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Photonic generation of versatile frequency-doubled microwave waveforms via a dual-polarization modulator

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ABSTRACT

We report a photonic approach to generate frequency-doubled microwave waveforms using an integrated electro-optic dual-polarization modulator driven by a sinusoidal radio frequency (RF) signal. With active bias control, two MZMs of the dual-polarization modulator operate at minimum transmission points, a triangular waveform can be generated by a parameter setting of modulation index. After introducing a broadband 90° microwave phase shifter, a square waveform can be obtained by readjusting the power relationship of harmonics. The proposal is first theoretically analyzed and then validated by simulation. Simulation results show that a 10 GHz triangular and square waveform sequences are successfully generated from a 5 GHz sinusoidal RF drive signal, and the performance of the microwave waveforms are not influenced by the finite extinction ratio of modulator.

1. Introduction

Photonic generation of microwave waveform has attracted much attention due to its wide applications in modern radar systems, wired and wireless communications, and all-optical microwave signal processing and manipulation [1–5]. Compared with the electrical techniques, which generate the microwave waveforms with the bandwidth limited to below 20 GHz, photonic generation of microwave waveforms has the advantages of wide bandwidth, large frequency tunable range, low loss, and immunity to electromagnetic interference. In recent years, various optical approaches have been reported to generate microwave waveform such as optical spectral shaping [6,7], frequency to time mapping [8], external modulation [9–14] and so on. Among them, the external modulation of a continuous wave (CW) optical signal exhibits the merits of low cost and high flexibility in wave shape.

The principle for microwave waveform generation using external modulation is based on controlling the phases and amplitudes of the optical sidebands generated due to the nonlinearity of the external modulator. It was reported that triangular waveform can be generated using a Mach–Zehnder modulator (MZM) combined with a dispersive element [9,10]. However, the repetition rate of the generated waveform is difficult to tune due to the use of fiber. In [11], a method to generate triangular waveforms using a MZM biased at the minimum transmission point and stimulated Brillouin scattering (SBS) in optical fiber was proposed. However, the SBS effect is sensitive to the environmental fluctuations, which may make the system unstable. In [12,13], triangular waveforms were generated using a dual-parallel MZM (DPMZM) in conjunction with a 90° electrical hybrid coupler or a tunable optical bandpass filter. However, the repetition frequency of the generated waveform equals to the frequency of RF drive signal, which is restricted by the bandwidth of modulator and electrical devices. In order to improve the frequency multiplication factor, a frequency-doubled triangular waveform was generated using a DPMZM combined with a frequency tripler [14]. Since the undesired optical sidebands are suppressed by controlling three bias points of DPMZM, the finite extinction ratio has an impact or distortion on the final waveform. Although the major distortion can be removed by bias control, the shape of the generated triangular waveform may be deteriorated when the extinction ratio is lower than 25 dB.

In this paper, we propose a versatile frequency-doubled microwave waveforms generation method based on a commercially available integrated dual-polarization modulator. By properly choosing the system parameters including the modulation index and the direct current bias phase shift, frequency-doubled triangular and square waveform sequences with a full duty cycle can be generated. A theoretical analysis leading to the operating conditions to obtain triangular and square waveforms is developed and demonstrated by simulation. Besides, the effects of several non-ideal factors on the performance of the microwave waveforms are analyzed. Simulation results show that the finite extinction ratio has no impact on the waveform.

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2. Principle

The schematic diagram of the proposed frequency-doubled microwave waveform generation scheme is shown in Fig. 1. A light wave from a laser diode (LD) is sent to a dual-polarization modulator via a polarization controller (PC). The dual-polarization modulator is a commercially available integrated device (FUJITSU FTM7980EDA) including a polarization beam splitter (PBS), a polarization beam combiner (PBC), and two MZMs. In the real system, the length of two paths are different, which leads to the phase imbalance between two radio frequency drive signals applied to two MZMs. In order to solve this problem, two tunable electrical phase shifters should be connected in the two paths to adjust the phase of two radio frequency drive signals. A low-frequency microwave signal is divided into two paths and applied to the two MZMs. By adjusting PC to let the polarization state of the input light wave an angle of 45° to one principal axis of the PBS, the output of the MZM1 can be expressed as.

$$E_x(t) = \frac{\sqrt{2}}{2}E_c e^{jw_c t}[\gamma_1 e^{jm_1 \sin(w_m t+\theta)} + (1-\gamma_1)e^{j\varphi_1}e^{-jm_1 \sin(w_m t+\theta)}] \tag{1}$$

where E_c and w_c are the amplitude and angular frequency of the optical carrier, w_m and θ are the angular frequency and initial phase of the input microwave signal, m_1 is the modulation index of the MZM1, $\gamma_1 = (1-1/\sqrt{\varepsilon_{r1}})/2$ is the power splinting or combining ration of arm two for the Y-branch waveguide of MZM1, ε_{r1} is the extinction ratio, and φ_1 is the direct current bias phase shift of MZM1.

Microwave driving signals of MZM1 and MZM2 are from the same RF signal. The difference is that a frequency tripler (FT) is used in the lower path, resulting in a frequency triple of the driving signal. Thus, the output of the MZM2 can be expressed as

$$E_y(t) = \frac{\sqrt{2}}{2}E_c e^{jw_c t}[\gamma_2 e^{jm_2 \sin(3w_m t+3\theta)} + (1-\gamma_2)e^{j\varphi_2}e^{-jm_2 \sin(3w_m t+3\theta)}] \tag{2}$$

where m_2 is the modulation index of the MZM2, $\gamma_2 = (1-1/\sqrt{\varepsilon_{r2}})/2$ is the power splinting or combining ration of arm two for the Y-branch waveguide of MZM2, ε_{r2} is the extinction ratio, and φ_2 is the direct current bias phase shift of MZM2.

Then, the output of MZM1 and MZM2 are combined by the PBC. After amplified by an erbium-doped fiber amplifier (EDFA) and detected by a square-law PD, the photocurrent can be expressed as

$$\begin{aligned} I(t) &= RG(|E_x(t)|^2 + |E_y(t)|^2) \\ &= \frac{RGE_c^2}{2}\{\gamma_1^2 + (1-\gamma_1)^2 + 2\gamma_1(1-\gamma_1)\cos[2m_1 \sin(w_m t+\theta) - \varphi_1] \\ &\quad +\gamma_2^2 + (1-\gamma_2)^2 + 2\gamma_2(1-\gamma_2) \\ &\quad \cos[2m_2 \sin(3w_m t + 3\theta) - \varphi_2]\} \end{aligned} \tag{3}$$

where R is the PD responsivity, and G is the EDFA gain.

If the MZM1 and MZM2 are biased at the minimum transmission points (φ_1=φ_2=π), using Jacobi–Anger expansion, the photocurrent can be rewritten as

$$\begin{aligned} I(t) = \frac{RGE_c^2}{2}\{&\gamma_1^2 + (1-\gamma_1)^2 - 2\gamma_1(1-\gamma_1)\{J_0(2m_1) + 2\textstyle\sum_{n=1}^{\infty} J_{2n}(2m_1)\cos[2n(w_m t+\theta)]\} \\ &+\gamma_2^2 + (1-\gamma_2)^2 - 2\gamma_2(1-\gamma_2)\{J_0(2m_2) + 2\textstyle\sum_{n=1}^{\infty} J_{2n}(2m_2) \\ &\cos[2n(3w_m t + 3\theta)]\}\} \end{aligned} \tag{4}$$

where J_n is the (n)th-order Bessel function of the first kind. It can be seen from Eq. (4) that only even-order harmonics are generated. We consider only frequency components up to second-harmonic, which means the modulation indices should be adjusted to relatively small value ($m_1 < 1$ and $m_2 < 1$). Eq. (4) becomes

$$\begin{aligned} I(t) = \frac{RGE_c^2}{2}\{&\gamma_1^2 + (1-\gamma_1)^2 - 2\gamma_1(1-\gamma_1)[J_0(2m_1) + 2J_2(2m_1)\cos(2w_m t + 2\theta)] \\ &+\gamma_2^2 + (1-\gamma_2)^2 - 2\gamma_2(1-\gamma_2)[J_0(2m_2) + 2J_2(2m_2)\cos(6w_m t + 6\theta)]\} \end{aligned} \tag{5}$$

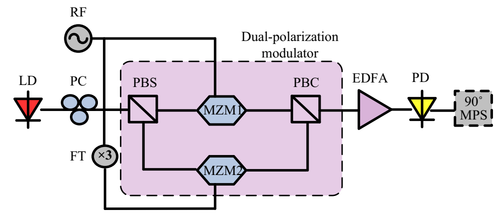

Fig. 1. Schematic diagram of a frequency-doubled microwave waveform generation system using a dual-polarization modulator. LD: laser diode; PC: polarization controller; PBS: polarization beam splitter; MZM: Mach–Zehnder modulator; PBC: polarization beam combiner; EDFA: erbium-doped fiber amplifier; PD: photodetector; RF: radio frequency; FT: frequency tripler; MPS: microwave phase shifter.

2.1. Triangular waveform

It is known that the Fourier series expansion of a triangular waveform can be expressed as

$$T_{tr}(t+t_0) = A_{tr} + B_{tr}\sum_{k=1,3,5}^{\infty}\frac{1}{k^2}\cos(k\Omega t + k\Omega t_0) \tag{6}$$

where A_{tr} and B_{tr} represent constant value, Ω is the fundamental angular frequency, t_0 is constant time. Eq. (6) shows that the shape of a triangular waveform is unchanged if phase shift $k\Omega t_0$ (k=1, 3, 5...) is introduced to the harmonic of the triangular waveform. Eq. (6) can not be satisfied for every k, an ideal triangular-shaped waveform is not feasible. However, a triangular waveform can be approximated by a finite number of the Fourier series components [15,16]. The triangular waveforms with and without considering fifth-order harmonic are simulated numerically, as shown in Fig. 2. It can be seen from Fig. 2 that the triangular waveforms with and without fifth-order harmonic are almost the same since the fifth-order harmonic is with very small amplitude and can not make a significant contribution to the final waveform. Thus, in our case, two Fourier components will be considered, and the modulation indices should be adjusted to meet the following relationship:

$$\gamma_1(1-\gamma_1)J_2(2m_1) = 9\gamma_2(1-\gamma_2)J_2(2m_2) \tag{7}$$

Eq. (5) can be rewritten as

$$I_{tr}(t) = A + B\left[\cos(\Omega t + \Omega t_0) + \frac{1}{9}\cos(3\Omega t + 3\Omega t_0)\right] \tag{8}$$

where

$$\begin{aligned} A = \frac{RGE_c^2}{2}\{&\gamma_1^2 + (1-\gamma_1)^2 - 2\gamma_1(1-\gamma_1)J_0(2m_1) + \gamma_2^2 + (1-\gamma_2)^2 \\ &- 2\gamma_2(1-\gamma_2)J_0(2m_2)\} \end{aligned} \tag{9}$$

$$B = -2RGE_c^2\gamma_1(1-\gamma_1)J_2(2m_1) \tag{10}$$

$$\Omega = 2w_m, \text{ and } \Omega t_0 = 2\theta$$

2.2. Square waveform

It is known that the Fourier series expansion of a square waveform can be expressed as

$$T_{sq}(t+t_0) = A_{sq} + B_{sq}\sum_{k=1,3,5}^{\infty}\frac{1}{(-1)^{\frac{k-1}{2}}k}\cos(k\Omega t + k\Omega t_0) \tag{11}$$

Similar to the triangular waveform, Eq. (11) can not be satisfied for every k, an ideal square waveform is not feasible. The square waveforms with and without considering fifth-order harmonic are simulated numerically, as shown in Fig. 3. In can be seen from Fig. 3 that the influence of the higher order harmonic on the waveform quality of the square waveform is obvious. Compared with the generation of a triangular waveform, the generation of a square waveform needs more