microwave frequencies. Furthermore, the use of lower frequency LO sources helps to avoid signal leakage problems. Moreover, it benefits from the fact that the RF and LO signals are normally relatively close in frequency (for a comparatively low IF) [3].

The subharmonic mixer scheme, shown in Fig. 2, is based on an antiparallel diode pair and use a triplexer structure to separate LO, RF and modulating signals. The antisymmetric current-voltage characteristic of the diode makes second-harmonic generation impossible if two diodes are perfectly balanced. Thus the short circuit $\lambda_{LO/4}$ stub at the LO port is a quarter of a wavelength long at the input frequency of the LO and so it is an open circuit. However, at RF this stub is approximately a half wavelength long, so providing a short circuit to the RF signal. Conversely, at the RF input the open circuit $\lambda_{LO/4}$ stub presents a good open circuit to the RF but is a quarter wavelength long at the frequency LO and so it is a short circuit. The IF is normally far enough away from the RF frequency to allow easy realisation of an IF filter presenting an open circuit output to RF port.

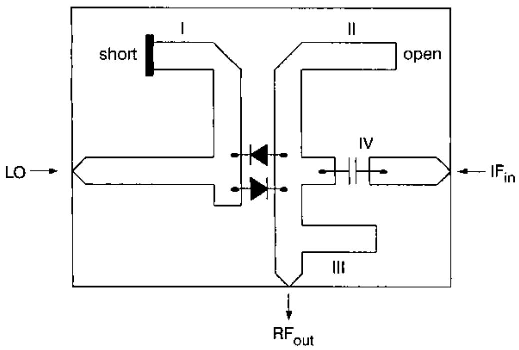

Fig. 2 *Subharmonic mixer layout*

The used beam-lead antiparallel diode pair is the HSCH-9551 manufactured by Agilent. The circuit has been realised on a thin-film Al_2O_3 substrate. Each device has been designed using Agilent-ADS. The LO power level required by the mixer is ~10 dBm for optimal performance in the operating band. The obtained conversion loss is ~9 dB.

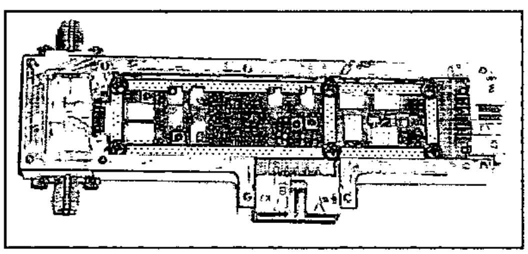

Fig. 3 *Photograph of realised modulator prototype*

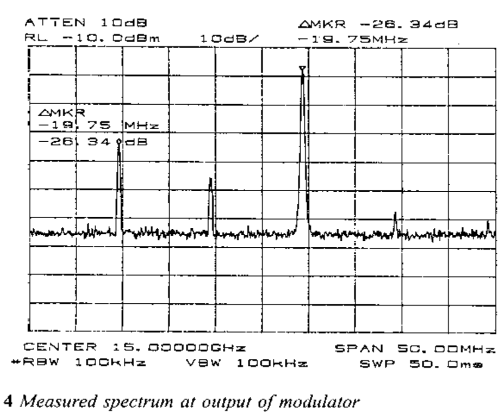

Fig. 4 *Measured spectrum at output of modulator*

Measurement results: The modulator has been characterised by sending on the *I* and *Q* inputs, two sinusoidal carriers at the same low frequency, having the same amplitude, and being strictly in quadrature. Fig. 4 shows the measured output spectrum. The modulator performance around 15 GHz in terms of measured image rejection is 26 dBc, while the carrier rejection is 37 dBc with an input LO power level of 13 dBm. The experimental conversion loss is ~10 dB coherently with the results of the individual mixers.

Successive analysis has been performed generating the *I* and *Q* signals using a modem (data rate = 16 × 2 Mbit/s; modulation = 16QAM) to examine the observance of the limits imposed by the 15 GHz ETSI standard. As shown in Fig. 5, the output spectrum is perfectly held in the ETSI mask.

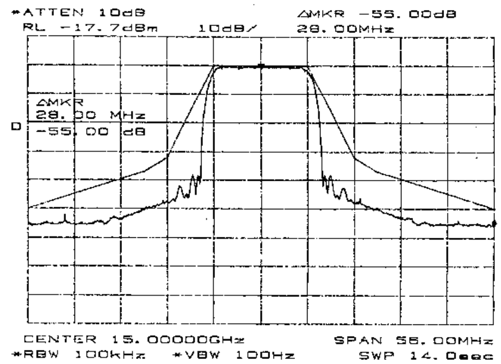

Fig. 5 *Comparison between measured modulated signal and ETSI mask*

Conclusion: Direct modulation of the carrier signal has been shown to be an attractive means of reducing the complexity of the RF hardware required in a communication transceiver. A simple technique to realise a direct IQ modulator using a subharmonic mixer has been described. The experimental results show that this new implementation scheme provides a low-cost and high-performance quadrature amplitude modulation (QAM) modulator.

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Adaptive SNR algorithm for turbo codes

Jia Hou and Moon Ho Lee

The sensitivity of the iterative maximum *a posteriori* probability (MAP) decoder to the initial channel values is investigated and a simple adaptive signal-to-noise ratio (SNR) algorithm operating within the delta variances from two MAP decoders is proposed. It can improve the BER by about 0.3 dB without much additional hardware complexity.

Introduction: Turbo codes with maximum *a posteriori* probability (MAP) decoding can achieve remarkable performance over additive white Gaussian noise (AWGN) channels [1]. The requirement is knowledge of the signal-to-noise (SNR), which determines initial values for MAP decoders. The SNR value is always given as constant in the conventional decoding scheme. However, because of the effect from the channels, the SNR of each frame is different and unequal to the given SNR value. Therefore, an algorithm based on extrinsic values was proposed to update the SNR value towards its optimum value for each decoder iteration or each turbo code frame [2]. In this Letter, we investigate the relation between the delta variances from two MAP decoders and the initial channel values, and propose a simple adaptive SNR algorithm to improve the BER by 0.3 dB from the BCJR algorithm and 0.2 dB from the algorithm in [2], respectively.

System model and review: A turbo code consists of two recursive systematic convolutional (RSC) codes with feedback. Let u_k, $k \in \{1, \ldots, N\}$, be the information bits, the code bits of which are binary phase shift keying (BPSK) modulated and transmitted through an $N(0, \sigma^2)$ AWGN channel. At the receiver, $(y_k^s, y_{1k}^p, y_{2k}^p)$ are signals corresponding to u_k, where y_k^s is the systematic signal, and y_{1k}^p and y_{2k}^p are parity for RSC1 and RSC2, respectively. They are sent to the MAP decoders DEC1 and DEC2 to produce the estimates $\hat{u}_k$ (see Fig. 1). At the ith iteration, let $L_j^{(i)}(\hat{u}_k)$ and $L_{ej}^{(i)}(\hat{u}_k)$ denote, respectively, the log-likelihood-ratio (LLR) and the extrinsic values of the estimated information bit $\hat{u}_k$ delivered by decoder j, with $j = 1, 2$. We have

$$L_1^{(i)}(\hat{u}_k) = L_{e2}^{(i-1)}(\hat{u}_k) + \frac{2}{\sigma^2} y_k^s + L_{e1}^{(i)}(\hat{u}_k) \tag{1}$$

$$L_2^{(i)}(\hat{u}_k) = L_{e1}^{(i)}(\hat{u}_k) + \frac{2}{\sigma^2} y_k^s + L_{e2}^{(i)}(\hat{u}_k) \tag{2}$$

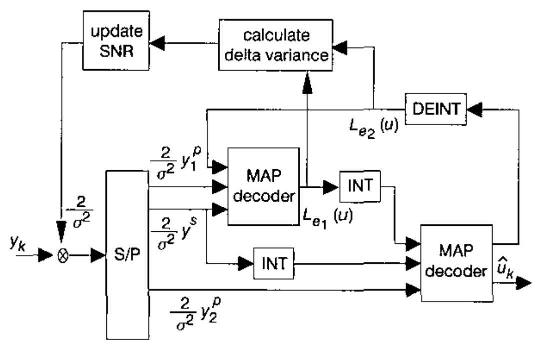

Fig. 1 *System model of proposed turbo decoders*

where $(2/\sigma^2)y_k^s$ is the initial value of MAP decoders. The channel reliability value σ is obtained from SNR (E_b/N_0) as

$$\sigma = \frac{1}{\sqrt{2 \times R \times (10)^{SNR/10}}} \tag{3}$$

where R is the code rate and SNR is defined by bit energy E_b and noise N_0. In addition, over different SNRs, the turbo codes have a different optimum number of iteration (see Fig. 2). Thus, [3] proposed a method to simply set the stop criterion based on SNR estimation [4] and properties of delta variance. There, the variances of the extrinsic information of $\hat{u}_k$ are

$$Var[L_{e1}^{(i)}] = \frac{\sum_{k=0}^{N-1} L_{e1}^{(i)}(\hat{u}_k)^2}{N} - \left(\frac{\sum_{k=0}^{N-1} L_{e1}^{(i)}(\hat{u}_k)}{N} \right)^2 \tag{4}$$

$$Var[L_{e2}^{(i)}] = \frac{\sum_{k=0}^{N-1} L_{e2}^{(i)}(\hat{u}_k)^2}{N} - \left(\frac{\sum_{k=0}^{N-1} L_{e2}^{(i)}(\hat{u}_k)}{N} \right)^2 \tag{5}$$

where N is the data length. The delta variance at the ith iteration is then defined as [3]

$$\Delta Var[L_e^{(i)}] = Var[L_{e2}^{(i)}] - Var[L_{e1}^{(i)}] \tag{6}$$

The change of BER with delta variances in Fig. 3 clearly demonstrates that the error-floor characteristic occurs after the maximum delta variance and the linear increasing property occurs before the optimum BER, this being because delta variances represent the reliability of the decoded bits.

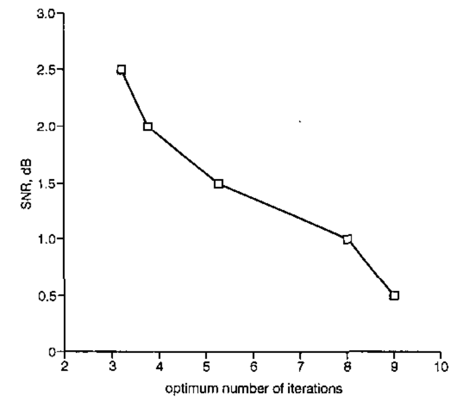

Fig. 2 *SNR and optimum number of iterations*

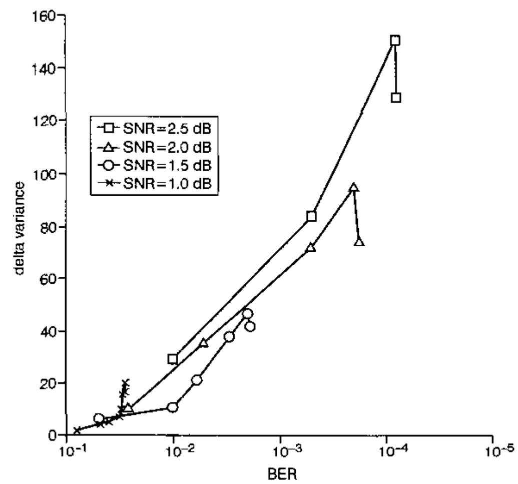

Fig. 3 *BER and delta variance*

Proposed turbo decoding scheme: A fixed SNR value is often chosen for any frame and any iteration in the conventional BCJR algorithm. However, this fixed SNR cannot accurately present channel information. Therefore, to improve the BER performance, we propose an update scheme from outputs of decoders. First, from Fig. 2, we have

$$SNR = f(I) \tag{7}$$

where I is optimum number of iterations, and $f()$ is a linear simple function. Furthermore, the I decides the optimum BER performance B as

$$I = \varphi(B) \quad \text{and} \quad B = \varphi^{-1}(I) \tag{8}$$

where $\varphi()$ is the simple function with one by one corresponded. Otherwise, we can construct a simple corresponding function with delta variance ΔVar and optimum performance B from Fig. 3:

$$\Delta Var = f'(B) \text{ and } B = f'^{-1}(\Delta Var) \tag{9}$$

where $f'()$ also is a simple linear function. Generalising (7)–(9), we then get

$$SNR = f(\varphi(f'^{-1}(\Delta Var))) \tag{10}$$

Obviously, it is a simple linear function, as the others. We can plot it after the first iteration, as in Fig. 4, and the other iterations have the same linear property with different amplitudes from the simulations.

Table 1: List of SNRs and simulation time

Average SNR (dB)	BCJR	Adaptive SNR [2]	Proposed scheme
0.5	0.5	0.45	0.47
1.0	1.0	0.93	0.92
1.5	1.5	1.31	1.47
2.0	2.0	2.18	2.12
2.5	2.5	2.70	2.67
Simulation time (min/100 frames)	9.4	10.3	10.1

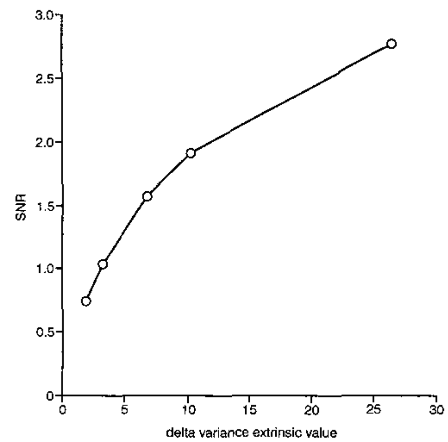

Fig. 4 *Delta variance after first iteration*

To accurately describe the channel optimum value for each frame from measurement of the MAP decoders and information bits, we propose an adaptive SNR algorithm using delta variance and its linear function to update the initial values after the first iteration. The estimation of updated SNR is denoted as

$$SNR \simeq \log(0.566 \times \Delta Var(L_e^{(1)}) + 0.996) \qquad (11)$$

where the calculation of delta variance is based on a given SNR which is same as in the conventional BCJR scheme. In general, updating the SNR from (11) for other iterations, we can obtain reliable channel values to improve the performance of turbo codes. The proposed system has simple calculation and little additional hardware complexity (see Fig. 1). The performance is shown in Fig. 5. We observe that it uses little time delay to improve 0.3 dB from the conventional BCJR algorithm [1] and 0.2 dB from the adaptive channel SNR scheme in [2].

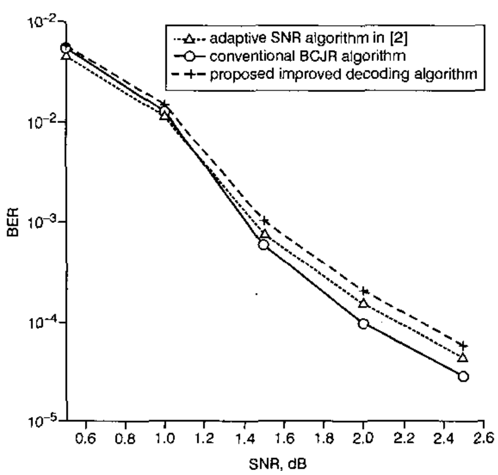

Fig. 5 *Performances of proposed turbo decoding scheme*

Conclusion: We have presented some properties of delta variance from MAP decoders, and propose a modified decoding algorithm over AWGN channels. The proposed algorithm obviously improves BER performance and reduces the time delay from the scheme in [2].

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Efficient stopping criterion for iterative decoding of turbo codes

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A simple and efficient stopping criterion for iterative turbo decoding is proposed, focusing on not only reducing the average number of iterations but also minimising performance degradation due to early stopping. A measure and two decision thresholds are introduced to enhance performance and simplify the stopping criterion. Simulation results show that the proposed method provides performance very close to that of a decoder with optimum stopping criterion and reduces average number of decoding iterations.

Introduction: Turbo code can achieve performance close to the Shannon limit by iterative decoding based on the soft-input and soft-output decoding algorithm [1]. It has been widely used as channel coding schemes of 3GPP and 3GPP2 standards for CDMA wireless communication systems. For a decoding method of turbo codes, various iterative decoding schemes are used since they provide better performances as the number of iterations increases. However, an iterative decoding has the drawback that it gives very little performance improvement with further iterations after a sufficient number of iterations which results in unnecessary decoding delays and computations. To solve this problem, several criteria have been devised to stop the iteration after sufficient iterative decoding [2, 3]. However, there are problems in using these methods in practical systems in that some criteria have complex computations and large memories and others have variable performance losses with respect to various frame size or signal-to-noise ratio (SNR). In this Letter, we propose a simple stopping criterion to solve this problem, based on Hagenauer's approximated cross-entropy (CE) stopping criterion, where the cross-entropy of distributions of the decoder's outputs is used as a measure of the closeness of them [2]. Frame error rate (FER) is provided to compare performance to Genie-aided decoding, where Genie stops the decoding when no errors occur in the frame. We show that the performance of the proposed scheme is comparable to Genie-aided decoding, to be feasible in practical systems, and is stable regardless of various frame sizes and SNR.

Stopping criterion: Stopping criterion is generally described as follows using a measure and a threshold:

$$\textit{Stop the decoding iteration if } M(i) < T(i)$$
$$(or M(i) > T(i)) \textit{ at the ith iteration}$$