eqn. 3 is $Z_{MF}[n] + Z[n]$, where $Z_{MF}[n] = \langle r[n], s_1 \rangle$ and $Z[n] = \langle r[n], x_1[n-1] \rangle$. So, if $A_1[n]$ is available, $\hat{\alpha} = A_1[n]\hat{b}_1[n] = A_1[n]\mathrm{sgn}(Z_{MF}[n] + Z[n])$. $A_1$ is a constant, and assuming that $x_1$ is chosen properly, $A_1[n] = A_1[n-1] \cong |Z_{MF}[n-1] + Z[n-1]|$. Then the estimate of $A_1[n]$ and $b_1[n]$ becomes $\hat{A}_1[n] = \mathrm{abs}(Z_{MF}[n-1] + Z[n-1])$, $\hat{b}_1[n] = \mathrm{sgn}(Z_{MF}[n] + Z[n])$, and $Z_{MF}$ does not depend on $\hat{b}_1[n]$, $\hat{A}_1[n]$. Finally, the adaptation rule for minimising eqn. 3 is as follows:

$$x_1[n] = x_1[n-1] - \mu(Z_{MF}[n] - \hat{\alpha} + Z[n])(r[n] - Z_{MF}[n]s_1^T) \quad (4)$$

where $\hat{\alpha} = \mathrm{abs}(Z_{MF}[n-1] + Z[n-1])\mathrm{sgn}(Z_{MF}[n] + Z[n])$, $x_1[0] = Z_{MF}[0] = Z[0] = 0$ and $\mu$ is the step size. Eqn. 4 is similar to a decision-directed unconstrained MOE. In [1], switching to decision-directed mode using conventional LMS was mentioned. However, it needs a typically unknown time to switch, and improper choice of the time may not guarantee convergence.

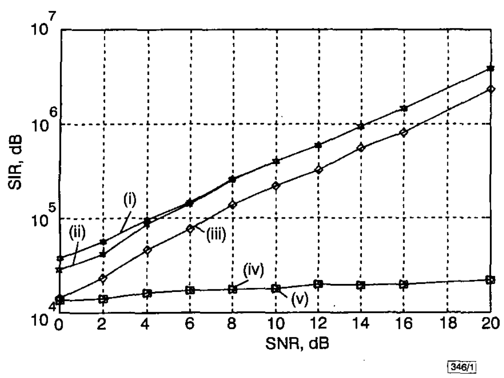


**Fig. 1** *SIR against SNR for user of interest*

(i) trained receiver
(ii) proposed receiver with known amplitude
(iii) proposed receiver with unknown amplitude
(iv) unconstrained MOE
(v) Griffith's algorithm

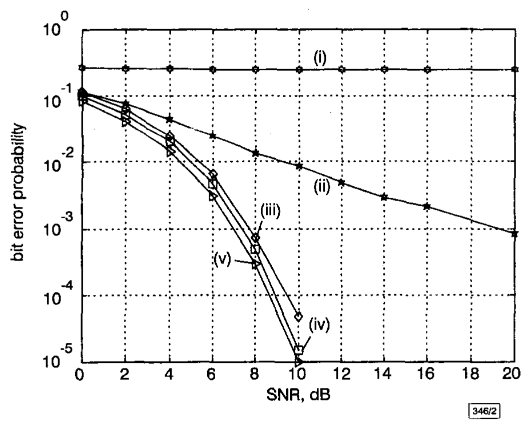


**Fig. 2** *BER against SNR for user of interest*

(i) matched filter
(ii) unconstrained MOE
(iii) proposed receiver with unknown amplitude
(iv) proposed receiver with known amplitude
(v) decorrelator

*Numerical results:* In the following numerical examples, the performances of the proposed detector in an AWGN are examined. Each user is assigned a unique signature sequence from a set of Gold codes of length $N = 31$ with binary phase shift keying (BPSK) modulation. The receiver samples the received signal four times per chip and $\mu = 0.0001$. The total number of active users is 21, and the power of all users is 10dB greater than the desired user's power. 30 simulation results are averaged. In Fig. 1, we show the SIR performance of the proposed detector. We define the SIR as $SIR(\mathrm{dB}) = 10\log_{10} \{(\langle s_1, c_1 \rangle)^2 / \Sigma_{k=2}^{K} (\langle s_k, c_1 \rangle)^2\}$. The decorrelator has a infinite SIR not shown in Fig. 1. To measure the SIR, we stop the adaptation after receiving 10000 symbols and make comparisons. As we can see, the SIR performance of the proposed receiver (unknown amplitude) is better than that for the unconstrained MOE and slightly worse than the trained receiver's performance (eqn. 4 with known bit and amplitude). The performance of Griffith's algorithm and the MOE is almost the same. Fig. 2 illustrates the BER performance. The performance of the decorrelator is slightly better than that of the proposed receiver, and there is no error floor in the high SNR region for the proposed receiver. Also the performance of the proposed receiver with known amplitude ($A_1$ is known) is almost identical to that of the trained receiver. We simulated with random codes and results not shown in this Letter were slightly worse than the Gold code results. Moreover, more time is needed for convergence.

*Conclusion:* In this Letter, we have proposed an adaptive decorrelating detector for CDMA systems. With low computational complexity, the BER performance of the proposed detector is almost identical to that of a decorrelator that utilises the assumed known signatures of the interferers. The only parameters for constructing the receiver are the signature sequence and symbol timing of the desired user.


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## High-frequency, low-crosstalk modulator arrays based on FTC polymer systems

A.H. Udupa, H. Erlig, B. Tsap, Y. Chang, D. Chang, H.R. Fetterman, Hua Zhang, Sang-Shin Lee, Fang Wang, W.H. Steier and L.R. Dalton

A novel electro-optic polymer, Furan TetraCyano Indane (FTC), has been utilised in the fabrication of low $V_\pi$, high-speed, travelling-wave Mach-Zehnder modulator arrays. The modulators were realised with microstrip transmission lines that were optimised resulting in a line loss of 3.64dB/cm at 40GHz. Mach-Zehnder modulators based on this design with 1.62cm interaction length resulted in devices with $V_\pi < 5$V and a reasonably flat frequency response up to 40GHz. Electrical crosstalk measurements were also conducted on adjacent modulators separated by 400μm in a modulator array. Without the implementation of any crosstalk reduction techniques, the measurement results indicated a crosstalk level of < –40dB in the frequency band between 0 and 40GHz. These impressive results were a consequence of the low dielectric constant of the polymer material at microwave frequencies and of the small lateral dimensions of the microstrip lines.

High-density arrays of high-speed modulators with low half-wave voltage ($V_\pi$) are of importance in a variety of applications, which include optical interconnects. The suitability of polymers for realising high-speed, travelling-wave modulators has already been demonstrated [1] utilising a PUR-DR19 polymer system. Furthermore, the electro-optic (EO) coefficient of polymer materials has increased significantly over the past few years, causing a drastic drop in $V_\pi$. The obtainable EO coefficient, $r_{33}$, has increased from 15pm/V for PUR-DR19 [2] to 45pm/V for FTC (Furan Tetra-Cyano Indane) [3]. High speeds and improved $r_{33}$ have enabled polymer materials to become competitive alternatives to $LiNbO_3$. However, in addition to higher individual performance, devices in modulator arrays are required to exhibit low crosstalk. In some situations, isolation > –20dB is required [4]. $LiNbO_3$ modulator arrays with –50dB crosstalk > 0–20GHz have been demonstrated [4]. However, in this case, the modulators were separated by 1.5mm and microwave shields and septa were used in order to obtain low crosstalk levels. In this Letter, we present the characterisation of an FTC Mach-Zehnder modulator and subsequent crosstalk measurements of a modulator array. For devices separated by 400μm, the crosstalk was < –50dB over 0-20GHz and < –40dB over 20–40GHz. These results were impressive considering that no crosstalk reduction techniques, such as buried microstrip lines [5], were implemented.

The modulators reported in this work were based on FTC, a new EO polymer. FTC has an $r_{33}$ of 45–57pm/V at a wavelength of 1.06μm. It exhibits an absorption maximum at 635nm and shows an intrinsic loss of ~ 0.75dB/cm at 1.3μm. FTC-based Mach-Zehnder modulators were fabricated on silicon substrates using the fabrication procedure outlined in [2]. For the work reported in [2], flexible Mylar substrates were used, which for this work were replaced by silicon rigid ones. Microstrip transmission lines of 50Ω impedance drove the modulators. The strip width was designed to be 32μm for a polymer thickness of 10μm. Reduction of electrical losses was a great concern in order to derive the bandwidth benefits [6] provided by the outstanding velocity match of polymer materials [1]. Simulations revealed a pronounced increase in line loss for a strip thickness < 1μm, which prompted the use of an up-plating technique. Gold plating of the strip to a thickness of 4.75μm resulted in a line loss of 3.64dB/cm at 40GHz. The effect of the lower ground plane thickness on the microstrip loss was also studied. Reducing this thickness from 2μm to 0.1μm showed a 1.2dB/cm increase in loss at 40GHz. With the above results in mind, the microstrip lines were fabricated with thick ground and top electrodes.

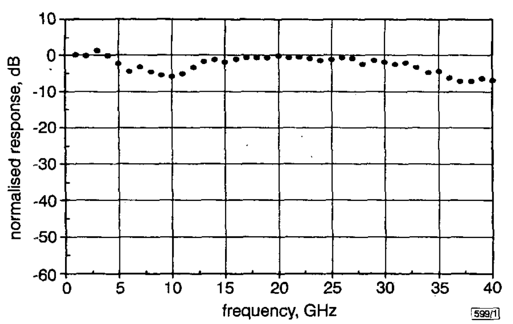


**Fig. 1** *Measured normalised frequency response for FTC Mach-Zehnder modulator*

Dips at 7.5 and 37.5GHz arose from resonance in microstrip line due to sharp bends

The frequency response of the FTC modulator was measured from 0 to 40GHz using a heterodyne technique similar to that described in [7]. The output of the diode-pumped Nd:YAG laser operating at 1.3μm was coupled into PM fibre after transmission through an intensity control unit. Subsequently, the fibre was butt-coupled to the modulator, the output of which was coupled to a PM 2 × 2 coupler. The other diode-pumped Nd:YAG laser was coupled into the remaining input port of the 2 × 2 coupler, enabling the optical heterodyne measurement. The operating frequencies of both lasers were offset by a controlled and known amount. One coupler output was sent to a fast photodetector, which was subsequently connected to a spectrum analyser. The modulator was driven by an electrical signal from an HP synthesiser using DC to V band air-spaced coplanar probes.

Mach-Zehnders with 1.62cm interaction lengths were characterised using the heterodyne setup. A $V_\pi$ of 4.7V was measured for these devices at 1.3μm, which was close to the calculated value of 3.6V. The measured normalised frequency response from 1 to 40GHz is shown in Fig. 1. The ripple in the frequency response at 7.5 and 37.5GHz arose from resonances due to the abrupt bends in the microstrip line (see Fig. 2). In the bands from 0 to 4GHz and from 13 to 27GHz the response remained relatively flat near a value of 0dB. Future work along these lines will involve the redesign of the microstrip line in order to eliminate resonances and thereby derive a virtually flat frequency response across the 40GHz bandwidth.

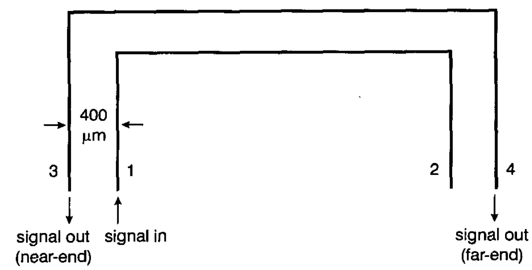


**Fig. 2** *Schematic diagram of the crosstalk measurement between adjacent microstrip lines in modulator array*

Coplanar probes are used to launch microwave signal at 1 and pick it up at either 3 or 4

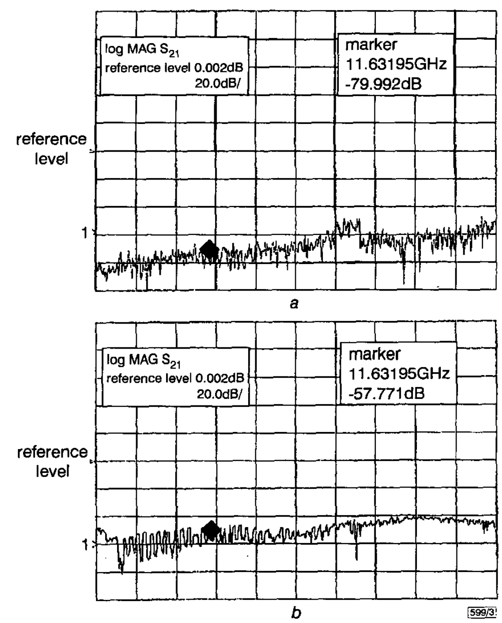


**Fig. 3** *Measurement of near-end crosstalk using network analyser*

*a* noise floor
*b* transmission between points 1 and 3 of Fig. 2
centre: 20.0225GHz
span: 39.999GHz

The schematic representation of the electrical crosstalk measurement between two adjacent modulators in a modulator array is shown in Fig. 2. A coplanar probe was used to launch the microwave signal at point 1 and another probe was used to measure the signal at point 3 or 4. An 8510 network analyser was used to determine the crosstalk as a function of frequency up to 40GHz. A measurement made between points 1 and 3 is referred to as the 'near-end crosstalk' and, as can be seen from Fig. 3, it is < –40dB over the entire band of 0–40GHz. The far-end crosstalk (between points 1 and 4) was < –45dB over 0–40GHz. These results were extremely encouraging considering that no crosstalk reduction measures were taken. The high level of isolation was attributed to the relatively small dielectric constant of polymers at microwave frequencies [8] and to the comparatively small lateral dimensions of the microstrip lines.

To study the effect of microwave packaging on the electrical crosstalk, adjacent modulators in the array were connectorised and the chip was enclosed in a metal casing. Preliminary measurements have shown that packaging has negligible effect on the crosstalk.

*Conclusion:* Mach-Zehnder travelling-wave polymer modulators were fabricated from FTC. These devices exhibited a $V_\pi$ of 4.7V. The flat frequency response of the modulators in the 0 to 40GHz band was compromised by the microstrip line geometry; however, this problem will be eliminated in future work. Without the implementation of any crosstalk reduction measures, the crosstalk level between adjacent modulators was < –40dB in the band between 0 and 40GHz. With a small half-wave voltage, impressive frequency response and high isolation levels, these devices are ideally suited for many photonic applications.


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## $L_p$-norm based minimisation algorithm for signal parameter estimation

Hao Zhang and Yingling Peng

A new $L_p$-norm-based minimisation algorithm for signal parameter estimation is proposed. This method can be used to obtain accurate results without any prior knowledge of the key coefficient $p$ or any adaptive process to find the optimum value of $p$. Simulation shows it to be an efficient and robust method.

*Introduction:* Recently the $L_p$-norm estimation technique has received much attention in the field of signal processing because of its robustness and efficiency [1]. It is especially suitable for non-Gaussian signal parameter estimation. Traditionally the least square method has been used to extract the useful information from noise, always assuming the noise is Gaussian for it is in this situation that the least square is optimal in the sense of maximum likelihood. As is well known, the performance of the least square method is significantly reduced in the case of non-Gaussian noise distribution. The $L_p$-norm estimation technique overcomes this problem. It can produce more accurate results than the least square method when the signal is contaminated by impulsive or heavy tail distributed noise.

Although $L_p$-norm estimation is attractive, it has one obvious drawback – how to determine the optimal exponential coefficient $p$. Many researchers have suggested that $p$ should be adjusted adaptively according to the kurtosis of the residual estimators error [2]. In addition to the ambiguous relation between the kurtosis of the residual error and coefficient $p$, there exist another two difficulties: accurate estimation of the kurtosis with short data samples and the convergence of the adaptive process.

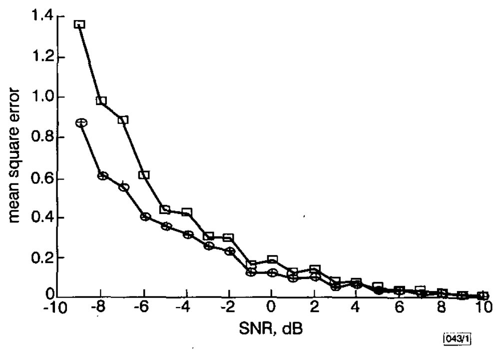


**Fig. 1** *Mean square error in the presence of double exponential noise*

□ least square method
○ $L_1$-norm estimation
+ MIRLS method

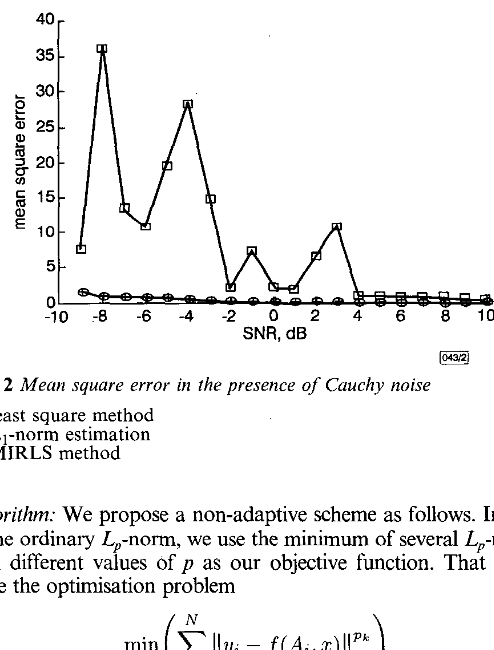


**Fig. 2** *Mean square error in the presence of Cauchy noise*

□ least square method
○ $L_1$-norm estimation
+ MIRLS method

*Algorithm:* We propose a non-adaptive scheme as follows. Instead of the ordinary $L_p$-norm, we use the minimum of several $L_p$-norms with different values of $p$ as our objective function. That is, we solve the optimisation problem

$$\min_{p_k}\left(\sum_{i=1}^{N}\|y_i - f(A_i, x)\|^{p_k}\right)$$

where $y_i$ is signal buried in noise, $A_i$ is the known signal component and $x$ is the unknown signal coefficient. Usually $f$ is a linear operator, in particular $f(A_i, x) = A_i x$.

The conventional method for solving the least $Lp$-norm optimisation problem is the IRLS (iteratively reweighted least square) method, but this is not applicable to our problem. We therefore use the so-called MIRLS (modified IRLS) method. The $k$th iterative step of the MIRLS method is described as follows:

step 1: $r_i(k) = (y - Ax(k))_i$
step 2: $W(k) = I_{N\times N}$