Optics Communications 410 (2018) 389–395

Contents lists available at ScienceDirect

Optics Communications

journal homepage: www.elsevier.com/locate/optcom

Simultaneously frequency down-conversion, independent multichannel phase shifting and zero-IF receiving using a phase modulator in a sagnac loop and balanced detection

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ARTICLE INFO

Keywords:
Frequency down-conversion
Multichannel phase shift
Zero-IF receiving
Phase modulator
Balanced detection

ABSTRACT

Photonic microwave frequency down-conversion with independent multichannel phase shifting and zero-intermediate frequency (IF) receiving is proposed and demonstrated by simulation. By combined use of a phase modulator (PM) in a sagnac loop and an optical bandpass filter (OBPF), orthogonal polarized carrier suppression single sideband (CS-SSB) signals are obtained. By adjusting the polarization controllers (PCs) to introduce the phase difference in the optical domain and using balanced detection to eliminate the direct current components, the phase of the generated IF signal can be arbitrarily tuned. Besides, the radio frequency (RF) vector signal can be also frequency down-converted to baseband directly by choosing two quadrature channels. In the simulation, high gain and continuously tunable phase shifts over the 360 degree range are verified. Furthermore, 2.5 Gbit/s RF vector signals centered at 10 GHz with different modulation formats are successfully demodulated.

1. Introduction

Microwave frequency conversion and phase shifting are two essential functions in the transmitter and receiver of many electronic systems, such as phase arrayed radar, wideband wireless communication, satellite payload, electronic warfare, and deep space exploration. For example, in millimeter-wave phase arrayed radar systems, on the one hand, a continuous full 360° tunable wideband phase shifter is a key component for phase-array beam pointing [1]. On the other hand, radio frequency (RF) signals should be down-converted into intermediate frequency (IF) signals at the receiver [2]. Thus, it is desirable to implement frequency down-conversion and phase shifting simultaneously. Besides, an important part of a modern radar system is the in-phase and quadrature (I/Q) detector, which enables the extraction of both real and imaginary parts of the reflected radar signal [3]. The image product is naturally canceled out in I/Q down conversion without filtering, so the receiver becomes frequency agile and versatile with different or multiple operating bands. Meanwhile, as compared with the IF signal in the super heterodyne receiver, the I/Q baseband signals have lower requirements for the operating frequency and sampling rate of the analog-to-digital converter (ADC).

One important characteristic of modern radar system is frequency agility [3], which means the receiver must be able to operate at any frequency in which the radar signal is being transmitted. The higher the frequency hopping range, the more difficult it is to intercept or jam the signal. Thus, it is necessary to realize wideband frequency down-conversion, multichannel phase shifting, and zero-IF receiving. The implementation of such a receiver using traditional microwave techniques could be difficult, particularly when operating over a broad frequency range, which is mainly due to the frequency-dependent characteristics of RF components. On the other hand, microwave photonic systems are able to become reconfigured electronically to operate at different frequencies over a broad frequency range and, thus, could be considered for such an application [4–7].

Recently, various photonic microwave frequency down-conversion methods have been proposed, including two cascaded MZMs biased at the quadrature points [8], dual-parallel MZM (DP-MZM) [9], optoelectronic oscillator based on a single dual-drive MZM (DD-MZM) [10]. Furthermore, many photonic microwave phase shifting schemes have also been studied, such as the combined use of a phase modulator and an optical filter [11], two cascaded polarization modulators (PolMs) [12], nonlinear polarization rotation in a highly nonlinear fiber [13], and so on. However, the above approaches cannot realize frequency down-conversion and phase shifting simultaneously. In fact, combining both functions into a simple compact configuration can not only simplify

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https://doi.org/10.1016/j.optcom.2017.10.044
Received 26 August 2017; Received in revised form 16 October 2017; Accepted 21 October 2017

the structure of the entire system, but also improve its performance. A photonic scheme to simultaneously achieve microwave frequency down-conversion and tunable microwave phase shift was proposed and demonstrated in [14]. RF and local oscillator (LO) signals drive two arms of the DD-MZM, respectively. A FBG reflects the first-order sidebands of both RF and LO frequency. After optical to electrical conversion, the IF signal is recovered, and the phase of the IF signal can be shifted continuously in full 360 degree. However, to realize independent multichannel phase shifting, multiple lasers or modulators are required, which increases the complexity of the system.

In this paper, we propose a simple photonic system that can implement frequency down-conversion, multichannel phase shifting, and zero-IF receiving concurrently. The key devices are a PM in a sagnac loop and an OBPF, which generate an orthogonal polarized carrier suppression single sideband (CS-SSB) signal with the frequency shifts equal to the frequencies of the RF and LO signals. By introducing phase difference between orthogonal polarized optical sidebands using a polarization controller (PC), dividing into two polarization directions via a polarization beam splitter (PBS), and beating at the balanced photodetector (BPD), frequency down-conversion with improved gain and independent multichannel phase shifting is achieved. In addition, zero-IF receiving can be also realized by choosing two quadrature channels. The proposed method is analyzed theoretically, demonstrated by simulation, and compared with the conventional two cascaded MZMs scheme.

2. Principle

2.1. Frequency down-conversion with multichannel phase shifting

Fig. 1 shows the schematic diagram of the proposed frequency down-conversion with multichannel phase shifting scheme based on bidirectional use of a PM in a sagnac loop and balanced detection. A light wave generated from a laser diode (LD) is injected to a polarization beam splitter (PBS_1) through a polarization controller (PC_1) and an optical circulator (OC). By adjusting PC_1, the light wave is split into two orthogonally polarized light waves that are equal in power. The PM is a commercial traveling-wave modulator with two electric ports, which are designed to have effective modulation for clockwise direction light waves at the first electric port (port 1) and have effective modulation for counter-clockwise direction light waves at the second electric port (port 2) [15]. For the clockwise direction, due to the velocity mismatch, the optical carrier is only modulated by the RF signal and the modulation by the LO signal is negligible. For the counter-clockwise direction, the optical carrier is only modulated by the LO signal and the modulation by the RF signal is negligible. Since the polarization states of the light waves from port 1 and port 2 are orthogonal, a 90° polarization rotation (PR) is introduced by a polarization rotator or a PC. Then, orthogonal polarized phase modulated signals reach the PBS_1 again, and amplified by the following erbium-doped fiber amplifier (EDFA). The optical bandpass filter (OBPF) is used to select upper optical sidebands of the orthogonal polarized phase modulated signals. The selected optical sidebands are sent to an optical splitter which is used to divide the optical signals into N channels. In each channel, a PC, a PBS and a balanced photodetector (BPD) are used to construct the rest of the photonic microwave phase shifter, and the phase of the IF signal after the BPD can be tuned easily by controlling the PC.

According to the bidirectional modulation characteristic of the PM, the optical signal at the output of the PBS_1 can be written as

$$\begin{bmatrix} E_{CK}(t) \\ E_{CCK}(t) \end{bmatrix} = \frac{\sqrt{2}}{2} E_c e^{jw_c t} \sqrt{l} \begin{bmatrix} e^{jm_{RF}\cos(w_{RF}t)} \\ e^{jm_{LO}\cos(w_{LO}t)} \end{bmatrix} \tag{1}$$

where E_c and w_c are the amplitude and angular frequency of the optical carrier, l is insertion loss of the PM, $m_{RF} = \pi V_{RF}/V_\pi$ represents the modulation index of the RF signal, V_{RF} is the RF receiving signal voltage applied to the PM, V_π is the half-wave voltage of the modulator, w_{RF} is the angular frequency of the RF receiving signal, $m_{LO} = \pi V_{LO}/V_\pi$ represents the modulation index of the LO signal, V_{LO} is the LO signal voltage applied to the PM, w_{LO} is the angular frequency of the LO signal.

Using Jacobi–Anger expansion, Eq. (1) can be approximately rewritten as

$$\begin{bmatrix} E_{CK}(t) \\ E_{CCK}(t) \end{bmatrix} = \frac{\sqrt{2}}{2} E_c e^{jw_c t} \sqrt{l} \begin{bmatrix} \sum_{n=-\infty}^{\infty} j^n J_n(m_{RF}) e^{jnw_{RF}t} \\ \sum_{n=-\infty}^{\infty} j^n J_n(m_{LO}) e^{jnw_{LO}t} \end{bmatrix} \tag{2}$$

where J_n is the (n)th-order Bessel function of the first kind.

As the modulation index is usually small, the second and higher-order Bessel function are ignored. Eq. (2) can be further simplified as

$$\begin{aligned} \begin{bmatrix} E_{CK}(t) \\ E_{CCK}(t) \end{bmatrix} &= \frac{\sqrt{2}}{2} E_c e^{jw_c t} \\ &\times \sqrt{l} \begin{bmatrix} jJ_1(m_{RF})e^{-jw_{RF}t} + J_0(m_{RF}) + jJ_1(m_{RF})e^{jw_{RF}t} \\ jJ_1(m_{LO})e^{-jw_{LO}t} + J_0(m_{LO}) + jJ_1(m_{LO})e^{jw_{LO}t} \end{bmatrix} \end{aligned} \tag{3}$$

It can be seen from Eq. (3) that the generated orthogonal polarized optical signal contains an optical carrier and ±1st order sidebands, as shown in Fig. 1A and B.

The orthogonal polarized optical signals are amplified by the EDFA and the upper sidebands are selected by the OBPF, as shown in Fig. 1C. It can be given as

$$E_{OBPF}(t) = \frac{\sqrt{2}}{2} E_c e^{jw_c t} \sqrt{lG_A} \begin{bmatrix} jJ_1(m_{RF})e^{jw_{RF}t} \\ jJ_1(m_{LO})e^{jw_{LO}t} \end{bmatrix} \tag{4}$$

where G_A is the gain of the EDFA.

Since the transformation matrix of the PC_n and PBS_n is given by

$$T_{PC-PBS} = \begin{bmatrix} \cos\alpha & -\sin\alpha e^{j\theta_n} \\ \sin\alpha & \cos\alpha e^{j\theta_n} \end{bmatrix} \tag{5}$$

where α is the angle between the principle state of the light and the principle axis of the PBS, and θ_n is the phase difference between the x and y polarization state induced by the PC_n. The output optical signal of the PBS_n can be written as

$$\begin{aligned} \begin{bmatrix} E_{PBS-1}(t) \\ E_{PBS-2}(t) \end{bmatrix} &= T_{PC-PBS} E_{OBPF}(t) \\ &= \frac{\sqrt{2}}{2} E_c e^{jw_c t} \sqrt{lG_A} \begin{bmatrix} \cos\alpha & -\sin\alpha e^{j\theta_n} \\ \sin\alpha & \cos\alpha e^{j\theta_n} \end{bmatrix} \begin{bmatrix} jJ_1(m_{RF})e^{jw_{RF}t} \\ jJ_1(m_{LO})e^{jw_{LO}t} \end{bmatrix} \\ &= \frac{\sqrt{2}}{2} jE_c e^{jw_c t} \sqrt{lG_A} \begin{bmatrix} \cos\alpha J_1(m_{RF})e^{jw_{RF}t} - \sin\alpha e^{j\theta_n} J_1(m_{LO})e^{jw_{LO}t} \\ \sin\alpha J_1(m_{RF})e^{jw_{RF}t} + \cos\alpha e^{j\theta_n} J_1(m_{LO})e^{jw_{LO}t} \end{bmatrix} \end{aligned} \tag{6}$$

It can be seen from Eq. (6) that two carrier suppression single sideband (CS-SSB) signals with frequency shift of w_{RF} and w_{LO} and phase difference of $\pi + \theta_n$ are obtained at the output of one principal axis of the PBS, and two CS-SSB signals with frequency shift of w_{RF} and w_{LO} and phase difference of θ_n are obtained at the output of the other principal axis of the PBS, as shown in Fig. 1D and E. After detected by a square-law BPD, the photocurrent can be written as

$$\begin{aligned} I_{BPDn}(t) &= R\left[\left|E_{PBS-1}(t)\right|^2 - \left|E_{PBS-2}(t)\right|^2\right] \\ &= -\frac{1}{2} RG_A l E_c^2 \begin{Bmatrix} \cos^2\alpha J_1^2(m_{RF}) + \sin^2\alpha J_1^2(m_{LO}) \\ \quad - \sin 2\alpha J_1(m_{RF})J_1(m_{LO}) \cos\left[(w_{RF} - w_{LO})t - \theta_n\right] \\ -\sin^2\alpha J_1^2(m_{RF}) - \cos^2\alpha J_1^2(m_{LO}) \\ \quad - \sin 2\alpha J_1(m_{RF})J_1(m_{LO}) \cos\left[(w_{RF} - w_{LO})t - \theta_n\right] \end{Bmatrix} \end{aligned} \tag{7}$$

where R is the BPD responsivity.

In order to suppress the DC components and improve the power of the IF signal, we set $\alpha = 45°$. Thus, Eq. (7) can be rewritten as

$$I_{BPDn}(t) = RG_A l E_c^2 J_1(m_{RF})J_1(m_{LO}) \cos\left[(w_{RF} - w_{LO})t - \theta_n\right] \tag{8}$$