IV. SUMMARY

The dispersion limited performance of 10 Gb/s lightwave systems using an adjustable chirp optical modulator and an analog tapped delay line equalizer has been examined. The improvement in system reach offered by a smart receiver optimized under a minimum bit error ratio criterion has been determined and an equalizer with three taps has been shown to be the best choice from a performance-complexity perspective.

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Heterodyne Detection of 310-Mb/s Quadriphase-Shift Keying Using Fourth-Power Optical Phase-Locked Loop

J. M. Kahn, A. M. Porter, and U. Padan

***Abstract*—We report an experimental synchronous heterodyne receiver for quadriphase-shift keying based upon a fourth-power optical phase-locked loop. Our transmitter and local-oscillator sources are 1.521-μm external-cavity semiconductor lasers, and the modulation is applied using a series combination of two traveling-wave LiNbO$_3$ phase modulators. In detection of 310-Mb/s QPSK, we have achieved a receiver sensitivity of 104 photons/bit. Using the same receiver, we have detected 155-Mb/s BPSK with a sensitivity of 65 photons/bit.**

I. INTRODUCTION

An optical quadriphase-shift keyed (QPSK) system with homodyne or heterodyne detection requires, for a given bit rate, half the optical and electrical bandwidth of its BPSK counterpart [1]. This bandwidth efficiency makes QPSK potentially useful for dense frequency-division-multiplexed or high-bit-rate systems. When chromatic dispersion is significant, QPSK systems are far less sensitive to degradation than BPSK or MSK systems of the same bit rate [2]. Dispersion-induced intersymbol interference (ISI) and dispersion-induced cross-channel interference (XCI) can be mitigated through electronic equalization at the receiver. In a QPSK homodyne receiver this requires complicated baseband cross-cancellation [2]. By contrast, in a QPSK heterodyne receiver it can be achieved by a single passband equalizer [3], making QPSK/heterodyne an attractive combination for long-

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J. M. Kahn is with the Department of Electrical Engineering and Computer Sciences, University of California, Berkeley, CA 94720.

A. M. Porter was with the Department of Electrical Engineering and Computer Sciences, University of California, Berkeley, CA 94720. She is now with GTE Laboratories, Waltham, MA 02254.

U. Padan is with the State of Israel, Ministry of Defense, RAFAEL, Haifa, Israel.

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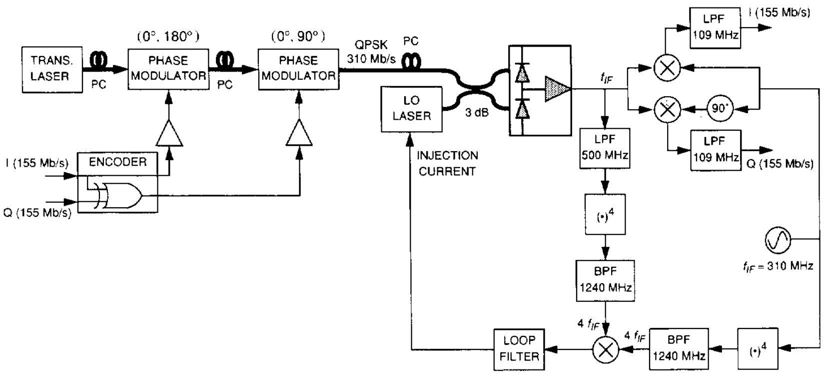

Fig. 1. Configuration of experimental 310-Mb/s QPSK heterodyne system using a fourth-power optical phase-locked loop.

span, high-bit-rate systems. Heterodyne receivers also use simpler optical hardware than homodyne receivers for QPSK. Heterodyne receivers combine the signal and local-oscillator (LO) light in a 3-dB coupler and require only a single balanced detector, while homodyne receivers require a 90° optical hybrid and two balanced detectors. Unlike BPSK, there is no theoretical sensitivity advantage of QPSK homodyne over QPSK heterodyne [1]; both techniques ideally can achieve a sensitivity of 18 photons/bit. It should be noted that coherent detection of QPSK poses stringent laser-linewidth requirements [4]–[6].

Derr has used He–Ne lasers to demonstrate a 100-Mb/s QPSK system using an intradyne optical receiver, achieving a sensitivity of 529 photons/bit [7]. Norimatsu *et al.* have employed external-cavity semiconductor lasers and a homodyne decision-directed receiver to obtain a sensitivity of 1136 photons/bit at 8 Gb/s [8]. Despite the potential advantages of QPSK heterodyne, there have been no reported demonstrations of this technique to date. In this letter, we report a 310-Mb/s QPSK system using external-cavity semiconductor lasers synchronized by a heterodyne fourth-power optical PLL (OPLL).

II. SYSTEM DESIGN

Our experimental QPSK heterodyne system is illustrated in Fig. 1. The transmitter and LO lasers are external-cavity InGaAsP devices [9] having a beat linewidth of 2.4 kHz. The LO laser is equipped with a two-electrode chip to provide flat frequency-modulation response for frequencies from dc to several tens of megahertz, allowing it to serve as an optical current-controlled oscillator.

We perform 310-Mb/s NRZ QPSK modulation using the series combination of two $LiNbO_3$ traveling-wave phase modulators [5]. The first modulator is driven with a phase shift of either 0° or 180°, and the second by a phase shift of either 0° or 90°. The two drive signals are derived by an encoder from the input in-phase (I) and quadrature (Q) data. In this experiment, the I and Q data are delayed copies of the same pseudorandom bit sequence (PRBS); a one-bit delay is sufficient to achieve the maximum possible decorrelation.

The received signal polarization is matched manually to that of the LO. The two signals are combined in a 3-dB coupler and are detected by a balanced InGaAs p-i-n/HEMT receiver having a quantum efficiency of 77%, a 3-dB bandwidth of 100 kHz-10.1 GHz, and an average equivalent input noise current density (one-sided) of 10.8 $pA/Hz^{1/2}$. The detected LO photocurrent of 1.8 mA induces a shot noise density (one-sided) of 24.0 $pA/Hz^{1/2}$.

The balanced-receiver output is set to an intermediate frequency (IF) of 310 MHz by phase-locking it to a 310-MHz reference oscillator, as now described. Part of the IF signal is low-pass-filtered by a 500-MHz, 3-pole Chebyshev filter to band-limit the noise. Using a passive frequency doubler, this signal is raised to the fourth power to wipe off the QPSK modulation, and is then band-pass filtered near $4f_{IF} = 1240$ MHz to remove unwanted signals at other harmonics of f_{IF}. Using a double-balanced mixer, the phase of this $4f_{IF}$ carrier is compared to the phase of the fourth power of the reference-oscillator signal. The phase-detector output is passed to a proportional-plus-integral loop filter, which controls the injection current to the LO laser. This forms a second-order OPLL characterized by a damping constant $\zeta = 0.66$, and a natural frequency ω_n, which is varied. The loop dc gain $K_v \cong 10^9$ Hz is sufficient to render negligible any static phase error, in view of the ~ 100 kHz/min free-running frequency drift between transmitter and LO lasers [9].

Another portion of the IF signal is demodulated by the reference oscillator and by a 90°-phase-shifted version of the reference to yield, respectively, the I and Q data streams. These are filtered by 5-pole Bessel filters having 3-dB bandwidths of 109 MHz.

III. PHASE ERROR MEASUREMENT

An important OPLL figure of merit is σ_ϵ, the standard deviation of the phase error $\epsilon(t)$ between the IF and reference signals at f_{IF}. In a QPSK system, a value of $\sigma_\epsilon = 2.97°$ is expected to induce a 0.5-dB penalty at a

BER of 10^{-9} [10]. According to theory [4]–[6], σ_ϵ depends on system and OPLL parameters through

$$\sigma_\epsilon^2 = \int_{-\infty}^{\infty} S_\phi(f)|1 - H(j2\pi f)|^2 df + \int_{-\infty}^{\infty} S_w(f)|H(j2\pi f)|^2 df \quad (1)$$

where $S_\phi(f)$ (rad²/Hz) $= \Delta\nu_{\text{beat}}/(2\pi f^2)$ is the (two-sided) power spectral density (PSD) of the phase noise and $S_w(f)$ (rad²/Hz) $= \eta T/M$ is the (two-sided) PSD of the effective shot noise $w(t)$ [6], which is essentially white within the bandwidth of interest. Here, $\eta \cong 0.55$ is the noise factor for the fourth-power loop [6], T is the symbol interval, and M is the number of detected photons/symbol. The closed-loop transfer function of a second-order OPLL with loop filter of the form $F(s) = (1 + s\tau_2)/(s\tau_1)$ is given by

$$H(s) = \frac{(\omega_n^2 + 2\varsigma\omega_n)G(s)}{(\omega_n^2 + 2\varsigma\omega_n s)G(s) + s^2} \quad (2)$$

where $G(s) \neq 1$ represents non-ideality within the loop (including propagation delay).

In this experiment, we cannot directly measure the phase error $\epsilon(t)$, but we can record the loop-filter input signal $e(t) = \epsilon(t) + w(t)$. Theory predicts the PSD of $e(t)$ to be

$$S_e(f) = \left[S_\phi(f) + S_w(f)\right]|1 - H(j2\pi f)|^2. \quad (3)$$

The variance of the true phase error can be estimated from this PSD using

$$\sigma_{\epsilon,\text{meas}}^2 = 2\int_0^{B_{\text{meas}}} \{S_e(f) - S_w(f)\} df \quad (4)$$

where we ignore any correlation between $\epsilon(t)$ and $w(t)$ under the assumption that $w(t)$ is white and that the loop contains a pure delay in $G(s)$. The measurement bandwidth B_{meas} must be sufficient to include all frequencies where the integrand of (4) is nonzero. We used a spectrum analyzer to measure $2S_e(f)$, and Fig. 2(a) presents spectra recorded with $\zeta = 0.66$, $\Delta\nu_{\text{beat}} = 2.4$ kHz, $M = 112$ detected photons/symbol, and several different values of ω_n. These spectra were recorded without data modulation present. Fig. 2(b) shows the PSD $2S_e(f)$ predicted by (3), using $G(s) = e^{-s\tau_d}/(1 + s\tau_p)$, with the delay $\tau_d = 20$ ns and the pole $\tau_p = 30$ ns being chosen based upon characterization of the loop components. Fig. 2(a) and (b) exhibit reasonable agreement between the positions of the peaks in $2S_e(f)$, but the shapes of the curves differ between experiment and theory.

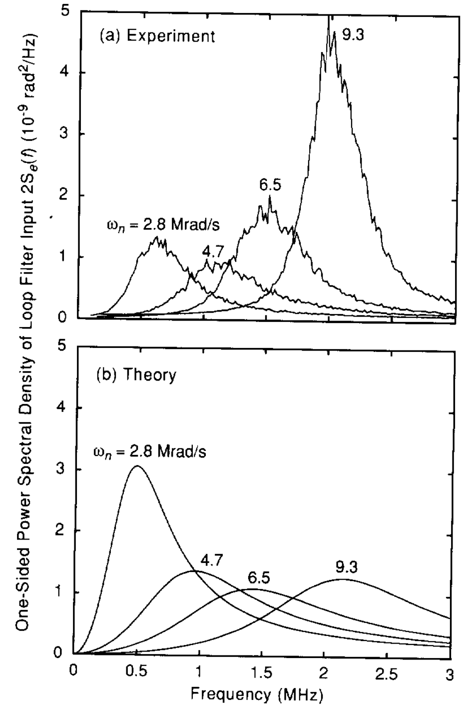

Fig. 2. One-sided power spectral density of loop-filter input signal, $2S_e(f)$: (a) experimentally measured without data modulation; (b) theoretically predicted.

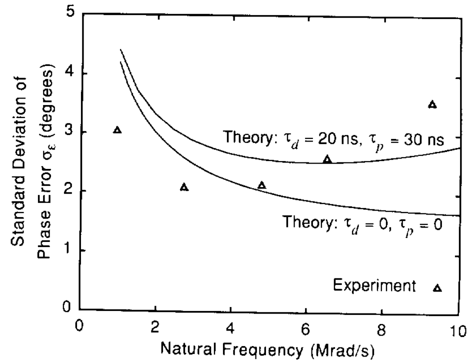

Fig. 3. Standard deviation of phase error: experimentally measured without data modulation and theoretically predicted, with and without inclusion of delay and extra pole in loop.

Fig. 3 presents the standard deviation of phase error σ_ϵ, as derived from experiment using (4). The measurement bandwidth was $B_{\text{meas}} = 5$ MHz, and $S_w(f)$ was estimated from the high-frequency asymptote of $S_e(f)$. No modulated data were present. We found a minimum phase error of $\sigma_{\epsilon,\min} = 2.1°$ at an optimum natural frequency $\omega_{n,\text{opt}} = 2.8$ Mrad/s. Our experimental results are compared in Fig. 3 to the predictions of (1), considering both $\tau_d = \tau_p = 0$ and $\tau_d = 20$ ns, $\tau_p = 30$ ns in the above expression for $G(s)$. For the latter, more realistic case, the theoretical predictions of σ_ϵ are of the same order as the experimental values and show the same general trend. With this nonzero choice of τ_d and τ_p, theory predicts $\sigma_{\epsilon,\min} = 2.5°$ at $\omega_{n,\text{opt}} = 6.2$ Mrad/s. Disagreement between theory and experiment in Figs. 2 and 3 may arise, in part, because of neglect of additional poles and zeros in the choice of $G(s)$.

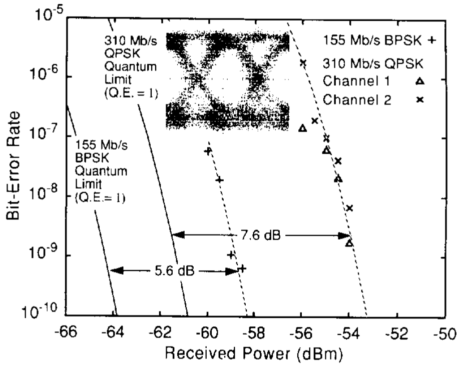

Fig. 4. Bit-error rate versus received optical power for reception of 155-Mb/s BPSK and 310-Mb/s QPSK using a $2^{15} - 1$ PRBS. Inset: eye pattern of channel 1 demodulated from a 310-Mb/s QPSK signal.

IV. System Performance

Fig. 4 presents the bit-error rate (BER) performance for 155-Mb/s BPSK and 310-Mb/s QPSK, measured using $\omega_n = 4.7$ Mrad/s, which was found to yield the lowest BER. For both modulation formats, the BER was found to be independent of the PRBS for word lengths between $2^7 - 1$ and $2^{15} - 1$, indicating that the fourth-power OPLL is effective in performing modulation wipe-off for BPSK and QPSK.

For BPSK, the receiver sensitivity (10^{-9} BER) was found to be -58.8 dBm or 65 photons/bit, which is 5.6 dB away from the quantum limit of -64.4 dBm. Estimated detected penalties include: optical coupling losses (1.3 dB), incomplete suppression of receiver thermal noise (1.0 dB), OPLL phase error (negligible), use of a 109-MHz, 5-pole Bessel filter in place of the matched filter (1.5 dB), and ISI caused by imperfect frequency response of the transmitter and receiver (the remaining penalty of 1.8 db).

In reception of QPSK, we found the sensitivity to be -53.8 dBm or 104 photons/bit. This lies 7.6 dB from the shot-noise limit of -61.4 dBm. For QPSK, we estimate that most of the detection penalties are close to those for BPSK, with two exceptions: OPLL phase error (0.4 dB) and ISI and XCI arising from transmitter and receiver imperfections (the remaining penalty of 3.4 dB).

V. Conclusion

We have demonstrated a synchronous heterodyne receiver for QPSK based upon a fourth-power OPLL. At a bit rate of 310 Mb/s, we achieved a receiver sensitivity of 104 photons/bit.

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