16-QAM signal pre-equalization using multiple binary analog FIR filters based on 28-nm FD-SOI for dispersion compensation

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Abstract: We propose a novel pre-equalizer for multi-level IQ-modulation signals. The pre-equalizer is composed of multiple binary analog finite impulse response (FIR) filters which can realize smaller integrated-circuit (IC) chip size. In the proposed design scheme, required complex-number operations are implemented by a butterfly construction of real-number FIR filters. The analog FIR filters were designed using 28 nm fully depleted silicon on insulator (FD-SOI) based complementary metal-oxide-semiconductor (CMOS) circuits. The performance was investigated using numerical simulation of a 40-Gb/s 16-ary quadrature amplitude modulation (16-QAM) standard single-mode fiber (SSMF) transmission system. The simulated CMOS circuits successfully compensated the 16-QAM signals distorted by chromatic dispersion. The error vector magnitude (EVM) of the transmitted 16-QAM signals was improved from 27% to 12%.

Keywords: equalization, FIR filter, 16-QAM, CMOS circuit

Classification: Fiber-Optic Transmission for Communications

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1 Introduction

Higher speed, larger capacity data transmission is required in fiber optic communication links to accommodate rapidly increasing data traffic. Multi-level modulation is one key technology to increase the transmission capacity. Quadrature amplitude modulation (QAM) including 16-QAM and 64-QAM have been experimentally investigated, where in-phase (I) and quadrature (Q) components of lightwave are modulated by multi-level signals [1, 2]. Higher order QAM schemes have also been studied to realize higher spectral efficiency [3]. However, these multi-level IQ-modulation signals are susceptible to waveform distortion caused by, e.g., chromatic dispersion (CD) in comparison with binary modulations due to the
smaller inter-symbol distance. Digital signal processing (DSP) plays an important role in compensating for waveform distortion caused by CD. Finite impulse response (FIR) filters are often used in such systems [4, 5]. On the other hand, analog FIR filters based on analog electronic circuit technology are also an attractive approach for realizing cost-effectiveness and low power consumption [6, 7, 8, 9, 10]. We have investigated binary analog FIR filters using binary delay-line components based on complementary metal-oxide-semiconductor (CMOS) inverters which can realize smaller integrated-circuit (IC) chip size [11]. Recently, we proposed a novel multi-level pre-equalizer using the binary FIR filters [12]. However, they can be used only for 1-dimensional (1D) modulation signals such as pulse-amplitude-modulation (PAM) because the analog FIR filter can only operate real numbers. In case of 2-dimensional (2D) modulations such as IQ-modulations, I- and Q-components of the signals and tap-coefficients of the FIR filters are expressed by complex numbers [4]. In this paper, we propose a novel pre-equalizer using multiple binary analog FIR filters for multi-level IQ-modulation signals. The FIR filters were designed using 28-nm fully depleted silicon on insulator (FD-SOI) based CMOS circuits. The performance was investigated by numerical simulations of an optical communication system for 40-Gb/s 16-QAM standard single-mode fiber (SSMF) transmission.

2 Construction of proposed pre-equalizer

Fig. 1(a) shows block diagram of an FIR filter with complex-number tap-coefficients used for IQ-modulation signals. Input signal, \( X(n) \), and output signal, \( Y(n) \), are given by complex numbers as

\[
X(n) = X_I(n) + jX_Q(n), \\
Y(n) = Y_I(n) + jY_Q(n),
\]

where \((X_I(n), X_Q(n))\) and \((Y_I(n), Y_Q(n))\) are the sets of I- and Q-components of the modulated signals. \( n \) is the number of symbol of the signals. \( k \)-th tap-coefficient of the FIR filter, \( h_k \), can be expressed as

\[
h_k = h_{r,k} + jh_{i,k},
\]

where \( h_{r,k} \) and \( h_{i,k} \) are real part and imaginary part of the tap-coefficient, respectively. Output, \( Y(n) \), can be calculated as shown below.

\[
Y(n) = \sum_{k=0}^{M-1} h_kX(n-k) \\
= \sum_{k=0}^{M-1} (h_{r,k} + jh_{i,k})\{X_I(n-k) + jX_Q(n-k)\} \\
= \sum_{k=0}^{M-1} h_{r,k}X_I(n-k) - \sum_{k=0}^{M-1} h_{i,k}X_Q(n-k) \\
+ j \sum_{k=0}^{M-1} h_{i,k}X_I(n-k) + j \sum_{k=0}^{M-1} h_{r,k}X_Q(n-k)
\]

Therefore, I- and Q-components of output signal can be expressed as
Equations (5) and (6) can be implemented with butterfly construction of four real-number FIR filters as shown in Fig. 1(b). The four terms on the right side of Eq. (5) and (6) are represented by the four blocks of FIR filters in the figure. In case of 16-QAM modulation, $X_I(n)$ and $X_Q(n)$ are expressed by 4-level (2 bit/Symbol) signals. We have proposed and investigated multi-level FIR filter composed of binary analog FIR filters as shown in Fig. 1(c) [12]. In the figure, $X_I(n)$ and $X_Q(n)$ are the most-significant bit (MSB) and least-significant bit (LSB) of the 4-level signal, respectively. One multi-level FIR filter consists of two parallel binary FIR filters and a 2/1-weighted adder. Therefore, the complex-number FIR filter shown in Fig. 1(a) can be implemented by using our proposed binary analog FIR filters. It should be noted that this proposed construction scheme is fully scalable to increase the multi-level of the modulation format. For example, an 8-level FIR filter is composed of three parallel binary FIR filters and a 4/2/1-weighted adder, which realize 64QAM modulation. The binary analog FIR filters are composed of multipliers implemented with Gilbert Cells, as shown in Fig. 1(d) and binary delay lines implemented with CMOS inverters, as shown in Fig. 1(e), realizing cost-effective-
ness and low power consumption. The circuits were designed using 28 nm FD-SOI based CMOS circuit [11]. The maximum output voltage and slew rate (SR) of the Gilbert Cells were about 200 mV and 5.2 mV/ps, respectively. It should be noted that the circuit has a sufficiently high response speed for 10 Gsymbol/s signals. The maximum output voltage and SR of the CMOS inverters were about 1.0 V and 81 mV/ps, respectively. This slow response speed of the inverters were used to realize the binary delay lines [11].

3 System setup

Fig. 2 schematically shows the system setup used in our numerical simulation for 40-Gb/s 16-QAM transmission over 50 km of SSMF. 16-QAM optical signal was modulated by a pair of 4-level signals which were pre-equalized by the analog FIR filter calculated by a SPICE circuit simulator, resulting in a pre-equalized optical 16-QAM signal. The 4-level signals were composed of four sets of PRBS $2^9 - 1$ binary data. The pre-equalizer had four taps whose tap coefficients had been determined by a least-mean-square (LMS) algorithm. The modulated optical signal was transmitted by a 50 km SSMF. The accumulated CD was 837.5 ps/nm. The received optical power was adjusted using an attenuator (ATT). The transmitted signal was detected by optical homodyne detection. Electrical noise was taken into account at PDs as thermal noise with power density of $1 \times 10^{-10}$ pW/Hz. Here, we assumed that the local oscillator (LO) was ideally synchronized to the optical signal.

![Fig. 2. System setup of 40-Gb/s 16-QAM transmission over 50 km SSMF](image)

4 Results and discussion

Figs. 3(a) to (f) show eye-diagrams and constellations of the transmitted 16-QAM signals. Figs. 3(a) and (d) show the eye-diagram and the constellation for the case where the pre-equalization was not employed. The waveform was completely distorted by CD in the SSMF. Figs. 3(b) and (e) show the eye-diagram and the constellation for the case where pre-equalization using the proposed analog FIR filter was employed. Clear eye-openings and clear constellation were achieved by using the pre-equalizer. Figs. 3(c) and (f) show the eye-diagram and the constellation for the case where pre-equalization using an ideal FIR filter was employed. By comparing the results of the analog and ideal FIR filters, we can observed some residual waveform distortion. The residual distortion was due to imbalances among the binary FIR filters used. The analog FIR filter consists of eight binary FIR filters connected by analog adders as explained in section 2. The distortion is caused by insufficient linearity of the adders. This problem, however, will be resolved by optimizing the circuit parameters of the analog FIR filter. Calculated EVM values
versus optical signal-to-noise ratio (OSNR) are shown in Fig. 3(g). In the figure, we also plotted the EVM for the case of the ideal FIR filter. Even when using the ideal FIR filter, the EVM did not become 0% due to the electrical noise of the receiver. In the case without the pre-equalization, the EVM did not decrease to less than 27%, even when we improved the OSNR. In the case with the pre-equalization, however, the EVM was improved to less than 13%. The degradation of the EVM from using the ideal FIR filter was only about 3 dB. The results clearly showed the effectiveness of our proposed pre-equalizer.

5 Conclusion

We proposed a novel pre-equalizer for multi-level IQ-modulation signals. The pre-equalizer is composed of multiple binary analog FIR filters which can realize smaller IC chip size. This equalization technology should encourage the adoption of analog FIR filters in future cost-effective and low-power-consumption optical coherent systems.

Fig. 3. Eye-diagrams, constellations of the transmitted signals and EVM characteristics.
Preliminary experiment of meteor burst communications in equatorial region

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Abstract: The aim of the study is to explore the possibility of the use of Meteor Burst Communications (MBCs) in equatorial regions. We installed the master and the remote stations at Yogyakarta, Java Island and Jimbaran, Bali Island, Indonesia, respectively. As a preliminary experimental result, we confirmed that some packet transmissions between the two stations were achieved through meteor burst channels.

Keywords: meteor burst communications, equatorial regions, Indonesia

Classification: Space Utilization Systems for Communications

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1 Introduction

When meteors enter into the Earth’s atmosphere, ionized gas columns, which are called “meteor bursts,” are generated by friction with the atmosphere. Meteor Burst Communications (MBCs) are non Line-of-Sight communication methods using reflection phenomena of low VHF band waves by the meteor bursts [1, 2, 3, 4, 5, 6, 7]. MBC systems can realize data transmission between stations apart from each other by at most 2000 km. Moreover, MBC systems can be easily and inexpensively constructed compared with satellite communications, etc.

Since the existence time of the meteor bursts is as short as about several hundreds milliseconds, MBCs are suitable for communications which need small data transmission and allow some delay. For this reason, MBC systems are applicable for data transmission for environmental and meteorological observation systems which allow some delay and do not require massive data transmissions.

A famous application of MBCs is SNOTEL operated by Natural Resources Conservation Service (NRCS), United States Department of Agriculture [3]. SNOTEL is a snow telemetry system and collects weather data from the Rocky Mountains and Alaska. About 800 remote stations of SNOTEL are located where access is often difficult or restricted. Several master stations collect the data from the remote stations and transmit them to the central computer of NRCS.

Some MBC experiments and operations have been also done in high-latitude regions [5, 6, 7]. The experiments exhibit some interesting results from various points of view such as within-day variation and seasonal variation. From the experiments in Antarctica, some unique results have been obtained, which are different from the results in mid-latitude regions.

However, to the best of our knowledge, there have been no MBC experiments and operations in equatorial regions. Some astronomical observation results in equatorial regions showed that the appearance of meteors has some difference with that in mid-latitude regions, e.g., there are smaller variations in the height of meteor bursts, smaller seasonal fluctuations, etc [8, 9]. Therefore, we expect to obtain unique properties about MBC in equatorial regions.

In this study, we try to conduct MBC experiments to obtain communication performance of MBCs and to explore the possibility of MBCs in equatorial regions. In May 2017, we have conducted a preliminary experiment about MBCs in...
Indonesia. In this article, we provide the result of the MBC experiment and show the possibility of using MBCs in equatorial regions.

2 Conditions of the experiment

2.1 The locations and the period of the experiment

The locations of the stations are listed below:

- The master station: Gadjah Mada University, Yogyakarta, Java Island, Indonesia (S 7°48’5”, E 110° 21’52”), time zone: +7.
- The remote station: Udayana University Jimbaran Campus, Jimbaran, Bali Island, Indonesia (S8°22’9”, E115°8’18”), time zone: +8.

The stations are located in approximately an east-west direction and the distance between them is about 530 km. We carried out the experiment from May 9 to May 11, 2017 including installation of the stations. The effective period of the experiment was about 26 hours, which was from May 10, 2017, 6:00 am UTC to May 11, 2017, 8:00 am UTC.

2.2 Equipment

The antennas of both the master and remote stations are Yagi antennas with 3 elements. The elevation angle of the antennas is set at 18°. The direction angle of the master and remote antennas are 97° and 276° clockwise from the North. Both stations used the same type of MBC modems: MRC565, Maiden Rock Communications, LLC. The specifications of the modems are: the carrier frequency is 48.375 MHz, the transmission rate is 4000 bps, the modulation scheme is DEPSK and the transmission power is set at 100 W. The remote station periodically makes 20-byte test data every five minutes for sending to the master station.

3 Data transmission procedure

Fig. 1 illustrates the packet transmission procedure. In this figure, $M \rightarrow R$ means packets from the master station to the remote station and $R \rightarrow M$ means packets in vice versa. The principal of the packet transmission procedure is explained as follows:

- The master station periodically sends Probe Packets (PPs) to the remote station.
- When the remote station successfully receives a PP, the remote station transmits a Data Packet (DP) to the master station.
- If the master station successfully receives the DP, the master station sends an Acknowledgment Packet (AP) to the remote station.
If the remote station receives the AP, the remote station removes the data which has been transmitted to the master station from the buffer in the remote station.

As shown in this figure, the time required for the successful packet transmission is about 250 ms.

We note that the shorter the period of transmission of PPs is, the better performance the master station yields. However, we set the duration of PPs to 500 ms to prevent a heat problem with the modem.

4 Experimental result

Fig. 2 shows the observed noise power at both of the stations during the experiment period. Since acceptable noise power for the MBC system is $-115$ dBm or less, we observed that most of the the experiment period was appropriate for the MBC experiment. However, at the master station, the noise power level was slightly higher than the acceptable noise level in the beginning of the experiment period.

Fig. 3 shows the number of received packets. By the transmission procedure, the received PPs and APs were counted at the remote station and the received DP was counted at the master station.

From this figure, we observe that, during most of the experiment period, PPs were received by the remote station even when the number of them was not so large. The result means that the reflection of PPs by meteor bursts occurred during most of the period.

We also observe that many data transmissions were achieved in the period from May 10, 8:00 pm to May 11, 4:00 am UTC. In order to clearly identify a reason of the phenomena, we would need further experiments to collect more data. For now we suppose the following possibilities.

One possible reason is that the number of meteors increased during the period. The period-of-interest was during the morning in Indonesia. Some past studies in mid-latitude regions showed that many meteor burst channels were established in the morning, especially, around the time near sunrise. Another possible reason is related to the noise level. From Fig. 2, we find that the noise power level of the period-of-interest was lower than other periods. We also need to take other possible reasons, e.g., sporadic E propagation, into account.

The total numbers of received PPs, DPs and APs were 390, 79 and 51, respectively. Since the transmission interval of PPs was set to 500 ms, the number of probe packets was about 7200 per hour. Therefore, the PP reception rate was about 0.208% ($= 390/(7200 \times 26)$).

Since the number of accepted PPs at the remote station should be the same as the number of data packets, 390 data packets have been transmitted from the remote station. Therefore, the acceptance ratio of the DPs at the master station was 20.3% ($= 79/390$). From the viewpoint of the remote station, out of 79 DPs, the remote station recognized that 51 data packets have been accepted by the master station. Thus the successful data transmission rate for the remote station was 64.6% ($= 51/79$).
5 Conclusion

In this article, we provided an overview of our preliminary MBC experiment in Indonesia and showed the possibility of applying MBC in the equatorial region. Since we expect to observe unique properties and/or communication performance of MBCs in equatorial regions, we will continue MBC experiments to obtain additional results.

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Joint rate and power optimization for SIC in uplink NOMA systems

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Abstract: To implement K-user multi-rate multi-power transmission in non-orthogonal multiple access systems with successive interference cancellation, a repeat accumulator code serially concatenated with a spreading is employed for each user to implement a variable, low-rate coding. A joint rate and power optimization (RPO) is proposed to maximize the sum rate with error free decoding. Numerical results show that our proposed coding-spreading scheme with joint RPO, supporting the multi-rate transmission with the same structure of encoder, approaches the Shannon limit.

Keywords: non-orthogonal multiple access systems, successive interference cancellation, joint rate and power optimization

Classification: Fundamental Theories for Communications

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1 Introduction

For non-orthogonal multiple access (NOMA) systems with successive interference cancellation (SIC), the Shannon limit can be achieved by power-only optimization
(PO) as long as each user has an ideal channel code. Previous works [1, 2, 3] are mainly based on this assumption of ideal code. It is intuitively to think that code design for NOMA systems is not necessary, since each user can employ the conventional well-designed channel codes directly.

However, NOMA systems need to support massive of users and multiple rates to meet the requirements for diverse services. An increase in the number of users will result in very serious multi-user interference, so that some users need low-rate codes to guarantee the decoding performance. Moreover, such systems should support various rates for the various multimedia services. Employing conventional well-designed low-rate LDPC or Turbo code, is impractical since users have to employ difference encoders to realize multi-rate transmission. It is required to design a bank of variable, low-rate channel codes with the same structure of encoder.

In this paper, for $K$-user multi-rate multi-power NOMA systems with SIC, we use a repeat accumulator (RA) code serially concatenated with a spreading for each user to implement a variable, low-rate coding, where the transmission rate can be changed flexibly by varying the repeat number and spreading length. Here, users employ the identical RA encoder, although their transmission rates are different. Moreover, both RA and repetition (spreading) codes are simple and computationally efficient. To enhance the system performance, we propose a joint rate and power optimization (RPO) to maximize the sum rate. Numerical results show that our proposed coding-spreading scheme with joint RPO, supporting the multi-rate transmission with the same structure of encoder, approaches the Shannon limit.

2 System model

In this section, we simply introduce $K$-user multi-rate multi-power NOMA systems in Fig. 1.

Suppose there are $K$ active users simultaneously transmitted information to a base station. For the $k$th user, we employ a rate-$1/q_k$ RA code serially concatenated with a length-$l_k$ spreading. The $k$th user’s transmission rate is $r_k = 1/(q_k l_k)$. Adjusting information bit lengths $B_k$ guarantees the same length of transmitted signal vectors $B_k q_k l_k \equiv N$ for each user and realizes multi-rate transmission. The sum rate is $R_{\text{sum}} = \sum_{k=1}^{K} r_k$. Let $p_k$ be the transmission power of the $k$th user. The total power is $P_{\text{sum}} = \sum_{k=1}^{K} p_k$. The receiver gets a superimposed signal vector $y = (y_1, \cdots, y_N)$ at time $j$ is
\[ y_j = \sum_{k=1}^{K} \sqrt{p_k x_j^{(k)}} + z_j, \quad j = 1, \cdots, N, \]  

(1)

where \( z_j \) is a zero-mean Gaussian variable with a variance of \( \sigma^2 \). The SIC performs the decoding process of each user successively [4].

### 3 Joint rate and power optimization (RPO)

In this section, for given \( K, P_{\text{sum}} \), and \( \sigma^2 \), we give the optimal rate and power profile, which achieves the optimal sum rate with error free decoding.

Before proceeding, let us consider a single-user system with a rate-1/q RA code serially concatenated with a length-l spreading under the interference power \( \zeta \) and noise power \( \sigma^2 \). For given transmission power \( p \), we find an optimal pair of \((q^*, l^*)\) that gives the maximum rate of \( 1/(q^*l^*) \) by extrinsic information transfer (EXIT) chart. Since the rate must be less the capacity, it holds that \( 1/(q^*l^*) \leq 0.5 \log_2(1 + p/(\zeta + \sigma^2)) \). Hereafter, for simplicity’s sake, we still use \((q, l)\) to represent \((q^*, l^*)\).

Now, let us turn back to our multi-rate multi-power NOMA systems. The SIC receiver decodes the \( K \) users’ information in the order of increasing user index by assumption. After recovering the information bits of the first \( k - 1 \) users, the receiver subtracts these \( k - 1 \) users’ signals from the received signal, and decodes the \( k \)th user’s information by regarding the remaining \( K - k \) users’ signals as interference with power \( \zeta_k = \sum_{k'=k+1}^{K} p_{k'} \).

For the joint RPO, let us first consider the \( K \) real powers. Multi-user information theory tells us that arbitrary power allocation can provide the sum rate to approach the capacity of \( K \)-user Gaussian multiple access channel [4]

\[ \sum_{k=1}^{K} 0.5 \log_2 \left( 1 + \frac{p_k}{\sum_{k'=k+1}^{K} p_{k'} + \sigma^2} \right) = 0.5 \log_2 \left( 1 + \frac{\sum_{k=1}^{K} p_k}{\sigma^2} \right), \]  

(2)

if \( K \) random codes are used. This implies a reasonable assumption of a equal-ratio power allocation, i.e.,

\[ p_k/p_{k+1} = \mu. \]  

(3)

Since \( P_{\text{sum}} = \sum_{k=1}^{K} p_k \), we have

\[ p_k = \begin{cases} 
\frac{\mu^{K-k}}{1-\mu^K} P_{\text{sum}}, & \mu \neq 1 \\
\frac{P_{\text{sum}}}{K}, & \mu = 1 
\end{cases} \]  

(4a, 4b)

with \( k = 1, \cdots, K \).

For the \( k \)th user with interference power \( \zeta_k \), given power \( p_k \) in (4), the EXIT chart analysis (see above) gives the optimal \((q_k, l_k)\) that provides the maximum rate \( 1/(q_k l_k) \) in \( \text{SINR}_k = p_k/(\zeta_k + \sigma^2) \). Since this optimization is given at the power ratio of \( \mu \), we denote this optimal rate and power profile by \( \Omega^\mu \triangleq ((q_1^\mu, l_1^\mu), \cdots, (q_K^\mu, l_K^\mu, p_K^\mu)) \), that provides the maximum sum rate \( R_{\text{max}}^\mu = \sum_{k=1}^{K} 1/(q_k^\mu l_k^\mu) \).

The assumption of equal-ratio power allocation above is based on the random codes, which have arbitrary rates and thus are flexible to their corresponding real
powers. In our work, the $k$th user’s rate $1/(q_k l_k)$ is discrete with the restriction of integers $q_k$ and $l_k$. Therefore, the simple equal-ratio power allocation may results in a sum-rate loss. For this reason, we set $\mu \in [\mu_{\text{min}}, \mu_{\text{max}}]$. The value of $\mu_{\text{min}}(\mu_{\text{max}})$ is determined by a minimum transmission rate requirement for each user. Over this range, we obtain the optimal rate and power profile $\Omega^{\mu^{*}}$ as

$$\Omega^{\mu^{*}} = ((q_1^{\mu^{*}}, l_1^{\mu^{*}}, P_1^{\mu^{*}}), \cdots, (q_K^{\mu^{*}}, l_K^{\mu^{*}}, P_K^{\mu^{*}})) = \arg \max_{\mu \in [\mu_{\text{min}}, \mu_{\text{max}}]} \sum_{k=1}^{K} \frac{1}{q_k^{\mu^{*}} P_k^{\mu^{*}}}$$

which gives the optimal sum rate $R_{\text{sum}}^{\mu^{*}} = \sum_{k=1}^{K} 1/(q_k^{\mu^{*}} l_k^{\mu^{*}})$ with error free decoding.

In our RPO, there are two special cases. When $\mu = 1$, users have the same power, and only rate profile is optimised [5]. In the case of each user’s rate being the same, information theory shows that its optimized power profile satisfies (3) with $\mu = \sqrt{P_{\text{sum}}/\sigma^2 + 1}$ [4].

### 4 Numerical results

In this section, we present some numerical results by the above joint RPO.

Let’s set $K = 10$, $P_{\text{sum}} = 10$, and $\text{SNR} = P_{\text{sum}}/\sigma^2 = 2$ dB. Our RPO provides sum rate $R_{\text{sum}}^{\mu}$ for $\mu \in [0.74, 1.55]$ (see Fig. 2(a)). The gap between the threshold and Shannon limit is also shown in Fig. 2(a). Among $R_{\text{sum}}^{\mu}$, $R_{\text{sum}}^{\mu^{*}} = 0.50$ is the maximum at $\mu^{*} = 0.76$ and thus is the optimal sum rate with the minimum gap of 1.98 dB. The corresponding optimal rate and power profile $\Omega^{\mu^{*}}$ and its threshold are illustrated in Table I. In $\Omega^{\mu^{*}}$, the first user’s transmission rate of 0.00625 is very low, which is implemented by the rate-1/5 RA code and the length-32 spreading.

Also in Fig. 2(a) and Table I, RO [5] ($\mu = 1.00$, equal power) gives sum rate $R_{\text{sum}}^{\mu} = 0.45$, the gap of 2.48 dB, and the corresponding optimal profile. On the other hand, PO [4] ($\mu = 1.10$, equal rate) also gives these optimal values. Compared with RO and PO, our RPO improves the maximum sum rate by 0.05 and 0.08, and decreases the gap by 0.50 and 0.39 dB, respectively at SNR = 2 dB. This improvement is also available for other SNRs. We omit the discussion at the SNRs due to limited space. The reason why RPO is superior to RO and PO is that RPO employs multiple equal-ratio power $\mu$, and choose the optimal ratio $\mu^{*}$, which gives the power profile well matched with the rate profile, and compensates the sum-rate loss due to the restriction of discrete rates of RA code and spreading.

### Table I. Optimal rate and power profiles (SNR = 2 dB)

| Sch. | $\Omega^{\mu^{*}} = ((q_1^{\mu^{*}}, l_1^{\mu^{*}}, P_1^{\mu^{*}}), \cdots, (q_K^{\mu^{*}}, l_K^{\mu^{*}}, P_K^{\mu^{*}}))$ | $R_{\text{sum}}^{\mu^{*}}$ | SNR_{th} | S. L. | Gap |
|------|-------------------------------------------------------------------------------------------------|---------------------|--------|------|-----|
| RPO  | $(5, 32, 0.21), (5, 23, 0.29), (5, 17, 0.38), (5, 13, 0.49), (5, 9, 0.65), (5, 7, 0.86)$       | $0.50$              | $2.00$  | $0.02$| $1.98$|
|      | $(4, 6, 1.13), (5, 3, 1.48), (5, 2, 1.95), (5, 1, 2.56)$                                      |                     |        |      |     |
| RO   | $(5, 7, 1.00), (4, 8, 1.00), (5, 6, 1.00), (4, 7, 1.00), (5, 5, 1.00), (6, 4, 1.00)$       | $0.45$              | $1.91$  | $-0.57$| $2.48$|
|      | $(5, 4, 1.00), (6, 3, 1.00), (5, 3, 1.00), (7, 2, 1.00)$                                      |                     |        |      |     |
| PO   | $(4, 6, 1.48), (4, 6, 1.34), (4, 6, 1.22), (4, 6, 1.11), (4, 6, 1.01), (4, 6, 0.92)$       | $0.42$              | $1.30$  | $-1.07$| $2.37$|
|      | $(4, 6, 0.84), (4, 6, 0.76), (4, 6, 0.69), (4, 6, 0.63)$                                      |                     |        |      |     |
It is interesting to give a comparison of sum rates between the coding-spreading and coding-only in Fig. 2(a). In the coding-only scheme, each user only employ a RA code \((l_k = 1)\), where \(q_k\) and \(p_k\) are jointly optimized. Obviously, our coding-spreading scheme outperforms the coding-only scheme. The reason is that in the lower SINR of \(p = (\zeta + \sigma^2)^{-1}\), the (single-user) coding-spreading works better than coding-only (see Fig. 2(b)). Note that, in the coding-only scheme, employing conventional well-designed low-rate LDPC or Turbo codes, may also approach the Shannon limit, but is impractical since users have to employ difference encoders.

**5 Conclusion**

In multi-rate multi-power NOMA systems with SIC, we employed an RA code serially concatenated with a spreading for each user to implement a variable, low-rate coding. We proposed a joint RPO to maximize the sum rate. Numerical results show that our proposed coding-spreading scheme with joint RPO, supporting the multi-rate transmission with the same structure of encoder, approaches the Shannon limit.
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Low computational complexity spectrum sensing based on cyclostationarity for multiple receive antennas

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Abstract: Low computational complexity spectrum sensing based on cyclostationarity for multiple receive antennas is proposed. The proposed technique does not calculate certain test statistics at all receive antennas, as opposed to conventional techniques that perform calculations at all receive antennas. Therefore, a low computational complexity can be achieved for sensing. Numerical examples verify that the proposed technique can obtain favorable sensing performance, even with the low computational complexity.

Keywords: cognitive radio network, cyclostationarity detection based spectrum sensing, space diversity

Classification: Terrestrial Wireless Communication/Broadcasting Technologies

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1 Introduction

Cognitive radio is the core technology for the efficient use of wide-frequency bands. In cognitive radio networks, secondary users (SUs) need to sense and monitor the radio spectrum around themselves for detecting and sharing the frequency bands that are not occupied by primary users (PUs). Because SUs must not inhibit the PUs’ communication, spectrum sensing techniques for PUs’ protection are very important in realizing cognitive radio networks [1]. It is well established in several spectrum sensing types that cyclostationarity-based spectrum sensing [2, 3] is robust against interference, although its computational complexity is not low. In order to improve performance, various techniques have been proposed for cyclostationarity-based spectrum sensing using multiple receive antennas (MRAs) [4, 5]. In conventional approaches to cyclostationarity-based spectrum sensing with MRAs, the computational complexity is proportional to the number of antennas, as statistics are computed at each antenna. This letter proposes computationally efficient spectrum sensing by means of test statistics sharing among MRAs.

2 Cyclostationarity based spectrum sensing

We consider that a PU transmits an orthogonal frequency division multiplexing (OFDM) signal, which is composed of the useful symbol duration $T_{FFT}$ and cyclic prefix (CP) duration $T_{CP}$, as well as the OFDM symbol duration $T_{OFDM} = T_{FFT} + T_{CP}$, and the SU detects the PU signal by means of spectrum sensing. Spectrum sensing of the OFDM signal is carried out in order to detect the signal cyclostationarity. This is a binary hypothesis testing problem, as follows:

$$
\mathcal{H}_0 : r_i(n) = v_i(n) \\
\mathcal{H}_1 : r_i(n) = h_i s(n) + v_i(n), \quad i = 1, \ldots, N_R,
$$

where $r_i(n)$, $v_i(n)$, $h_i$, $s(n)$, and $N_R$ are the received signal, additive white Gaussian noise (AWGN) following $\mathcal{CN}(0, \sigma_i^2)$, channel gain at the $i$th receive antenna, PU signal, and number of receive antennas, respectively. Moreover, the hypotheses $\mathcal{H}_0$ and $\mathcal{H}_1$ indicate that the PU is inactive and active, respectively. In cyclostationarity-based spectrum sensing, a cyclic autocorrelation function (CAF) is employed. Let $\hat{R}_{r_i}^{\text{ca},N}(n, \tau)$ denote an $N$ sample-computed approximated CAF for $r_i(n)$, lag parameter $\tau$, and cyclic frequency $\alpha$. Then, it can be written as

$$
\hat{R}_{r_i}^{\text{ca},N}(m, \tau) = \frac{1}{N} \sum_{n=m}^{m+N-1} r_i(n) r_i^*(n + \tau) e^{-j2\pi\alpha n \Delta t}, \quad i = 1, \ldots, N_R,
$$
where $\Delta t$ is the sampling period. The OFDM signal exhibits a cyclostationarity because of its own CP [2]. Let $a_k = k/T_{\text{OFDM}}$ ($k \in \mathbb{Z}$) and $\beta_k = (k + 0.5)/T_{\text{OFDM}}$, and the CAF of the OFDM signal has peaks at $a_k$ and $\tau = N_{\text{FFT}} := T_{\text{FFT}}/\Delta t$; moreover, the CAF of the OFDM signal has no peaks at $\beta_k$ [2]. Maximum cyclic autocorrelation selection (MCAS) [3] originally has low computational complexity, because the CAF must be computed at only a limited number of cyclic frequencies. In MCAS, the CAF values at certain cyclic frequencies, namely $a_1$ (peak) and $\beta_k$ ($k = 0, 1, \ldots, N_D - 1$) (non-peak), are compared in order to detect the PU signal, where $N_D$ is the number of non-peak CAFs to be used for sensing. In the single antenna case, MCAS carries out signal detection using peak and non-peak CAFs, as follows:

$$\left| \hat{R}_{r_1}^{a_1,N}(m, N_{\text{FFT}}) \right|_{\mathcal{H}_1} > \max_{N'} \sqrt{N'/N} \left| \hat{R}_{r_1}^{\beta_k,N'}(m + kN', N_{\text{FFT}}) \right|, \quad k = 0, 1, \ldots, N_D - 1, \quad (3)$$

where $N'$ is the number of samples for $\hat{R}_{r_1}^{\beta_k,N'}(m + kN', N_{\text{FFT}})$ computation. Furthermore, $N'$ is any integer satisfying $N'/N_D \leq N$; however, the MCAS signal detection performance is degraded when $N'$ is extremely small [6]. It should be noted that $\hat{R}_{r_1}^{a_1,N}(m, N_{\text{FFT}})$ and $\sqrt{N'/N} \hat{R}_{r_1}^{\beta_k,N'}(m + kN', N_{\text{FFT}})$ follow $\mathcal{CN}(0, \sigma_0^2/N)$ in $\mathcal{H}_0$ [3]. Because both of the probability density functions agree, MCAS can control the false alarm probability as $P_{\text{FA}} = 1/(N_D + 1)$.

### 3 Test statistics sharing among multiple antennas based on MCAS

As noted above, the non-peak CAF $\sqrt{N'/N} \hat{R}_{r_i}^{\beta_k,N'}(m + kN', N_{\text{FFT}}) \sim \mathcal{CN}(0, \sigma_c^2/N)$ in $\mathcal{H}_0$; furthermore, the non-peak $\sqrt{N'/N} \hat{R}_{r_i}^{\beta_k,N'}(m + kN', N_{\text{FFT}})$ is a zero mean complex Gaussian in $\mathcal{H}_1$. Therefore, it can be stated that the non-peak CAF $\sqrt{N'/N} \hat{R}_{r_i}^{\beta_k,N'}(m + kN', N_{\text{FFT}})$ can only take a random variable in, whether $\mathcal{H}_0$ or $\mathcal{H}_1$. These facts indicate that a comparison of the peak CAF and non-peak CAFs corresponds to a comparison of the peak CAF and certain random variables where the statistical property is known in MCAS. Based on the above consideration, the proposed technique computes $N_R$ number of CAFs at $\beta_k$ from among the $N_R \times N_D$ that are computable, as illustrated in Fig. 1(a), and the decision variables at $a_1$ and $\beta_k$ are defined as

$$Z_{a_1} = \left| \frac{1}{N_R} \sum_{i=1}^{N_R} \hat{R}_{r_i}^{a_1,N}(m, N_{\text{FFT}}) \right|, \quad (4)$$

$$Z_{\beta_k} = \left| \hat{R}_{r_i}^{\beta_k,N'}(m + kN', N_{\text{FFT}}) \right|, \quad i = 1, \ldots, N_R, \quad k = 0, 1, \ldots, N_D - 1. \quad (5)$$

Again, $N_D$ of $N_R \times N_D$ $Z_{\beta_k}$s are used for signal detection. The subscripts $(i, k)$, which are in $Z_{\beta_k}$, for the receive antenna and $\beta_k$ cyclic frequencies, are chosen without deviation for the receive antenna, e.g., $(1, 1), \ldots, (N_R, N_R), (1, N_R + 1), \ldots, (N_R, 2N_R), (1, 2N_R + 1), \ldots$ in the $N_R < N_D$ case. The decision criterion of the proposed technique is expressed by

$$Z_{a_1} \geq \max_{\mathcal{H}_0} \left| \frac{N'}{N_R} Z_{\beta_k} \right|, \quad k = 0, 1, \ldots, N_D - 1. \quad (6)$$
4 Computational complexity

The computational complexity required for the complex operation multiplication in the proposed technique is compared with that of the conventional techniques [4, 5]. We let $M_{\text{Proposed}}$, $M_{\text{Cho}}$, and $M_{\text{Huang}}$ denote the computational complexity of the proposed and conventional techniques [4, 5], respectively. These can be written as

\begin{align*}
M_{\text{Proposed}} &= 8NN_R + 4N'N_D + 2(N_D + 1), \quad (7) \\
M_{\text{Cho}} &= 2N_R \{4N + 2N_D(N' + 1) + 1\}, \quad (8) \\
M_{\text{Huang}} &= 8NN_R^2 + 2N_R(N + 2). \quad (9)
\end{align*}

Fig. 2 illustrates $M_{\text{Proposed}}$, $M_{\text{Cho}}$, and $M_{\text{Huang}}$ for different numbers of $N_R$. We set $N_R = 2\ldots8$, $N = 2560$, $N' = 134$, and $N_D = 9$. It can be seen that the computational complexity of the proposed technique is lower than that of [4, 5]; therefore, computational complexity can be reduced by the CAFs at $\beta_k$ sharing. Furthermore, it can be seen that the difference in the computational complexity increases as $N_R$ increases.

5 Numerical examples

For evaluation of the proposed technique, we set $T_{\text{FFT}} = 64\Delta t$, $T_{\text{CP}} = 16\Delta t$, $N = 2560$, $N' = 134$, $N_D = 9$, and a Rayleigh flat fading channel model is employed. Moreover, the proposed technique is also evaluated in the case of an automatic gain control (AGC). The proposed technique must compute only $N_D$ number of non-peak CAFs, under the assumption that the noise variances at all RF chains are the same. However, the AGC causes the AWGN variance itself with OFDM signals to change, and this results in the difference in noise variances at each RF chain, because the received signal amplitude at each receive antenna determines the AGC gain level at each receive antenna [7]. As a result, the variances of the non-peak CAFs may be also varied. Therefore, it is conceivable that non-peak CAF sharing...
affects signal detection performance. Thus, we evaluate the two cases: 1) without AGC and 2) with AGC in this section. In order to simulate the AGC behavior, each received signal is normalized by its maximum value for $N$ samples, and each signal is multiplied by a constant gain factor $G$. We assume that the RF received signal power is approximately $-80$ dBm, the characteristic impedance of the RF chains, baseband circuits, and so on is $50 \Omega$, and that the maximum amplitude of the baseband signal after passing through the AGC is $3$ V. From these, we employ $G = 10^5$.

Firstly, Fig. 3(a) illustrates the detection performance of the proposed and conventional techniques [4, 5] in the without AGC case. The performance of the proposed technique is slightly degraded in comparison with that of the conventional techniques, even though its computational complexity is lower (former: 84 and 70%; latter: 56 and 13% for $N_R = 2$ and 8). Next, Fig. 3(b) illustrates the detection performance of the proposed technique in the with AGC case. It can be seen that the performance degradation caused by the AGC is negligible. Moreover, it can be observed that the proposed technique can obtain almost the same performance as the conventional techniques [4, 5]. Finally, Fig. 3(c) illustrates the $P_{FA}$ performances of the proposed and conventional techniques. The proposed technique achieves the target $P_{FA}$ with an error of less than 1%.
Computationally efficient cyclostationarity-based spectrum sensing was considered. The proposed technique was able to reduce computational complexity by means of non-peak CAFs sharing among the MRAs. The results demonstrate that the sensing performance of the proposed technique is favorable, and its computational complexity is superior to that of the conventional technique [4], in proportion to the number of receive antennas, even though the received signal passed through the AGC. Future work will include the development of a weighted version of the proposed technique for improving detection performance.

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