)$. Note that the scattered wave includes those generated by the receiving elements themselves, i.e., the wave corresponding to mutual coupling.

Assuming that the array elements exhibit an isotropic antenna pattern, the signal received by the $m$th array element can be expressed as

$$S_m(\omega; \phi, z_c) = S_m^d(\omega; \phi, z_c) + S_m^s(\omega; \phi, z_c) \quad (2a)$$

$$S_m^d(\omega; \phi, z_c) = \frac{A(\omega; \phi, z_c; r_m)}{d(\phi, z_c; r_m)} \exp[-j k d(\phi, z_c; r_m)] \quad (2b)$$

$$S_m^s(\omega; \phi, z_c) = \int_{r_m} A(\omega; \phi, z_c; s) \frac{f(\omega; \phi, z_c; s, r_m)}{d(s, r_m)} \exp[-j k d(s, r_m)] ds \quad (2c)$$

where $A(\omega; \phi, z_c; \cdot)$ is the complex amplitude pattern of the transmitting antenna, $f(\omega; \phi, z_c; s, r_m)$ is the scattering coefficient of the scatterer related to the bistatic radar cross-section, $\Omega$ is the spatial domain of the integration, and $k = \omega/c$ ($c$ is the speed of wave propagation) is the wavenumber.

### 2.2 3D image reconstruction

Considering that the propagation phase corresponding to the wave from the transmitting antenna which is scattered at the location $p$ and received by the array element at $r_m$, image reconstruction can be achieved using the following integral transformation of the received signal.

$$g_m(p) = \int_{-\infty}^{\infty} \int_{0}^{2\pi} \int_{0}^{\infty} S_m(\omega; \phi, z) \exp[j k d(\phi, z; p)] \cdot \exp[j k d(p, r_m)] \omega d\omega d\phi dz \quad (3)$$

For each element ($m = 1, 2, \ldots$), we repeat the calculation expressed in Eq. (3) and obtain the corresponding spatial image. The final image can be created by coherently summing these images as follows:

$$g(p) = \sum_m g_m(p) \quad (4)$$

### 2.3 Direct wave filtering

The direct wave component must be suppressed for successful image reconstruction of the scattered wave, as stated in the literature [3], because the direct wave is generally much stronger than the scattered wave of interest.

To filter out the direct wave component from the total wave, we first remove the phase and attenuation factors corresponding to direct wave propagation, then compute the Fourier transform of the resultant signal with respect to azimuth angle $\phi$, as follows:

$$\tilde{S}_m(\omega; \xi, z_c) = \mathcal{F}(\phi) [S_m(\omega; \phi, z_c) d(\phi, z_c; r_m) \exp[j k d(\phi, z_c; r_m)]] \quad (5)$$

where $\mathcal{F}(\phi) \{\cdot\}$ is the Fourier transform with respect to the azimuth angle $\phi$, and $\xi$ represents the transformed domain.
Because the direct wave component in the transformed domain concentrates at \( \xi = 0 \), the direct wave can be suppressed by filtering the response around \( \xi = 0 \) \cite{3}. The filtered data are then transformed back to the original \((\omega; \phi, z_c)\) domain, and the propagation phase and attenuation are recovered. This operation is expressed as

\[
S'_m(\omega; \phi, z_c) = \mathcal{F}^{-1}_{(\xi)} \left[ S_m(\omega; \xi, z_c)H(\xi) \right] \frac{\exp[-jkd(\phi, z_c; r_m)]}{d(\phi, z_c; r_m)}
\]

where \( \mathcal{F}^{-1}_{(\xi)}[\cdot] \) represents the inverse Fourier transform with respect to the variable \( \xi \) and \( H(\xi) \) is the filter function represented as follows:

\[
H(\xi) = \begin{cases} 
0, & \text{for } |\xi| < \xi_0 \\
1, & \text{otherwise}
\end{cases}
\]

where \( \xi_0 \) is an positive constant that determines the notch width of the filter.

3 Simulation and experiment

3.1 Overview

The simulation model and measurement setup are shown in Figs. 2(a) and (b), respectively. The numerical simulation was performed using a numerical electromagnetic solver based on the method of moments. A four-element uniform linear array was used as the receiving antenna. The receiving elements were half-wavelength (at the center frequency) dipoles for the simulation and collinear antennas for the measurement. The operating frequency was 1.09 GHz, and the bandwidth was 400 MHz.

An infinitesimal current element was used as a transmitting antenna in the simulation, and a double-ridged guided horn was used for the measurement. The dipoles received radiated waves from the current element in the simulation, and induced currents on the centers of the dipoles were stored for image reconstruction. In the measurement, the horn antenna and the receiving elements were connected to the vector network analyzer.

The radius and height of the antenna-scanning cylinder were 3 and 1 m, respectively, with the antenna altitude ranging from -50 to 50 cm. The angular and height sampling intervals were 4° and 10 cm, respectively. In the simulation, the receiving elements and near-field scatterers were fixed, whereas the current element was moved over the antenna-scanning cylinder. In the experiment, the antennas and scatterers under test are rotated by a turntable, and the transmitting horn antenna...
was moved vertically to form a cylindrical scanning surface. As shown in Fig. 2, four scatterers were placed around the receiving elements.

### 3.2 Results and discussion

Figure 3 shows the reconstructed 3D image, where (a) and (b) represent the simulated images and (c) and (d) represent the experimental images; (a) and (c) show the $(x, y)$-plane images at $z = 0$, and (b) and (d) show the $(x, z)$-plane images at $y = -0.4$ m. These images were created by the formulation presented in Section 2. The solid line represents the location of the near-field scatterers. From these figures, clear peaks can be observed at the scatter location $s_n$ and at the location corresponding to the array element.

By comparing Figs. 3(a) and (d), we can observe from Fig. 3(d) that the image intensity of the scatterer on the right side is stronger than that of the scatterer on the left side, whereas the intensities are similar in the simulated image of Fig. 3(b). This result suggested that the strength of the mutual coupling between the right scatterer and the receiving elements was stronger than that between the left scatterer for the antennas used in the experiment. Thus, it was better to lower the near-field scatterer on the right side to reduce mutual coupling. In contrast, the effect of the scatterer height on the mutual coupling was minimal for the dipoles in the simulation. From these results, we can confirm that the proposed 3D imaging could determine a better height alignment for the scatterers.

![Fig. 3. Reconstructed image. (a) $(x, y)$-plane image at $z = 0$ (simulation). (b) $(x, z)$-plane image at $y = -0.4$ m (simulation). (c) $(x, y)$-plane image at $z = 0$ (experiment). (d) $(x, z)$-plane image at $y = -0.4$ m (experiment).](image)

### 4 Conclusion

We proposed a microwave 3D imaging method for near-field scatterers in the vicinity of an antenna array in this study. The method is validated using numerical simulation and measurement. The results show that the location and the scattering intensity of the scatterers can be identified; thus, the technique can be used to design and evaluate antennas which can reduce mutual coupling.
Majority vote reward scheme improves spatial location identification tasks

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Abstract: We evaluate accuracy and reliability in a spatial location identification task (SpLIT) when workers know that their reward scheme is a majority vote (MV) reward scheme before processing SpLIT and when workers do not know that their reward scheme is MV before processing SpLIT. This evaluation was made to clarify how MV affects accuracy and reliability in SpLIT. When MV motivates workers, accuracy and reliability will be different in each condition. The answers of workers who knew that their reward scheme was MV before processing SpLIT had lower accuracy but higher reliability than the answers of workers who did not know this. We concluded that MV influences SpLIT by motivating workers.

Keywords: SpLIT, crowdsourcing, reward scheme
Classification: Navigation, Guidance and Control Systems

References


1 Introduction

We clarified how a majority vote (MV) reward scheme affected accuracy and reliability in a spatial location identification task (SpLIT). In this paper, performance is defined as accuracy and reliability. In SpLIT, participants identify the location from where a camera view originates in the floor map of a facility. In SpLIT, it is assumed that processing is done by people, not computers, indicating that SpLIT could be
processed in crowdsourcing environments [1]. Crowdsourcing is the outsourcing of a piece of work to a crowd of people via an open call for contributions [2], and it is almost always used for processing done by people without special skills or knowledge, called workers. MV is a reward scheme for crowdsourcing workers. A reward scheme is the condition motivating a worker to obtain rewards in crowdsourcing, and in MV a worker is rewarded when his or her answer is close to the answers of other workers. The previous study [1] argued that MV improved performance in SpLIT, but it failed to clarify how such improvement was achieved.

SpLIT can solve the problem of identifying locations for navigation systems, and studies have been conducted on this [1, 3]. Although identifying the exact locations of pedestrians is needed for developing navigation systems, identifying locations indoors is complicated by insufficient landmarks and the low accuracy of global positioning systems (GPSs). Many automated systems have solved the identification of locations indoors, but most require predefined labels or expensive equipment such as RGBD sensors or lasers to scan environments. Our main motivation in this study is to exploit human workers who can process SpLITs without predefined labels or expensive equipment. Consequently, processing SpLIT can save time and reduce the cost of identifying indoor locations while outperforming methods that use only computers.

In SpLIT, workers identify the location from where a camera view originates in a two-dimensional floor map in the following three steps:

1. detect cues from a camera view (Fig. 1 (c));
2. map the detected cues to features on a floor map (Fig. 1 (a));
3. identify the location corresponding to the camera view.

Cues provide information for workers to identify the camera location on a floor map. In this study, cues are explicitly shown as text labels.

In a previous study [1], Rao et al. clarified that MV improved performance in SpLIT without specifying how such improvement was achieved. MV may improve task performance by motivating workers or filtering spammers [4], and it only motivates workers when those workers know that their reward scheme is MV before processing SpLIT. Therefore, we compared task performance when those workers know that their reward scheme is MV before processing SpLIT with task performance when those workers did not know this to clarify whether MV improved performance in SpLIT by motivating workers or by filtering spammers.

2 Previous study

In a previous study [1], Rao et al. measured accuracy and reliability when workers processed SpLIT in MV and in a ground truth (GT) reward scheme to explore how these two reward schemes would impact accuracy and reliability. GT is a reward scheme through which a worker is rewarded when the worker’s answer is close to the correct answer (Fig. 2, right). Rao et al. created two sets of instructions for experimental conditions to explain the two reward schemes. Workers were randomly assigned to one of the two reward schemes and given the reward scheme’s instructions before processing SpLIT. The answers of workers in MV showed higher
(a) Floor map

(b) Summary of camera views designed with different combinations of cues

<table>
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<th>Corners</th>
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(c) Camera views 1 to 5

**Fig. 1.** Object information shown to workers
accuracy and reliability than those in GT. Accuracy is the ratio of correct answers to total answers. Reliability is the distance between the correct location and the pin’s location indicating the worker’s answer.

Since the study of Rao et al. set a condition where the workers were informed of the assigned reward scheme before processing SpLIT, we cannot identify whether MV improved performance in SpLIT by motivating workers or by filtering spammers. To clarify this issue, we also evaluated accuracy and reliability by having participants process SpLIT without informing them of the assigned reward scheme before processing SpLIT.

3 Experiment

To evaluate the accuracy and reliability of participants’ answers, we published SpLIT on Amazon Mechanical Turk (MTurk) with three reward schemes: MV, GT, and no prior notice of reward scheme (NP). MTurk is a crowdsourcing platform used in large-scale crowdsourcing environments. Publish means to request workers on a crowdsourcing platform to process a task. In our experiment, the workers are requested to estimate the locations from which five camera views originated and to apply pins at these locations on the floor map. Outputs of the workers are the pin locations and explanatory text giving reasons for deciding on these locations. MTurk’s reward for every worker was $0.15 regardless of the workers’ reward schemes and pin locations. This condition probably did not affect the motivation of workers because they knew that they could still obtain rewards regardless of the reward schemes and pin locations after processing SpLIT. We published the tasks for 8 days, and 33 workers answered in MV, 35 in GT, and 34 in NP. To ensure quality control of the answers, we conducted preprocessing to exclude workers who met the following two conditions:

(a) workers with duplicate worker IDs;
(b) workers who pointed to almost the same location in all five camera views.

The number of workers for (a) was 2, and that for (b) was 1. In the case of (a), the first answer was adopted. After preprocessing, performances of the 31 workers answering in MV, the 34 in GT, and the 34 in NP are discussed in the next section.

In this experiment, we regarded any answer that corresponded to any view seen from a correct location as a correct answer. In order to determine the views seen from correct locations, we considered the width of the space captured by the camera.
view. We regarded any answer that corresponded to a location within $20\sqrt{3}$ pixels from a correct location as a correct answer.

4 Results

First, in terms of the relationship between the number of cues and accuracy, camera view 4 showed higher accuracy than camera view 3 (Fig. 3(a)). A chi-square test was performed, and the results show that the difference in accuracy between camera view 3 and camera view 4 was significant, where chi-square value $= 35.21$, $p < 0.005$. Camera view 4 provides fewer cues than camera view 3. The following observation results were grasped from the explanatory text giving reasons for deciding the pins’ locations for camera views 3 and 4 under the three reward schemes:

- Camera view 3: Many workers’ answers were excluded from being the correct location among the options because no cues were shown that were difficult to see from this camera view.

![Fig. 3. Experiment results](image-url)
Camera view 4: This camera view offered distinctly different cues than the other locations.

The accuracy was affected by the amount of semantic information that could be obtained from the cues, not the number of cues. Here, semantic information is defined as information that ranges from commonsense knowledge to domain-specific knowledge [1].

Second, in terms of the relationship between reward scheme and accuracy, MV generally showed the lowest accuracy, except for camera view 5 (Fig. 3(b)). A chi-square test was performed, and the results show that the difference in accuracy between MV and NP was significant, where chi-square value = 7.82, \( p < 0.005 \). The accuracy of MV is significantly lower than that of NP. We concluded that MV acted as a negative motivation and thus lowered accuracy.

Finally, in terms of the relationship between reward schemes and reliability, MV showed the minimum gap and the highest reliability (Fig. 3(c)). This is because it was necessary to determine the answer locations after anticipating what other workers would answer. Accurate workers responded after giving the task more consideration than workers under the other reward schemes. Accordingly, MV acted as a positive motivation for workers who answered correctly.

5 Conclusion

We evaluated accuracy and reliability in SpLIT when workers knew that their reward scheme was MV before processing SpLIT and when those workers did not know this. Our aim was to clarify how MV affected accuracy and reliability in SpLIT. Consequently, the workers who knew that their reward scheme was MV before processing SpLIT had lower accuracy but higher reliability than the workers without such knowledge in advance. We concluded that MV influenced the performance in SpLIT by motivating workers.
Near-field characteristics of a TEM horn used for radiated immunity tests

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Abstract: We have fabricated a transverse electromagnetic (TEM) horn for radiated immunity tests in close proximity as specified in standard IEC 61000-4-39 and compared its near-field characteristics with those of other typical test antennas such as broadband dipole and double-ridged guide (DRG) horn antennas. Experimental results show that the TEM horn generates a homogeneous field and maintains the field strength without a rapid change near the antenna.

Keywords: broadband dipole, immunity, radiated immunity test in close proximity, phase center, TEM horn, wave impedance

Classification: Electromagnetic Compatibility (EMC)

References

1 Introduction

Radiated immunity tests related to electromagnetic fields of electronic equipment or systems have been performed under far-field conditions. Recently, with the widespread use of portable wireless devices such as mobile phones, immunity requirements for electronic equipment with the aim of ensuring protection against portable transmitters used in close proximity have been specified in international standards [1, 2] and manufacturer standards. In near-field radiated immunity tests, TEM horns and dipole-like antennas such as sleeve antennas, monopole antennas, and broadband dipoles are used in accordance with test standards. The field uniformity and the propagation characteristics in proximity to the antenna affect the field strength reaching the active electronic components inside the electronic equipment depending on used antenna, even when the field strength is specified in the illuminated surface of the electronic equipment.

In this study, we fabricated a TEM horn that complies with IEC standards for proximity immunity tests [1]. The field characteristics of the TEM horn near the antenna were compared with those of the flat broadband dipole used in near-field immunity tests in automotive industry [2], the half-wave dipole used as a basic antenna, and the DRG horn widely used in EMC measurements. In addition, the far-field approximation from the characteristics of propagation over a distance and the wave impedance were evaluated considering the phase center.

2 TEM horn

The radiated immunity test in close proximity is carried out in a frequency range of 380 MHz to 6 GHz, which is used in portable transmitters such as mobile phones. The field-generating antenna is placed in a position 100 mm from the front of the equipment under test (EUT) and illuminates it with a specified field strength of 10 V/m to 300 V/m [1].

A shortened exponentially tapered TEM horn [3] we fabricated is shown in Figs. 1(a) and (b). The plate has a tapered structure designed to maintain the characteristic impedance of the exponential taper transmission line considering the matching from the feeding section to the aperture. The antenna length was decreased by 10% to improve the radiation directivity. Although the TEM horn is a balanced antenna, the leakage current flowing outside the outer conductor of the coaxial cable is suppressed using the balanced feeding mechanism without using the balun circuit [3, 4]. The 0.6-mm-thick plates of the antenna were made of brass, and the spacing between them was maintained using an expanded hard foam with a relative dielectric constant of 1.07.

2.1 Phase center

The phase center, which is defined as the center of curvature of the equiphase front in the far-field region, is effective for antenna measurements [5]. The mean of the phase centers of the electric (E) and magnetic (H) planes, \( d_{pc} \), coincides with the location of the amplitude center [6]; thus, it can be treated as an equivalent point source.
2.2 Wave impedance

The wave impedance of the electromagnetic wave radiated from the antenna is equal to the free space impedance of 377 Ω when it is sufficiently far from the antenna. In other words, as one of the conditions for the far-field region, the wave impedance should satisfy the free space impedance. The wave impedance on the axis of the antenna close to the antenna was calculated as the ratio of the E- to H-fields using the FIT solver. The calculation result of the wave impedance of the TEM horn is shown in Fig. 1(d) and compared with that of half-wave dipoles in the same figure. Although the wave impedance converges toward a constant value of 377 Ω as the distance from the antenna increases, it in close to the antenna differs depending on the frequency and type of the antenna. The wave impedance of the TEM horn does not show a rapid change as those of dipoles.
3 Experiments and results

The field characteristics in proximity were evaluated for the TEM horn and other typical EMC antennas, which are half-wave dipoles, the flat broadband dipole used in near-field tests for vehicles [2], and the DRG horn often used in EMC measurements. The experimental setup is shown in Fig. 2(a). A field-generating antenna and a single-axis E-field probe were separated by the distance \( r \) in a full anechoic chamber. The frequency-selective E-field probe with an optical fiber link was connected to a vector network analyzer (VNA). The E-field distribution on a plane of 400 mm \( \times \) 400 mm was measured along the antenna axis from 50 mm to 400 mm (= \( r \)) by scanning with the probe using an XYZ positioner at typical frequencies of 930 MHz, 2.45 GHz, and 5.8 GHz used for the immunity tests. The dimensions (\( W \), \( H \), and \( L \)) of the DRG horn used for comparisons were 244 \( \times \) 159 \( \times \) 279 mm. The broadband dipole had the flat element with dimensions of 109 \( \times \) 240 mm, and the element lengths of half-wave dipoles were 175 mm, 67 mm, and 29 mm.

Figures 2(b) to (d) show the measured results of the propagation characteristics along the axis of the antenna. The received field level was normalized at 100 V/m at a distance of 100 mm, which is the condition used in the proximity immunity tests. The measured values of the TEM horn and half-wave dipole are in good agreement with the results calculated using the FIT solver. These results indicate

![Fig. 2.](image)

(a) Experimental set-up. Field characteristics of TEM horn, half-wave dipole, flat broadband dipole, and DRG horn at (b) 930 MHz, (c) 2.45 GHz, and (d) 5.8 GHz.
that the field characteristics along the distance in close proximity to the antenna greatly differ among the antenna types. That is, the field strength to which the active electronic circuits are exposed inside the EUT changes depending on the antenna, even when the field strength was specified at the surface of the EUT. The TEM horn in close proximity showed field characteristics with a smaller reduction in field strength than other antennas, especially dipoles. For example, the distances where the field strength decreases by 4 dB from the specified position \((r = 100 \text{ mm})\) at 2.45 GHz are 209 mm, 54 mm, 68 mm, and 127 mm for the TEM horn, flat dipole, half-wave dipole, and DRG horn, respectively.

The far-field approximation by curve fitting for the TEM horn and half-wave dipole is plotted in Figs. 2(b)–(d). In the far-field region, the field strength decreases in inverse proportion to the distance \(r\) from the antenna. Therefore, as the second condition for the far-field region, the field strength along the distance from the antenna should fit the curve using \(1/r\). In particular, the length of the TEM horn cannot be ignored with respect to the measurement distance; thus, curve fitting was performed using \(1/(r + d_{pc})\) considering the phase center. The positions \((d_{pc})\) of the phase centers were 76 mm, 348 mm, and 313 mm inside from the aperture at 930 MHz, 2.45 GHz, and 5.8 GHz, respectively, as shown in Fig. 1(c). The minimum far-field distance was determined by curve fitting, and then the wave impedance was compared with that of 377 \(\Omega\) at that distance. The wave impedance of the TEM horn and half-wave dipole at a distance that can be regarded as the far field obtained by curve fitting is close to 377 \(\Omega\), as shown in Fig. 1(d) and Figs. 2(b)–(d). For example, the minimum far-field distances of the TEM horn are 350 mm at 930 MHz, 200 mm at 2.45 GHz, and 50 mm at 5.8 GHz, and the wave impedances at these distances are 368 \(\Omega\), 386 \(\Omega\), and 368 \(\Omega\), respectively. The fields at the location where the EUT is placed in the proximity test behave similarly to the near- or far-field region depending on the test frequency and the location when evaluating the field characteristics near the antenna under two far-field conditions on the basis of the propagation characteristics and the wave impedance. However, the far-field distances estimated on the axis of the antenna are different from those under the well-known far-field condition, which is based on the plane wave, of \(2D^2/\lambda\) (where \(D\) is the aperture size and \(\lambda\) is the wavelength) for aperture antennas, as the fields around antenna are spherical near the antenna.

Figure 3 shows the measured results of the field uniformity of the TEM horn and flat broadband dipole. The TEM horn has a small reduction for the given field strength even near the antenna, for example, the field strength at the distance of 400 mm at 2.45 GHz is more than 10 dB higher than that of the flat broadband dipole. Although the flat broadband dipole has two beams spread at 5.8 GHz, which results in lower strengths on the axis of the antenna as shown in Fig. 2(d), the TEM horn generated large homogeneous field areas over a wideband compared with the other antennas. Even near the antenna, the field strength of the TEM horn did not change as rapidly as those of dipole-like antennas.
4 Conclusion

In near-field immunity tests, TEM horns and various dipole-like antennas are used; however, antenna characteristics affect the test results. The field characteristics near the antenna were compared between the TEM horn that we fabricated and other typical test antennas, namely, a flat broadband dipole, a half-wave dipole, and a DRG horn. The results showed that the propagation characteristics greatly differ among the antennas. In particular, the TEM horn generated a homogeneous field and maintained the field strength with small reduction, whereas the dipole-like antennas showed rapid changes in field strength. In addition, the far-field approximation was performed considering the phase center using the propagation characteristics and the wave impedance, and the behavior of near- or far-field characteristics in proximity of antenna was clarified.