Special Cluster on Antennas and Propagation Technologies 2019

The objective of this special cluster is to report the advanced technologies on the antenna and propagation related to progressing technologies for 5G mobile communication systems, MIMO, PAN/BAN, and wireless power transfer and so on. In 2019, several conferences (AWAP2019 in Korea, IJAWT2019 in Indonesia, MJWRT2019 in Malaysia and ISAP2019 in China) were held. This special cluster has been published to raise the interest of the researchers in the field of antennas, propagation and the related topics.

This is the second special cluster in IEICE Communications Express (ComEX). Since review decision in the ComEX has no conditional acceptance (i.e. Acceptance or Rejection only), this cluster provided two submission deadlines as the same as the first cluster to make a chance for revised submission. For this cluster, 11 and 14 letters were submitted by the first and the second deadlines. After careful review, 11 letters were accepted in total.

I am grateful to the Editorial Committee of ComEX for deciding to publish this special cluster and their continuous support. I would like to express my sincere appreciation to all authors for their excellent contributions, and to the reviewers and editorial committee members for their great effort to make this second special cluster successful.

Finally, the third cluster in ComEX; “Special Cluster in Conjunction with IEICE General Conference 2020” has been already planned. Please refer the call for paper in ComEX web site.

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Comparison of SNR and channel capacity with micro and milli-meter wave bands based on outdoor propagation measurement

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Abstract: The 5-th generation mobile communication system (5G), which is the next generation communication system, is being developed. In 5G, it is assumed that multiple cells with different frequency bands will be used. The attenuation of signal power due to propagation loss in free space and the reflection of radio waves from buildings, etc., differs depending on the frequency. In this paper, a radio propagation measurement is performed for micro and milli-meter waves (5 and 19 GHz bands), which will be used in 5G. The outdoor propagation characteristics of each frequency band based on the measured channel state information are evaluated in terms of the signal to noise ratio and channel capacity.

Keywords: 5G system, micro wave, millimeter wave, outdoor propagation, propagation pathloss, channel capacity

Classification: Antennas and Propagation

References

[1] Cisco Systems Inc., “Cisco visual networking index: Global mobile data traffic forecast update, 2016–2021 white paper,” Mar. 2017.
[2] H. Papadopoulos, C. Wang, O. Bursalioglu, X. Hou, and Y. Kishiyama, “Massive MIMO technologies and challenges towards 5G,” *IEICE Trans. Commun.*, vol. E99-B, no. 3, pp. 602–621, Mar. 2016. DOI:10.1587/transcom.2015EBI0002
[3] E. G. Larsson, O. Edfors, F. Tufvesson, and T. L. Marzetta, “Massive MIMO for next generation wireless systems,” *IEEE Commun. Mag.*, vol. 52, no. 2, pp. 186–195, Feb. 2014. DOI:10.1109/MCOM.2014.6736761
[4] R. Taniguchi, K. Nishimori, R. Kataoka, K. Kameyama, K. Kitao, N. Tran, and T. Imai, “Evaluation of massive MIMO considering real propagation characteristics in the 20-GHz band,” *IEEE Trans. Antennas Propag.*, vol. 65, no. 12, pp. 6703–6711, 2017. DOI:10.1109/TAP.2017.2754441
1 Introduction

In recent years, there has been a significant increase in the volume of wireless traffic due to the spread of wireless communication devices such as smartphones. In particular, between 2016 and 2021, the total wireless traffic throughout the world is expected to increase sevenfold [1]. It is difficult for the current communication technology to cope with this increase in traffic volume. Therefore, the research and development of wireless communication technology that can realize high speed and high efficiency in a limited frequency band is essential. The 5-th generation mobile communication system (5G), which is the next generation communication system, has been developed in this regard. In 5G, multiple cells with different frequency bands are used to improve frequency utilization efficiency. The propagation loss in free space and the attenuation of signal power due to reflections and refractions on buildings, plants, etc., are different depending on the frequency band.

Until now, experiments have been reported in the 2 to 5 GHz band for 5G [2, 3]. Also, the direction of arrival of the microwave band was indicated in our previous study [4]. In this report, radio propagation measurement is performed for micro and milli-meter waves (5.12 and 19.55 GHz band), which are expected to be used in 5G. The outdoor propagation characteristics of each frequency band based on the measured channel state information (CSI) are evaluated in terms of the signal to noise ratio (SNR) and channel capacity.

2 Device configuration and measurement environment

At the transmitter, first, a PC generates a baseband orthogonal frequency duplex multiplexing (OFDM) signal. Then, the signal is D/A converted, quadrature modulated, and frequency converted to 2.425 GHz(L.O.1). Here, this device can be used even in the 5.12 GHz frequency band. Then, this signal is converted to 19.55 GHz(L.O.2), power amplified, and transmitted from the antennas. Also, by switching the L.O.1 frequency from 2.425 GHz to 5.12 GHz, the device can be operated at 5.12 GHz.

The signal received by the receiver is down converted from 19.55 GHz to 2.425 GHz(L.O.3). Even at the receiving side, evaluation of the 5.12 band is possible by switching the L.O.3 frequency from 2.425 GHz to 5.12 GHz. The signal, which is frequency converted from 2.425 GHz, is configured to be transferred to a PC after quadrature demodulation and A/D conversion.

The OFDM signal used consists of a short preamble and a long preamble. First, the short preamble is used to detect the beginning of the signal by autocorrelation processing. Then, CSI is estimated using the long preamble. Since each antenna has different signal transmission timing, it can obtain $4 \times 4$ channel matrix.

The parameters are as follows. The center frequencies are 5.12 GHz and 19.55 GHz, and the four transmitting and receiving antennas are sleeve antennas. The signal band is 20 MHz, the number of bits for A/D conversion and D/A conversion are both 12 bits, and the transmission power is 27 dBm.

In order to obtain the SNR and channel capacity in the outdoor small cell environment, the radio wave propagation measurement is performed on the uplink using center frequencies of 5.12 GHz and 19.55 GHz. In the measurement, the
The transmitter is moved linearly, and the signals are transmitted and received at intervals of 2 m.

Fig. 1 shows a diagram of the measurement environment; it shows courses 1 to 5 (with the transmitter moved linearly) and the point of the receiver. The transmitter antenna height is 1.5 m, assuming that the users are using the terminal. The receiving antenna height is 11.15 m, assuming the base station antenna height. In the measurement, four transmitting and receiving antennas were used to eliminate the effect of decrease in received power due to fading.

3 Characteristics of received SNR

In order to obtain the basic characteristics in the outdoor small cell environment, propagation measurements were carried out for the uplink with 5.12 GHz and 19.55 GHz [1]. Fig. 2 shows the received SNR of both frequencies for Course 1 to Course 5. These figures show that the received SNR with 5.12 GHz is much higher than that with 19.55 GHz. The SNR decreases according to the distance moved by the transmitter (D), when 5.12 GHz is used. On the other hand, the SNR obtained with 19.55 GHz decreases considerably at $D = 22$ m and between $D = 70$ and 90 m on Course 2. Similarly, the SNR obtained with 19.55 GHz decreases significantly between $D = 20$ and 70 m on Course 3. This is due to the influence of large tree and hedges between the transmitter and the receiver. Since the wavelength of the 19.55 GHz signal is about 25% shorter than that of the 5.12 GHz signal, it is considered that these radio waves cannot pass through plants, etc. In addition, the...
radio wave of 19.55 GHz has a high rectilinearity compared to that of 5.12 GHz, which is considered to be a factor that makes diffraction at an obstacle difficult.

![Fig. 2](characteristics_of_SNR.png)

**Fig. 2.** Characteristics of SNR.

### 4 Relationship between the distance and channel capacity

Fig. 3 shows the transmission and reception distance characteristics of the channel capacities of 5.12 GHz and 19.55 GHz based on the measured propagation. In this letter, the channel capacity is defined as follows so that the ratio band of 5.12 GHz is 1:

$$
C = \frac{f_c}{5.12} \log_2 \det \left( I_{N_R} + \frac{\text{SNR}}{N_T} \frac{f_c}{5.12} HH^H \right),
$$

The center frequency $f_c$ are 5.12 and 19.55 GHz. The number of transmitting and receiving antennas $N_T$ and $N_R$ are four. The SNR is based on the measured
propagation. The $H$ is propagation channel. $H$ is Hermitian transpose. The propagation model is i.i.d. Rayleigh fading channel to derive the basic characteristics. The number of trials is 1000, and its median is graphed. The bandwidth of 19.55 GHz is 3.82 times the bandwidth of 5.12 GHz, assuming that the available relative bandwidths of 19.55 GHz and 5.12 GHz are the same.

As can be seen from the graph, the channel capacity of the 19.55 GHz signal is more than that of the 5.12 GHz signal when the distance is less than 90 m. On the other hand, the channel capacity of the 5.12 GHz signal is more than that of the 19.55 GHz signal when the distance is more than 90 m.

For 19.55 GHz signals, the portion of the channel capacity less than about 40 bps/Hz is the result obtained from the measured SNR of the non-line-of-sight (NLOS) environment. In the line of sight (LOS) environment, that part of channel capacity is more than 40 bps/Hz, and the channel capacity of the 19.55 GHz signal is more than that of the 5.12 GHz signal. On the other hand, in the NLOS environment, the channel capacity of the 19.55 GHz signal is less than that of the 5.12 GHz signal.

From the above, when constructing a communication system using a millimeter wave band, it can be said that it is necessary to use a sufficient bandwidth, assuming use in a service area far from a base station or a service area where there is considerable shielding. On the other hand, in a service area closer to a base station, it can be said that sufficient communication can be realized even in the millimeter wave band without using a large bandwidth supplement. In other words, it is necessary to use different frequency bands depending on the environment. For example, in the LOS environment, the millimeter wave band is used when the distance is sufficiently close such as less than 40 m, and when the distance is long such as more than 40 m or in the NLOS environment, the microwave band is used.
5 Conclusion

In this study, outdoor wave propagation was measured in a small cell environment for 5.12 GHz, which is a microwave band, and 19.55 GHz, which is a millimeter wave band. The base station antenna height was 11.15 m, and the transmission distance was up to 100 m. The basic performance of the multiple-input and multiple-output (MIMO) for each frequency was compared and examined in terms of the SNR and channel capacity.

First, the device configuration and measurement environment was shown. Then, the SNR characteristics of the 5.12 GHz and 19.55 GHz signals were shown for each measurement course. It was found that the SNR decreased as the distance increased, and the millimeter wave band was significantly affected by the shielding of plants such as trees and hedges compared to the microwave band. Moreover, the performance of the MIMO channel capacity with respect to the transmission distance was confirmed when the fractional bandwidths of the 5.12 GHz and 19.55 GHz signals were identical. If the ratio bands of the millimeter wave band and the microwave band are the same, the microwave band is advantageous. However, it is found that in the NLOS environment, it is possible to obtain larger channel capacity in the millimeter wave band if the relative bandwidths are the same.

From the above, it was concluded that in an actual communication environment, it is necessary to choose between the millimeter wave band and the microwave band depending on the difference between the LOS/NLOS environment.

Acknowledgments

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Application of characteristic modes to design the platform-mounted inverted-F antenna

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Abstract: Many high-frequency antennas have significantly smaller dimensions than the operating wavelength, causing a narrow bandwidths. This paper examines characteristic modes (CMs) of platform-mounted antennas to increase the overall bandwidth of the system. We use the theory of CMs to identify the appropriate platform modes and determine the practical feeding circuits. We fabricate a prototype platform antenna using a scaled model at 860 MHz to verify the simulation results.

Keywords: characteristic modes, characteristic mode analysis, CMA, platform-mounted antenna, inverted-F antenna

Classification: Antennas and Propagation

References

[1] H. Wheeler, “Fundamental limitations of small antennas,” \textit{Proc. IRE}, vol. 35, no. 12, pp. 1479–1484, Dec. 1947. DOI:10.1109/JRPROC.1947.226199
[2] L. J. Chu, “Physical limitations of omni-directional antennas,” \textit{J. Appl. Phys.}, vol. 19, no. 12, pp. 1163–1175, Dec. 1948. DOI:10.1063/1.1715038
[3] T.-Y. Shih and N. Behdad, “Bandwidth enhancement of the platform-mounted HF antenna using the characteristic mode theory,” \textit{IEEE Trans. Antennas Propag.}, vol. 64, no. 7, pp. 2648–2659, July 2016. DOI:10.1109/TAP.2016.2543778
[4] K. Kawabata and H. Arai, “Efficient antenna design for platform-mounted HF antenna using characteristic mode theory,” iWEM, 2016.
[5] R. F. Harrington and J. R. Mautz, “Theory of characteristic modes for conducting bodies,” \textit{IEEE Trans. Antennas Propag.}, vol. 19, no. 5, pp. 622–628, Sept. 1971. DOI:10.1109/TAP.1971.1139999
[6] FEKO, EM Software and Systems (www.feko.info).

1 Introduction

The high-frequency (HF) band (3–30 MHz) is used for various applications. Since HF have a low frequency and a large wavelength, a sufficient antenna length cannot
be provided, and a shortened structure using matching circuit is often used [1, 2]. However, miniaturized antennas tend to be electrically small and have a narrow bandwidth. In the papers [3, 4], a method to enhance the bandwidth of the HF antenna is presented by using the platform modes. The platform has comparable size with operating wavelength, and the bandwidth of the antenna is enhanced if the platform works as a part of the antenna. The platform-mounted antennas are designed using characteristic mode theory [3, 5]. The Inverted-F antenna (IFA) is introduced to excite the platform modes by the simulation [4]. The proposed platform-mounted antenna does not need to use matching capacitances thanks to utilize IFA elements. At first, we verify the simulation result by measurement, then we show other excitation mode by IFAs.

2 Characteristic modes of the platform

This section, we examine the characteristic mode of a 80 mm × 180 mm × 70 mm size platform in Fig. 1(a) using FEKO [6]. Fig. 1(b) shows the modal significance (MS) for the first six modes of the platform. The mode becomes dominant as the MS approaches one [5] then we focus on that mode 1 to excite at 860 MHz. Figs. 1(c) and (d) show the current distribution and radiation pattern for the mode 1 at 860 MHz. The current distribution is the strongest at the edge of the platform, and the radiation pattern is omni-directional in the xy plane and a figure of eight in the zx plane. The next step is to use the IFA element to excite this radiation pattern.
3 Platform-mounted antenna

An IFA element is designed at 860 MHz. IFA uses copper wire with a diameter of 0.9 mm, the impedance of the IFA is adjusted at about 50 Ω. The length of the IFA element is approximately a quarter wavelength while the height is controlled for low profile structure. Four IFA elements are installed on the platform as shown in Fig. 2(a). The number and position of IFA element is adjusted to excite the similar radiation pattern of the mode 1 of the platform. This is proved through the S-parameters in four cases (Fig. 2(b)). The width of $S_{11}$ less than 10 dB is much improved when the IFA is placed at the edge of the platform. From the simulation result, the bandwidth of the platform-mounted four IFA elements are 5.70%. Only the simulation results are shown in [4], so this paper verifies them experimentally.

![Proposed model](image1)

![S11 w/ ideal power divider](image2)

![The fabricated model](image3)

![S11 w/ wilkinson power divider](image4)

![xy plane](image5)

![zx plane](image6)

Fig. 2. The platform-mounted antenna (mode 1).
Fig. 2(c) shows the fabricated model of the aluminum case covered with copper tape with IFAs. In order to use four IFA elements, a power divider is also made. Fig. 2(d) shows the measured S-parameter of the fabricated model. There is a small difference between the measurement and the simulation due to effect of the power divider. But the simulated S-parameter of the model including both the platform-mounted antenna and the power divider has a good agreement with the measurement as shown in Fig. 2(d). The measured radiation patterns in xy and zx planes in Figs. 2(e) and (f) agree well with simulation and the mode 1 radiation is excited well by this feeding method.

Next, we consider to excite mode 3, which is one of the higher order mode. Fig. 3(a) shows the current distribution of the mode 3 at 860 MHz. Its radiation...
pattern is a figure of eight in the yz plane and omni-directional in the zx plane. We use the same IFA while feeding to excite radiation pattern of mode 3. The position of IFA feeding point is optimised to excite mode 3 as shown Fig. 3(b), and fabricated model is shown in Fig. 3(c). The simulation bandwidth is 0.80%, while the measurement is expanded due to feeding network. Figs. 3(e) and (f) shows radiation pattern on the yz and zx planes. Measurement radiation pattern is agreed with simulation result. We verified that proposed IFA feeding also excites the mode 3.

4 Conclusion

This paper verified that radiation patterns of mode 1 and 3 are excited using IFA mounted platform. The bandwidth of the platform-mounted antenna are improved greatly. The platform-mounted antenna using mode 1 and 3 provide similar radiation patterns with CMs. We verified that the antenna performance is improved by designing the antenna using the excitation mode of the platform. In future work, the size of the antenna will be miniaturized.
Three-dimensional printed asymmetric biconical antenna for borehole concrete sensing application

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Abstract: An asymmetric biconical antenna is proposed to be used for a borehole radar system to inspect the corrosion of reinforcing bar inside a concrete structure. First, the antenna is designed by considering the hole needed for the borehole diameter. Then, the antenna is fabricated using a three-dimensional printer with polylactic acid plastic and coated with copper tape. Finally, the antenna is tested in free space and concrete environment. The diameter of the antenna is 20 mm and a length of 34 mm, with the usable frequency in free space is 2.4 GHz (2.1 GHz in concrete) to 8 GHz.

Keywords: biconical antenna, three-dimensional printer, concrete sensing, borehole radar system

Classification: Sensing

References

[1] M. Nishimoto and Y. Naka, “Analysis of transient scattering by a metal cylinder covered with inhomogeneous lossy material for nondestructive testing,” *IEICE Trans. Electron.*, vol. E101.C, no. 1, pp. 44–47, Jan. 2018. DOI:10.1587/ transele.E101.C.44

[2] J. L. Davis and A. P. Annan, “Ground-penetrating radar for high-resolution mapping of soil and rock stratigraphy,” *Geophys. Prospect.*, vol. 37, no. 5, pp. 531–551, July 1989. DOI:10.1111/j.1365-2478.1989.tb02221.x

[3] G. Roqueta, L. Jofre, and M. Q. Feng, “Analysis of the electromagnetic signature of reinforced concrete structures for nondestructive evaluation of corrosion damage,” *IEEE Trans. Instrum. Meas.*, vol. 61, no. 4, pp. 1090–1098, Apr. 2011. DOI:10.1109/TIM.2011.2174106

[4] J. H. Bungey, “Sub-surface radar testing of concrete: A review,” *Constr. Build. Mater.*, vol. 18, no. 1, pp. 1–8, Feb. 2004. DOI:10.1016/S0950-0618(03)00093-X

[5] M. Vona and D. Nigro, “Evaluation of the predictive ability of the in situ concrete strength through core drilling and its effects on the capacity of the RC columns,” *Mater. Struct.*, vol. 48, no. 4, pp. 1043–1059, Apr. 2015. DOI:10.1617/s11527-013-0214-2

[6] S. Zhang, C. C. Njoku, W. G. Whittow, and J. C. Vardaxoglou, “Novel 3D
1 Introduction

A non-destructive inspection by utilizing ground penetrating radar provides the possibility to inspect the corrosion of the reinforcing bar inside the concrete structure without exposing the reinforcing bar to the environment. Nishimoto et al. demonstrated the possibility to use an ultra-wideband impulse signal with the frequency range of 2 GHz to 9 GHz to inspect a corroded metal bar from 10% to 30% corrosion rate [1]. By using a multi-gigahertz electromagnetic wave, a higher spatial resolution can be observed from the target [2] and could differentiate a thin layer of iron oxide formation on the surface of the reinforcing bar [3]. However, the penetration of the electromagnetic wave inside the concrete is limited due to the high conductivity characteristic of the concrete [4]. The borehole radar system can be utilized to reduce the distance between the reinforcing bar and the antenna by creating a hole near the observed reinforcing bar in the concrete. However, the effect of the core drilling in the concrete beam could impact the concrete strength. This was demonstrated by Vona et al. with a 30% concrete strength reduction for a 100 mm hole on a 200 mm concrete beam [5].

In this paper, we propose a coaxial feed biconical antenna for borehole concrete sensing with a small diameter and ultra-wideband characteristics to transmit an impulse signal with the bandwidth of 2 GHz to 8 GHz. The antenna is designed to be printed using a three-dimensional (3D) printer and coated by using conductive tape. The antenna has been tested on the free space and concrete environment, and both of the simulated and measured results of the antenna design are presented.

2 Antenna design

We conducted parameter studies to determine the optimal diameter of the antenna and the optimal ratio between the length of both arms of the biconical antenna with EMPro finite-difference time-domain simulator from Keysight Technologies. Fig. 1(a) shows the antenna shape used in this study. The permittivity of the polylactic acid (PLA) plastic used in this study is based on [6], with relative dielectric permittivity of 2.72 and loss tangent of 0.008 (for 100% infill). The antenna is assumed to be matched to a 50 Ω system and simulated without considering the coaxial cable and connector used later in the fabrication process.

The parameters studies begin with the choice of the diameter $D$, and the ratio of top-section length $L_t$ and bottom-section length $L_d$ while using the plastic thickness of 2 mm and the antenna gap of 1.5 mm. The parameters are determined by choosing the best simulations results so that the bandwidth of the 10 dB return loss becomes as wide as possible. Based on [5] which describes the effect of the core drilling in the concrete beam on the concrete strength, we decided to reduce the diameter to be as small as possible (less than 20 mm) to reduce the hole’s...
diameter needed for the borehole sensing while maintaining the ultra-wideband characteristics.

The simulated return loss of the different antenna diameters and lengths ratio is shown in Fig. 1(b) and (c). Based on the simulation, the final geometrical parameters are as follows: $D = 20$ mm, $D_{in} = 2.3$ mm, $L_{t} = 25$ mm, $L_{d} = 7.5$ mm, $g = 1.5$ mm, $T_{p} = 2$ mm, $T_{c} = 0.06$ mm. From the simulation results, the bigger the antenna diameter the better the frequency response on the lower frequency. We decided to choose the $20$ mm diameter because it has a better characteristic when the asymmetric configuration is applied, and also the asymmetric configuration has the best bandwidth compared with the equal length of the antenna arm. We also simulated the radiation pattern for the free space and inside the concrete, the results are shown in Fig. 1(d). Overall, the antenna has an omnidirectional radiation pattern for both the free space and inside the concrete simulation.

3 Fabrication process

The 3D printer can deposit material in three-dimension space with high precision and can produce a complex geometrical object with rapid prototyping time and low manufacturing cost. The inner plastic core is printed by using Creality Ender 3 3D printer with PLA plastic. The overall dimension of the antenna is shown in
Another two cones are printed (see Fig. 2(b)) for attaching conductive tape on the inside surface of the plastic core. The 3D printer used has a 0.4 mm extruder diameter and using 1.75 mm standard PLA filament. We decided to build the antenna with a 0.6 mm layer height and 40 mm/s print speed to improve the surface roughness of the printed parts.

After the parts are printed (see Fig. 2(b)), the antenna was coated by a conductive spray made from silver and copper. However, the spray method was proven difficult to solder due to the thinness of the cured conductive layer and the inner plastic core is deformed by the high temperature due to low insulation between the conductive layer and plastic core. Even though we successfully make one prototype using this method, the conductive layer is fragile and easily flakes if the surface of the plastic core is not primed with enamel paint before sprayed with conductive spray.

To circumvent these difficulties, we decided to use conductive tape with 0.06 mm thickness made from copper foil and conductive adhesive backing. The conductive tape is coated to the surface of the parts and a rigid coaxial cable (see Fig. 2(c)) is soldered into the antenna, we are using ITT RG405 for the coaxial cable and Radiall SMA for the connector. By using this tape, the soldering process can be done without deformed the inner plastic and the bonding between parts is strong and able to hold the two-part antenna in place. The fully fabricated antenna is shown in Fig. 2(d).

4 Antenna characterization

We characterize the frequency behavior of the proposed antenna by using an Advantest R3768 vector network analyzer to measure the return loss of the antenna.
Both the measured and simulated return loss of the antenna is shown in Fig. 3(a). From the measured data, it can be seen that the usable bandwidth is from 2.4 GHz to 8 GHz in free space. The characteristic of the fabricated antenna is agreed well with the simulation result in free space. In addition, the return loss of the measurement results is better from 5.5 GHz to 8 GHz.

We also try to measure the return loss of the antenna in a concrete environment. In order to simulate the environment in concrete, 66 liters of sand and 1 liter of water were mixed and packed into a rectangular container with a 30 mm diameter hollow pipe penetrating in the center as a borehole for measurement (see Fig. 3(b)). The fabricated antenna was inserted into the center of the pipe and the characteristics were measured. Fig. 3(c) shows the measurement results of the concrete model.

Fig. 3. Measurement of the antenna characteristics. (a) Return loss in free space. (b) The measurement model of the concrete environment (size is in mm). (c) Return loss in the concrete model.
model. The simulation result is also shown in the figures. The lower frequency limit of the antenna in concrete is reduced to 2.1 GHz, which increases the available bandwidth. In addition, the return loss of the measurement results is slightly worse from 3.5 GHz to 4.8 GHz but is practically acceptable. From these results, we can confirm that the characteristics of the proposed antenna in the concrete model agree well with the simulation results.

5 Conclusion

We proposed a biconical antenna for borehole concrete sensing with small and ultra-wideband characteristics. For insertion into a borehole with a small diameter, the antenna was designed with a diameter of 20 mm and a length of 34 mm. In the simulation, the usable bandwidth was from 2.4 GHz to 8.0 GHz in free space and 1.9 GHz to 8.0 GHz in concrete. Next, we fabricated the designed antenna using a 3D printer with PLA plastic and measured its characteristics in free space and in a concrete model. As a result, we confirmed that the fabricated antenna had almost the same characteristics as the simulation results and it was available for inspection of reinforcing bar in concrete.
Beam selection for mm-wave massive MIMO systems using ACO & combined digital precoding under hybrid transceiver architecture

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Abstract: We propose a Wiener filter (WF) and a combined zero forcing (ZF)/WF precodings with ant colony optimization (ACO) algorithms as the digital precoder for millimeter wave multiple-input multiple-output (MIMO) systems. In the proposed schemes, beams are selected based on maximal magnitude (MM) criteria and the selection is optimized using ACO algorithm to achieve near optimal solution with highly reduced complexity. According to the simulation results, while ACO-WF scheme achieves higher sum rate in high inter-user interference (IUI) environments compared with conventional precoders, we confirm ACO-ZF/WF scheme can obtain a maximal sum rate in both low and high IUI environments.

Keywords: millimeter wave, MIMO, ACO algorithm, zero forcing, Wiener filter

Classification: Antennas and Propagation

References

[1] S. Qiu, K. Luo, and T. Jiang, “Beam selection for mmWave massive MIMO systems under hybrid transceiver architecture,” IEEE Commun. Lett., vol. 22, no. 7, pp. 1498–1501, July 2018. DOI:10.1109/LCOMM.2018.2829482

[2] D. Alimo and M. Saito, “Beam selection for massive MIMO using wiener filter under hybrid transceiver,” Proc. the 2019 IEICE Society Conference, Osaka, Japan, Sept. 2019.

[3] L. Liang, W. Xu, and X. Dong, “Low-complexity hybrid precoding in massive multiuser MIMO systems,” IEEE Wireless Commun. Lett., vol. 3, no. 6, pp. 653–656, Dec. 2014. DOI:10.1109/LWC.2014.2363831
1 Introduction

Beamspace massive MIMO, considered in this literature, is the hybrid architecture where conventional antenna space is converted into beamspace to produce pre-defined beams using discrete lens array (DLA) in the analog domain [1]. With millimeter wave (mm-wave) massive MIMO, the channel becomes sparse and not all the beams contribute to the overall system performance as such, beam selection is important to reduce hardware cost, complexity, and power consumption further while maintaining reasonable system performance.

Selecting beams using maximal magnitude (MM) has low complexity however, it ignores the inter-user interference (IUI) which limits its performance in multi-user systems. Although the best way to find optimal beams is to use exhaustive search, it is prohibited due to its high computational complexity in massive MIMO systems. Beam selection using MM, ant colony optimization (ACO) and zero forcing (ZF) as digital precoder (ACO-ZF) was proposed [1]. ZF precoder completely eliminates IUI and results in a higher sum rate when IUI level is low. However, the performance degrades when interference becomes very high because it reduces the received signal power at the receiver due to its noise enhancement feature.

In this study, we propose a Wiener filter (WF) based precoder and beam selection using MM and ACO (ACO-WF) [2]. Although WF-ACO produces higher sum rate in high IUI region than ACO-ZF does, it provides lower sum rate in low IUI region. Thus, we also propose a combined precoding scheme of ZF and WF as the dual digital precoder. Through computer simulation, the proposed combined precoding scheme results in a system sum rate superior to the system sum rate when ZF or WF precoding is used alone as a digital precoder in a low and high IUI environment.

2 System model

We consider a base station (BS) with \( N \) antennas serving \( K \) single antenna users in a downlink mm-wave beamspace massive MIMO multi-user scenario, which is the same as that in the literature [1] and its parameters used in simulation are as in Table I shown in Section 5. The users are randomly located in a circular area of radius \( L \) at a distance \( D \) from the BS.

The received signals by the \( K \) users are given by [1]:

\[
y = H^H U F_{BB} x + n
\]  

(1)

where \( H = [h_1, \ldots, h_K] \) is an \( N \times K \) channel matrix, where \( h_k \) for \( k = 1, \ldots, K \) is the channel vector for the \( k \)-th user, \( U \) is an \( N \times K \) discrete fourier transform (DFT) matrix as the analog precoder, \( n \) is an \( N \times 1 \) additive white Gaussian noise (AWGN) vector whose entries are Gaussian random variables with zero mean and variance \( \sigma^2 \), \( F_{BB} \) is an \( N \times K \) digital precoding matrix, \( x \) is an \( N \times 1 \) transmitted complex signal vector. We use the same channel model to generate the components of \( h_k \) as used in the literature [1].

The beamspace MIMO channel \( \hat{H} \) becomes:

\[
\hat{H} = [\hat{h}_1, \hat{h}_2, \ldots, \hat{h}_K] = [U^H h_1, U^H h_2, \ldots, U^H h_K]
\]  

(2)
where $\mathbf{h}_k$ is the beamspace channel vector of the $k$-th user. The elements of $\mathbf{h}_k$ represents the $N$ orthogonal pre-defined beams. The number of dominant beams is much smaller than $N$ due to the sparse structure of the beamspace channel. As such, $K$ beams are selected to reduce the MIMO system dimension to form $K \times K$ channel matrix $\mathbf{H}_r$ with no performance loss such that $\mathbf{H}_r = \mathbf{H}(m,:)\mathbf{r}^m$, where $R$ is the set of selected dominant beams based on MM criteria. The dimension reduced beamspace MIMO downlink received signals of $K$ users are given by:

$$\mathbf{\tilde{y}}_r = \mathbf{\tilde{H}}_r^H \mathbf{\tilde{F}}_{BB} \mathbf{x} + \mathbf{n}$$

where $\mathbf{\tilde{F}}_{BB}$ is the $K \times K$ dimension reduced precoding matrix and has the constraint of $E\{||\mathbf{\tilde{F}}_{BB}\mathbf{x}||^2\} = \rho$ and $\rho$ is the transmit power at the BS.

### 3 Precoder matrix and sum rate maximization

Low complexity BS system is the main focus of this work. For the purpose, ZF precoder is proposed as [1]:

$$\mathbf{F}_{ZF} = \alpha_{ZF} \mathbf{\tilde{H}}_r (\mathbf{\tilde{H}}_r^H \mathbf{\tilde{H}}_r)^{-1}, \quad \alpha_{ZF} = \sqrt{\frac{\rho}{\text{tr}(\mathbf{\tilde{H}}_r^H \mathbf{\tilde{H}}_r)^{-1}}}.$$  

(4)

To improve the sum rate performance in high IUI area, we propose Wiener filter based precoder as [2]:

$$\mathbf{F}_{WF} = \alpha_{WF} \mathbf{F}, \quad \mathbf{F} = (\mathbf{\tilde{H}}_r^H \mathbf{\tilde{H}}_r + \zeta \mathbf{I})^{-1} \mathbf{\tilde{H}}_r, \quad \zeta = \frac{\sigma^2 K}{\rho},$$

$$\alpha_{WF} = \sqrt{\frac{\rho}{\text{tr}(\mathbf{\tilde{F}}_{WF}^H \mathbf{\tilde{F}}_{WF})}}.$$  

(5)

Even though WF outperforms ZF at low and high SNR, the sum rate performance of WF becomes lower then ZF at low IUI environment and it increases higher than ZF when IUI becomes higher. Based on this characteristic, a combined ZF/WF precoder is also proposed in this work. According to used criteria, the precoders $\mathbf{F}_{ZF}$ and $\mathbf{F}_{WF}$ are replaced by $\mathbf{\tilde{F}}_{BB}$ in (3).

Assuming equal transmit power at the BS, the average rate of the $k$-th user is given by $R_k = \log_2(1 + \gamma_k)$, where $\gamma_k$ is the signal-to-interference plus noise power ratio (SINR) of the $k$-th user [3]. The sum rate becomes $R_{\text{sum}} = \sum_{k=1}^{K} R_k$. The optimal beams $\mathbf{\tilde{H}}_r^{\text{opt}}$ are selected based on:

$$\mathbf{\tilde{H}}_r^{\text{opt}} = \arg\max R_{\text{sum}}.$$  

(6)

The best solution to obtain $\mathbf{\tilde{H}}_r^{\text{opt}}$ is to use exhaustive search. However, since its computational complexity in massive MIMO systems is prohibitive, a low complexity beam selection algorithm is crucial to obtain a sub-optimal $\mathbf{\tilde{H}}_r^{\text{opt}}$.

### 4 Beam selection optimization using ACO algorithm

To achieve a near optimal solution of beam selection, we use ACO as a low complexity algorithm. Due to channel sparsity, $D_k$ dominant beams are used for the $k$-th user beam selection such that $D_k$ is smaller than the number of pre-defined beams. The index set is given by $D_k = \{I_{d_1}, \ldots, I_{d_{D_k}}\}$. For the $t$-th iteration, the cost function $d_{kd}^t$ is used to determine the suitability of the $d$-th beam in $D_k$. 

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From (4) and (6), the cost function for ACO-ZF scheme is given by [1]:

\[ d_{kd}^{t} = \text{tr}\left( (\tilde{H}_{t,k}^{-1})^{T} \tilde{H}_{t,k}^{-1} + \Pi \right) \]  

(7)

and from (5) and (6), the cost function for ACO-WF scheme is given by [2]:

\[ d_{kd}^{t} = \text{tr}\left( \left( \tilde{H}_{t,k}^{-1} \right)^{T} \left( \frac{(K-1)}{\rho} \mathbf{I} + \tilde{H} \right)^{-1} \right) \]  

(8)

where \( \tilde{H}_{t,k}^{-1} \) is a \((K-1) \times K\) matrix obtained by omitting the \(k\)-th row of \( \tilde{H}_{t}^{-1} \), which represents the channel matrix obtained at the \((t-1)\)-st iteration, and \( \tilde{H} = \tilde{h}(I_{kd},:)\tilde{h}(I_{kd},:) + \varsigma \mathbf{I} \) where \( \tilde{h}(I_{kd},:) \) is the \(I_{kd}\)-th row of the full dimension beamspace channel matrix \( \tilde{H} \) and \( \varsigma \) is a positive number to impose the existence of matrix inversion.

To determine the cost function for ACO-ZF/WF, we first set a threshold \( \psi \) from the sum rate performances, then select (7) if \( \psi \leq K \) or (8) if \( \psi > K \).

The cost function for ACO-ZF, ACO-WF, and ACO-ZF/WF is then converted into their heuristic values by:

\[ \eta_{kd}^{t} = \sqrt{e^{-d_{kd}^{t}/N^{2}}} \]  

(9)

Then the BS selects the beam for each user based on the probability:

\[ p_{kd}^{t} = \frac{[r_{kd}^{t-1}]^{a}[\eta_{kd}^{t}]^{b}}{\sum_{d=1}^{D_{k}}[r_{kd}^{t-1}]^{a}[\eta_{kd}^{t}]^{b}} \]  

(10)

where \( a \) and \( b \) are positive constants, and \( r_{kd}^{t-1} \) is the positive feedback of the previous selection. The beam with the highest probability is selected for the \(k\)-th user. Then the feedback is updated after the beam selection by:

\[ r_{kd}^{t} = (1 - \Gamma)r_{kd}^{t-1} + \omega \eta_{kd}^{t} p_{kd}^{t} \]  

(11)

where \( \omega \) is a weight factor and \( \Gamma \) is a decay parameter whose value is between 0 and 1. Higher probability and heuristic value increase the positive feedback faster and thus a sub-optimal solution can be reached faster with increasing iterations. The modified ACO algorithm is shown in Algorithm 1.

**Algorithm 1** ACO algorithm for beam selection

**Input:** \( \tilde{H}, D_{k}, \forall k, T_{\text{max}}, a, d, \Gamma, \omega, \psi \)

**Output:** \( \tilde{H} \)

1: \( R^{0} = \{I_{0}^{0}, \ldots, I_{K}^{0}\}; \tilde{H}^{0} = \tilde{H}(m,:)_{m \in R^{0}}; \text{tr} = \infty; \eta_{kd}^{t} = 1, \forall t, \forall k, \forall d \).
2: for \( t = 1 \) to \( T_{\text{max}} \) do
3: \( R^{t} = R^{t-1}; \)
4: for \( k = 1 \) to \( K \) do
5: for \( d = 1 \) to \( D_{k} \) do
6: \( \text{if } K \leq \psi \text{ then} \)
7: \( \text{Calculate } d_{kd}^{t} \text{ according to (7)}; \)
8: \( \text{else} \)
9: \( \text{Calculate } d_{kd}^{t} \text{ according to (8)}; \)
10: end if
11: \( \text{Calculate } \eta_{kd}^{t} \text{ and } p_{kd}^{t} \text{ according to (9) and (10), respectively}; \)
12: Update $f_{kd}'$ according to (11);
13: end for
14: Select maximum $p_{kd}'$, $\forall d$. Update index $d_{\text{max}}$ in $R^{l-1}$; $l_{d_{\text{max}}}' = I_{kd_{\text{max}}}$;
15: if $d_{kd_{\text{max}}} \leq \tau$ then
16: $\tau = d_{kd_{\text{max}}}$, $\tilde{H}_r = \widetilde{H}(m, :)$; 
17: else
18: $\tilde{H}_r = \widetilde{H}(m, :)$;
19: end if
20: end for
21: end for
22: Return $\tilde{H}_r$

| Table I. Simulation parameters |
|--------------------------------|
| Parameter                  | Value | Parameter                  | Value (ZF) | Value (WF) |
| Distance $D$               | 150   | $a$                        | 0.8        | 1          |
| Radius $L$                 | 10    | $b$                        | 0.4        | 2          |
| No. of cluster $N_c$       | 3     | $a_{k,0}$                  | $-3$ dB    | $-3$ dB    |
| $\omega$                   | 0.5   | $a_{k,i}$                  | $-5$ dB    | $-5$ dB    |
| Angular spread AS          | 5°    | Decay $\Gamma$            | 0.3        | 0.3        |
| No. of rays $N_r$          | $\mathcal{U}[1, 30]$ | $\sigma^2$ | 1          | 1          |

5 Numerical results

The sum rates of the proposed schemes ACO-WF [2] and ACO-ZF/WF are evaluated and compared to ACO-ZF [1], interference-aware (IA), and MM schemes in this section. The simulation parameters are listed in Table I. We use the same channel parameters as those in [1].

The BS is with $N = 30$ beams, while $K$ user terminals are with single antenna. Using ACO, 1 beam per user terminal is attained.

We show sum rates versus $K$ from 5 to 21 in Fig. 1a. From the figure, we can see that the sum rate by ACO-ZF is better than that by ACO-WF when $K \leq 8$. Thus, we set the threshold as $\psi = 8$ for ACO-ZF/WF. From Fig. 1a, it can be seen that the proposed scheme of ACO-ZF/WF achieves higher sum rate compared to IA and MM schemes. When $K \leq \psi$, ACO-ZF is selected as the precoder and when $K > \psi$, ACO-WF is selected as the precoder. By selecting one precoder at a time, the sum rate can be maximised in both low and high IUI environments.

In Fig. 1b, sum rate comparison against number of iterations is shown in a low IUI scenario, where $N = 30$ and $K = 5$, i.e., $K \ll N$. Then for ACO-ZF/WF, ZF would be the precoder of choice. This scheme only requires 3 iterations to reach the maximum sum rate.

In Fig. 1c, sum rate comparison against number of iterations is shown in high IUI scenario. In this scenario, we consider $N = 30$ and $K = 15$. The precoder of choice becomes WF in ACO-ZF/WF scheme due to the existence of high IUI. With only 2 iterations, a sum rate higher than the other schemes can be achieved.
6 Conclusion

Beams are selected using MM and optimized using ACO and the combined precoding scheme in the digital domain is proposed in mm-wave massive MIMO system using DLA in the analog domain. Through simulation results, the proposed ACO-ZF/WF scheme achieved higher sum rate in low and high interference scenarios compared to other schemes.

(a) Sum rate versus number of users $K$ when $N = 30$, $D_k = 8$, $T_{max} = 10$, and $\text{SNR} = 20$ dB.

(b) Sum rate versus number of iterations when $N = 30$, $K = 5$, $D_k = 8$, and $\text{SNR} = 20$ dB.

(c) Sum rate versus number of iterations when $N = 30$, $K = 15$, $D_k = 8$, and $\text{SNR} = 20$ dB.

![Graph](image-url)

**Fig. 1.** Sum rate comparisons of several beam selection schemes
Evaluation of mode vector for near-field 2-D target imaging using Khatri-Rao product extended array processing

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Abstract: Achieving high-precision and high-resolution imaging with millimeter-wave radar requires a large number of antennas and data volumes. In this study, we applied the Khatri-Rao (KR) product extended array processing to synthetic aperture radar (SAR) data to increase the number of antenna elements substantially and extend the synthetic aperture length. The target of interest in this study is relative short range target, which is several meters away from the radar, such as indoor human with a small object on his/her body. When the target is close to the radar, target imaging by SAR using the far-field approximation will often increase the error. In this case target-imaging by using range-migration is required. However, this processing often computationally inefficient. In this study, we evaluate image-quality of short-range 2-dimensional imaging with far-field and near-field mode vector with/without range migration. The image-quality is examined by peak power gain. Numerical results show that the near-filed mode vector works properly for out study without range migration. In addition, SAR imaging using KR product extended arrays shows that image quality can be improved compared to conventional uniform SAR.

Keywords: short-range target imaging, Khatri-Rao product, SAR, millimeter-wave radar, low-redundancy linear arrays

Classification: Antennas and Propagation

References

[1] H.-M. Chen, S. Lee, R. M. Rao, M.-A. Slamani, and P. K. Varshney, “Imaging for concealed weapon detection,” IEEE Signal Process. Mag., vol. 22, no. 2, pp. 52–61, Mar. 2005. DOI:10.1109/MSP.2005.1406480

[2] J. M. Lopez-Sanchez and J. Fortuny-Guasch, “3-D radar imaging using range migration technique,” IEEE Trans. Antennas Propag., vol. 48, no. 5, pp. 728–737, May 2000. DOI:10.1109/8.855491

[3] Y. Wakamatsu, H. Yamada, and Y. Yamaguchi, “MIMO Doppler radar using Khatri-Rao product virtual array for indoor human detection,” IEICE Trans. Commun., vol. E99-B, no. 1, pp. 124–133, Jan. 2016. DOI:10.1587/transcom.
1 Introduction

In recent years, applications of millimeter-wave (MW) radar for indoor human monitoring and security are expected [1]. In particular, fears of terrorism and suspicious person or objects are increasing, and the importance of a system to check the possession of dangerous materials is being reviewed. In this study, by applying Khatri-Rao (KR) product extended array processing for Synthetic Aperture Radar (SAR) data, we increased the number of antenna elements virtually and extended the synthetic aperture length. By using KR processing, the imaging with a small number of antenna elements became possible.

In addition, when acquiring 3D or 2D imaging for short-distance targets, it is necessary to consider near-field effect, since far-field approximation no longer holds [2]. If signal processing is performed using far-field approximation in such a situation, the imaging quality deteriorates and spurious responses often occur. Of course, joint estimation [3] based on strict location for 2D/3D imaging can yield good results, but it is computationally inefficient. Since we can use enough frequency bandwidth in the MW-FMCW radar, it is preferable to make 2D/3D target image after delay estimation.

In this report, we evaluated image quality of 2D estimation for short-range target by using SAR. In the imaging algorithm employed here, we first applied FFT to beat signal in each measurement point for distance estimation, then we carried out direction estimation. The image quality discussed here is concerned with a suitable mode vector in the direction (or azimuth location) estimation. We show imaging results of point targets by using far-field mode vector, near-field mode vector, and strict mode vector with range migration [4]. To evaluate the results quantitatively, peak power loss of target response was employed. From these results, we show that the imaging accuracy of the near-field mode vector without range migration could be acceptable for relative short-range target of our interests. In addition, SAR imaging using KR product virtual arrays could improve image quality in compared to that by the conventional uniform SAR.

2 Receiving data model

In this study, we considered 2D imaging, which is azimuth and range direction (or x-y plane), by one-dimensional scanning SAR. We also assumed that the range/distance estimation in each measurement point was done by FFT for simplicity. In the followings, we only focused on the direction/azimuth position estimation by using the array data by the range of interest.
Let us assume that a linear array of \( L \) data points obtained from \( i \)-th range bin having one target response for simplicity. The received signal \( x_i \in \mathbb{C}^{L} \) of the plane wave of complex amplitude \( s \) coming from the \( \theta \) direction is given by

\[
x_i = a(\theta)s + n,
\]

where \( n \) is the additive noise, and \( a(\theta) \in \mathbb{C}^{L} \) denotes the response of the array to the plane wave of arriving from direction \( \theta \). If the target is located in the far-field area, we can denote \( a(\theta) \) by using Far-field mode vector as follows:

\[
a(\theta) = [1, e^{-j\frac{2\pi}{\lambda}d \sin \theta}, e^{-j\frac{4\pi}{\lambda}d \sin \theta}, \ldots, e^{-j\frac{2\pi}{\lambda}(L-1)d \sin \theta}]^T,
\]

where \( \Delta d \) is array spacing. Note that the phase constant becomes doubled as compared to the mode vector in conventional array signal processing since SAR data acquisition is assumed in this study. When there exist \( K \) targets, it becomes:

\[
x_i = \sum_{k=1}^{K} a(\theta_k)s_k + n = As + n,
\]

\[
A = [a(\theta_1), a(\theta_2), \ldots, a(\theta_K)],
\]

\[
s = [s_1, s_2, \ldots, s_K]^T,
\]

where \( s_k(t) \) and \( \theta_k \) are the complex amplitude and arrival direction of each arrival wave, respectively. For near-field target, we should consider the spherical wave effect in the azimuth direction estimation. There are two ways: the first one is an approximate approach to just compensate phase shift by the spherical effect. This is proportional to delay time of the target. We will also refer to \( a(\tau) \in \mathbb{C}^{L} \) as Near-field approximated mode vector, and is given by

\[
a(\tau) = [e^{-j2\pi f\tau_1}, e^{-j2\pi f\tau_2}, \ldots, e^{-j2\pi f\tau_L}]^T
\]

where \( f_c \) is the center frequency and \( \tau_i \) \((i = 1, 2, \ldots, L) \) is round trip delay time to target at each antenna. Note that the data in \( x_i \) is extracted in the same range-bin. This is an approximated mode vector in this sense. Strictly speaking, range migration is required for near-field target. In this case, we extract the range-bin data corresponding to the delay time (or distance) from the radar to the target. In the following analysis, we denote this estimation as “near-field mode vector with range migration.”

### 3 Khatri-Rao product virtual array

Consider the case where a narrow band signal is received by the \( L \) elements Low-Redundancy Linear Arrays (LRLA) [5]. The received data are expressed by (3) by using corresponding array configuration, then the correlation matrix \( R_{xx} \) can be given by

\[
R_{xx} = E[xx^H] = ASA^H + R_N = \sum_{k=1}^{K} |s_k|^2 a(\theta_k)a(\theta_k)^T + R_N,
\]

where \( A \) is the mode matrix, \( S \) is the source correlation matrix, and \( R_N \) is the noise correlation matrix. By using a vectorization operation to the correlation matrix, we
can obtain the following virtual array data vector. Then vec[-] is a function to vectorize the matrix.

\[
z = \text{vec}[R_{\text{xx}}] = \sum_{k=1}^{K} |s_k|^2 \text{vec}[a(\theta_k)a(\theta_k)^H] + \text{vec}[R_N].
\]  

(8)

This is the basic concept of KR transform. We often take out the non-overlapping elements in \(\text{vec}[a(\theta_k)a(\theta_k)^H]\), and make corresponding mode vector \(b(\theta_k)\) as follows:

\[
z' = \sum_{k=1}^{K} b(\theta_k)|s_k|^2 + \text{vec}[R_N].
\]  

(9)

In this way, we can obtain virtually extended data vector.

The KR processing is straightforwardly applicable for 2D joint estimation [4]. However, when we apply combination of 1D estimation as the scenario in this letter, we must estimate the range first because phase information of targets is dropped by the KR processing.

4 Simulation results

4.1 Imaging quality comparison

In this section, we show some simulation results of 2D position estimation using SAR data obtained by virtually expanding the synthetic aperture length by applying the KR processing. The employed array in this simulation is LRLA whose element positions are \{0 1 3 6 13 20 27 34 41 45 49 50\} normalized by \(\lambda/4\). Using the mode vector in (2) and (6), respectively, we performed 2-D target location estimation on the x-y coordinate plane, where x and y corresponds to range and azimuth direction, respectively. Center Frequency of the radar is 79 GHz and Bandwidth is 3 GHz. In these simulations, we deployed three point targets on \((X, Y) = \#1 (0.5 \text{ m}, 2.5 \text{ m}), \#2 (0 \text{ m}, 3 \text{ m}), \text{ and } \#3 (1 \text{ m}, 3.5 \text{ m})\).

Imaging results are shown in Figs. 1(a), (b) and (c). Figs. 1(d), (e) and (f) show the results of DOA estimation on target (#1). We confirmed that the grating lobe could be suppressed simply by increasing the accuracy of the mode vector. Of course, the results using Near-field mode vector with range migration is the best. However, by using the Near-field mode vector without range migration, we can obtain a good imaging result in this simulation.

4.2 Evaluation of mode vector

To evaluate image quality quantitatively, we adopt peak power loss of point target. We got an indication of whether to use long-range mode vectors or short-range mode vectors at the distance between the target and the antenna. In this paper, we examine peak power loss in comparison to the peak power in a far-field target. Here we adopt the target at range of 40 m as a reference target. Propagation loss due to range distance is compensated in the following evaluation. We compared the antenna aperture lengths and compared the case of the Uniform Linear Array (ULA) without KR processing and the case of LRLA with KR processing. The
FM-CW radar parameter used here is Center Frequency of 79 GHz, Sweep Time of 5 msec., and frequency bandwidth of 1 GHz and 3 GHz, respectively. In this simulation, we assume a point target at \((x, y) = (0 \text{ m}, 0\sim10\text{ m})\) at 1 m intervals in \(y\). In Fig. 2, we show the simulation parameters. The LRLA element arrangements are selected to realize the same virtual aperture length to the corresponding ULA. The simulation results of evaluation mode vector is shown in Fig. 3, where (a) and (c) show the results of the Far-field mode vector, and (b) and (d) show the results of the Near-field mode vector. Figs. 3(a) and (b) show the results with ULA. On the other hand, Figs. 3(c) and (d) show the results of the LRLA with the KR precessing.

From these results, we have confirmed that even if the aperture length is large, the use of the near-field mode vector can suppress the deterioration of accuracy. It was also found that the accuracy is better for the virtual arrays by the KR processing having the same aperture length equivalently. In addition, we should
note that the accuracy depends on the frequency bandwidth of the MW-radar. Image deterioration becomes severe for wide-bandwidth radar.

5 Conclusion

In this report, we have evaluate imaging quality for short-range MW-SAR. The results show that the imaging quality can be good by using the near-field mode vector without range migration, as in the case of the situation put the target in 3 m (750λ) with 28 cm (70λ) aperture. It was also confirmed that a narrower frequency bandwidth provided better results. Furthermore, we showed that the imaging quality can be improved by the virtual array using the KR processing. This evaluation is applicable for 2D MIMO radar. It will be done in near future.

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Simplification method of 3D point cloud data for ray trace simulation in indoor environment

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Abstract: With the advent of IoT (Internet of Things) services in recent years, various industries are now working towards operating wireless networks and this in turn gives rise to the need for radio propagation simulations. However, in order to obtain reliable simulation results, an accurate model of the environment is often needed. This paper proposes a method for generating a plane model automatically, for use in ray tracing simulations from 3D point cloud data in order to ease the burden of creating models manually. Ray tracing simulation results using the generated model are compared with measurements and the comparison shows that the trend of normalized received power is similar.

Keywords: ray trace, point cloud data, indoor environment

Classification: Antennas and Propagation

References

[1] Z. Chen, K. Saito, W. Okamura, Y. Kishiki, and J. Takada, “3D point cloud data simplification method for electromagnetic simulation in indoor environment,” IEICE Technical Report, SRW2018-49, pp. 19–24, Jan. 2019.

[2] R. B. Rusu and S. Cousins, “3D is here: Point Cloud Library (PCL),” 2011 IEEE International Conference on Robotics and Automation, Shanghai, pp. 1–4, 2011. DOI:10.1109/ICRA.2011.5980567

[3] M. A. Fischler and R. C. Bolles, “Random sample consensus: A paradigm for model fitting with applications to image analysis and automated cartography,” \textit{Commun. ACM}, vol. 24, no. 6, pp. 381–395, 1981. DOI:10.1145/358669.358692

[4] K. Saito, J.-I. Takada, and M. Kim, “Dense multipath component characteristics in 11-GHz-band indoor environments,” \textit{IEEE Trans. Antennas Propag.}, vol. 65, no. 9, pp. 4780–4789, Sept. 2017. DOI:10.1109/TAP.2017.2728087

[5] E. Costa, “Ray tracing based on the method of images for propagation simulation in cellular environments,” Tenth International Conference on
1 Introduction

With the advent of IoT (Internet of Things) services in recent years, various industries are now working towards building and operating wireless networks for their own application. Traditionally, mobile operators are the ones who are largely involved in designing wireless networks, and they have widely used radiowave propagation simulation for this task.

In order to perform an accurate radiowave propagation simulation, a detailed environmental model is often required, especially so when simulating in the millimeter wave bands. As such, it is difficult for users to design their own wireless network and this poses a challenge for the dissemination of IoT services. This research aims to support these users by allowing them to design a wireless network with less effort.

Recently, point cloud data, which is a set of data points in 3D space acquired by scanning the environment, has become increasingly accessible. For radiowave propagation simulations like ray tracing, point cloud data cannot be used as is and needs to be simplified into planes. In this paper, in order to reduce the burden of constructing a model for ray tracing simulation, a method that automatically generates a simplified planar model using 3D point cloud data that can be easily acquired is presented. To check if the generated simplified model is appropriate for radiowave propagation, ray tracing simulation results using the generated simplified model are compared with channel sounding measurements.

This paper is organized in the following manner. Section 2 of this paper explains the method that was used to create a plane model that is suitable for use with ray tracing simulations. The details of the point cloud data acquisition and radiowave propagation measurement and simulation are presented in section 3 and section 4 evaluates the results. Finally, section 5 presents the conclusion.

2 Plane model creation method using point cloud data

For many simulation programs, 3D computer aided design (CAD) models are often more appropriate and convenient to use as compared to point cloud data. Despite this, 3D CAD models may not be readily available, especially for places that are built long ago. This gives rise to the need to create a 3D CAD model from scratch which could take a lot of time. In order to create 3D models from point cloud data, some researches focus on extracting planes to reconstruct the environment. However, they do not guarantee that the created model will be free of gaps which is important when doing radiowave propagation simulations like ray tracing. It also does not address the case when there are walls that extrude from the main wall. This paper proposes a method to fix these issues based on our previous work which used a laser scanner to acquire the point cloud data [1]. Furthermore, indoor environments are mostly composed of vertical and horizontal planes. Taking this into
consideration, this paper proposes to automate and simplify the generation of 3D indoor models from point cloud data suitable for ray tracing method. To do this, the following 8 processes were deployed using the Point Cloud Library (PCL) [2].

(1) Down sampling. One point cloud file contains millions of points. To speed up and decrease memory usage of the subsequent process, the Voxelized Grid approach is used as the down sampling filter.

(2) Segmentation. This process groups point cloud data into multiple clusters. The basic idea is to examine the curvature and normal values of surrounding points to determine if the points should be regarded as the same region.

(3) Plane detection. The Random Sample Consensus (RANSAC) [3] method was used to extract a finite plane model for each cluster. The plane model is defined by the following function:

\[ ax + by + cz + d = 0 \]  (1)

The boundaries of each plane are then determined by the outermost point of its corresponding cluster.

(4) Classification of planes. The planes that are detected in the previous step are categorized into horizontal, vertical or ceiling/floor. Ceiling/floor planes are special horizontal planes and are found by finding the 2 largest horizontal planes. The height of the room can then be estimated by taking the difference in height between these 2 planes. In the subsequent processing, the horizontal and ceiling/floor planes will not be used.

(5) Plane regeneration. All the main walls in a room are expected to start on the floor and end at the ceiling. In this step, a correction will be applied to alter the height of all the main walls. Main walls are defined as walls whose height is more than 95% of the room height. This step removes gaps between the main walls and the floor and ceiling.

(6) Wall combinations. In this step, the edges of wall planes are connected to remove gaps that appear due to inaccuracies in the point cloud data. For one edge of the plane, the nearest edge of the other planes is determined and another plane is then generated to connect the 2 planes. A distance threshold is set between 2 edges to prevent incorrect connections.

(7) Extension of planes. Vertical planes whose height are less than 95% of the room height are assumed to be planes that extrude from the main walls of the room. The algorithm in Fig. 1 is used to find these planes and extend it and connect it to the main wall planes. First, the projection of the small vertical plane on the wall plane is determined. Thereafter, the 4 planes that are required

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**Fig. 1.** Input and algorithm to find target wall for extension
to connect the edges of the small vertical plane to the main wall plane is subsequently determined. Lastly, the projection area of the small vertical plane is removed from the main wall plane. In this manner, main wall planes with holes can be correctly generated.

(8) Roof and ground filling. Using the vertices of the main wall planes, ceiling and floor planes are generated. It is assumed that both ceiling and floor planes are completely flat, i.e., existing beams or hanging lights are not reconstructed.

3 Measurement and simulation

The point cloud data that is used in this paper was acquired from a conference room as shown in Fig. 2(a) at Tokyo Institute of Technology, Ookayama Campus using a Structure Sensor (instead of a laser scanner in [1]) which can be easily mounted on an iPad and is relatively inexpensive. During data acquisition, there was no furniture. Data was taken from multiple positions, and the point cloud registration algorithm was used to merge the different data pieces together. Resolution of point cloud data ranges from 1 cm to 11 cm with an average of 8 cm. The size of the room is approximately 8.70 m × 17.60 m × 3.00 m.

![Figure 2](image)

(a) Measurement area  
(b) Propagation route  
(c) Measurement and simulation parameters

| Parameters                  | Measurement | Simulation               |
|-----------------------------|-------------|--------------------------|
| Center frequency            | 11 GHz      | 11 GHz                   |
| Bandwidth                   | 400 MHz     | -                        |
| Transmit power              | 10 mW       | 10 mW                    |
| Number of tones             | 2048        | -                        |
| Measurement resolution      | 2.5 ns      | -                        |
| BS/MS Antenna type          | 12-element circular antenna array (12 v-pol and 12 h-pol) | Omnidirectional antenna |
|                             | Beamwidth for each element: Horizontal: 100°; Vertical: 35° | Vertical polarization |
| BS/MS Antenna gain          | 6 dBi       | 0 dBi                    |
| BS/MS Antenna height        | 1.7 m       | 1.7 m                    |
| Model material              | -           | metal (relative permittivity = 1, conductivity = 1e7[S/m]) |
| Max number of Reflection,   | -           | 5                       |
| Max number of Diffraction,  | -           | 1                       |
| Max number of Ref + Diff    | -           | 5                       |

Fig. 2. (a) Photograph of measurement area. (b) Propagation route. (c) Measurement and simulation parameters.

In order to verify the usability of the generated simplified model in terms of radiowave propagation, a ray tracing simulation using RapLab was performed and
the results were compared to channel sounding measurements [4] in the same location. RapLab is a commercial ray tracing simulator based on the method of images [5]. To search for single reflection paths, an image of the receiver (Rx) with respect to a surface is created and a line is drawn from the transmitter (Tx) to the image of Rx to determine the intersection point on the surface. The reflection path will then be composed of Tx, the intersection point and Rx. The image method can be extended to determine multiple interactions.

The specifications of the channel sounder are shown in Fig. 2(c). The Tx/Rx antenna arrays are 12-element circular arrays with dual-polarized patch antennas. During measurement, there was no furniture and the mobile station (MS) was moved to the left towards the base station (BS) at a speed of approximately 0.25 m/s as shown in Fig. 2(b). The total number of MS location is 140. For the ray trace simulation, a maximum of 5 reflections and maximum of 1 diffraction was used. The front and back part of the room are walls mounted with metallic whiteboard. The floor is concrete while the ceiling is lined with fluorescent lights inside metallic casings. Since most materials are metal, the simulation model material used is metal.

4 Comparison of results with actual measurements

Fig. 3(a) shows the original point cloud data and Fig. 3(b) shows the generated simplified model, respectively. The size of the generated model is 8.40 m × 17.52 m × 2.90 m as compared to the actual size which is 8.70 m × 17.60 m × 3.00 m. There are 2 extended vertical planes shown on the right wall of Fig. 3(b). The simplified results give 3.50 m × 2.10 m × 1.02 m, and 5.23 m × 2.10 m × 1.20 m, as compared to the actual size of 6.14 m × 2.21 m × 0.96 m, and 5.84 m × 2.21 m × 1.06 m respectively. The whole program can be finished within a few seconds.
In terms of radiowave propagation, Fig. 3(c) compares the measured and simulated delay profiles when the MS is at 0 m (farthest location from BS). Since the measurement resolution is 2.5 ns, the simulated delay profiles are also combined for every 2.5 ns. For comparison, both measured and simulation results are normalized using their respective maximum power. It can be seen from the graph that the peak positions are in good agreement between the measurement and the simulation, and the specular multipath components (SMC) are well reproduced by the simulation. Despite this, it can also be seen that the normalized power is generally lower in the simulation as compared to the measurement. This is largely due to the fact that dense multipath components (DMC) due to scattering effects such as diffuse scattering are not considered in the simulation. However, since material properties such as wall roughness have a larger effect on DMC than the shape of the room, it can be concluded that the generated simplified model from point cloud data, which can reproduce the SMC results relatively well, is a good representation of the actual room for radio propagation simulations.

Fig. 3(d) shows a comparison of cumulative distribution function (CDF) obtained from the received power at all 140 measurement points for the measurement and simulation. Received power for each position $s$, $P_r(s)$, is divided by the received power of the nearest MS to BS, $P_r(s = s_1)$, to obtain $P'_r(s)$ as shown in Eq. (2). Note that $P_r(s)$ is the sum of the delay profile at position $s$.

$$P'_r(s) = \frac{P_r(s)}{P_r(s = s_1)}$$

Fig. 3(d) shows a comparison of $P'_r(s)$. The graph shows the range of the measurement received power and range of the simulated received power differs by less than 2 dB.

5 Conclusion

In this paper, a method was proposed to automatically generate a simplified planar model from 3D point cloud data gathered using relatively inexpensive and easy to use Structure Sensor, for use in ray tracing simulation. This could reduce the burden and time to construct 3D models for simulation. The simplified model could replicate the ceilings, floors, main walls and extend planes without gaps which is important for radiowave propagation simulations.

Ray tracing simulation results using the generated simplified model compared with measurements show that the specular multipath components are well reproduced by the simulation although the normalized power is generally lower in the simulation due to the absence of dense multipath components. The CDF for all measurement points show the same trend for both measurement and simulation, and the difference is less than 2 dB.

The proposed method assumes that the ceiling and floor of room are parallel and flat. Objects in the room are not modeled and only rectangular planes are constructed. Material properties are also not defined automatically. Future works will address the above constraints to develop a more generalized method.

Acknowledgments

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Doppler spectrum consideration on V2V communication for platooning

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\textbf{Abstract:} 5th generation mobile communication (5G) is being actively researched and developed. We focus on “Ultra-Reliable and Low-Latency Communication” and consider the application of truck platooning as a use case. The latest research on the Doppler spectrum of the vehicle-to-vehicle (V2V) direct communication of platooning vehicles has revealed that the direct wave is not Doppler-shifted and that a component with twice the maximum Doppler frequency was observed.

This paper focuses on road surface reflections and scattered waves, and shows that there is no Doppler shift near the regular reflection point and that the Doppler shift near the transmission/receiving point.

\textbf{Keywords:} 5G mobile communication system, V2V direct communication, platooning, Doppler spectrum, Doppler shift

\textbf{Classification:} Antennas and Propagation

\textbf{References}

[1] M. Mikami, R. Yamaguchi, and H. Yoshino, “Field experimental evaluation on 5G low latency communications for application to truck platooning,” IEICE Technical Report, RCS2018-210, pp. 181–186, Nov. 2018.
[2] K. Seki and M. Hamaguchi, “A study on inter-vehicle communication for truck platooning (2nd Report),” Proc. 17th ITS World Congress, Busan, South Korea, Oct. 2010.
[3] A. Keiji, “Current activities of development on the automated truck platoon—On development status of automated driving technology—,” IPSJ Magazine, vol. 54, no. 4, pp. 303–309, Sept. 2013.
[4] K. Serizawa, K. Tomimoto, M. Miyashita, and R. Yamaguchi, “Doppler spectrum evaluation on V2V communication for platooning,” IEICE Commun. Express, vol. 8, no. 5, pp. 184–189, May 2019. DOI:10.1587/comex.2019XBL0008
1 Introduction

The advent of 5th mobile communication (5G) has triggered intense research and development activities to understand the radio wave propagation characteristics of the new frequency band expected to be used. 5G has three main features: Enhanced Mobile BroadBand (eMBB), multiple simultaneous connections (mMTC), and high-reliability and low-latency (URLLC). Among them, we focus on URLLC, and considering truck platooning as a one of use case [1].

Truck platoons are controlled by electronic connection technology (direct vehicle-to-vehicle communication), and multiple trucks are packed together to form a train. Japan is investigating whether it is possible to man the lead vehicle with all following vehicles being unmanned. This goal is to significantly reduce labor costs, offset the driver shortage problem, increase operation efficiency, and reduce energy consumption [2, 3].

We measure the Doppler shift present in platooning V2V direct communication, and pay close attention to the spectral broadening. In previous letter, when directional antenna was used during platooning, twice the maximum Doppler frequency was observed, and it was described that multiple scattering by the road surface and vehicles was the cause [4]. In this letter, the spectrum spread up to the maximum Doppler frequency is observed, using an omnidirectional antenna. It is experimentally clarified that the cause is the difference between the incident angle/exit angle of the road scattered wave.

2 Convoy travel V2V Doppler spectrum

2.1 Measurement overview

Fig. 1(a) outlines the experimental apparatus used to assess the Doppler spectrum in platooning V2V direct communication. Two trucks were lined up, with the leading vehicle as the receiving system and the following vehicle as the transmitting system. For the transmission system, a standard signal generator (SG) was used to transmit an unmodulated continuous wave (CW signal). A spectrum analyzer (SA) was used as receiving system. In order suppress frequency stability concerns, rubidium oscillators using GNSS signals were installed in the transmission and reception systems to synchronize the reference signal.

Fig. 1(d) shows the measurement parameters. The measurements were made in the 28 GHz band (center frequency 27.9 GHz) used by 5G. Here, the maximum Doppler frequency $f_D$ is expressed by Eq. (1).

$$f_D = \left(\frac{v}{c}\right) \times f_c$$

$v$: travel speed, $c$: speed of light, $f_c$: Transmitter center frequency

The maximum Doppler frequency, $f_D$, is 775 Hz and 1550 Hz, as given by Eq. (1) when the center frequency is 27.9 GHz and the moving speed is 30 km/h and 60 km/h. The distance between vehicles in platooning is expected to be 4 m to 8 m so as to reduce air resistance and improve fuel efficiency, but for this measurement the separation was kept to about 10 m for safety as both vehicles were manned. Patterns were measured for two types of transmitting and receiving antennas: an omnidirectional antenna and a quad-ridge horn antenna. The antenna
height was 0.75 m at both the transmitter and receiver. Antennas are generally installed as high as possible, but as most trucks will carry containers owned by someone else, the antenna height is determined by the rear chassis arrangement. Using the same antenna height simplifies the measurement setup.

2.2 Platooning and Doppler spectrum
Both the leading car and the following car were driven at the same speed, and the Doppler spectrum was observed. Fig. 1(b) and 1(c) show the spectra observed when omnidirectional antennas and quad-ridge horn antennas were used, respectively. The horizontal axis shows the frequency shift (frequency difference from the transmission frequency), and the vertical axis shows the reception level. The spectrum without Doppler shift is considered to be the direct wave and road surface regular reflection waves. From Fig. 1(b), when the omnidirectional antenna is used, the spectrum spreads with a gentle Doppler shift of 1500 Hz and 3000 Hz, at the

![Experimental setup and measurement results in platooning](image-url)
driving speeds of 30 km/h and 60 km/h, respectively. In Fig. 1(c), a weak Doppler shifted spectrum was observed at about 1500 Hz and 3000 Hz [4]. These spectra are approximately twice the maximum Doppler frequency. These double waves are caused by multiple waves scattered from stationary features such as road surfaces [4]. Here, in order to clarify the cause of the gentle spectrum broadening in Fig. 1(b), we focused on the waves scattered from the road surface.

3 Experiment on road surface scattered waves

As a result of measuring the Doppler shift, spectral broadening was observed when the omnidirectional antenna was used. In order to identify the cause, we paid close attention to the incident angle/exit angle of the road surface reflection/scattered wave and confirmed the influence of Doppler shift.

3.1 Examination of road scattering model

Fig. 2(a) shows a Doppler shift model for platooning. Given that the following truck transmits while the leading truck receives, ①CW transmission signal with the carrier frequency $f_c$ is transmitted while the following vehicle is running. At this time, as the transmitting antenna is moving forwards the frequency is raised by $\Delta f$.

① As the receiving antenna is moving forwards at the same speed, the received frequency is lowered by $\Delta f$ so,

$$f_{R X} = f_c + \Delta f - \Delta f = f_c.$$ 

② The incident wave on the road surface changes frequency according to the angle between the road surface and the antenna, and becomes $f_{T X}$ as per Eq. (2).

$$f_{T X} = f_c + f_c \times (v/c) \times \cos \theta_{T X}$$ 

(2)

③ When the road surface wave reaches the receiving antenna, the frequency changes according to the angle between the road surface and the antenna, as in Eq. (2), and becomes $f_{R X}$ as per Eq. (3). At this time, the frequency is lowered because the receiving antenna is moving away from the reflection point.

$$f_{R X} = f_c + f_c \times (v/c) \times \cos \theta_{T X} - f_c \times (v/c) \times \cos \theta_{R X}$$

$$= f_c \{1 + (v/c) (\cos \theta_{T X} - \cos \theta_{R X})\}$$

(3)

The above formula shows that with normal reflection, $\theta_{T X} = \theta_{R X}$, there is no Doppler shift. Moreover, for the $(\theta_{T X}, \theta_{R X})$ values of $(90^\circ, 0^\circ)$, $(0^\circ, 90^\circ)$, the Doppler shift at the time of reception is roughly $-f_D, f_D$. As shown in Fig. 2(a), at each road surface scattering point, $\theta_{T X}, \theta_{R X}$ is uniquely determined, $-1 < (\cos \theta_{T X} - \cos \theta_{R X}) < 1$, the Doppler frequency received by the first car, $f_{R X}$, lies in the range $-f_D < f_{R X} < f_D$. Fig. 2(b) shows that for 10 m of Inter-vehicular distance, the distance from the point immediately below the transmitting antenna to the road surface scattering point and the normalized Doppler shift frequency related by $(f_D \times (\cos \theta_{T X} - \cos \theta_{R X}))$. From this figure, it can be seen that the road surface scattered wave doppler shifts by $-f_D$ near the transmission point and doesn’t doppler shift over a wide range near the normal reflection point as the intermediate point. On the other hand, it can be confirmed that a $+f_D$ shift occurs in near the
receiving point. This is a characteristic corresponding to the difference between the transmission angle $\theta_{TX}$ and the reception angle $\theta_{RX}$ as described above.

3.2 Verification of road surface scattering model

For the propagation paths from the transmitting antenna to the receiving antenna of $(90^{\circ}, 5.5^{\circ})$ and $(0^{\circ}, 5.5^{\circ})$ ($\theta_{TX}, \theta_{RX}$, respectively), the Doppler frequencies at the time of reception were assumed to be approximately $-f_D$ and $f_D$, respectively. This verification experiment used quad-ridge horn antennas used as the transmitting and receiving antennas, and the (transmitting antenna angle and receiving antenna angle) were set to $(90^{\circ}, 0^{\circ})$, $(0^{\circ}, 90^{\circ})$ to limit the road surface scattering path; measurements were conducted to determine if a Doppler shifted spectrum could be observed in the vicinity of $-f_D$ and $f_D$. Fig. 3(a) shows the measurement parameters for the verification experiment, while Fig. 3(b) and (c) show the measurement results. Fig. 3(b) shows that the received spectrum level near $-f_D$ is high because the transmitting antenna is tilted downwards at 90°. On the other hand, in Fig. 3(c), the level of the received spectrum near $f_D$ is high. Therefore, the scattered wave from points, especially near the transmission and receiving point, away from the median reflection point on the road surface are considered to be Doppler shifted.

(a): Measurement parameters

| Parameter         | Value          |
|-------------------|----------------|
| Center frequency  | 27.9 GHz       |
| Frequency span    | 10 kHz         |
| Transmitted signal| CW             |
| Transmission output| 15 dBm        |
| RBW               | 50 Hz          |
| VBW               | 50 Hz          |
| Sweep time        | 1000 ms        |
| Inter-vehicular distance | 10 m      |
| Antenna           | Quad-ridge horn|
| Antenna height    | 0.75 m         |
| Polarization plane| V polarization |

Fig. 2. Examinations of scattering path
From these results, the gradual spread of the spectrum during platooning when using omnidirectional antennas is caused by the road surface incident angle/exit angle, shifting from the median reflection point, which yields the Doppler shift found.

4 Conclusion

We measured the Doppler spectrum for V2V direct communication using two platooned trucks; the spectrum spreading caused by waves reflected from road surfaces points was examined. Measurements showed that the regular reflected wave (reflection point is equidistant from Rx and Tx) exhibits no spectrum spreading. Furthermore, it was shown that non-regular scattered waves exhibit a broadened spectrum due to the difference between the road incident angle and the exit angle. It was quantitatively shown that road reflection points closer to the Tx antenna lower the Doppler shift, and reflection points closer to the Rx antenna raise the Doppler shift. In the future, we plan to conduct more detailed verification tests in environments closer to real-world use cases, such as when a truck train is overtaken.

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Fig. 3. Verification setup and measurement results
Steering of the circularly polarized beam from a spiral antenna

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Abstract: The circularly polarized axial and conical beams from a two-arm spiral antenna are investigated and the radiation phase distributions around the antenna axis are clarified. A tilted beam is created by superimposing the conical beam onto the axial beam. The tilted beam is steered in the azimuth plane when the two arms are excited by different amplitudes and phases. The beam direction of the tilted beam estimated using formulated phase distributions agrees with that obtained from a numerical analysis. The estimation with the formulated phase is simple and has the superiority over the determination of the complex and time-consuming numerical analysis.

Keywords: beam-steering, circularly polarized wave, spiral antenna

Classified: Antennas and Propagation

References

[1] H. Zhou, A. Pal, A. Mehta, D. Mirshekar-syahkal, and H. Nakano, “A four-arm circularly polarized high-gain high-tilt beam curl antenna for beam-steering applications,” IEEE Antennas Wireless Propag. Lett., vol. 17, no. 6, pp. 1034–1038, June 2018. DOI:10.1109/LAWP.2018.2830121
[2] H. Nakano, T. Abe, and J. Yamauchi, “Planar reconfigurable antennas using circularly polarized metalines,” IEEE Antennas Wireless Propag. Lett., vol. 18, no. 5, pp. 1006–1010, May 2019. DOI:10.1109/LAWP.2019.2907533
[3] J. Kaiser, “The Archimedean two-wire spiral antenna,” IRE Trans. Antennas Propag., vol. 8, no. 3, pp. 312–323, May 1960. DOI:10.1109/TAP.1960.1144840
[4] CST Computer Simulation Technology, Darmstadt, Germany, Microwave Studio. [Online]. Available: http://www.cst.com, Accessed on: Jan. 2020.

1 Introduction

Modern communication systems often require an antenna system that has a circularly polarized (CP) beam-steering function. For this, reconfigurable antennas have been proposed, \textit{e.g.}, in [1] and [2]. In these antennas, the CP beam-steering in the azimuth plane is achieved by selecting an activated element with a switching circuit. Note that the antenna systems in [1] and [2] have four feed points. A question arises as to whether CP beam-steering can be performed using two feed...
points. If this is achieved, we have a CP beam-steering antenna with a simple structure. From this background, we present a CP beam-steering antenna based on a two-arm Archimedean spiral antenna [3]. A simple formula for estimating the beam direction is derived.

2 Antenna configuration

Fig. 1 shows the configuration of a two-arm Archimedean spiral antenna. The two spiral arms are printed on a dielectric substrate of relative permittivity $\varepsilon_{r\text{-sub}} = 2.6$ and thickness $B = 0.8\,\text{mm}$. The dielectric substrate is placed above a conducting cavity whose diameter is $D_{\text{CAV}} = 82\,\text{mm} = 1.23\lambda_0$, and height is $H_{\text{CAV}} = 7.0\,\text{mm} = 0.105\lambda_0$, with $\lambda_0$ being the free-space wavelength at a design frequency of 4.5 GHz. The radial distance from the coordinate origin to the centerline of the spiral arm, $r_{\text{arm}}$, is defined by the Archimedean function $r_{\text{arm}} = a_{\text{spiral}}\phi_{\text{winding}}$, where $a_{\text{spiral}}$ is the arm growth constant and $\phi_{\text{winding}}$ is the winding angle: $a_{\text{spiral}} = 1.273\,\text{mm}/\text{rad}$ and $0.5\pi\,\text{rad} \leq \phi_{\text{winding}} \leq 8.5\pi\,\text{rad}$. The arm width is $w = 2.0\,\text{mm}$. Two feed points of the spiral arms, $F_1$ and $F_2$, are excited by voltage sources $V_1\,\text{volt}$ and $V_2\,\text{volt}$, respectively. A small conducting disc behind the spiral arms has a diameter of $2r_{\text{disc}} = 12.0\,\text{mm}$. The distance from the spiral plane to the small disc is $d_{\text{disc}} = 1.0\,\text{mm}$. To decrease undesirable reflection currents from the spiral arm ends, a ring-shaped absorber (ABS, ISFA EM, TDK production: relative permittivity $\varepsilon_{r\text{-abs}} = 1.92$ and $\tan\delta = 1.15$ at 4.5 GHz) is placed along the cavity wall. The thickness of the absorber is $t_{\text{abs}} = 11.0\,\text{mm}$.

![Diagram of antenna configuration](image-url)
3 CP beam-steering

A CP beam-steering with the spiral antenna is discussed in this section. Note that the results in this section are those obtained using an electromagnetic analysis solver (CST [4]).

First, we clarify the basic radiation from the spiral antenna. Fig. 2(a) shows the radiation pattern at a design frequency of 4.5 GHz when the feed points \( F_1 \) and \( F_2 \) are excited in balanced mode \((V_1 = 1\angle0°, V_2 = 1\angle180°)\). The red solid line denotes the co-polarized component (right-handed CP component \( E_R \)) of the radiation and the blue dotted line denotes the cross-polarized component (left-handed CP component \( E_L \)). The co-polarized component \( E_R \) is radiated in the

![Balanced mode radiation](image1)

![Balanced mode amplitude |ERH-bal| and phase ∠ERH-bal](image2)

![Unbalanced mode radiation](image3)

![Unbalanced mode amplitude |ERH-unbal| and phase ∠ERH-unbal](image4)

![Tilted beam](image5)

![Phase difference ∠ERH-unbal − ∠ERH-bal](image6)

![Beam direction of CP wave from the spiral antenna](image7)

**Fig. 2.** Radiation field from the spiral antenna at a design frequency of 4.5 GHz.
broadside direction (+z-direction). The amplitude $|E_{\text{RH-bal}}|$ and phase $\angle E_{\text{RH-bal}}$ of the $E_R$ in balanced mode are shown in Fig. 2(b), where the spherical coordinate angle $\theta$ is fixed to be a representative value of $\theta = 30^\circ$. It is found that $|E_{\text{RH-bal}}|$ is almost unchanged around the antenna axis (z-axis), while $\angle E_{\text{RH-bal}}$ changes by $360^\circ$ in an almost linear fashion.

Fig. 2(c) shows the basic radiation pattern at 4.5 GHz when the feed points $F_1$ and $F_2$ are excited in unbalanced mode ($V_1 = 1 \angle 0^\circ$, $V_2 = 1 \angle 0^\circ$). The maximum radiation intensity of the co-polarized component $E_R$ appears off the z-axis. The amplitude $|E_{\text{RH-unbal}}|$ and phase $\angle E_{\text{RH-unbal}}$ of the $E_R$ are shown in Fig. 2(d). The gradient of the $\angle E_{\text{RH-unbal}}$ differs from that of the $\angle E_{\text{RH-bal}}$, i.e., $\angle E_{\text{RH-unbal}}$ changes by $720^\circ$ in an almost linear fashion around the z-axis.

Second, we superimpose the basic unbalanced mode radiation onto the basic balanced mode radiation. The superimposition forms a tilted beam at azimuth angle $\phi = 0^\circ$, i.e., in the +x-direction, as shown in Fig. 2(e). This is due to the fact that difference in the phases of the basic balanced mode radiation and basic unbalanced mode radiation becomes zero at $\phi = 0^\circ$, as shown in Fig. 2(f), i.e., the two mode radiation phases are in-phase.

Third, we consider beam-steering. For this, the voltages at feed points $F_1$ and $F_2$ are changed, as shown in Eq. (1).

$$ (V_1, V_2)^T = (1 \angle 0^\circ, r \angle (180^\circ + \delta))^T $$

where $T$ denotes the transposition operator of a matrix, $r$ is called the excitation voltage amplitude, and $\delta$ is called the deviation angle. Fig. 2(g) shows the maximum radiation azimuth angle, $\phi_{\text{RH-max-CST}}$, when $r$ is varied and $\delta$ is fixed to be $\pm 90^\circ$. The red and blue dots are for $\delta = +90^\circ$ and $\delta = -90^\circ$, respectively. Thus, the CST analysis reveals that the CP beam is steered in the azimuth plane with change in $r$.

### 4 Estimation for the beam direction

In the previous section, CP beam-steering with the spiral antenna is performed, where the beam direction is determined by the electromagnetic analysis solver CST. In this CST analysis, first, we analyzed 3D radiation pattern for each $r$. Second, we searched the beam direction from the obtained 3D radiation pattern. Note that the numerical analysis is repeated for each $r$. This process is quite time-consuming.

In this section, we present a simple formula for estimating the beam direction.

We decompose Eq. (1) into a balanced mode component $V_{\text{bal}}$ and an unbalanced mode component $V_{\text{unbal}}$.

$$ V_{\text{bal}} \equiv A_{\text{bal}} \angle \phi_{\text{bal}} $$

$$ V_{\text{unbal}} \equiv A_{\text{unbal}} \angle \phi_{\text{unbal}} $$

where $A_{\text{bal}}$ and $\phi_{\text{bal}}$ are called the balanced mode excitation amplitude and phase, respectively; and $A_{\text{unbal}}$ and $\phi_{\text{unbal}}$ are called the unbalanced mode excitation amplitude and phase, respectively. We choose $\delta$ to be $\pm 90^\circ$ to obtain $A_{\text{bal}} = A_{\text{unbal}}$. The following relationships are held for $\delta = \pm 90^\circ$. 
The difference of $A_{\text{unbal}}$ relative to $A_{\text{bal}}$ is defined as mode phase difference $\Delta \phi$. From Eqs. (5) and (6), $\Delta \phi$ is expressed as Eq. (7).

$$\Delta \phi \equiv \phi_{\text{unbal}} - \phi_{\text{bal}} = \tan^{-1}\left(\frac{\pm r}{1 - r^2}\right) \quad \text{for} \quad \delta = \pm 90^\circ$$

That is, $\Delta \phi$ is varied with $r$. Note that $\angle E_{\text{RH-bal}}$ in Fig. 2(b) and $\angle E_{\text{RH-unbal}}$ in Fig. 2(d) are obtained for a situation of ($r = 1$, $\Delta \phi = 0$).

We estimate a maximum radiation azimuth angle of $\phi_{\text{RH-max}}$. The maximum radiation component $E_R$ appears when Eq. (8) is satisfied.

$$\angle E_{\text{RH-unbal}}(\phi_{\text{RH-max}}) + \Delta \phi = \angle E_{\text{RH-bal}}(\phi_{\text{RH-max}})$$

As shown in Fig. 2(b), the basic balanced mode radiation phase $\angle E_{\text{RH-bal}}(\phi)$ changes by 360° in an almost linear fashion. Therefore, $\angle E_{\text{RH-bal}}$ is approximated as Eq. (9).

$$\angle E_{\text{RH-bal}}(\phi) = -\phi + \angle E_{\text{RH-bal}}(0)$$

where $\angle E_{\text{RH-bal}}(0)$ denotes the phase of the basic balanced mode radiation at an azimuth angle of $\phi = 0^\circ$.

On the other hand, the basic unbalanced mode radiation phase $\angle E_{\text{RH-unbal}}(\phi)$ changes by 720° in an almost linear fashion, as shown in Fig. 2(d). This is approximated as Eq. (10).

$$\angle E_{\text{RH-unbal}}(\phi) = -2\phi + \angle E_{\text{RH-unbal}}(0)$$

Substituting Eqs. (7), (9), and (10) to Eq. (8), $\phi_{\text{RH-max}}$ is given by

$$\phi_{\text{RH-max}} = \angle E_{\text{RH-unbal}}(0) - \angle E_{\text{RH-bal}}(0) + \tan^{-1}\left(\frac{\pm 2r}{1 - r^2}\right) \quad \text{for} \quad \delta = \pm 90^\circ$$

where $\angle E_{\text{RH-bal}}(0)$ and $\angle E_{\text{RH-unbal}}(0)$ are given by the values for the blue dotted lines at $\phi = 0^\circ$ in Figs. 2(b) and (d), respectively.

To confirm the validity of Eq. (11), we compare the formulated value $\phi_{\text{RH-max}}$ with the CST value $\phi_{\text{RH-max-CST}}$. The solid line in Fig. 2(g) shows $\phi_{\text{RH-max}}$ for $\delta = +90^\circ$ and the broken line shows $\phi_{\text{RH-max}}$ for $\delta = -90^\circ$. There is good agreement between $\phi_{\text{RH-max}}$ and $\phi_{\text{RH-max-CST}}$ for both $\delta$ cases. Thus, the validity of Eq. (11) is confirmed. Note that the solid line and broken line are symmetric with respect to a dotted line, $\angle E_{\text{RH-unbal}}(0) - \angle E_{\text{RH-bal}}(0)$, which is specified by symbols (x) in Fig. 2(g). The estimation of the beam direction by Eq. (11) is simple and leads to less computational burden, compared with the repeated CST numerical analysis.

For additional observation, the CST numerical and experimental radiation patterns in the azimuth plane (at $\theta = 30^\circ$) are shown in Fig. 3, where $\delta$ is fixed to be +90°. It is clear that the CP beam is steered in the azimuth plane with change in $r$. The experimental results agree with the CST numerical analysis results.
Some comments are made here. (1) the beam direction in the elevation plane $\theta_{RH-max-CST}$ is almost constant when the beam is rotated around the antenna axis. This is attributed to the fact that $A_{bal}$ equals $A_{unbal}$ irrespective of the value of $r$ for $\delta = \pm 90^\circ$. (2) the equalization of $A_{bal}$ to $A_{unbal}$ infers that the gain and axial ratio (AR) in the beam direction will be almost constant. The gain is approximately 7.0 dBi and the AR is approximately 0.8 dB. (3) the voltage standing wave ratio (VSWR) is less than two around the design frequency.

![Fig. 3. Radiation pattern in the azimuth plane where $r$ is changed and $\delta$ is fixed to be $+90^\circ$.](image)

5 Conclusion

Steering of the CP beam from a two-arm spiral antenna has been discussed. First, the radiation in balanced and unbalanced modes from the spiral antenna has been analyzed. Second, the phase distributions for these two modes have been formulated. Third, using the formulated phase distributions, we have derived a simple formula that estimates the steered beam direction $\phi_{RH-max}$ in the azimuth plane. The derived formula provides a good estimation for $\phi_{RH-max}$. The formula is simple and less computational burden, compared with the repeated numerical analysis using a commercially available EM solver.
Extended beamforming by optimum 2-D sparse arrays

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Abstract: This paper presents an extended beamforming method by two-dimensional (2-D) Sparse Arrays. The authors have proposed optimum 2-D sparse arrays and evaluated them in the sense of direction of arrival (DOA) estimation accuracy. We develop a way of beamforming by the difference co-array of 2-D sparse arrays, and evaluate their beamforming performance as well as bit error rate (BER) characteristics through computer simulation.

Keywords: 2-D sparse array, beamforming, BER characteristics

Classification: Antennas and Propagation

References

[1] H. L. Van Trees, Optimum Array Processing: Part IV of Detection, Estimation, and Modulation Theory, Wiley, 2002.

[2] W. K. Ma, T. H. Hsieh, and C. Y. Chi, “DOA estimation of quasi-stationary signals with less sensors than sources and unknown spatial noise covariance: A Khatri-Rao subspace approach,” IEEE Trans. Signal Process., vol. 58, no. 4, pp. 2168–2180, Apr. 2010. DOI:10.1109/TSP.2009.2034935

[3] C. L. Liu and P. P. Vaidyanathan, “Hourglass arrays and other novel 2-D sparse arrays with reduced mutual coupling,” IEEE Trans. Signal Process., vol. 65, no. 13, pp. 3369–3383, July 2017. DOI:10.1109/TSP.2017.2690390

[4] Y. Iizuka and K. Ichige, “Optimum linear array geometry for 2q-th order cumulant based array processing,” Proc. Int. Workshop. Compressed Sensing and its application to Radar, Multimodal Sensing, and Imaging, pp. 198–201, May 2016. DOI:10.1109/CoSeRa.2016.7745730

[5] S. Nakamura, S. Iwazaki, and K. Ichige, “Optimum 2-D sparse array and its interpolation via nuclear norm minimization,” Proc. Int. Sympo. Circuits and Systems, pp. 1–5, May 2019. DOI:10.1109/ISCAS.2019.8702741

[6] S. Nakamura, S. Iwazaki, and K. Ichige, “Optimization and hole interpolation of 2-D sparse arrays for accurate direction-of-arrival estimation,” submitted to IEICE Trans. Commun.

[7] S. Iwazaki and K. Ichige, “Extended beamforming by sum and difference composite co-array for real-valued signals,” IEICE Trans. Fundamentals, vol. E102-A, no. 7, pp. 918–925, July 2019. DOI:10.1587/transfun.E102.A.918

[8] B. D. Carlson, “Covariance matrix estimation errors and diagonal loading in adaptive arrays,” IEEE Trans. Aerosp. Electron. Syst., vol. 24, no. 4, pp. 397–401, July 1988. DOI:10.1109/7.7181

[9] S. Nakamura, S. Iwazaki, and K. Ichige, “A note on beamforming method by 2-D sparse arrays,” Proc. IEICE Society Conf., no. B-1-141, Sept. 2019 (in Japanese).
1 Introduction

Two-dimensional (2-D) array antenna plays an important role in radar, sonar, and indoor/outdoor wireless communications [1]. Sparse arrays are nowadays very much attracted which can create virtual difference co-array by the help of Khatri-Rao product [2] and achieve the degree of freedom (DOF) of $O(N^2)$, where $N$ and $O(\cdot)$ denote the number of physical sensors and the order of computational cost, respectively. Hourglass array [3] has been proposed as a 2-D sparse array which can accurately estimate direction of arrivals (DOAs) while preserving small mutual coupling effect. The hourglass array has a property that its difference co-array becomes uniform rectangular array (URA) without hole, therefore we can accurately estimate DOAs by spatial smoothing preprocessing (SSP)-based DOA estimation algorithms like ESPRIT.

The authors have already studied that the array configuration of the hourglass array can further be modified to reduce mutual coupling effect by applying simulated annealing (SA) [4], and confirmed that we can enhance DOA estimation accuracy [5, 6]. Besides, two of the present authors have developed an extended beamforming method for 1-D sparse arrays and achieved very precise beam patterns [7]. However, only the array configuration and DOA estimation accuracy was discussed in [4, 5, 6, 7]. Evaluation of beamforming performance is mandatory for communication applications of sparse array.

In this paper, we present an extended beamforming method for 2-D sparse arrays. Similarly to [7], we introduce diagonal loading (DL) operation [8] and minimum variance distortionless response (MVDR) beamforming to realize its main-beam and null steering that can greatly suppress interference waves. We evaluate the beamforming performance [9] and bit error rate (BER) characteristics of the 2-D sparse arrays through computer simulation.

2 Signal model

Suppose that $D$ uncorrelated signal sources impinge on a 2-D sensor array in an additive white Gaussian noise (AWGN) environment, where the signals and noises are statistically independent. The array aperture and the sensor location are respectively given by $N_x \times N_y$ and $n d$, where $n = (n_x, n_y) \in \mathbb{Z}^2$ is an integer-valued vector, and $d = \lambda/2$ is the minimum separation between sensors, and $\lambda$ is the wavelength of incoming sources. Assume that the sensor location $n$ forms a set $\mathcal{S}$, then the sensor input $x_{\mathcal{S}}$ on $\mathcal{S}$ can be modeled in a similar manner with [3] as

$$x_{\mathcal{S}} = \sum_{i=1}^{D} A_i \mathbf{c}_{\mathcal{S}}(\tilde{\theta}_i, \tilde{\phi}_i) + \mathbf{u}_{\mathcal{S}},$$

where the $i$-th source is with the complex amplitude $A_i \in \mathbb{C}$, the azimuth $\phi_i \in [0, 2\pi]$ and elevation $\theta_i \in [0, \pi/2]$. The element of the steering vector $\mathbf{c}_{\mathcal{S}}(\tilde{\theta}_i, \tilde{\phi}_i)$ as

$$\tilde{\theta}_i = \frac{d}{\lambda} \sin \theta_i \cos \phi_i,$$

$$\tilde{\phi}_i = \frac{d}{\lambda} \sin \theta_i \sin \phi_i,$$
corresponding to the sensor at \( \mathbf{n} = (n_x, n_y) \) is given by \( e^{j2\pi(n_x\phi_{y} + \phi_{n})} \). The mutual coupling matrix \( \mathbf{C} \) is characterized by its entries \( \langle \mathbf{C} \rangle_{n_1, n_2} \):
\[
\langle \mathbf{C} \rangle_{n_1, n_2} = \left\{ \begin{array}{ll}
c(||\mathbf{n}_1 - \mathbf{n}_2||_2), & \|\mathbf{n}_1 - \mathbf{n}_2\|_2 \leq B, \\
0, & \text{otherwise,}
\end{array} \right.
\]
where \( \mathbf{n}_1, \mathbf{n}_2 \in \mathbb{S} \) denote the sensor location, \( B \) is the maximum sensor separation where the mutual coupling effect exists, and \( c(\cdot) \) is the mutual coupling coefficient given by \( c(0) = 1 \) and \( |c(k)/c(\ell)| = \ell/k \) for \( k, \ell > 0 \) \[3\]. Here, the covariance matrix \( \mathbf{R}_S \) of the array \( \mathbb{S} \) is given by \( \mathbf{R}_S = \mathbb{E}[\mathbf{x}_0 \mathbf{x}_0^H] \).

We also define the difference co-array \( \mathbb{D} = \{\mathbf{n}_1 - \mathbf{n}_2 \mid \mathbf{n}_1, \mathbf{n}_2 \in \mathbb{S}\} \) for any 2-D sparse array \( \mathbb{S} \). Then the input vector of the difference co-array \( \mathbf{x}_D \in \mathbb{C}^{N_D \times 1} \) can be obtained by vectorizing the matrix \( \mathbf{R}_S \) while removing duplicated entries \[3\], where \( N_D \) denotes the number of elements in difference co-array. Note that the physical array itself can resolve up to \((N - 1)\) signals while that the higher DOF of the difference co-array \( \mathbb{D} \) enables us to identify \( O(N^2) \) uncorrelated signals.

### 3 Proposed method

We propose a beamforming method for difference co-arrays and introduce modulation and demodulation scheme of extended signals.

#### 3.1 MVDR beamforming with diagonal loading

The physical array output \( \mathbf{y}_S = \mathbf{w}_S^H \mathbf{x}_S \in \mathbb{C} \) is the inner product of the complex weight vector \( \mathbf{w}_S \in \mathbb{C}^{N \times 1} \) and the physical array input \( \mathbf{x}_S \in \mathbb{C}^{N \times 1} \). The weight \( \mathbf{w}_S \) can be calculated as an MVDR beamforming weight by
\[
\mathbf{w}_S = \frac{R_S^{-1}v(\theta, \phi)}{v_S(\theta, \phi)^H R_S^{-1}v_S(\theta, \phi)},
\]
where \( v_S(\theta, \phi) \in \mathbb{C}^{N \times 1} \) denotes the array steering vector of a desired wave.

We develop a beamforming method for the difference co-arrays of 2-D sparse arrays. The beamforming by the virtual array can be accomplished by defining the virtual steering vector \( \mathbf{v}_D(\theta, \phi) \) at the location \( (n'_x, n'_y) \) as \( e^{j2\pi(n'_x\phi_{y} + \phi_{n'})} \), where \( (n'_x, n'_y) \in \mathbb{D} \). Therefore we calculate the virtual array output \( \mathbf{y}_D = \mathbf{w}_S^H \mathbf{x}_D \in \mathbb{C} \) where the steering vector of the difference co-array \( \mathbf{v}_D(\theta, \phi) \in \mathbb{C}^{N_D \times 1} \) and
\[
\mathbf{w}_D = \frac{R_S^{-1}v_D(\theta, \phi)}{v_D(\theta, \phi)^H R_S^{-1}v_D(\theta, \phi)} \in \mathbb{C}^{N \times 1},
\]
\[
R_D = \mathbb{E}[\mathbf{x}_D \mathbf{x}_D^H] \approx \frac{1}{K} \sum_{k=1}^{K} \mathbf{x}_D(k) \mathbf{x}_D^H(k) \in \mathbb{C}^{N \times N_D},
\]
Note that we create the input vector of the difference co-array \( \mathbf{x}_D \) for each snapshot so that to be used in demodulation. Its number of snapshots becomes same with that of the physical array \( K \). The covariance matrix \( \mathbf{R}_D \) may become singular in case of small number of snapshots \( K \) (\( < D \)), then we cannot directly apply the MVDR beamforming of \( (6) \). In such case, we first apply the DL operation \[7\] to the covariance matrix \( \mathbf{R}_D \) for rank restoration, i.e.,
\[
\mathbf{R}_{DL} = \mathbf{R}_D + \delta \mathbf{I}_{N_D},
\]
where $\delta$ is the loading parameter, and $I_{N_0}$ denotes the $N_D \times N_D$ identity matrix. Using the matrix $R_{DL}$ and the steering vector $v_D(\theta, \phi)$, the MVDR beamformer weight $w_{DL}$ is given as

$$w_{DL} = \frac{R_{DL}^{-1}v_D(\theta, \phi)}{v_D(\theta, \phi)^H R_{DL}^{-1}v_D(\theta, \phi)},$$

(9)

and then we have the output: $y_{DL} = w_{DL}^H x_D \in \mathbb{C}$.

### 3.2 Modulation and demodulation of virtual signals

We introduce modulation and demodulation scheme of the signals at virtual array elements. Removing the mutual coupling matrix $C$ from (1) for simplicity, we can rewrite (1) as

$$x_0(k) = VA + u_0(k),$$

(10)

where $V = [v_0(\Delta f_1, \phi_1), \ldots, v_0(\Delta f_{N_D}, \phi_{N_D})]$ and $A = [A_1, \ldots, A_D]^T$. In case we generate input signals of difference co-array for each snapshot, the array input vector $x_D(k)$ at the time index $k$ can be represented as

$$x_D(k) = vec(R_D(k)) = (V^* \otimes V)p + \sigma^2 \cdot vec(I_N),$$

(11)

where $vec(\cdot)$ is vectorization operator, $\otimes$ is Khatri-Rao product operator [2], $\sigma^2$ denotes the noise power. Besides, the vector $p$ is given by $p = [\sigma_1^2, \ldots, \sigma_D^2]^T$, where $\sigma_i^2 = |s_i|^2$ denotes the power of the $i$-th incident signal. The desired signal power $\sigma_d^2$ can be extracted by beamforming, and is written as $\sigma_d^2 = |s_d|^2$ whereas $s_d$ denotes the desired signal. Note that $|s_d|^2$ becomes non-negative real and does not preserve the phase component of the original complex signal $s_d$, therefore any phase modulation are not applicable. Assume the case of BPSK modulation for simplicity, and we modify the signal representation as follows.

In transmitter scheme, BPSK symbol sequence $s_1(k)$ is generated where $s_1(k)$ takes either $-1$ or $1$. Basically we have BPSK transmission signal by applying cosine rolloff filter and multiplying carrier signal to $s_1(k)$. Instead, we employ shift operation $s_2(k) = s_1(k) + M$ before applying the filter, and then the sequence $s_2(k)$ takes either $(-1 + M)$ or $(1 + M)$ where $M$ denotes a positive integer shift parameter. By this operation, the transmitting signal becomes equivalent with ASK signals. Applying cosine rolloff filter to the sequence $s_2(k)$ and then we have $\tilde{s}_2(k)$, then the transmission signal $s(k)$ becomes $s(k) = \tilde{s}_2(k)e^{-j2\pi f_c k}$, where $f_c$ denotes carrier frequency.

In receiver scheme, the beamforming result of the extended array input signal $y_D(k) = w_{DL}^H x_D$ is first filtered, and then take a positive square-root operation $\sqrt{y}(k)$. Finally the obtained signal becomes inversely shifted version of $\tilde{s}(k)$ so that the symbol center becomes zero, and then $\tilde{s}(k)$ is demodulated.

### 4 Numerical examples

This section evaluates beamforming performance and BER characteristics of 2-D sparse arrays. The arrays to be compared are (a) URA, (b) Hourglass array [3], (c) the optimum array without holes (called “hole-free”), and (d) the optimum array with hole [5, 6], of which the number of physical elements are set to be common.
as 25. The aperture of the URA and that of the other arrays are (a) \(5 \times 5\) and (b)–(d) \(9 \times 9\), respectively. The total number of elements including virtual array is (a) \(9 \times 9 = 81\) for URA, (b), (c) \(17 \times 17 = 289\) for hole-free virtual URA, and (d) \(269\) for the URA-like difference co-array with 20 holes. Note that the holes in (d) are interpolated by nuclear norm minimization [6] before the beamforming operation.

We assume 1 desired and 23 interference waves to evaluate beamforming performance where the DOAs are equally distributed for azimuth and elevation angles as in Figs. 1 and 2, where the star at \((\theta, \phi) = (45^\circ, 150^\circ)\) indicates desired wave direction while the 23 circles denote interference wave directions. In case of evaluating BER characteristics, we assume 1 desired and 4 interference waves where the DOA of the desired wave is set to \((\theta, \phi) = (45^\circ, 150^\circ)\), and that of the interference waves are given by \((\theta, \phi) = (20^\circ, 30^\circ), (30^\circ, 200^\circ), (80^\circ, 90^\circ), (10^\circ, 300^\circ)\).

Both SNR and SIR are set to 0 dB, but the SNR is changed from \(-20\) to 10 dB in BER characteristics evaluation. The other parameters are set to as follows: \(K = 500\) snapshots, the DL parameter \(\delta = 10^5\), the signal shift parameter \(M = 2\), and the mutual coupling parameters \(c(1) = 0.3, B = 5\) as in [5, 6]. Note that the DL parameter \(10^5\) is determined based on the discussion in [7], where the values larger than \(10^5\) can work as well those less than \(10^5\) cannot well create null beams.

### 4.1 Beamforming performance

Figs. 1 and 2 respectively show the beamforming results in cases without/with the DL operation. We see from Figs. 1 and 2 that the sparse arrays (b), (c) and (d) can make more number of null-beams than URA due to larger aperture. However, the beam patterns in cases without DL operation in Fig. 1 cannot well suppress sidelobes because the covariance matrix becomes nearly singular as mentioned in

![Beamforming results in cases without DL operation.](image_url)
Section 3. The DL operation well improves this problem as in Fig. 2, we see that the large sidelobes in Fig. 1 are quite well suppressed.

4.2 BER characteristics

BER characteristics are also evaluated. Fig. 3 shows the behavior of SNR–BER characteristics in cases of (a) without DL and (b) with DL, respectively. We see from Fig. 3(a) that the demodulation is not at all accomplished.

On the other hand, Fig. 3(b) says that the BER characteristics becomes better as SNR gets larger. We emphasize that the optimum 2-D sparse arrays achieves better BER characteristics than URA and hourglass arrays. Note that the results in Figs. 2(b), 2(c) and 2(d) are almost same because of using the sparse arrays with a same aperture. The proposed optimum array has the advantages that it can suppress mutual coupling effect while improving BER performance.
5 Concluding remarks

This paper presented an adaptive beamforming method for the extended 2-D Sparse Arrays. We developed an MVDR-based beamforming method for the extended 2-D sparse arrays and evaluated its BER performance. We confirmed that the optimum 2-D sparse arrays can form a very minute beams and improve BER performance.
Evaluation of performance improvement of space selective modulation by increasing number of transmitting antennas

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Abstract: As a physical layer security technique, directional modulation (DM) using an array antenna was proposed. By the method, modulated signal can be correctly reconstructed in only the pre-specified direction from the array. On the other hand, Space Selective Modulation (SSM) was proposed. SSM shares the similar principle to DM. Using the technique, modulated signal is correctly reconstructed in only the specified space, not the specified direction by utilizing fading characteristics. The originally proposed SSM uses three or four transmitting antennas. However, the transmissible probability is not 100%. Because a modulated signal is formed by summing up phase-controlled channel coefficients in SSM, symbols of the all constellation points may not be formed depending on the channel conditions. It brings about a decrease in the secrecy capacity of the secure information transmission. In this paper, we consider solving the problem by increasing the number of antennas. We quantitatively evaluated the secrecy capacity by computer simulations. The results confirm the improvement of the total performance in the secrecy capacity of SSM.

Keywords: physical layer security, space selective modulation, directional modulation, radio propagation characteristics

Classification: Wireless Communication Technologies

References

[1] H. Sasaoka, “[Tutorial lecture] Information security at physical layer based on radio propagation characteristics,” IEICE Technical Report, vol. 115, no. 40, pp. 13–18, May 2015 (in Japanese).
[2] M. P. Daly and J. T. Bernhard, “Directional modulation technique for phased arrays,” IEEE Trans. Antennas Propag., vol. 57, no. 9, pp. 2633–2640, Sept. 2009. DOI:10.1109/TAP.2009.2027047
[3] J. Goto, K. Kihara, T. Takahashi, and H. Miyashita, “A study on modulation method for transmission system using phased array antennas,” IEICE Technical Report, vol. 113, no. 233, pp. 47–50, Oct. 2013 (in Japanese).
1 Introduction

Recently, physical layer security techniques based on information theoretic safety are attracting attention [1]. An example of the technique is Directional Modulation (DM) [2, 3, 4]. DM forms a desired modulated signal only in pre-specified direction using an array antenna. In the preceding studies of DM [2], the environment is assumed to be a line-of-sight propagation. So the DM scheme cannot be applied to non-line-of-sight propagation. On the other hand, a secure information transmission suitable for non-line-of-sight propagation has been proposed [5]. The scheme shares similar principle to DM. By using the unique channel coefficients the scheme can form the modulated signal only at legitimate receiver in non-line-of-sight fading environments. For this reason, it is called Space Selective Modulation (SSM). SSM decomposes a desired modulated signal into multiple phase modulated signals having constant amplitude based on the channel coefficient so that a desired signal is reconstructed at the legitimate receiver. Reference [5] showed a method of decomposing a desired modulated signal assuming three and four transmitting antennas. However, in some cases, the modulated signal cannot be successfully reconstructed. In SSM, a modulated signal is formed by the sum of phase-controlled channel coefficients. Therefore, the desired modulated signal may not be reconstructed depending on the fading characteristics. We call the unsuccessful probability as untransmissible probability. One of the challenges about the technique is to improve the probability.

We consider increasing the number of transmitting antennas to solve the problem. In this paper, we focus on the performance evaluation particularly when multilevel modulation method is adopted because increasing the number of antennas is especially effective for higher multilevel modulation. We quantitatively evaluate the improvement of the untransmissible probability and the secrecy capacity. The results show the improvement of the untransmissible probability and the secrecy capacity.

2 Principle of SSM

2.1 Signal decomposition method of SSM

In this section, we show the signal decomposition method of SSM shown in [5]. The configuration of the transmission system is shown in Fig. 1(a). The scheme consists of $N$ antennas at the transmitter and a single antenna at the receiver. The scheme decomposes a desired modulated signal into $N$ phase modulated signals so
that the desired signal is properly reconstructed by summing up the received signals at the legitimate receiver as shown in Eq. (1).

\[ s(t) = \sum_{i=1}^{N} h_i e^{j\theta_i(t)} \]  

\( s(t) \) represents the desired modulated signal. \( h_i \) is the channel coefficient between the \( i \)-th transmitting antenna and the receiving antenna. \( e^{j\theta_i(t)} \) signifies phase modulated signal where \( \theta_i(t) \) is the phase shift of the modulated signal. Eq. (1) indicates the desired modulated signal \( s(t) \) is formed at the legitimate receiver as a result of summing up the phase modulated signals that had been affected by the change of the amplitude and phase of the channel coefficient \( h_i \). In the DM scheme, a plane wave is assumed so the phases of the received signals transmitted from the different transmitting antennas are simply rotated without amplitude variation. On the other hand, the amplitude also fluctuates in non-line-of-sight fading environments for which SSM is expected to work. Therefore, it is necessary to consider the change of the amplitude and phase by the channel coefficients. The following briefly shows the signal decomposing method of SSM studied in [5]. We show the method for a case with four transmitting antennas \((N = 4)\). The channel coefficients are assumed to be ordered from the largest to the smallest. We assume the channel coefficients are obtained by some means before decomposing the signal.

\[ |h_1| \geq |h_2| \geq |h_3| \geq |h_4| \]  

The absolute value of the channel coefficient is also expressed as \( l_i \) \((i = 1, 2, 3, 4)\).

Fig. 1(b) and (c) are diagrams on the complex plane. \( s \) represents a desired modulated signal. A complex vector from the origin to \( s \) is decomposed into four complex vectors \( r_i \) \((i = 1, 2, 3, 4)\) by the signal decomposition method described below. Note that the length of \( r_i \) is \( l_i \). From the region surrounded by the bold line shown in Fig. 1(b), a point is randomly selected and defined as a point \( c \). The area surrounded by the bold line is the possible range of the point \( c \). Next, the vector from the origin to the point \( c \) is represented by combination of two vectors \( r_1 \) and \( r_2 \), and the vector from the point \( c \) to the modulated signal vector \( s \) is by two vectors \( r_3 \) and \( r_4 \) as in Fig. 1(c). The vectors \( r_i \), their arguments \( \phi_i \) and the angles \( \alpha \) and \( \beta \) are calculated by the following equation.

\[ \alpha = \cos^{-1}\left(\frac{|c|^2 + l_1^2 - l_2^2}{2l_1|c|}\right) \]  

\[ \phi_1 = \arg(c) \pm \alpha, \quad \phi_2 = \arg(c - r_1) \]  

\[ \beta = \cos^{-1}\left(\frac{|s - c|^2 + l_3^2 - l_4^2}{2l_3|s - c|}\right) \]  

\[ \phi_3 = \arg(s - c) \pm \beta, \quad \phi_4 = \arg(s - c - r_3) \]  

\[ r_i = l_i e^{j\phi_i} \]

The phase \( \theta_i \) of the transmitted signal from the \( i \)-th antenna is calculated by subtracting the phase of the channel coefficient from \( \phi_i \).

### 2.2 Required condition for signal decomposition in SSM

In SSM, the signal decomposition shown in Eq. (1) may not be feasible depending on channel conditions. This section describes the required condition to decompose
a desired modulated signal in SSM. The following shows the conditions. Eqs. (8) and (9) must be satisfied in order to decompose both of the maximum amplitude of the modulated signal $|s_{\text{max}}|$ and the minimum amplitude of the modulated signal $|s_{\text{min}}|$. Note that $s_{\text{max}}$ and $s_{\text{min}}$ are complex signals of the constellation of a modulation system on the complex plane.

$$\sum_{i=1}^{N} |h_i| \geq |s_{\text{max}}|$$

(8)

$$|h_i| - \sum_{i=1}^{N} |h_i| \leq |s_{\text{min}}|$$

(9)

The left terms of Eqs. (8) and (9) indicate the possible values as the minimum and the maximum, respectively, as the reconstructed signals after summing up the all received signals at the receiver. Therefore, if the amplitude of the desired modulated signal does not satisfy Eqs. (8) and (9), the signal could not be formed. If $|s_{\text{max}}|$ was set smaller than $\sum_{i=1}^{N} |h_i|$, the combination to reconstruct desired signal would not be unique. In other words, $\theta_i(t)$ can be variable. By changing the phase $\theta_i(t)$ while satisfying the above all conditions from Eqs. (1) to (9) for reconstructing the desired signal at the legitimate receiver, different signals are received at the eavesdropper due to different fading characteristics. As a result, demodulation at the eavesdropper becomes difficult. As described above, by intentionally suppressing the signal intensity from the maximum possible level, the degree of freedom of the signal decomposition can be improved, and it can be used for a countermeasure against eavesdropping. The parameter indicating the amount of the suppression is defined by the following equation.

$$A_t = 20 \log_{10} \left( \frac{\sum_{i=1}^{N} |h_i|}{s_{\text{max}}} \right)$$

(10)

We define $A_t$ to be greater than or equal to 0. So Eq. (8) is always satisfied. However, if $|s_{\text{max}}|$ were equal to $\sum_{i=1}^{N} |h_i|$, the phases of all $h_i e^{i\theta_i(t)} (i = 1, 2, \cdots, N)$ would be the same. In other words, $\theta_i(t)$ are uniquely determined. Therefore, the corresponding received signal at the eavesdropper is also uniquely determined. On the other hand, increasing $A_t$ extends the degree of freedom of $\theta_i(t)$. However, as $A_t$ increases, the amplitude of the modulated signal decreases and SNR (Signal to Noise power Ratio) also decreases. It brings about the deterioration of the BER performance at the legitimate receiver. Therefore, $A_t$ must be set appropriately because the sharing characteristic between legitimate users and eavesdropping resistance have such trade-off relationship.

$|s_{\text{max}}|$ is determined by the sum of the amplitude of all channel coefficients (left side of Eq. (8)) and the value of $A_t$. $|s_{\text{min}}|$ depends on $|s_{\text{max}}|$ and the modulation level. And when the modulation level is higher, $|s_{\text{min}}|$ becomes smaller assuming constant $|s_{\text{max}}|$ and, as the result, the probability of satisfying Eq. (9) decreases. As the countermeasure to the problem, we consider increasing the number of transmitting antennas. The untransmissible probability is improved by increasing the number of transmitting antennas. And the improvement is particularly clear for higher multilevel modulation. For this reason, in this paper we focus
on the performance evaluation of SSM when high multilevel modulation is adopted such as 256QAM and 1024QAM.

3 Extension of signal decomposition in SSM

In SSM, the untransmissible probability is expected to improve as the number of transmitting antennas increases. To realize the signal decomposition of SSM using more number of antennas, it is necessary to extend the conventional decomposition scheme. We describe the extended method in this section. The configuration of the method is fundamentally the same as the conventional method shown in Fig. 1(a). As in the existing SSM, we decompose a desired modulated signal into $N$ phase modulated signals based on the channel coefficient and transmit them.

Here we show a case with eight transmitting antennas as an example to show how to determine the transmission phase. In the method, a waypoint is set between the origin and the desired modulated signal. The waypoint is equivalent to point $c$ in Fig. 1(b). We select the top four antennas with the higher absolute value of the channel coefficient from the eight transmitting antennas. The vector from the origin to the waypoint is formed by summing up the received signals from the four transmitting antennas. And all other transmitting antennas form the vector from the waypoint to desired signal. The method shown in Section 2.1 is applied to determine the phase of the transmitted signal. The case of more transmitting antennas than eight can be realized by the same way.

4 Performance evaluation by computer simulation

4.1 Simulation system

In this section, we evaluate the performance of SSM using many transmitting antennas by computer simulation. Table I shows the specifications of the simulation system. The configuration of the legitimate users is the same as that of Fig. 1(a). The eavesdropper has the same configuration of the legitimate receiver. The channel coefficient between the legitimate users is $h_l$. Similarly, the channel coefficient between the $i$-th transmitting antennas and the eavesdropper is expressed as $h_{ei}$. The correlation of the fading variation between $h_l$ and $h_{ei}$ is assumed $\rho$. In
this paper, we assume a block Rayleigh fading channel where the amplitude and phase fluctuations are constant during a single block. And we assume 100 symbols are transmitted per block. As described in the above section, high-level modulation method (256QAM and 1024QAM) is assumed in this paper. Selection of the value of $A_t$ is important to realize the best performance of SSM. However, we emphasize the improvement by increasing antennas in this paper. So $A_t$ is simply fixed 1 dB.

| Table 1. Specifications of simulation |
|--------------------------------------|
| Modulation scheme | 256, 1024 QAM |
| Number of antennas | Transmission: 4, 8 |
| | Reception: 1 |
| Propagation path | Block Rayleigh fading channel |
| | Correlation coefficient $\rho$: 0.2, 0.6, 0.95 |

4.2 Simulation result
In this section, we evaluate the untransmissible probability in SSM and the secrecy capacity. The secrecy capacity is calculated from BER of the legitimate receiver and the eavesdropper based on the conditional entropies $H(X|Y)$ and $H(X|Z)$. Note that the information transmitted by the legitimate transmitter, the information received by the legitimate receiver and the eavesdropper are expressed by $X$, $Y$ and $Z$, respectively. The secrecy capacity is found from the difference between $H(X|Y)$ and $H(X|Z)$.

Fig. 2(a) shows the untransmissible probability in SSM. In the case of original four transmitting antennas, the untransmissible probability is high for both 256QAM and 1024QAM. However, when the number of transmitting antennas is increased to 8, even if 1024QAM is used, the probability is greatly reduced.

Fig. 2(b) shows the secrecy capacity for 256QAM. It is confirmed by the figure, as the number of the transmitting antennas increases, the secrecy capacity improves.

5 Conclusion
In this paper, we examined the performance of SSM using many transmitting antennas. Computer simulations show that the untransmissible probability and the overall secrecy capacity are improved by increasing the number of transmitting antennas.
A study on half-shaped printed monopole antenna with short stub for UWB system

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Abstract: In this paper, the authors propose a small printed monopole antenna corresponding to the frequency band of UWB (Ultra-wideband) system that use a frequency of 3.1 GHz to 10.6 GHz. This antenna has a structure in which a bell-shaped UWB monopole antenna formed on a dielectric substrate (FR-4) has a half shape and a short stub is attached to the antenna. The size of the antenna is $30 \times 12 \times 1.6 \text{ mm}^3$. The antenna is designed to be smaller than the conventional one proposed by the author, and can operate in the frequency band of the UWB system. From the simulation result, it is shown that good characteristics can be obtained to UWB application for high speed wireless communication.

Keywords: UWB, small antenna, short stub, impedance matching

Classification: Antennas and Propagation

References

[1] M. Z. Win, D. Dardari, A. F. Molisch, W. Wiesbeck, and J. Zhang, “History and applications of UWB,” \textit{Proc. IEEE}, vol. 97, no. 2, pp. 198–204, Feb. 2009. DOI:10.1109/JPROC.2008.2008762
[2] FCC, “1st Report and Order on Ultra-Wideband Technology,” Feb. 2002.
[3] R. Pyndiah, “An overview of UWB technology,” IET Workshop on Practical Applications for Wireless Technology, pp. 65–80, 2006.
[4] A. M. Abbosh, “Miniaturization of planar ultrawideband antenna via corrugation,” \textit{IEEE Antennas Wireless Propag. Lett.}, vol. 7, pp. 685–688, Nov. 2008. DOI:10.1109/LAWP.2008.2009323
[5] R. Azim, M. T. Islam, and N. Misran, “Compact tapered-shape slot antenna for UWB applications,” \textit{IEEE Antennas Wireless Propag. Lett.}, vol. 10, pp. 1190–1193, Oct. 2011. DOI:10.1109/LAWP.2011.2172181
[6] N. Takemura and S. Ichikawa, “Experimental study of bell-shaped monopole antenna with short stub for UWB applications,” International Symposium on Antennas and Propagation, Dec. 2017.
[7] I. Pele, A. Chousseaud, and S. Toutain, “Simultaneous modeling of impedance and radiation pattern antenna for UWB pulse modulation,” IEEE AP-S
1 Introduction

The Ultra Wideband (UWB) system is widely studied as a system that enables high-speed wireless communication by performing communication using a wide frequency band [1]. In 2002, The Federal Communication Commission (FCC) has approved the spectrum within the frequency 3.1 GHz to 10.6 GHz for ultra-wideband systems, and the design of an antenna covering those frequency band is required [2]. The antenna is required to have a small size, and it is desirable that the antenna have non-directional radiation characteristics assuming communication with devices in various directions [3]. Impedance matching technology used to improve antenna characteristics has been reported as a miniaturization technology for ultra-wideband antennas [4, 5]. A bell-shaped planar monopole antenna attached with a short stub has been proposed [6]. This antenna improves the impedance matching in the lower band by attaching a short stub and is an antenna with a wide band characteristic that satisfies the band of the ultra-wideband system. The size is $28 \times 20 \times 1.6 \text{mm}^3$. In this paper, we focus on the symmetrical structure of the previously proposed ultra-wideband antenna, and propose a half-shaped UWB monopole antenna with the antenna cut in half. By adjusting the parameters of the short stub and the ground conductor, the characteristics satisfying the frequency band of the ultra-wideband system are shown. In order to achieve impedance matching, the antenna is shaped by cutting off the microstrip line, and short stubs is useful for impedance matching in the low frequency band of the ultra-wideband system.

2 Antenna design

Fig. 1 shows the half-shaped printed monopole antenna with short stub for the UWB system. This antenna is formed on a dielectric substrate (FR-4: $\varepsilon_r = 4.4$, $\tan \delta = 0.02$) with a length $L = 30 \text{ mm}$, a width $W = 13 \text{ mm}$, and a thickness $t = 1.6 \text{ mm}$. The antenna is fed by microstrip transmission line to the monopole on a dielectric substrate, and the short stub is connected to the ground plane of the microstrip transmission line from the central lower part of the monopole through holes. The monopole and the ground conductor are half-shaped compared to the previously proposed antenna. The characteristic impedance is $50 \Omega$. In general, the equivalent circuit of the UWB monopole antenna can be considered as a series of multiple resonant circuits [7]. The short stub operates as a parallel impedance to the UWB monopole antenna, and the antenna’s input impedance can be matched by properly adjusting the short stub’s parallel impedance. This antenna structure provides broadband operation. The dimensions of the proposed antenna are length ($L_L = 10.0 \text{ mm}$), width of microstrip line ($W_L = 2.0 \text{ mm}$), width of antenna ($W_A = 12.0 \text{ mm}$), length of ground ($L_G = 10.5 \text{ mm}$), width of ground ($W_G = 12.0 \text{ mm}$), length of short stub
(L_s) = 2.5 mm, width of short stub (W_s) = 0.5 mm, and diameter of through hole (D_H) = 0.5 mm.

3 Simulation results

The antenna is analyzed using the electromagnetic field simulator Keysight EMPro. The antenna parameters are calculated using the same values as the reference values shown in chapter 1. Fig. 2 shows the calculation results of the characteristics of the antenna. In this figure, (a) shows the smith chart when the length of short stub is changed from 1.5 mm to 3.0 mm at 0.5 mm intervals, (b) shows the result when the width of short stub is changed from 0.3 mm to 0.9 mm at 0.2 mm intervals. And, (c) and (d) show comparison of s-parameters with short stub and without short stub in VSWR and smith chart. Calculation results with short stubs are indicated by blue solid lines, and calculation results without short stubs are indicated by red dotted lines. From Fig. 2(a), it is obtained that when the length of the short stub is increased, the capacitance component of the input impedance tends to increase. From Fig. 2(b), it is obtained that when the width of the short stub is increased, the inductance component of the input impedance tends to increased. The input impedance can be adjusted by changing the length of short stub (L_s) and the width.

Fig. 1. Antenna configuration
of short stub (Wₛ). It can be solved the problem that when the antenna is designed small, the frequency band shifts to a high frequency and impedance matching in the low frequency is degraded. The proposed antenna’s short stub is adjusted to
L_S = 2.5 mm and W_S = 0.5 mm for impedance matching. From the calculation results of VSWR, the proposed antenna showed good results that is the VSWR is 2 or less over the entire desired band of the ultra-wideband system. This result shows that even in the case of the half shape, almost the same characteristics as the conventional antenna could be obtained. The simulation result of radiation pattern is shown in Fig. 3. The radiation pattern in the x-y plane is shown. The calculation frequencies are 3.5 GHz, 5.0 GHz, 6.5 GHz, 8.0 GHz, 9.0 GHz and 10.0 GHz. From the calculation results, radiation in the cut direction at high frequencies is slightly degraded, however it can be confirmed that almost non-directional radiation patterns are obtained at each frequency. It can be confirmed that good results were obtained as an antenna for UWB high speed wireless communication applications.

4 Conclusion

A half-shaped printed monopole antenna corresponding to the desired frequency band of the UWB system is proposed, and the calculation results of the antenna characteristics are presented. By adjusting the length and width of the short stub, the input impedance of the antenna can be obtained a good value throughout the UWB system, and even if the antenna is designed to be small, a wideband characteristic that satisfies the desired frequency band of the UWB system is realized. It was also confirmed that nearly all of the omnidirectional radiation characteristics were obtained. The proposed antenna achieves miniturization with an antenna configuration that is 30% size reduction of the conventional antenna, and achieves wideband characteristics that operate in the UWB frequency band.

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Near-field leaky-wave focusing antenna with inhomogeneous rectangular waveguide

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Abstract: The near field leaky-wave focusing antenna using the inhomogeneous rectangular waveguide is proposed and a design method is presented. The height of broad-wall of waveguide is inhomogeneous to obtain focusing effect at a desired position. In the antenna design, a relation between the phase constant of traveling wave and the height of broad-wall of the waveguide is derived. Simulation results and measured results at Ka-band are presented to validate the proposed design method.

Keywords: focusing antennas, near-field antennas, leaky-wave antennas, waveguide antennas

Classification: Antennas and Propagation

References

[1] J. T. Loane and S.-W. Lee, “Gain optimization of a near-field focusing array for hyperthermia applications,” IEEE Trans. Microw. Theory Techn., vol. 37, pp. 1629–1635, Oct. 1989. DOI:10.1109/22.41011
[2] J. O. McSpadden and J. C. Mankins, “Space solar power programs and microwave wireless power transmission technology,” IEEE Microw. Mag., vol. 3, no. 4, pp. 46–57, Dec. 2002. DOI:10.1109/MMW.2002.1145675
[3] H.-T. Chou, T.-M. Hung, N.-N. Wang, H.-H. Chou, C. Tung, and P. Nepa, “Design of a near-field focused reflectarray antenna for 2.4-GHz RFID reader applications,” IEEE Trans. Antennas Propag., vol. 59, no. 3, pp. 1013–1018, Mar. 2011. DOI:10.1109/TAP.2010.2103030
[4] T. Okuyama, Y. Monnai, and H. Shinoda, “20-GHz focusing antennas based on corrugated waveguide scattering,” IEEE Antennas Wireless Propag. Lett., vol. 12, pp. 1284–1286, Oct. 2013. DOI:10.1109/LAWP.2013.2284278
[5] Y. F. Wu and Y. J. Cheng, “Proactive conformal antenna array for near-field beam focusing and steering based on curved substrate integrated waveguide,” IEEE Trans. Antennas Propag., vol. 67, no. 4, pp. 2354–2363, Apr. 2019. DOI:10.1109/TAP.2019.2891725
[6] Y. F. Wu and Y. J. Cheng, “Two-dimensional near-field focusing folded reversely fed leaky-wave antenna array with high radiation efficiency,” IEEE Trans. Antennas Propag., vol. 67, no. 7, pp. 4560–4569, July 2019. DOI:10.1109/TAP.2019.2906019
1 Introduction

The imaging technology that can detect dangerous items in clothes using non-contact/non-invasive manner is needed at airports and seaports. One of the main reasons is that the many dangerous items are miniaturized, easy to be concealed in clothes. In order to detect or to obtain images of dangerous objects in clothes, it is necessary to focus electromagnetic wave on a specific position of the surface of a human body. Recently, the technologies for microwave-focusing in the near field are drawing attention in various applications such as the imaging, the thermotherapy [1], WPT (wireless power transfer) [2], RFID reader [3]. In most cases, the handy type imaging device with compact structure is desired. In the case of the focal plane imaging, lens antennas or reflector antennas are used to obtain high resolution of images. However, these antennas are not suitable for compact devices because weight and size of the lens and the reflector are usually a big problem, especially in the application of handy type imaging device. In this paper, a waveguide leaky-wave antenna (LWA) is designed to address this problem.

LWA is one of traveling wave antennas having characteristics that the radiation direction is changed as the frequency changes which are useful to change the focusing position without using a dielectric lens or a metal reflector. In the previous studies, the focusing LWAs have been proposed in [4, 5, 6, 7]. In [4], the structure of the radiating grating elements is designed to obtain focusing effect. In [5], the desired phase difference for the focusing is obtained by changing the distance between the focal point and the slot elements located on the curved substrate integrated waveguide. In [6], the location of radiating slot elements are designed to obtain the desired phase distribution. This LWA is effective for handy type devices from the view point of its low weight. The rectilinear LWA in [7] adjusts the slit width and the position of slit on the dielectric to obtain the desired phase distribution, however, the experimental study was not provided.

In this paper, the phase constant distribution of leaky waves is controlled by changing the height of broad-wall of a rectangular waveguide inhomogeneously, and with homogeneous structure of the radiating slot elements. The structure is simple and easy to fabricated. The design concept and principle of focusing effect using a waveguide LWA are shown in section II. Our research was first reported in [8] which showed the focusing effect by simulations. In section III, the design
method to obtain focusing effect at desired point in the near field is presented. In section IV, experimental results of our proposed antenna are shown and the conclusion of this paper is shown in section V.

2 Concept

A waveguide LWA is a kind of a traveling wave antenna. Eq. (1) shows the relationship between phase constant \( \beta \) of the traveling wave in the waveguide and a radiation direction \( \theta_r \) is

\[
\theta_r \ [\text{rad.}] = \cos^{-1} \frac{\beta}{k_0}
\]

where \( k_0 \) is the wave number in vacuum. The radiation direction depends on the phase constant. Also, the phase constant depends on the frequency. Generally the phase constant and a wave number are constant for a fixed frequency, the radiation direction is constant and leaked waves become a plane wave. In order to perform imaging using a waveguide LWA, leakage radiations are required to be focused in the near field at a fixed frequency. One method to focus leaked waves in the near field is to change the phase constant of the traveling wave along the waveguide with changing the height of broad-wall of waveguide, gradually. Detail of the structure are discussed in the next section. Fig. 1 shows a principle of focusing effect. A traveling wave is excited from the end point of the waveguide LWA. In order to focus at a desired point S in the near field, the distribution of phase constant \( \beta(z) \) of the traveling wave should be the desired distribution.

3 Design method

In this section, the design method to obtain the focusing effect at a desired point in the near field is discussed. As shown in Fig. 1, the desired focus point and the desired phase constant distribution are indicated as \( S(z_s, x_s) \) and \( \beta(z) \). Providing a distance \( OS \) from origin \( O (z = 0) \) to focal point \( S \) and a distance \( AS \) from a certain point \( A (z = z_A) \) on the surface of LWA \((-0.5L \leq z \leq 0.5L)\), the distance \( AP \) is given by
The phase difference corresponding to distance AP is defined as

\[ \Delta \text{Phase}(z_A) = k_0 \text{AP}. \]  

Furthermore, the phase difference and the phase constant \( \beta(z) \) can be expressed in an integral form as

\[ \Delta \text{Phase}(z_A) = \int_0^{z_A} \beta(z) \, dz. \]  

Therefore, the phase constant distribution in the range of \((-0.5L \leq z \leq 0.5L)\) is expressed by the following equation as

\[ \beta(z) = \frac{d}{dz} \Delta \text{Phase}(z). \]  

Fig. 2(a) and Fig. 2(b) shows the simulation model of the waveguide LWA, and each parameters of the proposed antenna are \( a = 8.5 \, \text{mm}, \, b = 4 \, \text{mm}, \, g = 1 \, \text{mm}, \, l = 4 \, \text{mm}, \, s = 1 \, \text{mm}, \, p = 3 \, \text{mm} \) and \( L = 200 \, \text{mm} \). \( h(z) \) is design parameter. The design frequency is set as 27 GHz. The antenna has slot arraying on the narrow-wall surface of a rectangular waveguide. The structure of LWA part is based on the waveguide slot array proposed in [9]. A monopole located at Port 1 excites TE\(_{10}\) mode with cutoff frequency of \( f_c = c / 2a = 17.7 \, \text{GHz} \), then the design frequency is in the propagation mode. In order to obtain the relation between the radiation direction and the height of the broad-wall \( h \), the radiation patterns were evaluated using FDTD analysis. Fig. 2(c) shows the simulated radiation direction indicated as circle marks, when height of the broad-wall \( h \) is changed. The radiation directions

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**Fig. 2.** Simulation model and design procedure of focusing LWA.
decrease by increasing the height of broad-wall $h$. In Fig. 2(c), a fitted curve using logarithmic function is obtained, and the relation between the height of broad-wall $h$ and the radiation direction is derived as

$$h \ [\text{mm}] = e^{2.616-0.6455z} \ \text{[rad]}. \quad (6)$$

Substituting Eq. (1) into Eq. (6), the expression of the relation between the phase constant $\beta(z)$ and the height of broad-wall $h$ can be derived as

$$h \ [\text{mm}] = e^{2.616-0.645 \cos^{-1} \frac{z}{150}}. \quad (7)$$

Using Eq. (7), it is possible to design the inhomogeneous height distribution $h(z)$ from the desired phase constant distribution obtained by Eq. (5). Fig. 2(d) shows the desired phase constant distribution, and the designed height of broad-wall distribution $h(z)$ by using Eq. (7) when the focusing point is selected as $S(150 \text{ mm}, 195 \text{ mm})$ at 27 GHz. Furthermore, the simulated phase constant distribution is the result of designed height of broad-wall distribution $h(z)$. The desired phase constant distribution gradually decreases as $z$ increases. The designed height distribution of broad-wall also decreases as $z$ increases, gradually. Also, the simulation result of phase constant distribution roughly agrees with desired values. In this paper, the height $h$ is changed along the waveguide to obtain desired phase constant distribution, however, the cutoff frequency also changes. In this design, the frequency 27 GHz is not in the cutoff frequency region in the designed height range of 5.5 mm to 8.5 mm.

4 Experiment

This section describes the measured results of designed focusing antenna. Fig. 3(a) shows the birds view of fabricated leaky-wave focusing antenna. LWA composed of an aluminum plate with a large number of slots was provided on the narrow-wall of the waveguide. Fig. 3(b) shows the inside view of fabricated antenna. The height of the broad-wall was set to the distribution $h(z)$ in Fig. 2(d). The travelling wave was excited at Port 1. Port 2 was terminated by 50 $\Omega$. The near-field distribution was measured. An open ended waveguide (receiving antenna) was moved in $xz$-plane. The measurement area was $0 \text{ mm} \leq z \leq 300 \text{ mm}, \ 10 \text{ mm} \leq x \leq 300 \text{ mm}$, with at 5 mm intervals. Fig. 3(c) and Fig. 3(e) shows measurement result at 27 GHz and 29 GHz. In Fig. 3(c), the focusing effect at roughly the design point (the dashed lines intersection, $S(150 \text{ mm}, 195 \text{ mm})$) was observed. In Fig. 3(e), the focusing effect was also observed, and the focused area moved about 30 mm in $+z$ direction. Fig. 3(d) and Fig. 3(f) shows the $|E_x|$ transverse line distribution ($x = 195 \text{ mm}$). In Fig. 3(d), the maximum value was obtained at the desired position ($z = 150 \text{ mm}$), and it was found that there is a focal region. Both of in Fig. 3(d) and in Fig. 3(f), good agreement were observed in the depth and the width of focus and the sidelobe levels with the simulation results.
5 Conclusion

The near field leaky-wave focusing antenna using the inhomogeneous rectangular waveguide and its design method was proposed. Inhomogeneous distributions of the phase constant of traveling wave and the height of broad-wall of waveguide were provided to obtain focusing effect at a desired position. The fabrication of the leaky-wave focusing antenna was performed at 27 GHz. The focal region of simulation and measurement were roughly agreed with the desired position and the focal area moved with changing frequency.

Fig. 3. Measure electric field $|E_y|$ distribution at 27 GHz and 29 GHz.
Compressed sensing based low complexity 2D-DOA estimation by separation and pair-matching approach

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Abstract: Compressed sensing (CS) based direction-of-arrival (DOA) estimation has the advantage of high resolution and no need for wave number estimation. Although it is applicable for 2 dimensional estimation, the computation complexity is severe problem due to its increased number of array elements and search domain. This letter proposes a separation approach and pair-matching to resolve the above drawback. Since the CS can extract the original signal source, the pair-matching can be simply attained by cross correlation between source estimates of vertical and horizontal arrays. Computer simulation verifies the fundamental effectiveness of the proposed method.

Keywords: DOA estimation, compressed sensing, pair-matching

Classification: Antennas and Propagation

References

[1] R. Schmidt, “Multiple emitter location and signal parameter estimation,” IEEE Trans. Antennas Propag., vol. 34, no. 3, pp. 276–280, Mar. 1986. DOI:10.1109/TAP.1986.1143830

[2] M. Çetin, D. M. Malioutov, and A. S. Willsky, “A variational technique for source localization based on a sparse signal reconstruction perspective,” Proc. 2002 IEEE International Conference on Acoustics, Speech, and Signal Processing (ICASSP), pp. III-2965–III-2968, May 2002. DOI:10.1109/ICASSP.2002.5745271

[3] T. Terada, T. Nishimura, Y. Ogawa, T. Ohgane, and H. Yamada, “DOA estimation for multi-band signal sources using compressed sensing techniques with Khatri-Rao processing,” IEEE Trans. Commun., vol. E97-B, no. 10, pp. 2110–2117, Oct. 2014. DOI:10.1587/transcom.E97.B.2110

[4] T. Hoshikawa, T. Nishimura, T. Ohgane, Y. Ogawa, and J. Hagiwara, “Performance comparison of compressed sensing algorithms for DOA estimation of multi-band signals,” 2018 IEEE Workshop on Positioning, Navigation and Communications (WPNC), Oct. 2018. DOI:10.1109/WPNC.2018.8555749

[5] D. Geman and G. Reynolds, “Constrained restoration and the recovery of discontinuities,” IEEE Trans. Pattern Anal. Mach. Intell., vol. 14, pp. 367–383, Mar. 1992. DOI:10.1109/34.120331
1 Introduction

A number of researches on direction of arrival (DOA) estimation of radio waves are progressing in many fields, such as radar systems and positioning systems for wireless communication terminals, and applications to mobile communication. Improving estimation accuracy is essential for the rapid spread of wireless communication. Many techniques on DOA estimation have been investigated such as beamformer method and MUItiple SIgnal Classification (MUSIC) method [1]. These methods are frequently used nowadays. However, these methods require information on the number of incoming waves as precursor information.

In recent years, compressed sensing (CS) [2] is applied as a more accurate estimation method. CS can reconstruct the original source from a few observations in which most of components to be estimated are zero, i.e. called sparse. In the CS based DOA estimation, the space around the receiving antenna array is divided into small angular bins. DOAs and complex signals of incident waves can be estimated at the corresponding bins. Its notable feature is that one snapshot is sufficient. When the arrival waves are multi-band signals, the estimation accuracy can be improved more than the case of the single-band signals [3]. Further, it is known to possible to estimate the number of incident waves exceeding the physical degree of freedom of the antenna array. Several algorithms have been developed to obtain the solution of the CS. We focused on half-quadratic regularization (HQR) algorithm which is known to exhibit superior estimation accuracy [4]. However, there is a problem that the amount of calculation becomes enormous due to the increase of search domain, especially in the case of 2D-DOA estimation. Exploiting the nature of CS, this letter proposes a 2D-DOA estimation by separation and pair-matching approaches. The key idea is to calculate a cross correlation between complex-valued source estimates of vertical and horizontal arrays. Simulation result shows that the proposed approach can effectively work under the high SNR environment with reasonable computation complexity.

2 System description

Fig. 1 shows an L-shaped array. Receiver can be equipped with a planar array wherein two ULAs arranged with M and N elements in a row and column are utilized for 2D-DOA estimation. Each inter-element spacing is d. Suppose the antenna aperture faces the positive direction of the Y-axis perpendicular to the X–Z plane. K narrowband plane signals of sources impinging on the array yield distinct directions at elevation and azimuth angles \{\theta_k\}_{k=1}^K and \{\phi_k\}_{k=1}^K, respectively. The baseband array input at the t-th snapshot along the Z-axis is expressed as

\[ x_z(t) = \sum_{k=1}^{K} a(\theta_k)s_k(t) + n_z(t), \quad (1) \]

where \( a(\theta_k) \) is so-called steering vector represented as \( [a_1(\theta_k), \ldots, a_M(\theta_k)]^T \) with

\[ a_m(\theta_k) = \exp \left\{ -j \frac{2\pi}{\lambda} d(m - 1) \cos \theta_k \right\}. \quad (2) \]
\( \lambda \) denotes the wavelength and \( n_z(t) \sim \mathcal{CN}(0, \sigma_{n_z}^2) \) is an additive white Gaussian noise (AWGN) vector. On the other hand, the array input along the \( X \)-axis is also given by,

\[
x_x(t) = \sum_{k=1}^{K} a(\theta_k) s_k(t) + n_x(t),
\]

where \( a(\phi_k) = [a_1(\phi_k), \ldots, a_M(\phi_k)]^T \) and its each element is

\[
a_n(\phi_k) = \exp\left(-j \frac{2\pi}{\lambda} d(n-1) \cos \phi_k \right).
\]

\( n_x(t) \sim \mathcal{CN}(0, \sigma_{nx}^2) \) is also an AWGN vector. The elevation and azimuth angles can be estimated in a separation manner, however, these pair-matching is required.

### 3 Proposed method

First, we set initial problem to be solved. Let \( L, \hat{s} \in \mathbb{C}^{L \times 1} \) and \( A \in \mathbb{C}^{M(N) \times L} \) denote the number of bins, sparse vector containing source information and mode matrix, array input can be rewritten as

\[
x(t) = A\hat{s}(t) + n(t).
\]

Each element of \( A \) is given by \( a_m(\theta) \) or \( a_n(\phi) \). Here, HQR method is applied. It transforms a nonquadratic optimization problem into a series of quadratic problems. See [5] for its detailed derivation. Resulting iterative algorithm is given by

\[
H(\hat{s}^{(n)}) = E[A^H x(t)],
\]

where \( n \) is the iteration number and \( E[\cdot] \) denotes the ensemble averaging. \( H(\hat{s}^{(n)}) \) is then expressed as

\[
H(\hat{s}^{(n)}) \triangleq A^H A + \alpha \Lambda(\hat{s}^{(n)}),
\]

\[
\Lambda(\hat{s}^{(n)}) \triangleq \text{diag}\left(\frac{q/2}{(|\hat{s}^{(n)}|^2 + e)^{1-q/2}}\right),
\]

where \( \text{diag}(\cdot) \) is to compose a diagonal matrix \( \Lambda \in \mathbb{C}^{L \times L} \). Calculation of (5) is iteratively performed until
\[
\frac{\|\mathbf{s}^{(n+1)} - \mathbf{s}^{(n)}\|^2_2}{\|\mathbf{s}^{(n)}\|^2_2} < \delta,
\]  

(9)

is satisfied. \(\delta > 0\) is a small constant.

Original signal sources \(\hat{s}_{k,\theta}(t)\) and \(\hat{s}_{k,\phi}(t)\) can be obtained by applying HQR respective to array inputs \(x_z(t)\) and \(x_x(t)\). Obtained signals can be found in corresponding bins; DOAs \(\hat{\theta}\) and \(\hat{\phi}\) can be uniquely determined. Under these conditions, the question of interest is how to make pairs of the corresponding elevation and azimuth angles, i.e., \(\{\theta_1, \phi_1\}, \ldots, \{\theta_K, \phi_K\}\), in a computationally efficient manner.

To obtain the relation of the corresponding elevation and azimuth angles, we take the cross correlation matrix \(R_{\theta\phi}\) between \(\hat{s}_{k,\theta}(t)\) and \(\hat{s}_{k,\phi}(t)\), which are the estimation results of HQR.

\[
R_{\theta\phi} = E[\hat{s}_{\theta}(t)\hat{s}_{\phi}^T(t)].
\]  

(10)

Observing peak values of \(R_{\theta\phi}\), these row and column indices are pairs of elevation and azimuth angles for corresponding signal sources.

Since the computation for matrix inversion of \(H\) is the most dominant in HQR, its complexity is compared. Suppose the complexity order of Gaussian elimination based matrix inversion, straightforward 2D-DOA estimation requires \(O((L_\theta \times L_\phi)^3)\) whereas the proposed method can reduce it to \(O((L_\theta + L_\phi)^3)\). These operations must be repeated until the convergence condition is satisfied. Computation complexity becomes huge and thus it can be remarkably reduced by our proposed approach.

4 Computer simulation

4.1 Simulation parameters

This section examines the performance of the proposed 2D-DOA estimation. Evaluation metric is the probability of correct estimation. When the estimated DOA value is within the allowable range, it is determined as successful. Table I lists the simulation parameters. DOAs, \(\theta\) and \(\phi\), are uniformly distributed from 0° to 180°, respectively. The search domain for both for elevation and azimuth angles are also set to from 0° to 180° with 1° resolution; the space vectors have 181 bins. The probability of correct estimation is evaluated in terms of signal to noise power ratio (SNR), the number of antenna elements \((M, N)\), the number of snap shots \(P\), and modulation order. HQR parameters \((\epsilon, q, a)\) were empirically determined prior to the evaluation. All results were averaged via 10000 independent trials.

4.2 Simulation results

First, Fig. 2(a) visualizes the pair matching result of 2D-DOA. This case consider that three signal sources are arrived from \((\theta, \phi) = (160, 16), (119, 17), \) and \((42, 117)\). The number of antenna elements and snapshots are set to \(M = N = 20\) and \(P = 100\). SNR is 30 dB. The peak values in the figure indicate estimated DOAs. Pair matching is successfully accomplished by our proposed approach.

Following results show the probability of correct estimation when three incident waves are arrived at random angles. Fig. 2(b) shows the probability versus
Table 1. Simulation parameters

| Parameters                  | Values                                      |
|-----------------------------|---------------------------------------------|
| Data modulation             | BPSK, QPSK, 16QAM, 64QAM                    |
| SNR                         | 0, . . . , 30 dB                           |
| Inter-element spacing       | λ/2                                         |
| Antenna elements (M, N)     | (10, 10), (20, 20), (30, 30), (40, 40)     |
| Carrier frequency           | 28 GHz                                     |
| Number of sources K         | 3                                          |
| Number of snapshots P       | 10, . . . , 100                            |
| Search domain               | 0° < θ < 180°, 0° < φ < 180°              |
| Number of bins (Lθ, Lφ)     | (181, 181)                                 |
| HQR parameters (e, q, α)    | (1.0 × 10⁻⁶, 1.0 × 10⁻⁶, 1.0 × 10⁻³)        |
| Convergence condition δ     | 1.0 × 10⁻⁶                                 |

(a) Estimation results overview of 2D-DOA (BPSK, M = N = 20, K = 3, SNR = 30 dB).
(b) Probability of correct estimation v.s. SNR (M = N = 20, allowable range: ±1°).
(c) Probability of correct estimation v.s. number of snapshots (BPSK, M = N = 20, SNR = 30 dB).
(d) Probability of correct estimation v.s. SNR (BPSK, allowable range: ±1°).

Fig. 2. Simulation results

SNR with various modulation orders. As shown in the figure, BPSK achieves the highest probability about 77% at SNR = 30 dB when the allowable range is set to ±1°. It indicates that the pair-matching accuracy depends on the modulation order. Signal estimate by CS is also affected by the additive noise effect. The lower the
modulation order has its immunity thanks to the longer Euclidean distance between symbols.

Fig. 2(c) shows the probability versus the number of snapshots at SNR = 30 dB. Here, allowable range is also varied as ±0°, ±0.5°, and ±1°. In high SNR region, estimation accuracy remained almost unchanged even when the number of snapshots is changed [4]. It is possible to estimate DOA even at the small number of snapshots about 10. The advantage of CS-based DOA estimation is kept even in the proposed 2D-DOA separation and pair-matching approach. It should be noted that $K$ snapshots are required at least to realize pair-matching through a cross correlation calculation.

Finally, Fig. 2(d) shows the probability of correct estimation with the number of antennas elements. When sufficient snapshots are available, arbitrary number of signal sources can be estimated by the proposed method. Estimation accuracy can be improved as the number of antenna elements is increased. However, it tends to be saturated for more than 40 antenna elements per edge. Use of such large number of antenna elements also imposes impractical computation complexity in the conventional 2D-DOA estimation. The proposed method enables a considerable complexity reduction while achieving improved estimation accuracy. As a result, the proposed method is expected to be fundamentally effective approach in 2D-DOA estimation using compressed sensing.

5 Conclusion

In this letter, we proposed a practical approach that can simplify the compressed sensing based 2D-DOA estimation in separate manner and cross correlation. Key feature of the proposal is to exploit the nature of the compressed sensing that can extract original complex-valued signal source. It can realize the pair matching as well as computational complexity reduction. We can conclude that the proposed method is the most promising way for 2D-DOA estimation method towards 5G or beyond.

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Cross-tier interference mitigation considering pilot overhead for TDD MIMO heterogeneous networks

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Abstract: In a two-tier heterogeneous network, cross-tier interference mitigation between macro-cell (MC) and small-cell (SC) is an important issue, where MC and SC share the same time/frequency resources. In this article, we propose a method to mitigate downlink cross-tier interference between MC and SC, where macro-cell base-station (MBS) and small-cell base-stations have multiple antenna arrays and their precoding weights are optimized to mitigate mutual cross-tier interference under a given pilot overhead. Simulation results show that the proposed scheme improves the downlink sum-rate compared with conventional scheme that coordinates cross-tier interference from MBS to small-cell users under the same pilot overhead condition.

Keywords: two-tier heterogeneous network, massive MIMO, orthogonal pilot, pilot overhead

Classification: Wireless Communication Technologies

References

[1] F. Zhang, S. Sun, B. Rong, F. R. Yu, and K. Lu, “A novel massive MIMO precoding scheme for next generation heterogeneous networks,” IEEE GLOBECOM, San Diego, CA, pp. 1–6, 2015. DOI:10.1109/GLOCOM.2015.7417346

[2] K. Sundaresan and S. Rangarajan, “Efficient resource management in OFDMA femtocells,” Proc. ACM Intl. Symp. Mobile Ad Hoc Netw. Comput., pp. 33–42, May 2009. DOI:10.1145/1530748.1530754

[3] N. Saquib, E. Hossain, L. B. Le, and D. I. Kim, “Interference management in OFDMA femtocell networks: Issues and approaches,” IEEE Wireless Commun., vol. 19, no. 3, pp. 86–95, June 2012. DOI:10.1109/MWC.2012.6231163

[4] P. Zhao, Z. Wang, C. Qian, L. Dai, and S. Chen, “Location-aware pilot assignment for massive MIMO systems in heterogeneous networks,” IEEE Trans. Veh. Technol., vol. 65, no. 8, pp. 6815–6821, Aug. 2016. DOI:10.1109/TVT.2015.2480965

[5] W. Hao, O. Muta, H. Gacanin, and H. Furukawa, “Performance analysis on
uplink pilot allocation in TDD massive MIMO heterogeneous networks.” *IEICE Trans. Fundamentals*, vol. E100-A, no. 11, pp. 2314–2322, Nov. 2017. DOI:10.1587/transfun.E100.A.2314

[6] T. L. Marzetta, “Noncooperative cellular wireless with unlimited numbers of base station antennas,” *IEEE Trans. Wireless Commun.*, vol. 9, no. 11, pp. 3590–3600, Sept. 2010. DOI:10.1109/TWC.2010.092810.091092

1 Introduction

Two-tier heterogeneous network (HetNet) is a promising framework for next-generation wireless communication systems, where short-range small-cell (SC) base-stations are overlaid on a coverage area of macro-cell (MC) [1]. In particular, in two-tier HetNets, radio spectrum can be utilized more effectively if MC and SC share the same time/frequency resources [2]. However, cross-tier interference coordination between MC and SC is necessary to fully harvest potential performance of the HetNets [3, 4, 5].

Adaptive beam-forming using massive multi-input multi-output (mMIMO) technology [6] is an effective approach to mitigate inter-cell interference, e.g., when mMIMO is equipped at MC base station (MBS), the extra spatial degree of freedoms of mMIMO can be utilized to effectively mitigate cross-tier interference from MBS to SC users (SUs) if cross-channel state information (CrCSI) of the path between MC and SC is available. In [4], a location-aware pilot reuse scheme is proposed to coordinate uplink pilot interference. However, downlink cross-tier interference mitigation between MC and SCs is not considered. In [5], a downlink cross-tier interference mitigation scheme was proposed for HetNets with a time-division-duplex (TDD) mode, where CrCSI is estimated at MBS by overhearing uplink pilots from SUs and designing the precoding weight that cancels downlink cross-tier interference to SUs, where the pilot assignment to SUs is optimized under a given pilot overhead. Although this scheme is effective in improving SU’s rate while minimizing the required pilot overhead, cross-tier interference mitigation from SC base-stations (SBSs) to macro-users (MUs) is not considered.

In this article, based on works in [5], we propose an extended cross-tier interference mitigation scheme that coordinates not only cross-tier interference from MBS to SU but also the interference from SBS to MU in HetNets with TDD where mMIMO is equipped at MBS.

2 System model

We consider a TDD downlink HetNet in Fig. 1, where MBS has massive antenna array with $N_M$ elements and serves $M$ single-antenna MUs while each SBS has $N_S$ antennas and serves $M_S$ SUs. $K$ SBSs are randomly deployed in coverage area of the MC, where $N_M \gg N_S > M_S$ and $N_M \gg M$. MBS and SBSs estimate channel state information (CSI) by observing their users’ uplink orthogonal pilots and then calculate their downlink precoding weights.
The received signal at the $m$-th MU is expressed as

$$y_m^M = \sqrt{P_M}h_{0,0,m}^{M,M}m_{m}x_m + \sum_{i=1,i\neq m}^{M} \sqrt{P_M}h_{0,0,m}^{M,M}m_{i}x_i$$

Desired signal

$$+ \sum_{k=1}^{K} \sum_{j=1}^{M_k} \sqrt{P_S}h_{0,0,m}^{M,M}k,j,v_{k,j},s_{k,j} + n_m^M$$

Inter-user interference

where $P_M$ and $P_S$ denote the transit powers of MBS and SBS, respectively. $n_m^M$ denotes additive white Gaussian noise (AWGN) at $m$-th MU. $h_{a,b,c}^{M,M}$ and $h_{a,b,c}^{S,S}$ ($a \in \{0, 1, 2, \cdots, M\}$, $b \in \{0, 1, 2, \cdots, K\}$, $c \in \{0, 1, 2, \cdots, M_S\}$, $A, B \in \{M, S\}$) are pathloss coefficient and fading vector between base stations $A$ and users $B$, respectively. $A = M$ and $A = S$ denote MBS and SBS, respectively. $B = M$ and $B = S$ denote MU and SU, respectively. Here, subscripts $a$, $b$, and $c$ denote base station index ($a = 0$: MBS, $a \geq 1$: SBSs), cell index ($b = 0$: MC, $b \geq 1$: SC), and user index per cell. $w_i$ and $x_i$ denote $N_M \times 1$ precoding vector and transmit data for $i$-th MU, respectively. $v_{k,j}$ and $s_{k,j}$ are $N_S \times 1$ precoding vector and transmit data for $j$-th SU at the $k$-th SC, respectively.

Similarly, the received signal at $j$-th SU in the $k$-th SC can be written as:

$$y_{k,j}^S = \sqrt{P_S}h_{k,k,j}^{S,S}v_{k,j} + \sum_{l=1,l\neq k}^{K} \sqrt{P_S}h_{k,k,j}^{S,S}v_{l,n} + \sum_{u=1}^{M} \sqrt{P_M}h_{0,0,j}^{S,S}w_{u} + n_{k,j}^S$$

Desired signal

$$+ \sum_{i=1}^{M} \sum_{n=1}^{M_k} \sqrt{P_S}h_{k,k,j}^{S,S}v_{k,j},s_{k,j,n} + \sum_{m=1}^{M} \sqrt{P_M}h_{0,0,j}^{M,M}m_{m}w_{m} + n_{k,j}^S$$

Inter-user interference

Co-tier interference

Cross-tier interference

where $n_{k,j}^S$ denotes the AWGN.

Let $H^{M,M} = [(h_{0,0,1}^{M,M})^T, (h_{0,0,2}^{M,M})^T, \cdots, (h_{0,0,M}^{M,M})^T]^T \in \mathbb{C}^{N_M \times M}$ denote channel matrix estimated at MBS, where $(\cdot)^T$ denotes transpose of a vector $a$. To mitigate inter-user interference, ZF precoding matrix at MBS, $W \in \mathbb{C}^{N_M \times M}$, is given as

$$W = (w_1, w_2, \cdots, w_M)$$

$$= \frac{1}{\| (H^{M,M})^H (H^{M,M}(H^{M,M})^H)^{-1} \|_2} (H^{M,M})^H (H^{M,M}(H^{M,M})^H)^{-1},$$

where $(\cdot)^H$ denotes conjugate transpose of a complex matrix. $\| A \|_2$ denotes the Frobenius norm of $A \in \mathbb{C}$.

Similarly, let $H_{k,k}^{S,S} = [(h_{k,k}^{S,S})^T, \cdots, (h_{k,k,M_S}^{S,S})^T]^T \in \mathbb{C}^{M_S \times N_S}$ denote channel matrix between $k$-th SBS and associated SUs. ZF precoding matrix at $k$-th SBS $V_k \in \mathbb{C}^{N_S \times M_k}$ is written as:

$$V_k = (v_{k,1}, v_{k,2}, \cdots, v_{k,M_k})$$

$$= \frac{1}{\| (H_{k,k}^{S,S})^H (H_{k,k}^{S,S}(H_{k,k}^{S,S})^H)^{-1} \|_2} (H_{k,k}^{S,S})^H (H_{k,k}^{S,S}(H_{k,k}^{S,S})^H)^{-1}.$$

For example, $h_{0,0}^{M,M}$ stands for channel coefficient between MBS and $c$-th MU, while $h_{0,0}^{S,S}$ stands for that between $a$-th SBS and $c$-th SU in $b$-th SC.
Hence, signal to interference plus noise ratio (SINR) at \( m \)-th MU and that at \( j \)-th SU in \( k \)-th SC, are given below, respectively.

\[
\text{SINR}_m^M = \frac{P_M \beta_{0,0,m} |h_{0,0,m}^M w_m|^2}{\sum_{k=1}^{K} \sum_{j=1}^{M_k} P_S \beta_{k,0,m}^S |h_{k,0,m}^S v_{k,j}|^2 + \sigma^2},
\]

\[
\text{SINR}_{k,j}^S = \frac{P_S \beta_{k,k,j}^S |h_{k,k,j}^S v_{k,j}|^2}{\sum_{l=1,l \neq k}^{K} \sum_{n=1}^{M_l} P_S \beta_{l,k,j}^S |h_{l,k,j}^S v_{l,j}|^2 + \sum_{u=1}^{M} P_M \beta_{0,k,j}^M |h_{0,k,j}^M w_u|^2 + \sigma^2}.
\]

\section{Conventional scheme}

This section explains a pilot-user allocation and cross-tier interference mitigation scheme in [5]. In this scheme, all MUs and \( K_O \) SUs are allocated orthogonal pilots, where \( 0 \leq K_O \leq KM_S \). The other SUs are allocated the same orthogonal pilot, where \( KM_S \) denotes the total number of SUs. Hence, the number of required orthogonal pilots is \( M + K_O + KM_S \). MBS can distinguish not only \( M \) MU pilots but also \( K_O \) SUs’ pilot, because they are orthogonal each other. Thus, CSI of the path between MBS and \( K_O \) SUs can be estimated at MBS by extracting \( K_O \) SUs’ pilots. Hence, cross-tier interference from MBS to \( K_O \) SUs can be canceled by designing MBS precoding matrix \( \mathbf{F} \in \mathbb{C}^{NM \times K_O} \) as

\[
\mathbf{F} = (\mathbf{f}_1, \mathbf{f}_2, \cdots, \mathbf{f}_{K_O})
\]

\[
= \frac{1}{\|H^{MS}(H^{MS})^H\|_2^{-1}} (H^{MS})^H (H^{MS}(H^{MS})^H)^{-1},
\]

where \( H^{MS} = [(h_{0,0,1}, h_{0,0,2})^T, (h_{0,0,K_O})^T, (h_{M_S,0})^T]^T \in \mathbb{C}^{K_O \times NM} \) denotes interference channel matrix, and \( h_{0,0,j} \) denotes CSI of the path between MBS and \( j \)-th SU. \( \mathbf{f}_j \in \mathbb{C}^{NM \times 1} \) denotes the precoding vector to cancel cross-tier interference to \( j \)-th SU (\( j \leq K_O \)). From (3) and (7), the concatenated ZF precoding matrix at MBS, \( \mathbf{W}^M \in \mathbb{C}^{NM \times (K_O + M)} \), can be written as

\[
\mathbf{W}^M = (\mathbf{W}, \mathbf{F}) = (\mathbf{w}_1, \mathbf{w}_2, \cdots, \mathbf{w}_M, \mathbf{f}_1, \mathbf{f}_2, \cdots, \mathbf{f}_{K_O}).
\]

From (5) and (8), \( m \)-th MU’s rate is lower-bounded as [5]
\[ R^M_{\gamma_m} = \left[ 1 - \frac{K_O + M + M_S}{L} \right] \log_2 \left[ 1 + \frac{P_M \beta^M_{0,0,m}(N_M - M - K_O)}{\sum_{k=1}^{K} \sum_{l=1}^{M_S} P_S \beta^S_{k,l,0,m} + \sigma^2} \right]. \] (9)

where, \( L \) is the number of symbols per frame. From (6) and (8), \( j \)-th SU rate that uses the orthogonal pilot is lower-bounded as [5]

\[ R^S_{\gamma,k,j} = \left[ 1 - \frac{K_O + M + M_S}{L} \right] \log_2 \left[ 1 + \frac{P_S \beta^S_{k,k,j}(N_S - M_S)}{\sum_{l=1,l\neq k}^{K} \sum_{n=1}^{M_S} P_S \beta^S_{l,n,k,j} + \sigma^2} \right]. \] (10)

On the other hand, the other SU’s rate affected by cross-tier interference from MBS is lower-bounded as [5]:

\[ \hat{R}^S_{\gamma,k,j} = \left[ 1 - \frac{K_O + M + M_S}{L} \right] \log_2 \left[ 1 + \frac{P_S \beta^S_{k,k,j}(N_S - M_S)}{\sum_{l=1,l\neq k}^{K} \sum_{n=1}^{M_S} P_S \beta^S_{l,n,k,j} + \sum_{m=1}^{M} P_M \beta^M_{m,k,j} + \sigma^2} \right]. \] (11)

It is clear from Eqs (9), (10), and (11) that pilot overhead (i.e., pilot length) increases as \( K_O \) increases while SINR at each SU is improved by mitigating cross-tier interference. This implies that a trade-off relationship exists between pilot overhead and cross-tier interference mitigation capability.

The downlink sum-rate can be given as

\[ R = \sum_{m=1}^{M} R^M_{\gamma_m} + \sum_{k,j \in K_O} R^S_{\gamma,k,j} + \sum_{k,j \notin K_O} \hat{R}^S_{\gamma,k,j}, \] (12)

where \( K_O \) denotes a set of user indices who are allocated orthogonal pilots. The cardinality of \( K_O \) is \( K_O \). The above sum-rate can be maximized by optimizing pilot allocation \( K_O \) [5].

This scheme mitigates cross-tier interference from MBS to SUs while minimizing the required pilot overhead. However, cross-tier interference mitigation from SBS to MUs are not considered.

4 Proposed scheme

Based on works in [5], we propose an extended cross-tier interference mitigation scheme that coordinates not only cross-tier interference from MBS to SU but also that from SBS to MUs, where SBS estimates CrCSI of the path between SBS and MUs by overhearing MU’s pilots. The proposed scheme can mitigate cross-tier interference from SBS to MUs without changing the pilot allocation (i.e., without increasing the required pilot overhead).

\(^2\)The details of pilot allocation optimization algorithm is given in [5].
Similarly to the conventional scheme in [5], we assume that all MUs and $K_O$ SUs are allocated the orthogonal pilots and the remaining SUs are allocated the same pilots. Since all MUs’ pilots are orthogonal to SUs’ ones, each SBS can distinguish MUs’ and SUs’ pilots. Thus, SBS can estimate cross-channel state information (CrCSI) between SBS and MUs by receiving orthogonal pilots from MUs. More concretely, let $K_{IC}$ denote the number of MUs who are strongly affected by cross-tier interference from SBS. By detecting $K_{IC}$ MUs’ pilots, CrCSI between $k$-th SBS and $K_{IC}$ MUs can be estimated as

$$\mathbf{H}^S_M = [(\mathbf{h}_{k,0}^{SM})^T, (\mathbf{h}_{k,0}^{SM})^T, \cdots, (\mathbf{h}_{k,0}^{SM})^T]^T \in \mathbb{C}^{K_{IC} \times N_S},$$

where $0 \leq K_{IC} \leq N_S - M_S$. Hence, the precoding matrix at $k$-th SBS to cancel cross-tier interference to $K_{IC}$ MUs $\mathbf{G}_k \in \mathbb{C}^{N_S \times K_{IC}}$ can be given as

$$\mathbf{G}_k = (\mathbf{g}_{k,1}, \mathbf{g}_{k,2}, \cdots, \mathbf{g}_{k,K_{IC}}) = \frac{1}{\|[(\mathbf{h}_{k}^{SM})^H (\mathbf{h}_{k}^{SM})^H]^{-1}\|_2} (\mathbf{H}_{k}^{S,M} (\mathbf{H}_{k}^{S,M})^H)^{-1},$$

where $\mathbf{g}_{k,m} \in \mathbb{C}^{N_S \times 1}$ denotes the precoding vector of $k$-th SBS for $m$-th MU. From (4) and (14), the concatenated precoding matrix at $k$-th SBS, $\mathbf{W}^S_k \in \mathbb{C}^{N_S \times (M_S + K_{IC})}$, can be given as

$$\mathbf{W}^S_k = (\mathbf{v}_k, \mathbf{G}_k) = (\mathbf{v}_{k,1}, \mathbf{v}_{k,2}, \cdots, \mathbf{v}_{k,M_S}, \mathbf{g}_{k,1}, \mathbf{g}_{k,2}, \cdots, \mathbf{g}_{k,K_{IC}}).$$

Assuming $N_S > M_S$, each SBS can cancel $K_{IC} (\leq N_S - M_S)$ cross-tier interference at maximum at expense of decreasing the precoding gain to served SUs. More concretely, MUs’ sum-rate can be improved by increasing $K_{IC}$ but SU’s rate is decreased due to the decreased precoding gain. Thus, it is necessary to find the optimum value of $K_{IC}$ that maximizes the total sum-rate. In this article, each SBS mitigates cross-tier interference to the nearest MU.

### 5 Performance evaluation

We consider a single MC with a radius of 1000 meters and there is no user within 100 meters of MC center area, while the radius of each SC is 30 meters and there is no user within 5 meters of SC center area [5]. The minimum distance between two SBSs is 120 meters [5]. $P_M = 46$ [dBm] and $P_S = 26$ [dBm]. Noise power is $-104$ [dBm]. The other parameters are $N_M = 70$, $N_S = 10$, $M = 18$, $M_S = 2$ and $K = 13$. The number of symbols per frame is $L = 200$. The received signal is affected by pathloss and independent equal-level two-path Rayleigh fading. Pathloss between MBS and MU/SU is $27.3 + 39.1 \log 10(d)$, and pathloss between SBS and MU/SU is $36.8 + 36.7 \log 10(d)$ [5]. We consider two scenarios in Fig. 1. In scenario 1 (Fig. 1(a)), MUs are uniformly distributed in the MC, while scenario 2 (Fig. 1(b)) assumes that one MU is located near each SC and the rest of MUs are uniformly distributed within the MC, where $R_{edge}$ is 5 meters.

Fig. 2 shows the downlink sum-rate versus $K_O$ for the proposed method (solid line) and the conventional method (dotted line) [5]. Here, the conventional method corresponds to $K_{IC} = 0$, while the proposed scheme assumes $K_{IC} = 1$. In Fig. 2(a),

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This article assumes that each user equips single antenna. If each user has $N_A$ antennas, the numbers of antennas at MBS and each SBS must be $N_M \geq N_A (M + K_O)$ and $N_S \geq N_A (M_S + K_{IC})$. 

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there is optimal $K_O$ that maximizes SUs’ sum-rate. Since SBS uses spatial degree of freedom for mitigating cross-tier interference to MUs, SU’s downlink rate slightly decreases in both scenarios. In Fig. 2(b), as $K_O$ increases, the downlink rate of MUs decreases monotonously due to the increase of pilot overhead. MU’s downlink rate can be improved by applying the proposed scheme especially in scenario 2, because dominant cross-tier interference to MUs is mitigated. In Fig. 2(c), the downlink total sum-rate (i.e., MU rate + SU rate) can be improved by interference mitigation at SBS. This is because the effect of cross-tier interference mitigation to MUs is higher than the SUs’ rate degradation.

Fig. 3 shows downlink sum-rate as a function of $K_{IC}$, where $K_O = 4$ is chosen as the best value to maximize the achieved sum-rate of the proposed method. In both scenarios, as $K_{IC}$ increases, MU’s downlink sum rate is improved monotonously, while SU’s sum rate is degraded slightly due to the reduced precoding gain at each SBS. The total sum-rate can be improved by choosing the proper $K_{IC}$ in the propose scheme especially when MU is located near SC.
6 Conclusion

In this article, we have proposed a downlink cross-tier interference between MC and SCs. The results show the effectiveness of the proposed approach in terms of total sum-rate in two-tier TDD-HetNets.

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Wireless energy harvesting networks with multiple filter-and-forward relays

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Abstract: This letter proposes a filter-and-forward relay scheme with multiple relay nodes using wireless energy harvesting. A relay node processes the received signal by using a time-domain filter to suppress inter-symbol interference and forwards its outputs using the energy harvested from the received signals that arrive from the source node and from all the relay nodes. We consider the filter design method based on the signal-to-interference-plus-noise power ratio maximization problem under multiple constraints for relay transmit power. Simulation results show the effectiveness of the proposed scheme owing to the cooperation of multiple relay nodes.

Keywords: filter-and-forward relaying, wireless energy harvesting, relay networks, inter-symbol interference

Classification: Wireless Communication Technologies

References

[1] T. D. P. Perera, D. N. K. Jayakody, S. K. Sharma, S. Chatzinotas, and J. Li, “Simultaneous wireless information and power transfer (SWIPT): Recent advances and future challenges,” IEEE Commun. Surveys Tuts., vol. 20, no. 1, pp. 264–302, 2018. DOI:10.1109/COMST.2017.2783901

[2] A. A. Nasir, X. Zhou, S. Durrani, and R. A. Kennedy, “Relaying protocols for wireless energy harvesting and information processing,” IEEE Trans. Wireless Commun., vol. 12, no. 7, pp. 3622–3636, July 2013. DOI:10.1109/TWC.2013.062413.122042

[3] Y. Zeng and R. Zhang, “Full-duplex wireless-powered relay with self-energy recycling,” IEEE Wireless Commun. Lett., vol. 4, no. 2, pp. 201–204, Apr. 2015. DOI:10.1109/LWC.2015.2396516

[4] H. Chen, A. B. Gershman, and S. Shahbazpanahi, “Filter-and-forward distributed beamforming in relay networks with frequency selective fading,” IEEE Trans. Signal Process., vol. 58, no. 3, pp. 1251–1262, Mar. 2010. DOI:10.1109/TSP.2009.2035986

[5] J. Furukawa, T. Miyajima, and Y. Sugitani, “Filter-and-forward relay with self-energy recycling in frequency-selective channels,” Proc. IEEE GCCE 2019, pp. 1126–1130, Oct. 2019. DOI:10.1109/GCCE46687.2019.9015295

[6] V. Havary-Nassab, S. Shahbazpanahi, A. Grani, and Z. Q. Luo, “Distributed beamforming for relay networks based on second-order statistics of the channel state information,” IEEE Trans. Signal Process., vol. 56, no. 9, pp. 4306–4316, Sept. 2008. DOI:10.1109/TSP.2008.925945
1 Introduction

Recently, wireless energy harvesting (WEH) has attracted attention owing to its ability to prolong the lifetime of energy-constrained wireless nodes in small-scale cooperative wireless networks such as low power sensor networks. Simultaneous wireless information and power transfer (SWIPT) can be considered as one of the most promising applications of WEH and can be of fundamental importance in the 5G era [1]. A potential application of SWIPT is wireless powered relay networks where relay nodes assist in the communication between source and destination nodes while harvesting the energy from radio frequency (RF) signals.

Various schemes of wireless powered relay networks with WEH have been considered [2, 3]. In [2], an amplify-and-forward (AF) relaying scheme with WEH was proposed, in which a relay node simply forwards properly scaled and phase-shifted versions of the received signal using the energy harvested only from the received signal arriving from the source node. In [3], a new WEH technique called self-energy recycling was proposed, in which an AF relay node operating in full-duplex mode harvests the energy of the received signal arriving from the source node as well as its own transmitted signal. AF relaying schemes work well in frequency-flat channels, but generally suffer from performance degradation due to the influence of inter-symbol interference (ISI) in frequency-selective channels. An approach to overcome ISI is to employ a filter-and-forward (FF) relaying scheme [4], in which a relay node processes the received signal by a finite impulse response (FIR) filter and forwards its outputs. In [5], we have proposed to combine the FF relaying scheme and the self-energy recycling technique, and demonstrated its effectiveness in frequency-selective channels. In [5], however, we limited our research to the case of a single FF relay node with self-energy recycling. If the single relay node suffers from severe shadowing, it may lead to system failure. One would generally expect to avoid such failures if multiple relay nodes are available. However, thus far, there has been no study concerning WEH relay networks with multiple FF relay nodes.

In this letter, we propose WEH relay networks with multiple FF relay nodes using self-energy recycling. We design the filter coefficients such that they maximize the signal-to-interference-plus-noise power ratio (SINR) at the destination node under multiple constraints that limit the transmitted power of each relay node to its harvested power. Unfortunately, the filter design method in [5] cannot be applied to the case considered here owing to the multiple constraints and use of self-energy recycling. Thus, we consider employing a semi-definite relaxation to obtain the filter coefficients.

2 System model

Fig. 1(a) depicts a relaying network consisting of a pair containing a single-antenna source node (S) and single-antenna destination node (D) and \(N_R\) FF relay nodes \((R_i, i = 1, \cdots, N_R)\) equipped with a receive antenna and a transmit antenna. We assume that there is no direct path between \(S\) and \(D\) owing to the severe path loss and shadowing, and thus \(S\) and \(D\) communicate via \(R_i.s\). We further assume that single-carrier transmission is employed, and all the channels are frequency-selec-
tive and quasi-static during data transmission. Each FF relay node has an FIR filter to cooperatively suppress ISI occurring at $D$ and harvests energy from its received signals. We also assume that $D$ computes the filter tap coefficients and feeds them back to $R_i$ using a low-rate feedback link. As shown in Fig. 1(b), a time slot of duration $T$ consists of two phases of $T/2$ each. In the first phase, $R_i$ receives a signal from $S$, and harvests energy from a portion of the received signal and processes the remaining portion as an information-bearing signal. In the second phase, $R_i$ operating in the full-duplex mode transmits its filter outputs toward $D$, while it receives signals from all the relay nodes.

First, we explain the information processing aspect. In the first phase, the discrete-time complex envelope of the received signal of $R_i$ is represented as

$$r_{i,j}[n] = \sum_{l=0}^{L_f-1} f_{i,j}[n-l] + m_{i,j}[n],$$

where $f_{i,j}$ is the $l$th tap of the channel impulse response (CIR) between $S$ and $R_i$, $L_f$ is the length of $f_{i,j}$, $s[n]$ are i.i.d. transmitted symbols, and $m_{i,j}[n]$ is the noise introduced by the receive antenna of $R_i$. We apply the power splitting scheme which divides the received signal into a $\lambda$ portion for energy harvesting (EH) and a $1-\lambda$ portion for information processing, where $0 < \lambda \leq 1$ is the power splitting ratio. The baseband output signal of the FIR filter at $R_i$ is written as

$$t_l[n] = \sum_{j=0}^{L_w-1} w_{i,j} \sqrt{1-\lambda} r_{i,j}[n-l] + z_i^R[n],$$

where $w_{i,j}$ is the $l$th tap filter coefficient of $R_i$, $L_w$ is the length of the filter, and $z_i^R[n]$ is the noise introduced by the RF band to the baseband signal conversion at $R_i$ [2]. In the second phase, $R_i$ transmits its filter outputs $t_l[n]$ using the harvested power $p_{EH,i}$ described later. The transmitted signal $t_l[n]$ passes through a frequency-selective channel and then reaches $D$. The baseband received signal at $D$ can be written as

$$y[n] = \sum_{i=1}^{N_R} \sum_{l=0}^{L_g-1} g_{i,j} t_l[n-l] + v_D[n],$$

where $g_{i,j}$ is the $l$th tap of the CIR between $R_i$ and $D$, $L_g$ is the length of $g_{i,j}$, and $v_D[n]$ is the sum of the antenna and conversion noises at $D$. Substituting (1) into (2) and performing simple mathematical manipulations, we obtain
\[ y[n] = w^H G(\sqrt{1 - \frac{\alpha}{\beta}} \tilde{s}[n]) + \sqrt{1 - \frac{\alpha}{\beta}} \tilde{h}^T \hat{w}[n] + v^D[n], \]

where \( w \triangleq [w_{1,0} \ w_{2,0} \cdots w_{N_R L_w, 0}]^T \), \( \tilde{s}[n] \triangleq [s[n] \cdots s[n - L_f - L_w - L_g + 3]]^T \), \( \tilde{h}^T \triangleq [m_{1,1}^R \cdots m_{N_R L_w, 0}^R \cdots m_{N_R L_w, N_R}]^T \), \( \hat{w}[n] \triangleq [z_0^R[n] \cdots z_{N_R L_w}^R[n - L_w - L_g + 2]]^T \), \( \hat{v}^D[n] \triangleq [z_0^D[n] \cdots z_{N_R L_w}^D[n - L_w - L_g + 2]]^T \), \( G \triangleq [I_{L_w} \otimes G_0 \cdots I_{L_w} \otimes G_{L_w - 1}] \), \( G_l \triangleq \text{diag}\{[g_{1j} \cdots g_{N_R l}]\} \), \( \tilde{\bar{F}} \triangleq \text{toeplitz}(\tilde{F}, L_w) \), \( \tilde{\tilde{F}} \triangleq \text{toeplitz}(F, L_w) \), \( F \triangleq [f_{10} \cdots f_{N_R l}]^T \), the operator \( \otimes \) denotes the Kronecker product, and \( \text{toeplitz}(A_{N \times M}, L) \triangleq [A_0^T \cdots A_{L - 1}^T]^T \) is a \( NL \times (M + L - 1) \) block-Toeplitz matrix with \( A_i \triangleq [0_{N \times l} \ A_{N \times M} 0_{N \times (L - 1 - l)}] \). The first term in (3), \( w^H \tilde{\bar{F}} \tilde{s}[n] \), contains the desired and ISI components. We assume that the \( d \)th entry \( s[n - d + 1] \) of \( \tilde{s}[n] \) is the desired symbol. For later convenience, we denote a vector obtained by removing \( s[n - d + 1] \) from \( \tilde{s}[n] \) as \( \hat{\tilde{s}}[n] \). Also, we denote the \( d \)th column of \( \tilde{\tilde{F}} \) as \( \tilde{\tilde{f}} \) and the remaining matrix as \( \tilde{\bar{F}} \).

Next, we explain the EH operation performed at \( R_i \). In the first phase, \( R_i \) harvests the power from a portion of the signal transmitted from \( S \). The harvested power in the first phase is given as \( p_{1,i} = \eta E[|\sqrt{\eta} r_{1,i}[n]|^2] \), where \( 0 < \eta \leq 1 \) is the power conversion efficiency. In the second phase, \( R_i \) receives the signals transmitted from all the relay nodes. The complex envelope of the received signals of \( R_i \) is given as \( r_{2,i}[n] = \sum_{j=1}^{N_R} \sum_{l=0}^{L_g - 1} h_{j,i} t_j[n - l] + m_{2,i}^R[n] \), where \( h_{j,i} \) is the \( j \)th tap of the CIR between \( R_j \) and \( R_i \), \( L_h \) is the length of \( h_{j,i} \), and \( m_{2,i}^R[n] \) is the noise introduced by the receive antenna. Then, \( R_i \) harvests the power from the signals \( r_{2,i}[n] \). The harvested power in the second phase is given as \( p_{2,i} = \eta E[|r_{2,i}[n]|^2] = \eta w^H H_i \tilde{\bar{D}}_i^H w \), where

\[
\begin{align*}
\tilde{\bar{D}} & \triangleq I_{L_w} \otimes \tilde{\bar{D}}, \quad \tilde{\bar{D}} \triangleq [\tilde{D}_1, \cdots, \tilde{D}_{N_R}], \quad \tilde{D}_i \triangleq [\tilde{D}_i, \cdots, \tilde{D}_i], \\
D_{i,j} & \triangleq P_s(1 - \rho) A_i \tilde{\tilde{F}}^H A_j^H + \sigma_m^2 (1 - \rho) A_i A_j^H + \sigma_z^2 A_i A_j^H, \\
D_{i,j} & \triangleq P_s(1 - \rho) A_i \tilde{\bar{F}}^H A_j^H, \quad A_i \triangleq I_{L_w} \otimes E_i, \quad E_i \triangleq \text{diag}\{e_i\}, \\
H_i & \triangleq h_i^T \otimes I_{N_R L_w}, \quad h_i \triangleq [h_{1,i} \cdots h_{N_R,i}]^T, \\
P_s & \triangleq E[|s[n]|^2], \quad e_i \text{ is the } i \text{th column of the identity matrix, and } \sigma_m^2, \sigma_z^2 \text{ are the powers of } m_0^R[n], z_0^D[n], \text{ respectively. We assume that the harvested power originating from the noises is very small and thus can be ignored [2]. Finally, the total harvested power } p_{EH,i} \text{ at } R_i \text{ is the sum of } p_{1,i} \text{ and } p_{2,i}, \text{ i.e., } p_{EH,i} = p_{1,i} + p_{2,i}. \text{ The purpose of the FIR filters } w \text{ at relays is to suppress ISI occurring at } D, \text{ while the desired component is to be enhanced when the available transmit power is limited to } p_{EH,i}. \end{align*}
\]

### 3 Filter design

We consider to determine the filter coefficients vector \( w \) to maximize the received SINR at \( D \). The optimization problem can be formulated as

\[
\max_w \text{ SINR} \quad \text{s.t.} \quad p_{1,i} \leq p_{EH,i}, \quad i = 1, \cdots, N_R
\]

where \( p_{1,i} = E[|t_i[n]|^2] = w^H D_{i,j} w \) is the power consumed at \( R_i \) for signal transmission. The multiple constraints ensure that the transmitted power of each relay node does not exceed its harvested power.
To solve the problem (4), we take a similar approach as in [6]. From (3), we can rewrite (4) as
\[
\max_w \frac{w^H Q_w}{w^H Q_w + w^H Q_s + \sigma^2_{\nu,0}} \quad \text{s.t.} \quad w^H D_i w \leq p_{1,i}, \quad i = 1, \ldots, N_R
\]
where \( Q_s \triangleq P_s (1 - \lambda) \tilde{G} \tilde{F}^H \tilde{G}^H \), \( Q_i \triangleq P_s (1 - \lambda) \tilde{G} \tilde{F}_i \tilde{G}^H \), \( Q_n \triangleq \{ (1 - \lambda) \sigma^2_m + \sigma^2_{\nu,0} \} \tilde{G} \tilde{H}^H \), \( \tilde{D}_i \triangleq (D_{ij} - \eta H_i \tilde{H}^H_j) \), and \( \sigma^2_{\nu,0} \) is the power of \( \nu \tilde{D}^H \). We define a semi-definite matrix \( W \triangleq w w^H \) and an auxiliary variable \( \tau > 0 \) corresponding to SINR. Using a semi-definite relaxation [6], we have
\[
\max_{\tau, W} \quad \text{s.t.} \quad \text{tr}((Q_s - \tau (Q_s + Q_n)) W) \geq \tau \cdot \sigma^2_{\nu,0} \\
\text{tr}((\tilde{D}_i W) \leq p_{1,i}, \quad i = 1, \ldots, N_R \\
W \geq 0,
\]
where \( W \geq 0 \) implies that \( W \) is a positive semi-definite matrix. We solve this optimization problem by using a bisection technique. The optimization procedure is summarized as follows:

**Step1)** Set the initial interval of \( \tau \) as \([\tau_l, \tau_u] = [0, \tau_0]\).

**Step2)** Set \( \tau := (\tau_l + \tau_u)/2 \).

**Step3)** Solve the following feasibility problem.
\[
\text{find } W \quad \text{s.t.} \quad \text{tr}((Q_s - \tau (Q_s + Q_n)) W) \geq \tau \cdot \sigma^2_{\nu,0} \\
\text{tr}((\tilde{D}_i W) \leq p_{1,i}, \quad i = 1, \ldots, N_R \\
W \geq 0.
\]
If (7) is feasible, then \( \tau_l := \tau \), otherwise \( \tau_u := \tau \).

**Step4)** Repeat Step2) and Step3) until \((\tau_u - \tau_l) < \varepsilon\).

The error tolerance value \( \varepsilon \) is set to small value. After we obtain the solution \( W_{\text{opt}} \), we apply the randomization procedure [6] to obtain the filter coefficients vector \( w \).

Using the eigen-decomposition \( W_{\text{opt}} = T \Sigma T^H \), we generate \( N_p \) candidate vectors as \( w_p = \sqrt{T} \Sigma_p \), \( p = 1, 2, \ldots, N_p \), where \( \Sigma_p \) is a vector with zero-mean unit-variance complex Gaussian random variables. Finally, we choose the best solution among these candidates that maximizes the SINR while satisfying the constraints in (4).

### 4 Simulation

We conducted simulations to evaluate the performance of the proposed system. The relay nodes are placed at equal intervals along the line that is orthogonal to the line through \( S \) and \( D \). The closest distance between \( S \) and \( R \) is 10 m. The closest distance between \( R \) and \( D \) is 10 m. The distance between the transmit antenna to the receive antenna of \( R_i \) is 0.3 m. The distance between adjacent relays is 1 m. The modulation scheme used is QPSK. The power of all noises is −90 dBm. Taking into account the pass loss and Rayleigh fading, we can model the CIR coefficients \( f_{ij}, g_{ij}, h_{ij} \) as zero-mean complex Gaussian variables with the variance of \( \Lambda_{XY} d_{XY}^m \) where \( \Lambda_{XY} \) is the average signal power attenuation which depends on the antenna gains and carrier frequency. We set \( \Lambda_{S,R_i} = 10^{-3} \) and \( \Lambda_{S,R_i} = \Lambda_{R_i,D} = 10^{-2} \). The average BER is obtained from \( 10^3 \) trials, where each trial has an independent channel realization and \( 10^6 \) data symbols. \( \tau_0 \) is set to the maximum
SINR at $D$ when the relay transmit power is $P_s$. Other simulation parameters are as follows: $m = 3$, $P_s = 30$ dBm, $L_f = L_g = 3$, $L_h = 2$, $L_w = 3$, $\lambda = 0.9$, $\eta = 0.5$, $\varepsilon = 10^{-4}$, and $N_p = 10^4$.

Fig. 2 depicts the BER performances against the source transmit power $P_s$ for varying number of relay nodes $N_R$. As can be seen, the performance improves as $P_s$ increases. This is because the available transmit power of $R_i$ increases as $P_s$ increases. Moreover, it can be observed that the performance improves as $N_R$ increases owing to the cooperative diversity achieved by spatially distributed relay nodes.

Fig. 3 depicts the BER performances against the length of filter $L_w$ for varying $N_R$. The performance of the AF scheme corresponding to the case with $L_w = 1$ scarcely improves despite the increase in $N_R$. This is because AF cannot compensate for ISI. Alternatively, we observe that the use of FIR filter ($L_w \geq 2$) significantly improves the performance as $N_R$ increases.

5 Conclusion

In this paper, we proposed a filter-and-forward relay network for single-carrier transmissions in frequency-selective channels, where multiple relay nodes harvest the energy from the RF signals. Our simulation results show that the proposed relay network can efficiently suppress ISI and achieve a significant performance improvement owing to distributed multiple relay nodes.
Novel sensing techniques of chipless RFID sensor for infrastructure

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Abstract: Monitoring applications using sensors are being led by the government to address the problem of aging infrastructure in Japan. There are various methods for sensing but the durability of general infrastructures and buildings have already exceeded 30 years which makes it difficult to address the issue. As one of the solutions we focused on the technology of “Chipless sensor” that applies “Chipless RFID” technology composed only of metal patterns without using electronic components such as IC chips. This report describes a sensor that can detect moisture with just metal patterns by supplying power from the reader.

Keywords: chipless RFID, sensor, moisture detection, infrastructure

Classification: Sensing

References

[1] Y. Liu, F. Deng, Y. He, B. Li, Z. Liang, and S. Zhou, “Novel concrete temperature monitoring method based on an embedded passive RFID sensor tag,” Sensors, vol. 17, no. 7, 1463, 2017. DOI:10.3390/s17071463
[2] N. C. Karmakar, E. M. Amin, and J. K. Saha, Chipless RFID Sensors, Wiley, New York, NY, USA, 2016. DOI:10.1002/9781119078104
[3] A. Vena, E. Perret, and S. Tedjini, “High-capacity chipless RFID tag insensitive to the polarization,” IEEE Trans. Antennas Propagat., vol. 60, no. 10, pp. 4509–4515, Oct. 2012. DOI:10.1109/TAP.2012.2207347
[4] M. Zomorrodi and N. C. Karmakar, “Cross-RCS based, high data capacity, chipless RFID system,” Proc. IEEE MTT-S Int. Microw. Symp. (IMS), Tempa, FL, USA, pp. 1–4, 2014. DOI:10.1109/MWSYM.2014.6848402
[5] A. Vena, E. Perret, and S. Tedjini, “A depolarizing chipless RFID tag for robust detection and its FCC compliant UWB reading system,” IEEE Trans. Microw. Theory Techn., vol. 61, no. 8, pp. 2982–2994, Aug. 2013. DOI:10.1109/TMTT.2013.2267748
[6] H. Nakajima, M. Takahashi, K. Saito, and K. Ito, “Simple urination detection system using RFID,” IEICE Commun. Express, vol. 2, no. 3, pp. 98–103, 2013. DOI:10.1587/comex.2.98
1 Introduction

With the expansion of IoT technology, various products equipped with communication and sensor functions have been introduced. Under such circumstances, RFID technology that can operate under a little power solely by radio waves is attracting attention, and expected to be introduced into infrastructure facilities due to its little energy operation [1]. On the other hand, the durability of infrastructure and buildings has already exceeded 30 years, so even a battery-less RFID sensor is difficult to meet the durability. In order to solve this challenge, we looked into a possibility of “Chipless RFID sensor (Chipless sensor)” [2] technology that can detect the sensor value only with metal pattern. By irradiating electromagnetic waves, the metal pattern resonates and can detect the sensor value. There is no need for electronic components such as IC chips on the tag side. Therefore, long-term durability can be expected. In this paper, we report a sensor that detects the moisture by “Chipless sensor” technology in the real environment. And those researches would pave the way for detecting water leakage of the infrastructure over long period.

2 Chipless sensor detection system

Fig. 1(a) shows the configuration of the chipless sensor system. The tag has a metal etched pattern on the base material. In these researches, we adopted the frequency-domain system where IDs are generated as resonance peaks in specific frequency
band [2]. Electromagnetic waves were irradiated toward the chipless sensor using a network analyzer (KEYSHIGHT: E5071C), and the reflected waves were detected as resonant peak of Radar Cross Section value using reference sphere. Since the wide frequency band could generate multiple resonance peaks, the UWB band (7 to 10 GHz) is utilized which is legally allocated wide band frequency in Japan. Horn antenna is used for Tx and Rx antenna (Band: /Gain15dBi @ 10 GHz).

2.1 Chipless tag
Details of the tag are shown on the Fig. 1(b). The tag has a three-layer structure composed of copper foil and low dielectric loss material. A low dielectric loss material (MEGTRON6 R-5775N: 250 µm thick) sandwiches the copper pattern and solid copper foil (Copper: 18 µm thickness). Electromagnetic waves coming from Tx antenna resonate with each chipless sensor, and each reflects waves received by Rx antenna. The reflected wave is not modulated and is disturbed by the attached material and the surrounding environment. Since the tag has a sandwich structure for generating an electric field between the upper and lower copper foils, the influence on the back side is mitigated [3].

2.2 A depolarizing chipless RFID tag for robust detection
However, when a tag is attached on the actual objects such as wood wall, cardboard box and concrete blocks, tag peaks are buried by standing waves of objects. Thus, technologies that mitigate the effects of standing waves have been proposed [4, 5]. By making Tx and Rx orthogonal, each electric field plane also becomes orthogonal, making the standing wave of the object smaller, and the tag peak clearly detectable. Fig. 1(c) indicates two tag peaks of the Tx and Rx electric field in parallel (V) and orthogonal (H) to each other, and the parallel is flat and almost has no peak, while the orthogonal is above the peak at around 8.6 GHz. It is, therefore, easy to detect the peak by making Tx and Rx orthogonal.

3 Angle control for fine detection
Fig. 2(a) shows a chipless sensor that consists of 12 elements in one plane, assuming moisture detection such as a water leakage. The Radar Cross Section results of the electromagnetic simulation (ANSYS Electronics) are shown on the right side of the Fig. 2(a), and it can be confirmed that 12 peaks appear between 7.5 and 9.5 GHz. For easy understanding, red dots are marked on the each peaks at Fig. 2(a). However, in actual measurement, Tx and Rx are broadband horn antenna, the distance between the antenna and the tag is 50 cm, and the transmitted and received angle is set at 30 degrees. As shown in Fig. 2(c), the transmitted and received angle is defined with reference to a line passing through the center of objects. As a result is shown in Fig. 2(b), sensor peak could not be detected well, especially on the high frequency side. Because of the frequency characteristic, it is assumed that there is an influence of object reflection. Therefore, we tried to confirm the reflection of the tag and the object. The reflection intensity was measured by changing the transmitted angle, fixed with received angle as 30 degrees. The result shown in Fig. 2(d), confirmed that the tag had a strong peak.
at the 0° direction, whereas the object had directivity at the 30° direction. In order to evaluate the angle effect of the chipless sensor as shown in Fig. 2(d), the reflection intensity was read by placing the Rx so that received angle was 0°, 30°, and 60° in front of the Tx toward the object. When it was set at 0°, 12 peaks could be clearly detected. On the other hand, at 30° and 60°, the peaks at low and high frequencies were particularly dull. This is because some sensor elements at both ends carrying low frequency and high frequency side could not be received by the surrounding noise.

![Figure 2](image)

**Fig. 2.** a) 12 elements of sensor tag (tag components, simulation results), b) 12 elements of sensor tag peak of actual environment, c) Overview of directivity measurement, d) Reflection intensity of Tag and Cardboard, e) 12 elements of sensor tag peak when transmitted angle 0, 30, 60 degree

## 4 Water detection on concrete blocks

Assuming installation on an infrastructure, the chipless sensor tag were attached to a 300 × 300 × 50 mm concrete block. As shown in Fig. 3(a), when water droplets were dropped on several sensor elements, the sensor value could be detected as lost...
peaks. In this system, the peak of the element with water disappears, and the peak without water remains unchanged, so the position with water can be clearly detected. Sensors using RFID could also perform water detection due to the loss of tag functionality [6], however chipless sensors could identify the position of water adhesion. And this is the key feature of chipless sensor.

Furthermore, when a large amount of water is dripped on the assumption of water leakage, the sensor value as shown in Fig. 3(b) was confirmed. It was found that even when water entered the back of the element, since the floor level was not greatly lowered, moisture on the surface alone could be detected accurately.

**5 Conclusion**

In this paper, we showed the possibility of utilizing chipless sensor for infrastructure by adjusting the reading angle.

In the future, it will be necessary to structure a system that can detect the peak in concrete or in some other objects in consideration to embed in the structures.

In addition, since the outdoor use of the UWB band investigated this time is legally limited to few frequencies, it is necessary to solve this point and problems when using it in infrastructure applications.
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