Replicating a rendezvous node for a core-based tree multicast protocol in NDN networks for providing low latency

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Abstract: Since data obtained by Internet of Things devices is correlated with their locations, location-based services of exploiting such locality are promising. An example service is that a driver, i.e., a collector of data, asks cars on the road 2 km ahead of her or him how many cars are driving there. Since such a service requires low latency multicasting communication, this paper designs a low latency publish/subscribe mechanism by leveraging name-based communication like Named Data Networking.

Keywords: publish/subscribe communication, Named Data Networking, multicast protocol and low latency communication

Classification: Network

References

1 Introduction
Data obtained by Internet of Things (IoT) devices is correlated with their locations, and hence location-based services (LBSs) of exploiting such locality are promising [1]. Key requirements to LBSs are summarized as data retrieval from multiple IoT
devices and low latency communication. Hereafter, such data is called location data. Since location data retrieval has inherently a multicasting nature, we choose Content Oriented Publish/Subscribe Communication System (COPSS) [3] which provides publish/subscribe communication in Named Data Networking (NDN) networks for data retrieval. Its name-based communication avoids round trip delays for resolving names of IoT devices to physical addresses.

Despite the advantage of avoiding name resolution, a core-based multicast protocol adopted by COPSS incurs additional hops to the shortest paths between a collector and IoT devices because multicast trees are constructed from a common root called a rendezvous node (RN). The paper reduces such additional hops by placing multiple RNs so that a collector of data and each IoT device rendezvous to their closest RN.

2 Related work

Publish/subscribe communication protocols are used for IoT data retrieval. MQTT [2] is a protocol in IP networks and a central server plays a role of an RN. However, TCP/IP incurs long latency due to its rich flow control. On the contrary, COPSS [3], which does not have such control and name resolution, provides shorter latency than MQTT. The paper proposes placing multiple RNs to reduce hops inspired by an anycast mechanism in IP networks [4].

3 Architecture

Fig. 1 shows the architecture for location data retrieval.

3.1 Naming scheme

A name of location data is specified by a sequence of a location name and a data name, delimited by the reserved word #dat. A location name is specified according to Z-order as our previous work proposes [5]. A Z-order number is a quaternary number, \( z_1z_2\ldots z_m \). The most significant digit \( z_1 \) corresponds to one of the four squares divided from the largest square. Each square is divided into four squares and they are numbered from \( z_10 \) to \( z_13 \). Smaller squares are recursively numbered in the same way. Z-order quaternary numbers are naturally mapped onto name components like /\( z_1 / z_2 / \ldots / z_m \). An example location data name is /3/0/0/#dat/car/speed.

Fig. 1. Architecture for location data retrieval
3.2 Data retrieval by leveraging COPSS

COPSS [3] consists of the two procedures: First, subscribers send a subscribe packet to the RN so that a multicast tree is constructed from the RN to them. Its destination is a location data name, e.g., /3/0/0/#dat/car/speed in Fig. 1. The name of RN, i.e., /RN, is added to it so that the subscribe packet is delivered to the RN. COPSS routers on the path from each subscriber to the RN records a pair of a location data name and an outgoing face in its subscription table (ST). Second, a publisher sends a publish packet to the RN with a location data name as a destination. After receiving it, the RN forwards this packet after removing the RN name. The publish packet is forwarded by downstream COPSS routers according to their STs.

A collector retrieves location data in the two steps assuming that all IoT devices subscribe to their location data names and a collector subscribes to its name, e.g., /Collector. First, a collector sends a publish packet to all IoT devices of which location data names are of interest. Second, each of the IoT devices sends a publish packet with location data to the name.

4 Anycast mechanism

In the case of native COPSS, a single RN is deployed, as shown in Fig. 1. To avoid additional hops incurred by the single RN, we place a group of RNs at edge routers, as illustrated in Fig. 2, wherein they are called an anycast group.

The idea behind the RN replication is summarized below: A multicast tree which has multiple roots is constructed. Any RN anycasts a received publish packet to the other RNs so that the publish packet is delivered to all the subscribers whichever RN receives it. An important assumption is that a collector retrieves location data which is geographically adjacent to it. Since a geographical topology and a router topology are similar, many pairs of collectors and IoT devices rendezvous to their closest RN to each other.

1) Multicast tree construction: Two types of STs are used for subscribers and the other RNs in the anycast group as shown in Fig. 2. This paragraph explains how the two STs of RN1 are updated when it receives subscribe packets. Each RN has a name in the form of /RN/RN, where /RN is the common prefix of all the RN names. First, a subscriber sends a subscribe packet to /RN, precisely speaking /RN/3/0/0/#dat/car/speed. The packet is forwarded to the closest RN, i.e., /RN/RN1. RN1 records the pair of the location data name and the face #0 at the ST for /RN/RN1 and then anycasts the subscribe packet to the other RNs. Second, when RN1 receives a subscribe packet with the name /RN/3/0/2/#dat/car/speed from other RN, e.g., RN2, it records the pair of the name and the RN’s name, i.e., /RN/RN2, at the ST for /RN.

Although the replication reduces the average path length, it increases the number of ST updates of each RN because subscribe packets triggered by device movements are delivered from the closest RN to the device to all the other RNs. In Section 5, we evaluate their tradeoff relations.

2) Publish packet forwarding: When a collector, e.g., the publisher in Fig. 2, sends a publish packet to a location data name, i.e., /RN/3/0/0/#dat/car/speed,
it is forwarded to the closest RN, i.e., /RN/RN1. Then the RN forwards the packet to the faces corresponding to the location data name, and anycasts it to the faces to the other RNs. If the upstream RNs of the publisher and the subscriber are the same, the publish packet is forwarded via the RN which is the closest to both of them.

5 Evaluation through simulations

This section evaluates path lengths and ST updates through simulations.

5.1 Simulation conditions

The target area is the special wards of Tokyo, i.e., a square area of 32 km width. The area is recursively divided eight times. The smallest square has an 8-digit Z-order number and its width is 0.125 km. We construct a router-level topology on the basis of the locations and the coverage areas of telephone exchange buildings of NTT East Corporation. The router-level topology is a tree, and its depth is three. The root router, i.e., the first-level router, is placed at the building in Otemachi. The six second-level routers are placed at buildings near the 6 big terminal stations. The third-level routers are placed so that the target area is covered uniformly, and the routers are connected with the closest second-level routers. The fourth-level routers are placed at all the buildings and they are connected with the closest third-level routers.

Moving and stationary devices are deployed as follows: 2.4 million cars, which move at the average speed of 30 km/h, are deployed on realistic roads in Tokyo. Besides, 1 million stationary sensors are randomly deployed. In the evaluation, drivers obtain location data of cars in a circle of 200 m radius located at 2 km ahead.

5.2 Path stretch due to RN deployment

The average path lengths are 3.35, 2.98 and 3.30 in the cases where RNs are deployed at the 2nd, 3rd and 4th-level routers, respectively. The path lengths are averaged over these 1,000 requests and the path length is defined as the number of hops from a car to cars with speed sensors. 1,000 cars are randomly selected as collectors. The number of hops to retrieve location data is much smaller than that of the case of the single RN. In the case where a single RN is deployed at the first level router, the number of hops between cars and the RN is four and thus the path length is always eight.
5.3 Frequency of ST updates

Fig. 3 shows the average frequencies of ST updates which add and delete entries of location data names in each RN in the case of various numbers of cars. The average frequency decreases as the cars’ number increases. It means that the larger the density of cars becomes, the smaller the frequency of ST updates becomes. This is because, in the case that cars are densely deployed, even if all cars move, location data names recorded at edge routers do not frequently change. This reduces the number of subscribe packets forwarded by edge routers to RNs. In addition, even if cars are sparsely deployed, the number of ST updates in a minute is at most 150 in total. Hence, packets generated to update STs is not a serious issue.

![Graph showing the frequency of ST updates](image)

**Fig. 3.** The frequency of ST updates

6 Conclusion

The paper designs an anycast mechanism to enhance a name-based public/subscribe protocol to achieve low latency multicasting.

Acknowledgments

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Abstract: This paper presents a user throughput performance analysis of a coordinated radio-resource scheduler with a hardware accelerator for the fifth generation of mobile communications systems (5G). To ensure fairness among mobile terminals (MTs) in 5G ultra-high-density distributed antenna systems, the transmission weight matrixes for a huge number of possible combinations of antennas and MTs have to be computed in a coordinated radio-resource scheduling process. To accelerate the computation, we devised a scheduler that executes the computation on dedicated hardware. System-level simulations reveal that the scheduler with a hardware accelerator composed of 5.3 Mgates obtains around 20 Mbps at the 5th percentile of average user throughput for 256 MTs. Computational resources are hence the critical factor in obtaining high average user throughput and ensuring fairness among MTs.

Keywords: 5G mobile communications systems, resource scheduling, distributed antenna systems, hardware acceleration

Classification: Wireless Communication Technologies

References


1 Introduction

Ultra-high-density distributed antenna systems have been studied for the fifth generation of mobile communications systems (5G) [1, 2]. To accommodate rapidly increasing mobile data traffic, multiple antennas will be deployed in an ultra-dense pattern [Fig. 1(a)]. As shown in Fig. 1(b), the radio transmissions of all antennas are controlled at a centralized baseband unit equipped with a radio-resource scheduler.

In mobile communications systems, the scheduler determines a combination of antennas and mobile terminals (MTs) that are granted the right to transmit data. An optimal combination raises the sum of data throughputs from all antennas to the MTs (i.e., system throughput). At the same time, the scheduler must ensure fairness among MTs. To search for the optimal combination, it probes a combination and computes a metric based on MT selection algorithms, such as proportional fairness (PF) or the maximum carrier-to-interference ratio (Max C/I). By iterating such probes and computations, the scheduler determines the optimal combination.

In addition, this determination must be completed within a scheduling period of 1 millisecond in the long-term evolution (LTE) specifications [3]. If the scheduler cannot find the optimal combination within this time, the system cannot achieve higher system throughput and cannot ensure fairness among MTs.

To suppress interference among multiple antennas in 5G systems, a coordinated radio resource scheduling algorithm has been proposed [1]. The algorithm computes a transmission weight matrix from the channel matrixes reported from the MTs. Then, it estimates the data throughput from the transmission weight matrix and computes a metric for a combination. When the number of MTs is small, the scheduler executes this computation using software-based processing as in a...
conventional scheduler, and it can find the optimal combination within the scheduling period. However, it is difficult for a software-based scheduler to find the optimal combination when the number of MTs is large because it takes longer to compute the transmission weight matrix and it cannot probe a sufficient number of combinations.

We previously revealed in [4, 5, 6, 7] that system throughput in ultra-high-density distributed antenna systems can be increased by using a scheduler equipped with a hardware accelerator (HWA) using the Max C/I MT selection algorithm. In this paper, we focus on the user throughput performance and discuss fairness among MTs. The average user throughput required for providing 5G services, such as high-definition video, is around 20 Mbps even for MTs in poor channel conditions. To ensure fairness among MTs, we have devised a coordinated radio resource scheduler with an HWA that accelerates the computation of a PF metric by using dedicated hardware. In addition, we have clarified the dependences of the computational resources on the average user throughput.

2 Coordinated radio-resource scheduler with HWA

To ensure fairness among MTs, we devised a coordinated radio-resource scheduler with an HWA that runs a PF MT selection algorithm. As shown in Fig. 1(c), it consists of software and an HWA comprising three parts: a combination generation part, a total PF metric computation part, and a combination decision part. At the beginning of the scheduling period, the software sets the possible MTs for each antenna and average throughput of each MT in the HWA. To accelerate the processing, the HWA computes the transmission weight matrix for the generated possible combinations in parallel. To compute the PF metric of each MT, the scheduler estimates the data throughput from the computed transmission weight matrix and divides the estimated data throughput by the average data throughput. Then, it sums the PF metrics of all MTs in a generated combination (i.e., computes the total PF metric, Total.PF) as
\[
Total_{PF} = \sum_{n=1}^{N} \frac{TH_{n,m}}{Avg\_TH_m}
\]  
where \(TH_{n,m}\) denotes the computed data throughput of MT \(m\) served from antenna \(n\), and \(Avg\_TH_m\) represents the average user throughput of MT \(m\).

3 Evaluation

To examine how the required computational resources vary with the average user throughput, we carried out a system-level simulation. According to [7], the processing time to compute the metric of a combination on a scheduler with HWA having 2.9 Mgates or 5.3 Mgates is 9.2 \(\mu\)s and 1.9 \(\mu\)s, respectively. Moreover, the processing time without acceleration of 16 or 64 cores is 73 \(\mu\)s and 18 \(\mu\)s, respectively.

3.1 Simulation conditions

Table I shows the simulation conditions. We evaluated how the computational resources affect the average user throughput when using the PF and the Max C/I MT selection algorithms. Thirty-two antennas were deployed in a circle with a radius of 100 m. The density of MTs ranges from a dense urban scenario to an indoor hotspot scenario in 3GPP’s 5G use case model [8]. The approximate search described in [4] was used to find the optimal combination.

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3.2 User throughput analysis

Fig. 2(a) to (f) shows cumulative distribution functions (CDFs) of average user throughput for 64 MTs and 128 MTs when the Max C/I MT selection algorithm is
Fig. 2. CDFs of average user throughput for (a) Max $C/I$ at 64 MTs, (b) Max $C/I$ at 128 MTs, (c) PF at 32 MTs, (d) PF at 64 MTs, (e) PF at 128 MTs, (f) PF at 256 MTs, and (g) dependence of 5th percentile average user throughput on number of MTs.
used and for 32 MTs, 64 MTs, 128 MTs, and 256 MTs when PF is used. These results show that the scheduler with the HWA can obtain higher average user throughput for both algorithms regardless of the number of MTs. To discuss the fairness among MTs, we compared the 5th percentile average user throughput. In the case of the Max C/I, the 5th percentile throughput is below 20 Mbps even for the scheduler with the 5.3-Mgate HWA. In contrast, the PF scheduler with the same HWA obtains around 20 Mbps at the 5th percentile throughput even at 256 MTs.

Now, let us discuss the 5th percentile average user throughput in order to clarify the computational resources required for ensuring fairness among MTs. Fig. 2(g) shows the dependences of the 5th-percentile throughput on the number of MTs. To provide 20 Mbps at this percentile, the scheduler without acceleration from 16 or 64 cores is acceptable when the number of MTs is 32 and 64, respectively. As the number of MTs becomes larger, the scheduler needs to use the HWA in order to provide 20 Mbps at the 5th percentile. A scheduler with a 2.9-Mgate HWA suffices for 128 MTs, while one with an HWA of 5.3 M gates or more can handle 256 MTs. Thus, a system with a scheduler equipped with an HWA can accommodate several times as many MTs in the same area compared with a system that uses a scheduler without acceleration.

These simulation results show that computational resources are the critical factor in obtaining high average user throughput and ensuring fairness among MTs. An appropriate scheduler should be selected according to the system requirements.

4 Summary

We discussed the average user throughput of a coordinated radio-resource scheduler for 5G ultra-high-density distributed antenna systems. In 5G mobile communications systems, the transmission weight matrix and the PF metric for a huge number of possible combinations of antennas and MTs have to be computed in the scheduling process. To accelerate the computation, we devised a scheduler equipped with an HWA that computes the transmission weight matrix and total PF metric on dedicated hardware. System-level simulations reveal that, compared with a scheduler that does not use an HWA, this scheduler can obtain higher average user throughput for both the PF and Max C/I MT selection algorithms regardless of the number of MTs. In addition, a scheduler with a 5.3-Mgate HWA obtains around 20 Mbps at the 5th percentile of average user throughput when the number of MTs is 256. Moreover, a smaller HWA with 2.9 M gates is suitable for accommodating up to 128 MTs. Computational resources are hence the critical factor in obtaining high average user throughput and ensuring fairness among MTs.

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Low latency IR-HARQ for multiband WLAN system

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Abstract: This paper proposes a low latency incremental redundancy hybrid ARQ (IR-HARQ) scheme for a multiband wireless local area network (WLAN) system. The proposed scheme can reduce a retransmission delay and ensure backward compatibility with WLAN devices by using a clear to send (CTS) frame of a current WLAN system and rate compatible low-density parity-check (RC-LDPC) codes. Additionally, it can minimize the retransmission delay by selecting an optimal retransmission data length on the basis of the accumulated mutual information (ACMI) for available frequency bands. The results of computer simulations show that the multiband WLAN system using the proposed scheme has shorter retransmission delay and higher system throughput than the current WLAN system.

Keywords: wireless LAN, multiband, IR-HARQ, RC-LDPC, ACMI

Classification: Wireless Communication Technologies

References


1 Introduction

As wireless local area networks (WLANs) have increasingly been used, congestion and shortage of frequency resources regularly occur at 2.4 and 5 GHz. To overcome this problem, a multiband WLAN system has recently been proposed that simultaneously transmits data packets of frequency bands in the current IEEE802.11 WLAN system [1]. This system can improve the receiving performance and band occupancy time by using simultaneous transmission data packets that divide the original data into frequency bands and add a media access control (MAC) header. However, the MAC header makes the retransmission overhead in this system larger than that in the current WLAN system. Especially, when the number of retransmissions is increased by a multipath fading channel that causes an uncorrected modulation and coding scheme (MCS) selection, the larger overhead degrades system throughput.

A chase-combining hybrid ARQ (CC-HARQ) and an incremental redundancy hybrid ARQ (IR-HARQ) are generally valid approaches to improve the receiving and decoding performance in retransmission [2, 3]. In the CC-HARQ scheme, the receiving performance can be improved by combining recently received packets with the same packet received in the past. On the other hand, in the IR-HARQ scheme, the decoding performance improves as the number of retransmissions increases since some redundancy bits are retransmitted by rotation. Therefore, the IR-HARQ scheme has smaller retransmission overhead than the CC-HARQ scheme. However, the current WLAN system does not support the HARQ scheme. For the current WLAN system to support the HARQ scheme, we have proposed a low latency IR-HARQ scheme [4]. This scheme can ensure backward compatibility with WLAN devices by using a control frame of the current WLAN system and rate compatible low-density parity-check (RC-LDPC) codes that are designed on the basis of IEEE802.11n LDPC codes.

To reduce the retransmission delay in the multiband WLAN system, this paper proposes a low latency IR-HARQ scheme for the multiband WLAN system. The proposed scheme can minimize the retransmission delay by selecting an optimal retransmission data length on the basis of the accumulated mutual information (ACMI) for available frequency bands. In addition, it can ensure the above backward compatibility by adopting RC-LDPC codes and reporting the retransmission data length of the IR-HARQ scheme by using a clear to send (CTS) frame. Results of computer simulations show that the multiband WLAN system using the proposed scheme has a lower average transmission delay and higher system throughput than the current WLAN system that adopts IEEE802.11n LDPC codes [5].
2 Proposed low latency IR-HARQ scheme

This section describes the proposed low latency IR-HARQ protocol and the method for calculating the optimal retransmission data length on the basis of the ACMI for the multiband WLAN system.

2.1 IR-HARQ protocol for the multiband WLAN system

Fig. 1 shows the retransmission protocol for the conventional scheme in the current WLAN system and the proposed scheme for the multiband WLAN system. As shown in Fig. 1(a), the conventional scheme must transmit a request to send (RTS) and CTS frames before the transmission of data packets to avoid packet collision. In addition, if the retransmission user does not have the minimum random back-off time during a contention window (CW) after failing to receive a data packet, the retransmission delay increases. On the other hand, as shown in Fig. 1(b), the proposed scheme can give the retransmission packet the highest priority without a packet collision by informing all users about the retransmission by using the CTS/NACK frame. The retransmission procedure of the proposed IR-HARQ scheme is as follows.

Step 1. A source transmits the RTS frame at desired frequency bands.
Step 2. A destination transmits the CTS frame at available frequency bands that received the RTS frame.
Step 3. The source simultaneously transmits the divided data packets at frequency bands that received the CTS frame. In the following steps, all transmission and receiving processes use the same frequency bands.
Step 4. The destination combines the demodulated data sequences for the above divided packets and decodes the combined sequence after storing the packets in a HARQ buffer. If the retransmission packet is received, the destination decodes the data sequence that combines the stored data of the HARQ buffer and the demodulated data for the retransmission packet.
Step 5. The destination decides whether the result of a frame check sequence (FCS) check for received packets contains errors.
Step 6. If the result of the FCS check contains no errors, the destination transmits an acknowledgement (ACK) frame and clears the HARQ buffer.
Step 7. If the result of the FCS check contains errors, the destination transmits the CTS/NACK frame after setting a value of a network allocation vector (NAV) that is calculated from an optimal retransmission data length.
Step 8. If the source receives the ACK frame or a retransmission counter has reached an upper limit, the source finishes the retransmission process after clearing the retransmission counter.
Step 9. If the retransmission counter has not reached the upper limit, the source transmits the retransmission data packets that have the calculated retransmission data length from the NAV value of the CTS/NACK frame. After increasing the retransmission counter by one, return to Step 4.

In Step 5, if the result of the FCS check contains errors, the destination reports not only the retransmission request to the source but also a prohibition period of data
transmission to other stations by the CTS/NACK frame. As a result, the proposed scheme can quickly transmit the retransmission packet without a CW period and RTS/CTS sequence in the retransmission.

2.2 Optimization of the retransmission data length

In the IR-HARQ scheme, to minimize the retransmission delay, the retransmission data length must be minimized in accordance with the channel condition. In the multiband WLAN system, to succeed in decoding the combined sequence in the n-th transmission, the amount of ACMI, which is the left term in (1), has to be larger than the target data rate in the first transmission $R_1$.

$$\left( \sum_{m=1}^{n} N_m \sum_{f=1}^{F} \tilde{C}_{m,f} \right) > N_1 > R_1.$$  \hspace{1cm} (1)

Therefore, if the decoding of the n-th transmission fails, the proposed scheme can minimize the retransmission delay by selecting the minimum value for the next transmission data length $N_{n+1}$ that satisfies (2).

$$N_{n+1} \sum_{f=1}^{F} \tilde{C}_{n+1,f} > N_1 R_1 - \sum_{m=1}^{n} N_m \sum_{f=1}^{F} \tilde{C}_{m,f},$$  \hspace{1cm} (2)

$$\tilde{C}_{m,f} = \frac{1}{N_{sc,f}} \sum_{i=1}^{N_{sc,f}} \left( L - \int_{v \in \mathbb{C}} \frac{e^{-|v|^2}}{\pi 2^L} \sum_{x \in \mathbb{C}} \sum_{i=1}^{L} \log_2 A_{i,x,f}(SNR_{k,f})dv \right),$$  \hspace{1cm} (3)

where $N_m$ is the number of data symbols for the $2^L$-ary modulation scheme $\chi$ of the n-th transmission, $N_{sc,f}$ is the number of subcarriers at the f-th frequency band, and C is the complex field. If the bit-interleaved coded modulation (BICM) is used.

Fig. 1. Retransmission protocols: (a) conventional, (b) proposal.
in the $m$-th transmission, the capacity $\hat{C}_{m,f}$ is given by (3) and (4). For other coding scheme besides the BICM, the capacity is given by (3) and (5), which approximates (4) by the signal to noise ratio (SNR) offset $\alpha$.

$$A_{v,x,i}(SNR_{k,f}) = 1 + \left( \sum_{B_{i}(x)=1} e^{-\frac{|x-i|}{\sqrt{SNR_{k,f}+\alpha}}} \right) \left( \sum_{B_{i}(x)=0} e^{-\frac{|x-i|}{\sqrt{SNR_{k,f}+\alpha}}} \right), \quad (4)$$

$$A_{v,x,i}(SNR_{k,f}) = 1 + \left( \sum_{B_{i}(x)=1} e^{-\frac{|x-i|}{\sqrt{SNR_{k,f}+\alpha+v}}} \right) \left( \sum_{B_{i}(x)=0} e^{-\frac{|x-i|}{\sqrt{SNR_{k,f}+\alpha+v}}} \right), \quad (5)$$

where $B_{i}(x)$ is the $i$-th bit of the symbol $x$ and $SNR_{k,f}$ is the estimated SNR in the $k$-th subcarrier at the $f$-th frequency band to accommodate frequency selective fading.

### 3 Performance evaluation

This section evaluates the average transmission delay and the system throughput for the conventional and proposed schemes. In this letter, the conventional scheme (Conventional) is the retransmission scheme in Fig. 1(a) with IEEE802.11n LDPC codes, in which the coding rate is supported from 0.5 to 0.833. In contrast, the proposed scheme (Proposal) is the retransmission scheme in Fig. 1(b) with designed RC-LDPC codes [4], in which the coding rate is supported from 0.5 to 0.875. In addition, Conventional and Proposal calculate the retransmission data length by (4) and (5) with $\alpha = 1.0\sim1.4$ [4], respectively. This evaluation adopts a layered belief propagation (LBP) decoding [6] that uses 15 iterations; QPSK, 16QAM, and 64QAM modulations; the IEEE802.11a standard frame format with 20MHz bandwidth; the packet payload of 157∼882 bytes; and a 6-path exponential decay model with 2 dB decay. The maximum number of retransmission is four.

Fig. 2 shows the transmitter and receiver (TRX) block diagram for the multi-band WLAN system with the proposed IR-HARQ scheme. In this system, the demodulated data sequence for received data packets is decoded after the packets are combined and stored in the HARQ combiner. After the transmission of the CTS/NACK or ACK frame is requested from the HARQ controller at the destination in accordance with the result of the FCS check, the CTS/NACK or ACK frame is transmitted to the source. If the source receives the CTS/NACK frame, after the HARQ controller in the source reports the optimal retransmission data length that is calculated from the NAV value of the CTS/NACK frame to the encoder in the source, the additional parity bits are transmitted to the destination.

**Fig. 2.** TRX block diagram for multiband WLAN system with proposed IR-HARQ scheme.
Fig. 3 shows the average transmission delays and system throughputs for Conventional and Proposal with an adaptive modulation and coding scheme (AMC) at 2.4 and 5.7 GHz. In this figure, the X-axis is the average SNRs at 5.7 and 2.4 GHz bands. Additionally, Y-axis is the average transmission delay, i.e., the average time from when a packet is generated until it is successfully received or the retransmission limit is reached. As shown in Fig. 3, Proposal reduces the average transmission delay by up to 33% and increases the system throughput by 150% on average compared with Conventional. The reason for the improved system throughput at the average SNR < 10 dB is the improved average transmission delay. On the other hand, the reason for the improved system throughput at the average SNR > 10 dB is the required SNR that improved by the IR-HARQ scheme.

4 Conclusion

This paper proposed a low latency incremental redundancy hybrid ARQ (IR-HARQ) scheme for a multiband wireless local area network (WLAN) system and evaluated the average transmission delays and system throughputs for the proposed scheme and the current WLAN system. The results of computer simulations showed the multiband WLAN system using the proposed scheme has up to 33% lower average transmission delay and on average 150% higher system throughput than the current WLAN system.

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Adaptive MMSE-SVD for OFDM downlink MU-MIMO in a high mobility environment

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Abstract: In this paper, we propose an adaptive minimum mean square error filtering combined with singular value decomposition (MMSE-SVD) for OFDM downlink multi-user multi-input multi-output (MU-MIMO) transmission. During data transmission, base station (BS) updates the multi-user MMSE transmit filter by using linear channel prediction, while user equipments (UEs) update their eigenmode receive filters, constructed by SVD, by using decision feedback adaptive channel estimation. The uncoded BER performance achievable by adaptive MMSE-SVD is evaluated by computer simulation. It is shown that proposed adaptive MMSE-SVD can increase an allowable maximum Doppler frequency \(f_D T\) for keeping BER < 10^{-2} by about 4 times.

Keywords: MU-MIMO, MMSE-SVD, decision feedback channel estimation, linear prediction

Classification: Wireless Communication Technologies

References

1 Introduction

In the 5th generation (5G) networks, broader data services and higher link capacity than 4G networks are required [1]. A promising approach under the limited radio bandwidth is multi-user multi-input multi-output (MU-MIMO) [2]. MU-MIMO can further improve the spectrum efficiency without bandwidth expansion. Recently, the authors proposed a minimum mean square error filtering combined with singular value decomposition (MMSE-SVD) for spatially multiplexing user equipments (UEs) [3]. For performing MMSE-SVD, the MIMO channel state information (CSI) must be shared by the base station (BS) and UEs to construct the transmit and receive filters prior to the data transmission. Assuming the time-division duplex (TDD), BS and UEs can share the MIMO CSI without feedback.

Recently, we proposed a TDD subframe structure which simplifies the MIMO CSI sharing. The proposed TDD subframe consists of uplink pilot slot, downlink pilot slot, and 12 user data slots as shown in Fig. 1 [4]. Firstly, UEs, to be spatially multiplexed, transmit the frequency-division multiplexed (FDM) uplink pilot. Then, BS transmits the FDM downlink pilot. By doing so and exploiting channel reciprocity, both BS and UEs sides can share the MIMO CSI between transmit antennas and UEs’ receive antennas prior to data transmission without feedback. However, in a high mobility environment, MIMO CSI acquired by using uplink and downlink pilots will become outdated during data transmission period. In this paper, we propose an adaptive MMSE-SVD for OFDM downlink. In the adaptive MMSE-SVD, the BS updates the multi-user MMSE transmit filter using linear channel prediction while UEs update their eigen-mode reception filters, constructed by SVD, using decision feedback adaptive channel estimation.

2 Adaptive MMSE-SVD for high mobility environment

(a) Principle of MMSE-SVD

Multi-user spatial multiplexing of \( U \) UEs, each is equipped \( N_{\text{ue}} \) antennas, is considered. BS simultaneously transmits \( N_{\text{strm}} \) data streams for each UE (therefore, a total number of streams becomes \( U \cdot N_{\text{strm}} \)) from \( N_{\text{bs}} \) transmit antennas using \( N_{c} \) subcarriers. In OFDM downlink, the transmit and receive filter matrices for MMSE-SVD are respectively expressed as [3]

![Fig. 1. Subframe structure (UL pilot + DL pilot + 12 data slots)](image-url)
b) Updating the MMSE transmit receiver structure of adaptive MMSE-SVD. Fig. 2 illustrates a transmitter degradation in a high mobility environment, in this paper, we propose an adaptive filter and the actual propagation MIMO channel. To avoid the BER degradation in a high mobility environment, in this paper, we propose an adaptive MMSE-SVD for high mobility environment. By substituting the received signal vector of the previous slot into eq. (1), W\text{mmse}(k) is obtained by applying SVD to H_u(k) as

\[
\text{H}_{ul}(k) = \text{diag}(\text{H}_0(k), \cdots, \text{H}_u(k), \cdots, \text{H}_{U-1}(k)),
\]

where A_u(k) is the N_{strm} × N_{strm} eigenmode diagonal matrix. U^H(k)H(k) is the equivalent channel when each UE applies eigenmode reception (i.e. UE uses U^H(k) as the receive filter matrix). P(k) = diag[P_0(k), \cdots, P_u(k), \cdots, P_{U-1}(k)], and P_u(k) of size N_{strm} × N_{strm} represents the water filling based power allocation [5] across eigenmodes and subcarriers. BS estimates MIMO CSI \hat{H}_{bs}(k; t = 0) using uplink pilot signal at time slot t = 0, and UEs estimate each MIMO CSI \hat{H}_{ue, u}(k; t = 1) by using downlink pilot signal at time slot t = 1 [4]. Accordingly, BS constructs the transmit filter W_{mmse}(k; t = 0) and UEs construct the receive filter W_{svd, u}(k; t = 1) by using \hat{H}_{bs}(k; t = 0) and \hat{H}_{ue, u}(k; t = 1) instead of H(k) in eq. (1), respectively. In a high mobility environment, when W_{mmse}(k; t = 0) and W_{svd, u}(k; t = 1) are used continuously during data transmission (i.e. t = 2~13), BER performance degrades significantly due to the filter mismatch among the transmit filter, the receive filter and the actual propagation MIMO channel. To avoid the BER degradation in a high mobility environment, in this paper, we propose an adaptive MMSE-SVD for high mobility environment. Fig. 2 illustrates a transmitter/receiver structure of adaptive MMSE-SVD.

(c) Updating the MMSE transmit filter

BS applies a linear prediction to obtain the channel estimate \hat{H}_{bs}(k; t) at \(t = 2\sim13\) using \(\hat{H}_{bs}(k; t = 0)\) and \(\hat{H}_{bs}(k; t = -N_{slot})\) (which is the channel estimate using the uplink pilot in previous subframe (\(N_{slot} = 14\) slots)) as

\[
\hat{H}_{bs}(k; t) = \hat{H}_{bs}(k; t = 0) + \frac{\hat{H}_{bs}(k; t = 0) - \hat{H}_{bs}(k; t = -N_{slot})}{N_{slot}} \times t.
\]

By substituting \(\hat{H}_{bs}(k; t)\) into eq. (1), \(W_{mmse}(k; t)\) is updated.

(c) Updating the eigenmode reception filter

Each UE estimates the equivalent channel \(H_{ue, u}(k; t) = H_u(k; t)W_{mmse, u}(k; t)\) by applying decision feedback channel estimation. Assuming the time-varying of fading over consecutive slots is small, an N_{ue} × N_{strm} matrix \(\hat{R}_u(k; t)\) representing the received signal vector of the previous N_{strm} time slots is expressed as

\[
\hat{R}_u(k; t) = [R_u(k; t - N_{strm}), \cdots, R_u(k; t - 1)]
\]

\[
= \sqrt{\frac{2E_s}{T_s}} H_u(k; t) W_{mmse, u}(k; t) \bar{D}_u(k; t)
\]

\[
+ \sqrt{\frac{2E_s}{T_s}} \sum_{t' = 0}^{U-1} H_u(k; t) W_{mmse, u'}(k; t) \bar{D}_{u'}(k; t) + \hat{N}_u(k; t)
\]
where $ \mathbf{R}_u(k;t)$ is an $N_{ue} \times 1$ receive signal vector of the $u$th UE. $ \hat{\mathbf{D}}_u(k;t) = [\mathbf{D}_u(k;t-N_{strm}), \ldots, \mathbf{D}_u(k;t-1)]$ is an $N_{strm} \times N_{strm}$ transmitted data symbol matrix consisting of the transmitted data symbol vectors of previous $N_{strm}$ time slots and $ \hat{\mathbf{N}}_u(k;t) = [\mathbf{N}_u(k;t-N_{strm}), \ldots, \mathbf{N}_u(k;t-1)]$ is an $N_{ue} \times N_{strm}$ noise matrix. The estimated equivalent channel $ \hat{\mathbf{H}}_{eq,u}(k;t)$ is obtained by multiplying inverse matrix of decision feedback data symbol matrix $ \hat{\mathbf{D}}_u(k;t) = [\mathbf{D}_u(k;t-N_{strm}), \ldots, \mathbf{D}_u(k;t-1)]$ to the right side of eq. (4) as

$$
\hat{\mathbf{H}}_{eq,u}(k;t) = \begin{cases} 
\mathbf{R}_u(k;t) \hat{\mathbf{D}}_u^{-1}(k;t) & \text{rank}(\hat{\mathbf{D}}_u(k;t)) = N_{strm} \\
\hat{\mathbf{H}}_{eq,u}(k;t-1) & \text{rank}(\hat{\mathbf{D}}_u(k;t)) < N_{strm}.
\end{cases} \tag{5}
$$

Then, frequency-domain moving average filtering is applied to $ \hat{\mathbf{H}}_{eq,u}(k;t)$, yielding

$$
\check{\mathbf{H}}_{eq,u}(k;t) = \frac{1}{Q} \sum_{q=-Q/2}^{Q/2} \hat{\mathbf{H}}_{eq,u}(k+q;t). \tag{6}
$$

Each UE generates the receive filter using equivalent channel $ \check{\mathbf{H}}_u(k)$ as

$$
\mathbf{W}_{svd,u}(k;t) = (\check{\mathbf{H}}_{eq,u}(k;t))^H \left( \check{\mathbf{H}}_{eq,u}(k;t)(\check{\mathbf{H}}_{eq,u}(k;t))^H + \left( \frac{E_x}{N_0} \right)^{-1} \mathbf{I}_{N_{ue}} \right)^{-1}. \tag{7}
$$

The updated filter $ \mathbf{W}_{svd,u}(k;t)$ has smaller filter mismatch than $ \mathbf{W}_{svd,u}(k;t-1)$, but it could produce the noise enhancement. Then, each UE selects the receive filter as

$$
\mathbf{W}_{svd,u}(k;t) = \arg \min_{\mathbf{W}_{svd,u}(k';t) \in \{\mathbf{W}_{svd,u}(k';1), \ldots, \mathbf{W}_{svd,u}(k';t)\}} \left\| \mathbf{W}_{svd,u}(k';t) \mathbf{R}_u(k;t) - \mathbf{D}_u \right\|, \tag{8}
$$

where $\| \cdot \|$ represents the Euclidean norm and $ \mathbf{D}_u \in \Psi_{mod}^{N_{strm} \times 1}$ is an $N_{strm} \times 1$ vector consisting of $N_{strm}$ candidate symbols in the modulation constellation set $\psi_{mod}$.

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**Fig. 2.** Transmitter/receiver structure of adaptive MMSE-SVD
Finally, each UE obtains the received signal vector $\hat{\mathbf{d}}_u(k; t) = \hat{\mathbf{W}}_{\text{svd},u}(k; t) \mathbf{R}_u(k; t)$ to perform symbol decision.

### 3 Monte-Carlo computer simulation

We evaluate the uncoded BER performance of adaptive MMSE-SVD by computer simulation. $U = 2$ UEs having $N_{\text{ue}} = 2$ antennas are spatially multiplexed. BS simultaneously transmits $N_{\text{term}} = 2$ data streams for each UE from $N_{\text{bs}} = 4$ transmit antennas. Each data stream consists of $N_c = 128$ data-modulated symbols, where the modulation is assumed to be 16QAM. Pilot is generated using Zadoff-Chu sequence. Assuming that the channel is composed of $L = 16$ distinct paths, the transfer function $H_u(k; n_{\text{ue}}, n_{\text{mbs}})$ between $n_{\text{ue}}$th antenna of $u$th UE in macro-cell and $n_{\text{mbs}}$th transmit antennas can be represented as

$$H_u(k; n_{\text{ue}}, n_{\text{mbs}}) = \sum_{l=0}^{L-1} \tilde{\xi}_{u,n_{\text{ue}},n_{\text{mbs}}}(l) \exp\left(-j \frac{2\pi k r_{u,n_{\text{ue}},n_{\text{mbs}}}(l)}{N_c}\right),$$

where $\tilde{\xi}_{u,n_{\text{ue}},n_{\text{mbs}}}(l)$ and $r_{u,n_{\text{ue}},n_{\text{mbs}}}$ are respectively the complex-valued path gain and the time delay of the $l$th path with $E[\sum_{l=0}^{L-1} |\tilde{\xi}_{u,n_{\text{ue}},n_{\text{mbs}}}(l)|^2] = 1$ for all $u$, $n_{\text{ue}}$, $n_{\text{mbs}}$. We assume a sample-spaced time delay (i.e., $r_{u,n_{\text{ue}},n_{\text{mbs}}} = l$ for all $u$, $n_{\text{ue}}$, $n_{\text{mbs}}$).

Fig. 3(a) plots average uncoded BER as a function of average transmit $E_s/N_0$ with normalized maximum Doppler frequency ($f_D T$) as a parameter, where $T$ represents the slot length. It is seen from Fig. 3, BER degradation and error floor occur in the case of conventional MMSE-SVD without transmit/receive filter updating due to the filter mismatch among the transmit filter, the receive filter and the actual propagation MIMO channel. On the other hand, our proposed adaptive MMSE-SVD with transmit/receive filter updating can reduce the error floor.

Fig. 3(b) plots the average uncoded BER as a function of $f_D T$. BER performance of the conventional MMSE-SVD without transmit/receive filter updating degrades BER performance when $f_D T$ is higher than 0.0001. Below, we will discuss the performance improvement achievable if either one of transmit and receive filters is updated. If the receive filter is updated only, almost no performance improvement is obtained, i.e., the allowable $f_D T$ for keeping the BER below $BER < 10^{-2}$ is 0.001, which is almost the same as using the conventional MMSE-SVD. This is because the receive filter updating has no effect to reduce the inter-user interference (IUI) while it can mitigate the inter-antenna interference (IAI). If the transmit filter is updated only, the BER performance improvement obtained and accordingly, the allowable $f_D T$ can be increased to 0.003. This performance improvement is because the transmit filter updating can mitigate both IAI and IUI. On the other hand, if updating both the transmit and the receive filters, this is our proposed adaptive MMSE-SVD, updating the transmit filter mitigates IUI and IAI and updating the receive filter mitigate IAI. Therefore, a further performance improvement is obtained and the allowable $f_D T$ increases to 0.004. Note that $f_D T = 0.001 (0.004)$ corresponds to the velocity of 14.4 km/h (57.6 km/h) when assuming 5 GHz carrier frequency and subcarrier spacing of 75 kHz.
In this paper, we proposed an adaptive MMSE-SVD for OFDM downlink. In the adaptive MMSE-SVD during data transmission, MBS updates transmit filter using linear channel prediction while each UE updates receive filter using decision feedback channel estimation. It is shown by computer simulation that proposed adaptive multi-user MMSE-SVD can increase the allowable $f_D T$ for keeping the BER below $BER < 10^{-2}$ about 4 times. Applying the adaptive MMSE-SVD to the uplink is our future work.

4 Conclusion

In this paper, we proposed an adaptive MMSE-SVD for OFDM downlink. In the adaptive MMSE-SVD during data transmission, MBS updates transmit filter using linear channel prediction while each UE updates receive filter using decision feedback channel estimation. It is shown by computer simulation that proposed adaptive multi-user MMSE-SVD can increase the allowable $f_D T$ for keeping the BER below $BER < 10^{-2}$ about 4 times. Applying the adaptive MMSE-SVD to the uplink is our future work.

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Comparative analysis of various AC/coaxial adapters for LISN calibration up to 1 GHz

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Abstract: There are various types of AC/coaxial adapter for line impedance stabilization network (LISN) calibration because the AC plug used differs among countries. The difference in the characteristics of the adapters may be large at a high frequency such as 1 GHz; thus, their characteristics should be compared. In this paper, the transmission characteristics of one-phase AC/coaxial adapters for Japan (A-type) and UK and some Asian countries (BF and B3-types, respectively) are compared using their equivalent circuit models. The following two results are obtained: (1) the BF-type most strongly affects the LISN calibration, and (2) the transmission coefficient of the A-type is the least affected by the reference ground.

Keywords: AC/coaxial adapter, calibration, equivalent circuit, line impedance stabilization network (LISN), transmission line theory

Classification: Electromagnetic Compatibility (EMC)

References

1 Introduction

A line impedance stabilization network (LISN) is used for conducted disturbance measurement on mains lines, and its upper limit frequency is 30 MHz [1]. In recent years, however, the LISN for measurements up to 1 GHz has been studied because measurements above 30 MHz have also been required owing to the electromagnetic interference over a wide frequency range from kilohertz to gigahertz caused by green electronics [2]. LISN calibration is important to obtain accurate disturbance measurement results. As an example of the calibration, an LISN impedance measurement setup is shown in Fig. 1. In the measurement, the AC/coaxial adapter shown in the figure is needed to connect a vector network analyzer to the AC outlet of the LISN. Since the measured LISN characteristics include the adapter characteristics, the latter should be evaluated and eliminated from the measured LISN characteristics for the accurate LISN calibration.

Although various studies on the evaluation of adapters up to 30 MHz have been reported [3, 4], the evaluation above 30 MHz has hardly been carried out. Moreover, there are various types of adapter because the AC plug used differs among countries, but their characteristics have hardly been compared. The characteristics differ for each adapter, and the difference in these characteristics may become large with increasing frequency; thus, their comparison at high frequencies is important. The authors proposed an equivalent circuit model of an adapter up to 1 GHz [5]. The proposed model treats the one-phase AC plug for Japan as the transmission line model because it consists of three-conductors with a uniform cross-section. The AC plugs used in other countries also consist of conductors with a uniform cross-section as with the Japanese AC plug; therefore, the proposed model is useful for the characterization of various adapters.

In this paper, the transmission characteristics of three one-phase adapters are compared using the proposed model.
2 Equivalent circuit model of AC/coaxial adapter

Fig. 2(a) shows the structure of an AC/coaxial adapter. This adapter consists of two coaxial connectors, an AC plug, and a metal plate with two strip lines. The strip lines embedded in the metal plate connect the neutral/phase lines (#1 and #2) and each inner connector. The protective earth (PE) line (#3) connects the outer conductors of the connectors via the metal plate. Fig. 2(b) shows the dimensions of three AC plugs. As an example, the three adapters for Japan (A-type) and UK and some Asian countries (BF- and B3-types, respectively) are chosen. Here, \( h \) is defined as the height between the PE line and the reference ground (GND). Fig. 2(c) shows the equivalent circuit of the adapter [5]. This equivalent circuit is the cascade arrangement of three circuits: the coaxial connector, the metal plate, and the AC plug. The AC plug is represented as the three-conductor transmission lines above the GND, in which the currents flowing in the neutral/phase lines return to the PE line. The circuit of the metal plate consists of the two strip lines and the two capacitors \( C_{N}^{M} \) and \( C_{P}^{M} \) respectively expressing the coupling between the neutral/phase lines and the metal plate. In this paper, the equivalent circuit is considered as a lossless.
3 Comparison of transmission characteristics

3.1 Evaluation method

To obtain the matrix of the $ABCD$ parameters for the adapter $F^{\text{adapter}}$, the matrices of the $ABCD$ parameters for the coaxial connector, the metal plate, and the AC plug, namely, $F^c$, $F^M$, and $F^p$, respectively, are derived [5]. The matrix of the $ABCD$ parameters for the transmission line can be derived by the state variable method [6]. When $x^p$ is the line length of the AC plug, $F^p$ is given by

$$F^p = \exp \left( j\omega x^p \begin{bmatrix} O & L^p \\ C^p & O \end{bmatrix} \right)$$

(1)

where $O$ is the square zero matrix; $L^p$ and $C^p$ are the per-unit-length inductance and capacitance matrices of the AC plug given by the following symmetric matrices:

$$L^p = \begin{bmatrix} L^p_{11} & L^p_{12} \\ L^p_{21} & L^p_{22} \end{bmatrix}, \quad C^p = \begin{bmatrix} C^p_{11} + C^p_{21} & -C^p_{21} \\ -C^p_{21} & C^p_{22} + C^p_{21} \end{bmatrix}$$

(2)

with the self- and mutual inductances $L_{ij}^p$ and the capacitances $C_{ij}^p$. In the same manner, $F^c$ is given by
where \( x_c \) is the line length of the coaxial connector; \( L^c \) and \( C^c \) are the per-unit-length inductance and capacitance matrices of the two coaxial connectors:

\[
L^c = \begin{bmatrix} L^c & 0 \\ 0 & L^c \end{bmatrix}, \quad C^c = \begin{bmatrix} C^c & 0 \\ 0 & C^c \end{bmatrix}.
\] (4)

Moreover, \( F^M \) can be obtained as the following equation from the circuit of the metal plate shown in Fig. 2(b):

\[
F^M = \exp \left( j \omega x_s \right) \begin{bmatrix} O & L_M^s \\ C_s^M & O \end{bmatrix} \begin{bmatrix} U & 0 \\ 0 & U \end{bmatrix}.
\] (5)

Here, \( U \) is the unit matrix; \( x_s \) is the line length of the strip line; \( L_M^s \) and \( C_s^M \) are the per-unit-length inductance and capacitance matrices of the two strip lines; \( C^M \) is the matrix of the two capacitors \( C_N^M \) and \( C_P^M \). \( L_s^M \), \( C_s \), and \( C^M \) are given by

\[
L_s^M = \begin{bmatrix} L_s^M & 0 \\ 0 & L_s^M \end{bmatrix}, \quad C_s = \begin{bmatrix} C_s^M & 0 \\ 0 & C_s^M \end{bmatrix}, \quad C^M = \begin{bmatrix} C_N^M & 0 \\ 0 & C_P^M \end{bmatrix}.
\] (6)

Therefore, \( F_{\text{adapter}} \) is obtained using the following matrix:

\[
F_{\text{adapter}} = F^c \cdot F^M \cdot F^p.
\] (7)

Finally, the matrix of the \( S \)-parameters for them when the reference impedance is 50 ohm is derived from \( F_{\text{adapter}} \).

3.2 Evaluation result

Fig. 3(a) shows the calculated transmission characteristics of the three AC/coaxial adapters at \( h = 10 \text{ mm} \). Here, the circuit parameters of the coaxial connector and the strip line are selected so that their characteristics impedance is 50 ohms. The two capacitances \( C_N^M \) and \( C_P^M \) between the neutral/phase lines and the metal plate are 1 pF, as an example, and the circuit parameters of the three AC plugs are obtained using the circuit parameter extraction simulator Q3D Extractor [7]. Since the three transmission coefficients \( |S_{31}| \) remain at 0 dB up to 100 MHz, the three adapters hardly affect the LISN calibration below 100 MHz. In contrast, above 100 MHz, their coefficients decrease with increasing frequency, so the three adapters seriously affect the LISN calibration. Moreover, it has been found that these decreases differ for each adapter, where that of the BF-type is the largest. Therefore, the BF-type the most strongly affects the LISN calibration.

The impact of the GND on the adapters is also investigated. Fig. 3(b) shows the deviations in each transmission coefficient at \( h = 10 \text{ mm} \) from the coefficient of each adapter without the GND. Here, a large deviation value means that the GND strongly affects the coefficient. As shown in the figure, since the deviation of the A-type is the smallest, it has been found that the A-type is the least affected by the GND.
4 Conclusion

In this paper, the transmission characteristics of three one-phase AC/coaxial adapters for Japan (A-type) and UK and some Asian countries (BF and B3-types, respectively) were compared using their equivalent circuit models formed by the transmission line theory. It was found that the three adapters affected the LISN calibration above 100 MHz. It was also found that the impact of the calibration differs for each adapter and that the BF-type most strongly affects the calibration. Moreover, the impact of the reference ground on the adapters was investigated. The result indicated that the A-type is the least affected by the ground. In future works, the impact of the difference in these characteristics on the calibration will be evaluated.

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Direct technique for measuring chromatic dispersion of high-order mode in a few-mode fiber

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Abstract: We propose an interferometric technique for directly measuring the chromatic dispersion of the high-order mode in a few-mode fiber (FMF) by using a mode converter (MC). We successfully estimate the chromatic dispersion of the LP\(_{11}\) mode in the test fiber by using the proposed technique.

Keywords: interferometer, chromatic dispersion, few-mode fiber, mode converter

Classification: Fiber-Optic Transmission for Communications

References


1 Introduction

Mode division multiplexing (MDM) has been developed intensively by using few-mode fibers (FMF) to expand their transmission capacity [1, 2]. It is known that differential group delay (DGD) between modes, their chromatic dispersions, and optical loss are important parameters when designing MDM systems.

Many reports have been published on techniques for measuring the dispersion characteristics of FMF, including optical frequency-domain reflectometry [3], phase-sensitive optical low-coherence reflectometry [4], and modal interferometer method [5]. According to [5], the chromatic dispersion in the LP_{11} mode can be estimated by measuring the differential group delay between the LP_{01} and the LP_{11} modes in conjunction with the group delay time in the LP_{01} mode.

In this paper, we propose an interferometric technique for directly measuring the chromatic dispersion of the high-order mode in an FMF by using a mode converter (MC). We successfully estimate chromatic dispersion of the LP_{11} mode in the test fiber by using this technique. This technique can be applied to chromatic dispersion measurements of high-order modes in FMFs by using a suitable MC, if the mode coupling between modes is ignored.

2 Measurement principle

Fig. 1 shows a diagram of an interferometer for measuring group delay/transit time difference between two optical paths in an optical fiber and a wave propagating in the air [6]. Before the polarizer, P_{1}, an MC is inserted to convert the LP_{01} mode to the high-order mode. The interferometer facilitates measurement of the wavelength dependence of DGD and transit time delay of the high-order mode.

The beam of high-order mode emitted from the MC is divided into two beams by using a half mirror HM_{1}. One of these beams is launched into the test FMF, while the other is propagated through air. The direction of the beam in air is changed by the mirrors M_{1} and M_{2}, and it is superposed on the beam emitted from the FMF by another half mirror HM_{2}. The moving mirror M_{3} is adjusted so that the two optical paths can be made identical, which corresponds to \( d = 0 \), where \( 2d \) is the group time delay of the path difference between the two optical paths. The polarizer P_{2} is adjusted to achieve the maximum dip in the optical spectrum analyzer (OSA) trace. Here, DGD is defined as \( \Delta \tau = \tau_h - \tau_{air} \), where \( \tau_h \) and \( \tau_{air} \) are the group time delays of the high-order mode and the wave propagating in the air, respectively.
The DGD $\Delta \tau$ between the two optical paths is expressed as follows [5]:

$$\Delta \tau = \tau_h - \tau_{air} = -\lambda^2/(cL\Delta \lambda) + 2d/(cL)$$  \hspace{1cm} (1)

where $c$ is the velocity of light in free space, $L$ is the fiber length, $\lambda$ is the center wavelength of the adjacent minima, and $\Delta \lambda$ is the difference between the wavelengths of the adjacent minima in the trace.

Therefore, the chromatic dispersion $D_h$ of the high-order mode can be obtained by differentiating $\tau_h$ with respect to the wavelength as follows:

$$D_h = d\tau_h/d\lambda$$  \hspace{1cm} (2)

$D_h$ can be estimated experimentally by differentiating the wavelength dependence, $\tau_h - \tau_{air}$, with respect to wavelength, as shown in (2). Chromatic dispersion in the LP$_{01}$ mode can be estimated experimentally in the same way by launching the LP$_{01}$ mode into the test FMF through MC and P$_1$.

3 Experimental results

We used a super-luminescent diode (SLD) operating at a center wavelength of 1540 nm as the light source in our experiments. We prepared the test FMF with chromatic dispersion of LP$_{11}$ mode of 17.3 ps/nm/km at 1.55 µm, which was calculated taking the index profile into account. The MC (Optoquest Co., Ltd.) was used to covert the LP$_{01}$ mode to the LP$_{11}$ mode.

Fig. 2 shows the interference trace between the LP$_{11}$ mode in the FMF and the wave in air obtained using 0.78 m long FMF. The interference fringe was clearly visible in the trace. We observed the field pattern of the light wave emitted from the test FMFs by using an infrared vidicon camera (Hamamatsu Photonics), as shown in the inset of Fig. 2. The field pattern coincides with that of the LP$_{11}$ mode. We confirmed that the spectrum shown in Fig. 2 was the interference between the LP$_{11}$ mode in the FMF and the wave traveling in air.

The transit time delay difference $\tau_{11} - \tau_{air}$ can be obtained from Fig. 2 by using Eq. (1). Fig. 3(a) shows the DGD properties of the test fiber. The open circles represent the experimental results obtained using the test fiber. The solid line shows the best fit of the experimental results by using a quadratic function. The chromatic dispersion of the LP$_{11}$ mode can be computed using Eq. (2) and Fig. 3(a). Therefore, the chromatic dispersion can be estimated by differentiating the fitting curve.
shown in Fig. 3(a), with respect to wavelength. Fig. 3(b) shows the chromatic
dispersion plotted as a function of wavelength.

The red line shows the chromatic dispersion of the LP$_{11}$ mode measured using
the technique proposed herein. The blue line shows the calculated chromatic
dispersion plotted as a function of wavelength, which was calculated taking the
index profile into account. The chromatic dispersion of the test fiber differs slightly
from the calculated value.

Fig. 2. OSA trace obtained for transit time delay difference $\tau_{11} - \tau_{air}$

Fig. 3. (a) DGD as a function of wavelength. (b) Chromatic dispersion
of the LP$_{11}$ mode in the test FMF.
Discrepancies between the two sets of chromatic dispersion values obtained at 1.55 µm are shown in Fig. 3(b). The intensity distribution shown in Fig. 2 suggests that these discrepancies arise from ambiguity with respect to the determination of the positions of interference signal peaks.

As a result, we confirmed that the proposed technique can be used directly to measure the chromatic dispersion of the high-order mode, as well as the fundamental mode, in FMFs given an MC that can convert the LP_{01} mode to the desired mode.

4 Conclusion

We proposed an interferometric technique for measuring the chromatic dispersion of the high-order mode of a FMF by using an MC. The chromatic dispersion of the LP_{11} mode in the test FMF was successfully measured using this technique. It was confirmed that the proposed technique can be applied to direct measurement of the chromatic dispersion of the high-order mode in FMFs, given mode converters that can convert the LP_{01} mode to the desired mode.

Acknowledgment

This research is partially supported by the National Institute of Information and Communications Technology (NICT), Japan, under “Research on Innovative Optical Fiber Technology.”
Specific absorption rate and temperature increase in pregnant women at 13, 18, and 26 weeks of gestation due to electromagnetic wave radiation from a smartphone

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Abstract: Nowadays, mobile communication technology has become essential; thus, it is important to consider the effect of electromagnetic radiation from mobile terminals such as smartphones on the human body. In this study, we calculated the specific absorption rate (SAR) and temperature increase in pregnant women exposed to a smartphone’s radiation at 13, 18, and 26 weeks of gestation. The results indicate that the SARs were much lower in fetuses than in pregnant women under all of the calculation settings in this study. Moreover, the maximum temperature increases in fetuses were half of those in pregnant women.

Keywords: specific absorption rate, temperature increase, smartphone

Classification: Electromagnetic Compatibility (EMC)

References


1 Introduction

With the continuous expansion of mobile communication networks and the development of new mobile terminals such as smartphones and tablet PCs, people’s lives have improved. We have previously used mobile terminals for voice calls. However, we frequently used novel mobile terminals not only for voice calling but also for data communication. If pregnant women perform data communications near the abdomens using the novel mobile terminals, the distance between the fetuses and the terminals become very close. In this kind of environment, the numerical dosimetry for fetuses during pregnancy is one of the most important issues in electromagnetic (EM) field safety [1, 2]. Tateno et al. consider SARs and temperature rises in bodies during pregnancy radiated by simplified EM source such as a dipole antenna [3]. Chiaramello et al. evaluating SARs in a fetus body radiated by tablet PC, but they did not consider the temperature increase [4]. Therefore, in this study, we estimate SARs and temperature increase by using high-resolution computational smartphone model.

2 Calculation models

In this study, we used computational whole-body pregnant woman models for gestational ages of 13, 18, and 26 weeks [5, 6], and a computational smartphone model [7]. Fig. 1(a) shows the pregnant woman model at 26 weeks of gestation. The pregnant woman models are composed of 56 types of tissues and organs. The abdominal dimensions of the models for each week of pregnancy are adjusted to those of average Japanese pregnant women. Masses of fetal tissues are also adjusted to anatomical reference data [8]. Fig. 1(b) shows the smartphone model.
that was developed on the basis of an actual smartphone sold in Japan. The validity of the smartphone model was confirmed by comparing the SAR distributions between the smartphone model and the actual equipment [7].

The calculation model is shown in Fig. 1(c), (d). The smartphone model was placed horizontally at the front of the navel of the pregnant woman model. The shortest distance between the surface of the body and the edge of the smartphone was fixed at 10 mm. A perpendicular bisector of the short side of the smartphone is orthogonally crossed to the midsagittal plane of the body. Operating frequencies of the smartphone were 900 MHz and 2 GHz, and the maximum radiated power was fixed at 0.25 W in consideration of the maximum power of the third-generation mobile communication system. We calculated SARs and temperature increases in the calculation models using XFDTD ver. 7.5 (Remcom Inc., State College, PA, USA). The calculation region was divided into approximately 200 million cuboids varying from 0.2 to 1 mm for each edge. The tissue properties of the pregnant woman models were described by Tateno et al. [3]

3 Results and discussion

The peak 10-g-averaged SARs in the pregnant women models are shown together with the SARs in non-pregnant woman model [7] in Table I(a). The SARs were higher at 900 MHz than at 2 GHz in the fetuses, although the SARs in the pregnant women were higher at 2 GHz. The reason for this difference can be explained by penetration depth of the EM waves. For example, penetration depth of the muscle at 900 MHz and 2 GHz are 42 and 26 mm, respectively. Thus, the penetration depths of the anatomical tissues at 2 GHz are shallower than those at 900 MHz. Therefore, EM energy at 2 GHz were mainly absorbed in the maternal body, whereas the EM energy at 900 MHz could reach the fetal body more than 2 GHz.

The SARs in the pregnant women increased with the progress of pregnancy regardless of the operating frequencies of the smartphone. On the other hand, the SARs in the fetuses showed tendencies different from those in the pregnant women, and they were the highest at the gestational age of 18 weeks. In this study, the placement of the smartphone was determined on the basis of the navel of the pregnant women. In the case of 18 weeks of gestation, the distance between the smartphone and the fetus was the closest. Therefore, the SARs in the fetuses are the highest at 18 weeks of gestation. However, the highest fetal SAR was much lower (approximately up to one-fourth) than the maternal SAR under the same configuration and operating frequency. We also confirmed that the SARs in pregnant women and fetuses were much lower than the basic restriction of the general public recommended by the International Commission on Non-Ionizing Radiation Protection (ICNIRP) [9].

We compared the 10-g-averaged SARs of this study with the results by simplified sources, namely, dipole antennas and planar inverted-F antennas (PIFAs) with a metallic case [3]. The peak maternal SARs due to EM radiation from the dipole antennas and the PIFAs were 4.70 and 1.66 W/kg, respectively. Moreover, the peak fetal SARs were approximately 1.88 and 0.83 W/kg, respectively. In the previous study [3], the positions of the wave sources were determined in relation to
the position of the fetal head. The peak 10-g-averaged SARs when the smartphone model was placed in the same condition as [3] are shown in Table I(b). The SARs in the pregnant women and the fetuses for the smartphone radiation were the highest at 26 weeks of gestation, like the results for simplified sources. Moreover, the SARs in the pregnant women and the fetuses for the smartphone were much lower than those for the simplified sources. This is because of the difference in the near field pattern between the simplified sources and the smartphone.

Table I(c) shows peak temperature increases in the pregnant women and the fetuses. The temperature increases in the fetuses as well as the SAR results were highest at 18 weeks of gestation, and the maximum increase was 0.017 K. Therefore, the temperature increases in the fetuses were significantly less than temperature increase (1 K) which is known to cause adverse health effects such as growth retardation for the fetuses of many animals [10]. We also confirmed that the temperature increases in the fetuses were less than half of those in the pregnant women.

Fig. 2 shows the distributions of temperature increases on the midsagittal plane of the pregnant women at 18 and 26 weeks of gestation due to EM exposure from the smartphone operated at 2 GHz. We confirmed that the temperature increases
were suppressed in the placenta due to a large amount of blood flow. It is suggested that the placenta provides a strong cooling effect for both the maternal body and the fetus. Therefore, when we evaluate the temperature increases in a fetus, the positional relationships among the smartphone, the placenta, and the fetus are of considerable importance. We also confirmed that the temperature increases in

Table I. Peak values of SAR and temperature increase in pregnant women and fetsuses

(a) Peak 10-g-averaged SAR [W/kg] in the case:
   smartphone was placed at the front of the maternal navel

<table>
<thead>
<tr>
<th>@ Mother (@ Fetus)</th>
<th>Non-pregnant [7]</th>
<th>13 weeks</th>
<th>18 weeks</th>
<th>26 weeks</th>
</tr>
</thead>
<tbody>
<tr>
<td>900 MHz</td>
<td>0.12</td>
<td>0.12</td>
<td>0.13</td>
<td>0.15</td>
</tr>
<tr>
<td></td>
<td>(−)</td>
<td>(0.004)</td>
<td>(0.035)</td>
<td>(0.012)</td>
</tr>
<tr>
<td>2 GHz</td>
<td>0.19</td>
<td>0.26</td>
<td>0.22</td>
<td>0.33</td>
</tr>
<tr>
<td></td>
<td>(−)</td>
<td>(0.001)</td>
<td>(0.025)</td>
<td>(0.006)</td>
</tr>
</tbody>
</table>

(b) Peak 10-g-averaged SAR [W/kg] in the case:
   smartphone was placed in the same condition as [3]

<table>
<thead>
<tr>
<th>@ Mother (@ Fetus)</th>
<th>13 weeks</th>
<th>18 weeks</th>
<th>26 weeks</th>
</tr>
</thead>
<tbody>
<tr>
<td>900 MHz</td>
<td>0.15</td>
<td>0.11</td>
<td>0.35</td>
</tr>
<tr>
<td></td>
<td>(0.011)</td>
<td>(0.024)</td>
<td>(0.051)</td>
</tr>
<tr>
<td>2 GHz</td>
<td>0.18</td>
<td>0.18</td>
<td>0.41</td>
</tr>
<tr>
<td></td>
<td>(0.005)</td>
<td>(0.018)</td>
<td>(0.060)</td>
</tr>
</tbody>
</table>

(c) Peak temperature increase [K] in the case:
   smartphone was placed at the front of the maternal navel

<table>
<thead>
<tr>
<th>@ Mother (@ Fetus)</th>
<th>13 weeks</th>
<th>18 weeks</th>
<th>26 weeks</th>
</tr>
</thead>
<tbody>
<tr>
<td>900 MHz</td>
<td>0.025</td>
<td>0.030</td>
<td>0.017</td>
</tr>
<tr>
<td></td>
<td>(0.002)</td>
<td>(0.013)</td>
<td>(0.006)</td>
</tr>
<tr>
<td>2 GHz</td>
<td>0.053</td>
<td>0.057</td>
<td>0.043</td>
</tr>
<tr>
<td></td>
<td>(0.001)</td>
<td>(0.017)</td>
<td>(0.007)</td>
</tr>
</tbody>
</table>

Fig. 2. Distributions of temperature increase on the midsagittal plane of pregnant woman models.
pregnant women and fetuses were much lower than the value (1 K) known to cause possibly adverse effects.

4 Conclusions

In this study, we estimated the SARs and temperature increases in the pregnant women and their fetuses when they use a smartphone close to the abdomen. Results showed that the 10-g-averaged SAR in the maternal body increased with gestational progress. The SARs in the fetus changed depending on the positional relationship between the smartphone and the fetus, and they were highest at 18 weeks of gestation in this study. Moreover, we confirmed that the temperature increases in the maternal and fetal bodies were higher at 18 weeks of gestation than at 26 weeks of gestation, because the placenta provided a strong cooling effect for both the maternal body and the fetus. This indicates that the positional relationship among the smartphone, the placenta, and the fetus is of considerable importance. We also confirmed that the SARs for the smartphone radiation were the much lower than those for the simplified sources radiation.

Finally, we confirmed that the 10-g-averaged SARs in pregnant women were much lower than the basic restriction of the general public recommended by the ICNIRP. We also found that the temperature increases in fetuses were significantly less than the value (1 K) known to cause possibly adverse effects.

Acknowledgments

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Study of relationship between physiological index and quality of experience for video streaming service

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Abstract: While Quality of Service (QoS) is a service quality indicator that is used as objective evaluation by service providers, Quality of Experience (QoE) is a quality indicator subjectively evaluated by users. However, subjective evaluation of QoE for every service imposes the users a great load. Therefore, it is needed to estimate QoE from a physiological index without every subjective evaluation. In this research, we focus on skin conductance activity (SCA) as a physiological index and develop an experimental scheme to clarify the relationship between QoE and SCA when abrupt playback pause degrades the video quality. As the results of experiment and data analysis, QoE and SCA are correlated differently according to the video categories such as sport and news.

Keywords: video streaming, Quality of Experience, subjective quality evaluation, physiological indices, skin conductance

Classification: Multimedia Systems for Communications

References


1 Introduction

In these days, video streaming services have become one of the most popular Internet services. Owing to development of smartphones and tablets as well as wireless environments, users experience the video streaming service even in a mobile environment, where communication conditions are somewhat unstable. Therefore, the service providers try to keep the service quality as good as possible. A great amount of effort has been invested in subjective evaluation of Quality-of-Service (QoS) for video services [1]. The QoS parameters of video services are usually the frame size (resolution), the frame rate, and the quantization scale. In the mobile environment, however, abrupt playback pauses possibly occur as more critical service degradation. Such abrupt service degradation should be evaluated based on Quality-of-Experience (QoE) instead of QoS. Recently, QoE is focused as a subjective metric to reflect user experience not only for video services [2] but also for Web services [3], cloud services [4], and so on.

The generic method for QoE evaluation is Mean Opinion Score (MOS), which is the average of values that subjects assign their individual opinion based on the predefined scale [5]. MOS cannot, however, record instantaneous evaluation of the users. Moreover, MOS cannot capture the user’s evaluation over time. Namely, MOS is a static evaluation and time-varying video quality is not evaluated. Recently, instead of MOS, an advanced approach using physiological measures is being watched with intense interest to evaluate the video streaming services [6]. Particularly, electroencephalogram (EEG) is often used to evaluate multimedia quality [6, 7].

In this paper, Electro Dermal Activity (EDA) is used as a physiological index, because the measurement device is smaller compared with EEG and it is more suitable for wearable devices. EDA is categorized into Skin Potential Activity (SPA) and Skin Conductance Activity (SCA). SCA, which is used in our experiments, changes depending on skin sweating and is suitable to measure the degree of psychological stress.

We present a novel approach to evaluate the video streaming service quality from the viewpoints of QoE and SCA signal simultaneously. Then analytical models are proposed for both measured data. It is clarified that QoE and SCA
are well-correlated and significantly effected by two factors: service quality degradation and video category.

2 Experimental conditions

19 university students (9 men and 10 women) participated in the experiment as subjects. Each subject was asked to watch a video content. During watching the video content, the subjective QoE was measured by Single Stimulus Continuous Quality Evaluation (SSCQE) method and SCA was measured by the small electrodes attached to the subject’s index and middle fingers continuously.

SSCQE method is defined by International Telecommunication Union-Radio-communication Sector (ITU-R) [8]. In SSCQE method, a video sequence is shown to a subject and the subject needs to evaluate the video quality instantaneously by continuously adjusting a slider. It is recommended that the slider should have the scale range between 0 and 100, which shows that larger the scale, the higher the evaluation. In our experiments, the slider of 10 cm length was implemented in a tablet with an 8-inch touch screen monitor. The subjects manipulated the slider by his/her dominant hand during watching the video sequence and the sampling frequency was 10 Hz.

SCA was measured by a combination of MP150, GSR100C, and TSD203 of BIOPAC Systems Inc. The electrodes were attached to the subject’s non-dominant hand and the sampling frequency was 1000 Hz, which was down-sampled to 10 Hz corresponding to SSCQE method. The measurement methods were synchronized for QoE and SCA.

Two categories of video contents were used in the experiments: sport and news. Each original video content is 2 minutes long and has the resolution of $1280 \times 720$. The original content is called as Video N. As the quality degradation, abrupt playback pauses were added to Video N. Three types of pause time were used according to the degree of quality degradation, that is 5, 10, and 15 seconds and they are called asVideos $D_1$, $D_2$, and $D_3$ in order. Each subject watched 8 videos, namely a set of Videos $N$, $D_1$, $D_2$, and $D_3$ from two video categories, but the order of video presentation was random. During watching a video, the subject recorded QoE by using the slider according to SSCQE method and his/her SCA signal was measured simultaneously.

3 Experimental results and analysis

The synchronized data of QoE evaluation and measured SCR signals were obtained for each video content from 19 subjects. First, the data for the quality-degraded videos should be modeled for further analysis. Fig. 1 shows the models of QoE values and SCA signals.

In the QoE model shown in Fig. 1(a), the playback pause period is presented as Quality Degradation Section. QoE evaluation deterioration delays than the video quality degradation, because perception in brains is needed before moving the slider. Therefore, we define the QoE evaluation for the degraded quality as the value recorded at the end of Quality Degradation Section.
On the other hand, the SCA signal is modeled as Fig. 1(b). Since stress causes sweating, the skin conductance increases in general. Such a general signal pattern is obtained during the period of Quality Degradation Section from all of the measured data. Thus, two inflection points are detected; one is the begging point of the skin conductance increasing \( t(L) \) and the other is the maximum point of the skin conductance increasing \( t(P) \). Then Value of Stress (\( \text{VoS} \)) is defined as follows,

\[
\text{VoS} = \frac{SC_t(P)}{SC_t(L)},
\]

where \( SC_t \) is the value of the skin conductance at time \( t \).

\( \text{VoS} \) is, however, the actual measured value and the amplitude of \( \text{VoS} \) differs from each subject since physiological reaction depends on the individual. Moreover, each subject watched 3 resolution-degraded videos from two video categories. Therefore, 12 \( \text{VoS} \) values for the quality-degraded videos were collected from one subject and these values are normalized to mitigate individual difference as follows,

\[
\text{DoS} = \frac{\text{VoS} - \text{VoS}_{\text{min}}}{\text{VoS}_{\text{max}} - \text{VoS}_{\text{min}}},
\]

where \( \text{VoS}_{\text{max}} \) and \( \text{VoS}_{\text{min}} \) are the maximum and minimum values among 14 \( \text{VoS} \) values and \( \text{DoS} \) means the degree of Stress.

According to the above definitions, \( \text{QoE} \) and \( \text{DoS} \) in the period of Quality Degradation Section are obtained for each subject. Hereafter, the measured \( \text{QoE} \) values by SSCQE method are described as \( \text{QoE} \). \( \text{QoE} \) is the normalized value in the same way shown in (2). These data are averaged for each quality-degraded video,
that is $D_1$, $D_2$, or $D_3$, and for each video category, that is sport or news. Then the relationship between QoE and DoS is depicted in Fig. 2. The correlation coefficients shown in Fig. 2 are $-0.835$ for the sport videos and $-0.987$ for the news videos. In addition, since the total correlation coefficient is $-0.878$, negative correlation is observed between QoE and DoS.

As shown in Fig. 2, both QoE and DoS tend to be influenced by the video category or the playback pause period. Therefore, we apply the two-way analysis of variance (ANOVA) to the data of QoE and DoS respectively, where the first factor is the video category and the second one is the pause period (the video degradation). The null hypotheses are set as follows:

H1: The population means of the first factor are equal.
H2: The population means of the second factor are equal.
H3: There is no interaction between the two factors.

The results of ANOVA are shown in Table I, where (a) corresponds to the result for QoE and (b) does to the one for DoS. In Table I, SS stands for the sum of squares, DF does the degrees of freedom, and MS does the Mean square.

From the results of ANOVA, first of all, The null hypothesis H3 is not rejected even for the significance level of 0.1 (10.0%). Secondary, on the other hand, the null hypotheses H1 and H2 are rejected for the significance level of 0.05 (5.0%).

![Fig. 2. Relationship between QoE and DoS.](image)

**Table I.** The results of the two-way analysis of variance

<table>
<thead>
<tr>
<th>Factor</th>
<th>SS</th>
<th>DF</th>
<th>MS</th>
<th>F ratio</th>
<th>p value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Video category</td>
<td>0.769</td>
<td>1</td>
<td>0.769</td>
<td>12.496</td>
<td>0.002</td>
</tr>
<tr>
<td>Pause period</td>
<td>0.508</td>
<td>2</td>
<td>0.254</td>
<td>6.358</td>
<td>0.0043</td>
</tr>
<tr>
<td>Interaction</td>
<td>0.153</td>
<td>2</td>
<td>0.076</td>
<td>1.936</td>
<td>0.1591</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Factor</th>
<th>SS</th>
<th>DF</th>
<th>MS</th>
<th>F ratio</th>
<th>p value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Video category</td>
<td>0.858</td>
<td>1</td>
<td>0.858</td>
<td>8.116</td>
<td>0.0107</td>
</tr>
<tr>
<td>Pause period</td>
<td>1.237</td>
<td>2</td>
<td>0.618</td>
<td>4.891</td>
<td>0.0132</td>
</tr>
<tr>
<td>Interaction</td>
<td>0.036</td>
<td>2</td>
<td>0.018</td>
<td>0.266</td>
<td>0.7712</td>
</tr>
</tbody>
</table>
That is to say, both of QoE and DoS are significantly different for the video category as well as for the pause period. Moreover, the null hypotheses H1 and H2 for QoE can be rejected up to the significance level of 0.005 (0.5%). Finally, it is suggested that QoE can be estimated from the measured DoS data, since these are correlated. Also, the levels of measurement can be differentiated by the video category and the playback pause period.

4 Conclusion

We developed a novel experimental scheme to record QoE and SCA data for video streaming services simultaneously. The SCA data is a physiological index which is normally unconscious, while QoE is a subjective evaluation and recorded as conscious. The proposed experimental scheme can be applied to evaluate the abrupt playback pause of video streaming. We examined three kinds of playback pause period for two video categories: sport and news. After the measured QoE and DoS data are modeled, it has been clarified that they are correlated and differentiated by the video category and the playback pause period as the results of ANOVA analysis.
Dual-axis monopulse direction-of-arrival estimation planar antenna employing multilayer structure

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\textsuperscript{2}Faculty of Electrical and Computer Engineering, Chittagong University of Engineering and Technology, Chittagong–4349, Bangladesh
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Abstract: A novel multilayer structure of a direction-of-arrival (DOA) estimation antenna is proposed in this paper. This antenna provides dual-axis DOA estimation, while conventional monopulse DOA estimation antennas determine arrival angle in a single plane. The proposed antenna has a multilayer structure and consists of four ring-slot antennas and four magic-Ts. The proposed antenna provides both $xz$- and $yz$-plane dual-axis DOA estimation by using the monopulse mechanism.

Keywords: dual-axis, DOA estimation, multilayer, ring slot, magic-T

Classification: Antennas and Propagation

References

1 Introduction

In wireless communication systems, direction-of-arrival (DOA) estimation plays a vital role in order to achieve better reception of incoming signals. The research area in the field of DOA estimation has become wider in the last six decades. It has been applied in various applications including wireless communications, radar, etc. The role of the DOA estimation in wireless communications is essential since it helps to estimate the direction of the incoming signal. The estimated signal direction will be used to point the array beam towards the direction. The DOA estimation of the RF signals has been an interesting area of research as it offers several interesting benefits in terms of improved Quality-of-Service, such as better coverage, more reliable communication and higher data rates [1]. Furthermore, the DOA information can also be used for positioning or localization in a wireless cellular network.

Planar DOA estimation antennas using magic-Ts were already proposed [2, 3, 4, 5]. These DOA estimation antennas can determine the angle of arrival in a single plane by employing the phase monopulse mechanism.

In this paper, a new dual-axis DOA estimation antenna which can calculate the angle of arrival in two planes is proposed. The basic concept has already discussed in [6]. The proposed antenna employs a multilayer structure and four annular ring-slot antennas are used as antenna elements. The proposed antenna contains both sided microwave integrated feedline circuits including magic-Ts [7]. The ability to measure the direction of arrival waves in dual planes makes the proposed antenna unique.

2 Operating principle

Fig. 1(a) illustrates the basic block diagram of the proposed antenna. Three antenna elements are arranged along the \( x \)- and \( y \)-axis with the antenna separation of \( d_x \) and \( d_y \). The antenna elements along the \( x \)- and \( y \)-axis connect to Magic-T1 and -T2, respectively. As the magic-T provides in-phase or anti-phase power combination according to the input and output ports, the sum (\( \Sigma \)) and difference (\( \Delta \)) of the signals received by the two antenna elements are separately obtained [8]. Therefore, the sum and difference signals in the \( x \)-axis, i.e. \( \Sigma_{xz} \) and \( \Delta_{xz} \) are obtained from Magic-T1. Similarly, \( \Sigma_{yz} \) and \( \Delta_{yz} \) are obtained from Magic-T2. The arrival angle of the received signals can be determined by the amplitude of the sum (\( \Sigma \)) and difference (\( \Delta \)) of the two received signals using the monopulse mechanism.

Fig. 1(b) shows the coordinate system where \( \theta_{xz} \) and \( \theta_{yz} \) are defined as arrival angles in the \( xz \)-plane and \( yz \)-plane, respectively. In case of the dual-axis DOA estimation, the arrival angle in \( xz \)-plane, \( \theta_{xz} \), and \( yz \)-plane, \( \theta_{yz} \), can be expressed by following equations:
where $\lambda$ is the wavelength. $\Sigma_{xz}$ and $\Delta_{xz}$ are the $\Sigma$ and $\Delta$ signals in $xz$-plane and $\Sigma_{yz}$ and $\Delta_{yz}$ are the $\Sigma$ and $\Delta$ signals in $yz$-plane, respectively. Therefore the proposed array antenna provides DOA estimation in both $xz$-plane and $yz$-plane by using the monopulse mechanism.

### 3 Antenna structure

Fig. 2(a) shows the schematic layout of the proposed multilayer dual-axis DOA estimation antenna. The proposed antenna comprises four ring-slot antenna elements and four magic-Ts. The complete structure of the antenna is designed in three metal layers. The ring-slot antennas are formed in Layer 2 and the feed networks are distributed on two different layers (i.e., Layer 1 and Layer 3). Two separate feed networks are designed in two separate planes so that the arrival angle estimation operation can be obtained in two orthogonal planes (i.e., $xz$-plane and $yz$-plane). One of the feed networks with Port 1 and 3, which respectively provide $\Sigma$...
and $\Delta$ signals in the $xz$-plane is arranged on Layer 1. Layer 3 consists of another feed network with Port 2 and 4, which provide $\Sigma$ and $\Delta$ signals in the $yz$-plane, respectively. Two slot lines arranged orthogonally are also formed with four slot rings in Layer 2. In this structure, Layer 1 detects only horizontal polarization ($x$-polarization) while Layer 3 detects only vertical polarization ($y$-polarization). Therefore, the arriving waves have to have both $x$- and $y$-polarization to properly estimate the arrival angle in the dual-axis. In case of circular polarizations, this antenna works perfectly.

### 3.1 Prototype antenna

Fig. 2(b) shows the photograph of the fabricated 5.8-GHz dual-axis DOA estimation antenna in three different layers. $\Sigma$ and $\Delta$ signals of the $xz$-plane from Port 1 and Port 3 are obtained at Layer 1, and Port 2 and Port 4 at Layer 3 provide $\Sigma$ and $\Delta$ signals of the $yz$-plane, respectively. In this design, a Teflon fiber ($\varepsilon_r = 2.15$, thickness = 0.8 mm) is used as a substrate material. The size of the prototype
antenna is 88 × 88 mm. Each ring-slot antenna has an outer radius 7 mm and inner radius 6 mm resulting a slot width of 1 mm.

4 Results and discussion

Fig. 3(a) shows the measured reflection coefficient plots of Port 1, 2, 3 and 4 of the designed antenna. Better than 10-dB return loss is observed at 5.64 GHz for all four ports.

Fig. 3(b) illustrates the measured radiation patterns of the $\Sigma$ and $\Delta$ signals for the $xz$- and $yz$-plane. The measured gain of this antenna is 5.32 dBi for $\Sigma_{xz}$ and 6.84 dBi for $\Sigma_{yz}$. The arrival angles in $xz$- and $yz$-plane can be determined using the values of the $\Sigma$ and $\Delta$ with eqs. (1) and (2). The concept of the dual-axis DOA estimation can be confirmed by the experimental result.

5 Conclusion

A multilayer structure of a DOA estimation antenna was discussed in this paper. The proposed antenna can determine the arrival angles of the received signals in two orthogonal planes. The antenna has been fabricated and the performance has been measured to confirm the proposed DOA estimation concept. It is found that
the antenna can determine the arrival angles in the $xz$- and $yz$-planes according to the theory.

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Feasibility evaluations of three-dimensional-printed high-gain reflectarray antenna for W-band applications

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Abstract: An evaluation reveals the feasibility of a high-gain and low-fabrication-cost reflectarray antenna for W-band millimeter-wave radar applications. This reflectarray antenna offers the advantage of a simple, flat structure, and can be fabricated at low cost by three-dimensional (3D) printing. This paper describes the fabrication of a 3D-printed reflectarray antenna, whose gain is more than 30 dBi in the W-band. Firstly, the reflection-signal phases of dielectric plates of different thicknesses are evaluated using finite-difference time-domain (FDTD) analysis to obtain the characteristics of the reflectarray design. Then, an eight-zone Fresnel reflectarray, in which a phase difference of 45° separated each zone, is analyzed and fabricated. The results of the FDTD analysis shows a 31.6 dBi antenna gain at 78.5 GHz. Finally, the designed reflectarray is fabricated using a 3D printer. Measurements indicate that the achieved 30.4 dBi antenna gain nearly equaled the FDTD analysis value of 31.6 dBi at 78.5 GHz.

Keywords: acrylonitrile butadiene styrene plastic, finite-difference time-domain method, millimeter-wave radar, reflectarray antenna, three-dimensional printer, W-band

Classification: Antennas and Propagation

References


1 Introduction

Reflectarray antennas offer the advantages of a simple, small-volume, and flat structure [1, 2, 3]. We have been developing a millimeter-wave radar system for civil aviation uses, including helicopter collision-avoidance onboard radar [4]. A reflectarray antenna with a quasioptical approach is one of the most important options supporting high-performance radar systems usable for millimeter-wave radar applications [3]. In addition, we are interested in low-cost fabrication methods that maintain high performance [5]. Three-dimensional (3D) printers can reduce the cost of fabrication [2]. However, as far as we know, reflectarray antennas produced in this fashion are difficult to demonstrate high-gain characteristics above 30 dBi in the W-band.

This paper discusses the feasibility evaluations of a high-gain reflectarray antenna for W-band millimeter-wave radar applications. Our discussion confirms that 3D-printed fabrication produces a reflectarray antenna with gains exceeding 30 dBi. Firstly, the reflection-signal phases of dielectric plates of varying thicknesses are investigated numerically in the W-band using the finite-difference time-domain (FDTD) method. Next, an eight-zone reflectarray is designed to operate at 78.5 GHz and then analyzed to evaluate the antenna characteristics. Finally, the reflectarray antenna is fabricated with a 3D printer. Then, the antenna characteristics are evaluated in an anechoic chamber.

2 Analysis of the reflection-signal phase of dielectric plates

The reflection-signal phase must be determined to finalize the design of a reflectarray antenna. The thickness of the dielectric plate is the major design parameter for a 3D-printed reflectarray antenna. An FDTD analysis of the dielectric plate is carried out to obtain the quantitative value of the phase shift of the dielectric material.

Fig. 1(a) shows the FDTD analysis model used to obtain the reflection-signal phases. These reflection-signal phases of the plane-incident wave are analyzed for different dielectric-plate thicknesses. If one assumes an infinite periodical space, the top and bottom walls each consist of perfect electric conductor (PEC). In addition, each side wall consists of a perfect magnetic conductor (PMC). Also, the PEC is attached to the back of the dielectric plate. Commercially available FDTD analysis
Software (SEMCAD X, Schmid & Partner Engineering AG, Zürich, Switzerland) is employed. Fig. 1(b) shows the FDTD analysis parameters. Fabrication of the reflectarray with a 3D printer employs the dielectric constants of acrylonitrile butadiene styrene (ABS) plastics. The relative permittivity and loss tangent are 2.3 and 0.1, respectively.

Fig. 1(c) shows the analyzed reflection-signal relative phase shift for the different dielectric-plate thicknesses at 73.5 GHz, 78.5 GHz, and 83.5 GHz. In this case, the amounts of the phase shift are relative values compared without any dielectric materials. The thickness of the 0.1 mm step is analyzed. Results confirmed that the phase shift is proportional to the frequency. Because the center frequency of the reflectarray is 78.5 GHz, a 360° phase shift is obtained at a dielectric-plate thickness of approximately 3.3 mm.

3 Design and analysis of the reflectarray antenna

The W-band reflectarray antenna is designed using the analysis results of the dielectric-plate reflection phase. The reflectarray design is derived from the Fresnel equation.
\[ r_n = \sqrt{\frac{2nf\lambda}{P} + \left(\frac{n\lambda}{P}\right)^2}, \tag{1} \]

where \( r_n \) and \( f \) are the \( n \)th radius and focal length, respectively. In addition, \( P \) and \( \lambda \) are the number of Fresnel zones and wavelength of the incident wave, respectively.

A comparison with the conventional reflectarray fabricated by the metallic patch on the dielectric substrate [3] shows that the number of Fresnel zones \( P \) and focal length \( f \) were 8 and 75 mm, respectively. Then, the outer diameter of the reflectarray was 154 mm. The designed center frequency was 78.5 GHz.

The FDTD analysis results described in the previous section support determining the thickness of each zone to configure an 8-zone reflectarray in which phase shifts were 0°, 45°, 90°, 135°, 180°, 225°, 270°, and 315° at 78.5 GHz. The required thickness of the dielectric plate is obtained from Fig. 1(a). For example, the dielectric thicknesses for phase shifts of 0°, 90°, and 180° are 3.3 mm, 2.8 mm, and 1.7 mm, respectively. The dimensions of the reflectarray are 154 mm x 154 mm x 3.3 mm.

Fig. 2(a) shows the numerical model of the designed eight-zone reflectarray antenna with a WR-10 waveguide antenna serving as an primary source. Then, the antenna characteristics are analyzed by FDTD analysis. The constants of the

![Fig. 2. Analysis of 3D-printed reflectarray antenna in the W-band. (a) Eight-zone reflectarray antenna analysis model. (b) Analyzed \( \gamma z \)-plane electric-field strength at 78.5 GHz. (c) Analyzed radiation patterns at 78.5 GHz.](image-url)
Reflection array dielectric material are the same as in Fig. 1(b). In addition, a PEC is attached behind the reflectarray. Fig. 2(b) and 2(c) show the analyzed yz-plane electric-field strength and radiation patterns of the reflectarray antenna at 78.5 GHz. The radiation conditions for the reflectarray surface are confirmed by the electric-field strength. In addition, the maximum antenna gain is found to be 31.6 dBi. The half-power beamwidth (HPBW) measures for azimuth and elevation are 1.8° and 1.6°, respectively.

4 Fabrication and measurement of the reflectarray antenna

The designed reflectarray antenna was fabricated using a commercially available 3D printer (Afinia H800 3D Printer, Afinia 3D, Chanhassen, MN). Fig. 3(a) shows the printing process of the eight-zone reflectarray. The printing resolution of the vertical axis is 0.1 mm using an ABS plastics filament. In addition, the total printing time is 9 hours for the designed 3.3-mm-thick reflectarray. Because the ABS plastics filament is one of the most common materials for 3D printing, the cost of the used filament is less than a few US dollars.

Then, Fig. 3(b) shows the measurement setup of the fabricated reflectarray antenna for the W-band. The surface of the reflectarray do not receive any processing, such as smoothing or other postprinting corrections. The 0.1 mm aluminum tape is attached to the back of the reflectarray. The primary source of
the antenna is the WR-10 open-ended waveguide, which is the same used in the
analysis. In addition, the focal length of the antenna can be adjusted at the
micrometer level to clarify the sensitivity depending on the radiation point.

Fig. 3(c) shows the analyzed and measured radiation patterns for the azimuth
plane at 78.5 GHz. The measured radiation characteristics at focal lengths of 75 mm
and 77 mm are shown. The focal length of the original design is 75 mm. In addition,
the 77 mm focal length produce the maximum measured antenna gain. The
measured antenna gains for different focal lengths appear in Fig. 3(d). The antenna
gain increased as the focal length increased. The maximum antenna gain is obtained
at a 77 mm focal length. The FDTD analysis shows a 31.6 dBi antenna gain; the
measured antenna gains of the 75 mm and 77 mm focal lengths are 27.9 dBi and
30.4 dBi, respectively. This 2 mm difference in focal lengths is attributable mainly
to the differences in the material constant across the reflectarray. The 2.3 dielectric
constant of the ABS plastics in the analysis is slightly higher than the dielectric
constant of the ABS plastics filament employed in the fabrication. The HPBW of
the measured azimuth radiation patterns is 1.7° for the focal lengths of 75 mm and
77 mm. On the other hand, the analyzed HPBW was 1.6°. The demonstrated
antenna characteristics of the fabricated reflectarray antenna agree well with the
FDTD analysis results. In addition, the effectiveness and validity of the analysis are
confirmed by the measurement.

The reflectarray antenna based on the printed substrate showed a maximum
gain of 33 dBi at 72 GHz, which is shifted from the gain at 78.5 GHz [3]. The
proposed 3D-printed reflectarray antenna also achieved a maximum antenna gain
greater than 30 dBi. These measurements confirm the feasibility of the 3D-printed
millimeter wave antenna exhibiting both high gain and low cost.

5 Conclusion

3D printing of a W-band reflectarray antenna was proposed and demonstrated to be
a practical method for producing a high-gain and low-cost millimeter-wave antenna
that could be used in radar systems.

Firstly, the reflection-signal phase shifts, attributable to different dielectric-plate
thicknesses, were investigated using FDTD analysis that assumed the use of
ABS plastics. Secondly, a 154-mm-diameter eight-zone reflectarray antenna was
designed using the analysis results of the reflection phase. The results of the
reflectarray antenna analysis showed an antenna gain greater than 30 dBi at
78.5 GHz. Then, the reflectarray antenna was fabricated using a commercially
available 3D printer and common ABS plastic filaments. Finally, the measured
antenna characteristics showed a 30.4 dBi antenna gain at 78.5 GHz. The feasibility
of producing high-gain performance above 30 dBi with low-cost fabrication was
confirmed by these results.

A conformal antenna, produced by a 3D printer at 0.1 mm resolution, will be
investigated next to obtain additional high-gain characteristics.