Bandwidth efficient QAM schemes for Rayleigh fading channels

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Indexing terms: Modulation, Transmitters

Abstract: Techniques to improve the BER performance of 16-level QAM transmission over Rayleigh fading channels for PCN systems are presented. Simulations were carried out with a carrier frequency of 1.9 GHz, data rate of 64 kbit/s and mobile speed of 30 mph. The residual BER was reduced by over an order of magnitude, by introducing a circular constellation coupled with differential amplitude and phase encoding, compared with a square QAM constellation. When an oversampling and interpolation technique was combined with the circular constellation with differential encoding another order of magnitude reduction in residual BER was obtained. By expanding the number of QAM levels to 64, and on using the two extra bits gained for block coding, the BER was reduced to 10^{-6} for channel SNRs in excess of 35 dB. Decreasing the data throughput to 48 kbit/s using a $\frac{3}{4}$ rate RS(60, 44, 12) code, 16-level QAM, interleaved over 40 ms, transmitted over a Rayleigh fading channel yielded a BER of 10^{-6} for channel SNRs above 25 dB.

1 Introduction

Quadrature amplitude modulation (QAM) is a bandwidth efficient transmission method for digital signals. It is expected that the available spectrum for the proposed personal communications network (PCN) will soon be at a premium as the number of subscribers increases, and changing from binary modulation to QAM may significantly ease the problem. The severe amplitude and phase changes introduced by the fading channels make low error transmission of QAM difficult to achieve, unless procedures are introduced at both the transmitter and the receiver to combat the fading.

Methods of improving the bit error rate (BER) of QAM using various forms of coding on an increased symbol set are described. This means that the data throughput, symbol rate and transmission power are unaffected, although the transmitter and receiver are made considerably more complex. In the deliberations Rayleigh, rather than the less severe Rician, fading channels are considered to obtain worst case performance estimates of mobile radio communications.

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2 QAM transmissions over Rayleigh fading channels

Digital communications using 16-level QAM signals and conventional receiver techniques have unacceptably high bit error rates (BER) in a Rayleigh fading environment [1]. The problem is the inability to track absolute phase during fades with the result that on emergence from a fade the phase locked loop (PLL) in the receiver locks onto a different quadrant than that required [2]. Differential encoding can reduce this false phase locking problem, but the standard square QAM constellation suffers from possible false lock positions at 26° and 53°. At these angles, given appropriate amplitude scaling, more than half the original constellation points can be successfully mapped onto the rotated constellation points. The points that cannot be successfully mapped cause random fluctuations in the error signal in the clock recovery loop but do not actually drive the system off lock. Data mapped onto these points will nearly always be in error and thus this problem cannot be overcome with differential coding. Unfortunately, false locking occurs fairly frequently as the PLL and automatic gain control (AGC) tend to drive the system towards these lock points. The AGC can pose a problem for the square constellation as it has to act very fast to follow the fades, yet at the same time it must maintain a high degree of accuracy to allow amplitude information to be correctly decoded.

2.1 Star QAM

A constellation having no false lock positions is introduced to overcome these deficiencies. This constellation is called 'star QAM', and is essentially a twin eight level phase shift keying (PSK) constellation, as shown in Fig. 1. This constellation does not have a minimum least free distance between points in the strict sense, but does allow efficient differential encoding and decoding methods to be used which go some way towards mitigating the effects of Rayleigh fading.

2.1.1 Effect of differential coding: Some form of differential encoding is essential with PLLs as the Rayleigh fading channel can introduce phase shifts in excess of 50° between consecutive symbols, making it extremely difficult to establish an absolute phase reference. The encoding method is straightforward. Of the four bits in each symbol, b_1, b_2, b_3 and b_4, the first is differentially encoded onto the QAM phasor amplitude so that a '1' causes a change to the amplitude ring which was not used in the previous symbol, and a '0' causes the current

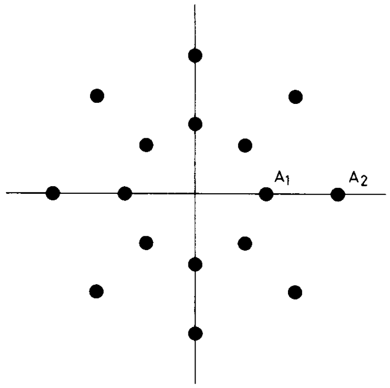

Fig. 1 *Star 16 QAM constellation at transmitter*

symbol to be transmitted at the same amplitude as the previous symbol. The remaining three bits are differentially Gray encoded onto the phase so that, for example, '000' would cause the current symbol to be transmitted with the same phase as the previous one, '001' would cause a 45° phase shift relative to the previous symbol, '011' a 90° shift relative to the previous symbol, and so on. Decoding data is now reduced to a comparison test between the previous and current received symbols. Suppose the transmitted rings are fixed at amplitude levels A_1 and A_2 as shown in Fig. 1. Let the received phasor amplitudes be Z_t and Z_{t+1} at time t and $t + 1$, respectively. The demodulator must identify whether there has been a significant change in amplitude to regenerate a logical '1'. The algorithm employed at the demodulator uses two adaptive thresholds to make its decision if

$$Z_{t+1} \geqslant \left(\frac{A_1 + A_2}{2}\right) Z_t \tag{1}$$

or if

$$Z_{t+1} < \left(\frac{2}{A_1 + A_2}\right) Z_t \tag{2}$$

then a significant change in amplitude is deemed to have occurred and bit b_1 is set to logical '1' at time $t + 1$. The thresholds are dependent on Z_t, and as the amplitude of the phasors change in fading conditions so do the thresholds. Should both of eqns. 1 and 2 fail to be satisfied, b_1 is assigned logical '0'.

If the received symbol phases are θ_t and θ_{t+1} at time t and $t + 1$, respectively, the demodulated angle is

$$\theta_{dem} = (\theta_{t+1} - \theta_t) \bmod 2\pi \tag{3}$$

This angle is then quantised to the nearest multiple of 45° and a lookup table consulted to derive the remaining three output bits, b_2, b_3 and b_4. This system considerably improves the BERs compared with those for the square constellation because it eliminates long error bursts that occur when a false lock has been made. Of considerable importance is that with differential amplitude encoding there is no longer any need for AGC. This not only simplifies the circuit, but also removes errors caused by an inability of the AGC to follow the fading envelope.

2.1.2 Effect of oversampling: Although the objectionable false locking characteristic of the square QAM constellation has been removed, the fading continues to cause problems because there are changes in channel amplitude and phase over consecutive symbol periods, which in general moves the differentially decoded phasors nearer to the decision boundaries. The most likely cause of error in the star QAM system is when both the noise, and the change in the phasor amplitude or phase caused by fading, combine to drive the incoming signal level over a decision boundary. As the Rayleigh fading envelope is crudely predictable, particularly at low vehicular speeds, a correction factor may be applied to the incoming signal to compensate for the changes in the fading envelope over the last symbol period. This should not be too dependant on the previous observations as they will allow errors to propagate. The system must be fast acting so that the sudden change from an amplitude decrease to an amplitude increase experienced at the bottom of a fade can immediately be detected and compensated. Current methods, such as using the differential of the PLL error signal to change the step size of the phase correction signal, do not fulfil these criteria as they tend to overshoot on sudden changes and exhibit damped second-order system behaviour. A simple oversampling receiver is used to overcome this problem. In this system n observations equally spaced in time are made per symbol period. When E_b/N_0 is low (below 30 dB for 16 QAM) the first observation from the current symbol is compared with the last observation from the previous symbol period. This reduces the magnitude of any changes in the channel phase and amplitude and is similar to the advantage gained by transmitting at a higher rate without the problems of frequency selective fading and the loss of a large number of bits during a fade. At higher E_b/N_0 ratios the current phasor can be modified to compensate for the fading. This is achieved by finding the change in the incoming symbol phase and amplitude over the current symbol period. This is performed by subtracting the phase at the end of the symbol from the phase at the beginning of the symbol, and subtracting the amplitude at the end of the symbol from the amplitude at the beginning of the symbol. The situation is shown in Fig. 2. The same notation is used as before, but with the addition of subscript n to the symbols, to signify the nth observation of the symbol at time t. Thus

$$Z_{t+1,diff} = Z_{t+1,n} - Z_{t+1,1} \tag{4}$$

where *diff* means differential, and

$$\theta_{t+1,diff} = \theta_{t+1,n} - \theta_{t+1,1} \tag{5}$$

These changes can be used to extrapolate back from the first observation in the current symbol to the point in time when the last observation from the previous symbol was made.

To improve the accuracy of eqns. 1 and 2, Z_t is replaced by the last sample value, namely $Z_{t,n}$ and an estimate of Z_{t+1} is used which is formed by extrapolation as

$$Z_{t+1,ext} = Z_{t+1,1} - \frac{Z_{t+1,diff}}{n} \tag{6}$$

rather than Z_t. Notice that $Z_{t+1,1}$ is the first sample of Z_{t+1} and is closest in time to $Z_{t,n}$. By subtracting the average change in Z_{t+1} over a symbol period, i.e. $z_{t+1,diff}/n$ from $Z_{t+1,1}$, an estimate of Z_{t+1} had it been transmitted at time t is obtained. This is beneficial as there may have been significant amplitude and phase changes between time t and $t + 1$ as can be seen in Fig. 2.

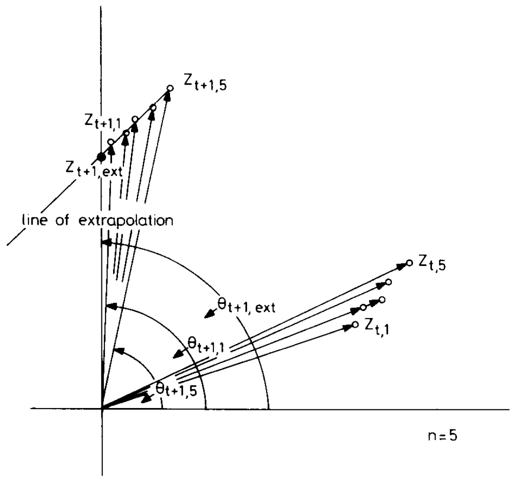

Fig. 2 *Correction for changes in fading envelope*

After determining bit b_1 with the aid of eqns. 1 and 2 with the modified Z_t and Z_{t+1} values, bits b_2, b_3 and b_4 are determined by formulating

$$\theta_{dem} = (\theta_{t+1,ext} - \theta_{t,n}) \bmod 2\pi \quad (7)$$

where $\theta_{t+1,ext}$ and $\theta_{t,n}$ are the phase angles associated with $Z_{t+1,ext}$ and $Z_{t,n}$. Again θ_{dem} is quantised and used to address a lookup table which provides values of b_2, b_3 and b_4.

It has been assumed that no pulseshaping or filtering was employed so that square pulseshapes result. This is impractical because of the spectral spillage caused by such modulation. The oversampling system can work with practical pulseshaping systems. The use of nonlinear filtering (NLF) is preferable because of its simplicity and only slight bandwidth increase over complicated partial response raised-cosine modulation schemes [8]. With pulseshaping the oversampling system needs to perform an inverse operation before predicting the change undergone in the channel. At the receiver a NLF waveform is reconstructed based on the absolute values of the previous and current waveforms, and the deviation from this reconstructed waveform is used to estimate the average change. Simulations showed the degradation in BER over unfiltered transmissions to be negligible.

This oversampling method performs worse at low E_b/N_0 ratios because the noise tends to render the predictions inaccurate. In practice oversampling is only worth implementing for E_b/N_0 ratios in excess of 30 dB, where substantial improvements can be made. Previous QAM systems tend to exhibit a residual BER at high E_b/N_0 ratios caused by the rapidly changing Rayleigh channel, rather than the additive noise. With this system the residual BER is reduced by approximately two orders of magnitude.

2.2 Simulation results

Pseudo-random data generated at 64 kbit/s were organised into four bit words having bits b_1, b_2, b_3 and b_4 in each word and mapped onto the 16-level star QAM constellation shown in Fig. 1. Whenever b_1 was a logical '1' the current phasor was transmitted at a level of A_1 or A_2 if the previous phasor amplitude was A_2 or A_1, respectively. If b_1 was a logical '0' then the amplitude of the transmitted phasor was unchanged from that of the phasor previously transmitted. When the bits $b_2 b_3 b_4$ were 000, 001, 011, 111, 101, 100, 110 and 010 the current phasor was transmitted with an angle of 0, 45, 90, 135, 180, 225, 270 and 315°, respectively, relative to the angle of the previous phasor. The QAM carrier was 1.9 GHz, and the mobile's speed was 30 mph. The channel exhibited flat Rayleigh fading generated using the normal twin white noise sources passed through Doppler filters and then combined, in quadrature, with additive white Gaussian noise (AWGN). The receiver operated as described above, with the oversampling ratio set at $n = 8$. Nonlinear filtering was employed at the transmitter. Both the transmitter and the receiver used a fourth-order Butterworth lowpass filter with a 3 dB point at 1.5 times the Baud rate, i.e. 25 kHz. The variation of BER with E_b/N_0 is shown in Fig. 3. The results for the square QAM

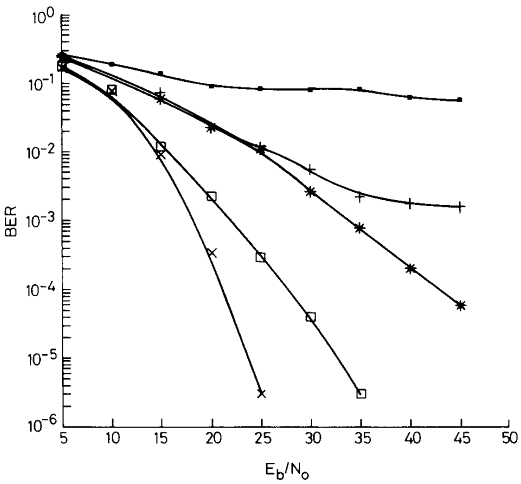

Fig. 3 *BER performance for QAM constellations*

—■— square
—+— star
—*— oversample
—□— diversity
—×— AWGN only

constellation and the star QAM with differential encoding are also shown as bench marks both over a Rayleigh fading channel, and the star QAM with differential encoding over an AWGN channel. The performance of the system having differential encoding and oversampling can be considerably enhanced by the use of spatial diversity, where two antennas and receiver circuits are used. Switched diversity was used for these simulations. The receiver with the incoming phasor of largest magnitude was selected for each phasor received. Both receivers must have their own differential decoders. The performance of the system is also shown.

A very substantial improvement in BER has been obtained over the conventional square QAM by using star QAM. At high E_b/N_0 ratios the oversampling technique gave a further significant gain. These changes to the basic QAM were achieved for only a small increase in receiver complexity. The system operated with a channel SNR of only 5 dB above that for an AWGN channel for a BER of 10^{-3} by introducing second-order switched diversity.

3 Trellis code modulation for QAM

Papers published by Ungerboeck and others [3] suggest that for an AWGN channel, significant coding gains can be achieved by expanding the symbol set size and using the extra bit(s) for channel coding. In its most common form this is known as trellis code modulation (TCM). The QAM symbol set size is doubled and the extra bit gained is used to convolutionally encode some of the lower integrity bits in the symbols.

TCM was conceived to operate with modems over AWGN channels. It could be assumed that an accurate phase reference could be established and maintained and so there was no problem with false locking. One of the properties of TCM systems is that they exhibit invariance to 180° rotations, i.e. if lock is established 180° from the correct lock position operation will be unchanged. Systems have been suggested [4] having nonlinear convolutional encoders which exhibit invariance to all 90° rotations, but no system exhibiting invariance to 45° rotations has been published. This means that TCM cannot be used as it stands with the star QAM system, as half of the possible lock points will lead to incorrect system operation. To maintain phase invariance to 45° rotations, it is necessary to introduce differential coding whereby each TCM word is associated with a phase and amplitude change, rather than with an absolute phase and amplitude as is normally done. The constellation chosen is a four ring star with eight points on each level, i.e. the 32-level star QAM shown in Fig. 4. This arrangement does not have a constant minimum free distance, but it does enable the TCM to be optimally and differentially mapped onto the constellation. Sub-optimal mappings involving constellations, for example, that have 4, 6, 10 and 12 points on the first, second, third and fourth rings, respectively, where the rings are numbered starting with the inner ring, have poorer results than a constellation having eight points on each of its four rings. Simulations of TCM revealed marginal improvements in BER at very high E_b/N_0 ratios compared with the 16-level QAM, although significant degradation in BER was experienced at the lower E_b/N_0 ratios encountered in practice. This degradation at lower E_b/N_0 ratios is because the BERs on which the convolutional decoder was operating were so high that the decoder often chose the incorrect path through the trellis and thus increased the number of errors. It was not until the BER fell below a certain level that the convolutional coding system was able to reduce the number of errors. TCM systems only proved beneficial when switched diversity was used, achieving a residual BER of 5×10^{-5} at E_b/N_0 ratios above 35 dB. This is because current TCM systems have been optimised for AWGN. Current research into TCM schemes for Rayleigh fading may provide systems which will be more successful [5].

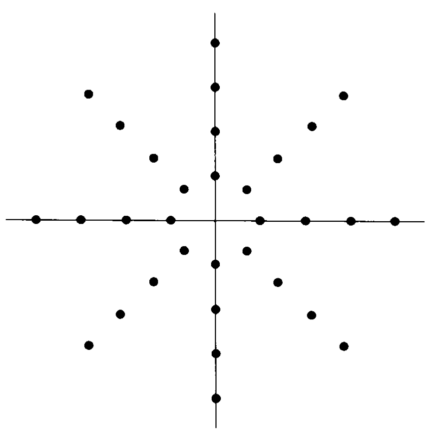

Fig. 4 *Star 32 TCM constellation*

4 Block coding

Block coding, rather than convolutional coding, was employed by expanding the signal set to cope with the extra bits required for the code. A 2/3 coding rate was considered to be appropriate, causing the number of QAM levels to increase from 16 to 64. An extension of the star constellation was selected having four amplitude rings with 16 points equispaced on each ring. This is not optimal with regard to the minimum distance criterion, but simulations showed it to work better in fading environments than constellations which were optimal for AWGN.

For block codes to perform well in the Rayleigh fading environment it is necessary to add interleaving to the system to randomly distribute the errors in time. The block code and the interleaving process introduce a delay, and the maximum permissible delay depends on the type of information to be transmitted. The integrity constraints of computer data transmissions are at least three orders of magnitude higher than those of digital speech transmissions. Data channels can accept longer interleaving delays which allows an effective randomisation of the bursty error statistics of the Rayleigh fading channel. This considerably increases the ability of the forward error correction (FEC) decoder to decrease the BER. For each transmission rate, propagation frequency, vehicular speed and interleaver algorithm there is a minimum interleaving depth or delay to transform the BER statistics of a fading channel into a good approximation of those encountered in a Gaussian channel. An interleaved FEC codeword has to overbridge the channel fades for a specific combination of system parameters. For the emerging personal communication network (PCN) this is a difficult requirement, since a high proportion of mobile subscribers are moving at walking pace and therefore may spend a relatively long time in a deep fade. One proposed solution, which is to be incorporated into the GSM system, is the use of frequency hopping. This sets an approximate limit on the length of time that can be spent in a fade. Should a mobile be stationary and in a fade, then the probability of hopping to another fade is very remote. Thus hopping produces a similar effect to setting a lower limit on the mobile speed and reduces the interleaver depth required. Given an appropriate hop rate and a powerful enough interleaver the channel errors can be sufficiently randomised for channel coding to work at any mobile speed. The results presented here would apply to all mobile speeds with the proviso that frequency hopping is carried out often enough to allow the interleaver to work correctly.

For the propagation frequencies (1.7–1.9 GHz) used for PCN there is approximately one fade every 10 cm distance. If the data transmission rate is 64 kbit/s, yielding a signalling rate of 16 ksample/s for the uncoded 16-level QAM, then when the mobile station is travelling at

30 mph, there are approximately 200 QAM samples transmitted between two deep fades. Interleaving over three fades randomises the bursty error statistics and has a delay of approximately 600 QAM samples corresponding to less than 40 ms. By increasing the interleaving depth to much higher values as are acceptable for data transmission the BER is further improved. For infinite delays the performance of the memoryless AWGN channel is achieved and is shown in Fig. 3.

The 40 ms delay quoted for three fades would be acceptable for PCM, ADM and ADPCM coded speech transmissions as the encoding delay introduced by the speech codec is negligible. Unfortunately, for complex speech coders, e.g. the regular pulse excited (RPE) speech codec used by GSM which can produce toll quality speech at 13 kbit/s, the delay introduced by the codec is sufficiently large that when it is added to the interleaving delay conversational speech becomes unacceptable.

Bose–Chaudhuri–Hocquenghem (BCH) block codes have favourable properties for PCN transmissions. They can correct both random and bursty errors and their error detection capability allows the BCH decoder to know when the received codeword contains more errors than the correcting power of the code can cope with. Provided a systematic BCH code is used, the information part of the coded word can be separated from the parity bits so that in code overload the information bits are not corrupted by the decoding process.

A special subclass of BCH codes is the maximum minimum distance Reed–Solomon (RS) codes. These codes operate on non-binary symbols and have identical error-locator and symbol fields. The non-binary RS codes are optimum because of their maximum distance properties, and may also be sufficiently long to over-bridge channel fades. They are more complex than binary BCH codes. An extremely long RS code was selected to explore both ends of the complexity/performance trade-off, the RS (252, 168, 42) code over GF (256) using eight-bit symbols, the moderately long RS (44, 30, 7) code over GF (64), as well as the short binary BCH (63, 45, 3) code. The nomenclature used here is RS (*m*, *n*, *k*) where *m* is the number of encoded symbols, *n* is the number of information symbols and *k* is the number of corrected symbols in a codeword.

5 64-level TCM

A 64-level 2/3 rate convolutional coding system was devised as a comparison test. This was based on the 32-level TCM. TCM cannot be used directly as it doubles the symbol set size, and here there is a quadrupling of the set size from 16 to 64. The principles of TCM can still be retained. The 64 possible code words were divided into sixteen groups of four called D groups [3]. The way that TCM works is to only add coding to those bits with the highest probability of error. With four input bits and six bits per symbol, half-rate convolutional coding can be added to two bits and the other two left uncoded. The two coded bits plus the two coding bits can specify one of sixteen groups, and the uncoded two bits can specify one of four points within this group. The four points in each group are spaced as far apart across the constellation as possible so that there is little chance of one of these four points being mistakenly received as another in the presence of noise. Fig. 5 shows the constellation of 64 points representing the 64 possible transmitted phasors. Fig. 5 may also be viewed as a lookup table for the differential encoding. The six bits span the decimal range 0–63 as they vary from 000000–111111. If this decimal number is in the range 0–3, 4–7, 8–11, ..., 60–63 subgroup D0, D1, D2, ..., D15 is selected, respectively. In each subgroup, each of the four possible numbers is mapped onto one of the four points in that subgroup. This is done by starting

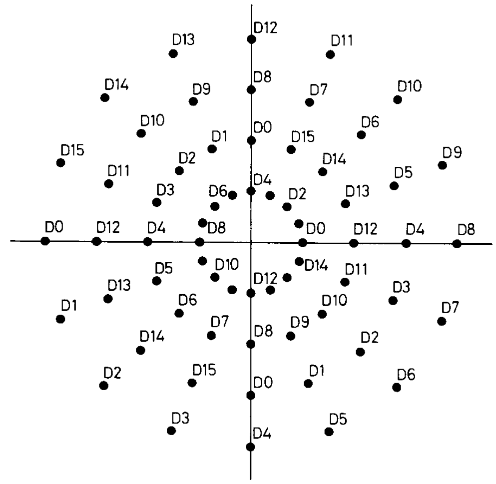

Fig. 5 *Star 64 TCM differential constellation*

on the positive x-axis and rotating anti-clockwise. The first point in the subgroup is given the lowest number for that subgroup. Consider subgroup D0, the point on the positive x-axis would be assigned the value 0, i.e. it is the point mapped to when the incoming six bits are 000000. The point on the positive y-axis becomes 1 corresponding to incoming bits 000001. The point on the negative x-axis becomes 2 corresponding to 000010, and the point on the negative y-axis becomes 3 corresponding to 000011. This is repeated for all subgroups. The D4 point on the negative y-axis would correspond to 19 or 010011. Once the phasor for the incoming codeword has been identified, it can be used to derive the differential signal which must be added to the previous constellation point transmitted to derive the constellation point for transmission. The representation is as follows.

A point on the inner ring causes the transmission of the current symbol with an identical amplitude to the previous symbol transmitted.

A point on the second ring causes the transmission of the current symbol with an amplitude on the next amplitude ring up from the previous symbol transmitted. If the previous symbol amplitude was that of the outermost ring then a change is made to the innermost ring, i.e. wrapround.

A point on the third ring causes an increase of two rings in amplitude from the previous amplitude transmitted again with wrapround, and so on.

A point at 0° (on the positive x-axis) represents no phase change from the previous point transmitted. A point at 22.5° represents a phase change of $+22.5°$ relative to the previous point transmitted. A point at 45° represents a phase change of $+45°$ relative to the previous point transmitted and so on.

As an example, if the last phasor transmitted was that corresponding to the point labelled D13 at 22.5° on the second ring and the input bits were 000010, these input bits are decoded as the D0 point on the negative x-axis. Thus the ring size is increased by three, wrapping round from the D13 point to the innermost ring, and rotating this point by 180° to end up at the unlabelled D9 point on the innermost ring at −157.5°. This is the phasor transmitted.

This representation has been used as it allows set partitioning into the D groups to be carried out in the same way as for normal (non-differential) TCM.

The half rate coder chosen was of constraint length $K = 5$ with generator polynomials

$$g1 = 1 + g_3 X^3 + g_4 X^4$$

and

$$g2 = 1 + g_1 X + g_2 X^2 + g_4 X^4$$

as recommended for the GSM system [7]. These generator polynomials are optimal for random error statistics. The encoded bits are interleaves prior to the mapping shown in Fig. 5. This interleaving was the same as used with the block codes.

6 Bandwidth efficient coding results

The TCM systems performed worse than the 16-level QAM at all E_b/N_0 ratios, as shown in Fig. 6. Normal convolutional coding may perform better than the TCM used here but block coding is favoured because of its overload detection properties. The BER using block coding dramatically decreased compared with the TCM and uncoded QAM schemes as shown in Fig. 7. The BCH (63, 45, 3) code and the RS (44, 30, 7) code have nearly identical performances, inspite of the considerably higher block length and complexity of the RS code. The RS (252, 168, 42) code offers an extra 2 dB coding gain for a large increase in complexity. The BCH code which provides virtually error-free communications for E_b/N_0 in excess of 30 dB, a value that may be realisable in the small microcells to be ultimately found in a fully developed PCN, is recommended. By using error correction coding and compensating for the increased bit rate by using higher level modulation, transmissions of higher integrity can be provided, but with identical bandwidth compared with an uncoded system.

7 Overall coding strategy

16 and 64-level QAM schemes which are error protected are considered, where the error coding decreases the useful information transmission rate and increases the integrity of the transmitted data. The 64-level QAM having a 2/3 rate code had its code rate reduced to 1/2, decreasing the information content rate by 1/4. This overall 1/2 coding rate was selected as it is widely used in mobile radio [6]. A further coding rate reduction does not bring corresponding coding gains and squanders channel capacity which cannot be compensated for by using more modulation levels. A 3/4 rate code with the 16-level QAM is also employed to provide the same overall bit rate as the coded 64-level QAM.

The results are depicted in Fig. 8. The uncoded 16-level QAM and the 64-level QAM/RS (44, 30, 7) curves are repeated for comparison along with the 64-level QAM/RS (60, 30, 15) and 16-level QAM/RS (60, 44, 12) arrangements. There is a consistent and remarkable improvement in both the 16-level QAM and 64-level QAM performance. For E_b/N_0 in excess of 25 dB there is an almost 5 dB extra coding gain improvement caused by the stronger RS code in the case of the 64-level QAM scheme. The performance of the 16-level QAM arrangement is dramatically improved by employing error cor-

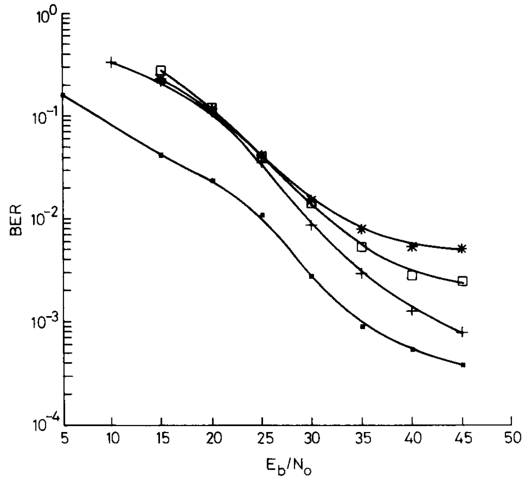

Fig. 6 *BER performance for 32 and 64 TCM*

—■— 16 star —✱— 64 TCM; $K = 4$
—+— 32 TCM —□— 64 TCM; $K = 5$

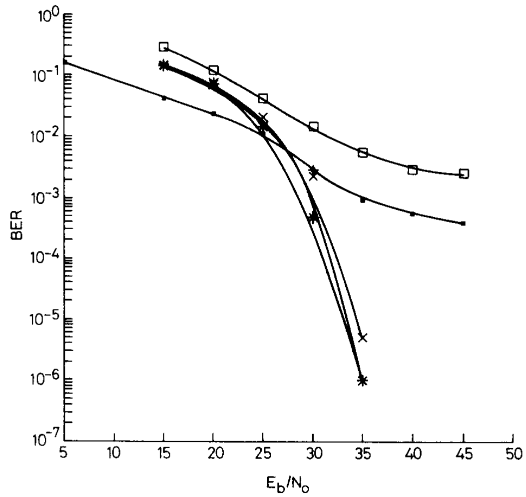

Fig. 7 *BER comparison of coded 64 and 16 QAM*

—■— 16 star —□— 64 TCM, $K = 5$
—+— BCH (63, 45, 3) —×— RS (252, 168, 42)
—✱— RS (44, 30, 7)

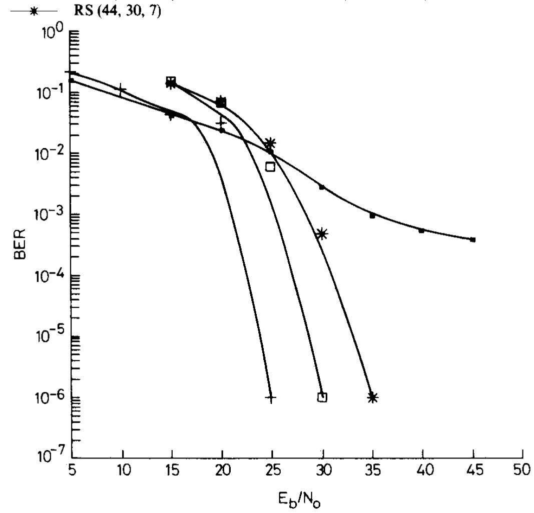

Fig. 8 *BER comparison of overlaid coding schemes*

—■— 16 star —✱— 64 + RS (44, 30, 7)
—+— 16 + RS (60, 44, 12) —□— 64 + RS (60, 30, 15)

rection coding and becomes superior to that of the 1/2-rate coded 64-level QAM system. An E_b/N_0 value of 25 dB is sufficient for reliable signalling using the 16-QAM/RS (60, 44, 12) system depending on the integrity required.

8 Summary and conclusions

Various signal processing techniques have been proposed to facilitate reliable QAM transmissions in Rayleigh fading environments. The findings indicate that these bandwidth efficient modulation schemes can be used for the emerging high traffic density PCNs. The well-known 16-level square constellation was found to be unsuitable for the PCN environment. The sub-optimal star scheme with differential encoding and oversampling signal estimation dramatically improved the BER perfoı.nance, rendering the channel appropriate for speech transmission. The lower BERs essential for data transmissions were achieved by expanding the 16-level QAM signal set to 64 levels and using the extra channel capacity acquired for error correction coding. For E_b/N_0 values in excess of 25 dB the performance of the coded 64-level QAM was superior to that of the 16-level QAM and at values of E_b/N_0 in excess of 30 dB it was virtually error free. When the overall coding rate was lowered to allow the 16-level QAM scheme to incorporate a 3/4 rate RS code and the 64-level QAM system had a 1/2 rate RS code, the 16-QAM arrangement out-performed the more complex 64-level scheme. A virtually error free performance was achieved above $E_b/N_0 = 25$ dB, a value that should be easy to achieve in the small microcells expected in PCN.

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