PERFORMANCE EVALUATION OF AN OOK COHERENCE MULTIPLEX RECEIVER BASED ON 4×4 PHASE DIVERSITY DETECTION

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ABSTRACT

Coherence multiplexing (CM) is a relatively cheap method for multiplexing multiple channels onto a fiber-optic link. Most previously published CM systems use either single-ended or balanced detection, which requires phase-locking for a stable output signal. Recently, we proposed a scheme in which the receiver output signal is stabilized by a phase diversity network. In this paper, the interferometric noise-limited performance of a CM receiver based on 4×4 phase diversity detection is analyzed and compared to a receiver based on balanced detection. The results will show that phase diversity introduces only a slight degradation of the transmission capacity of the system.

1 COHERENCE MULTIPLEXING (CM)

Coherence multiplexing (CM) is a form of CDMA which is especially suited for optical transmission [1]. Although CM cannot compete with WDM as far as transmission rates are concerned, it might be favourable from a cost point of view, as broadband light sources (for instance LEDs) and only simple components are required for crosstalk-free transmission. Therefore, CM is particularly suitable for small-scale networks like LANs and access networks. CM is based on the principle of distinguishing between coherent and incoherent mixing of lightwaves, as illustrated in Fig. 1. The idea is that each transmitter launches both a BPSK modulated version and an unmodulated version of the broadband carrier into the common fiber. These two carriers are made mutually incoherent (uncorrelated) by delaying them with respect to each other by a timeshift T_{tr} which is much larger than the coherence time of the source τ_{c}. In the receiver, the correct channel can be selected by mixing the received signal with the same signal being delayed by a timeshift T_{re}. Mixing is performed by a balanced detector, which consists of a 2×2 coupler and two photodiodes. If $T_{\mathrm{re}} = T_{\mathrm{tr}}$ then the lightwave taking the upper path in the transmitter and the lower path in the receiver mixes coherently with the lightwave taking the lower path in the transmitter and the upper path in the receiver, as their mutual time delay is zero. Since one of these lightwaves is modulated, the mixing product is an antipodal baseband signal proportional to the modulating signal $m(t)$. If the delays of all the transmitters are spaced apart much more than the coherence time of the sources, then it can be simply verified that all the lightwaves from interfering transmitters mix incoherently, resulting in broadband interferometric noise. Moreover, the coherent mixing term suffers from source intensity

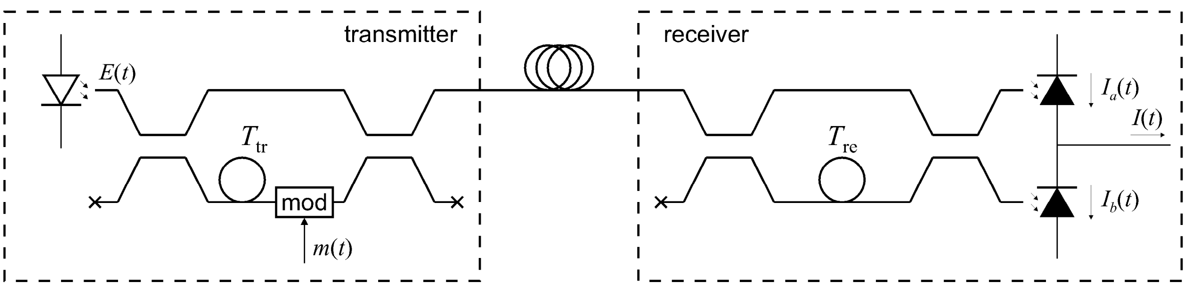

Figure 1: A coherence multiplex system with one transmitter and one receiver

noise. Both noise currents have a very broad power spectral density function which is proportional to the square of the received power. The power spectral densities of shot noise and thermal receiver noise are proportional to and independent of the received power, respectively. Hence, for large received powers, they can be neglected. Assuming that the complex envelope of the field emitted by the broadband source can be modelled as a circular complex Gaussian process with a Gaussian spectral profile, and that all the fields in the receiver have matched polarization states, the signal-to-noise ratio after matched filtering can be shown to be [1]

$$SNR = \frac{2}{4M^2+2M+1}\frac{T_\mathrm{b}}{\tau_\mathrm{c}} = \frac{\sqrt{\frac{2\pi}{\ln 2}}}{4M^2+2M+1}\frac{\Delta f}{R_\mathrm{b}} \tag{1}$$

where M is the number of active users, T_b is the bit-time of the modulating signal $m(t)$, Δf is the 3 dB linewidth of the source and R_b is the bitrate. As the matched filter bandwidth will generally be much smaller than the bandwidth of the noise, the decision samples can be assumed to be Gaussian distributed, so the bit error rate is given by

$$\mathrm{P_e} \approx \frac{1}{\sqrt{2\pi}}\int_{\sqrt{SNR}}^{\infty} \exp\left(-\frac{x^2}{2}\right)\,\mathrm{d}x \equiv \mathrm{Q}\left(\sqrt{SNR}\right) \tag{2}$$

2 STABILITY ISSUE

The problem with this balanced receiver is that the coherent lightwaves have to be mixed exactly in phase in order to achieve constructive interference. Therefore, even a small drift in the delay in either the transmitter or receiver will introduce a drop in the amplitude of the output signal, by a factor $\cos\Big(2\pi f_\mathrm{c}(T_\mathrm{tr}-T_\mathrm{re})\Big)$, where f_c is the carrier frequency of the light [2]. In most publications on CM, this practical problem is solved by phase-locking the coherent lightwaves. This can be done by applying a feedback loop from the detected signal to the delay T_re, which adapts the refractive index, for example by means of the thermo-optic or electro-optic effect [3]. Recently, we proposed how stabilized detection without phase-locking can be performed using a 3×3 phase diversity scheme [2]. Phase diversity is a detection method which has been successfully applied in numerous types of coherent optical systems [4]. In this paper, a coherence multiplex receiver based on a 4×4 phase diversity network is introduced. Capacity bounds of the resulting system will be given, and compared to the case in which balanced detection was used.

3 PERFORMANCE OF AN OOK 4×4 PHASE DIVERSITY RECEIVER

Consider the coherence multiplex receiver based on a 4×4 phase diversity network, as given in Fig. 2. In this receiver, the balanced mixer is replaced by a 4×4 optical hybrid (for instance a multimode interference coupler) and two differential pairs of photodiodes. The geometry of the hybrid is assumed to be such that the power of the input signals is equally divided over the output ports, and moreover, that the mixing phases are given by ϕ, $\phi+\frac{\pi}{2}$, $\phi-\frac{\pi}{2}$ and $\phi+\pi$, where ϕ is an arbitrary phase that depends both on the phase difference between the input signals (which changes with $|T_\mathrm{tr}-T_\mathrm{re}|$) and on the absolute phase transfer of the hybrid. As a result,

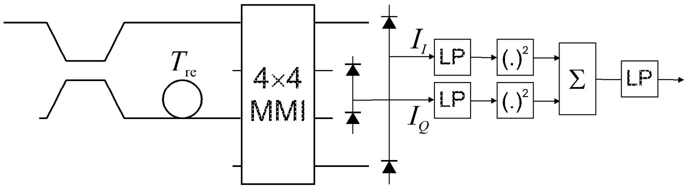

Figure 2: A coherence multiplex received based on a 4x4 phase diversity network

the output currents of the differential photodiode pairs $I_{\mathrm{I}}(t)$ and $I_{\mathrm{Q}}(t)$ both consist of desired signals $I_{\mathrm{I,s}}(t)$ and $I_{\mathrm{Q,s}}(t)$, respectively, and broadband noise $I_{\mathrm{I,n}}(t)$ and $I_{\mathrm{Q,n}}(t)$, respectively. It can be shown that the desired signals are given by $m(t)A\cos\phi$ and $m(t)A\sin\phi$, respectively, where the amplitude A incorporates transmitted power, splitting and coupling losses, and the responsivity of the photodiodes. To minimize the degradation due to the noise, both signals are first matched filtered by the low-pass filters, before they are squared and added, which results in an output signal that does not depend on ϕ, as the difference in mixing phase is $\frac{\pi}{2}$ radians. Therefore, it is not possible to detect BPSK modulated signals using this phase diversity receiver. This can be solved by replacing the BPSK modulation in Fig. 1 by on-off keying (OOK), which can be performed by directly modulating the source. Since transmitting a '0'-bit incorporates switching off the source, the variance of the interferometric noise depends on the number of users that are simultaneously transmitting a '1'-bit; therefore the signal-to-noise ratio can be considered as a random process $\gamma(t)$ which is varying with time. The distribution of $\gamma(t)$ can be determined by assuming that the received signals from both the matched transmitter and the interfering transmitters are bit-synchronized, such that for a particular bit k, the signal-to-noise ratio is a discrete random variable γ_k that can take one out of M possible values. All transmitters are assumed to transmit '0' and '1' bits with an equal probability of $\frac{1}{2}$, so the number of interfering transmitters N_k that is transmitting a '1'-bit, is binomially distributed. As a result, the average bit error probability can be calculated as follows:

$$\mathrm{P_e} \approx \sum_{n=0}^{M-1} \mathrm{P}_{\mathrm{e}|N_k=n}\mathrm{P}[N_k=n] = \left(\frac{1}{2}\right)^{M-1} \cdot \sum_{n=0}^{M-1} \mathrm{P}_{\mathrm{e}|N_k=n} \begin{pmatrix} M-1 \\ n \end{pmatrix} \tag{3}$$

Now consider the two mixing currents $I_{\mathrm{I}}(t)$ and $I_{\mathrm{Q}}(t)$. Both currents consist of a desired part $I_{\mathrm{I,s}}(t)$ and $I_{\mathrm{Q,s}}(t)$, respectively, interferometric noise $I_{\mathrm{I,in}}(t)$ and $I_{\mathrm{Q,in}}(t)$, respectively, and source intensity noise, $I_{\mathrm{I,sin}}(t)$ and $I_{\mathrm{Q,sin}}(t)$, respectively. As the complex envelopes of the electrical fields corresponding to the source lightwaves are considered as circular complex Gaussian processes with a spectral profile that is symmetric around the carrier frequency, the quadrature components of the electrical field are independent. Therefore, the interferometric noise components $I_{\mathrm{I,in}}(t)$ and $I_{\mathrm{Q,in}}(t)$ can be assumed to be uncorrelated. Now let I_k and Q_k denote the output samples of the matched filters for a particular bit k. Then it can be shown that I_k and Q_k contain information terms $I_{\mathrm{s},k} \equiv \mathrm{E}[I_k] = m_k A \cdot T_{\mathrm{b}} \cos\phi$ and $Q_{\mathrm{s},k} \equiv \mathrm{E}[Q_k] = m_k A \cdot T_{\mathrm{b}} \sin\phi$, respectively, and source intensity noise terms $I_{\mathrm{sin},k} = A_{\mathrm{sin},k}\cos\phi$ and $Q_{\mathrm{sin},k} = A_{\mathrm{sin},k}\sin\phi$, respectively, where $A_{\mathrm{sin},k}$ is zero-mean Gaussian with variance $\sigma^2_{\mathrm{sin},k} = m_k A^2 T_{\mathrm{b}}\tau_{\mathrm{c}}$. Finally, the interferometric noise terms $I_{\mathrm{in},k}$ and $Q_{\mathrm{in},k}$ are mutually uncorrelated zero-mean Gaussian distributed with equal variance $\sigma^2_{\mathrm{in},k} = \frac{1}{2}\left(4N_k^2 + (8m_k+2)N_k + 5m_k\right)A^2 T_{\mathrm{b}}\tau_{\mathrm{c}}$, where m_k is the corresponding information bit (which is either 0 or 1). As $A_{\mathrm{sin},k}$, $I_{\mathrm{in},k}$ and $Q_{\mathrm{in},k}$ are jointly Gaussian and mutually uncorrelated, they are independent. Now let S_k denote the output signal of the phase diversity receiver at the optimum sampling instant, so S_k is given by

$$S_k = I_k^2 + Q_k^2 = \Big((m_k A \cdot T_{\mathrm{b}} + A_{\mathrm{sin},k})\cos\phi + I_{\mathrm{in},k}\Big)^2 + \Big((m_k A \cdot T_{\mathrm{b}} + A_{\mathrm{sin},k})\sin\phi + Q_{\mathrm{in},k}\Big)^2 \tag{4}$$

When a binary '0' is transmitted, S_k can be proven to have a probability density function f_{S_k} which is central chi-squared with two degrees of freedom. If the detection threshold is denoted by s_{th}, then the probability of error for a binary '0' can be proven to be

$$\mathrm{P}_{\mathrm{e}|m_k=0,N_k=n} = \int_{s_{\mathrm{th}}}^{\infty} f_{S_k|m_k=0,N_k=n}(s)\,\mathrm{d}s = \exp\left(-\frac{s_{\mathrm{th}}}{(4n^2+2n)\,A^2 T_{\mathrm{b}}\tau_{\mathrm{c}}}\right) \tag{5}$$

For large signal-to-noise ratios, the output signal S_k in case of a binary '1' can be proven to be approximately non-central chi-square distributed with one degree of freedom, and the probability of a bit error in that case is

$$\mathrm{P}_{\mathrm{e}|m_k=1,N_k=n} = \int_0^{s_{\mathrm{th}}} f_{S_k|m_k=1,N_k=n}(s)\,\mathrm{d}s \approx \mathrm{Q}\left(\frac{A\cdot T_{\mathrm{b}} - \sqrt{s_{\mathrm{th}}}}{A\sqrt{\dfrac{4n^2+10n+7}{2}T_{\mathrm{b}}\tau_{\mathrm{c}}}}\right) \tag{6}$$

As identical symbol probabilities are assumed, the total bit error probability $\mathrm{P}_{\mathrm{e}|N_k=n}$ can be minimized with respect to s_{th} by setting $s_{\mathrm{th}} \approx \left(\frac{1}{2}A \cdot T_{\mathrm{b}}\right)^2$. Substituting the resulting conditional error probability in (3) gives

$$\mathrm{P_e} \approx \left(\frac{1}{2}\right)^M \cdot \sum_{n=0}^{M-1} \binom{M-1}{n} \left(\exp\left(-\frac{T_{\mathrm{b}}}{4\,(4n^2+2n)\,\tau_{\mathrm{c}}}\right) + \mathrm{Q}\left(\sqrt{\frac{T_{\mathrm{b}}}{2\,(4n^2+10n+7)\,\tau_{\mathrm{c}}}}\right)\right) \tag{7}$$

which can be numerically evaluated. Now we will compare this result to the average probability of bit error in case of balanced detection. To make a fair comparison, it should be assumed that OOK is also used in this case. Applying a similar reasoning as in the case of the phase diversity detector, the average bit error probability for balanced detection of OOK is easily found to be

$$\mathrm{P_e} \approx \left(\frac{1}{2}\right)^M \cdot \sum_{n=0}^{M-1} \binom{M-1}{n} \left(\mathrm{Q}\left(\sqrt{\frac{T_{\mathrm{b}}}{2\,(4n^2+2n)\,\tau_{\mathrm{c}}}}\right) + \mathrm{Q}\left(\sqrt{\frac{T_{\mathrm{b}}}{2\,(4n^2+10n+7)\,\tau_{\mathrm{c}}}}\right)\right) \tag{8}$$

Fig. 3 shows the network capacities one can obtain for a given number of users M at an average bit-error rate of 10^{-9}. It can be verified that these results also apply to the 3×3 phase diversity receiver.

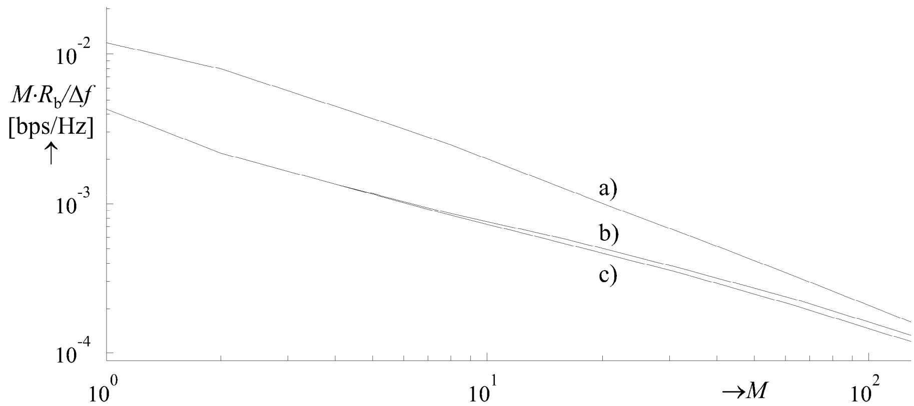

Figure 3: Network capacity over linewidth at a bit error rate $\mathrm{P_e} = 10^{-9}$
a) BPSK modulation and balanced detection
b) OOK modulation and balanced detection
c) OOK modulation and phase diversity detection

4 CONCLUSION

A phase diversity network in a CM receiver provides stabilization of the output signal without requiring phase-locking. At low bit error rates, phase diversity detection introduces only a slight capacity degradation compared to balanced detection when OOK modulation is used.

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