Dynamic range enhancement of a novel phase-locked coherent optical phase demodulator

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Abstract: We report on a novel cancellation technique, for reducing the nonlinearity associated with the tracking phase-modulator in recently proposed phase-locked coherent demodulator for phase modulated analog optical links. The proposed cancellation technique is input RF signal power and frequency independent leading to a significant increase in dynamic range of the coherent demodulator. Furthermore, this technique demonstrates that large values of the signal-to-intermodulation ratio of the demodulated signal can be obtained even though the tracking phase modulator is fairly nonlinear, and thereby relaxing the linearity requirements for the tracking phase modulator. A new model is developed and the calculated results are in good agreement with measurements.

OCIS codes: (060.1660) Coherent communications; (120.5060) Phase modulation

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1. Introduction

The use of optical links for the transmission of RF signals is a subject of considerable interest for future commercial and military systems [1, 2]. Intensity modulated analog optical links have been limited in performance by the nonlinear response of optical modulators [1]. The underlying reason for this is that the response of optical intensity modulators is 'hard-limited' by zero and full transmission. In contrast, optical phase modulation has no fundamental limit to modulation depth besides that given by the available modulation range in optical phase modulators. The challenge to implement a linear phase-modulated optical link lies in the receiver structure. A traditional coherent receiver has a sinusoidal response limiting the overall dynamic range of the optical link [3]. We have recently proposed, theoretically investigated and experimentally demonstrated a novel coherent optical phase-locked demodulator with feedback [4, 5, 6, 7] resulting in 15 dB of SFDR improvement compared to the traditional approach. The concept of this novel receiver is illustrated in Fig. 1. The output from the phase demodulator (a balanced optical mixer) is amplified and filtered by electronics, and then feed back to a local tracking phase modulator. Within the loop bandwidth, the effect of the feedback is to reduce the difference in phase between the local optical wave and the incoming wave. Therefore, the effective swing across the phase demodulator is reduced, resulting in an improved SFDR. This reduction could also be obtained by reducing the modulation depth at the transmitter but the signal-to-noise ratio (SNR) is reduced as a consequence. In contrast, in the proposed receiver, both the signal and the noise swings are reduced by the same factor (loop gain), retaining the SNR while improving the SFDR as shown in [4, 6]. However, to achieve a high bandwidth phase-locked receiver, compact semiconductor phase modulators have to be used to keep loop delay sufficiently low. These modulators can have fairly nonlinear response significantly limiting the dynamic range of the receiver [9]. Furthermore, the modulator distortion usually dominates over photodiode distortion and compensating for the modulators nonlinearities is therefore of significant importance [1]. In this paper, the impact of system nonlinearity, associated with tracking phase modulator, on the demodulated signal are determined in terms of the signal-to-intermodulation ratio. Furthermore, we propose a method to cancel out nonlinearities associated with the tracking phase modulator and inherently nonlinear response of the balanced detector.

2. Novelty of the work

A review of different linearization (cancellation) techniques can be found in reference [8]. So far, linearization techniques have been applied to intensity modulated analog optical links and mostly concentrated on the transmitter side. In many cases, the linearizer circuit was design to cancel either quadratic or cubic nonlinearity and cancellation of nonlinearities occurred in relatively narrow band (input RF signal power and frequency) [8]. We show that we can simultaneously cancel nonlinearities associated with the balanced receiver and tracking LO phase modulator by purely adjusting the loop gain and tailoring the nonlinearities of the tracking LO phase modulator. No extra circuitry is needed in order to obtain the cancellation. The proposed cancellation technique in this paper is frequency and power independent and has not been reported previously. Furthermore, the overall receiver concept is novel for linear optical phase demodulation, and as such, the first paper providing detailed analysis and deeper understanding of the receiver must therefore be not only novel but also very useful.

3. Model set-up

The set-up of the phase-locked optical demodulator, on which we base our model, is shown in Fig. 1.

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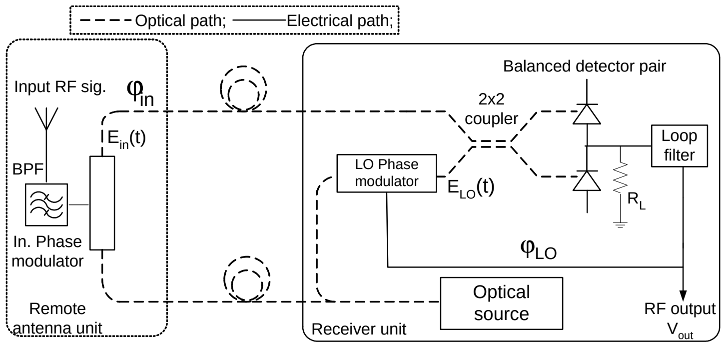

Fig. 1. General outline of phase-modulated optical link and phase-locked optical demodulator at the receiver unit.

The received RF signal, $V_{in}(t)$ is used to directly modulate an optical phase modulator at the remote antenna unit. The corresponding optical signal $E_{in}(t)$, see Fig. 1, is then written in complex notation as [1]:

$$E_{in}(t) = \sqrt{P_{in}} e^{j(\omega_0 t + \phi_{in}(t))} \tag{1}$$

where ω_0 is the optical frequency and P_{in} is the power of the optical field. Taking into consideration the nonlinearities associated with the (input) phase-modulator located at the remote antenna unit the phase of the optical signal, $\varphi_{in}(t)$, is expressed as:

$$\phi_{in}(t) = \frac{\pi V_{in}(t)}{V_{\pi,in}} \left(1 + \frac{a_2}{a_1 V_{\pi,in}} V_{in}(t) + \frac{a_3}{a_1 V_{\pi,in}^2} V_{in}^2(t) \right) \tag{2}$$

where $V_{\pi,in}$ is the voltage of the input phase-modulator, in order to obtain π phase shift and a_1, a_2 and a_3 represent the terms of the polynomial expansion of the input modulator nonlinear phase response. In order to characterize dynamic range of the demodulator, the input RF signal $V_{in}(t)$ is assumed to consist of relatively closely spaced tones [10]:

$$V_{in}(t) = V_1 \sin[\omega_1 t] + V_2 \sin[\omega_2 t] \tag{3}$$

where V_1 and V_2 are the amplitudes of the input RF signals and ω_1 and ω_2 are the input RF signal frequencies. The optical signal $E_{in}(t)$ is then transported to the receiver unit where its phase, is compared to the phase of the local optical signal $E_{LO}(t)$, using the balanced detector pair with load resistance R_L. A single optical source is used for both the remote antenna and the receiver unit. The optical LO signal, $E_{LO}(t)$, is thereby expressed as:

$$E_{LO}(t) = \sqrt{P_{LO}(t)} e^{j(\omega_0 t + \phi_{LO}(t))} \tag{4}$$

where $\phi_{LO}(t)$ is reference phase (signal) and is function of the feedback loop parameters, see Fig. 1 and P_{LO} is the power of the optical field[2]. Following Fig. 1 after the 3-dB coupler, we have in one arm:

$$E_1(t) = \frac{1}{\sqrt{2}} \sqrt{P_{in}} e^{j(\omega_0 t + \phi_{in}(t) - \pi/2)} + \frac{1}{\sqrt{2}} \sqrt{P_{LO}(t)} e^{j(\omega_0 t + \phi_{LO}(t) - \pi)} \tag{5}$$

[1] The scalar notation is used for both $E_{in}(t)$ and $E_{LO}(t)$ by assuming that the two fields are identically polarized.
[2] Due to the residual amplitude modulation of the tracking LO phase modulator P_{LO} will be time dependent. This is explained in more details later in the text

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The output of the second arm is:

$$E_2(t) = \frac{1}{\sqrt{2}}\sqrt{P_{in}}e^{j(\omega_0 t+\varphi_{in}(t)-\pi)} + \frac{1}{\sqrt{2}}\sqrt{P_{LO}(t)}e^{j(\omega_0 t+\varphi_{LO}(t)-\pi/2)} \tag{6}$$

where in equations (5) and (6) an ideal coupler has been assumed, i.e. equal splitting ratios. Taking into account the non-linearities associated with the photodetectors, the photocurrents generated in each branch of the balanced receiver, $I_1(t)$ and $I_2(t)$, containing the phase difference between $\phi_{in}(t)$ and $\phi_{LO}(t)$ are then expressed as:

$$\begin{aligned} I_1(t) &= R_{pd}|E_1(t)|^2\left(1+\frac{b_2|E_1(t)|^2}{b_1}+\frac{b_3|E_1(t)|^3}{b_1}\right) \\ &= R_{pd}\sum_{n=1}^{3}\frac{b_n}{b_1}\left(\frac{1}{2}P_{in}+\frac{1}{2}P_{LO}(t)-\sqrt{P_{in}P_{LO}(t)}\sin[\phi_{in}(t)-\phi_{LO}(t)]\right)^n \end{aligned} \tag{7}$$

$$\begin{aligned} I_2(t) &= R_{pd}|E_2(t)|^2\left(1+\frac{b_2|E_2(t)|^2}{b_1}+\frac{b_3|E_2(t)|^3}{b_1}\right) \\ &= R_{pd}\sum_{n=1}^{3}\frac{b_n}{b_1}\left(\frac{1}{2}P_{in}+\frac{1}{2}P_{LO}(t)+\sqrt{P_{in}P_{LO}(t)}\sin[\varphi_{in}(t)-\varphi_{LO}(t)]\right)^n \end{aligned} \tag{8}$$

where R_{pd} is the responsivity of the photodetectors and is assumed equal for both photodetectors. b_n represent the terms of polynomial expansion of the non-linear response of the photodiodes and n is an integer. In practise, it is only necessary to consider $n = 1..3$.The output signal from the balanced photodetector pair with load resistance R_L contains the phase difference between $\phi_{in}(t)$ and $\phi_{LO}(t)$ is expressed as:

$$V_{pd}(t) = R_L\left(I_2(t) - I_1(t)\right) \tag{9}$$

The signal $V_{pd}(t)$ is then used to control the feedback loop. (After the loop has acquired lock, the phase difference $\varphi_{in}(t) - \varphi_{LO}(t)$ will approach zero.) The signal $V_{pd}(t)$ is then passed through the loop filter (low pass) and amplified:

$$\frac{dV_{out}}{dt} = A\left[\frac{V_{pd}(t) - V_{out}(t)}{\tau_{LF}}\right] \tag{10}$$

where $\tau_{LF} = 1/2\pi f_{LF}$ is inversely proportional to the bandwidth of the loop filter and A is the gain of the loop filter. $V_{out}(t)$ is the output of the loop filter and the desired demodulated RF signal. $V_{out}(t)$ is then applied to the tracking LO phase modulator. The phase vs. voltage characteristic of the LO phase modulator is nonlinear. In practice, for the semiconductor phase modulators the quadratic and cubic nonlinearity terms will dominate over the higher order terms, and the phase-voltage relation can thereby be expressed as:

$$\phi_{LO}(t) = \frac{\pi V_{out}(t)}{V_{\pi,LO}}\left(1+\frac{c_2}{c_1 V_{\pi,LO}}V_{out}(t)+\frac{c_3}{c_1 V_{\pi,LO}^2}V_{out}^2(t)\right) \tag{11}$$

where c_1, c_2 and c_3 represent the terms of the polynomial expansion of the LO phase modulator response. In addition to nonlinearity associated with phase vs. voltage characteristic of the LO phase-modulator, any residual amplitude modulation, as would be expected in practice, may affect the performance of the demodulator in an adverse way. The normalized E-field amplitude of the (optical) LO signal can therefore be expressed as:

$$\sqrt{\frac{P_{LO}(t)}{P_0}} = \frac{A_{LO}(t)}{A_0} = 1 + D_1 V_{out}(t) + D_2 V_{out}^2(t) + D_3 V_{out}^3(t) \tag{12}$$

where A_0 if the E-field amplitude of the LO signal in the absence of amplitude modulation. In order to determine the overall dynamical response of the loop, the total phase error is defined as:

$$\phi_e(t) = \varphi_{in}(t) - \varphi_{LO}(t) \tag{13}$$

Taking the derivative of equation (13), (11) and (2), we obtain the differential equation describing the total phase error in the loop expressed in its general form as:

$$\begin{aligned} \frac{d\phi_e}{dt} &= \frac{\pi}{V_{\pi,in}} (\omega_1 V_1 \cos[\omega_1 t] + \omega_2 V_2 \cos[\omega_2 t]) \\ &\cdot \left(1 + \frac{2a_2}{a_1 V_{\pi,in}} (V_1 \sin[\omega_1 t] + V_2 \sin[\omega_2 t]) + \frac{3a_3}{a_1 V_{\pi,in}^2} (V_1 \sin[\omega_1 t] + V_2 \sin[\omega_2 t])^2 \right) \\ &- \frac{A\pi}{\tau_{LF} V_{\pi,LO}} \left(1 + \frac{2c_2}{c_1 V_{\pi,LO}} V_{out}(t) + \frac{3c_3}{c_1 V_{\pi,LO}^2} V_{out}^2(t)\right) \left(V_{pd}(t) - V_{out}(t)\right) \end{aligned} \tag{14}$$

It should be noted that Eq. (14) includes the effects of cascaded sources of nonlinearities associated with the input phase modulator, sinusoidal response of the balanced receiver, photodetectors and tracking LO phase modulator, giving a good insight into demodulators dynamics. We are therefore going to base our model on Eq. (14) and (10). By solving Eq. (10) together with Eq. (14), the desired demodulated signal, $V_{out}(t)$, is obtained characterizing the overall nonlinear response of the loop. Eq. (14) and (10) are first order non-linear differential equations and their solutions can be obtained numerically.

When the input RF signal consists of relatively closely spaced frequencies, the nonlinear response of the loop components will result in intermodulation distortion of the demodulated signal. 3rd order intermodulation products are especially important because they may set the Spurious Free Dynamic Range (SFDR) of the system [10]. The demodulated signal, $V_{out}(t)$ is then characterized by the Signal-to-Intermodulation Ratio (SIR) which is the ratio between the power of the demodulated signal (ω_1 or ω_2) and 3rd order mixing product ($2\omega_1 - \omega_2, 2\omega_2 - \omega_1$). Loop gain is defined as: $K = (\pi\sqrt{P_{in}P_0} A R_{pd} R_L)/V_{\pi,LO}\tau_{LF}$.

4. Linearity analysis based on perturbation theory

The time domain numerical model, based on nonlinear differential Eq. (14) and (10) is of rather complex nature and there are many parameters involved. The model is therefore detailed and in good agreement with experimental results, as it will be shown in section 5. However, due to the complex behavior of the nonlinear systems, it may be cumbersome to interpret its results. In this section, we therefore derive simple approximate analytical expressions which are qualitatively in good agreement with time domain numerical model. The analytical expressions are going to be used to interpret results obtained by the detailed time domain numerical model. For simplicity, the loop filter and nonlinearities associated with the photodetectors are not considered. Furthermore, V_π is assumed equal for the input and tracking LO phase modulator.

Let $V_{ref}(t)$ denote the signal incident at the tracking LO phase modulator when the loop is open:

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$$
\begin{aligned}
V_{ref}(t) &= 2R_{pd}\sqrt{P_{in}P_{LO}}R_L A \sin\left[\frac{\pi}{V_{\pi,LO}}V_{in}(t) - \frac{\pi}{V_{\pi,LO}}\left(V_{out}(t) + c_2' V_{out}^2(t) + c_3' V_{out}^3(t)\right) + \phi_0\right] \\
&\approx 2R_{pd}\sqrt{P_{in}P_{LO}}R_L A \sin[\phi_0] + G_1\left[V_{in}(t) - V_{out}(t) - c_2' V_{out}^2(t) - c_3' V_{out}^3(t)\right] \\
&- G_2\left[V_{in}(t) - V_{out}(t) - c_2' V_{out}^2(t) - c_3' V_{out}^3(t)\right]^2 \\
&- G_3\left[V_{in}(t) - V_{out}(t) - c_2' V_{out}^2(t) - c_3' V_{out}^3(t)\right]^3
\end{aligned}
\tag{15}
$$

where $c_2' = c_2/(c_1 V_{\pi,LO})$, $c_3' = c_3/(c_1 V_{\pi,LO}^2)$ and ϕ_0 is a constant. $G_1 = (2R_{pd}\sqrt{P_{in}P_{LO}}R_L A\pi \cos[\phi_0])/V_{\pi,LO}$, $G_2 = (2R_{pd}\sqrt{P_{in}P_{LO}}R_L A\pi^2 \sin[\phi_0])/2V_{\pi,LO}^2$ and $G_3 = (2R_{pd}\sqrt{P_{in}P_{LO}}R_L A\pi^3 \cos[\phi_0])/6V_{\pi,LO}^3$. However, since the DC term is out of the signal band and it is filtered away when the loop is locked, we chose not to consider the DC term. Since the response of the optical phase demodulator will be nonlinear due to the inherently nonlinear response of the balanced receiver (tracking signal $V_{ref}(t)$ is a nonlinear function), the output signal of the demodulator, $V_{out}(t)$, after locking the loop can be approximated as [10]:

$$
V_{out}(t) = A_1 V_{in}(t) + A_2 V_{in}^2(t) + A_3 V_{in}^3(t) \tag{16}
$$

where A_1, A_2 and A_3 are constants. We assume that 3rd mixing product $A_3 V_{in}^3(t)$ will have larger impact on the SFDR than the second order mixing product $A_2 V_{in}^2(t)$ and we chose therefore not to consider the second order mixing product. Furthermore, the input signal $V_{in}(t)$ consists of closely spaced tones: $V_{in}(t) = V_1 \sin[\omega_1 t] + V_1 \sin[\omega_2 t]$. In order to find A_1 and A_3, we lock the loop $V_{ref}(t) = V_{out}(t)$ and insert Eq. (16) in (15). Using the method of harmonic balance the coefficients A_1 and A_3 are found. The demodulated signal, $V_{out}(t)$ is then expressed as:

$$
V_{out}(t) = \frac{G_1}{1+G_1}V_{in}(t) - \frac{(c_3' G_1^4 + G_3 - 2G_2 c_2' G_1^2)}{(1+G_1)^4}V_{in}^3(t) \tag{17}
$$

It is observed from Eq. (17), that for a specific c_2' by adjusting the loop parameters, the 3rd order mixing product of the demodulated signal can be (theoretically) brought to zero, i.e. $A_3 = 0$. In other words the feedback circuit in combination with second order nonlinearity associated with tracking LO phase modulator response, results into cancellation of 3rd order mixing product of the demodulated signal.

In addition to nonlinearities of the LO phase-modulator, any residual amplitude modulation would be present as explained in section 3. We therefore need to investigate the impact of the residual amplitude modulation on the SIR. Inserting Eq. (12) and (16) in (15), and locking the loop $V_{ref}(t) = V_{out}(t)$, the output signal can be expressed as:

$$
V_{out}(t) = \frac{G_1}{1+G_1}V_{in}(t) - \left(\frac{2G_2 c_2' G_1^2 + D_1 G_1^4 c_2' + G_3 + c_3' G_1^4 - D_2 G_1^3 - D_1 G_1 G_2}{(1+G_1)^4}\right)V_{in}^3(t) \tag{18}
$$

The result presented in Eq. (18) is encouraging since the nonlinearities associated with the tracking LO phase modulator can be cancelled out. Either loop gain can be used in order to

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obtain cancellation of the nonlinearities, or for a specific loop gain the tracking LO can be tailored such that the nonlinearities cancel out.

5. Experimental results

The simulation results shown in this section and next sections are obtained by the detailed time domain numerical model based on Eq. (14) and (10). The experimental set-up, similar to Fig. 1, was constructed in order to verify the model [4]. The experimental bandwidth is limited by the time delay imposed by the discrete components of the receiver. An integrated version of the receiver is necessary to scale to GHz operation.

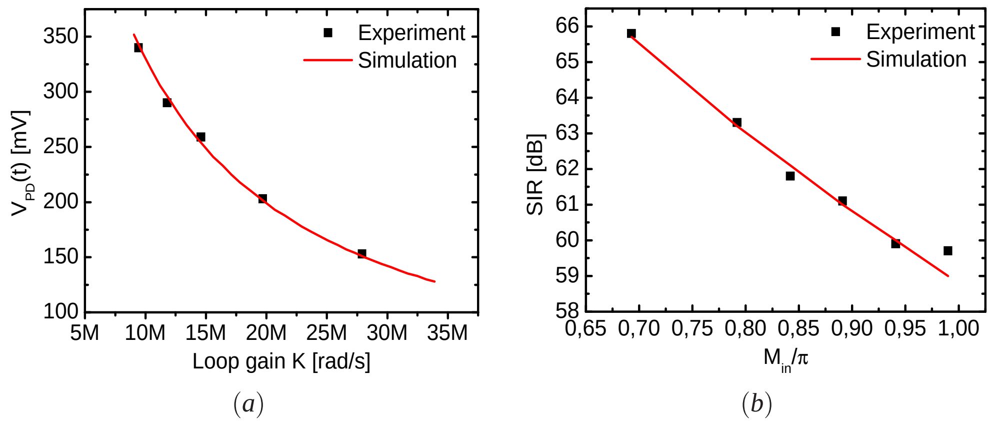


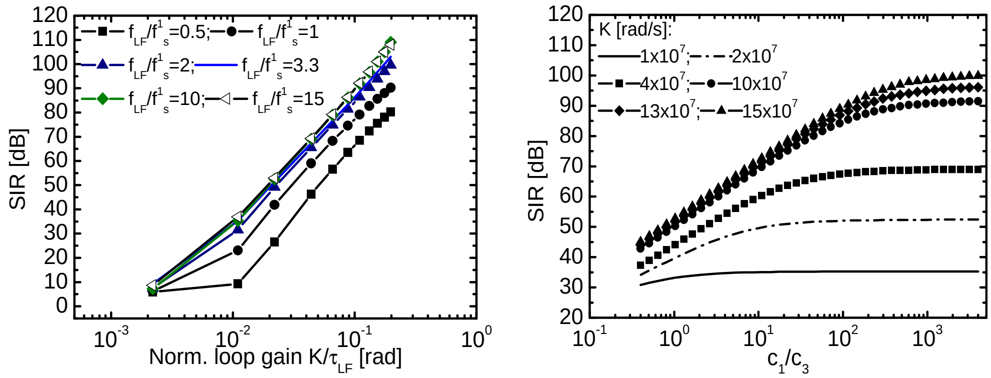
Fig. 2. (a) One tone measurement. Output of the balanced photodetector, $V_{pd}(t)$, as a function of loop gain. (b) Two tone measurement. SIR as a function of input signal modulation depth, M_{in}.

In Fig. 2 (a), a one-tone measurement is shown together with simulation results. Due to the experimental bandwidth limitation the input RF signal frequency is only $f_1 = 150$ kHz and the loop filter bandwidth is 1.1 MHz. The amplitude of the signal after balanced photodetection, $V_{pd}(t)$, is plotted as a function of the loop gain, K. Experimental and simulation results show that as the loop gain is increased, the amplitude of $V_{pd}(t)$ is reduced, i.e. the linearity of the demodulator is improved. Good agreement between the experimental and simulation results is obtained for one tone measurement. In Fig. 2 (b), results of the two tone measurement are shown together with the simulation results. The SIR is plotted as a function of the modulation depth, $M_{in} = (\pi/V_{\pi,in})V_1$, of the input RF signal. V_1 is the amplitude of the input RF signal and is assumed equal for both tones. The input RF signal frequencies are: $f_1 = 150$ kHz and $f_2 = 170$ kHz. As expected, the SIR decreases as M_{in} is increased. Once again good agreement between the model and experimental results is observed.

6. Effects of loop gain and LO phase-modulator nonlinearities

We set a goal of 90 dB of the SIR for the modulation depth of $\pi/2$. As mentioned in the introduction the modulator distortion, especially of the tracking LO phase modulator in the considered case, will usually dominate over the photodiode distortion. We therefore assume that the tracking LO phase modulator is much more nonlinear than the photodiodes, i.e. $c_2/b_2 >> 1$ and $c_3/b_3 >> 1$. Furthermore, electronics nonlinearities can be suppressed by the feedback loop and only need to be lower than the nonlinearities of the tracking LO phase modulator response. In contrast, the tracking LO phase modulator nonlinearities are not suppressed and must therefore be carefully considered. However, a linear input phase modulator is considered.

In Fig. 3 (a), Signal-to-Intermodulation Ratio (SIR) is computed as a function of the loop gain when the ratio between loop filter bandwidth and the RF input signal frequency, (f_{LF}/f_1), is varied. Input RF signal $V_{in}(t)$ includes two closely spaced frequencies: $\omega_1/2\pi = f_1$ and $\omega_2/2\pi = f_2$, as shown in Eq. (3). Linear tracking LO phase-modulator is assumed in Fig. 3 (a). The intermodulation is the magnitude of the mixing terms $(2f_1 - f_2, 2f_2 - f_1)$.



(a) (b)

Fig. 3. RF input signal modulation depth $M_{in} = \pi/2$. (a) SIR of the demodulated signal as a function of normalized loop gain for selected values of the ratio f_{LF}/f_1. (b) SIR of the demodulated signal as a function of the ratio c_1/c_3 of the LO phase-modulator. Quadratic term $c_2 = 0$.

Figure 3 (a) illustrates that as the loop gain K is increased, the performance of the phase-locked demodulator improves in terms of SIR, i.e. the SIR of the demodulated signal increases. As the ratio, (f_{LF}/f_1), is significantly increased, the SIR converges. Furthermore, the slope of the SIR line is approximately 3. As observed in Fig. 3 (a) relatively large values of the SIR can be obtained provided large loop gain and linear tracking LO phase-modulator. Using Eq. (17) and setting $c_2, c_3 = 0$, the expression for the SIR can be obtained:

$$SIR = 20\log\left[\frac{8(1+G_1)^3}{M_{in}^2}\right] \tag{19}$$

Even though Eq. (19) is derived for the loop without the loop filter, it is in good correspondence with Fig. 3 (a). Eq. (19) and Fig. 3 (a) show that the SIR increases with loop gain with a slope of 3.

One of the key challenges in creating a linear demodulator is the linearity of the tracking LO phase-modulator. The phase-change vs. voltage characteristic of the LO phase-modulator is nonlinear and thereby reducing the SIR of the demodulated signal. First, we are going to investigate the impact of cubic nonlinearity on the SIR. In Fig. 3 (b), SIR is computed as the ratio between the linear term (c_1) and cubic term (c_3) of the LO phase-modulator response for selected values of the loop gain. In general, the SIR decreases as the ratio c_1/c_3 decreases. The values of c_1/c_3 for which SIR starts to decrease are loop gain dependent since the nonlinearities of the LO phase-modulator become more enhanced as the loop gain is increased. This is also in accordance with Eq. (17), i.e. as the loop gain is increased 3rd order mixing product

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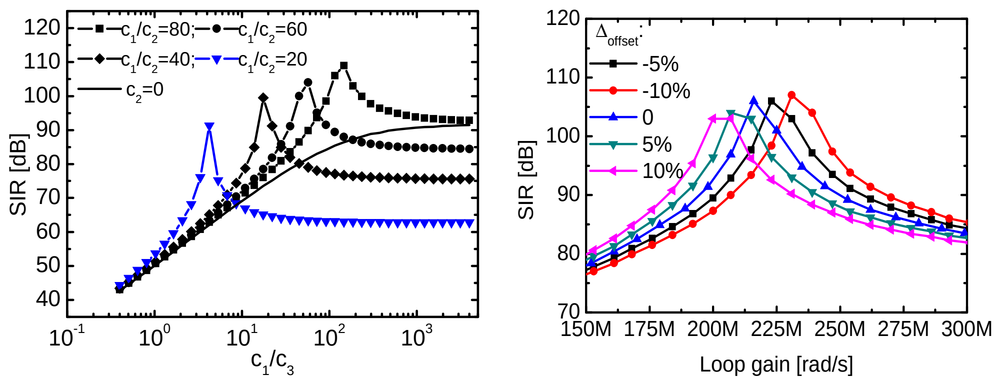
increases as well. For the loop gain of $K = 10 \cdot 10^7$ rad/s, the ratio c_1/c_3 needs to be > 200, to maintain the SIR of 90 dB. Very recently, the the ratio between cubic term and linear term c_1/c_3 of the semiconductor phase modulator has been measured to be 26 [9]. Obviously, a lot of improvement is needed in order to obtain the $c_1/c_3 = 200$.

Next, we investigate how the quadratic term c_2 of the nonlinear response of the phase modulator can be used to cancel out 3rd order mixing product of the demodulated signal.



(a) (b)

Fig. 4. SIR of the demodulated signal as a function of c_1/c_3. The ratio c_1/c_2 takes values: $80, 60, 40, 20$. (b) SIR of the demodulated signal as a function of loop gain, K. $c_1/c_2 = 40$ and $c_1/c_3 = 20 + \Delta_{offset}$.

In Fig. 4 (a), SIR of the demodulated signal is computed as a function of the ratio c_1/c_3 for the selected values of c_1/c_2 ratio. For the reference a SIR is also plotted for the case when $c_2 = 0$. The selected values of the ratio c_1/c_2 are experimentally obtainable for electrical circuits [10]. To the author's best knowledge the measurement of the quadratic term for the semiconductor phase-modulators has not been performed yet, however, some initial measurements are under constructions by the authors group. We have therefore varied the ratio c_1/c_2 in order to cover low and high values. Fig. 4 (a) shows that as the ratio c_1/c_2 is decreased the SIR becomes severely limited by the second order nonlinearity. Furthermore, for low values of the third order nonlinearity of the LO phase modulator, the SIR becomes completely dominated by the second order nonlinearity, i.e. independent of the c_1/c_3 ratio. This is in accordance with Eq. (17), i.e. if c_3 is low 3rd order mixing product of the demodulated signal is dominated by c_2. However, the results in Fig. 4 (a) also show that for non-zero values of c_2, there exist a combination of c_2 and c_3 for which the 3rd order mixing product is minimized, i.e. peaking (resonance) of the SIR. This implies that the combined effects of the nonlinearities associated with the balanced receiver and the nonlinear LO phase-modulator response plus the gain contribution from the feedback loop, result in cancellation of 3rd order mixing product. This is also observed from Eq. (17) as stated earlier. Furthermore, the resonance peak moves towards lower higher of c_3 as c_2 is increased. So, by having second order nonlinearity associated with tracking LO phase modulator, we can tolerate more cubic nonlinearity, c_3. Having the ratio c_1/c_3 of only 20 and $c_1/c_2 = 40$, the SIR of 90 dB can still be obtained.

Suppose that we want to tailor the phase-modulator such that SIR peaking is obtained. In practise, we may not be able to match exactly the required values of c_1/c_3 and c_1/c_2 in order to obtain SIR peaking, so we need to investigate what happens if we are slightly off. In Fig. 4

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(b), the SIR is computed as a function of loop gain when the ratio c_1/c_3 is varied from the exact value of c_1/c_3 for which the SIR peaking is obtained, i.e. $c_1/c_3 = 20 + \Delta_{offset}$. The ratio $c_1/c_2 = 40$ is held constant. Fig. 4 (b) shows that the SIR peaking is dependent on the loop gain and it occurs in a relatively wide band of the loop gain. It is also noticed that as the ratio c_1/c_3 is varied, the resonant peak of the SIR moves as well. So, by adjusting the loop gain resonant peaking of the SIR can be re-obtained. Another, thing which should be addressed is frequency dependence of c_3 if the demodulator is operated over wide frequency range. Frequency dependence will cause the c_1/c_3 to vary, and Fig. 4 (b) can be used to observe the effect of varying c_1/c_3. If the ratio c_1/c_3 varies with frequency for a specific loop gain, we will move away from the resonant peaking of the SIR. One solution could be to re-adjust the loop gain or to design wide band tracking LO phase modulator. Furthermore, the demodulator could be designed to operate in narrow frequency band.

7. Effects of residual amplitude modulation

In this section, the impact of residual amplitude modulation on the SIR is considered. Nonlinear phase response associated with the tracking LO phase modulator is set to zero, i.e. $c_2 = c_3 = 0$. From Eq. (18) it can be observed that D_1 and D_2 will have impact on the 3rd order mixing product of the demodulated signal. However, as seen in Eq. (18) D_3 has no impact on the 3rd order mixing product of the demodulated signal. In practice, coefficients D_1, D_2 and D_3 can be related to the absorption coefficient of the tracking LO phase modulator.

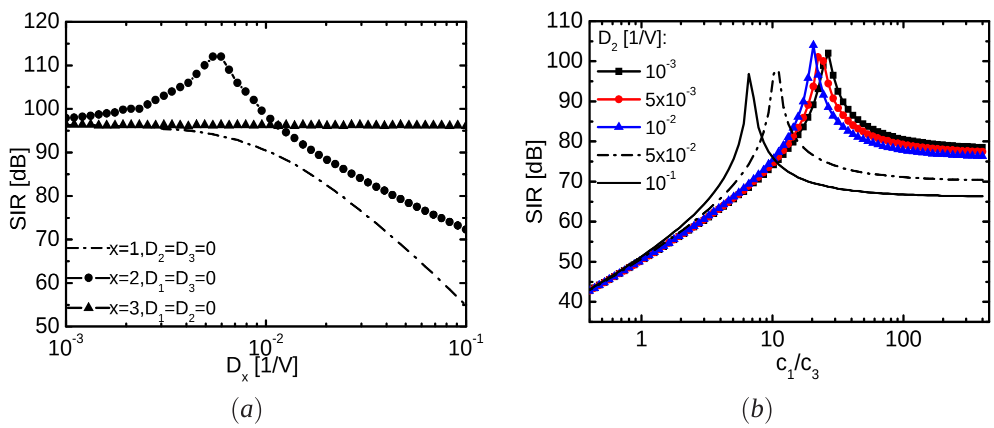

Fig. 5. (a) SIR of the demodulated signal as a function of D_x for x=1,2,3. $(c2, c3) = 0$. (b) SIR of the demodulated signal as a function of c_1/c_3 for selected values of the quadratic term, D_2, of the residual amplitude modulation. $c_1/c_2 = 40$ and $D_1 = 0.03$ 1/V.

In Fig. 5 (a), the SIR of the demodulated signal is computed as a function of D_x where $x = 1,2$ and 3. We assume that terms D_x where $x > 3$ are negligible, as it would be expected in practice. Fig. 5 shows that as D_1 $(D_x = 0, x = 2,3)$ is increased beyond $5 \cdot 10^{-3}$ 1/V, the SIR starts to decrease. When D_2 is varied $(D_x = 0, x = 1,3)$ resonant behavior of the SIR, similar to Fig. 4(b), is observed, i.e. there exist a value of D_2 for which the 3rd order mixing product is minimized. Furthermore, we observe, Fig. 5 (a), that the effect of D_1 is more deteriorating than that of D_2. For the case when D_3 is varied $(D_x = 0, x = 1,2)$, the SIR is not affected and this is in accordance with Eq. (18). Usually, for the semiconductor phase modulator, the quadratic term of the amplitude modulation is more difficult to minimize than the linear term. We therefore need to concentrate on the quadratic term of amplitude modulation.

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8. Combined effects of nonlinearities and their cancellation

In this section, the combined effect of nonlinearities (nonlinear phase response and residual amplitude modulation) associated with tracking LO phase modulator are considered. We pay more attention on the quadratic term, D_2, of the residual amplitude modulation since it is more difficult to reduce (in practice) for semiconductor phase modulators, compared to the linear term, D_1.

In Fig. 5 (b), SIR is computed as a function of the the ratio c_1/c_3 when D_2 is varied from 10^{-3} 1/V to 10^{-1} 1/V. The ratio c_1/c_2 is set to 40 and $D_1 = 0.03$ 1/V. It is observed from Fig. 5 (b) that for low values of cubic nonlinearity (c_3) of the phase modulator response, the SIR is fully limited by the quadratic term of residual amplitude modulation, D_2. However, Fig. 5(b) also shows that there exists a combination of c_3, c_2, D_1 and D_2 for which the 3rd order mixing product is minimized, i.e. peaking (resonance) of the SIR. The resonant peak moves towards lower values of c_1/c_3 ratio as D_2 is increased. Fig. 5 (b) is in accordance with Eq. (18) which states that 3rd order mixing product can be reduced by proper combination of the loop gain and tracking LO phase modulator nonlinearities. The cascaded sources of nonlinearities associated with balanced detector, phase and amplitude modulation of the tracking LO phase modulator cancel each other. Fig. 5 (b) illustrates that even though the ratio c_1/c_3 is low for the measured semiconductor phase modulator ($c_1/c_3 = 26$ [9]), high values of the SIR are still obtainable.

Next, we are going to investigate how the resonant peaking of the SIR scales with input RF signal voltage and frequency. We must be sure that the cancellation of the nonlinearities does not only occur at single power level or single frequency. Nonlinearities associated with the tracking LO phase modulator are chosen such that resonant peaking of the SIR is obtained, see Fig. 5 (b). In Fig. 6 (a), we plot the amplitude of the fundamental of the demodulated signal (f_1) and amplitude of the 3rd order intermodulation product $(2f_1 - f_2)$ as a function of input signal voltage, V_{in}. The amplitude of the fundamental and 3rd order intermodulation product (IM_3 curve) of the demodulated signal increases with the input signal voltage, as expected. It should be observed that there are no dips in the IM_3 curve as the input signal voltage is varied. This means that the cancellation of the nonlienarities associated with the balanced receiver and phase modulator occurs over broad range of input RF signal powers. In Fig. 6 (b), the SIR is computed as a function of input RF signal frequency as the input signal modulation depth, M_{in} takes values from π to $\pi/7$. Fig. 6 (b) shows that the SIR remains constant as the input RF signal frequency is increased. As the input RF signal modulation depth decreases, the SIR increases as expected.

9. Conclusion

A novel cancellation technique, which is RF signal power and frequency independent, for the recently demonstrated phase-locked coherent optical phase demodulator is proposed and numerically investigated. It has been shown that the interplay between the loop gain of the feedback circuit and nonlinearities of the tracking LO phase modulator may result in cancellation of the third order intermodulation product of the demodulated signal. The cancellation of the third order intermodulation product can be achieved by carefully tailoring the nonlinearities of the tracking LO phase modulator, or for the specific set of nonlinearities loop gain can be adjusted such that cancellation is obtained. The proposed technique significantly reduces linearity requirements for the tracking LO phase modulator and large values of the signal-to-intermodulation ratio can thereby be obtained even though the tracking LO phase modulator is fairly nonlinear.

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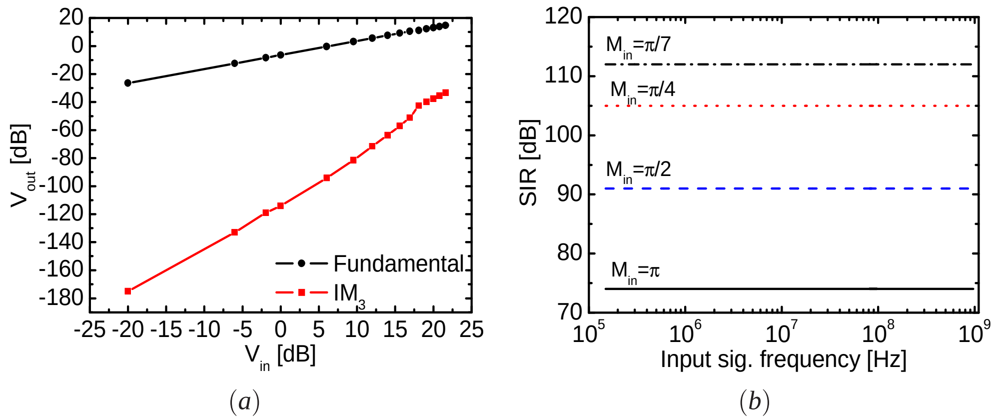

Fig. 6. $c_1/c_3 = 10$, $c_1/c_2 = 40$, $D_1 = 0.03$ 1/V and $D_2 = 0.05$ 1/V. (a) Amplitude of the fundamental and 3rd order mixing product (IM_3 curve) of the demodulated signal as a function of input signal voltage, V_{in}. (b) SIR of the demodulated signal as a function of input signal frequency for selected values of modulation depth M_{in}

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