Effectiveness of transmission-quality-aware online network design and provisioning enabled through optical performance monitoring

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Abstract: The use of the higher-order modulation format and narrower channel bandwidth can enhance the spectral efficiency of photonic networks. However, such spectrally efficient schemes cannot be utilized when the large margin is required to tolerate uncertainties in the performance of system components. Furthermore, the disaggregation of optical transmission system increases the required margin. In this paper, we propose highly spectrally efficient network architecture, where the required margin is minimized by transmission-quality-aware modulation-order/channel-bandwidth assignment based on optical performance monitoring. Up to 82% improvement in spectral efficiency for 400-Gbps DP-QPSK/DP-16QAM signals is confirmed through transmission experiments and network simulations.

Keywords: networks, wavelength routing, network survivability

Classification: Fiber-Optic Transmission for Communications

References


1 Introduction

To enhance the spectral efficiency of photonic networks, bandwidth-variable transponders (BVTs) are highly effective as they make it possible to use higher-order modulation formats and denser wavelength-division multiplexing (WDM) [1]. However, such spectrally efficient systems are vulnerable to transmission impairments such as fiber nonlinearity, amplifier noise, and signal-spectrum narrowing. To maximize the spectral efficiency under the various impairments, the possible highest modulation order and narrowest channel bandwidth should be assigned to each optical path depending on the path’s transmission conditions. Unfortunately, to allow for the uncertainties in the performance of components including optical fibers, erbium-doped fiber amplifiers (EDFAs), and wavelength-selective switches (WSSs), transmission performance analyses based on off-line design tools must be quite conservative, i.e. higher than actually needed margins [2]. As recently highlighted, an additional source of margin stems from the disaggregation of the optical transport system (OTS). In presently deployed all-in-one network systems, a universal margin is set by relying on the experiences of vendors and network providers. In contrast, disaggregation forces margins to be set on a component basis, which can substantially enlarge the total margin [3]. For reliable network operation, this implies that excessive margins need to be reserved to accommodate the impairment uncertainties and disaggregation effects, which may hinder the use of higher-order modulation formats and degrade the spectral efficiency.

In this paper, we demonstrate the effectiveness of transmission-quality-aware online network design and provisioning in terms of margin minimization; such design and provisioning are enabled by using the recently developed optical performance monitor (OPM) [4]. The network spectral efficiency is maximized by properly assigning the modulation order and channel bandwidth according to the actually measured performance, since the OPM can properly assess transmission performance in a path-by-path manner, a benefit not possible with off-line design tools. Based on the measured performance, the possible highest modulation order and narrowest channel bandwidth can be assigned to each path. This paper starts by experimentally evaluating the Q-factor values of 400-Gbps signals as determined by OPMs; we examine four-subcarrier 32-Gbaud dual-polarization (DP) QPSK signals within 150/162.5/175 GHz and two-subcarrier 32-Gbaud DP-16QAM signals within 75/87.5/100 GHz. Second, we analyze the fiber-utilization efficiency attained by the OPM-enabled network. The results show that the number of fibers needed to accommodate a certain amount of traffic is reduced by up to 45%. The corresponding improvement in spectral efficiency is +82%.
2 Transmission-quality-aware network design and provisioning based on optical performance monitoring

2.1 Margin minimization by optical performance monitoring with Q-factor estimation

The margin is classified into design margin, unallocated margin, and system margin as summarized in Table I [2]; the system margin is prepared for slowly varying impairments, fast-varying impairments, and operator’s buffer. The disaggregated OTS also needs a disaggregation margin [3]. The accurate transmission performance determined by OPM enables precise margin management in a path-by-path manner, and hence the required margin can be reduced. Our recently developed built-in-receiver OPM estimates the Q factor by counting the number of bits regenerated in the forward-error correction (FEC) process [4] and hence the performance of optical path candidates can be calculated in a fast, accurate, and cost-effective manner. As a result, we can reduce the design margin assigned to handle uncertain transmission impairments and the system margin for slowly varying impairments and disaggregation; the unallocated margin and system margins for rapidly varying impairment and operator’s buffer are still needed.

2.2 Margin reduction using transmission-quality-aware path establishment enabled by OPM

The unallocated margin originates from the discrete data rate, channel bandwidth, and transmission reach, which can be reduced by using VBTs. The distance-adaptive modulation networks that use VBTs [1] adopt the highest-order modulation format possible that suits the transmission distance of each path. To realize reliable distance-adaptive modulation networks, the optical channel guard-band bandwidth must be broad enough so as to avoid the spectrum narrowing caused by WSS traversal; otherwise, an extra margin must be set to account for the penalty imposed by spectrum narrowing. A desirable solution is to assign the modulation order and channel bandwidth according to both the transmission distance and hop count of each path. This scheme, however, necessitates extensive analyses of the

<table>
<thead>
<tr>
<th>Margin classification</th>
<th>Origins</th>
<th>Scenario A</th>
<th>Scenario B</th>
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<td>Reduced</td>
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<td>Slowly varying impairments</td>
<td>Aging and data loading (nonlinearity)</td>
<td>Unchanged</td>
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<td>Fast-varying impairments</td>
<td>Polarization impairments etc.</td>
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<td>Disaggregation of the OTS</td>
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<td>Operator’s buffer</td>
<td>Extra buffer determined by the operator</td>
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transmission performance yielded by all path candidates; a large number of network parameters need to be considered and this will yield unfeasible computational loads. To realize adaptive control of the modulation order and channel bandwidth in a simple and reliable manner, we utilize the Q factor determined by OPM. With the measured Q factors, we can assign the highest modulation order and narrowest channel bandwidth for each path while satisfying the Q-factor requirement. Note that OPM based on spectrum analysis is not effective when the signal spectrum is narrowed by WSS traversal. In contrast, Q factor output by our OPM takes all impairments into consideration; therefore, we can realize highly dense WDM networks in which the signal spectrum is severely narrowed. Our proposed scheme thus can enhance the spectral efficiency without precisely considering each impairment.

3 Transmission experiments

We measured the Q factor versus hop count while changing modulation order and channel bandwidth. We examined the 400-Gbps signal formats recommended by OIF [5], i.e., four-subcarrier 32-Gbaud DP-QPSK signals and two-subcarrier 32-Gbaud DP-16QAM signals. The subcarriers of all signals are aligned with 37.5-GHz spacing. The channel bandwidth is set to 150/162.5/175 GHz for the QPSK signal and 75/87.5/100 GHz for the 16QAM signal. The signals traversed a 40-km single-mode fiber (SMF) and following variable optical attenuator (VOA); the span loss of an 80-km SMF was emulated. The signal was delivered to the next node that used an express switch comprised of 1×8 splitters and 8×1 WSSs connected in a broadcast-and-select manner. The target-signal spectrum was narrowed with each node traversal. After traversing multiple nodes, the target signal was dropped and detected with a digital coherent receiver. Finally, the Q factor was measured by the OPM.

Fig. 1 plots the measured Q factor versus hop count; the parameters are the assigned modulation order and channel bandwidth. We find that channel bandwidths of 162.5 GHz for the QPSK signal and 87.5 GHz for the 16QAM signal are sufficient to prevent the spectrum narrowing effect. In contrast, we observe serious signal-quality degradation caused by the spectrum-narrowing effect when the assigned channel bandwidths are 150 GHz for the QPSK signal and 75 GHz for the 16QAM signal. In the next section, the spectral efficiency is maximized by assigning the modulation orders and channel bandwidths according to these experimental results.

Fig. 1. Q factor measured as a function of hop count.
4 Network simulations

To demonstrate the effectiveness of our proposed network, we evaluate fiber-utilization efficiency in three scenarios summarized in Table I. Scenario A: distance-adaptive network without OPM, where a modulation format is assigned to each path according to just the transmission distance obtained from simulation tools. In Scenario B: distance-adaptive network with OPM, where the optimum modulation format is selected in accordance with the Q factor obtained from OPM. Scenario C: our proposed transmission-quality-aware network with OPM, where the optimum modulation format and channel bandwidth are selected according to the Q factor obtained from OPM. All inter-node distances are assumed to be 80 km to match the experimental setup; note that non-uniform inter-node distances are also possible since our OPM can estimate the Q-factor value exactly from the received signal. The path demand is uniformly and randomly distributed and the average number of path demands between each node pair is set to 5. The available bandwidth of each fiber is 4,400 GHz (i.e. 352 12.5-GHz slots in the C band). We assume the use of 400-Gbps QPSK and 16QAM signals as in the experiments. We determined the assigned bandwidth by the experimental results shown in Fig. 1. In Scenario A and B, the assigned bandwidths are 162.5 GHz for DP-QPSK signals and 87.5 GHz for DP-16QAM signals. In Scenario C, the assigned bandwidths are selected from 150/162.5 GHz for DP-QPSK signals and 75/87.5 GHz for DP-16QAM signals. As mentioned above, the bandwidths per channel that can completely avoid the spectrum narrowing effect are 162.5 GHz in the QPSK system and 87.5 GHz in the 16QAM system; therefore, 175 GHz in the QPSK system and 100 GHz in the 16QAM system are excluded from the candidates in the scenarios examined.

Fig. 2 shows simulation results for each tested topology: the tested network topologies (Fig. 2(a)), the number of fibers necessary for accommodating all path demands (Fig. 2(b)), and the ratios of assigned modulation orders/channel bandwidths (Fig. 2(c)). The horizontal axis denotes the actually required total margin, which includes a 2-dB margin for the fast-varying impairments and operator’s buffer. The Q factor at the FEC threshold is set to 5.7 dB. As for Scenario A, the additional margin of 4 dB (i.e. constant 6 dB in total) is assumed considering design errors, aging, network loading, and disaggregation. In Scenario A, 16QAM signals cannot be assigned due to the large margin required. As a result, the number of fibers needed is large. Compared to Scenario A, Scenario B can reduce the necessary numbers of fibers by 0–27% for JPN48 and 0–43% for Telefonica depending on the actually required margin; Scenario C, our proposal, can reduce the number of necessary fibers by 2–32% for JPN48 and 6–45% for Telefonica. As for the ratios of assigned modulation orders/channel bandwidths, Scenario B and C can utilize 16QAM signals because the margin can be reduced thanks to the use of OPM; moreover, Scenario C can reduce the channel bandwidth more effectively thanks to adaptive bandwidth allocation. Our scheme thus can attain the highest fiber utilization.
5 Conclusion

We demonstrated the effectiveness of real-time optical performance monitoring in realizing transmission-quality-aware online network design and provisioning. The precise performance estimation enables adaptive control of modulation order and channel bandwidth for each path. The scheme can substantially reduce margin needed for uncertain transmission impairments and system disaggregation and as a result the network’s spectral efficiency can be maximized. The effectiveness was confirmed by transmission experiments and network analyses. The number of fibers was reduced by up to 45%; the improvement in spectral efficiency corresponds to 82% (i.e. $100/(100 - 45) \sim 1.82$). The unallocated margin can be further reduced by increasing the number of modulation formats and channel bandwidths that can be selected.

Acknowledgments

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Time-domain distortion of a pulse-operated high-power GaN amplifier and a reduction method

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Abstract: This paper describes pulse waveform distortion by a high-power GaN amplifier and a method for reducing it in a solid-state radar that performs pulse compression. When a pulse signal is input to the amplifier, the output pulse waveform differs from the input form because of the influences of nonlinearity and transient response, which are remarkable at the time of rise and fall when the amplitude has a temporal slope. In the measurement results, the temporal slope and the pulse width increased by 79.6% and 1.5%, respectively, compared with the input waveform. On processing the received signal, these influences degraded the signal-to-noise ratio after pulse compression by about 5 dB. In the proposed method, these influences were reduced by adding a dummy signal before and after the desired signal. This method is advantageous in that it does not require the addition of a feedback circuit for distortion compensation and it is not easily influenced by changes in the signal settings and the ambient temperature.

Keywords: radar, GaN amplifier, pulse compression, nonlinearity, transient response

Classification: Sensing

References


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1 Introduction

Modern radar generally uses a semiconductor amplifier to generate the transmitted signal. Furthermore, to achieve detection performance comparable with that of a conventional electron tube, radar often uses a pulsed chirp signal as the transmitted signal and pulse compression to process the received signal. Gallium nitride (GaN) is used as the amplifying semiconductor because it has excellent thermal conductivity, band gap, and breakdown voltage and can handle higher output power compared with gallium arsenic (GaAs) in the same size.

In a GaN amplifier, the gain and efficiency (among other properties) change according to the output power [1], and gain compression in excess of 3 dB is not uncommon in high-power operation [2]. Also, any change in the power efficiency changes the amount of heat generated around the device. The change in heat generation is determined by the structure of the semiconductor and does not coincide with the envelope of the radiofrequency (RF) signal. This induces hysteresis in the GaN amplifier, known as the memory effect, and affects the linearity of the output signal [3]. Under pulse operation in which the output level changes with time, careful attention is required [4]. Techniques such as pre-distortion and feed forward can be used to reduce such distortion. However, assuming that the distortion changes with the signal settings and the ambient temperature, both those techniques require a feedback circuit.

In this study focused on solid-state radar, we measured the pulse waveform distortion in a high-power GaN amplifier and investigated its influence on pulse compression. Additionally, we propose a simple method for reducing the distortion.

2 Pulse waveform distortion of a high-power GaN amplifier and its influence on pulse compression

For the investigation, we used an X-band high-power GaN amplifier for marine radar. It was designed using a GaAs amplifier and two GaN amplifiers in series and operated with gate-bias switching for low power consumption. This amplifier had a maximum output power of 51 dBm (126 W), and gain of 61 dB at 9.41 GHz under a 4 µs pulse width and 2600 Hz repetition frequency. At this time, the gain difference from the linear region was 5 dB.

Fig. 1(a) shows two output waveforms, one at maximum output (51 dBm) in red and the other at an output of 50 dBm in green; the waveform of the amplifier
input is also shown in blue. To see any differences more clearly, normalized amplitudes are shown. In the measurements, the gate bias was also pulsed, but a sufficient margin (0.5 µs) was maintained to obtain pure RF characteristics.

This figure shows that the temporal slope of the amplitude steepened at the rise and fall compared with the input. In addition, the behavior at the edge was such that the pulse became wider. Fig. 1(b) details these comparisons numerically, wherein a number in parentheses indicates the percentage change compared with the amplifier input. At the maximum output, the temporal slope and the pulse width increased by 79.6% and 1.5%, respectively.

Fig. 1(c) shows the results of pulse compression. The pulse waveform distortion described above appeared as a degradation of the signal-to-noise ratio (SNR) upon pulse compression and deteriorated with increase of the operating range in the nonlinear region. A mismatched filter was used for the pulse compression [5].

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<tbody>
<tr>
<td>Amp in</td>
<td>101.1 (+33.5%)</td>
<td>90 (+50.0%)</td>
<td>4.467</td>
</tr>
<tr>
<td>Out (50 dBm)</td>
<td>135.0 (+33.5%)</td>
<td>135.0 (+50.0%)</td>
<td>4.467 (0%)</td>
</tr>
<tr>
<td>Out (51 dBm)</td>
<td>161.7 (+59.9%)</td>
<td>161.7 (+79.6%)</td>
<td>4.533 (+1.5%)</td>
</tr>
</tbody>
</table>

(b) Temporal characteristics

Fig. 1. Pulse waveform distortion and its influence on pulse compression.

3 Proposed method for reducing distortion

In Fig. 2(a), to understand the influence of nonlinearity, the values in Fig. 2(c) were applied to the amplifier input of Fig. 1(a). When considering only the influence of nonlinearity, the temporal slope steepened with the output power, the pulse tended to widen, but there was no delay relative to the input signal. On the other hand, when considering only the delay, phenomena such as pulse widening would naturally not be seen. This behavior was considered to be the influence of both
nonlinearity and transient response as shown in Fig. 2(b). A symmetric transient response appeared also at the fall time in Fig. 1(a), so it was presumed that a change in the output level was involved rather than a simple delay.

Fig. 2. Consideration of measurement results and summary of proposed method.
In the proposed method, a dummy signal that is separable by processing the received signal is added before and after the desired signal. By changing the amplitude of the dummy signal according to the desired signal, the sum amplitude is held constant while transmitting the desired signal. After passing through the amplifier, the distortion appears in only the dummy signal. The desired output waveform is obtained by separating the dummy signal at the receiver. In this method, no feedback circuits are required for distortion compensation. Furthermore, because the desired signal is handled in an electrically and thermally stable region, it is hardly affected by changes in the signal settings and the ambient temperature.

4 Effect of proposed method

In this study, a non-modulated pulsed signal whose frequency differed from that of the desired signal was used as the dummy signal, and the dummy signal was removed by the band-pass filter at the receiver. Fig. 3(a) shows the comparison before and after applying the proposed method at the maximum output. In particular, the transient response was greatly improved and there was a higher degree of similarity with the input signal. However, further investigation is required to explain the slight remaining differences. Fig. 3(b) shows the results of pulse compression. The difference in characteristics was almost eliminated, and an SNR improvement of about 5 dB was obtained compared with the results in Fig. 1(c).
5 Conclusion

This paper used measurement results to describe the pulse waveform distortion at the rise and fall of the signal in a high-power GaN amplifier, showing that the distortion appeared as a deterioration of the SNR due to pulse compression. In addition, a method was proposed for reducing this distortion, and its efficacy was confirmed. Although this method involves a wider pulse, operation in the electrically and thermally stable region can be secured with no feedback circuit. In future work, we intend to study in more detail the signal specifications necessary for the dummy signal, such as the insertion timing.

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Digital post-processing against inline spectrum narrowing caused by optical node traversals

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Abstract: This paper evaluates performance of digital coherent receivers that executes partial-response spectrum shaping and MLSE to mitigate the inline-spectrum-narrowing impairment caused by traversing multiple optical nodes in photonic networks. Extensive computer simulations confirm that the maximum transmission distances and hop counts of DP 4-QAM and 16-QAM signals in highly dense WDM networks are substantially extended.

Keywords: digital coherent receiver, partial-response coding, spectrum narrowing

Classification: Fiber-Optic Transmission for Communications

References


1 Introduction

To accommodate the growing network traffic, spectral efficiency of photonic networks must be improved. The spectral efficiency can be improved by the use of highly dense wavelength-division multiplexing (WDM) and the higher-order modulation formats with coherent detection. In such spectrally efficient systems, the signal suffers from severe signal-spectrum narrowing caused by traversing the wavelength-selective switches (WSSs) within optical nodes [1]. The inline spectrum narrowing might occur at every optical-node traversal. The spectrum narrowing results in intersymbol interference (ISI) as the global transfer function of concatenated WSSs does not satisfy the Nyquist ISI criterion. Furthermore, the narrowed signal spectrum is contaminated by amplified-spontaneous-emission (ASE) noise generated from the optical amplifiers. As the signal repeatedly undergoes such filtering-and-amplification processes, the ISI and ASE noise impair the signal in an interactive manner. The ISI itself can completely be equalized by digital filters in digital coherent receivers [2]. However, such linear equalizers enhance the in-band noise power because the low-power frequency components of the filtered signal are emphasized for ISI compensation [1]. Maximum likelihood sequence estimation (MLSE) can perform symbol decision under the influence of ISI without the noise enhancement [3]. However, the computational complexity of MLSE is high owing to the uncertain impulse response of the channel. Another candidate is partial-response coding, e.g. duo-binary, in conjunction with MLSE [4, 5]; it realizes a narrow signal bandwidth by deliberately imposing deterministic ISI at the transmitter side. The partial-response-coded signal possesses high spectrum-narrowing tolerance; however, its short intersymbol distances on the constellation map result in low ASE-noise tolerance. In addition, the noise enhancement accompanying ISI compensation is inevitable [6, 7].

Given this background, we showed effectiveness of receiver-side partial-response filters in conjunction with MLSE decoding in highly dense WDM networks using wavelength routing [6, 7]. The 4/16/64-QAM (M-QAM) signal spectrum narrowed by traversing optical nodes is shaped with a digital filter in the digital
coherent receiver as if the signal is coded by a two-tap partial-response filter; in other words, the $M$-QAM signal is converted into the two-tap partial-response-filtered (PRF) $M$-QAM signal at the receiver side. By allowing the deterministic two-symbol-length ISI, the noise enhancement inherent in the conventional systems can be suppressed. The symbols are then decoded by the simple two-symbol MLSE. Although this scheme can extend the maximum transmission distances and hop counts, the noise enhancement still occurs if the received signal spectrum is extremely narrow due to a large number of node hops.

This paper extended the allowable ISI length to 3 and 4, where spectrum shaping is performed as if the received signals are coded by the $n$-tap ($n = 2, 3,$ or 4) partial-response filter best suited to the received signals. As a result, the impact of noise is reduced. Then, simple $n$-symbol MLSE conducts symbol decision under the residual deterministic $n$-symbol-length ISI. The maximum transmission distances and hop counts of dual-polarization (DP) $M$-QAM signals in highly dense WDM networks are more extended when the impacts of spectrum narrowing is severe.

### 2 Digital signal processing for inline spectrum narrowing

Fig. 1 shows the post-processing scheme in the digital coherent receiver; the conventional scheme is also illustrated. The $M$-QAM signal carried by the network experiences multiple filtering-and-amplification processes along the channel. In the conventional scheme, the received signal is then equalized by a digital filter in the digital coherent receiver. In the equalization process, noise enhancement is unavoidable and the signal-to-noise ratio (SNR) is degraded. In contrast, the spectrum of the received $M$-QAM signal is shaped to yield an $n$-tap PRF $M$-QAM signal spectrum using a digital filter in the digital coherent receiver. This filtering process allows deterministic $n$-symbol-length ISI and so can suppress the noise enhancement inevitable in the conventional scheme. Then, the simple $n$-symbol MLSE conducts symbol decision under the residual deterministic ISI. We can thus alleviate the impairment stemming from the spectrum narrowing. Fig. 2 shows the trade-off between spectrum-narrowing tolerance and noise tolerance regarding the number of taps of partial-response filter $n$. As $n$ increases, noise is suppressed more since the filter shape becomes narrower. However, the shorter intersymbol distance due to deterministic ISI results in lower noise tolerance as shown in the lower row. Hence the optimum $n$ should be selected so as to maximize the transmission performance.

It has been proven that the receiver-side spectrum shaping followed by MLSE can improve the transmission performance of point-to-point non-coherent systems in which the signal spectrum is severely narrowed due to the transmitter/receiver electrical band limitation [8, 9]. In contrast, the subjects of this paper are inline spectrum narrowing caused by WSSs and the mitigation scheme executed in the digital coherent receiver. Hence, the filtering source, noise source, and detection scheme differ from those discussed in the literatures [8, 9].
Simulations

We executed computer simulations assuming nonlinear transmission systems. A transmitter generates a power-optimized 32-Gbaud DP 4/16/64-QAM (M-QAM) signal. The signal is then combined with closest eight wavelength channels through a WSS. The grid interval is 37.5 GHz considering highly dense WDM networks. The signals are then launched into a fiber link. Each link consists of a 50-km or 100-km single-mode fiber (SMF) and an erbium-doped fiber amplifier (EDFA). The loss coefficient, dispersion parameter, and nonlinear coefficient of the SMF are 0.2 dB/km, 17 ps/nm/km, and 1.5/W/km, respectively. The noise figure of the EDFA is 5 dB. The power of each channel is then amplified to −1 dBm and the signal enters the next node. The route-and-select (R&S) node architecture (WSS-WSS) is assumed [10]; the node loss is 15 dB. The passband of each WSS is calculated by convoluting a rectangular function with 37.5-GHz bandwidth and a Gaussian function with 10-GHz 3-dB bandwidth; here, the 3-dB bandwidth of the

3 Simulations

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Gaussian function determines the cut-off characteristics. After multiple node hops, the target signal is dropped and detected by a digital coherent receiver. After compensation for chromatic dispersion, the DP $M$-QAM signal is polarization-demultiplexed by a two-by-two butterfly-structured finite-impulse-response (FIR) filter with 16 tap coefficients adapted by the least-mean-square algorithm [2]; at the same time, the signal spectrum is converted into a proper $n$-tap PRF $M$-QAM signal spectrum. After that, the $n$-tap PRF $M$-QAM signal is decoded by the simple $n$-symbol MLSE [3]. Finally, the bit-error ratio (BER) is calculated. The target BER is $10^{-2}$ considering the use of forward error correction (FEC). For reference, conventional demodulation schemes were also evaluated.

Fig. 3 depicts BER versus transmission distance and hop count, where the modulation order $M$ and inter-node distance are changed. We find that improvement in the 50-km-span systems is more prominent than that in the 100-km-span systems. This is because the signal spectrum is more narrowed in the 50-km-span systems with the given transmission length. Compared to the conventional scheme, the proposed scheme can attain much longer transmission distance and larger hop count when $M = 4$ and $M = 16$. As for the cases where $M = 64$, less improvement is observed because such systems are more susceptible to ASE noise and fiber nonlinearity rather than to spectrum narrowing. Such features are in good agreement with the qualitative prediction shown in Fig. 2. When the spectrum narrowing is the dominant cause of performance degradation, the BER is substantially improved; for example, the maximum hop count of the 4-QAM signal can be extended by 16 for the 50-km-span system and by 8 for the 100-km-span system; the corresponding distance increases are 800 km and 800 km. Thus, the use of appropriate partial-response filter can substantially extend the transmission distances and hop counts.
4 Conclusions

We confirmed the effectiveness of a combination of partial-response spectrum shaping and MLSE to alleviate the impact of the spectrum narrowing caused by traversing multiple optical nodes. Extensive computer simulations confirmed the effectiveness in highly spectrally efficient WDM networks based on wavelength routing. For example, we can extend the maximum hop count of a 32-Gbaud DP-QPSK signal by 8, when the inter-node distance is 100 km; the distance increase is 800 km. The scheme can utilize conventional coherent transmitters; only the receiver’s digital signal processing needs to be modified.

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Robust design of radar absorbent material with broadband characteristics

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Abstract: Recently, radar absorbent materials (RAM) with various characteristics have been developed according to applications. Since it is difficult to satisfy the required performance when errors occur in them during fabrication, it is necessary to have robustness of the electrical material constant. This paper shows robust design of RAM with broadband characteristics. The change magnification of relative permittivity is set to 0.3 of the limit value and that of thickness is set to 0.05. As a result, even when the design variables take the worst values with respect to the objective function, it is possible to realize broadband characteristics of relative bandwidth of 164.3% with respect to wave-absorption amount ($W_a$) of 10 dB.

Keywords: radar absorbent material, robustness, broadband characteristics, optimization

References


1 Introduction

Recently, radar absorbent materials (RAM) with various characteristics, such as broadband, oblique incidence, and polarization characteristics, have been developed according to applications. RAM with multilayer structure is suitable for broadband characteristics [1]. Since design calculation is complicated by increasing RAM layer, optimization is necessary to determine the electrical material constant with respect to the received frequency and incidence angle, which determine wave-absorption amount ($W_a$). Moreover, those optimal solution are obtained as fixed values in optimized design [2]. However, it is difficult to satisfy the required performance when errors occur in them during fabrication. Therefore, it is necessary to have robustness of the electrical material constant and to satisfy broadband characteristics. This paper indicates robust design of dielectric RAM with broadband characteristics at vertical incidence.

2 RAM configuration

Fig. 1 shows the configuration of two flat layers RAM composed of dielectric materials. We use two flat layers RAM with a flat plate structure in order to overcome structural restrictions and widen the range of applications. Flat layer RAM has a structure in which the RAM is installed on a metal plate, and it becomes possible to reduce its energy by controlling the amplitude and phase of the reflected wave. As shown in Fig. 1, the electrical material constant are the values of the complex relative permittivity and thickness. Considering the frequency dependence of complex relative permittivity, we use those values at 1 GHz and the decay coefficients [3]. Therefore, the number of design variables $x$ is 10 in total. For the range of $x$, the real part of permittivity of each layer at 1 GHz is 1 to 20, the imaginary part of permittivity is 0 to 20, the values of decay coefficients are 0 to 1, and the thickness of each layer is 1 mm to 20 mm.

![Fig. 1. Configuration of two flat layers RAM composed of dielectric materials.](image-url)
3 Robust design method

Robust design is a design in which the minimum value of the objective function is the maximum within the variation section of the design variables, and it is possible to be formulated as a problem to maximize the worst value of the objective function. Assuming that the normalized input impedance of the RAM is $Z_{in}$, $Wa$ is shown as Eq. (1):

$$F_1(x) = -20 \log_{10}(\text{abs}((Z_{in} - 1)/(Z_{in} + 1))) = Wa$$

where the unit of $Wa$ is dB. Here, when the change magnification of $x$ is $3\sigma$, the objective function $F(x)$ is expressed by the Eq. (2) using Eq. (1), the design variables $x$, and the change magnification $3\sigma$. It is possible to be formulated as a problem to maximize the minimum value of $Wa$. As shown in Eq. (3), the design variables $x$ are possible to move within the range of $3\sigma$ from the average values $x_0$, because the variable $u$ exists for $3\sigma$ and $u$ moves within the range from $-1$ to $+1$. In the proposed robust optimization method, $u$ is optimized to give the worst value for the objective function firstly, and then the average values $x_0$ are optimized so as to obtain the best result even if they are multiplied by the worst value of $u$.

Maximize: $F(x) = \min \{ F_1(x) \mid x \in [x \times (1 - 3\sigma) \leq x \leq x \times (1 + 3\sigma)] \}$

$$x = x_0 \times (1 + 3\sigma \times u), \quad -1 \leq u \leq 1$$

We set the change magnification $3\sigma$ of $x$ to 0.3 for the real and imaginary parts of relative permittivity and 0.05 for the thickness. The larger the change magnification of the design variables, the wider the allowable range when the variation occurs. Therefore, the value of $3\sigma$ is gradually increased, and its limit value is confirmed. Because the relative permittivity not only relates to fabrication error but also whether or not there is actually a substance having a relative permittivity close to it, the larger the value of $3\sigma$ to be set, the more advantageous the design. On the other hand, with respect to the thickness, since only the manufacturing error is related to $3\sigma$, it is set within 5% considered as the error range. The optimization calculation is performed from frequency 3 GHz to 20 GHz in steps of 0.25 GHz, and the incident angle is examined only when it is 0 deg. We focus on the minimum value of $Wa$ obtained at the specified frequency and maximize it. Here, the setting of the initial point is important in order not to fall into a local solution. Therefore, we decide 30 candidate initial points so that they are evenly scattered within the design variable space, and adopt the initial point from which the best solution was obtained among them by using Hammersley sampling methods [4].

4 RAM with broadband characteristics

Fig. 2 indicates frequency characteristics of $Wa$ of two flat layers RAMs by optimized design and robust design. Fig. 2(a) shows optimized design results when $3\sigma$ is 0. The solid line ($x$._best1) shows the result of the optimized design variables which give the best result to the objective function. The broken line ($x$._worst1) shows the result when the objective function takes the worst value when giving variation of $3\sigma$ to the optimized design variables. Fig. 2(b) indicates robust design results when the design variables are optimized by giving variation of $3\sigma$. The solid line ($x$._best2) shows the result of the center design variables at which the best result
is obtained for the objective function. The broken line \((x_{\text{worst2}})\) shows the result when the objective function takes the worst value when the design variables deviate from the center values most. Table I shows values of design variables of \(x_{\text{best1}}, x_{\text{worst1}}, x_{\text{best2}}\) and \(x_{\text{worst2}}\). While the proposed robust design method maximizes the worst value of \(W_a\) from the calculated frequency 3 GHz to 20 GHz, the worst value of the broken line in Fig. 2(b) is 15.2 dB at 20 GHz and that of the broken line in Fig. 2(a) is 13.1 dB at 3 GHz. Thus, the former robust design is bigger. Therefore, it is possible to be said that robust design is achieved. Moreover, \(W_a\) is very high when there are no variations in the design variables, and the RAM has broadband characteristics for \(W_a\) of 10 dB or more in Fig. 2(a). On the other

![Optimized design results](image1)

![Robust design results](image2)

Fig. 2. Frequency characteristics of \(W_a\) of two flat layers RAMs.
hand, when variations occur in design variables, the characteristics cannot be maintained and the required performance is not sufficiently satisfied.

Next, when comparing whether or not there are variations in design variables in Fig. 2(b), the shapes of both graphs do not change significantly and both cases satisfy the broadband characteristics. Therefore, it is obvious that the required performance can be maintained regardless of the presence or absence of variations in design variables. In addition, when the index of $W_a$ is set to 10 dB, the minimum frequency satisfying 10 dB is 2.39 GHz in the optimized design, while it is 1.96 GHz in the robust design. As a result, RAM by the robust design satisfies more broadband characteristics, that shows relative bandwidth of 164.3%. For the above reasons, the proposed robust design method is effective.

### 5 Conclusion

In this paper, robust design of RAM with broadband characteristics was investigated. As the change magnification of the design variables is larger, the allowable range widens when the variation occurs. As a result of checking the limit value, the change magnification of the relative permittivity was 0.3. In addition, the change magnification of the thickness was set to 0.05 considered as the range of manufacturing error. We designed two flat layers RAM with robustness to design variables using these change magnifications. As a result, even in the case of the design variables take the worst values with respect to the objective function, it is possible to realize broadband characteristics of relative bandwidth of 164.3% with respect to $W_a$ of 10 dB. From the above, this method is effective in RAM design, and is possible to maintain broadband characteristics even when there are variations in design variables.

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<th>Table I. Values of design variables</th>
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Interferometric angle-of-arrival estimation using a simple weight network

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Abstract: This paper presents a method to estimate the Angle of Arrival (AOA) for incident waves, using an RF-based interferometric monopulse technique. We used an AOA network to yield a weight function for the received signals, in which the phase of the received signals for each element is delayed in accordance with the coordinate of the elements using the differing lengths of microstrip transmission lines. Based on electromagnetic field analysis and experimental results, it is clarified that the proposed method achieves a high degree of accuracy for AOA estimation.

Keywords: interferometric angle of arrival estimation, circular array, weight network, connected car

Classification: Antennas and Propagation

References


1 Introduction

In the automotive industry, there is a significant interest in realizing new technologies such as self-driving cars, Internet of Things (IoT) for automobiles, and commercial vehicle management [1]. The direction of the incoming wave, i.e., the communications counterpart, such as another connected car or traffic signal, changes significantly according to the dynamic movement of the automobile. Therefore, the connected car must estimate the direction of incident wave autonomously. We are currently developing a multiple-input multiple-output (MIMO) antenna system that utilizes the circular phased array technology for use in connected cars [2]. This MIMO antenna system is realized using the capability of the angle of arrival (AOA) estimation function [3].

As for an AOA network, a Butler matrix is commonly used for giving an appropriate phase delay to the received signals [4]. However, a Butler matrix has the drawback of being a bulky microwave component, which leads to hardware complexity. The AOA antenna proposed in this paper employs an RF-based interferometric monopulse technique without a Butler matrix, which reduces hardware complexity [5].

This paper presents an experimental verification of AOA estimation that employs an RF-based interferometric monopulse technique. We used an AOA network for yielding a weight function to the received signals, in which their phase for each element is delayed in accordance with the coordinates of the elements, using the differing lengths of microstrip transmission lines. The experimental results are in good agreement with the results obtained from the geometrical relationship between the direction of incoming waves and the location of antenna elements.

2 Principle of the AOA estimation

As shown in Fig. 1(a), the AOA antenna is composed of nine elements, eight of which (#1–#8) are arranged at equal angular intervals of 45° on a circle of radius $a$, with the final element (#9) located at the center of the circle. When a radio wave originates from the $\phi$ direction, measured from the x-axis, the voltage induced on the $i$th element located on the circle with radius $a$ is calculated using the following equation:

$$V_i = e^{j2\pi a \cos \frac{2\pi i}{8} \phi}$$  \hspace{1cm} (1)

where all the elements are assumed to be isotropic antennas, and $\lambda$ denotes the wavelength of the received signal.

The voltage $V_i$ is multiplied by the weight function $W_i$ defined by the following equation:

$$W_i = e^{-j \frac{2\pi i}{8} \pi}$$  \hspace{1cm} (2)
The summation of the total voltage induced on the eight elements can therefore be written as:

\[ E_\Delta = \sum_{i=1}^{8} e^{j\left(\frac{\pi}{\lambda} \cos\left(\frac{2\pi}{\lambda} \sigma - \phi\right) - \frac{2\pi}{\lambda} \sigma\right)} \]  

(3)

Now, we consider the case where the number of elements is infinite and \( E_\Delta \) converges to the Bessel function expressed by the following equation:

\[ E_\Delta = e^{-j(\phi - \gamma)} \cdot 2\pi J_1 \left(\frac{2\pi}{\lambda} a\right) \]  

(4)

The received signal at the ninth element is represented by \( E_{10} \). Furthermore, phases of \( E_\Delta \) and \( E_{10} \) are denoted by \( \angle E_\Delta \) and \( \angle E_{10} \), respectively. The phase difference is calculated as follows:

\[ \phi_m = \angle E_\Delta - \angle E_{10} \]  

(5)

The phase difference expressed by Eq. (5) is approximately proportional to \( \phi \) (the actual direction of an incident wave).

Fig. 1(b) shows the summation of the whole voltage denoted by Eq. (3) as a function of the array radius, together with the Bessel function. The frequency used in the analysis is 2 GHz. The magnitude of \( E_\Delta \) reaches the maximum value when the radius is 4.4 cm. This configuration of narrow element spacing may lead to a significant gain reduction in the antenna, due to a strong electromagnetic mutual coupling. However, the effect of mutual couplings is not adequately considered in the theoretical formula.

To investigate the effects of mutual couplings on the characteristics of the AOA antenna, half-wavelength dipole antennas are used for the nine array elements. All dipoles are terminated at 50 \( \Omega \). The complex field pattern is calculated using the method of moments. The analytical results reveal that even in the presence of mutual couplings, the radiation gain of the antenna is nearly the same as that of a half-wavelength dipole antenna [6]. We have verified that this phenomenon is attributed to the decoupling mechanism caused by the cancellation of signals, with the weighted phase excitation induced in 50 \( \Omega \) terminals.

Fig. 1(c) shows the error of the estimated angle as a function of the array radius. The error is reduced by decreasing the array radius, with an error of less than 0.1° when the array radius is 4.9 cm. The maximum error decreases with a decrease in the distance between the adjacent elements. Hence, we determined the array radius to be 4.9 cm, based on the performance of the MIMO array [2]. Therefore, the distance between the adjacent elements is 3.75 cm, which is equivalent to a quarter wavelength at 2 GHz. In this case, the directivity of the subarray is expected to have a cardioid radiation pattern.
In this paper, we introduce the weight functions defined by Eq. (2), allowing the use of a simple AOA network. Fig. 2 shows the network used for estimating the AOA that processes the received signals [7]. Fig. 2(a) is a diagram, and Fig. 2(b) depicts a fabricated microwave circuit using a double-sided printed circuit board (an FR4-based PCB) with a thickness of 1 mm and a relative permittivity of 4.2.

In Fig. 2(b), the phase of the received signals for each element is delayed in accordance with Eq. (2), using the differing lengths of microstrip transmission lines. The use of a $180^\circ$ rat-race hybrid in Fig. 2(b) allows a $180^\circ$ phase reversal to be involved in the two signals combined by the three Wilkinson power combiners located on the right and left side of Fig. 2(b), enabling a small-sized AOA network to be realized.

Fig. 2(c) shows the phase characteristics of the fabricated AOA network. The phase difference between Port $#i$ ($i = 1, 2, \ldots, 8$) and Port $E_A$ was measured at 2 GHz using a vector network analyzer. In Fig. 2(c), Port #1 is used as a reference of the phase. In Fig. 2(c), the measured phase decreases by $45^\circ$ intervals with an increase in the element number. Furthermore, the difference between the measured and designed phases is $5.3^\circ$ on average. These facts indicate that the fabricated AOA network works properly to achieve the operation of the weight function given by Eq. (2).
4 Experimental results

In our previous study [7], the impact of the array radius on the operating characteristic has not been investigated because the radius was fixed at 4.9 cm. In this paper, an experimental verification of the proposed AOA estimation method with regard to the different array radius was conducted in an anechoic chamber. Fig. 3(a) shows a photograph of the AOA antenna, with the network depicted in Fig. 2(b). Hereafter, we consider an AOA configuration where eight half-wavelength dipole antennas (#1–#8), designed at 2 GHz, are mounted on a circle with a radius $a$ and an additional dipole antenna (#9) is located at the center of the circle, as shown in Fig. 3(a).

Fig. 3(b) shows the radiation patterns of $E_\Delta$ and $E_\Omega$ defined in the $xy$-plane. The solid and dashed lines depict the measured and calculated results, respectively. The blue lines depict $E_\Delta$, whereas the red lines represent $E_\Omega$ with a radius of 4.9 cm, respectively. The black lines depict $E_\Delta$ with a radius of 8.5 cm. As shown in Fig. 3(b), the radiation gain of $E_\Delta$ at $a = 4.9$ cm is approximately 0 dBd. A larger radiation gain is achieved with a radius of 4.9 cm than with 8.5 cm. This behavior is in good agreement with the results obtained from Fig. 1(b). The amplitudes of $E_\Delta$ and $E_\Omega$ have omnidirectional patterns, with a radius of 4.9 cm. Hence, the fabricated AOA antenna can be received signal that the communication target radiates.

Fig. 3(c) shows the phase radiation characteristics. The solid and dashed lines depict the measured and calculated results, respectively. The phase of $E_\Delta$ is approximately proportional to the actual direction of an incident wave $\phi$. On the
other hand, the phase of $E_{12}$ remains constant regardless of the azimuth angle. These facts indicate that the fabricated antenna works well as AOA function.

Fig. 3(d) shows the estimated angle as a function of the AOA for an incident wave. Fig. 3(d) shows that measured results are proportional to the angle of an incident wave. The error in the red and black curves was found to be 6.1° and 11.2° on average, respectively. In Fig. 1(c), the estimated angle is characterized by a higher accuracy when the radius is 4.9 cm than when it is 8.5 cm. The same results as in the analysis were obtained by actual measurements. However, the estimation error is larger in the actual measurements. A possible cause of this error is the phase error arising from the fabricated AOA network, as shown in Fig. 2(c).

5 Conclusion

This paper presents an experimental verification of AOA estimation that employs an RF-based interferometric monopulse technique. The experimental results are in good agreement with the results obtained from the geometrical relationship between the direction of incoming waves and the location of antenna elements.

Acknowledgments

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A list Viterbi equalizer with simple linear channel interpolation for acoustic communications

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Abstract: This paper discusses a list Viterbi equalizer (LVE) with linear channel interpolation (LCI), which can cope with time and frequency selective channels, i.e., doubly-selective channels of acoustic communications. Features of the proposed LVE are (1) to cope with delay spread over 127 bits with small computational complexity, (2) to cope with time selective channels by simple channel estimation and its LCI and (3) to cope with non-minimum phase condition by metric-criteria combining. Next, computer simulation results show that the proposed LVE has better performance on doubly-selective channels than a decision-feedback equalizer employing recursive least square (RLS) algorithm. Finally, field evaluation results confirm that the proposed LVE is suitable for acoustic communications environment.

Keywords: acoustic communications, doubly-selective channels, list Viterbi algorithm, linear channel interpolation, offset-quadrature phase-shift keying

Classification: Wireless Communication Technologies

References


1 Introduction

Recently, acoustic communications have been studied in land and underwater environment [1, 2]. A propagation environment of acoustic communications suffers from large delay spread and fast time-varying channels, i.e., severe doubly-selective channels.

In mobile communications, channel state information (CSI) varies frequently between minimum phase condition and non-minimum one. In order to maintain good communication quality, equalizers have to track and compensate these channel conditions. In these environments, a decision-feedback equalizer with a feedforward filter (FFF-DFE) employing recursive least square (RLS) algorithm has good performance [3]. However, computational complexity of RLS algorithm increases in proportion to the number of required taps. From an implementation point of view, it is important for equalizers to reduce computational complexity even on large time dispersive channels.

This paper proposes a list Viterbi equalizer (LVE), which employs simple channel estimation and its linear channel interpolation. The list Viterbi algorithm (LVA) is one of state-reduction algorithms including decision-feedback sequence estimation (DFSE) and M-algorithm, and it can reduce computational complexity, keeping good performance. In addition, the LVE with the metric-criteria combining (MCC) scheme [4, 5] has excellent performance in non-minimum phase condition environment. This paper evaluates performance of the proposed LVE on doubly-selective channels. Finally, field experimental results confirm that the proposed LVE is suitable for acoustic communications.

2 Communication system model

Fig. 1 shows a communication system model in the presence of inter-symbol interference (ISI), where $k$ denotes discrete time, $b_k (b_k \in +1, -1)$ is a transmitted information sequence, $r_k$ is a received sequence and $\hat{b}_k$ is an estimated information information sequence.
sequence. This paper employs offset-quadrature phase shift keying (OQPSK) in order to suppress influences of non-linear distortion of the amplifier and sampling timing error [5]. The information sequence $b_k$ and the modulated signal $x_k$ are described as follows:

$$u_k = \exp(j\pi b_k),$$

$$x_k = u_k \exp\left(j\frac{\pi}{2}k\right).$$

Bit rate sampling of the OQPSK signals generates to $\pi/2$-shift BPSK signals in the presence of ISI. Demodulators employing equalizers can compensate ISI caused by both OQPSK and channels. The received sequence $r_k$ is as follows:

$$r_k = \sum_{i=0}^{L} h_k[i]x_{k-i} + \eta_k,$$

where $L$ denotes a channel memory length, $h_k[i]$ denotes channel impulse response (CIR) including ISI caused by OQPSK, channels and phase rotation of $\pi/2$ at the receiver side, and $\eta_k$ denotes additive noise.

Maximum delay spread in acoustic communications is much larger than that of radio frequency (RF) band wireless communications. This paper assumes that the symbol rate is 8 kbps and the maximum delay spread is from several msec to several tens of msec. This paper selects $L = 127$ in order to cope with the delay spread of about 15 msec.

## 3 Functions of receiver part

A demodulation part employs a reverse phase rotator, a timing estimator, a channel estimator, LVE and forward error correction (FEC) processing.

### 3.1 Channel estimation based on synchronization word (SW)

Assuming 8th-order M-sequence with $L$ bit cyclic prefix (CP), the estimated CIR $\hat{h}[i]$ can be simply derived without performance degradation as follows [5]:

$$\hat{h}[i] = \frac{1}{28} p_i + \frac{1}{215} \sum_{s=0}^{L} p_s,$$

$$p_s = \sum_{i=0}^{2^s-2} r_{s+i}\chi_i^M,$$

where $\chi_i^M$ is the M-sequence of the transmitted sequence, and $p_s$ is a cross-correlation value.

### 3.2 List Viterbi equalizer

The LVE is a generalized concept including DFSE and M-algorithm, where $V$ is a memory length, $2^V$ is the number of states and $S$ is the number of surviving paths. The LVE employs the metric-criteria combining scheme $\Gamma_k^{MCC}$ which combine the squared Euclidean metric-criterion, $\Gamma_k^{SQR}$, with the modified metric-criterion based on the matched filter, $\Gamma_k^{MOD}$, as follows:

$$\Gamma_k^{MCC} = \Gamma_k^{SQR} + \lambda \Gamma_k^{MOD}$$
\[ \Gamma_k^{\text{SQR}} = |r_k - \hat{r}_k|^2 \]  
\[ \hat{r}_k = \sum_{i=0}^V h_i \tilde{u}_{k-i} + \sum_{i=V+1} L h_i \tilde{u}_{k-i}^{\text{SV}} \]  
\[ \Gamma_k^{\text{MOD}} = -2 \Re \{ y_k \tilde{u}_{k}^* \} + s_0 |\tilde{u}_{k}|^2 + 2 \Re \left[ \sum_{j=1}^L s_j \tilde{u}_{k-j}^* \tilde{u}_{k-j} \right] \]  
\[ y_k = \sum_{i=0}^L \hat{h}_i^* r_{k+i} \]  
\[ s_j = \sum_{i=0}^{L-j} \hat{h}_i^* \hat{h}_{i+j} \],

where \( \tilde{u}_k \) is a candidate of the transmitted information sequence, \( \tilde{u}_k^{\text{SV}} \) is a candidate of the transmitted information sequence based on the surviving path and \( \lambda (\lambda > 0) \) is a parameter of the combining ratio.

### 4 Linear channel interpolation (LCI)

This section proposes LCI in order to track time-varying channels. LCI derives the estimated CIR \( \hat{h}_k^{\text{LCI}} \) interpolating the estimated CIR \( \hat{h}_k^{\text{pre}} \) and \( \hat{h}_k^{\text{post}} \), where \( \hat{h}_k^{\text{pre}} \) is estimated from SW at the present frame, and \( \hat{h}_k^{\text{post}} \) is estimated from SW at the next frame. Let us define that \( T_{\text{frame}} \) is a frame duration, \( k'(0 \leq k' \leq T_{\text{frame}}) \) is the discrete time number at the present frame. The proposed LCI tap coefficient \( \hat{h}_k^{\text{LCI}} \) is derived as follows:

\[ \hat{h}_k^{\text{LCI}} [i] = \frac{T_{\text{frame}} - k'}{T_{\text{frame}}} \hat{h}_k^{\text{pre}} [i] + \frac{k'}{T_{\text{frame}}} \hat{h}_k^{\text{post}} [i]. \]  

### 5 Computer simulation

This section compares packet error rate (PER) performance of MCC-LVE and SQR-DFSE, where MCC-LVE is the LVE with the MCC scheme \( \Gamma_k^{\text{MCC}} \), SQR-DFSE is the DFSE with only the squared Euclidean metric-criterion \( \Gamma_k^{\text{SQR}} \). We also evaluate influence of LCI. This paper assumes that a modulation scheme is OQPSK, transmit/receive filters are 20% root cosine roll-off, the number of received antennas \( N_R \) is 1 and FEC is Reed-Solomon code \((240, 192, 8)\). Channels suffer from 2-path independent Rayleigh fading with the delay spread of 32 bits and desired to undesired signal power ratio of 6 dB, where the maximum Doppler frequency normalized by symbol rate, \( f_{DT}T \), of 0% corresponds to quasi-static fading channels. FFF-DFE with RLS algorithm in a decision-directed mode is also evaluated for reference, where the number of feedforward filter taps is 64 and the number of feedback filter taps is 64. Fig. 2(a) shows PER performance as a function of average \( E_S/N_0 \) on quasi-static channels, Fig. 2(b) shows PER performance in the absence of noises as a function of \( f_{DT}T \). From Figs. 2(a) and 2(b), we can obtain the following results:

- MCC-LVE has better PER performance than SQR-DFSE;
- the equalizers employing the proposed LCI have better tracking performance than the others;
- FFF-DFE suffers from PER floor on the quasi-static channels; this is because the tap coefficients do not sufficiently converge, and FFF-DFE suffers from the error propagation.

Thus, computer simulation results show that the proposed MCC-LVE employing the LCI can improve not only tracking performance but also PER performance in the required signal to noise power ratio (SNR), keeping small computational complexity.

### 6 Field experiment

This section evaluates PER performance of the proposed LVE on actual field experiment. In this experiment, a carrier frequency is 22.6 kHz and a transmission rate is 8 ksps. We employ PM0.3H (FOSTEX) of a speaker and LS-100 (OLYMPUS) of a microphone. Fig. 3(a) shows a sounded CIR on quasi-static channels where the distance between the speaker and the microphone is 10 m and Fig. 3(b) shows PER performance as a function of $f_D T$ at the same point. This section evaluates the same equalizers in 5. From Fig. 3(b), we can obtain the following results:

- similar to the simulation results, the equalizers employing the proposed LCI improve tracking performance;
- in actual fields experiment, FFF-DFE seriously degrades tracking performance; FFF-DFE seems to suffer from severe error propagation compared with the simulation result.

These results confirm that the proposed LVE is suitable for acoustic communications environment.
7 Conclusion

This paper has proposed the LVE employing the LCI for large delay time-dispersive channels of acoustic communications. It can reduce computational complexity without performance degradation. In addition, computer simulation results have confirmed that the proposed LVE employing the LCI shows better tracking performance and better performance in the required SNR than FFF-DFE employing RLS algorithm. Finally, field evaluation results have confirmed the proposed LVE is suitable for acoustic communications.

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Single-coil dual-band antenna design for wireless capsule endoscopic communication in MHz band

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Abstract: Inductive coupled coil is an antenna structure that can easily realize multiple resonant frequencies with planar structure for MHz band wireless capsule endoscopic (WCE) communication. However, since one resonant frequency is generated by one coil, the antenna size is large when designing multi-band antennas. In this paper, the antenna matching circuit is modified to be a series circuit with an LC tank so that one coil can generate two resonant frequencies. For the receiving antenna, this improvement can reduce half the number of coils and achieve smaller antenna size. For the transmitting antenna that is difficult to use multiple coils due to the strict size constraints of the WCE, this improvement can achieve two resonant frequencies through one coil. In order to evaluate the feasibility and whether this improvement will reduce the antenna transmission performance, a single-coil dual-band antenna operating in the MHz band is designed and manufactured, and the simulated and measured reflection and transmission coefficients are evaluated by a biological tissue equivalent phantom. By comparing with the previously proposed antenna results, it can be known that the transmission performance is not deteriorated, indicating that the improvement is feasible for the WCE transmitting and receiving coil antennas.

Keywords: wireless capsule endoscopic communication, low frequency communication, multi-band communication, coil antenna, matching circuit

Classification: Antennas and Propagation

References


1 Introduction

Wireless Capsule Endoscope (WCE) is a device that can conduct gastrointestinal (GI) tract inspection without the use of cables [1]. Images can be transmitted wirelessly through a transmitting antenna inside WCE and a receiving antenna on the surface of the patient’s body [2]. The Medical Implant Communication System (MICS) is widely used for medical communication (core band: 402–405 MHz), but narrow bandwidth limits the transmission rate of the device [3]. On the other hand, according to Japanese law, the 10–60 MHz band is license-free, and wide frequency band offers the possibility of high-speed medical communication [4].

In [5], Wang, et al. designed and manufactured a transceiver for 10–60 MHz band human body communication. Impulse radio (IR) signal is adopted to decrease the size of the transceiver. The transceiver output power is shown in Fig. 1. It can be seen that the transceiver contains five signal peaks (19.3, 28.9, 38.5, 48.2 and 57.6 MHz) in the 10–60 MHz band. In order to obtain the strongest signal, the antenna resonant frequencies are usually designed to match the signal peaks of the transceiver. In [6], the corresponding WCE transmitting and receiving antennas are designed with the resonant frequency of approximately 30 MHz. Both antennas are

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capsule-shaped, which achieves a total data rate of 1.25 Mbps through the transceiver in [5]. However, the receiving antenna is not suitable for design in a capsule shape as it is difficult to be mounted on the surface of the patient’s body during the inspection. In addition, the two antennas can only resonate at one signal peak of the transceiver, which fails to fully utilize signal energy of the transceiver.

As another antenna structure used for power and signal transmission in the MHz band, the inductive coupling coil can easily realize the planar antenna structure and multiple resonant frequencies [7, 8, 9, 10]. In our previous study [9, 10], the authors designed and fabricated planar-structured dual-band and tri-band coil receiving antennas for WCE signal reception in the 10–60 MHz band, and a single-band transmitting antenna for communication performance evaluation. However, since one coil generates only one resonant frequency, the use of multiple coils results in a relatively large size of the receiving antenna. In addition, due to the strict size limitations of the capsule, it is difficult for the transmitting antenna to use multiple coils, therefore only one resonant frequency can be generated.

In this paper, by using a series circuit with an LC tank in the antenna matching circuit, the authors attempt to reduce the total number of coils while maintaining the same number of resonant frequencies. In order to evaluate the feasibility and whether this improvement will reduce the antenna transmission performance, a single-coil dual-band antenna is designed and fabricated for WCE communication in the 10–60 MHz. By comparing with the previously proposed antenna results [6], it is possible to evaluate the transmission performance difference and the feasibility of the improvement.

2 Antenna structure with simulated and measured results

Instead of using a series and parallel circuit, the improved antenna uses a series circuit with an LC tank so that two resonant frequencies can be realized by one coil. When the coil operating frequency is lower or higher than LC tank resonant frequency, the LC tank can be equivalent to an inductor or a capacitor, and two resonant frequencies can be obtained by connecting the inductive or capacitive LC tank with the series circuit. Detailed principle and the resonant frequency calculation formula can be obtained in [11].

In Fig. 2(a), the structure and dimensions of the proposed antenna are shown, which is optimized by Computer Simulation Technology Microwave Studio 2018.
Photographs of the fabricated antenna top and bottom side are shown in Fig. 2(b). The antenna is made of a 1.6 mm thick FR-4 substrate ($\varepsilon_r = 4.3$; $\tan \delta = 0.035$), a coil (width: 0.5 mm; pitch: 1 mm; copper thickness: 0.017 mm), two metal vias (diameter: 0.5 mm), with a matching circuit and GND at
the bottom of antenna. Inductance and capacitance values of the matching circuit for the simulation are: $C_1 = 3.4 \, \text{pF}$, $C_2 = 4.7 \, \text{pF}$, $L_1 = 1 \, \mu\text{H}$, and $L_2 = 0.95 \, \mu\text{H}$, respectively. Inductance and capacitance values of the matching circuit for the measurement are: $C_1 = 3 \, \text{pF}$, $C_2 = 4.7 \, \text{pF}$, $L_1 = 1 \, \mu\text{H}$, and $L_2 = 0.82 \, \mu\text{H}$, respectively. By re-adjusting the antenna inductance and capacitance values after fabrication to ensure that the antenna can resonate at the peak of the transceiver.

A biological-equivalent phantom is used for antenna simulation and measurement, as shown in Fig. 2(c). Since the authors only designed the receiving antenna, two identical proposed antennas are used to evaluate the antenna transmission performance, with 1 mm gap from the phantom and 50 mm distance of two antennas. In the measurement, the distance between the antenna and the phantom can be adjusted by covering the phantom with a plastic wrap, and the electrical properties of the plastic wrap are negligible. The measured relative permittivity and conductivity of the phantom are 72.9 and 1.21 S/m at 50 MHz, with small change in the 35–65 MHz frequency range. To simplify antenna simulation, the measured electrical properties are used for antenna simulation in the entire frequency range.

The simulated and measured antenna reflection and transmission coefficients are shown in Figs. 3(a) and 3(b). Two resonant frequencies that match the signal peak of the transceiver can be observed at 38.5 and 57.6 MHz, with 50-mm transmission coefficients of $-18.6$ and $-26.9$ dB for the simulated results, and $-14.1$ and $-25.8$ dB for the measured results, respectively. Since there is a weak coupling between the coaxial cables at low frequencies, the measured transmission coefficient is slightly higher than the simulated result [9]. The measured reflection coefficient is lower than the simulated result as measurement errors occur when plastic wrap is used to adjust the distance between the antenna and the phantom.

![Fig. 3.](image-url)

(a) Simulated antenna reflection and transmission coefficient.
(b) Measured antenna reflection and transmission coefficient.
(c) Simulated reflection coefficient of different $C_2$.
(d) Simulated reflection coefficient of different $L_2$. 

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As a comparison in [6], two resonant frequencies that match the signal peak of the transceiver can be observed at 48.2 and 57.6 MHz, with 50-mm transmission coefficients of $-19$ and $-37.9$ dB for the simulated results, and $-14.3$ and $-32.7$ dB for the measured results, respectively. It can be seen that the transmission performance at the signal peak of the transceiver is not deteriorated, indicating that the improvement is feasible for the WCE transmitting and receiving coil antennas.

In order to clarify whether different capacitance and inductance values of the LC tank can influence the antenna resonant frequency, the authors simulated the antenna transmission coefficient with different inductance and capacitance values of the LC tank, and results are shown in Figs. 3(c) and 3(d). It can be seen that as the capacitance and inductance values increase, the higher resonant frequency of the antenna (57.6 MHz) decreases, while the lower resonant frequency of the antenna (38.5 MHz) hardly changes. The simulated results show that different capacitance and inductance values of the LC tank will influence the resonant frequency of the antenna, and the degree of two resonant frequencies is different. This is good news for antenna frequency adjustment and can be used to better guide future antenna designs.

3 Conclusion

In this paper, the authors improved the antenna matching circuit by using a series circuit with an LC tank to reduce the total number of coils while maintaining the same number of resonant frequencies. A single-coil dual-band antenna is designed and fabricated to evaluate the feasibility of the improvement. The simulated and measured results show that two resonant frequencies can be realized by one coil and the antenna transmission performance at the signal peak of the transceiver is not deteriorated, indicating that this improvement can be used to design WCE transmitting and receiving antennas.

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Experimental study on polarimetric-HoloSAR

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Abstract: This paper pursuits a possibility of fully polarimetric and three-dimensional (3-D) imaging by Holographic Synthetic Aperture Radar Tomography (HoloSAR). HoloSAR is 3-D extension of Circular Synthetic Aperture Radar (CSAR) along the vertical direction whose trajectory is a circle in a horizontal plane. By adding fully polarimetric capability to HoloSAR, it becomes a perfect 3-D imaging radar system which reveals scattering mechanisms from object through scattering power decomposition. This paper attempts to achieve such a prototype system in an anechoic chamber. The experimental result successfully yielded a 3-D image reconstruction of objects with exhibiting polarimetric scattering mechanism.

Keywords: Synthetic Aperture Radar (SAR), CSAR, HoloSAR, scattering mechanism

Classification: Sensing

References


1 Introduction

With the advent of Synthetic Aperture Radar (SAR), various functions to SAR systems and data acquisition methods have been applied and added to improve the system performances. Recent SAR advances are well explained in [1], including polarimetry, interferometry, tomography, etc. One goal of SAR applications will be 3-D reconstruction of objects under imaging. Since precise 2-D imaging around 360° is available now by Circular SAR whose trajectory is circular in a horizontal plane, it will be possible to use multiple CSAR in the vertical direction for 3-D imaging.

The pioneering HoloSAR work [2] explains the principle and 3-D reconstruction method in detail together with related problems. This work intends the HoloSAR by adding polarimetric scattering power decomposition scheme [3] to achieve advanced 3-D image reconstruction system. If this kind of Polarimetric-HoloSAR system becomes available, precise 3-D image reconstruction with scattering mechanisms will be available for future applications.

In the following, a brief HoloSAR principle using backprojection method is described in Sect. 2, followed by polarimetric scattering power decomposition scheme in Sect. 3, and the experimental results in Sect. 4. Sect. 5 is the conclusion.

2 HoloSAR methodology

Among various CSAR schemes, we use Step Frequency Continuous Wave (SF-CW) radar system [4] for the measurement based on a vector network analyzer.

2.1 Formulation

Fig. 1 shows the HoloSAR measurement configuration. The aspect angle $\varphi$, and height $Z_c$ are defined as

$$\varphi = k\Delta\varphi \quad (k = 0, 1, 2 \ldots, N_{\varphi})$$
$$Z_c = Z_{\min} + l\Delta Z_c \quad (l = 0, 1, 2 \ldots, N_{Z_c})$$

where $\Delta\varphi$ and $\Delta Z_c$ are sampling intervals, $N_{\varphi}$, $N_{Z_c}$ are number of sampling points, and $Z_{\min}$ is the minimum height. Step frequency $f$ is defined by

$$f = f_0 - \frac{B}{2} + m\frac{B}{N_f} \quad (m = 0, 1, 2 \ldots, N_f)$$

where $f_0$ is the center frequency, $B$ is bandwidth, and $N_f$ is number of steps. Transmitted impulse signal $s$ of SF-CW radar is written as

$$s(t, \omega) = \exp(j\omega t).$$

In the model shown in Fig. 1, the received signal $p$ for the point target $(x_n, y_n, z_n)$. $p$ is represented by

$$p(t, \omega, \varphi, Z_c) = \sum_{n=1}^{N_{tgt}} g_n \exp\{j\omega(t - \tau_n(\varphi, Z_c))\}$$

where $N_{tgt}$ is number of point objects, $g_n$ is reflection coefficient, $\tau_n$ is round trip time delay from antenna position to the position $(x_n, y_n, z_n)$ of object $n$. The time delay $\tau_n$ is represented by
where $R_g$ is trajectory radius. The transmission and reception ratio $S$ becomes

$$S(\omega, \varphi, Z_c) = \frac{p(t, \omega, \varphi, Z_c)}{s(t, \omega, \theta, Z_c)} = \sum_{n=1}^{N_\omega} g_n \exp\{-j\omega \tau_n(\varphi, Z_c)\}. \quad (7)$$

The HoloSAR reconstruction is achieved by calculating $g$ in (8) using the back-projection method. $g$ is represented by

$$g(x, y, z) = \sum_{k=0}^{N_f} \sum_{l=0}^{N_p} \sum_{m=0}^{N_c} S(\omega, \varphi, Z_c) \exp\{j\omega t(x, y, z, \varphi, Z_c)\}$$

$$t(x, y, z, \varphi, Z_c) = \frac{2\sqrt{(x - R_g \cos \varphi)^2 + (y - R_g \sin \varphi)^2 + (z - Z_c)^2}}{c} \quad (9)$$

where $t$ in exponential term is round trip time delay from antenna position to space $(x, y, z)$.

2.2 Image reconstruction

HoloSAR reconstruction is implemented with the following steps using back-projection scheme.
1. Range compression of raw data

\[ S(t) = F_{(t)}[S(\omega, \varphi, Zc)]^{-1} \]

where, \( F_{(t)}[.]^{-1} \) is fast time inverse Fourier transform.

2. Approximation of round trip time delay from antenna position to space

\[ t(x, y, z, \varphi, Zc) = t(x_p, y_q, z_r, \varphi, Zc) \approx \frac{2}{c} \sqrt{(x_p - R_e \cos \varphi)^2 + (y_q - R_e \sin \varphi)^2 + (z_r - Zc)^2} \]

where, \( x_p, y_q, z_r \) are represented as followed using size of one boxel \( \Delta x \times \Delta y \times \Delta z \).

\[ x_p = x_{min} + p\Delta x \quad (p = 0, 1, 2 \ldots N_p) \]
\[ y_q = y_{min} + q\Delta y \quad (q = 0, 1, 2 \ldots N_q) \]
\[ z_r = z_{min} + r\Delta z \quad (r = 0, 1, 2 \ldots N_r) \]

where, \( x_{min}, y_{min}, z_{min} \) are minimum analysis area, \( N_p, N_q, N_r \) are number of boxels in each direction.

3. Taking the sum

\[ g(x_p, y_q, z_r) = \sum_{p=0}^{N_p} \sum_{q=0}^{N_q} S(t)(\omega, \varphi, Zc) \exp\{j\omega t(x_p, y_q, z_r, \varphi, Zc)\} \frac{1}{N_p N_q}. \]

### 3 Polarimetric scattering power decomposition

By the polarimetric measurement using the combination of Tx and Rx with H-pol. and V-pol. basis, it is possible to obtain scattering matrix. This matrix can be transferred to Pauli vector,

\[ k_p = \frac{1}{\sqrt{2}} \begin{bmatrix} O_{HH} + O_{VV} \\ O_{HH} - O_{VV} \\ 2O_{HV} \end{bmatrix} \]          \tag{10} \]

where \( O_{HH}, O_{HV}, O_{VH}, O_{VV} \) are observed scattering matrix elements after the reconstruction processing. The Pauli vector is used to create coherency matrix which retain the second order statistics of polarimetric information. The ensemble average of the coherency matrix, denoted as \( \langle \{T\} \rangle \), is defined by,

\[ \langle \{T\} \rangle = \frac{1}{N_{ave}} \sum_{k} k_p^{*} k_p \]

where \( \dagger \) is complex conjugate and transpose, and \( N_{ave} \) is the number of averaged boxels. The scattering power decomposition method expand the measured coherency matrix into sub-matrices representing scattering models [5], as follows:

\[ \langle \{T\} \rangle = f_s\{T\}_{\text{surface}} + f_d\{T\}_{\text{double}} + f_v\{T\}_{\text{volume}} + f_{\ell}\{T\}_{\text{helix}}. \]

By comparing the quantity of the measured data and scattering model, we can retrieve the following 4-scattering powers to be used in the next section.
The surface scattering power $P_s$ (Blue color assigned)
The double bounce power $P_d$ (Red)
The volume scattering power $P_v$ (Green)
The helix scattering power $P_c$ (Yellow)

4 Experimental results

In this section, we pursue a possibility of Polarimetric-HoloSAR to retrieve scattering mechanism from reconstructed images of concrete blocks in an echoic chamber measurement.

4.1 Measurement specification

Measurement specifications and object configuration are shown in Fig. 2. We used several concrete blocks with cube of 0.1 m, placed on an iron plate. The combination of these blocks were modeled as a building, collapsed building, and four isolated houses as shown in Fig. 2(b) for top view. Fig. 2(c) is a photo taken from $\varphi = 0^\circ$, along the x axis of Fig. 2(b). Fig. 2(d) is a model of collapsed building whose maximum height is 0.3 m. Elevation angle of four isolated houses are $0^\circ$, $10^\circ$, $20^\circ$, $30^\circ$ from the top of Fig. 2(b) with respect to x-axis. The origin of antenna height is set on the iron plate surface at $z = 0$. At each height, we used Inverse SAR (ISAR) scheme [6] to obtain fully polarimetric data set in a horizontal plane with circular trajectory.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Center frequency</td>
<td>15 GHz</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>4 GHz</td>
</tr>
<tr>
<td>Number of steps</td>
<td>401 steps</td>
</tr>
<tr>
<td>Off-nadir angle</td>
<td>60°</td>
</tr>
<tr>
<td>Trajectory radius</td>
<td>2.25 m</td>
</tr>
<tr>
<td>Height interval</td>
<td>0.02 m</td>
</tr>
<tr>
<td>Antenna height</td>
<td>0.8~1.8 m</td>
</tr>
<tr>
<td>Aspect angle interval</td>
<td>0.2°</td>
</tr>
</tbody>
</table>

(a) Radar parameters

Fig. 2. Experimental situation
4.2 3-D image reconstruction

The data sets of fully polarimetric data by HoloSAR measurement were used to create 3-D image using the following parameters,

- Size of one boxel: $\Delta x \times \Delta y \times \Delta z = 0.01 \times 0.01 \times 0.01$ [m]
- Analysis area: $-1 \leq x \leq 1$, $-1 \leq y \leq 1$, $-0.5 \leq z \leq 0.5$ [m]
- Averaging size: $x \times y \times z = 3 \times 3 \times 3$ [boxel]
- Threshold: maximum power [dB] $-20$ [dB]

Fig. 3 shows a 3-D and fully polarimetric scattering power image of the scene under test (Fig. 2). The movie file Fig. 3(a) shows the reconstructed image viewed from $360^\circ$ directions. Fig. 3(b) shows the image seen from top, while (c) is side view image (seen at $\varphi = 0^\circ$).

As can be seen in Fig. 3, the image shows accurate mapping in the horizontal plane. The plotted box position is corresponding to the actual position in the measurement situation. The accurate 2-D positioning is the most advantageous of multi-lateral observations of HoloSAR.

The second advantage is an estimation of the maximum object height by vertical information of HoloSAR measurement. Also in the block with elevation
angles of 20° and 30°, strong surface scattering \( P_s \) is seen. Since the elevation angle of a typical houses are 15° to 30°, the house height can be detected.

These above two advantages may be common to HoloSAR systems. The most important and third advantage of Polarimetric-HoloSAR is correct building detection capability by strong scattering power \( P_d \) caused by the double bounce scattering mechanism. By circular trajectory observation, it is possible to receive very strong double bounce scattering generated by the right angle structure of building walls and the ground. This happens four times if the building is rectangular. By adding the polarimetric capability, it becomes possible to classify or identify objects more accurately.

5 Conclusion

In this paper, we examined the possibility of fully polarimetric and HoloSAR to retrieve 3-D scattering information from objects under imaging scene. Basic experimental results showed advantages of HoloSAR for 1) accurate 2-D positioning in the horizontal plane, and 2) height information in the vertical direction. Furthermore, by polarimetric measurement, the capability of scattering mechanism information retrieval is added to the HoloSAR system, which leads to classify or identify objects accurately. If this kind of polarimetric HoloSAR systems becomes available, precise 3-D image reconstruction will be available for future applications.