Experimental evaluation of fault recovery methods in Elastic Lambda Aggregation Network

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Abstract: In Elastic Lambda Aggregation Network (E\(\lambda\)AN), fault recovery using an Optical Line Terminal (OLT) in a different Central Office (CO) is applicable due to virtualization of OLT and active Optical Distribution Network (ODN). In addition, the Time Division Multiple Access (TDMA)-based OLT sharing method has been proposed to maintain minimum connectivity of the excessive number of Optical Network Units (ONU) from a single OLT. In this letter, we report the experiments of these two fault recovery methods implemented on the software-based E\(\lambda\)AN prototype system. The required time from failure detection to connection recovery was evaluated in these experiments.

Keywords: optical access network, aggregation network, disaster recovery

Classification: Network

References


1 Introduction

With the popularization of Fiber To The Home (FTTH) in recent years, the Internet has become an essential infrastructure for our life. In today’s optical access networks, Passive Optical Network (PON) systems are widely deployed as a cost-efficient FTTH solution. In the PON system, a single Optical Line Terminal (OLT) and multiple Optical Network Units (ONU) are connected by using optical fiber cables and optical splitters. In the commercialized PON systems, several protection methods that deploy backup optical fiber cables, OLTS, and ONUs have been standardized to cope with a network failure. However, multiple failures on these components can happen simultaneously in a time of disaster, such as a major earthquake. The protection method becomes unable to operate if main and backup components fail at the same time.

Elastic Lambda Aggregation Network (EλAN) has been studied as an access/aggregation integrated network that realizes high availability [1, 2]. In EλAN, optical access paths that have flexible bandwidth are set on active Optical Distribution Network (ODN) to connect OLTS and ONUs. ODN consists of all-optical devices such as Bandwidth Variable Wavelength Cross Connects (BV-WXC), and provides various topologies such as point-to-point path, point-to-multipoint tree, and ring.

In addition, OLTS and ONUs in EλAN have programmability in order to support multiple functions for various network services. Logical OLTS (L-OLT) and ONUs (L-ONU) are configured within Programmable OLTS (P-OLT) and ONUs (P-ONU), respectively. The migration method of L-OLT between P-OLTS has been proposed to achieve more flexible utilization of P-OLTS [3, 4, 5, 6]. When a failure occurs in a P-OLT or optical devices on an access path, connectivity of P-ONUs can be restored by migrating a L-OLT to another P-OLT and reconfiguring the access path. Hereinafter, this fault recovery method is referred to as the “inter-OLT restoration”. Network Management System (NMS) performs coordinated control of P-OLTS and ODN. Specifically, NMS manages resources of P-OLTS (i.e., hardware and software for realizing L-OLTS), calculates routes and bandwidth of access paths for each network service, and reconfigures optical devices on ODN.
Generally, there is a limit on the number of ONU s that a single OLT can accommodate (e.g., 256 ONUs in E\(\lambda\)AN) due to limitations of buffer memory, link IDs, and so on. In a case of massive failure, there is a possibility that the number of surviving OL Ts is very limited and some ONUs cannot be restored even if these ONUs are available and have reachability to the OL Ts. To maintain connectivity of excessive ONUs from a single OLT in E\(\lambda\)AN, the TDMA-based OLT sharing method has been proposed [7, 8]. This method enables an OLT to maintain minimum connectivity of more ONUs in a time of massive failure.

In this letter, we report the experiments of the inter-OLT restoration and the TDMA-based OLT sharing method by using the software-based E\(\lambda\)AN prototype system.

### 2 Operation of fault recovery methods

In this section, we overview the inter-OLT restoration and the TDMA-based OLT sharing method in E\(\lambda\)AN.

#### 2.1 Inter-OLT restoration method

As mentioned in the previous section, the inter-OLT restoration method is applicable in E\(\lambda\)AN due to active ODN and L-OLT migration.

When a central office (CO) containing multiple P-OLTs becomes incapacitated, P-ONUs connecting with these P-OLTs become unable to receive network services. To restore the connectivity of P-ONUs, NMS calculates the new placement of L-OLTs in available P-OLTs and determines if the access paths can be configured on ODN. If the access paths can be configured successfully, NMS executes L-OLT migration so that the P-ONUs receive the network services via a P-OLT in another CO. If the access paths cannot be configured, NMS iterates the calculation of L-OLT placement and the request of access path configuration until the solution is found.

#### 2.2 TDMA-based OLT sharing method

NMS determines to apply the TDMA-based OLT sharing method when the appropriate placement of L-OLTs to restore the connectivity of all P-ONUs cannot be found by the inter-OLT restoration method. P-ONUs are divided into several ONU groups, each of which includes equal or less than 256 P-ONUs. Timeslots to communicate with the L-OLT are allocated to each ONU group in rotation. The access path on ODN is reconfigured periodically in synchronization with the timeslots to connect the L-OLT and every ONU groups.

In the TDMA-based OLT sharing method, the communication interval of each ONU group becomes large, such as few seconds. During a timeslot allocated to one ONU group, downstream data frames to other ONU groups are discarded in the L-OLT due to its buffer capacity. To solve the problem of the buffer limit and stabilize throughput, a proxy of L-OLT [7, 8] is introduced. The proxy is generated in the layer-2 network that connects the core network and P-OLTs by exploiting Network Function Virtualization (NFV) and service function chaining. The proxy buffers downstream data frames to each ONU group and transfers them to the
L-OLT in synchronization with switching of ONU groups so that frame loss in the L-OLT is avoided.

3 Experimental evaluation

In this section, we report the experiment results of the inter-OLT restoration and the TDMA-based OLT sharing method. Fig. 1 shows the software-based EλAN prototype system implemented for the experiments.

The software-based EλAN prototype system consists of two P-OLTs, two P-ONUs, layer-2 and layer-1 switches, and the proxy utilized in the TDMA-based OLT sharing method. We assumed that the two P-OLTs were deployed in different COs. In addition, NMS and a frame generator (not presented in Fig. 1) were connected to this system.

P-OLTs and P-ONUs were implemented by using commodity Linux servers. L-OLTs that provide buffering, header modification, and forwarding of Ethernet frames were configured in each P-OLT as software on Docker containers. The layer-2 switch transfers data frames to/from appropriate P-OLTs. We assumed that P-ONU #1 and #2 belong to different ONU groups #1 and #2, respectively. Instead of ODN, an electrical layer-1 switch that transfers data signals without any processing was deployed in this experiment. This layer-1 switch can be utilized as an alternative to ODN since it is a full-mesh crosspoint switch that can connect any two ports. NMS collected the operation status information including traffic amount and alarms periodically from each L-OLT. Each of recovery methods was triggered when a L-OLT generated a Loss of Signal (LoS) alarm and NMS caught it. The frame generator sent and received two different traffic flows distinguished by VLAN IDs for ONU group #1 and #2.

3.1 Inter-OLT restoration method

At first we conducted the experiment of the inter-OLT restoration method. Fig. 2(a) shows the network configuration. When NMS received a LoS alarm from the
L-OLT, NMS calculated the placement of L-OLTs and determined the migration of L-OLT #1 from P-OLT #1 to P-OLT #2. Then NMS calculated the route of access path and reconfigured the layer-1 switch. After that, L-OLT #1 was migrated to P-OLT #2 so that the connectivity of P-ONU #1 was restored.

Fig. 2(b) shows the traffic rate of the flow for ONU group #1 measured at the receiving port of the frame generator. After the migration of L-OLT #1 and the reconfiguration of layer-1 switch, the traffic flow for ONU group #1 was restored through P-OLT #2 with the same traffic rate as before. The inter-OLT restoration procedure took 40.5 seconds on average from the detection of LoS to the resumption of traffic. The majority of the service interruption time was spent on the migration of the Docker container, which took 37.4 seconds on average. There is room for reducing the service interruption time to few seconds by using another L-OLT migration method, such as process migration [6].

3.2 TDMA-based OLT sharing method

Next we conducted the experiment of the TDMA-based OLT sharing method. Fig. 3(a) shows the network configuration. In this experiment, the proxy was realized as software on a fixedly-placed Linux server. The proxy software was disabled in the normal operation, and data frames only passed through the server without any buffering and processing.

When NMS received a LoS alarm, NMS tried to calculate the placement of L-OLTs at first. However, in this experiment, there was no room in P-OLT #2 to configure a new L-OLT. Therefore, NMS decided to execute the TDMA-based OLT sharing method that L-OLT #2 accommodated both ONU groups to maintain the connectivity. NMS controlled the operation mode of L-OLT #2 periodically, in other words, allocated timeslots to ONU group #1 and #2 in rotation. NMS also turned on the proxy software, and controlled the layer-1 switch and the proxy at the same time in order to transfer the traffic flows to the appropriate ONU groups.

In this experiment, the duration time of a timeslot for each ONU group was set to 10 seconds.

Fig. 3(b) shows the traffic rate measured at the receiving port of the frame generator. At first, traffic of both flows was observed at the same time. After the transition to the TDMA-based OLT sharing, each of traffic flows was appeared alternately at 10-seconds interval. The transition time from the normal operation to
the TDMA-based OLT sharing was 1.55 seconds on average. Note that ONU group #2 had to wait additional 10 seconds to restart the communication with L-OLT #2. The switching time of ONU groups was 1.45 seconds on average.

4 Conclusion

The inter-OLT restoration and the TDMA-based OLT sharing method have been proposed as fault recovery methods in EλAN. In this letter, we reported the experimental evaluation of both methods conducted by using the software-based EλAN prototype system. In the inter-OLT restoration method, 40.5 seconds were required from failure detection to connection recovery, but this required time is expected to be reduced to few seconds. In the TDMA-based OLT sharing method, we confirmed that ONU groups can keep the connectivity by sharing a single L-OLT, with 1.45 seconds of the ONU group switching time.

Acknowledgments

This work is supported by “Research and Development of Elastic Optical Aggregation Network,” the commissioned research of National Institute of Information and Communications Technology (NICT).
Facility implementation of adaptive clutter suppression to an existing wind profiler radar: First result

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Abstract: Adaptive clutter suppression (ACS) is a technique that mitigates signals from undesired objects by using subarrays. However, because many of existing wind profiler radars (WPRs) have a single-channel receiver, ACS cannot be applied to them. Aiming at implementing ACS capability to existing WPRs, a pilot system composed of auxiliary subarrays, Universal Software Radio Peripheral (USRP), and a workstation, was developed. The use of USRP enabled multi-channel reception and easy development of the software for real-time data processing by using C++. An example, in which ACS was applied to signals collected by the pilot system, is presented.

Keywords: radar, clutter, adaptive array, software-defined radio

Classification: Sensing

References

1 Introduction

Wind profiler radar (WPR) receives signals scattered by radio refractive index irregularities (clear-air echo), and measures height profiles of vertical and horizontal wind in the clear air. Owing to the capability of measuring wind in the clear air, WPRs are utilized for studying dynamical processes in the atmosphere [1]. WPRs are also used for monitoring wind variations routinely. In Japan, a nationwide WPR network, referred to as Wind Profiler Network and Data Acquisition System (WINDAS), is operated in order to provide upper-air wind data to the numerical weather prediction [2]. A wind profiler network is also operated in Europe [3].

Received signal from an undesired object (i.e., clutter) contaminates a frequency spectrum of received signal (i.e., Doppler spectrum). The contaminated Doppler spectrum cannot be used frequently because the clutter contamination inhibits accurate measurement of wind velocity. Therefore clutter signals need to be mitigated. Clutter signals received by a sidelobe of a radar beam are able to be mitigated by using an adaptive array technique referred to as adaptive clutter suppression (ACS). By using atmospheric radars with a multi-channel receiver, ACS has been studied (e.g., [4, 5, 6, 7]). However, because many of existing WPRs have a single-channel receiver, ACS cannot be applied to them. In this study, it is demonstrated that ACS facility can be implemented to an existing WPR by additionally installing a simple system.

2 Facility for adaptive clutter suppression (ACS)

In order to implement ACS capability to an existing WPR referred to as LQ-13 [8], a pilot system (hereafter ACS system) was developed. LQ-13 has the center
frequency of 1357.5 MHz and the peak transmission power of 5.2 kW. The radar beam of LQ-13 is able to point to vertical and four oblique directions (i.e., toward north, east, south, and west with a zenith angle of 14°). Fig. 1 shows a block diagram of the ACS system. The ACS system is composed of three auxiliary subarrays (SAs), analog unit, three signal samplers referred to as Universal Software Radio Peripheral (USRP) X310 [9], and a workstation (WS). Because objects at the surface (buildings, trees, cars, and so on) are major sources of clutters, the ACS system mainly aims at mitigating clutters existing on the ground. The beam pattern of the SAs has the sensitivity maximum at the 0° elevation angle, has the sensitivity minimum at the zenith direction, and is omni-directional in the horizontal plane. In order to mitigate a clutter signal from an arbitrary horizontal direction, three SAs are used in the ACS system. The SAs are installed outside the clutter fence of LQ-13. The analog unit amplifies signals from the SAs and converts their frequency from the radio frequency (RF) of 1357.5 MHz to the intermediate frequency (IF) of 130 MHz. The USRP X310s digitize signals from the main antenna of LQ-13 (hereafter the main antenna) and those from the SAs. Then the USRP X310s perform digital quadrature detection. They also collect a trigger signal produced by the timing controller of LQ-13. The trigger signal and the complex time series are transferred to the WS with a transfer rate of 10 mega-samples per second through the 10 Gigabit Ethernet. The 10 Gigabit Ethernet is used for attaining the data transfer rate necessary for the multi-channel signal collection [10]. In order to synchronize frequencies of the USRP X310s with those of the transmitter and receiver of LQ-13, 10-MHz reference signals and 1 pulse per second (PPS) signals are supplied from the GPS receiver installed in LQ-13.
The WS carries out the real-time data processing after the quadrature detection. The operating system of the WS is Ubuntu 14.04 LTS. The software used for the real-time data processing is written in C++, and was developed by updating the software for the digital receiver which uses USRP as a signal sampler [11]. By the update, the facility of processing multi-channel signals was implemented. Because C++ is a general-purpose programming language, the use of C++ facilitated the update of the software. The software executes two kind of threads; the data taking thread (DT) and the signal processing threads (SPs). In order to avoid memory conflict and to carry out continuous data collection without an interruption by data transfer between the threads, DT and SPs are interfaced with the shared memory. Using the trigger signal, DT carries out ranging. Ranging is carried out on the WS because USRP X310 transfers data only sequentially (i.e., both during transmission and reception) [8]. SPs carries out real-time data processing necessary for reducing the data amount. In order to process signals from the main antenna and those from the three SAs simultaneously, four SPs are executed in parallel.

The ACS system as developed is easy to be installed to existing WPRs. Requirement for its installation is only supply of the following signals from an WPR to the ACS system; 10-MHz reference signal, 1 PPS signal, trigger signal for transmission or reception, and received signal from the main antenna. C++ library referred to as Universal Hardware Driver (UHD) [12] is used to control the USRP X310s. By using UHD, all the controls of the USRP X310s, including the configuration of the measurement parameters (number of receive channels, receive frequency and data transfer rate), are executed by the software written in C++. Therefore it is not necessary to use Hardware Description Language and other tools for designing the logic on the Field Programmable Gate Array in the USRP X310s.

3 Measurement result

In this section, an example of a measurement result is presented. As an algorithm for adaptive signal processing, directionally constrained minimization of power with constrained norm (DCMP-CN) [5, 6, 7] was used. In the DCMP-CN used in this study, the value of the directional constraint was 1 for signals from the main antenna and 0 for signals from the SAs. The values were used because the SAs are not sensitive at the beam directions of the main antenna (i.e, at the zenith angle of 0° or 14°) [5]. Therefore, under the condition that the weight for signals from the main antenna is unchanged and that the norm maximum of weight vector is limited, the DCMP-CN minimizes the power of the synthesized signal. The norm constraint value of 1.5 was used. The amplitude of the time series from the main antenna and those from the SAs were normalized so that they have the same noise power. The direct current (DC) components of the times series were removed because the removal gave the better ACS performance for the antenna configuration described in Section 2 and for the location of LQ-13.

Fig. 2 shows an example of time series. The beam direction and the transmitted subpulse width were eastward and 1 µs, respectively. The signal from the main antenna is shown in Fig. 2a. The time series has two components; a clear-air echo from the main beam and a clutter signal from a sidelobe. The clutter signal changed
slowly with time, and the slow change is clearly seen especially in the Q component. On the other hand, the clear-air echo varied more with time. The weighted sum of the signals collected by the three SAs is shown in Fig. 2b. Because the SAs are not sensitive at the beam direction of the main antenna, only the signal from the clutter was measured. The sum of the signals from the main antenna (Fig. 2a) and those from the SAs (Fig. 2b) is shown in Fig. 2c. It is the final result obtained by applying the DCMP-CN, and shows that the signals from the SAs were weighted so that they cancel the clutter signal from the main antenna.

Fig. 3 displays the Doppler spectrum of the time series shown in Figs. 2a and that of the time series shown in Fig. 2c. The Doppler spectrum point at 0 Doppler velocity is not plotted for the unsmoothed plots (i.e., blue and black curves)
because their values are small. When the DCMP-CN is not applied, the clutter signal is dominant at the ranges at which Doppler velocity is close to 0 (blue curve). Owing to the contamination of the clutter signal, the estimated Doppler velocity is as small as $-0.06 \text{ m s}^{-1}$. However, it becomes $-0.48 \text{ m s}^{-1}$ by applying the DCMP-CN (black and red curves). With an assumption that the clutter signal was completely removed by the DCMP-CN, the signal to clutter ratio is estimated to be $-10.4 \text{ dB}$.

It is noted that the phase differences of the SAs caused by the hardware need not be calibrated because the signals from the SAs do not contain the clear-air echo (see Fig. 2b). This advantage also contributed the easy installation of the ACS system.

4 Conclusion

In this study, we demonstrated that ACS facility can be implemented to an existing WPR by additionally installing a simple system comprised of auxiliary subarrays, USRP, and a workstation. Because increasing number of SAs gains adaptability of sidelobe control, experiments using more number of SAs are useful for evaluating performance of ACS. Such experiments are easy to be realized because the use of USRP and the software written in C++ facilitate changes of the ACS system. By taking advantage of the availability of the ACS system, further studies, which aim at implementing ACS capability to existing WPRs, are going to be carried out.

Acknowledgment

This study is supported by KAKENHI Grant Number 26281008 funded by the Japan Society for the Promotion of Science.
All-optical feedback gain control of remote optically pumped amplifiers

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Abstract: A novel all-optical feedback automatic gain control scheme for remote optically pumped amplifiers has been proposed. The static and dynamic gain control characteristics of the remotely pumped erbium-doped fiber amplifier have been experimentally clarified. The maximum gain excursion has been successfully reduced by a factor of \sim10 from 2.1 dB to \sim0.2 dB using the proposed gain control scheme.

Keywords: automatic gain control, ROPA, EDFA, all-optical

Classification: Fiber-Optic Transmission for Communications

References


1 Introduction

Compared to conventional lumped erbium-doped fiber amplifiers (EDFAs), remote optically pumped amplifiers (ROPAs) can significantly enhance the transmission distances/capacities in long-haul transmission systems (unrepeatered and repeatered systems) [1, 2, 3, 4, 5, 6, 7]. This is because the ROPAs can achieve higher optical signal-to-noise ratios (OSNRs) of more than ~5 dB over that of the lumped EDFAs. The automatic gain control (AGC) of the ROPA is indispensable in the wavelength routing/switching systems of photonic networks in future [8, 9, 10]. Moreover, realizing core-by-core AGC of a remotely pumped multicore EDFA, which has a shared pump light source for several cores of a multicore EDFA, is a crucial issue in a spatial division multiplexing transmission system [8, 9, 10]. An ROPA consists of a remotely pumped EDFA (RP-EDFA) section and a distributed Raman amplification (DRA) section [6]. Therefore, the gain of each amplifier section must be dynamically kept constant against the change in the number of the wavelength division multiplexing (WDM) signal channels. In this paper, we propose an all-optical feedback AGC (FB-AGC) scheme applicable to both single-core and multicore ROPA transmission systems and report experimental results on the AGC characteristics of ROPA transmission systems for the first time, to our knowledge. Using a novel RP-EDFA module, a dynamic FB-AGC operation with gain excursions of less than ~0.2 dB has been successfully achieved. Note that some preliminary experimental results on the FB-AGC scheme were previously reported in [11, 12].
2 Experimental configuration

The experimental configuration of our proposed all-optical FB-AGC scheme for an ROPA is shown in Fig. 1(a). Here, we assumed a 40-channel WDM system in the experiment. We launched a signal power of 1 mW per channel into the transmission fiber span. The WDM signal lights consisted of a surviving light and four saturation lights. The wavelength of the surviving light was 1550.0 nm, and those of the saturation lights were 1532.7, 1539.0, 1546.9, and 1552.5 nm in the C-band. The saturation lights were periodically added and dropped by an acousto-optic modulator (AOM) to evaluate the dynamic AGC characteristics of the scheme.

The transmission fiber span consisted of a variable optical attenuator (VOA\textsubscript{span}), two 30-km-long transmission fibers (Fiber-1 and -2), and an RP-EDFA module between the transmission fibers. We used VOA\textsubscript{span} instead of a transmission fiber for experimental convenience. We varied the loss of VOA\textsubscript{span} to adjust the loss of the transmission fiber span (L\textsubscript{span}). Distributed Raman amplification was generated in Fiber-2. We investigated the AGC characteristics of the ROPA system for two typical values of L\textsubscript{span} of 25.4 and 28.2 dB, corresponding to span lengths of 134 and 149 km, respectively, assuming that the loss coefficient of the transmission fiber at 1550.0 nm is 0.19 dB/km. The gain excursion is expected to be larger with a smaller span loss which corresponds to a shorter span length. Therefore, the case of 25.4-dB span loss is worst for AGC operation. The RP-EDFA module had a fiber-ring-laser circuit, which included an RP-EDF, an isolator, a variable optical attenuator (VOA\textsubscript{ring}), and two wavelength selective couplers (WSC\textsubscript{F} and WSC\textsubscript{R}). Both ends of the RP-EDF, whose length was 3.2 m, were fusion-spliced with standard single mode fibers. We adjusted the loss of the VOA\textsubscript{ring} to obtain a flat gain spectrum of the RP-EDFA module over the C-band. The net flat gain of the RP-EDFA module was \sim 14 dB. We used the WSC\textsubscript{F} and WSC\textsubscript{R} to couple and separate, respectively, the WDM signal lights and laser light. WSC\textsubscript{F} and WSC\textsubscript{R} were dielectric multilayer couplers for the 100- and 200-GHz grid WDM system, respectively. They had 3 ports, referred to herein as the signal, laser, and common port. A pump light from a pump light source (Pump LS) with a wavelength (\lambda\textsubscript{p}) of 1490 nm (except in the measurement of the static gain characteristics of the RP-EDFA module, which will be described in the next section) was launched into Fiber-2 via a coupler. The power and spectra of the WDM signal lights were measured by a sampling oscilloscope and an optical spectrum analyzer.

The loss spectra of WSC\textsubscript{F} and WSC\textsubscript{R} are shown in Fig. 1(b)–(d) and (e)–(g), respectively. Fig. 1(b) and (e) shows the spectrum of WSC\textsubscript{F} and WSC\textsubscript{R}, respectively, between the common and signal ports. The losses of WSC\textsubscript{F} and WSC\textsubscript{R} for the WDM signal lights were less than 0.5 dB. The low losses for the WDM signal lights are required to achieve high OSNRs. The loss of WSC\textsubscript{R} for the pump light was also less than 0.5 dB. The low loss for the pump light is required to achieve a high pumping efficiency. The pump power penalty to employ the FB-AGC scheme was equal to the pump loss of the WSC\textsubscript{R}. Fig. 1(c) and (f) show the detailed loss characteristics of the rejection band of WSC\textsubscript{F} and WSC\textsubscript{R}, respectively. The losses of WSC\textsubscript{F} and WSC\textsubscript{R} within a 1-dB bandwidth of WSC\textsubscript{F} for the laser light transmission were more than 11 and 22 dB, respectively. High attenuation at the
laser light wavelength ($\lambda_1$) is necessary for WSCR to avoid degradation of the signal light propagating in Fiber-2 owing to nonlinear effects caused by co-propagating laser light at high power. Fig. 1(d) and (g) shows the loss spectra of WSCF and WSCR, respectively, between the common and laser ports. $\lambda_1$ was within the pass band of WSCF, which was narrower than that of WSCR. The center wavelength and 1-dB bandwidth of the pass band of WSCF were 1559.3 and 0.5 nm, respectively.

3 Experimental results

First, we measured the OSNRs of the systems with and without the ROPA scheme. We used an input signal power of 0 dBm launched into the transmission fiber span and a noise band width of 0.1 nm. For an $L_{\text{span}}$ value of 25.4 dB, the OSNRs without ($OSNR_{w/o}$) and with ($OSNR_w$) the ROPA scheme ranged from 26.3 to 27.6 dB and 31.0 to 32.8 dB, respectively. The ONSR improvements ($\Delta OSNR$) ranged from 4.7 to 5.1 dB. Here, $\Delta OSNR$ was defined as the difference between $OSNR_w$ and $OSNR_{w/o}$. On the other hand, for an $L_{\text{span}}$ value of 28.2 dB, the $\Delta OSNR$ values ranged from 4.7 to 5.3 dB. The data indicate that $\Delta OSNR$ values of $\sim$5 dB were achieved across the C-band by employing the ROPA scheme for both cases of $L_{\text{span}}$. 

Fig. 1. (a) Experimental configuration of all-optical FB-AGC scheme for an ROPA. Loss spectra of (b)–(d) WSCF and (e)–(g) WSCR.
Next, we investigated the static gain characteristics of the RP-EDFA module for two values of the wavelength of the surviving light ($\lambda_{\text{surv}}$). Fig. 2(a) and (b) show the dependences of the gain on the total input signal power ($P_{\text{sin}}$) at typical pump powers ($P_p$) from 131 to 170 mW for $\lambda_{\text{surv}}$ of 1550.0 and 1531.0 nm, respectively. $P_{\text{sin}}$ and $P_p$ were the powers launched into the RP-EDFA module in this measurement. $P_{\text{sin}}$ was the sum of the power of the surviving light ($P_{\text{surv}}$) and that of the saturation light ($P_{\text{sat}}$). $P_{\text{surv}}$ was kept at 0.039 mW while $P_{\text{sin}}$ was adjusted by varying $P_{\text{sat}}$. For experimental convenience, we used $\lambda_p$ of 1.48 µm in this measurement. In Fig. 2(a), the maximum static gain ($G_{s,\text{max}}$) was $\sim$13.5 dB at $P_{\text{sin}}$ of 0.039 mW for all values of $P_p$. The gains decreased with increasing $P_{\text{sin}}$ and increased with $P_p$. The minimum static gain ($G_{s,\text{min}}$), which depended on $P_p$, ranged from $\sim$13.2 to $\sim$13.3 dB at $P_{\text{sin}}$ of 1.51 mW. The gain difference ($\Delta G_s$), the difference between $G_{s,\text{max}}$ and $G_{s,\text{min}}$, ranged from 0.2 to 0.3 dB for the tested pump powers $P_p$ with $\lambda_{\text{surv}}$ of 1550.0 nm. On the other hand, $G_{s,\text{max}}$ was $\sim$14.4 dB, and $G_{s,\text{min}}$ was in the range from $\sim$13.8 to $\sim$13.9 dB with $\lambda_{\text{surv}}$ of 1531.0 nm, as shown in Fig. 2(b). $\Delta G_s$ ranged from 0.5 to 0.6 dB. The data indicate that the average value of $\Delta G_s$ at 1531.0 nm (0.55 dB) was $\sim$2 times larger than that at 1550.0 nm (0.25 dB). It is considered that $\Delta G_s$ was large at near the gain peak wavelength ($\sim$1532 nm) due to the spectral hole burning effect [11].

Finally, we evaluated the dynamic characteristics of the FB-AGC scheme of the ROPA. $P_{\text{sin}}$ launched into the transmission fiber span was 16.0 dBm in this experiment. For simplicity, $P_{\text{surv}}$ launched into the RP-EDFA module was set at $-17.7$ dBm in both cases of $L_{\text{span}}$. $P_p$ launched into the transmission fiber were 315 and 170 mW for the cases of $L_{\text{span}}$ of 25.4 and 28.2 dB, respectively. Raman gains in the DRA section were 4.3 and 2.3 dB, respectively. Fig. 3(a) shows the transient gain excursions of the surviving light with and without the FB-AGC scheme at $L_{\text{span}}$ of 25.4 dB. To evaluate the AGC characteristics of the ROPA system, the gain was measured at the end of Fiber-2. The saturation lights were dropped at the times ($t$) of 1.3 and 6.3 ms, and added at 3.8 ms. The gain difference as a function of $t$, $\Delta G_d(t)$, is defined as the gain at $t$ minus the gain in the steady state for the case of adding. Without the FB-AGC scheme $\Delta G_d(t)$ increased (decreased) when the saturation lights were dropped (added) owing to the gain saturation effect of the ROPA. The maximum value of $\Delta G_d(t)$ ($\Delta G_{d,\text{max}}$) was 2.1 dB. On the other hand,
with the FB-AGC scheme, the gain excursion was significantly suppressed. Fig. 3(b) shows the transient gain excursion with the FB-AGC scheme in the short time range for the case of dropping. Oscillation in the gain excursion was observed when the saturation lights were dropped. It is considered that the gain-excursion oscillation was caused by the relaxation oscillation of the laser light in the fiber-ring-laser circuit in the RP-EDFA module. \( \Delta G_{d,max} \) was 0.20 dB with the FB-AGC scheme, \sim 10\) times smaller than that without the scheme.

On the other hand, \( \Delta G_{d,max} \) without and with the FB-AGC scheme at \( L_{\text{span}} \) of 28.2 dB were 1.7 and 0.14 dB, respectively, which was \sim 12\) times smaller than that without the scheme. \( \Delta G_{d,max} \) was decreased with increasing \( L_{\text{span}} \) of 2.8 dB (28.2–25.4 dB), because the signal power launched into the RP-EDFA module was decreased by 2.8 dB.

![Fig. 3. (a) Transient gain excursions of the ROPA with and without the FB-AGC scheme. (b) Oscillation in the gain excursion with the FB-AGC scheme.](image)

### 4 Conclusion

We have proposed a novel all-optical feedback AGC scheme for ROPAs. We experimentally clarified the static and dynamic gain control characteristics of the RP-EDFA. The OSNR improvements of \sim 5\) dB were achieved across the C-band by employing the ROPA scheme. The maximum gain excursion was significantly reduced from 2.1 dB without AGC to \sim 0.2\) dB with the all-optical feedback AGC scheme for the worst case of 25.4-dB span loss.

### Acknowledgments

This work was supported in part by JSPS KAKENHI Grant Number JP16K06355.
Implementation and verification of PCI express interface in a SoC

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Abstract: This paper describes the implementation and design of interface for peripheral component interconnect express (PCIe) interconnect and memory in a complex system on chips (SoC). PCIe bus traffic is made of a series of PCIe bus transactions. The direction of the data will be from initiator to completer (for write transaction) or vice-versa (for read transaction). The interface will read the command of the master and send corresponding response to the master. The major objective of the project is performance verification of SoC on a dedicated channel between PCIe end point and memory using performance models. We are using direct memory access (DMA) type of requests and bandwidth is measured at bottleneck for different PCIe generations, lane configurations and payloads. Bandwidth obtained is being compared with theoretical peak bandwidth calculated.

Keywords: PCIe, memory, SoC, DMA

Classification: Network

References

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1 Introduction

Over the past years, processor performance is going on improving, but the technological progress of system interconnects are not matching with them. Comparing with existing system interconnects such as Ethernet and InfiniBand, the PCI Express (PCIe) device has advantages such as higher protocol efficiency, low-power consumption per port, and low unit price [1]. To help the processor and interconnect have balanced performance, we propose a system interconnection device using PCIe. PCIe is high-speed point-to-point connection technology that comes from the typical PCI bus technology. It is used as the primary interface between processor and IO devices. It is offering high-speed data transfers over a physical link composed of multiple lanes that are scalable (from ×1 to ×32). Starting from the first generation (GEN 1) with 2.5 Giga Transfers per second (GT/s) speed per lane, the PCIe has been continually enhancing its performance and reached 16 GT/s per lane speed as its fourth generation [2]. Second generation (GEN 2) and third generation (GEN 3) of PCIe have speeds as 5 GT/s and 8 GT/s respectively.

2 Review on PCIe bus

PCIe employs point-to-point interconnects for communication between two devices as against its predecessor buses that used multi-drop parallel interconnect. A point-to-point interconnect implies limited electrical load on link allowing higher frequencies to be used for communication. Because of the serial interconnection, the board design cost and complexity has reduced considerably.

PCIe can be demonstrated as layered architecture. It has three different logical layers: Transaction layer (TL), Data link layer (DL), and Physical layer (PHY). PCIe uses packet to transmit information from a transmitter to receiver. Hence it is known as a packet based protocol. The packet is transmitted through different layers. Each layer adds its additional information as described below [3].

Transaction layer adds the Transaction Layer Packet (TLP) header to the data payload. Data link layer adds the sequence id and link level cyclic redundancy check bits to the TLP. At physical layer, the framing bits are added to the packet coming from data link layer. The conversion of logical signals to electrical signals happens in physical layer.

3 Interface design

PCIe interface is a device that acts as a transitional medium between initiator and completer devices. In our design we have PCIe as the initiator or master and memory as completer or slave. Here, we can read data from the memory and can also write data from PCIe to memory [4]. The read request gets a response data back from the memory to PCIe. While the write request gets an acknowledgement signal.

There are different signals that are modelled in interface block. They are named below along with brief functionality.

- Valid – It is the activation signal. When it is low the device is valid to perform.
  - Data transmission starts only after high signal value of valid.
• **Command** – It is a 6 bit bus. If this is 000100 then it indicates that the request command is the read. If it is equal to 101100 then it indicated that the request command is write.

• **Address** – It is a 36 bit bus. The request address is transmitted over this signal.

• **Length and Virtual Channels** – Both are 4 bit fields. Length is amount of data per channel. Virtual Channels (VC) is the number of channels incorporated in this. Product of both the values gives the actual amount of data.

• **Ready** – This is a single bit bus. There are 2 types of ready signals. One is initiator ready and other one is completer ready.
  - Initiator ready (Irrdy) – If it is high then the initiator is ready to perform an action.
  - Completer ready (Crrdy) – If it is high then the completer is ready to perform an action.

• **Tag** – Count of number of requests issued.

### 4 FSM implementation

The Finite State Machine (FSM) implementation is designed as shown in Fig. 1.

**Fig. 1.** FSM state diagram for PCIe interface.

**A. State_idle (S0):** In this state all the signals are in idle mode. Or if they are not in idle state then we are resetting all the signals to idle state. That means all the signals are inactive to perform operation and wait for the master device activation signal. When the valid signal goes high, it activates the master device that means the PCIe is now active to read or write data to slave devices [4, 5].

**B. State_initiator_ready (S1):** If Irrdy changes its state from idle state that is 1, it also changes the state and goes to the next state which is state_completer ready. But
if the Irrdy signal is not high then it waits for this signal to be high. And if the valid again changes its state from high to low, it goes back to the idle state [4].

C. State completer_ready (S2): In this state, the Crrdy signal is checked. If it becomes high, then the state change is happened to the state_send_address. If the initiator ready is low again, then is initiator is not ready for the transmission. Hence the state changes back to state_initiator_ready.

D. State_send_address (S3): The fourth state is the state send address. In this state at first it checks that weather it is a write transaction or a read transaction. If it is a write transaction and the slave device is memory, then it makes the other internal signals as its need and makes the memory activation signal chip select (CS) as low and write enable high, thus the transaction occurs. If it is a read operation from memory, then again it makes the other internal signals as its need and makes the memory activation signal cs low and read enable high.

E. State_final (S4): The tag value is compared with the total number of transactions specified (N). If the tag equals N, then state goes back to idle state. This process will continue until or unless the valid signal goes high. If it goes high, then it stops the transaction and goes back to the idle state.

5 Simulation results

Compilation and Simulation of the design is done on Synopsys Verilog Compiler Simulator (VCS) tool. Simulation waveforms for Direct Memory Access (DMA) write and read are as shown in Fig. 2. First set of waves are for DMA write transaction and second set represents the DMA read transaction.

![DMA transactions simulation waveforms](image)

Considering the parameters described in previous sections, different test cases are formulated and made to be run on SOC. It is possible to calculate or at least
obtain reasonable estimates of the performance values. In general, the bandwidth is total amount of data transferred in particular amount of time. Overall bandwidth includes the bandwidth values generated by different types of transactions [6].

The theoretical performance values of different DMA cases obtained using formula. For GEN 3 128b/130b encoding is used, while 8b/10b encoding scheme is used for GEN 1 and GEN 2. Encoding scheme cannot be altered unlike some performance factors [6].

\[
\text{Bandwidth} = \left( \frac{\text{Total data transferred}}{\text{Total transfer time}} \right) \times \frac{\text{GB}}{s}.
\]

Hence the final bandwidth values depend on the number of outstanding transactions sent from the PCIe endpoint and the initial and final time instances of the transactions. The total number of outstanding transactions gives a measure of the data transferred. Obtained values are being compared with theoretical maximum values to verify the performance as shown in Fig. 3.

6 Conclusion

In this paper, we have presented a PCIe interface design to efficiently transmit the initial upstream request information to the memory and downstream response information to the device. We have implemented the interface design using a FSM for DMA transactions initiated from PCIe endpoint to the memory. Bandwidth measured itself is concluding that the PCIe interface is a promising system interconnect for complex SoC’s. Test case scenarios are developed with different configurations to perform the memory access routine. The current performance testing results demonstrate that PCIe link with DMA feature enabled outperforms the remaining interconnects.
Characteristics of ultra-wideband radar echoes from a drone

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Abstract: This letter proposes to use an ultra-wideband (UWB) radar for drone detection and experimentally investigates feasibility of the proposal. First, the radar cross section (RCS) of a typical quadrotor drone is clarified. Next, radar range profiles of a flying drone are discussed. As a result, we have confirmed that an UWB radar with high range resolution can detect a drone and observed a unique feature of the echoes reflected from the drone’s rotor blades. The feature will enable us to distinguish a drone from other flying objects.

Keywords: ultra-wideband, radar, drone, RCS, measurement

Classification: Sensing

References

1 Introduction

In recent years, small drones which have multiple rotor blades are widely used for various usages. Those usages include package delivery and observations of disaster-stricken areas [1]. The increase in drone is causing social problems. The case of a drone that fell to the Japanese Prime Minister’s official residence is still fresh in our mind. Since a drone has payload of several hundred grams to several kilograms even in small size, it has a danger of being used for terrorism. Therefore, it is an urgent issue to realize a system that detects malicious drones and captures them quickly.

Radar is one of the technologies that can remotely detect drones. Unlike camera images or sound concentrating microphones, radar is available in all weather conditions, because it is rarely hindered by rain or fog. In [2] and [3], for the drone detection, the uses of a continuous wave (CW) radar and a narrowband pulse radar have reported, respectively. These studies tried to detect and distinguish a drone from other flying objects such as birds and insects, by analyzing the Doppler signatures of the drone.

Ultra-wideband (UWB) radar is a technology that offers high precision ranging and high range resolution by transmitting and receiving ultrashort pulses with a bandwidth of 0.5 GHz or more [4]. In [5], a method for identifying a type of a car by using an UWB radar has been proposed. The method detects the echoes reflected from each part of the car with high accuracy. We thought that UWB radar could capture unique features of a drone such as the echoes from the rotor blades. In this letter, we propose to use an UWB radar for drone detection and experimentally investigate feasibility of the proposal. First, the radar cross section (RCS) of a typical quadrotor drone is clarified. Next, radar range profiles of a flying drone are discussed. In general, the higher the frequency, the easier for the radar to observe the target’s characteristics. Therefore, we used the submillimeter wave band (@ 24/26 GHz), which is relatively high frequency and is also employed for automotive radars.

2 RCS measurements of a typical quadrotor drone

RCS is the measure of a target’s ability to reflect echo signals toward the radar. The RCS value, which vary considerably with the target’s size, material, viewing aspect and frequency of radar radio wave, affects the radar performance. Therefore, it is important to measure the RCS of a target for radar system design. The RCS of a rotor blade of a drone has been reported so far [6]. In this report, the RCSs of the rotor blade of two kinds of materials (metal, carbon) were measured in three frequency bands (L-band: 1–2 GHz, S-band: 2–3 GHz, C-band: 5–6 GHz). However, the RCS of a drone including the entire body has not been clarified. In order to calculate RCS in the actual environment, it is necessary to consider losses due to...
measurement system. To take the losses into consideration, in general, the RCS of a target $\sigma$ is calculated by

$$\sigma = \left( \frac{P}{P_{\text{ref}}} \right) \cdot \left( \frac{R}{R_{\text{ref}}} \right) \cdot \sigma_{\text{ref}}$$ (1)

using a standard reflector whose RCS is known [7]. Where $\sigma_{\text{ref}}$ represents the RCS value of the standard reflector. $R$ and $R_{\text{ref}}$ are the distance to the target and that to the standard reflector, respectively. $P$ and $P_{\text{ref}}$ are the received power from the target and that from the standard reflector, respectively.

A quadrotor drone (DJI Phantom 3 with plastic rotor blades) was mounted on a turntable in an anechoic chamber as shown in Fig. 1(a). Three axes ($x$, $y$, and $z$ axes) fixed to the drone’s body are defined as shown in Fig. 1(b). Measurements were carried out with H-H polarization in two cases. In the first case, we placed the drone on the turntable so that the drone’s $z$ axis was vertical, as shown in the Fig. 1(b). Then we rotated the drone $\theta$ degrees around the $z$ axis. In the second case, the drone was placed on the turntable so that its $y$ axis was vertical, and we rotated the drone $\phi$ degrees around the $y$ axis. The received signal power from the target was measured using a vector network analyzer (VNA) while rotating the turntable by 360 degrees in increments of 0.6 degrees. The center frequency $f_c$ was set to 24 GHz. For comparison, measurements, where $f_c$ was set to 2.4 GHz, were also carried out. The bandwidth $BW$ was set to 0.5 GHz. Besides, a metal sphere with a diameter of 0.24 m ($\sigma_{\text{ref}} = -13$ dBsm) was used as the standard reflector.

Figs. 1(c) and 1(d) show the measured values of the RCS patterns in the two cases. The average value of RCS at $f_c = 24$ GHz is $-13$ dBsm in both cases. We can see that the RCS pattern of $f_c = 24$ GHz changes more sharply depending on
the angle than that of $f_c = 2.4$ GHz. This is because the wavelength is shorter than the size of the drone when $f_c = 24$ GHz, thus reflections from each part of the drone’s frame are obtained. Furthermore, in the case of $f_c = 24$ GHz in Fig. 1(c), there is a peak at 90 degrees. The reason is probably because there is a large reflective surface on the side of the drone mounted camera.

3 UWB radar measurements for a flying drone

We also measured radar range profiles, which indicate intensity of echo signals as a function of distance. Using a VNA’s time-domain function, the range profile measurements were carried out for 30 seconds while we let the drone fly. We conducted this experiment in a laboratory in order to avoid the influence of wind on the drone’s flight. Fig. 2 shows the measurement system, where $f_c = 24$ GHz, $BW = 3, 1, 0.5$ GHz. Standard gain horn antennas (25 dBi) with H-H polarization were used and the height was set 1.2 m above the floor. In this experiment, we set two measurement scenarios: the case of hovering at a distance of 3 m and the case of flying round-trip at a distance of 3 m to 7 m. In both cases the altitude was set at 1.2 m above the floor.

Fig. 3(a) shows the measured range profile at $BW = 3$ GHz when the drone is hovering. We can see an intense echo at 3 m distance and also unsteady fluctuation in distance due to hovering. Furthermore, there are two echoes that similarly fluctuate in the distances just before and after the intense echo. This is because the echoes, which are reflected from the drone body and the rotor blades, can be separated with the high range resolution capability. These echoes from the rotor blades are mainly reflected signals from all of the motor parts (rotors) that rotate the blades. Because rotors are always installed in a drone, it is expected that such feature is observed even with other types of drones. This is a unique feature of a drone with rotor blades and it will be useful for distinguishing a drone from the other flying objects. For example, it can be expected to extract the feature of the range profile using Hough transform as in [8]. Fig. 3(b) shows the measured range profile at $BW = 3$ GHz when the drone is flying round-trip. We can see that the change in the distance due to the drone moves is detected accurately. Since the operator was fine-tuning the drone’s direction of flying, there are variations in the distance. The measured range profiles at $BW = 1$ GHz and 0.5 GHz are shown in
Figs. 3(c) and 3(d), respectively, when the drone is hovering. In contrast to the case in Fig. 3(a), the echoes reflected from the rotor blades cannot be separated with these bandwidths. This is due to degradation of range resolution. From the above measurement results, although it depends on the size of drones, the bandwidth (range resolution) of 3 GHz or more is probably necessary in order to obtain the unique feature of a drone.

4 Conclusion

In this letter, we have proposed the use of an UWB radar for drone detection and investigated its feasibility experimentally. Firstly, a typical quadrotor drone’s radar cross section (RCS) that is an important value for radar system design was measured. As a result, we found that the average value of the RCS was $-13$ dBsm (@ 24 GHz). Secondly, range profile measurements for a flying drone were conducted. As a result of the experiment, we confirmed that the UWB radar with high range resolution could detect a flying drone and observe its unique feature that was produced by rotor blades. This feature must be useful for distinguishing a drone from other flying objects such as birds and insects.

In future research, we will investigate a technique to distinguish between a drone and the other flying objects, and the detection performance when millimeter wave band (@ 79 GHz) is used. We will also investigate whether the unique feature can be observed for other types of drones, as well as the drone used in this experiment.
Partial scrambling selected mapping for PAPR reduction of OFDM signals

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Abstract: To reduce peak-to-average-power ratio (PAPR) of OFDM signals, a self-synchronized-scrambler-based selected-mapping (SS-SLM) has been investigated, where the transmit sequence is scrambled to reduce PAPR. In this method, the receiver needs to use descrambler to recover the original data. However, the descrambling causes error propagation which degrades bit error rate (BER). In this paper, we propose a partially self-synchronized-scrambler-based selected-mapping (PSS-SLM) which scrambles only most significant bits allocated to I-phase and Q-phase of QAM symbols. Numerical results show that the proposed PSS-SLM achieves better BER than that of SS-SLM in QAM-OFDM systems while keeping the same PAPR reduction capability as SS-SLM.

Keywords: OFDM, peak-to-average power ratio (PAPR), selected mapping (SLM), scrambler

Classification: Wireless Communication Technologies

References

1 Introduction

Orthogonal frequency division multiplexing (OFDM) is an effective scheme for wireless transmission over frequency-selective fading channels [1]. One of the major drawbacks in OFDM is its high peak-to-average power ratio (PAPR) which causes nonlinear distortion at the output of power amplifier. To reduce PAPR of OFDM signals, several approaches have been presented such as clipping-and-filtering (C&F) and subcarrier-phase-control [2, 3, 4, 5]. C&F is a simple technique. However, it causes nonlinear distortion which degrades bit error rate (BER) especially in the case with higher order quadrature amplitude modulation (QAM) [3].

One of the latter approaches based on subcarrier-phase-control is selected-mapping (SLM) that reduces the probability of high peak power occurrence [4]. In SLM, the transmit sequence is converted to multiple randomized sequences and the best candidate that minimizes the PAPR is transmitted. However, the receiver needs to recover the original information from the received randomized sequence, e.g., by sending redundant side information to the receiver. As a method to recover the original data without explicit side information, a self-synchronized-scrambler-based SLM (SS-SLM) has been proposed [5], where the transmit data sequence is randomized using a self-synchronized-scrambler. In this method, the receiver needs to use the corresponding descrambler to recover the original data from the received scrambled sequence. However, when transmission errors occur, the descrambling process causes error propagation which increases BER.

In this paper, we propose a partially self-synchronized-scrambler-based selected-mapping (PSS-SLM) where only significant bits of QAM symbols are scrambled to reduce PAPR of OFDM signal. In the proposed PSS-SLM, error propagation occurred at the descrambler is effectively mitigated when higher order modulation is used, while keeping the same PAPR reduction capability. We approximately analyze the bit error rates (BERs) of OFDM systems with PSS-SLM and clarify performance improvements of the proposed approach compared with SS-SLM.

2 Partially self-synchronized scrambler-based SLM method

Fig. 1(a) shows a block diagram of OFDM system with the proposed PSS-SLM. Similarly to the general structure in SLM [4], the transmit signal is converted to multiple randomized sequences by using a self-synchronized scrambler. Structures of the scrambler and its descrambler are illustrated in Fig. 1(b), where the same polynomial function is used in both the scrambler and the descrambler [6]. On the transmitter side, the best signal which exhibits the minimum PAPR is selected among $U$ candidates and then transmitted. Unlike the conventional SS-SLM that scrambles entire bits of transmit sequence, the proposed PSS-SLM works so that only the most significant bits (MSBs) of in-phase (I-phase) and quadrature-phase (Q-phase) components of each QAM symbol are transformed to a randomized one. Note that the remaining bits are not scrambled and no need exists to descramble them on the receiver side. After demodulating the received symbols at the receiver, only MSBs are descrambled to obtain the original information. Since the bit error
probability for the MSBs is lower than those of the other remaining lower order bits, the probability of error propagation occurrence can be effectively decreased using the proposed method.

The details are explained as follows. Let \( a = (a_1, \cdots, a_{N_b}) \) and \( s^{(u)} = (s^{(u)}_1 \cdots s^{(u)}_{N_r}) \) denote the original binary data sequence and the redundant header bits, respectively. Here, \( N_b \) denotes the number of information bits per data-block. \( N_r \) is the header size. Note that the header bits \( s^{(u)} \) has the same role as side-information in the general SLM. \( u \) denotes the candidate index. The maximum number of the candidates is given as \( U = 2^{N_r} \). In Fig. 1(a), the \( u \)-th header \( s^{(u)} \) is inserted into the head of the original binary sequence \( a \) so that the \( u \)-th input sequence of the scrambler is given as \( a^{(u)} = (s^{(u)} a) \), \( u = 1, \cdots, U \). Then, each input sequence \( a^{(u)} \) is randomized by the scrambler, where the input sequence length is \( N_b + N_r \). Since the scrambler uses a feedback-type structure, \( U \) different randomized sequences \( \{a^{(u)} = ( \hat{a}^{(u)}_1 \cdots \hat{a}^{(u)}_{N_b+N_r}) \} \), \( u = 1, \cdots, U \), are generated with \( U \) different header inputs followed by the same original data. The \( u \)-th candidate \( \hat{a}^{(u)} \) is mapped onto QAM symbols as \( X^{(u)} = (X^{(u)}_1 \cdots X^{(u)}_N) \), where \( N \) denotes the number of subcarriers. \( X^{(u)} \) is modulated with inverse fast Fourier transform (IFFT) to generate an OFDM candidate signal \( x^{(u)} = (x^{(u)}_1 \cdots x^{(u)}_N) \), \( u = 1, \cdots, U \). The best OFDM signal which exhibits the minimum PAPR is selected among \( U \) candidates and transmitted.

Figs. 2(a) and (b) show examples of Gray-coded 16QAM symbol constellations in case with PSS-SLM and SS-SLM, respectively. In these figures, red and blue circles denote original signal point (before scrambling) and those after scrambling, respectively. As illustrated in these figures, when only MSBs are scrambled with PSS-SLM, the original signal point is transited to one of the four
constellation points which correspond to the number of MSBs of I-phase and Q-phase (i.e., 2 bits). In contrast, after being scrambled with SS-SLM, the original signal point is transited to one of all sixteen constellation points. Note that \( U \) different candidate sequences can be generated even when only MSBs are scrambled as far as the number of header bits \( N_r \) is greater than or equal to \( \log_2 U \). Thus, we can expect that the proposed PSS-SLM provides good PAPR reduction capability comparable to the conventional SS-SLM.

On the receiver side, after the demodulation, the information data bits are recovered by descrambling the received randomized sequence if no transmission error occurs, where the descrambler uses the same polynomial function as the scrambler. If transmission error occurs, the descrambling process causes error propagation which degrades BER performance as discussed in the next section.

### 3 Bit error rate analysis

In this section, we consider Gray-coded 16QAM as an example of the subcarrier modulation schemes. Note that discussions in this section can be extended to other QAM cases. Let \( M \) be the number of taps in the scrambler with a given scrambler polynomial function. BERs of higher order bit and lower order bit of Gray-coded 16QAM are given respectively as

\[
P_H = \int_{-\infty}^{0} (p(x) + q(x)) \, dx
\]

\[
P_L = \int_{-\infty}^{-2\delta} p(x) \, dx + \int_{2\delta}^{\infty} p(x) \, dx + \int_{-2\delta}^{2\delta} q(x) \, dx,
\]

where

---

**Fig. 2.** Constellations and possible transitions of Gray-coded 16QAM with SS-SLM and PSS-SLM.
\[ p(x) = \frac{1}{\sqrt{2\pi\sigma}} \exp\left(\frac{(x-\delta)^2}{2\sigma^2}\right), \quad q(x) = \frac{1}{\sqrt{2\pi\sigma}} \exp\left(\frac{(x-3\delta)^2}{2\sigma^2}\right), \]

and \(2\delta\) denotes distance between two neighboring constellation points as illustrated in Fig. 2(a). Here, \(\frac{E_b}{N_0} = \frac{\frac{\gamma}{2\sigma^2}}{2\sigma^2}\). Supposing that the descrambling process causes error propagation and increases the number of bit errors by \(M\) times greater than the ideal case (i.e., without error propagation), we can approximately give the theoretical BER of Gray-coded 16QAM with PSS-SLM in AWGN condition as

\[ P(M) \approx \frac{M}{2} P_H + \frac{1}{2} P_L, \quad (3) \]

where the first and second terms of right-hand side are denoted respectively as error probabilities of higher and lower order bits in 16QAM symbols. Equation (3) shows that error probabilities of the highest order bits in I-phase and Q-phase of each modulated symbol are increased by \(M\) times compared to the ideal case. Since error rate of the highest order bits is less than that of other lower order bits (i.e., \(P_H < P_L\)), it is clear that scrambling higher order bits is more effective approach than that of scrambling lower order bits.

On the other hand, we can approximately give the theoretical BER of Gray-coded 16QAM with SS-SLM in AWGN condition as

\[ Q(M) \approx \frac{M}{2} (P_H + P_L). \quad (4) \]

Unlike the proposed PSS-SLM, Eq. (4) shows that the conventional SS-SLM increases error probabilities of both higher order bits and lower order bits by \(M\) times than the ideal case.

4 Performance evaluation

We evaluate the performance of OFDM systems with the proposed PSS-SLM. The number of subcarriers is 64. Subcarrier modulation scheme is QPSK, 16QAM, 64QAM, and 256QAM. Block-diagram of OFDM system is the same as that shown in Fig. 1(a). The polynomial function used at the scrambler is \(g(z) = z^6 + z^4 + 1\).

The number of header bits is \(N_r = 2, 4, \) and 6 bits, which correspond to the number of the candidates in SLM, \(U = 2^{N_r} = 4, 16, \) and 64, respectively.

We assess statistical characteristic of the instantaneous power of OFDM signals by using CCDF defined as a complementary function of cumulative distribution function. Fig. 3(a) plots CCDF of the normalized instantaneous power of the OFDM signal with the proposed PSS-SLM. For comparison purpose, we also show those of original OFDM without any PAPR reduction and case with SS-SLM, respectively. From the result, we can see that the instantaneous power at CCDF \(= 10^{-4}\) is reduced using the PSS-SLM by about 1.9 dB, 2.9 dB, and 3.4 dB compared with the original OFDM case, when \(U = 4, 16, \) and 64, respectively. In this figure, the same results are obtained for cases with different subcarrier modulations when \(U\) is fixed. We can confirm that the proposed method achieves almost the same amount of PAPR reduction as cases with the SS-SLM.

Fig. 3(b) shows BER performance of OFDM systems with PSS-SLM and SS-SLM, where QPSK/16QAM/64QAM/256QAM are used. For comparison, BER
From this figure, we can confirm that theoretical BERs of OFDM with PSS-SLM and SS-SLM show good agreement with its simulation results. It can also be seen that the proposed PSS-SLM achieves better BER performance than SS-SLM in the case with 16/64/256QAM, because error propagation at the descrambler is mitigated using PSS-SLM. Note that PSS-SLM is equivalent to SS-SLM when performance of ideal case, i.e. in which error propagation never occurs, is plotted.

**Fig. 3.** Performance of OFDM system with the proposed method, where subcarrier modulation schemes are QPSK, 16QAM, 64QAM, and 256QAM.
QPSK-OFDM is used. The result shows that BER of PSS-SLM is improved as the modulation order increases. Thus, the proposed method is more effective for OFDM systems using higher order QAM.

5 Conclusion

In this paper, we have proposed a PSS-SLM where only MSBs of QAM symbols are scrambled to reduce PAPR of OFDM signal. We confirmed that theoretical BERs of OFDM with PSS-SLM show good agreements with simulation results. The results demonstrated that the proposed PSS-SLM achieves better BER than that of SS-SLM in QAM-OFDM systems while keeping the same PAPR reduction capability.

Acknowledgments

This research was partially supported by JSPS KAKENHI (17K06427) and the Telecommunications Advancement Foundation.
Wideband waveguide antenna using stepped L-shaped probe for wide-angle circular polarization radiation

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Abstract: A circularly polarized (CP) waveguide antenna using L-shaped probe is presented. The proposed antenna can cover the UWB (Ultra Wideband) high-band (34.3%: 7.25 GHz–10.25 GHz in Japan) with 3-dB axial ratio (AR) having a wide range of angle in the radiation patterns. For obtaining this performance, a stepped structure is introduced in the L-shaped probe. The optimization of the stepped structure can control the amplitude ratio of the two orthogonal modes of electric field. This can enhance the AR bandwidth of the previously proposed waveguide antenna using an L-shaped probe and parabolic short wall. Finally, the proposed structure achieves 39% of 3-dB bandwidth keeping a wide angle for radiating CP in an angle of around 135°.

Keywords: circular polarization, wideband antenna, waveguide antenna, L-shaped probe, axial ratio

Classification: Antennas and Propagation

References

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1 Introduction

Circularly polarized (CP) antennas have advantages, such as reducing effects of multipath fading and no need of considering antenna alignment. This is because the amplitude of electric field is constant at any angle in the wave front, and the sense of CP is switched from left-hand (LH) CP to right-hand (RH) CP and vice versa as long as the incident angle is less than the Brewster angle. To make the maximum use of such advantages, the cross-polarization (XPOL) should be reduced to decrease the AR.

Furthermore, broadband antennas are required to increase channel capacity in wireless telecommunications. Some CP waveguide antennas using L-shaped probes have been recently reported [1, 2, 3, 4, 5, 6]. The L-shaped probe is bent at a bending point to make the 90-degree phase difference between the two orthogonal modes. However, its asymmetrical structure easily generates XPOL in the off-boresight directions. In order to reduce the XPOL to widen the angle for radiating CP, a waveguide CP antenna with a parabolic shorting wall [4] was proposed. To reduce the AR in this structure, a narrowed aperture has been also proposed in [5]. This structure has low AR (<1 dB) and wide angle for CP, however the 3-dB AR bandwidth is narrower than UWB highband and the AR bandwidth of previous structures [4]. This letter presents that an optimized stepped structure of the L-shaped probe is effective to obtain a wideband AR characteristics for UWB High-band in the waveguide antenna keeping sufficient low XPOL in off-boresight directions.

2 Antenna structures and characteristics

Fig. 1 shows the proposed antenna structure with a narrowed aperture and a parabolic wall as a short wall. For this antenna, the feeding L-shaped probe has a stepped structure in the probe structure with a size of the narrower part of $K$. In addition to this, $R$ is the diameter of the thick part of probe. Furthermore, a stepped aperture is introduced in order to reduce higher order modes [3]. The proposed antenna is fed through an inductive coaxial structure so as to cancel the capacitive impedance. Considering the required UWB high band from 7.25–10.25 GHz, the
waveguide diameter has been set at 27.0 mm in order to have a cutoff frequency at 6 GHz. For the center frequency at 8.75 GHz, the length between the probe and the tip of the parabolic wall should be around $\lambda_g/4$, where the $\lambda_g$ is the effective wavelength. The length of two sides of the probe should be chosen in order to generate the TE$_{11}$ mode crossing spatially at the right angle having a 90-degree phase difference. The narrowed aperture structure is effective to reduce the AR and widen the angle for CP in the radiation pattern. This design has been discussed in [5].

In addition to this, the propose antenna has three techniques to reduce XPOL in a wideband frequency. At first, a parabolic short wall with $x = 0.8(y^2 + z^2)$ is installed based on the parabolic shorting wall reported in [4]. This reduces XPOL in wide angle ranges in the radiation patterns, because the variation in distance between the bent point of L-shaped probe and shorting wall can be almost constant with respect to any angle centered at a bent point. As the second, the entire length $L$ along $x$ direction is optimized to be chosen at 43.7 mm considering the AR behavior. With a decrease in $L$ from 57.7 mm to 43.7 mm, AR in higher frequency band around 9.40 GHz is gradually reduced with a sinusoidal variation. As the third technique, the narrowed aperture is used, that is, we have made the diameter of aperture $D$ smaller to be 27.6 mm than the inside diameter [5]. The probe structure in Fig. 1(b) is identical with the probe in [5] when $K = 13.5$ mm and $R = 1.0$ mm. However, for $K = 7.5$ mm and $R = 1.8$ mm, −10-dB $S_{11}$ bandwidth of 44.6% (6.7–10.6 GHz) is obtained as shown in Fig. 2(a). The bandwidth has been expanded a little compared to that in [5]. AR characteristics is obtained as in Fig. 2(b) with a 3-dB AR ratio of 39% (7.0–10.4 GHz) covering the UWB high band. Choosing the $K$ and $R$ as the above values, the widest AR bandwidth is obtained. Furthermore, Fig. 2(b) also shows gain characteristics along the boresight direction. The gain of the proposed antenna has the close gain characteristics to [5]. The measured gain is slightly lower by 1 dB at most in the higher frequency. This is
mainly due to an uncertainty of the phase center regarding the proposed antenna and the standard antenna, where our standard antenna is available for ≤10 GHz.

Fig. 2(c) shows the variation in amplitude characteristics as a function of $K$ with $R = 1.8$ mm. An increase in $K$ from 4.5 mm to 7.5 mm controls mainly $E_{\theta}$ in the lower frequency. Furthermore, with an increase in $R$ from 1.2 mm to 1.8 mm, the amplitude of $E_{\phi}$ increases in the far field as shown in Fig. 2(d). The both

(a) Simulated and measured $S_{11}$ characteristics with simulated characteristics in the reference [5].

(b) Simulated and measured AR and gain characteristics with simulated characteristics in the reference [5].

(c) Amplitude characteristics of $E_{\theta}$ and $E_{\phi}$ as a function of $K$ ($R = 1.8$ mm).

(d) Amplitude characteristics of $E_{\theta}$ and $E_{\phi}$ as a function of $R$ ($K = 7.5$ mm).

(e) Radiation patterns at 8 GHz and 9 GHz.

Fig. 2. Antenna performances.
amplitude characteristics are sensitive to $R$ throughout the frequency. As a result, choosing the suitable $K$ and $R$, we can obtain the widest AR bandwidth of 39\% (7.0–10.4 GHz) with $R = 1.8$ mm and $K = 7.5$ mm covering the UWB high-band.

3 Radiation patterns

Fig. 2(e) shows radiation patterns at 8 GHz and 9 GHz for both $xy$- and $zx$-planes. The measured results (mea.) shows good agreements with simulated (sim.) results. For 3-dB AR, XPOL should be smaller than co-polarization by more than 15 dB. Considering this, 3-dB AR is obtained covering 150° and 160° in $xy$- and $zx$-planes, respectively, at 8 GHz. At 9 GHz, 135° and 125° are available for CP in $xy$- and $zx$-planes, respectively. Although the bandwidth for 3-dB AR has been expanded, the obtained angles are comparable with those in [5].

4 Comparison of antenna performances

As discussed above, the antenna performance is sensitive to the probe structure. The performance of the discussed structure is compared with some other CP cavity-type or waveguide antennas [5, 7, 8, 9]. However, some have high gain with a horn, and others have low gain. For a comparison with a fair condition, we try discussing the comparison using products of antenna gain $G$ and angle $\theta_c$ (in $xy$-plane for [5] and the proposed structure) in which CP with the 3-dB AR is radiated, as we know empirically that these two parameters have a trade-off relation mutually. The performances of the related antennas are summarized in the Table I, where the frequency is a representative value around the center frequency of operating band, and the ARBW is the 3-dB bandwidth of AR in \% at the center frequency of the respective bandwidths.

<table>
<thead>
<tr>
<th>Ref.</th>
<th>Freq. [GHz]</th>
<th>ARBW [%]</th>
<th>$G$ [dBic]</th>
<th>$\theta_c$ [°]</th>
<th>$G\theta_c$ [°]</th>
</tr>
</thead>
<tbody>
<tr>
<td>[7]</td>
<td>20</td>
<td>-</td>
<td>14.5</td>
<td>30</td>
<td>848.5</td>
</tr>
<tr>
<td>[8]</td>
<td>-</td>
<td>35</td>
<td>7.8</td>
<td>140</td>
<td>843.5</td>
</tr>
<tr>
<td>[9]</td>
<td>2.2</td>
<td>26</td>
<td>15</td>
<td>34</td>
<td>1075.2</td>
</tr>
<tr>
<td>[5]</td>
<td>9.0</td>
<td>35</td>
<td>9.6</td>
<td>146</td>
<td>1331.5</td>
</tr>
<tr>
<td>Proposed</td>
<td>9.0</td>
<td>39</td>
<td>9.3</td>
<td>135</td>
<td>1149.0</td>
</tr>
</tbody>
</table>

In the table, the antenna in [7] is fed with a 2-point feeding structure, and the antenna in [8] is fed with a 4-point feeding structure and have a wider AR bandwidth than [7], however their $G\theta_c$ products are close to each other but no more than 850 degrees. An antenna with a truncated horn antenna and 1-point feeding structure has a higher $G\theta_c$ product [9]. Compared these antennas, the antenna with an L-shaped probe in [5] has a higher $G\theta_c$ product with a wider AR bandwidth which is comparable with the AR in [8] having the 4-point feeding. The proposed antenna has a little bit smaller $G\theta_c$ product than [5], however the proposed antenna can achieve a wider AR bandwidth with a small sacrifice of $G$. 

© IEICE 2017
DOI: 10.1587/comex.2017XBL0085
Received May 24, 2017
Accepted June 8, 2017
Publicized June 23, 2017
Copyedited September 1, 2017

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Furthermore, according to the comparison in Fig. 2(b), the proposed antenna has higher ARs but the highest one is still less than 3 dB in the required frequency. From the discussion above, we can choose the better AR or $G\theta_\epsilon$ by optimizing the probe structure. As a result, the proposed antenna is the only structure that can cover UWB high band and high $G\theta_\epsilon$ product in cavity-type antennas.

5 Conclusion

In this letter, a design technique of the L-shaped probe is presented to improve the AR bandwidth of the waveguide antenna in [5]. The probe design can change the amplitude characteristics of the orthogonal electric field for CP. The proposed probe structure can achieve CP in a wideband frequency range of 39% for covering UWB high-band (7.25–10.25 GHz in Japan) keeping a wide range of angles around 120°–150° in the radiation patterns.

Acknowledgments

This work was supported by the Ministry of Education, Culture, Sports Science and Technology (MEXT) Grant-in-Aid for Scientific Research (C), and the Telecommunication Advancement Foundation.
Via-less and small-radiation waveguide to microstrip line transition for millimeter wave radar modules

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Abstract: We propose a waveguide to microstrip line transition which perpendicularly connects one waveguide and two microstrip lines. Because the proposed transition consists of only a waveguide and a dielectric substrate, the configuration can be simplified. Additionally the transition does not need via holes on the substrate. This configuration reduces manufacturing complexity and cost. We assume to co-locate the transition and a patch array antenna on the dielectric substrate for millimeter wave automotive radar modules. Because the proposed transition suppress undesired radiation by multi-stage impedance transformers, the desired radiation pattern from the array is not worsen. The small radiation from the transition is shown by simulation. Additionally we fabricate 4 × 4 planar patch array antenna with the proposed transition. Measured and simulated radiation patterns are introduced.

Keywords: waveguide-microstrip line transition, patch array antennas

Classification: Antennas and Propagation

References
1 Introduction

For safety driving and cruise control of vehicles, various sensors have been adopted. Millimeter wave (mm-wave) radars have advantages of long sensing distance, precise ranging and applicability at night. This paper proposes a waveguide (WG) to microstrip line (MSL) transition which is an essential component for mm-wave radars.

Fig. 1 shows mm-wave module’s configurations. The module means a hardware transmitting/receiving mm-wave which includes frequency conversion and signal processing. Fig. 1(a) is the typical configuration. Antennas are connected to one RF-IC transmitting/receiving millimeter wave signal. Unfortunately this configuration demands long MSL. It seriously attenuates mm-wave. On the other hand, we have developed the modules like Fig. 1(b). WG-MSL transitions are placed for each channel. Additionally each WG formed in PWB is connected to RF-IC at just behind the transition as per Fig. 1(c). Because this configuration drastically reduces MSL length, mm-wave signal’s loss can be minimized. It improves SNR (signal to noise ratio) and sensing range. Additionally this configuration can choose arbitral number and position of T/Rx channels with minimum loss.

This configuration needs the WG-MSL transition which perpendicularly connects the WG and the MSL. Though this type of transition basically requires a backshort for the WG, backshort-less transitions have been proposed in [1, 2]. It reduces complexity and manufacturing cost. [1, 2] also eliminated via connecting both sides of the substrate because virtual electrical short is realized by a choke [1] or open stubs [2]. Via-less is strong advantage for cost reduction and reliability improvement against a potential risk of breaking via. However the chokes and the stubs generally cause radiation at their edge. In this paper, we also propose a backshort-less and via-less WG-MSL transition. Additionally we eliminate the chokes or the stubs; the proposed transition consists of only a WG and a MSL on a dielectric substrate. This concept is similar to an aperture-coupled patch antenna [3] that the patch that the MSL are electromagnetically coupled through the slot. In our proposal, the patch is replaced with the WG.

![Fig. 1. Module configurations](image-url)
In automotive radar applications, the WG-MSL transition and the patch array antennas are co-located as per Fig. 1(b). In such situation, undesired radiation from the transition should be reduced in order to maintain required radiation pattern. We adopt multi stage impedance transformers to improve transmitting efficiency and to reduce undesired radiation. We proposed basic structure of the proposed transition in [4]. In this paper, we introduce the detail and the performance comparison. Additionally, we fabricate 4 × 4 planar patch array antenna with the proposed transition and demonstrate the radiation pattern.

2 Proposed waveguide-microstrip line transition

Fig. 2(a) shows the proposed WG-MSL transition. It consists of a straight waveguide and a dielectric substrate with copper foils. Via holes are not needed for the substrate. In the upper copper foil, only a MSL with impedance transformers is formed. The lower copper foil is a ground except a slot. The WG and the MSL are coupled through the slot.

Fig. 2(b) shows the top view. H-shape slot is used to improve transmission from WG to MSL by concentrating the electric field in center. Though Trans A is one of impedance transformers, the main purpose is to change mm-wave in the WG to the MSL. The width of Trans A is relatively wide. It may not be reasonable for impedance matching because the wide MSL yields lower impedance while the impedance of the WG is high. Therefore narrow MSL is suitable for impedance matching, the structural difference between the MSL and the slot causes undesired radiation from the slot. It also worsens the transmission efficiency (S21, S31). On the other hand, very wide Trans A forces drastic impedance change to final MSL (0.04`). Because large impedance change leads structural change in the MSL section, it also causes undesired radiation. Therefore the width of Trans A is determined by numerical simulation in order to reduce the radiation. Remaining Trans B, C and D are impedance transformers to connect between Trans A and final MSL. Though the MSL width (impedance) variation, Trans B (wide, low impedance) - Trans C (narrow, high impedance) - Trans D (wide, low impedance), is not reasonable for impedance matching, they reduce structural difference and the undesired radiation. The proposed structure narrows frequency bandwidth. However the impact is small because the bandwidth of the proposed transition supports the bandwidth of patch antennas. If the undesired radiation is tolerable, the impedance matching without Trans C and D is possible as per Fig. 2(c). This structure is used for the performance comparison.

We simulated the proposed transition of Fig. 2(a) and (b). The model has one WG port and two MSL ports. The upper side of the substrate is filled with air which is not displayed. Relative permittivity of the substrate is around 3. Dielectric loss and conductor loss are not included. In order to compare the performance, we also simulated the 2-stage transformer in Fig. 2(c). In the model, Trans C and D were not used, only Trans B was adjusted for the impedance matching. Fig. 2(d) and (e) show the simulated reflection coefficient and transmission, respectively. A blue plot is the proposed structure, while a red plot is the case of 2-stage transformers for the comparison. The latter case yields broad frequency characteristic because of natural
Impedance transform. In Fig. 2(e), the proposed transition yielded superior S21 near center frequency because the undesired radiation is suppressed. Because dielectric and conductor losses are not included in the simulation, radiation loss is calculated by $1 - (|S11|^2 + |S21|^2 + |S31|^2)$. The radiation loss of the proposed transition was $-17$ dB while that of the comparison was $-14$ dB.

Fig. 2. Proposed WG-MSL transformer and performance comparison
The undesired radiation pattern on x-z plane is shown in Fig. 2(f). The radiation of the proposal is smaller than the comparison across most of angles. The undesired radiation of the proposal is drastically suppressed on y-z plane as per Fig. 2(g). The wave sources of the undesired radiation which are edges of MSL and the slot align a line on x-axis and they form a kind of a linear array. Therefore if the radiation at $\theta = 0 \text{deg}$ is suppressed, the radiation is able to suppressed across all angles on y-z plane. Worst undesired radiation of the proposal is $-10.7 \text{dBi}$. It is acceptable because desired gain from the patch array explained in next section is sufficiently high and small radiation from the transition does not affect total radiation pattern including antennas.

3 Fabrication of patch array with proposed transition

We demonstrate the proposed transition. A planar patch array antenna was connected to the transition because the stand-alone transition needed 3-port measurement (1-WG + 2-MSLs) and the measurement was difficult. Therefore we evaluated total antenna performance including the transition and the antenna. We fabricated $4 \times 4$ planar patch array antenna with the proposed transition as per Fig. 3(a). The transition and the antenna are co-located on the substrate like Fig. 1(b). The element number and the element spacing are just one example for demonstration. The MSL corresponding to Trans C in Fig. 2(b) was bent to connect the patch array. Though the bends slightly increase the undesired radiation, arbitrary bends are possible within tolerable undesired radiation. MSL feeding circuit from the transition to each row of the array was optimized so as to excite them with equal amplitude and co-phase. Though the proposed transition outputs reversed phase to two MSLs, the reversal is also compensated by MSL length. The slot on the reverse copper layer and the WG at reverse side were not displayed.

The measured and simulated reflection coefficients are shown in Fig. 3(b). Both approximately matched with small shift about 0.5%. The possible factor of the discrepancy is permittivity difference between the simulation and the actual material. Fig. 3(c) shows the radiation pattern on vertical (x-z) plane. The side lobe level is higher than that of horizontal (y-z) plane because of uniform excitation. The measured pattern and the simulated pattern roughly matched even if upper two rows and lower two rows were connected to different outputs of the transition. It means that the phase reversal of the transition was compensated and all elements were excited with co-phase. Fig. 3(d) shows the radiation pattern on horizontal (y-z) plane. Because amplitude taper was applied to each 4-element array on y-direction, this pattern yielded the low sidelobe. The measured pattern and the simulated pattern roughly matched and it exhibits that the proposed transition successfully worked. Though it is difficult to divide the desired radiation from patch antennas and the undesired radiation from the transformer, the result implies that the undesired radiation was small. A possible factor of the discrepancy of the measured and the simulated patterns comes from structural difference; an infinite ground was used in the simulation while actual antenna had ground edges.

We also simulated the $4 \times 4$ array with 2-stage transformer which is not shown in this paper. Though the gain was slightly lower than the proposal, radiation
The pattern difference was relatively small. The sidelobe level in the condition of the $4 \times 4$ array was sufficiently higher than the undesired radiation from the transition, e.g., the simulated sidelobe was $-5$ dBi at 35 deg in Fig. 3(d) while the undesired radiation was $-28$ dBi (proposed)/$-12$ dBi (2-stage transformer) in Fig. 2(g). When antennas with low side lobe are needed, especially the required side lobe level is lower than the undesired radiation of 2-stage transformer, the proposed transition becomes more effective. Though we demonstrated the proposed transition in this paper, confirming proposal’s potential of very low side lobe is future work.

### 4 Conclusion

We proposed a waveguide to microstrip line transition without via holes. We simplified the structure by eliminating electrical short structures like chokes and stubs. This configuration leads cost reduction and reliability improvement. Un-desired radiation from the transition could be reduced by using only impedance transformers with small structural variations. Small undesired radiation was confirmed by simulation. Additionally we fabricated $4 \times 4$ planar patch array antenna with the proposed transition and confirmed that the simulated and measured radiation patterns approximately matched.
Model for estimating the properties of mechanically induced long-period fiber grating based on polarization and applied pressure

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Abstract: This study proposes a model for estimating the properties of mechanically induced long-period fiber gratings (MLPFGs) based on polarization and applied pressure. The model arises from the linear relation between the applied pressure and coupling coefficients of MLPFGs through utilization of Jones matrix analysis. The transmission property of a fabricated MLPFG can be easily estimated using the relation between the applied pressure and coupling coefficients obtained from the measured spectra of MLPFGs. The estimated spectra around the resonance wavelengths based on our proposed model show good agreement with the measured ones.

Keywords: long-period fiber grating, polarization dependence, applied pressure dependence

Classification: Sensing

References


1 Introduction

Mechanically induced long-period fiber gratings (MLPFGs) have an attractive property; their transmission spectra strongly depend on the applied pressure from their surroundings because the periodic refractive index modulations of MLPFGs vary directly with pressure [1]. Moreover, MLPFGs have unique advantages including simple fabrication, immunity to electromagnetic interference, resistance against fire or sparks, and compatibility with remote and multiplex operation [2]. Therefore, MLPFGs have been extensively studied for use in cost-effective and highly functional optical sensors or devices [2, 3, 4, 5].

The transmission spectrum of a MLPFG strongly depends on the state of polarization (SOP) of the incident light applied to the MLPFG [6, 7]. Therefore, for estimating the polarization dependence of resonance loss in MLPFGs, considering the optical sensors or devices based on MLPFGs is important. When designing MLPFGs, it is also important to understand the relation between the applied pressure and the induced change in the refractive index in the core of the fiber [8]. However, characterizing this relation is difficult because it strongly depends on the material properties of the MLPFG, such as that of the fibers, coating materials, and pressurized substances, as well as on the placement of the materials.

In this study, we propose a model for estimating the properties of an MLPFG based on polarization and applied pressure. Our model arises from the linear relation between applied pressure and coupling coefficients of MLPFGs and utilizes Jones matrix analysis [9]. The transmission spectrum of the fabricated MLPFG at any applied pressure and SOP of the incident light can be easily estimated from the measured spectra of the MLPFG using our model.

2 Proposed model for estimating properties based on polarization and applied pressure

Propagation through the MLPFG can be expressed by a Jones matrix, $[J_p]$, as follows:
where $\phi_p$ denotes the induced relative phase difference between the x- and y-polarization. $a_x$ and $a_y$ are the transmittances for the x- and y-polarized electric fields, respectively, at a wavelength of $\lambda$. In the followings subscript $i$ ($i = x$ or $y$) denotes the value for $i$-polarization.

The transmitted power through the MLPFG can be expressed by the following equation [8]:

$$|a_i|^2 = 1 - \frac{\sin^2(\kappa_i L(1 + (\sigma'_i/\kappa_i)^2)^{0.5})}{(1 + (\sigma'_i/\kappa_i)^2)} \quad (i = x, y),$$

where $L$ is the length of the MLPFG. $\kappa$ is the ac cross-coupling coefficient between the core and cladding modes [8] and $\sigma'$ is a general dc self-coupling coefficient [8] defined as

$$\sigma'_i = \frac{1}{2}(\beta_{ei} - \beta_{ci}) - \frac{\pi}{\Lambda} + \sigma_i = \frac{\pi \Delta n_{eff,i}}{\lambda} - \frac{\pi}{\Lambda} + \sigma_i \quad (i = x, y),$$

where $\beta_{ei}$ and $\beta_{ci}$ are the propagation constants of the core and cladding modes, respectively. $\sigma$ is the dc coupling coefficient [8]. $A$ and $\Delta n_{eff}$ are the grating pitch and effective index difference between the core and cladding modes, respectively.

Here we assume that the wavelength dependence of the coupling coefficients $\kappa$, $\sigma$, and $\Delta n_{eff}$ is small and that these coefficients change linearly with the applied pressure, $P$ [8, 9]. Thus, each coefficient can be represented as a function of the applied pressure as follows:

$$\kappa_i = A_i P + B_i,$$
$$\sigma_i = C_i P + D_i,$$
and
$$\Delta n_{eff,i} = E_i P + F_i \quad (i = x, y, A, B, C, D, E, F : \text{const}).$$

The minimum transmittance of the MLPFG is

$$|a_i|^2_{\min} = 1 - \sin^2(\kappa_i L) \quad (i = x, y),$$

which occurs at the wavelength $\lambda_p$ when

$$\sigma'_i = \frac{\pi \Delta n_{eff,i}}{\lambda_p} - \frac{\pi}{\Lambda} + \sigma_i = 0 \quad (i = x, y).$$

The maximum transmittance of the MLPFG occurs at $\lambda_0$ when

$$(\sigma'_i)^2 + \kappa_i^2)^{0.5}L = \pm \pi \quad (i = x, y),$$

and therefore,

$$\frac{\pi \Delta n_{eff,i}}{\lambda_0,i} - \frac{\pi}{\Lambda} + \sigma_i = \left(\frac{\pi^2}{L^2} - \kappa_i^2\right)^{1/2} \quad (i = x, y).$$

From (8) and (10), the following equations hold:

$$\sigma_i = \frac{\lambda_0,i - \lambda_{p,i}}{\lambda_0,i - \lambda_{p,i}} \left(\frac{\pi^2}{L^2} - \kappa_i^2\right)^{1/2} + \frac{\pi}{\Lambda} \quad (i = x, y).$$
\[ \Delta n_{\text{eff},j} = \left( \frac{\pi}{\lambda} - \sigma_j \right) \frac{\lambda_{p,j}}{\pi} \ (i = x, y). \tag{12} \]

The Jones vector of the incident light, \([E_{\text{in}}]\), can be written as
\[
[E_{\text{in}}]' = \begin{bmatrix} \cos \theta \exp \left( j \frac{\phi_{\text{in}}}{2} \right) & \sin \theta \exp \left( -j \frac{\phi_{\text{in}}}{2} \right) \end{bmatrix}. \tag{13} \]

The Jones vector of the output light from the MLPFG, \([E_{\text{out}}]\), and the transmitted power through the MLPFG, \(I_{\text{out}}\), can be expressed using the following equations:
\[
[E_{\text{out}}] = [J_p][E_{\text{in}}] = \begin{bmatrix} \alpha_x(\lambda) \cos \theta \exp \left( j \frac{\phi_p(\lambda) + \phi_{\text{in}}}{2} \right) \\ \alpha_y(\lambda) \sin \theta \exp \left( -j \frac{\phi_p(\lambda) + \phi_{\text{in}}}{2} \right) \end{bmatrix} \tag{14} \]
\[
I_{\text{out}} = |\alpha_x(\lambda) \cos \theta|^2 + |\alpha_y(\lambda) \sin \theta|^2. \tag{15} \]

Unpolarized light, such as white light, vibrates in any direction at a given time. Therefore, \(\phi_{\text{in}}\) and \(\theta\) vary randomly for unpolarized light. Hence, the average transmitted power for unpolarized light through the MLPFG, \(I_{\text{out dep}}\), can be written as
\[
I_{\text{out dep}} = \frac{(|\alpha_x(\lambda)|^2 + |\alpha_y(\lambda)|^2)}{2} = \sum_{i=x}^{y} \left( 2 - \frac{\sin^2(\kappa_i L (1 + (\delta_i/k_i)^2)^{0.5})}{(1 + (\delta_i/k_i)^2)} \right) / 2. \tag{16} \]

The measured ac coupling coefficient, \(\kappa\), at an applied pressure, can be determined using (7) and the measured resonance loss of the MLPFG. Thus, \(\kappa\) can be estimated as a function of applied pressure, \(P\), by determining the constants \(A\) and \(B\) in (4) using the measured ac coupling coefficient of the MLPFG. The measured dc coupling coefficient, \(\sigma\), for an applied pressure can be calculated using the measured resonance wavelength, \(\lambda_p\), minimum loss wavelength, \(\lambda_0\), measured ac coupling coefficient, \(\kappa\), and the parameters of MLPFG based on (11). Thus, \(\sigma\) can be estimated as a function of applied pressure by determining the coefficients \(C\) and \(D\) in (5) from the measured dc coupling coefficients \(\sigma\). The effective index difference, \(\Delta n_{\text{eff}}\), between the core and cladding modes can be estimated as a function of applied pressure by determining the constants \(E\) and \(F\) in (6) from \(\lambda_p\) and \(\sigma\) using (12). The transmitted power of the MLPFG for a given SOP and pressure can be estimated by combining (2), (3), (4), (5), (6), (11), (12), (13), (15), and (16).

3 Experimental results

Fig. 1(a) shows the experimental setup for measuring the transmission spectra of the MLPFG at various pressures. A MLPFG was fabricated using a standard telecommunication optical fiber (ITU-T. G.652), a metallic screw, an aluminum plate, and weights in the range of 15.0 to 30.0 kg. The pitch and the length of the screw were 500 \(\mu\text{m}\) and 5.0 cm, respectively. We measured the applied pressure dependence of the transmission spectra of the MLPFG by varying the amount of weight applied. The transmission spectra with unpolarized light were measured.
using a white light source and an optical spectrum analyzer (OSA). The transmission spectra for x- and y-polarized light were measured using a polarizer, a polarization controller, a superluminescent diode, and an OSA. Fig. 2 shows the

![Fig. 1. Experimental setup.](image)

**Fig. 1.** Experimental setup.

![Fig. 2. Simulated and experimental transmission spectra of the MLPFG at various pressures (a) for x- and y-polarized incident light and (b) for unpolarized incident light.](image)

**Fig. 2.** Simulated and experimental transmission spectra of the MLPFG at various pressures (a) for x- and y-polarized incident light and (b) for unpolarized incident light.

![Fig. 3. Pressure dependence of the coupling coefficients: (a) ac coupling coefficient, κ, (b) dc coupling coefficient, σ, and (c) effective index difference, Δn_{eff}.](image)

**Fig. 3.** Pressure dependence of the coupling coefficients: (a) ac coupling coefficient, κ, (b) dc coupling coefficient, σ, and (c) effective index difference, Δn_{eff}.
pressure dependence of the transmission spectra of the MLPFG for x- and y-polarized light and those for unpolarized light. The solid and dotted lines show the measured and simulated spectra, respectively. The red, blue, and black lines show the transmission spectra for x-polarized, y-polarized, and unpolarized light, respectively. Note that the transmission spectra of the MLPFG with the microbending loss removed are shown in Fig. 2. We estimated the coupling coefficients, $\kappa$ and $\sigma$, as a function of the applied pressure according to (4) and (5) using the measured spectra. Fig. 3 shows how $\kappa$, $\sigma$, and $\Delta n_{eff}$ vary with the pressure. The circles and crosses show the estimated parameters for x- and y-polarized light, respectively. The solid lines are fitted to the experimental data. The estimated coefficients ($A_x$, $B_x$, $A_y$, $B_y$, $C_x$, $D_x$, $C_y$, $D_y$, $E_x$, $F_x$, $E_y$, and $F_y$) were obtained (0.251, 9.18, 0.272, 8.51, −0.1296, 9478, −0.1908, 9235, 7.87 $\times$ $10^{-8}$, −0.0016, 1.40 $\times$ $10^{-7}$, and −0.0015, respectively). It is confirmed that these parameters are directly proportional to the applied pressure as expected. The spectrum around the resonance wavelengths can be successfully estimated using our model and the estimated spectra are in good agreement with those measured experimentally.

4 Conclusions

We proposed a model for estimating MLPFG properties based on polarization and applied pressure. We fabricated a simple MLPFG using an aluminum plate, a metallic screw, a standard telecommunication fiber, and some weights. Then, we measured the applied pressure dependence of the transmission spectrum of the MLPFG. The estimated spectra around the resonance wavelengths based on our proposed model agree with those measured experimentally. An estimation error occurs at wavelengths far from the resonance wavelengths. Further studies are required to determine the underlying cause for this estimation error. However, the resonance loss and resonance wavelength are the most critical for MLPFGs that are used as sensors [2, 3, 4, 5]. Therefore, we believe that our proposed model is useful for estimating the properties of fabricated MLPFGs.

Acknowledgments

The authors would like to thank Dr. I. Yamashita at the Kansai Electric Power Co., Inc. for his valuable advice during discussions. This work was supported by JSPS KAKENHI Grant Number 16K06307.
Numerical estimation for TEM horn antennas with transmission line taper shapes

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Abstract: The design of transverse electromagnetic (TEM) horn antennas has to consider impedance matching between the feeding point and the aperture. To achieve good impedance matching, several typical tapered transmission lines are applied to the taper shape of TEM horns in this study. The numerical estimation of the antenna characteristic of each TEM horn is performed using the finite integration method. An exponential taper is most applicable to the antenna shape among those transmission line tapers. An exponentially tapered TEM horn, however, has problems in maintaining a single main lobe in the radiation pattern. We propose a shortened exponentially tapered TEM horn in which has a simple structure that improves the radiation directivity, and the estimation results show its effectiveness as a broadband antenna.

Keywords: TEM horn antenna, transmission line taper, impedance matching, finite integration method, exponentially tapered TEM horn, shortened exponentially tapered TEM horn

Classification: Electromagnetic Compatibility (EMC)

References

1 Introduction

TEM horn antennas are widely used for UWB (ultra-wideband) systems such as radar and wireless communication and for electromagnetic compatibility (EMC) estimation due to their low pulse distortion. Recently, this antenna type has been explicitly specified in an international EMC standard [1] as a field generating antenna used for radiated immunity testing in close proximity. A basic TEM horn consists of linearly tapered plates. For impedance matching of the feeding point and the aperture, a linearly tapered TEM horn with continuous resistive loading has been proposed [2, 3]. A TEM horn with an exponentially tapered shape has also been designed, and which has the advantage of obtaining smooth impedance variations without resistive loading [4]. This result suggests the possibility of applying other well-known tapered transmission lines, such as the Klopfenstein taper, to the antenna shape. However, the effectiveness of these transmission lines as broadband antennas has not been sufficiently confirmed. Although the TEM horn with an exponentially tapered is useful for a broadband antenna, it does not maintain a single main lobe in the radiation pattern over some part of its frequency range. An exponentially tapered TEM horn with arc curvature was subsequently proposed to remove the fluctuations of the main lobe [5].

In this paper, we discuss the antenna characteristics of TEM horns with antenna shapes based on typical tapered transmission lines, including exponential, triangular [6], Klopfenstein [7], and Hecken [8] tapers, and verify whether they are effective as broadband antennas. In this numerical estimation using the finite integration method (FIM), the TEM horns were designed for EMC measurements from 400 MHz to 6 GHz. From the results of numerical analysis for the TEM horns, we propose a shortened exponentially tapered TEM horn, in which has a very simple structure, optimized to maintain a single main lobe.

2 TEM horn

A TEM horn antenna consists of two tapered metal plates fed by a coaxial line. To obtain good reflection characteristics, the antenna structure has to be designed considering impedance matching of the feeding point and the aperture of the antenna. For a TEM horn with a linearly tapered shape, a resistively loaded taper is used to reduce reflections [2, 3]. To achieve good impedance matching without resistive loading, typical tapered transmission lines are applied to the antenna shape, i.e., the separation of two plates. The taper shapes we compared are exponential [6], triangular [6], Klopfenstein [7], and Hecken [8] types. The variations of the characteristic impedance $Z(z) (0 \leq z \leq L)$ for each taper of length $L$, source impedance $Z_0$, and load impedance $Z_L$ are expressed as follows:
exponential taper [6]:
\[ Z(z) = Z_0 e^{z/L \ln(Z_L/Z_0)}, \]  
\[(1)\]

triangular taper [6]:
\[ Z(z) = \begin{cases} 
Z_0 e^{2(z/L)^2 \ln(Z_L/Z_0)} & (0 \leq z \leq L/2), \\
Z_0 e^{(4z^2/L^2 - 2z^2/L^2 - 1) \ln(Z_L/Z_0)} & (L/2 \leq z \leq L),
\end{cases} \]
\[(2)\]

Klopfenstein taper [7]:
\[
\ln Z(z) = \frac{1}{2} \ln(Z_0 Z_L) + \frac{\Gamma_0}{\cosh A} A^2 \phi\left(\frac{2z}{L} - 1, A\right),
\]
\[(3)\]
where
\[
\phi(x, A) = \int_0^x \frac{I_1(A \sqrt{1 - y^2})}{A \sqrt{1 - y^2}} dy,
\]
\[
\Gamma_0 = \frac{Z_L - Z_0}{Z_L + Z_0},
\]
where \(I_1\) is the first kind of modified Bessel function and \(A\) is a parameter for the tapered line curve, and

Hecken taper [8]:
\[
\ln Z(z) = \frac{1}{2} \ln(Z_0 Z_L) + \frac{1}{2} \ln\left(\frac{Z_L}{Z_0}\right) \frac{B}{\sinh B} \varphi\left(\frac{2z}{L} - 1\right),
\]
\[(4)\]
where
\[
\varphi(B, x) = \int_0^x I_0(B \sqrt{1 - y^2}) dy,
\]
\[
\text{where } I_0 \text{ is the modified zero-order Bessel function and } B \text{ is a parameter for the tapered line curve.}
\]

Fig. 1. Analytical model for TEM horn with transmission line taper shape.
In this numerical estimation, the TEM horn was designed for EMC measurements in the frequency range from 400 MHz to 6 GHz. The antenna length \( L \) and the aperture dimension are set to half the wavelength of the lowest frequency. The other antenna parameters are the input impedance at the coaxial feeding point of 50 \( \Omega \) (= \( Z(0) \)) and the characteristic impedance at the square aperture of 377 \( \Omega \) (= \( Z(L) \)). The variations of the characteristic impedance for each taper shape is indicated in Fig. 1(a). Assuming that the TEM horn consists of minute parallel plates, as shown in Fig. 1(b), the width \( w(z) \) at the location \( z \) of a taper plate can be approximated as follows from the plate separation \( h(z) \) and characteristic impedance \( Z(z) \) [4]:

\[
w(z) = \frac{h(z)}{Z(z)} 120\pi.
\]  

(5)

Each antenna model is composed of the taper structure determined using Eqs. (1)–(5), as shown in Fig. 1(c). The antenna characteristics for TEM horns with transmission line tapers were estimated numerically using a full-wave electromagnetic solver (CST MW-Studio [9]) based on FIM. The TEM horn has a balanced structure with parallel plates and is fed by an unbalanced coaxial cable. Therefore, a broadband balun is generally needed to feed the antenna. In this simulation, TEM horns are directly fed without a coaxial cable.

Fig. 2(a) and (b) show the calculation results of the reflection (\(|S_{11}|\)) and the gain characteristics of the TEM horns, respectively. Each antenna exhibits a satisfactory reflection characteristic over the broadband frequency range owing to the effects of impedance matching by the tapered transmission lines. By applying
transmission line tapers to the antenna shape, impedance matching without resistive loading is possible. However, as shown in Fig. 2(b), the antennas applying the triangular and Klopfenstein tapers have many large ripples in the gain profile. On the other hand, the antenna with the exponential taper has flat gain characteristics, allowing it to be used as a broadband antenna. When the constant $B$ is small for the Hecken taper (e.g., $B = 0.1$), the shape is in agreement with that of the exponential taper. When the Hecken taper has a slightly different shape from the exponential taper (e.g., when $B = 2$), the gain characteristic deteriorates at higher frequencies. The exponential taper is therefore the most suitable as the taper structure of a TEM horn among the investigated tapered transmission lines. However, the maximum gain of the antenna is not located in front of the antenna at over some part of its frequency range, as shown in Figs. 2(c) and (d).

3 Improved TEM horn

For antennas used for EMC measurements, the direction of maximum radiation is commonly designed to be in front of the antenna. We propose an improved TEM horn that resolves the directivity problem of the exponentially tapered TEM horn. As a similar approach to improve the directivity, a TEM horn with arc curvature shape has been proposed [5]. The antenna shape that we designed has a very simple structure compared with the arc curvature shape. The modified exponential taper is shortened by removing approximately 10% of the antenna length from the aperture, as shown in Fig. 3(a). The cut length ($dl$) was determined by optimization of the directivity by numerical simulation. That is, the difference between the on-axis gain and the maximum gain was calculated for each cut length, as shown in Fig. 3(b).

![Fig. 3. Shortened exponentially tapered TEM horn.](image)
When the cut length was 6%, i.e., $0.06 \times L(= dl)$, or more, the on-axis gain matched the maximum gain. The gain characteristic and radiation pattern of the exponentially tapered TEM horn with a cut length of 10% are respectively shown in Figs. 3(c) and (d). The direction of maximum gain is directly in front of the antenna for a given frequency range, and also the radiation pattern maintains a single main lobe. Although the gain is decreased at lower frequencies resulting from the reduced antenna length, a relatively flat characteristic is obtained. A cut length of 6% to 12% is suitable considering the effect of gain reduction at lower frequencies. This upper limit of the cut length (12%) was determined so that the gain reduction did not exceed 3 dB.

4 Conclusion

In this study, we evaluated TEM horn antennas with a taper shape applied several typical transmission line tapers, including exponential, triangular, Klopfenstein, and Hecken types. Numerical analyses of TEM horns designed for EMC measurements from 400 MHz to 6 GHz were performed using FIM. Each TEM horn exhibited a satisfactory reflection characteristic owing to the effect of impedance matching by the transmission line. However, only a TEM horn with an exponentially tapered shape had a gain characteristic that enables its use as an antenna. The single main lobe of the exponential-type horn, meanwhile, is not maintained at some part of frequency range. We proposed a shortened exponentially tapered TEM horn that resolves this directivity problem, and the results obtained show its effectiveness for use as, for example, an EMC broadband antenna. The remaining problem is to design an adequate broadband balun for an unbalanced TEM horn and then to confirm the numerical results experimentally.