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一种用于光线性采样的载波相位估计算法

李飞涛[1,2],赵卫[1],刘元山[1]
(1 中国科学院西安光学精密机械研究所 瞬态光学与光子技术实验室,西安 710119)
(2 中国科学院大学,北京 100039)

摘 要:提出了一种改进的相位盲搜索算法,该算法通过两次寻找二次最小欧拉距离得到载波相位噪声的估计值.为验证该算法,搭建了一套光线性采样仿真系统,对符号率为10 Gbaud/s的16QAM光调制信号的星座图进行了测量,采样源为脉宽为500 fs、周期为10 ns的光频梳.仿真结果表明,当两个激光器线宽和小于10MHz时,经过该算法对相位噪声进行补偿后,10G-Gb/s的16QAM信号的星座图能够清晰展示.

关键词:相干光通信;线性采样;正交幅度调制;数字信号处理;载波相位噪声
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A Phase Estimation Algorithm for Optical Linear Sampling

LI Fei-tao[1,2], ZHAO Wei[1], LIU Yuan-shan[1]
(1 *State Key Laboratory of Transient Optics and Photonics*, *Xi'an Institute of Optics and Precision Mechanics of CAS*, *Xi'an* 710119, *China*)
(2 *University of Chinese Academy of Sciences*, *Beijing* 100039, *China*)

Abstract: An improved Blind Phase Search algorithm was proposed. In this algorithm, carrier phase noise was estimated by searching minimum Euler distance twice. To verify this algorithm, an optical linear sampling simulation system was set up. The constellation diagram of 10Gbaud/s 16QAM signal was simulated by using an optical frequency comb with pulse width of 500fs and duration of 10ns as optical sampling source. The simulation result shows that, the phase noise between the signal source and optical sampling source can be compensated effectively by this algorithm when the sum linewidth of the two lasers is less than 10MHz. Finally, the constellation diagram of 10Gbaud/s 16QAM signal is displayed clearly.

Key words: Optical coherent communication; Optical linear sampling; Quadrature amplitude modulation; Digital signal processing; Carrier phase noise
OCIS Codes: 060.0060; 060.1660;060.4080; 060.2300;060.7140

0 Introduction

The transmission capacity of multiple terabits per second could be achieved by using a Wavelength Division Multiplexing (WDM) with tens and hundreds of channels in combination with the spectrally efficient advanced modulation format[1-2]. The spectral efficiency of the Quadrature Amplitude Modulation (QAM) formats has N times than that of the Binary Phase Shift Keying (BPSK). Therefore, they are regarded as the attractive candidates for meeting such requirements[3-5]. In order to realize the large capacity and high spectral

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First author: LI Fei-tao (1983－), male, Ph. D. degree candidate, mainly focuses on linear optical sampling technology. Email: lihua_5188@163.com
Supervisor(Corresponding author): LIU Yuan-shan (1975－), male, associate professor, Ph. D. degree, mainly focuses on ultra-short optical pulse generation, radio-over-fiber transmission system, high－stability mode-locked fiber lasers, optical fiber frequency comb, ultra-fast all-optical sampling techniques, and ultra-wide bandwidth optical measurement instruments. Email: yuanshanliu@yahoo.com.
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efficiency optical communication systems in the near future, in addition to have significant requirement for the transmitters and receivers, such systems also require sophisticated and high bandwidth characterization tools which can be able to detect the complex field of the optical transmission signal. Especially with the bit rates increasing rapidly, characterizing optical fields in telecommunications has become more and more important. Optical sampling oscilloscope with a bandwidth on the order of 1THz has been achieved using the nonlinear optical sampling techniques. However, this approach attains a poor sensitivity and it is also not able to detect the phase of the optical waveform[6-7]. However, instead of gating the temporal intensity information of the data signal source, coherent linear optical sampling gates the electric field of the data signal using its interference with the electric field of sampling pulses. Since coherent optical linear sampling is a linear process, therefore, it not only can provide a measurement sensitivity (3×10^{3} m · W^{2}) nearly three orders of magnitude better than that of the nonlinear optical technique (10^{6} m · W^{2}), but also obtain a high temporal resolution by utilizing ultra-short optical pulses in coherent linear optical sampling[8-10]. Furthermore, coherent linear optical sampling has the ability to offer a timely diagnostic technology for studying advanced modulation formats such as differential (quadrature) phase shift keying (DPSK/DQPSK)[11-12]. Recently, the generation of Square-16-QAM signal has been extensively investigated in optical transmission systems since 16-QAM can offer a high spectral efficiency. In general, its constellation diagram monitoring is completed by using the coherent detection combine with a high sampling rate Analog to Digital Conversion (ADC)[13-15]. Unfortunately, the cost of the high sampling rate ADCs with a wide band width front end is costly. Furthermore, the standard could not meet the application requirements, for example, for a multilevel modulation signal > 50 Gbaud. Although Wen He *et al*[16] have experimentally achieved the constellation diagram measurement of a square-16-QAM signal by using a linear optical sampling technique, whose signal under test and sampling pulses source in experiment are implemented by using the same Continuous Wave laser. It is obvious that this approach is lack of generality. Linear optical sampling can be used for measuring the constellation diagram of QAM signals, however, some new challenges are still encountered in this scheme in order to achieve an accurate constellation diagram measurement of QAM signal. In addition to require a low amplitude and phase noise, wavelength stabilized signal under test and optical sampling pulse laser, especially for higher order QAM constellations, a lower spacing among adjacent constellation points lead to a smaller phase noise and amplitude fluctuation tolerance for optical sampling pulse laser. These problems can be alleviated due to new availability of frequency-stable and phase-stable optical frequency comb[17-18] and a high-speed digital signal processing.

In this paper, the constellation diagram measurement process of Square-16-QAM signal was analyzed by using the coherent linear sampling and point out the actual mechanism for phase noise which has an influence on constellation diagram measurement of square-16-QAM signal. In addition, we propose a digital signal processing algorithm for the phase compensation based on two minimum distances. Finally, the compensation effectiveness of this algorithm is verified by numerical simulations.

1 Principle

In general, the optical Square-16-QAM signal is achieved by modulating the optical carrier (emitted from narrow linewidth Continuous Wave (CW) laser) in an optical In-phase Quadtraure-modulator. The electrical driving signal for Square-16-QAM signal is defined as

$$u_{\mathrm{I}}(t)=-V_{D\pi}+\frac{2V_{\pi}}{\pi}\sum_{k}\left[\arcsin(i_k)\cdot p(t-kT_{\mathrm{S}})\right]\qquad i_k\in\left(-1,-\frac{1}{3},\frac{1}{3},1\right)\tag{1}$$

$$u_{\mathrm{Q}}(t)=-V_{D\pi}+\frac{2V_{\pi}}{\pi}\sum_{k}\left[\arcsin(q_k)\cdot p(t-kT_{\mathrm{S}})\right]\qquad q_k\in\left(-1,-\frac{1}{3},\frac{1}{3},1\right)\tag{2}$$

where i_k and q_k represent the normalized symbol coordinates. For the square-16-QAM signal, it holds $i_k\in\{-1, -1/3, 1/3, 1\}$ and $q_k\in\{-1, -1/3, 1/3, 1\}$. $V_{D\pi}$ and V_{π} describes the Mach-Zehnder Modulator and the peak-to-peak modulation for the phase of π respectively. $p(t)$ is the pulse shape of the electrical signals. We assume that the electrical driving signals for the Square-16-QAM transmitter are ideal in order to simplify the derivation. When ignoring any insertion loss, the field of the In-phase Quadtraure-modulator output can be shown as

$$\frac{E_{\mathrm{out}}(t)}{E_{\mathrm{in}}(t)}=\frac{1}{2}\left\{\left[\cos(\pi\frac{u_{\mathrm{I}}(t)}{2V_{\pi}})\right]+\mathrm{j}\left[\cos(\pi\frac{u_{\mathrm{Q}}(t)}{2V_{\pi}})\right]\right\}\tag{3}$$

When neglecting the polarization issues, which can be treated as polarization diversity, the normalized electrical field of the incoming optical advanced modulation formats and the sampling pulse laser can be written in a complex notation (neglecting transmission impairments) as

$$E_D(t)=a(t)\cdot e^{j\varphi(t)}\cdot\sqrt{P_D}\cdot e^{j(-\omega_D t+\varphi_D+\varphi_{ND})}=\varepsilon_D(t)\cdot\sqrt{P_D}\cdot e^{j(-\omega_D t+\varphi_D+\varphi_{ND})}=(I(t)+jQ(t))\cdot\sqrt{P_D}\cdot e^{j(-\omega_D t+\varphi_D+\varphi_{ND})}=E_I(t)+jE_Q(t) \tag{4}$$

$$E_S(t)=\sum_N \varepsilon_S(t-NT_s)\exp\cdot[j(-\omega_{SC}(t-NT_s)+N\Delta\varphi_{ce}+\varphi_{S0}+\varphi_{NS})] \tag{5}$$

In Eqs. (4) and (5), P_D represents the CW power, ω_D is the carrier frequency of the signal laser and when the optical frequency comb pulse train can be decomposed into a slow varying envelope around an optical carrier of center frequency, $\omega_{SC}=2\pi m_0/T_s$ will be the center frequency. φ_{D0} and φ_{S0} are the initial phases, and φ_{ND} and φ_{NS} is the phase noise of signal under test and sampling pulse laser, respectively. In Eq. (4), the signal $\varepsilon_D(t)$ represents all the modulations performed on the CW laser, $a(t)$ and $\varphi(t)$ are the amplitude and phase, $I(t)$ and $Q(t)$ is the in-phase and quadrature components of optical advanced modulation formats signal respectively. $E_I(t)$ and $E_Q(t)$ is the in-phase and quadrature normalized electrical field components of optical advanced modulation formats signal respectively. In Eq. (5), ε_S represents the envelope function of each sampling pulse, T_S describes the repetition rate of the sampling laser, and $N\Delta\varphi_{ce}+\varphi_S$ is the relative phase between the analytic signal and the carrier for pulse N. In the frequency domain, the components of the optical frequency comb are divided by f_{rep}. The position of the entire optical frequency comb is offset integer multiples of f_{rep} by an offset frequency[18] $\delta=\Delta\varphi_{ce}f_{rep}/2\pi$, i. e. $\nu_m=mf_{rep}-\delta$, which is relative to the pulse-to-pulse phase shift. Without an active controlling, δ is a dynamic quantity, which is very sensitive to perturbation of the laser. So the relative phase ($\Delta\varphi_{ce}$) from pulse-to-pulse in an un-stabilized lase changes in a non-deterministic manner.

As depicted in Fig. 1(a), the data under test and sampling pulses laser have the same polarization, the signal wave and a train of sampling pulses are combined in an optical 2×4 90°-hybrid and yield the output fields

$$\begin{bmatrix}E_1\\E_2\\E_3\\E_4\end{bmatrix}=\frac{1}{2}\begin{bmatrix}1&1\\1&j\\1&-1\\1&-j\end{bmatrix}\cdot\begin{bmatrix}E_D(t)\\E_S(t)\end{bmatrix}=\begin{bmatrix}E_D(t)/2+E_S(t)/2\\E_D(t)/2+jE_S(t)/2\\E_D(t)/2-E_S(t)/2\\E_D(t)/2-jE_S(t)/2\end{bmatrix} \tag{6}$$

The upper balanced detector is used for detecting E_1 and E_3, and the lower Balanced Detector (BD) is exploited for detecting E_2 and E_4, the photocurrents are obtained

$$I^*(t)=R\cdot\left\{\mathrm{real}\left[\int_{-\infty}^{+\infty}E_D(t)\cdot E_S^*(t)\,dt\right]\right\}+n_1(t)-n_3(t) \tag{7}$$

$$Q^*(t)=R\cdot\left\{\mathrm{imag}\left[\int_{-\infty}^{+\infty}E_D(t)\cdot E_S^*(t)\,dt\right]\right\}+n_2(t)-n_4(t) \tag{8}$$

In Eqs. (7) and (8), $n_i(t)$, $i=1,2,3,4$, represent the shot noise of four photodiodes, R is the responsivity. Let's assume that the relative phase between the two fields is zero, which mainly depending on the optical path differences between the splitter and combiners. The two measured signals can be combined to describe

$$S^*(t)=R\cdot\left[\int_{-\infty}^{+\infty}E_D(t)\cdot E_S^*(t)\,dt\right]+n_k^{tot} \tag{9}$$

In Eq. (9), $n_k^{tot}=(n_{1,k}-n_{3,k})+i(n_{2,k}-n_{4,k})$ represents the complex shot noise. Usually, the bandwidth of the photodetectors not only require much higher than the repetition rate of the sampling source, but also require much smaller than the bandwidth of the two laser sources, therefore, the output of the balanced detector can include many of well separated sampling events synchronized with pulses from the sampling source. Furthermore, the pulse-width of sampling pulses must be much shorter than the integration time of the photodetectors, so that the signal of Eq. (7) and (8) can be written as an integral from $-\infty$ to $+\infty$. In the following equations such integration will be described as $\int$ for the sake of clarity. The measured sample by the Nth sampling pulse can be described as

$$S^*(N)=R\cdot\sqrt{P_D}\cdot\exp[i((\omega_{SC}-\omega_D)t-N(\omega_{SC}T_S+\Delta\varphi_{ce})+\varphi_{D0}-\varphi_{S0}+\varphi_{ND}-\varphi_{NS})]\cdot\int\varepsilon_D(t)\cdot\varepsilon_S^*(t-NT_S)\,dt+n_k^{tot} \tag{10}$$

Now we suppose that the carrier frequency of the data signal can be included in the spectrum of the sampling optical comb laser at an arbitrary position and near one of the modes, as is showing in Fig. 1(c), ω_{sc} is the center frequency of the optical frequency comb, ω_D is the optical carrier frequency of signal under test, carrier frequency detuning, $\Delta\omega$ is the detuning of the optical carrier between signal and the center frequency of optical frequency comb laser. Therefore $\omega_D=\omega_{s,n}+\Delta\psi$, where $\Delta\psi$ is the carrier frequency offset. Then Eq. (10) becomes

$$S^*(N)=R\cdot\sqrt{P_D}\cdot\exp[i((\omega_{sc}-\omega_{s,n})t-\Delta\Psi t-N(\omega_{SC}T_S+\Delta\varphi_{ce})+\varphi_{D0}-\varphi_{S0}+\varphi_{ND}-\varphi_{NS})]\cdot\int\varepsilon_D(t)\cdot\varepsilon_S^*(t-NT_S)\,dt+n_k^{tot}=R\cdot\sqrt{P_D}\cdot\exp[i(-N(\omega_{SC}T_S+\Delta\varphi_{ce})+\varphi_{D0}-\varphi_{S0}+\varphi_{ND}-\varphi_{NS})]\cdot\int\varepsilon_D(t)\cdot\varepsilon_S^*(t-NT_s)e^{i\Delta\omega t}\,dt+n_k^{tot} \tag{11}$$

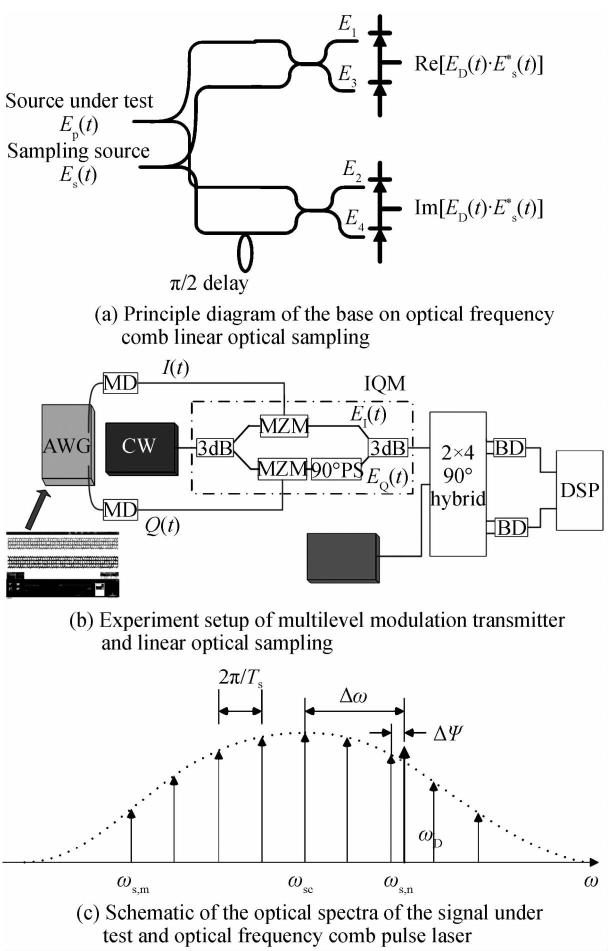
In Eq. (11), the fact that $\Delta\omega=\omega_{sc}-\omega_D$ is the carrier frequency detuning between the signal laser and the optical frequency comb are used. Let's suppose that the pulse-width of sampling pulses (Δt_S, FWHM) is much smaller than the pulse-width (Δt_D) of the data signal in order to make sure that the envelope and phase of the data pulse are almost constant during sampling. Then Eq. (11) becomes

$$S^*(N)=R\cdot\sqrt{P_D}\cdot\exp[\mathrm{i}(-N(\omega_{SC}T_s+\Delta\varphi_{ce})+\varphi_{D0}-\varphi_{S0}+\varphi_{ND}-\varphi_{NS})]\cdot\varepsilon_D(NT_s)\cdot\int\varepsilon_S^*(t)\mathrm{e}^{\mathrm{i}\Delta\omega t}\mathrm{d}t+n_k^{tot}=R\cdot\sqrt{P_D}\cdot\exp[\mathrm{i}(-N(\omega_{SC}T_s+\Delta\varphi_{ce})+\varphi_{D0}-\varphi_{S0}+\varphi_{ND}-\varphi_{NS})]\cdot\varepsilon_D(NT_s)\tilde{\varepsilon}_S^*(\Delta\omega)+n_k^{tot}\qquad(12)$$

In Eq. (12), the $\tilde{\varepsilon}_S(\omega)$ is Fourier transformation of the slowly varying envelope $\varepsilon_s(t)$. Such a diagnostic there can be used for testing samples of Square-16-QAM signal at the times NT_s.

2 Experiment setup

The schematic diagram of multi-level- modulation transmitters and optical linear sampling are shown in Fig. 1(b). The components include: Arbitrary Waveform Generator (AWG), Modulator Driving (MD), Continuous Laser (CW), Mach-Zehnder Modulator (MZM), Phase Shift (PS), Optical Frequency Comb pulse laser (OFC), Balanced Detector (BD), Digital Signal Processing (DSP). The optical part of the transmitters is common for every M - PSK and M - QAM modulation format in addition to the electrical driving signal. To realize a different modulation format, only the electrical driving signal needs to be adjusted. For Square-16-QAM, 4-ary electrical driving signals are need for the MZM in two arms of the IQ modulator. The electrical driving signals of optical Square-16-QAM are generated by using AWG (arbitrary waveform generator) and amplified with the MD. In fact, a bias controller (not shown in Fig. 1 (b)) is required to control the bias positions for MZM of IQ modulator. It simultaneously sets the first and second MZM at Null points and the third modulator (PM) at quad point. The phase and amplitude of the optical Square-16-QAM signal are measured by observing its interference with two orthogonal quadratures of the sampling optical frequency comb pulses in 90° hybrid (in Fig. 1), and its output are detected with 2 pairs of BD. The output samples of two BDs are sampled by 2 ADCs. Finally, the samples were loaded to personal computer for further digital signal processing.



(a) Principle diagram of the base on optical frequency comb linear optical sampling

(b) Experiment setup of multilevel modulation transmitter and linear optical sampling

(c) Schematic of the optical spectra of the signal under test and optical frequency comb pulse laser

Fig. 1 Optical linear sampling

3 Results and discussion

Eq. (12) is the main result of the analysis of amplitude and phase information of the electric field of a data-encoded telecommunication channel when the linear optical sampling process is used. Several important conclusions can be derived from this Equation. Compared with the original information, the measured samples are corrupted due to the phase noise, phase offset, carrier envelope phase and shot noise. First, carrier-envelope phase is a random variation in the conventional mode-locked lasers, more efforts are underway to achieve a stabilization of the carrier-envelop phase[18]. Therefore, the influence of the phase $N(\omega_{sc}T_s+\Delta\varphi_{ce})$ was removed by experimentally setting $\omega_{sc}T_s+\Delta\varphi_{ce}$ to a multiple of 2π, which has been investigated in Ref. [5]. Second, the period of the sampling pulse is chosen to be exactly equal to the proper integer multiple period of the data signal when we pay attention to testing the constellation diagram of the data signals, the measured samples depend on the optical carrier frequency detuning only and is independent of carrier frequency offset, which is different from the eye diagram measurement analysis in Ref. [7]. In fact, the repetition rate of the sampling pulse laser has a certain drift with the temperature changing of the external environment. Without active stabilization, it is difficult to let the sampling pulse's period exactly equal to proper integer multiple of the data signal's period. Therefore, an ideal sampling pulse source is repetition rate stabilization pulsed laser. Third, the shot noises of the photodetector can be corrected using a pair of excellent matching photodetector in the balanced detector. Fourth, the initial phases for both signal laser (usually, narrow-linewidth laser) and mode-locked laser are constant, their effect on optical linear sampling can be compensated by software. In this paper, we assume that both initial phases are zero. Optical frequency comb with frequency and phase stabilization is considered as an ideal sampling pulse source since it meets these requirements for sampling source. Finally, the phase noise on sampling pulses is related to the coherence of the sampling source. The coherence of the source under test is limited by phase noise of the signal laser during the measurement of a representative set of samples of its electric filed[20]. Both affect the phase modulation signal. The amplitude noise from the sampling source (*i. e.* fluctuations of the amplitude of the electric field of sampling pulses) will lead to identical fluctuation on the amplitude of the measured samples. These noises have been compensated by using a phase estimation in coherence optical communication systems using the recent availability of Gbit-speed digital signal processing. The Viterbi & Viterbi algorithm and Maximum Likelihood phase estimation are conventionally used when the data is embedded in White Gaussian Noise (WGN) and corrupted by a quasi-constant phase offset. The phase noise of both lasers may change dynamically among different symbols. Phase noise are usually considered as a Wiener process [21]

$$\sigma_f^2 = 2\pi(\Delta f \cdot T_s) \tag{14}$$

However, the numerical simulation was performed in order to prove the analytic results presented above. The data rate of pseudo-random 4-ary electrical driving signals sequence from AWG was set to be 10 Gbit/s and sampled by the real-time oscilloscope at the sampling rate of 40 Gsamples/s. The corresponding eye-diagram is shown in the Fig. 1. The pulse-width of optical frequency comb pulses is set to 500 fs and the repetition rate is 100 MHz, the corresponding pulse and spectral shape of the optical frequency comb is shown in Fig. 2. The pulse-width of sampling pulses (Δt_S, FWHM) is much smaller than the duration (Δt_D) of the data signal, the amplitude of the sampling pulses can be considered as constant at the sampling instant. Therefore, the in-phase and quadrature elements of the Square-16-QAM signal at the sampling instant, $I^*(NT_S)$ and $Q^*(NT_S)$, can be obtained. Under an optimal condition, however, the in-phase and quadrature samples at the sampling instant only include phase noises of both lasers. In coherent optical communication systems, this phase noise can be compensated with digital signal processing in order to accurately recover the constellation diagram of Square-16-QAM. In optical multi-level modulation coherent communication, Blind Phase Search (BPS) algorithm has been consider to be a benchmark in phase noise estimation algorithms[21], however, its hardware

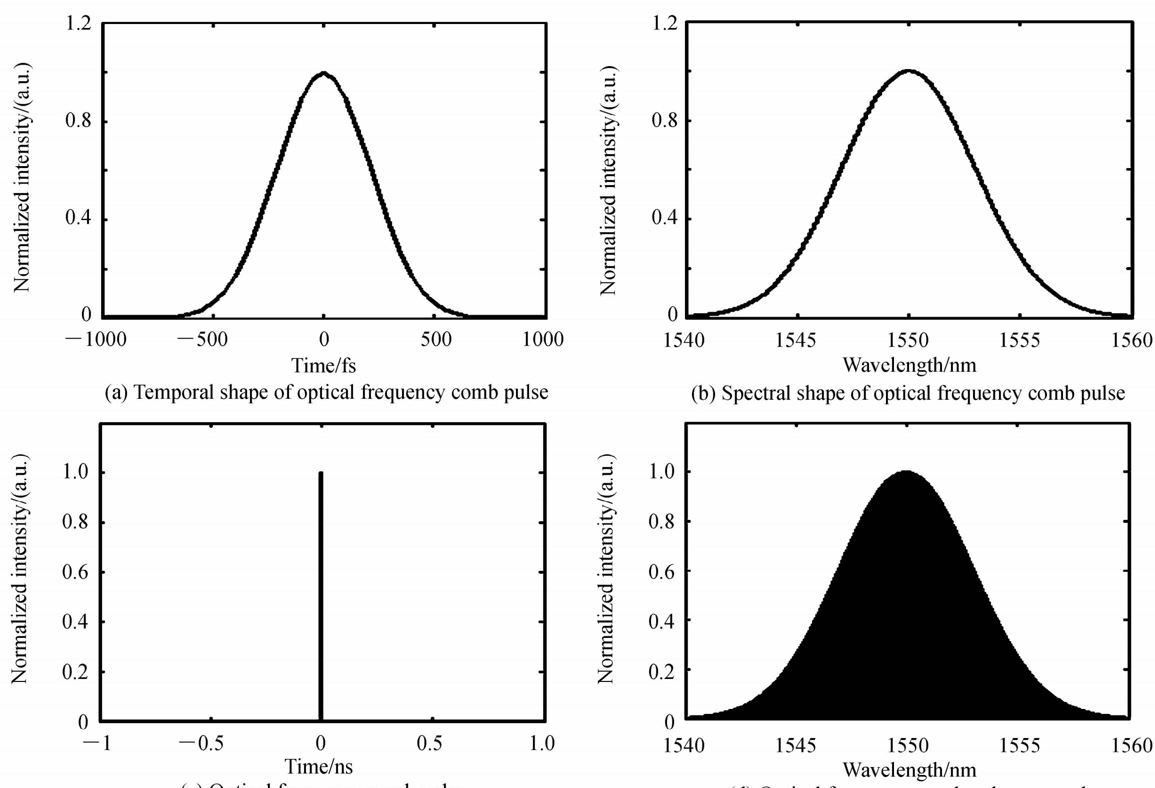


Fig. 2 Time domain and spectral domain diagram of simulation of the optical frequency comb sampling pulse

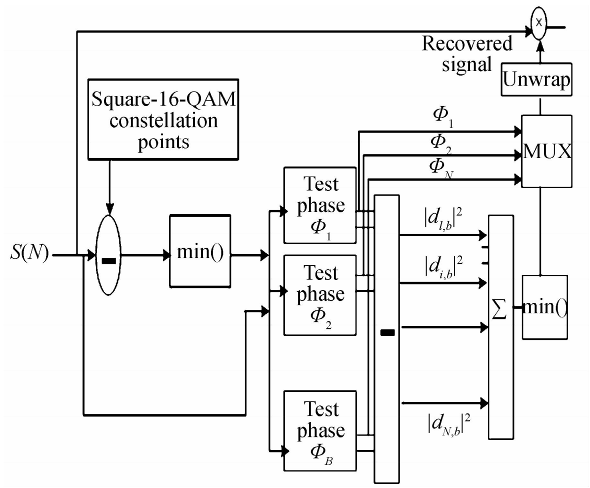


Fig. 3 Two-stage blind phase search method

implementation is very complexity. In this paper, we propose an algorithm, shown in Fig. 3, which employs a two-stage process for finding the optimal estimated phase. Firstly, we find the minimum Euclidean distance between the samples and constellation points, the constellation points denotes the output of the decision device. Secondly, the samples and constellation points are simultaneously multiplied by the test phase, calculate the square of Euclidean distance among them, put every N samples into a block and find the minimum among them and unwrap. The corresponding results is the estimated phase , which is very similar to Blind Phase Search (BPS) algorithm. The difference is the method of decision. The estimated phase is achieved with matlab software implementation to overcome the hardware implementation complexity. The flow chart of signal processing is drawn in Fig. 3. Therefore, constellation diagrams of Square-16-QAM signal are reconstructed, *i. e.* a full statistical representation of samples of the complex electric field.

In this section, some simulations results are discussed. The simulations were performed with an Arbitray Waveform Generator (AWG) and matlab to prove the proposed phase estimation algorithm. With the AWG, the Square-16-QAM signal is configured. The data rate is 10 Gbaud (corresponding to 40 Gbit/s). The sum line-width signal and Optical frequency comb are varied (100 kHz, 1 MHz and 10 MHz). The Wiener noise with zero mean, the varied sum line-width of signal and Optical frequency comb, variance of $2\pi \cdot \Delta f \cdot T_s$ is embed into the Square-16-QAM signal. The samples of signals with different Wiener noises are processed by our proposed phase estimation algorithm implemented in MATLAB (see Fig. 3). This phase estimation algorithm can also be exploited for the offline processing in the optical linear sampling experiment. The main results of the simulation are shown in Fig. 4, 5, 6, respectively.

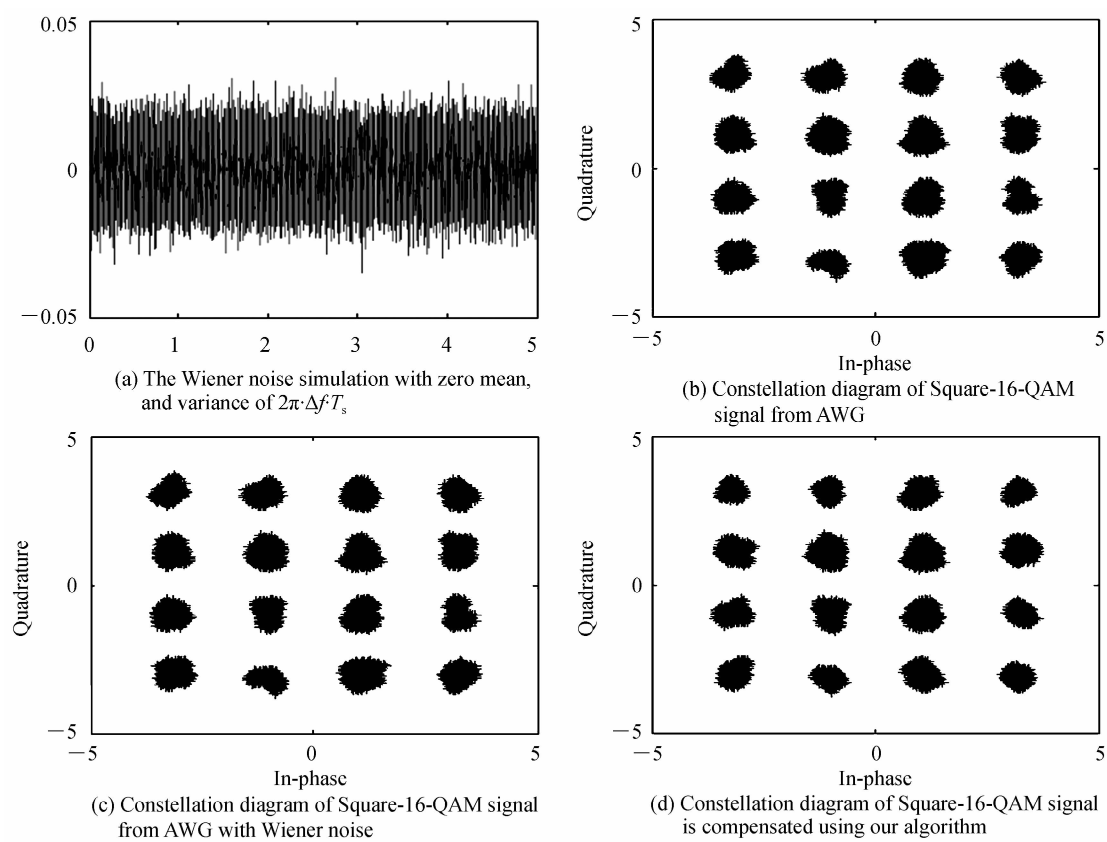


Fig. 4 The sum linewidth of signal and optical frequency comb is 100 KHz

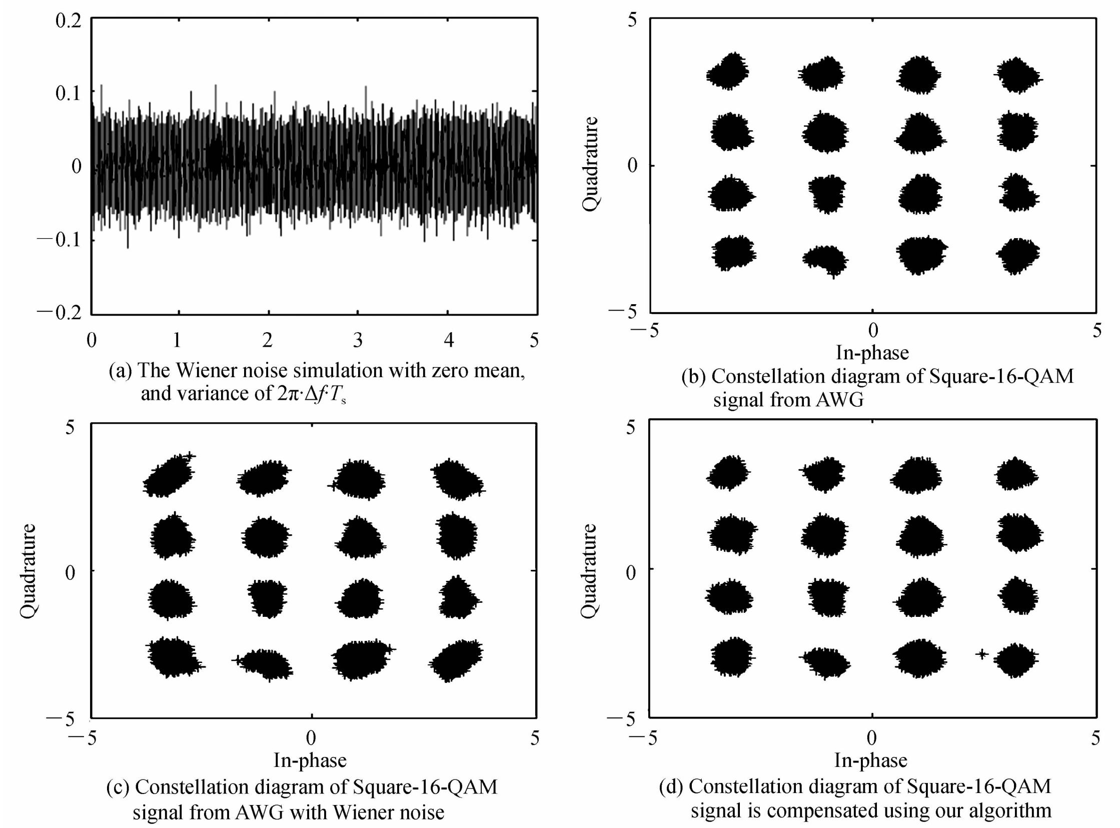


Fig. 5 The sum linewidth of signal and optical frequency comb is 1 MHz

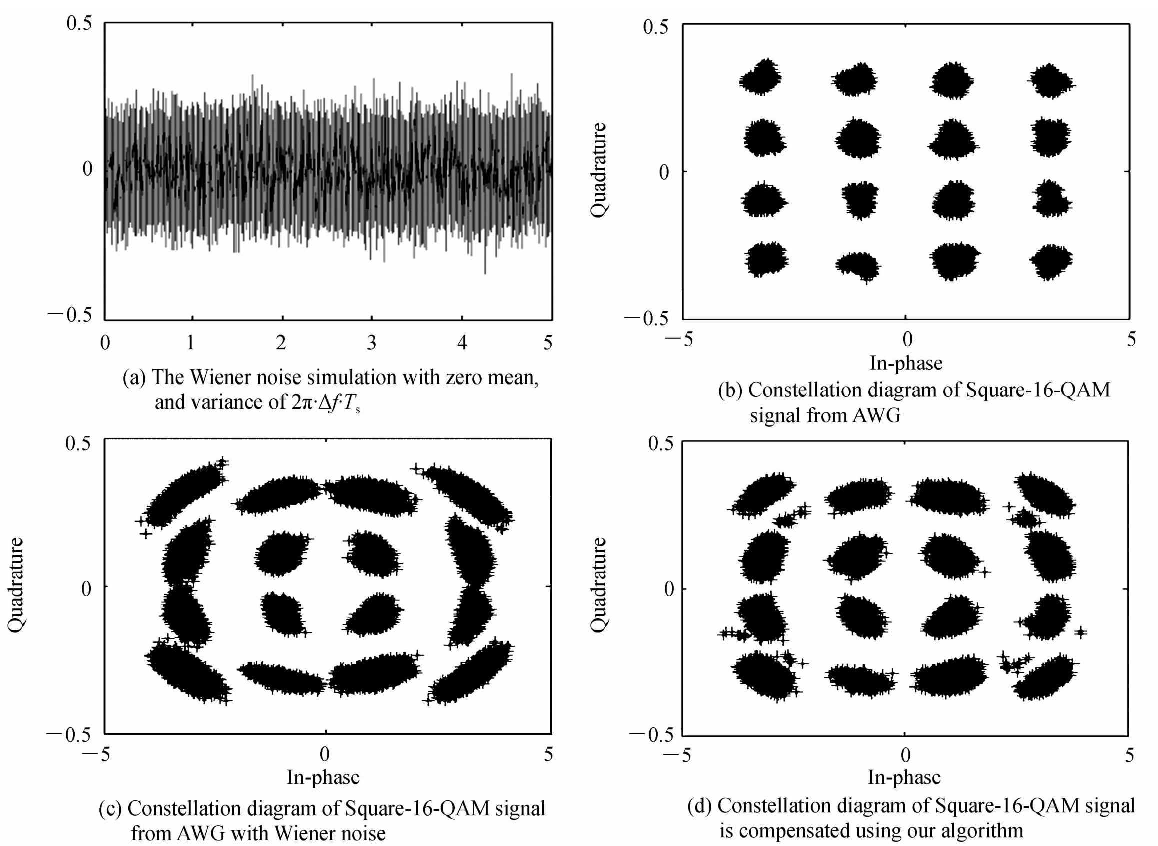


Fig. 6 The sum linewidth of signal and optical frequency comb is 10 MHz

In fact, when the advanced modulation formats that encode information on both amplitude and phase of the optical wave are detected using the linear optical sampling, it is found that the optical frequency comb is as an ideal sampling pulse source due to its frequency and phase stabilization. However, optical frequency combs may show strong phase noise, which makes its application be limited in a high-speed data transmission and an optical linear sampling as a sampling source. Recently, more efforts are paid for realizing the lower phase noise optical frequency comb. More works will publish on the application of optical frequency comb when optical frequency comb meet the highly demanding requirements of coherent communication or higher phase noise is able to be compensated by using a digital signal processing. In fact, optical pulses signal propagation in a network may be affected by a wide variety of phenomena, such as chromatic dispersion in fibers and components, polarization-mode dispersion, amplified spontaneous emission and nonlinear interactions. Therefore, special algorithms have to be proposed, which are adapted to the special environment of optical transmission. Optimized algorithm of unknown transmission noise of Square-16-QAM signals will be investigated in the future.

4 Conclusion

In conclusion, we analyze the Square-16-QAM constellation diagram measurement process by using the coherent optical linear sampling. In the linear optical sampling process, an optical frequency comb pulse with a pulse-width of 500fs and a duration of 10ns is used as an optical sampling source. Meanwhile, the phase noise between the signal and the sampling source lasers is compensated by using the algorithm. The results show that the phase noise is compensated effectly when the combined linewidth of the signal and optical sampling source is 100kHz , 1 MHz, 10MHz by this algorithm.

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