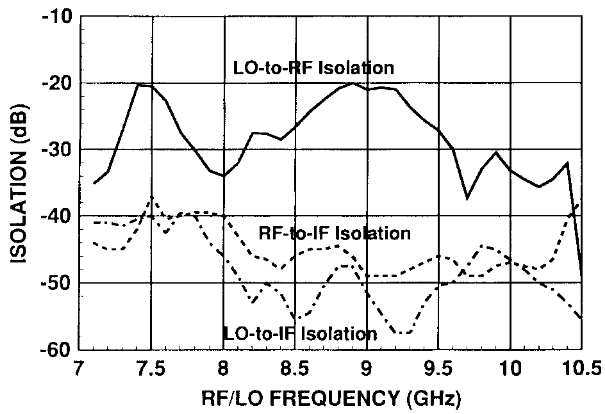

Fig. 6. Measured interport isolations.

dBm, and it does not vary when the LO power level increases from 3.5 to around 11 dBm. Fig. 6 shows the measured interport isolations of the mixer. The LO-to-RF isolation is higher than 20 dB, and the RF-to-IF and LO-to-IF isolations are better than 36 dB. Return losses at the three ports were also measured. When the LO power is 3.5 dBm, the LO port exhibited a return loss of better than 10 dB from 6 to 8 GHz. For the LO frequency of 7 GHz used in the foregoing data, a 20-dB return loss was obtained. The return losses at the RF port were from 11 to 27 dB over 7.1 to 10 GHz and 7 to 11 dB between 10 and 10.5 GHz. The measured IF port's return losses range from 11 and 17 dB over 0.1 to 3.5 GHz. These RF and IF return losses were measured when the mixer diodes were pumped by an LO power of 3.5 dBm at 7 GHz.

IV. Conclusion

A new broad-band uniplanar singly balanced diode mixer has been developed. A conversion loss from 6 to 10 dB (where the RF is swept from 7 to 10.5 GHz and the LO is fixed at 3. 5 dBm at 7 GHz) has been measured for the first-iteration mixer design. More than 20 and 36 dB have also been achieved for the LO-to-RF and LO(RF)-to-IF isolations, respectively. Good return losses at the RF, LO, and IF ports have also been measured. This mixer has the advantages of wide bandwidth, good interport isolations, and simple circuit design. The complete uniplanarity of the mixer is very suitable for low-cost MIC and MMIC manufacturing.

Acknowledgment

The authors wish to thank L. Fan and M. Li for their technical support.

References

[1] S. A. Maas and K. W. Chang, "A broad-band, planar, doubly balanced monolithic Ka-band diode mixer," *IEEE Trans. Microwave Theory Tech.*, vol. 41, pp. 2330–2335, Dec. 1993.

[2] C. H. Ho, L. Fan, and K. Chang, "New uniplanar coplanar waveguide hybrid-ring couplers and magic-T's," *IEEE Trans. Microwave Theory Tech.*, vol. 42, pp. 2440–2448, Dec. 1994.

[3] C. Nguyen and K. Chang, "Design and performance of millimeter-wave end-coupled band-pass filters," *Int. J. Infrared Millim. Waves*, vol. 6, no. 7, pp. 497–509, July 1985.

A Two-Channel Optical Downconverter for Phase Detection

Paul D. Biernacki, Lee T. Nichols, D. G. Enders, Keith J. Williams, and Ronald D. Esman

***Abstract*—Experimental results for a two-channel optical downconverter link operating from 2 to 18 GHz are presented. Using low-noise preamplifiers results in a noise figure (NF) of 8.5–14 dB over the frequency range of 2–18 GHz. For the first time, relative phase measurements between optically downconverted signals have been performed. An in-phase/quadrature phase-measurement technique indicates a phase precision of $\pm 2°$ with as little as -60 dBm radio frequency (RF) received power. Comparing the optical microwave downconverter to an electrical microwave downconverter in terms of phase detection reveals similar performance between the two systems.**

***Index Terms*—Downconversion, fiber-optic link, microwave phase detection, photonics, remoting.**

I. Introduction

Radio frequency (RF) receiving or transmitting systems with multigigahertz bandwidth operation can be more easily and reliably implemented using commercial off-the-shelf optoelectronic components. A downconverting optical system capable of detecting the relative phase of microwave signals offers the possibility of reduced weight, smaller size, and fewer components at the antenna location, while still providing essential microwave functionality. The attractiveness of fiber-optics technology primarily comes from its exceptional RF isolation, its ability to remote microwave signals great distances, and its immunity to electromagnetic interference (EMI). Additionally, use of a lightweight optical modulator as a microwave sensor makes it possible to remotely operate the downconverter since the local oscillator (LO) power can be carried through an optical fiber.

We demonstrate that by using cascaded modulators [1], optical amplification [with an erbium-doped fiber amplifier (EDFA)], balanced detection, and low-noise preamplifiers (DBS #DB96-0625) in our system configuration, a 22-dB noise figure (NF) improvement over our previous system [2]–[3] is achieved. Additionally, the two-channel configuration offers the ability to remotely determine the relative phase of microwave signals. To our knowledge, this is the first ever two-channel optically downconverting detection system. To characterize this system, we used an in-phase/quadrature phase (I/Q) measurement technique. Phase detection is a basic function required for sensitive direction-finding antenna systems utilizing phase interferometry [4]. Phase detection also enables the demodulation of complex microwave communication signals such as phase shift keying (PSK), quadrature PSK (QPSK), or many phase-level PSK (M-ary PSK). Therefore, I/Q measurements give us a good indication of how well these functions could be performed by photonic systems.

Manuscript received October 8, 1997; revised June 25, 1998. This work was supported by the Office of Naval Research.

P. D. Biernacki, L. T. Nichols, K. J. Williams, and R. D. Esman are with the Naval Research Laboratory, Code 5672, Washington, DC 20375 USA.

D. G. Enders is with the Naval Research Laboratory, Code 5726, Washington, DC 20375 USA.

Publisher Item Identifier S 0018-9480(98)08010-7.

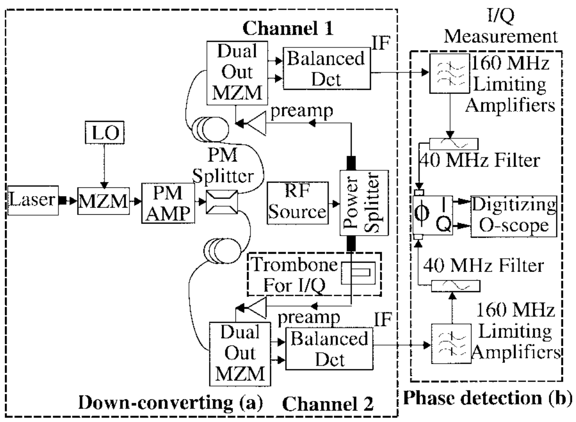

Fig. 1. Two-channel optical downconverting link. Configuration (a) is the basic balanced detection optical link, while configuration (b) adds the phase detection system performance metrology.

II. The Downconverter System

The system configuration used is shown in Fig. 1. It is essentially the same as previously reported [3], except for the insertion of microwave preamplifiers into the system and the construction of two channels to perform phase detection. A 50-mW 1550-nm single-frequency laser is fusion-spliced to a Mach–Zehnder modulator (MZM). After laser light is amplitude modulated by an LO and optically amplified, a directional coupler equally splits the modulated output into two correlated LO signals. Cascading the common LO modulator with the RF modulators in each channel results in mixing between the LO signal and input RF signals. This mixing process produces both downconverted and upconverted output frequencies $|\omega_{\rm RF} \pm \omega_{\rm LO}|$. The balanced detection process [5]–[6] exploits the 180° relative phase shift between the outputs of the dual-output MZM's to electrically cancel EDFA noise as well as relative intensity noise (RIN) in the downconverted intermediate frequency (IF) signal near 160 ± 15 MHz. Since the EDFA noise and RIN are effectively canceled, the only significant noise sources remaining are thermal and shot noise [3]. In order for additive electrical power combination to occur at the detector receivers, the dual outputs need to be path length-matched within each channel. The path length match was experimentally determined to be within 1.8°±1° at a frequency of 2 GHz. Thereby, the IF power at 160 MHz was sufficiently optimized. All modulators were biased at quadrature during testing.

To utilize fiber-optic links (FOL's) successfully in transmit and receive systems, the NF and minimum detectable signal (MDS) should be competitive with those available in microwave systems. If our FOL system is biased at quadrature, the transduction of a microwave signal onto an optical carrier results in an optical power intensity variation given by [1]

$$P_O = \frac{L_{\rm OC} P_{\rm in}}{2} \left(1 - \sin\left[\frac{\pi V}{V_\pi(\omega)} \sin[\omega t + \theta(\omega)]\right]\right) \tag{1}$$

where $P_{\rm in}$ is the input optical power, $L_{\rm OC}$ represents optical coupling and transmission losses, V is the voltage amplitude driving the MZM modulators (nominally into 50 Ω), $V_\pi(\omega)$ is the frequency dependent half-wave voltage of the modulator, and $\theta(\omega)$ contains the phase response of the modulator representing the delay in the modulated signal due to the optical-microwave mismatch. Arranging two modulators in a series cascaded configuration results in mixing of the LO signal (modulator one) and RF input signal (modulator two) giving an output optical power [1]

$$\begin{aligned} P_O &= \frac{L_{\rm OC} P_{\rm in}}{4} \left(1 - \sin[X_{\rm LO} \sin[\omega_{\rm LO} t + \theta(\omega_{\rm LO})]]\right) \\ &\quad \cdot \left(1 - \sin[X_{\rm RF} \sin[\omega_{\rm RF} t + \theta(\omega_{\rm RF})]]\right) \\ X_{\rm RF/LO} &= \frac{\pi V_{\rm RF/LO}}{V_{\pi {\rm RF/LO}}(f_{\rm RF/LO})}. \end{aligned} \tag{2}$$

Expanding this equation in terms of Bessel functions and dropping third-order and higher terms gives

$$\begin{aligned} P_O = \frac{L_{\rm OC} P_{\rm in}}{4} &\cdot \{1 - 2J_1(X_{\rm RF}) \sin[\omega_{\rm RF} t + \theta(\omega_{\rm RF})] \\ &- 2J_1(X_{\rm LO}) \sin[\omega_{\rm LO} t + \theta(\omega_{\rm LO})] \\ &+ 2J_1(X_{\rm RF}) J_1(X_{\rm LO}) \\ &\cdot \cos[(\omega_{\rm LO} - \omega_{\rm RF})t + \theta(\omega_{\rm LO}) - \theta(\omega_{\rm RF})] \\ &- 2J_1(X_{\rm RF}) J_1(X_{\rm LO}) \\ &\cdot \cos[(\omega_{\rm LO} + \omega_{\rm RF})t\theta(\omega_{\rm LO}) + \theta(\omega_{\rm RF})]\} \end{aligned} \tag{3}$$

where J_1 is the Bessel Function of order one. Finally, since R is the responsivity of the detector (Amps/Watt), the total detected photo current is given by $I_{\rm DC} = R L_{\rm OC} P_{\rm in}/4$. Therefore, the IF (160 MHz) electrical power delivered to a 50-Ω load is given by

$$P(\omega_{\rm LO} - \omega_{\rm RF}) = \frac{I_{\rm DC}^2}{2} \left(2J_1(X_{\rm RF}) J_1(X_{\rm LO})\right)^2 \cdot 50. \tag{4}$$

The most salient feature of this equation is the dependence of IF gain on both the optical photo current as well as the LO drive voltage. The measured V_π at RF of the modulators was ~16 V over 2–18 GHz. With RF power $P_{\rm RF}$ launched into the MZM, the conversion gain of the downconverter is

$$G = \frac{P(\omega_{\rm LO} - \omega_{\rm RF})}{P_{\rm RF}} \tag{5}$$

and the NF, given by

$$\mathrm{NF} = \frac{N_{\rm out}}{G \cdot kTB} \tag{6}$$

is measured to be only 11.6 dB. The NF and conversion gain versus frequency for our cascaded modulator FOL system is shown in Fig. 2. The NF varies from 8.5 to 14 dB over the frequency range of 2–18 GHz, while the conversion gain varies from 14.5 to 9.5 dB over the same frequency range. These conversion-gain measurements agree reasonably with the calculated value of ~12 dB. It should be

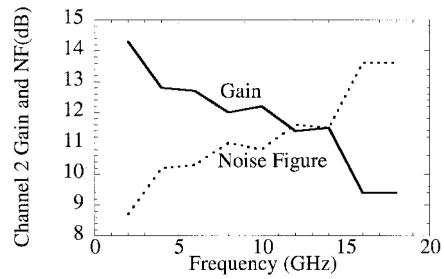

Fig. 2. Gain and NF of the optical downconverter versus frequency.

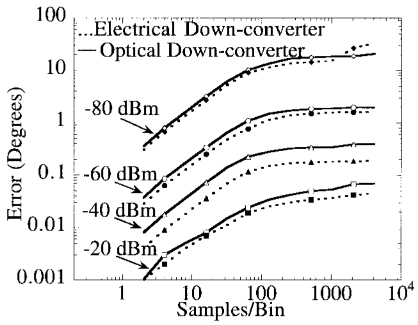

Fig. 3. Mean of standard deviation rms phase error for various RF power levels for the optical and electrical downconverter. The mean of the standard deviation indicates how quickly and confidently phase measurements can be acquired.

noted that the system NF improvement is influenced primarily by the addition of the preamplifiers. In particular, the previous [3] NF of 34 dB was decreased by the gain of the preamplifiers, but increased due to the lower detector current (the partitioning of the available optical power from the EDFA). The 4-dB NF of the preamplifier obviously also added noise to the system. The total average measured current through each arm of each balanced detector was 11 mA, giving a total electrical current of 22 mA for each channel. The current was limited by the available EDFA power of only 150 mW. High current is desirable since it allows near-shot noise-limited performance. In terms of system noise, the equivalent input noise (EIN) was measured to be -162 dBm/Hz. In this system, the LO power level was $+24$ dBm. The MDS can be defined differently depending on the system bandwidth and how a system is used. Here, we set the $\mathrm{MDS} = -87$ dBm since in a 30-MHz bandwidth this gives a signal-to-noise (S/N) ratio of 1. The 1-dB compression dynamic range of this system can be defined as the ratio of the saturation input power of the preamplifiers to the system MDS. With a saturation input of -14 dBm within a 30-MHz bandwidth, the compression dynamic range is approximately 72 dB. Alternatively, by setting the MDS equal to the minimum signal that yields $< \pm 2^\circ$ phase error, the sensitivity of the phase measurements between two channels in the optical downconverter can be measured. These phase measurements indicate a $\pm 2^\circ$ phase resolution down to -60-dBm received RF power and will now be discussed in greater detail.

III. I/Q Phase Measurement Results

Phase detection of incoming signals provides information that can be used for direction-finding, directional tracking, and reception in communication systems using complex modulation. The experimental configuration used for measuring and evaluating system phase-detection performance for the two-channel downconverter is shown in Fig. 1. The I/Q phase detector is fed by limiting amplifiers centered around 160 MHz and a two-channel 1-GSample/s digitizing oscilloscope with an 8-bit analog/digital (A/D) converter processes and evaluates the phase measurements. An electrical microwave downconverter was used for performance comparisons. The electrical downconverter uses two microwave mixers (Avantek DBX186L) fed by the LO and RF inputs to produce the IF signals, which are then detected by the same limiting amplifier/phase-detector combination shown in Fig. 1(b). A greater LO power ($+24$ dBm) was used in the optical downconverter system compared to $+7$ dBm used in the electrical downconverter. An optical modulator requiring a lower switching voltage (V_π) as well as higher optical power could lower the LO needed in the optical modulator without sacrificing performance.

In order to evaluate system performance in terms of phase-detection precision, we measure the statistical properties of the relative phase computed directly from the sampled I and Q values with the expression

$$\Delta\phi = \arctan\left(\frac{Q}{I}\right) \tag{7}$$

where I and Q are the in-phase and quadrature-phase voltage outputs of the phase detector given by

$$I = K\cos(\Delta\phi) \quad Q = K\sin(\Delta\phi) \tag{8}$$

K is a scaling constant and is approximately 31 mV/° for our I/Q phase detectors used. Applying (3) to our dual channel case, as well as noting that the downconverted signal is limited to the 160 ± 15-MHz band, the phase difference $\Delta\phi$ is given by

$$\Delta\phi = \theta(\omega_{\mathrm{RFCH1}}) - \theta(\omega_{\mathrm{RFCH2}}). \tag{9}$$

It was also assumed above that the LO path length-match input is identical for both modulators. To interpret the results, we need to consider the IF bandwidth and noise power within that bandwidth as well as the signal strength. Uncorrelated phase noise at the MZM inputs adds noise to the downconverted IF, which results in an amplitude noise term in the computed phase

$$\Delta\phi = \Delta\phi_o + \phi_n(t) \tag{10}$$

where $\phi_n(t)$ is given by [7]

$$\phi_n(t) = \sqrt{2}\ \tan^{-1}\sqrt{\frac{N_o}{S}} \tag{11}$$

where N_o is the output noise power (-87 dBm in 30 MHz) and S is signal power. The phase noise at the IF of 160 MHz is bandlimited to 30 MHz by the IF filters. Consequently, as acquisition times decrease to less than 66 ns, we expect the root mean square (rms) phase-detection error to decrease because these acquisition times are much shorter than the period of the highest frequency noise components. Restated another way, frequency components of the noise will be coherent in the I/Q system over this short time [7]. Therefore, any amplified noise will now produce an output voltage noise power in the phase detector $[K\Delta\phi_n(t)]$. It should also be noted that there is an additional rms amplitude fluctuation in the output noise that will also add to the phase error [7]. The source of the noise originates primarily from the noncommon elements in our system, such as the preamplifiers and modulators. As long as the limiting amplifiers are operated within their dynamic range, they will not significantly contribute to the noise. Statistical results, shown in Fig. 3, are derived from a total of 4096 samples acquired from the oscilloscope at a sampling rate of 1 GHz. This sample space is then divided into N smaller bins (sample sizes) containing 4096/N samples, as shown in Fig. 4(a). Within each bin, the standard deviation is calculated. Then, the mean of the resulting N values is calculated and plotted, as shown in Fig. 3. The mean of the resulting N values reveals that: 1) the optical downconverter compares well with the electrical downconverter with only a maximum error difference of 0.2° (@ -40 dBm of RF power) and 2) about 100 samples (100 ns) are needed to make an accurate phase measurement. The frequencies used were 8 GHz for the LO and 8.16 GHz for the RF signals (160 MHz IF) for both the optical and electrical downconverter. This mean value can also be computed using (11), and also accounts for the additional amplitude noise contributed fluctuations. The implications of using this method [see Fig. 4 (a)] are faster measurement acquisition with the tradeoff being that the RF power must be known in order for the true phase error to be known. Alternatively, the mean can be computed first within each bin [see Fig. 4 (b)] and then the standard

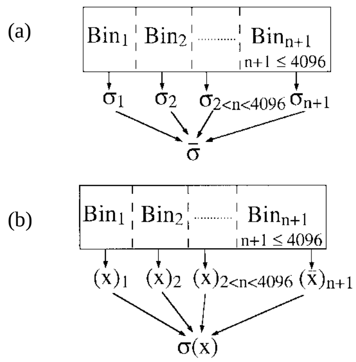

Fig. 4. Mathematical representation of average standard deviation and standard deviation of the average value by partitioning bin sizes. There are a total of 4096 points corresponding to 1 sample point/ns.

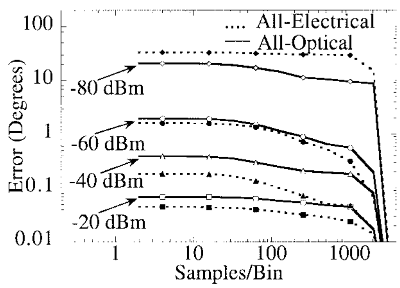

Fig. 5. Standard deviation of mean rms phase error for various RF power levels for the optical and electrical downconverter. The standard deviation of the average indicates the stability of the phase error over certain time intervals.

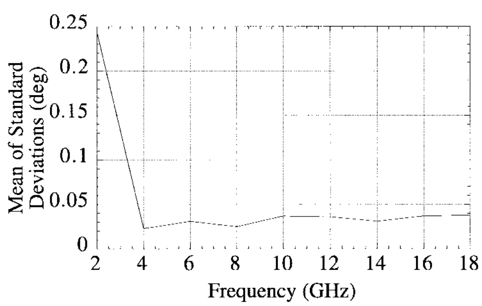

Fig. 6. Mean of the standard deviation of the optical downconverter versus frequency. The RF power level was 0 dBm.

deviation of the resulting N means can be plotted, as shown in Fig. 5. This technique has the advantage of providing a consistent measure of the maximum detection error even for very brief acquisition times. Note that the deviation decreases slightly with increasing bin size as expected for the statistics used. Figs. 3 and 5 show how the statistics change for varying input signal strengths for both the optical and electrical downconverter. From these graphs, the system phase error is determined to be only $\pm 2^\circ$ for as little as -60 dBm input RF power in the optical downconverter. This is nominally the same accuracy provided by the electrical downconverter, although the electrical downconverter used an LO power of $+7$ dBm. The time needed for assuring that the detected phase error is within a certain deterministic error value can be determined by the two statistical error methods used. Particularly, if the received RF power is known, the accuracy of the phase measurement can be determined by setting the acquisition measurement time.

The phase-detection accuracy as a function of RF input frequencies indicated RF and LO spectral independence (see Fig. 6) with less than 0.25° phase deviation over the frequency range of 2–18 GHz. Anomalous behavior at 2 GHz is due to the preamplifier's less than ideal performance at this frequency. The calculated mean of the standard deviation used 128 samples/bin and an RF power level of 0 dBm was input into the phase detector. A more accurate A/D converter could be used to boost the phase sensitivity below -60 dBm, as well as better V_π modulators, greater optical power, and greater LO power.

IV. Conclusion

In conclusion, a two-channel optical downconverter for phase detection was constructed and tested. Use of low-noise microwave preamplifiers resulted in an NF of 8.5–14 dB over a frequency range of 2–18 GHz. I/Q measurements indicate a phase precision within $\pm 2^\circ$ with as little as -60 dBm RF received power. Comparing the optical downconverter to an electrical microwave downconverter in terms of phase detection reveals similar performance between the two systems, except that with the modulators used, the optical downconverter requires more LO power.

Acknowledgment

The authors thank Dr. P. J. Matthews, who provided much assistance in the area of systems design and component assembly.

References

[1] G. K. Gopalakrishnan, W. K. Burns, and C. H. Bulmer, "Microwave-optical mixing in $LiNbO_3$ modulators," *IEEE Trans. Microwave Theory Tech.*, vol. 41, pp. 2383–2391, Dec. 1993.
[2] G. K. Gopalakrishnan, R. P. Moeller, W. K. Burns, K. J. Williams, and R. D. Esman, "A low-loss downconverting analog fiber-optic link," *IEEE Trans. Microwave Theory Tech.*, vol. 43, pp. 2318–2323, Sept. 1995.
[3] K. J. Williams, and R. D. Esman, "Optically amplified downconverting link with shot-noise-limited performance," *IEEE Photon. Technol. Lett.*, vol. 8, pp. 148–150, Jan. 1996.
[4] D. D. Vaccaro, *Electronic Warfare Receiving Systems*. Norwood, MA: Artech House, 1993, pp. 235–245.
[5] G. L. Abbas, V. W. S. Chan, and T. K. Lee, "A dual-detector optical heterodyne receiver for local oscillator noise suppression," *J. Lightwave Technol.*, vol. LT-3, pp. 1110–1122, Oct. 1985.
[6] S. Yamashita and T. Okoshi, "Suppression of common-mode beat noise from optical amplifiers using a balanced receiver," *Electron. Lett.*, vol. 28, pp. 21–23, 1992.
[7] W. P. Robins, *Phase Noise in Signal Sources*. Stevenage, U.K.: Peregrinus, 1982, pp. 6–8.