Complementary frequency shifter based on polarization modulator used for generation of a high-quality frequency-locked multicarrier

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Received October 22, 2013; revised December 9, 2013; accepted January 20, 2014; posted January 27, 2014 (Doc. ID 199879); published March 12, 2014

A novel polarization-modulator-based complementary frequency shifter (PCFS) has been proposed and then used to implement the generation of a frequency-locked multicarrier with single- and dual-recirculating frequency shifting loops, respectively. The transfer functions and output properties of PCFS and PCFS-based multicarrier generator have been studied theoretically. Based on our simulation results through VPItransmissionMaker software, 100 stable carriers have been obtained with acceptable flatness while no DC bias control is required. The results show that the proposed PCFS has the potential to become a commercial product and then used in various scenarios. © 2014 Optical Society of America

OCIS codes: (060.2330) Fiber optics communications; (060.2630) Frequency modulation; (230.2090) Electro-optical devices.

http://dx.doi.org/10.1364/OL.39.001513

The optical frequency-locked multicarrier with/without fixed phase relation, namely, optical frequency comb, has been broadly applied to optical arbitrary waveform generation [1], precise optical metrology [2], high accuracy optical sensor [3], ultrafast optical signal processing [4], photonic microwave signal processing [5,6], and high data rate optical fiber transmission [7–9]. There are many ways used to realize the generation of these multicarriers. Mostly, overdrive single and cascaded Mach–Zehnder modulators and/or phase modulators can obtain high-stability optical-frequency comb lines [10,11]. However, these methods need a high-power radio-frequency (RF) amplifier, which is the key component to obtain a larger number of comb lines. Additionally, one can apply the method based on the polarization modulator (PolM) to obtain a flat optical comb [12] while the number of comb lines is also limited. Recently, the recirculating frequency shifter (RFS)-based multicarrier generator (MCG) utilizing the single-side-band (SSB) modulator has been demonstrated extensively [7,8,13,14]. By using this method, especially, hundreds of stable comb lines can be achieved with dual-loop configuration based on an SSB modulator or complementary frequency shifter (CFS) [15], which is initially used to form a dual-carrier IQ modulator in [16]. However, the biggest disadvantage of this scheme is the requirement of three direct-current (DC) bias controls to make the modulator operate at the right point [17].

In this Letter, inspired by the configuration of CFS, we propose a novel PolM-based CFS polarization-modulator complementary frequency shifter (PCFS) to achieve the desired complementary SSB frequency shift where no DC bias control is required. Moreover, the proposed scheme also has the advantages of fewer phase modulators, broadband tunable frequency spacing, and lower requirement of RF power amplifier when compared with the conventional LiNbO3 CFS (CCFS). Hence, this PCFS would be attractive and cost-effective when used as a dual-carrier modulator or optical frequency shifter. Then, the PCFS-based multicarrier generator (PCMCG) used to generate hundreds of stable carriers has been implemented. The basic principle of the novel PCFS and the transfer function of PCMCG have been analyzed theoretically. Simulations of generating hundreds of carriers have been implemented by applying the VPItransmission-Maker software [18]. According to our previous works [13,15,17], the PCMCG can achieve the flexible frequency-spacing carrier number and full C-band covered multicarrier output with improved properties through the simulation.

The basic principle of the PCFS is shown in Fig. 1. It consists of a power splitter, two PolMs, a polarization-maintained optical coupler (PMOC), two PCs, two polarizers, and an RF source. The input RF signal $f(t)$ with its frequency of $f_m$ is split into two equal parts with a $\pi/2$ phase shift and then drives respective PolMs. Assuming the frequency of the input seed laser $E_{in}(t)$ is $f_0$, the novel PCFS can achieve a desired frequency-spacing SSB-shift with frequencies at both sides of the seed laser ($f_0 \pm f_m$) simultaneously by tuning the PCs. We represent the input linearly polarized seed laser from a tunable laser source as $E_{in}(t) = E_0 \exp(j\omega_0 t)$. The light-wave with its state of polarization (SOP) is aligned at an angle of $\alpha_t$ relative to one principal axis of the PolMs and then projected to

![Fig. 1. Principle of the novel polarization-modulator-based CFS (PCFS).](http://dx.doi.org/10.1364/OL.39.001513)
the two orthogonal directions. At the output of each PolM, the two modulated light waves are combined with a PMOC and then sent to a polarizer via a PC. The principal axis of the respective polarizer is aligned at α2 relative to the x-axis of the PolM. The two angles α1 and α2 can be tuned by the corresponding PCs. We also represent the input RF signals used to drive PolM1 and PolM2 as \( f_1(t) = V_{pp} \cos(\omega_m t) \) and \( f_2(t) = V_{pp} \sin(\omega_m t) \), respectively. To obtain the output of x- and y-polarization components of each output port, we adopt the transfer-matrix-based method, i.e., using \( H_{in} \), \( H_{PMOC} \), and \( H_{DPol} \) to represent the input optical signal and functions of PolMs, PMOC, and two polarizers, as shown in Eq. (1):

\[
H_{in} = \begin{bmatrix} \cos \alpha_1 \\ \sin \alpha_1 \end{bmatrix}, \quad H_{DPol} = \begin{bmatrix} \cos \alpha_2 \sin \alpha_2 \\ 0 \end{bmatrix}, \quad H_{PMOC} = \begin{bmatrix} 1 & 0 & j \\ 0 & 1 & 0 \end{bmatrix}, \quad H_{PolM} = \begin{bmatrix} H_{1x} \\ 0 \end{bmatrix}, \quad H_{PolM} = \begin{bmatrix} H_{1x} \\ 0 \end{bmatrix}, \quad H_{PolM} = \begin{bmatrix} H_{2x} \\ 0 \end{bmatrix}, \quad H_{PolM} = \begin{bmatrix} H_{2x} \\ 0 \end{bmatrix},
\]

where \( V_x \) is the half-wave voltage of PolM, and \( V_{pp} \) is the amplitude of the RF drive voltage. The parameter \( \phi \) is the static-phase difference between the x- and y-polarization components, which can also be tuned by the PC before the Pol. Then, the normalized transfer function of PCFS can be expressed mathematically as follows:

\[
T_{PCFS} = H_{DPol}H_{PMOC}H_{in} = \begin{bmatrix} T_{o1} & T_{o2} \end{bmatrix},
\]

\[
T_{o1} = C_1(H_{1x} + jH_{2x}) + C_2(H_{1y} + jH_{2y}) \]

\[
= [C_1 \exp(j\delta_m \cos(\omega_m t)) + C_2 \exp(-j\delta_m \cos(\omega_m t))] + j[C_1 \exp(j\delta_m \sin(\omega_m t)) + C_2 \exp(-j\delta_m \sin(\omega_m t))],
\]

\[
T_{o2} = C_1(H_{1x} + jH_{2x}) + C_2(H_{1y} + jH_{2y}) \]

\[
= [C_1 \exp(j\delta_m \cos(\omega_m t)) + C_2 \exp(-j\delta_m \cos(\omega_m t))] + j[C_1 \exp(j\delta_m \sin(\omega_m t)) + C_2 \exp(-j\delta_m \sin(\omega_m t))],
\]

\[
C_1 = \cos \alpha_1 \cos \alpha_2, \quad C_2 = \sin \alpha_1 \sin \alpha_2 e^{j\phi},
\]

where \( \delta_m = \pi V_{pp}/V_x \) is the phase modulation depth of PCFS. Note that this modulation depth is two times the CFS used in [15], which means that the costly RF power amplifier can be used. However, to achieve the desired dual-SSB frequency shift, the required condition deduced from Eq. (2) should be satisfied as shown follows:

\[
C_1 + C_2 = \cos \alpha_1 \cos \alpha_2 + \sin \alpha_1 \sin \alpha_2 e^{j\phi} = 0.
\]

Thus, the simplest case, which can satisfy the condition above, can be obtained as shown in Eq. (4):

\[
\alpha_1 - \alpha_2 = \frac{\pi}{2}, \quad \phi = 0.
\]

By applying this guiding result, the final transfer function of PCFS can be expressed as follows:

\[
T_{PCFS} = C_{1/2} \begin{bmatrix} \sin(\delta_m \cos(\omega_m t)) + j \sin(\delta_m \sin(\omega_m t)) \\ j \sin(\delta_m \cos(\omega_m t)) + \sin(\delta_m \sin(\omega_m t)) \end{bmatrix}.
\]

Obviously, the two outputs of PCFS have the same property of CCFS and therefore can realize the desired dual-SSB frequency shift. The significant issue is that the necessary DC bias control used in CCFS has been removed. Then, the transfer function of the PCMCG shown in Fig. 2(b) with dual-loop configuration can be obtained in the same way as in [15] as shown in the following:

\[
T_{PCMCG} = [T_1 \quad T_2] \quad \alpha \quad \begin{bmatrix} g_1 J_1(\delta_m) \exp(j\omega_m t) + b \exp(-j3\omega_m t) \exp(j\phiRT1) \\ jg_2 J_1(\delta_m) \exp(-j\omega_m t) + b \exp(j3\omega_m t) \exp(j\phiRT2) \end{bmatrix} \approx \begin{bmatrix} \exp(-j\omega_m t) + b \exp(j3\omega_m t) \exp(j\phiRT1) \\ j \exp(j\omega_m t) + b \exp(-j3\omega_m t) \exp(j\phiRT2) \end{bmatrix},
\]

where \( b = -J_3(\delta_m)/J_1(\delta_m) \) and \( J_{1/3}(\cdot) \) are the third-order crosstalk coefficient and the first kind odd-order Bessel functions. \( T_{1/2} \) and \( \phiRT1,2 \) are the transfer function and phase delay per RT in loop1 and loop2, respectively. For simplicity we have assumed that the optical amplifiers in loop1 and loop2 can compensate for the total losses exactly, including the insertion losses of all components and modulation loss. Note that the dual-loop configuration can be easily transformed to a single-loop configuration by choosing one output of the PCFS as shown in Fig. 2(a).

According to [15], when the desired carrier’s number is 2N for the dual-loop configuration, the stable output of PCMG after the N-th RT can be obtained by assuming that the bandwidth of respective optical band-pass filter in loop1 and loop2 cover the frequency range of \( (f_0, f_0 + NF_m) \) and \( (f_0 - NF_m, f_0) \), shown as follows:

\[
E_{in}(t) = E_{in}(t) + \sum_{n=1}^{N} \text{desired signals in loop1} \]

\[
E_{in}(t) = E_{in}(t) + \sum_{n=1}^{N} \text{desired signals in loop2} \]

\[
E_{in}(t) = E_{in}(t) + \sum_{n=1}^{N} \text{crosstalk components in loop1} \]

\[
E_{in}(t) = E_{in}(t) + \sum_{n=1}^{N} \text{crosstalk components in loop2} \]

where \( C_{n1} \) and \( C_{n2} \) are the normalized coefficients of crosstalk components to the desired signal components at their corresponding frequencies, shown as follows:
These results are coincident with those in [15]; however, there is still a slight difference in the coefficient of phase modulation depth shown in Eq. (2). Figure 3 shows the crosstalk coefficient \(|b|\) versus the drive voltage \(V_{pp}\). Clearly, the required RF drive voltage of PCFS is only one half of that of CCFS, which hints that the requirement of high RF power amplifier is reduced, e.g., \(V_{pp} = 0.27V_z\) is enough to make the crosstalk coefficient be around \(-15\) dB for PCFS.

To validate our proposed PCFS and PCMCG, the simulation results have been implemented by using the VPItransmissionMaker. Here, the frequency and voltage of drive RF signal \(f_m = 12.5\) GHz and \(V_{pp} = 0.27V_z\), respectively, and the seed laser with its center wavelength, linewidth and input power are assumed as \(1552.52\) nm, \(10\) MHz and \(0\) dBm, respectively. Figure 4 shows the output spectra of the PCFS transfer function at output port 1 [Figs. 4(a) and 4(c)] and port 2 [Figs. 4(b) and 4(d)]. Among those, Figs. 4(a), 4(b), and 4(d) show the outputs when \(\alpha_1 = 0^\circ\), \(\alpha_2 = -90^\circ\) which means that the condition shown in Eq. (4) is satisfied. Instead, Fig. 4(c) shows the outputs when \(\alpha_1 = 0^\circ\), \(\alpha_2 = -60^\circ\) for output1. Obviously, the desired carrier-suppressed PCFS can only be obtained when the condition is satisfied. The desired signal-to-crosstalk power ratio is 30 dB, which is in agreement with the theoretical analysis shown in Fig. 3. On the contrary, the desired SSB frequency shift will not be realized if the condition is not satisfied. As shown in Fig. 4(c), some unwanted-crosstalk components have appeared, such as the zero-, second-, and third-order harmonics.

Figures 5(a) and 5(b) show the 50-carrier output spectra of PCMCG with single-loop configuration for frequency up- and down-shifting. Here, the drive voltage of the RF signal \(V_{pp} = 0.27V_z\) is used above. The optical bandpass filter used in simulation is an ideal rectangle model with 630 GHz bandwidth which covers the desired 50-carrier frequency range. The stop-band attenuation assumed is 40 dB, which is a common parameter and can be realized by the commercial products, such as WaveShaper 4000 S and OTF-350. The optical amplifier we used is an EDFA with black-box model, which operated at saturation mode with 23 dBm output power and 4 dB noise figure. By sweeping the PC in the loop, the final 50-carrier outputs with a flatness of \(~4\) dB has been obtained for frequency up- and down-shifting as shown in Figs. 5(a) and 5(b), respectively. But, the flatness of the output carriers can be further improved by wave-shaping devices such as the wavelength-selective-switch shown in the dash-box in Fig. 2.

To show the output property of PCMCG with dual-loop configuration as shown in Fig. 2(b), the simulation results have also been obtained. As shown in Fig. 6(a), the final 100 carriers, except the seed laser, which cover a wavelength range of 10 nm, have been generated successfully. But, the flatness of output carriers is not...
so good due to the third-order crosstalk and the imperfect compensation gain of EDFA. As stated above, the flatness of output carriers can be further improved by applying wave-shaping devices. As shown in Fig. 6(b), the output 100-carrier with a flatness of 3 dB (shown in inset) can be realized. At the same time, the carrier-to-noise ratio (CNR) is around 25 dB due to the accumulated amplified spontaneous emission noise power.

Therefore, the proposed MCG based on PCFS has the significant advantage of no DC bias control to achieve a large number of frequency-locked multicarrier generation with high stability and CNR. In addition, compared with the conventional integrated LiNbO$_3$ SSB modulator, e.g., IQ modulator, PCFS will only require two phase modulators, which also can be fabricated with integrated structure. Even more, the typical operation frequency range of PolM is larger than 40 GHz, which means the output carriers with large tunable frequency spacing can be achieved. Moreover, the demand for large RF power amplifiers can be further reduced. All of these advantages will make the proposed scheme valuable and cost-effective. Although, just as in the simulation obtained here, this configuration has application feasibility based on our previous demonstration if the PCFS are available as a commercial product.

In conclusion, a novel CFS based on a polarization modulator has been proposed. The frequency-locked multicarrier generator based on the novel CFS with single- and double-recirculating loops has also been analyzed. The basic principle and simulation results have been implemented by applying the VPItransmissionMaker software. According to our analysis, high quality of hundreds of frequency-locked carriers can be achieved. The results show that the proposed cost-effective scheme will find some potential applications in various scenarios.

The authors would like to acknowledge the support of the National Basic Research Programme of China (973) Project (No. 2012CB315603), National Natural Science Foundation of China (NSFC) under Grant No. 61307092, the Open Fund of State Key Laboratory of Information Photonics and Optical Communications (Beijing University of Posts and Telecommunications), the Fundamental Research Funds for the Central Universities, and the Program for New Century Excellent Talents in University (NCET-12-0679) in China.

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