A Re-configurable Cordless Telecommunications Scheme

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1 Introduction

Recently there has been a considerable effort devoted to intelligent joint speech coding and forward error correction (FEC) coding [1] as well as to equal [2] and un-equal protection coded modulation [3]. In reference [4] Hagenauer introduced an intelligent framework for considering source significance-, channel state- and decoder reliability-information in order to achieve the best possible performance over fading channels.

In this contribution we propose a complete re-configurable transceiver, which employs unequal protection pilot symbol assisted block coded modulation embedded in the source codec. A unique feature of the transceiver is that it can be reconfigured under network control as a lower speech quality, more robust or higher speech quality, less robust terminal in order to accommodate more benign or more hostile channel conditions, while using the same system bandwidth.

Section 2 briefly describes the re-configurable transceiver scheme, Section 3 details the proposed dual-rate ACELP codec, followed by a short discussion on bit sensitivity issues in Section 4, on modulation as well as packet reservation multiple access (PRMA) in Section 5 and source-matched embedded error protection in Section 6. Finally, before concluding, the system performance is characterised in Section 7.

2 System Architecture

As seen in **Figure 1**, a Voice Activity Detector (VAD) similar to that of the Pan-European GSM system [5] enables or disables the dual-rate 4.7/6.5 kbits/s (kbps) Algebraic Code Excited Linear Predictive (ACELP) encoder [6, 10] and queues the active speech frames in the Packet Reservation Multiple Access (PRMA) [7, 9] contention buffer for transmission to the base station (BS). The speech frames are mapped according to their error sensitivities to n protection classes and source sensitivity-matched binary Bose-Chaudhuri-Hocquenghem (BCH) encoded [8].

After BCH encoding the 4.7 and 6.5 kbps speech bits they are converted to 4- or 6-bit symbols, which modulate a re-configurable coherent 16- or 64-level diversity- and Pilot Symbol Assisted Quadrature Amplitude Modulation (PSAQAM) scheme [9]. In contrast to coded modulation [2], where the expanded constellation is used to accommodate parity bits for forward error correction (FEC) coding, here we exploit the expanded 64-QAM constellation to support

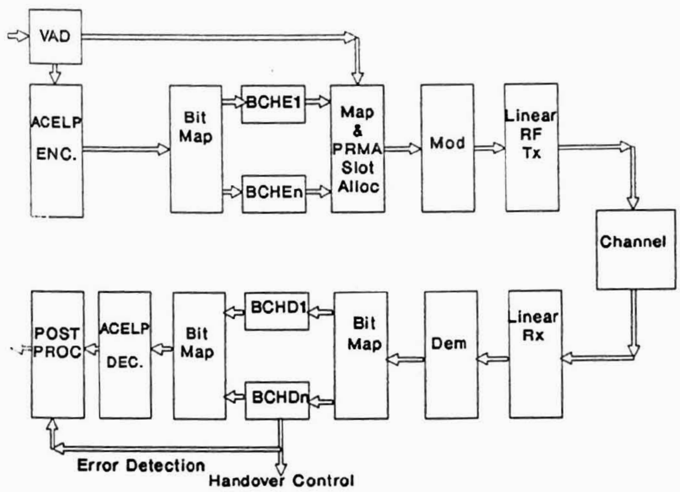

Figure 1: Transceiver Schematic.

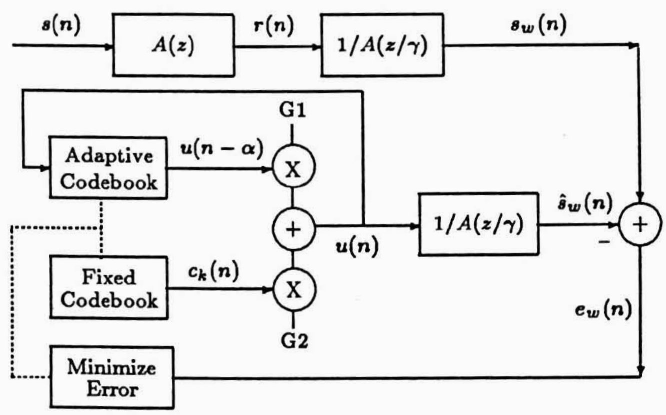

Figure 2: ACELP Encoder Structure

the transmission of higher quality 6.5 kbps-coded speech within the same bandwidth. This re-configurable scheme ideally suits a PRMA-assisted system, which constantly classifies its idle time-slots as top-grade, medium-grade and low-grade slots on the basis of the interference received within their duration. Then the higher speech quality but less robust 6.5 kbps/64-QAM stream is transmitted over top-grade time-slots, 4.7 kbps/16-QAM is sent using medium-grade slots, while the interfered low-grade slots can be disabled. Furthermore, the more robust 4.7 kbps/16-QAM mode can be activated for example upon leaving a benign pico-cellular indoors cell, when the portable station (PS) is handed over to a less friendly street micro-cell.

3 The Dual-rate Speech Codec

The schematic of the proposed dual-rate speech codec is shown in **Figure 2**. In code excited linear predictive (CELP) codecs a Gaussian process generated by a fixed codebook and having a slowly varying power spectrum controlled by the fixed codebook gain G_2 is superimposed on the output signal $u(n-\alpha)$ of the adaptive codebook in order to construct the excitation signal $u(n)$. The weighted synthetic speech waveform $\hat{s}_w(n)$ is generated by filtering the excitation $u(n)$ through the time-varying weighted synthesis filter $W(z) = 1/A(z/\gamma)$ [10]. Then $\hat{s}_w(n)$

Bit Index	Parameter	No. of Bits
1 to 34	LSFs	34
35 to 41	Adaptive Code-book Delay	7
42 to 44	Adaptive Code-book Gain	3
45 to 56	Fixed Code-book Index	12
57	Fixed Code-book Gain Sign	1
58 to 61	Fixed Code-book Gain	4
	Total per Sub-segment	27

Table 1: Bit Allocation Scheme

is compared to the weighted original speech $s_w(n)$ and the optimum combination of the fixed and adaptive codebook entries, minimising the weighted mean squared error is encoded for transmission to the decoder.

The adaptive codebook parameters are determined first under the assumption of $c_k = 0$, since c_k is not yet known. In order to proceed with the determination of c_k let $x(n)$ be the weighted original speech after removing the memory contribution of the filter $W(z)$ from previous frames and $h(n)$ be the impulse response of the filter $W(z)$. Then the mean squared weighted error (mswe) between the original and synthesised speech is given by:

$$E = \sum_{n=0}^{N-1} [x(n) - G_2 c_k(n) * h(n)]^2, \tag{1}$$

where * represents convolution. Setting $\partial E/\partial G_2 = 0$ leads to the mean square weighted error expression [10]:

$$E_{min} = \sum_{n=0}^{N-1} x^2(n) - \frac{\left[\sum_{i=0}^{N-1} \psi(i) c_k(i)\right]^2}{\sum_{i=0}^{N-1} c_k^2(i)\phi(i,i) + 2\sum_{i=1}^{N-2}\sum_{j=i+1}^{N-1} c_k(i)c_k(j)\phi(i,j)}, \tag{2}$$

where $\psi(i)$ represents the correlation between the impulse response $h(n)$ and the signal $x(n)$, given by $\psi(n) = x(n) * h(-n)$, while $\phi(i,j)$ represents the covariances of $h(n)$:

$$\phi(i,j) = \sum_{n=0}^{N-1} h(n-i)h(n-j). \tag{3}$$

The best innovation sequence is constituted by that codebook entry c_k with index k ($k = 1 \dots L$), which minimises the mean squared weighted error in *Equation 2*.

Since $\psi(n)$ and $\phi(n)$ are computed outside the error minimisation loop, the computational complexity is predetermined by the number of operations needed to evaluate the second term of *Equation 2* for all the codebook entries. In our transceiver we favour an algebraic CELP (ACELP) speech codec [6], which has a high innate robustness against channel errors and due to its algorithmic efficiency it allows us to use a codebook with 4096 entries. Hence it guarantees high speech quality at a much lower complexity than conventional CELP codecs.

Furthermore, the ACELP codec can be readily configured as a dual-rate codec. The excitation signal $u(n)$ is determined for each 5 or 7.5 ms subsegment of a 30 ms speech frame, depending on the targetted output bit rate and it is described in terms of the parameters listed in **Table 1**. Explicitly, a total of 27 bits are needed to represent the excitation signal $u(n)$ for each 5 or 7.5 ms subsegment. Thus for the low-rate mode we have a total of (34+4· 27)=142 bits per 30 ms

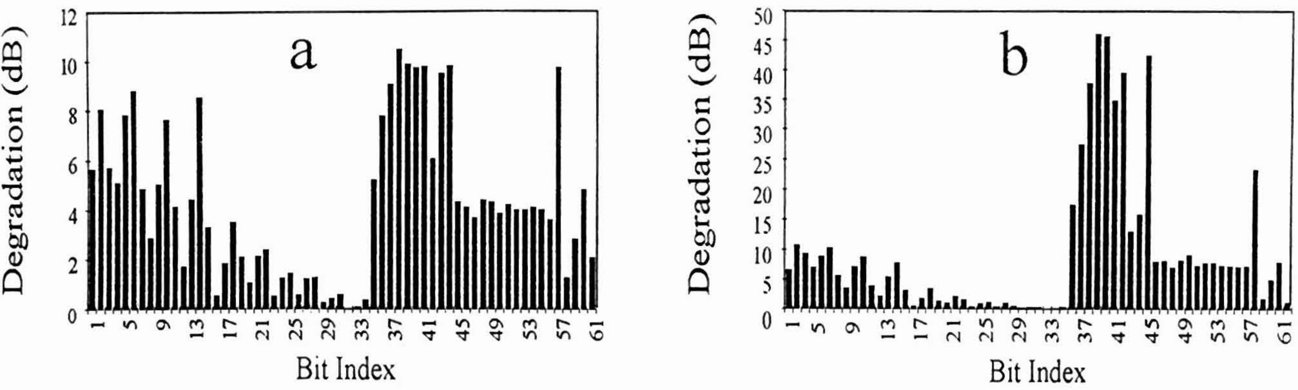

Figure 3: SEGSNR Degradations a) due to 100% Bit Error Rate and, b) due to Single Errors in Various Bits

frame, or a rate of about 4.7 kbits/s, while in the high rate mode a total of $(34+6 \cdot 27)=196$ bits per 30 ms frame, yielding a data rate of just over 6.5 kbits/s.

Having described the dual-rate ACELP codec let us now briefly focus our attention on its robustness against channel errors.

4 ACELP Error Sensitivity

In a first step towards achieving the highest possible system robustness we exploited the ordering property of the LSFs [10] and smoothed the evolution of all the parameters that have predictable time-domain behaviour. A commonly used approach of quantifying the error sensitivity of various coded bits is, to consistently invert a specific bit in every frame and evaluate the SEGSNR degradation. The error sensitivity for our 4.7 kbits/s ACELP codec measured in this way is shown in **Figure 3a** for the 34 LSF bits and the 27 first sub-frame bits. The bit allocation scheme used is shown in **Table 1**. The problem with this approach of quantifying error sensitivity is that it does not take adequate account of the different error propagation properties of different bits. This means that if, instead of corrupting a bit in every frame it is corrupted randomly with some error probability, then the relative sensitivity of different bits will change.

Therefore we propose another error sensitivity measure, which takes account of the error propagation effects. Accordingly, for each bit we find the average SEGSNR degradation due to a single bit error both in the frame in which the error occurs and in consecutive frames. The total SEGSNR degradation is then found by adding or integrating the degradations caused in the erroneous frame and in all the following frames which are affected by the error. **Figure 3b** demonstrates the total SEGSNR degradations evaluated in this way. The graph is significantly different from that in **Figure 3a**. In particular, the importance of the adaptive code-book delay bits due to their memory propagation becomes more explicit. In contrast, the significance of the LSFs is reduced, since they are re-computed for every new frame.

Our final error sensitivity measure is based on the total SEGSNR degradations described above and on a similar measure for the total Cepstral Distance degradation. The two sets of degradation figures are given equal weight and combined to produce an overall sensitivity measure, which is invoked to assign the speech bits to various bit protection classes.

Having documented the proposed speech codec we now concentrate on the modulation aspects of our transceiver.

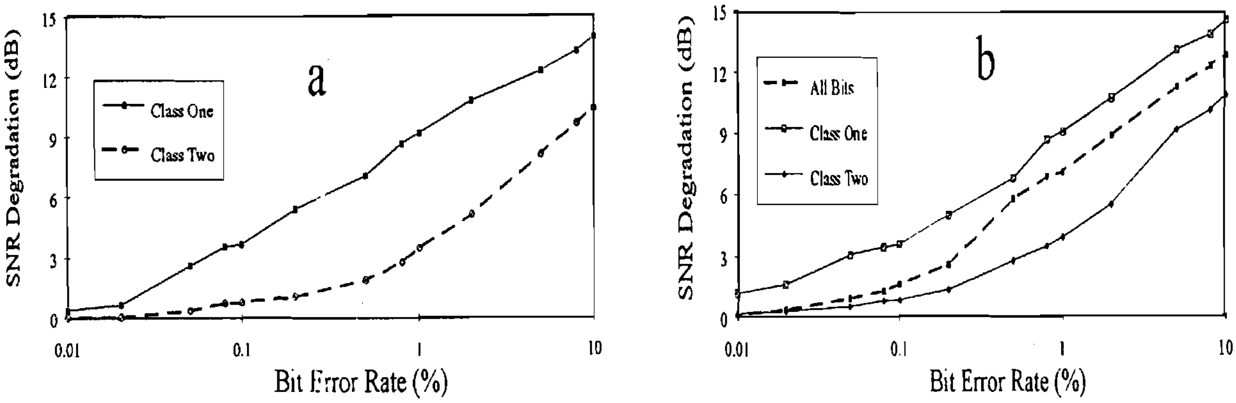

Figure 5: SEGSNR Degradation versus Bit Error Rate for the 4.7 kbps ACELP codec a) when mapping 71 ACELP bits to both Classes One and Two and, b) when mapping 57 and 85 ACELP bits to Classes One and Two, respectively

For the full-class system we have decided to use the approximately half-rate BCH(127,71,9) codec in both subchannels, which can correct 9 errors in each 127-bits block, while encoding 71 primary information bits. The coding rate is R=71/127 $\approx$ 0.56 and the error correction capability is about 7 %. Observe that this code curtails error propagation at the speech frame boundaries by encoding each 142-bit speech frame using two BCH(127,71,9) frames.

In order to design the twin-class system, initially we divided the ACELP bits in two sensitivity classes, Class One and Class Two, and evaluated the SEGSNR degradation inflicted by certain fixed channel BERs maintained in each of the classes using randomly distributed errors, while keeping bits of the other class intact. These results are shown in **Figure 5a**. Observe that an approximately ten-fold lower BER is required by the more vulnerable Class One bits in order to limit the SEGSNR degradations to values similar to those caused by the more robust Class Two bits. Recall from **Figure 4** that the 16-QAM C1 and C2 subchannel BER ratio was limited to about a factor three. Hence we decided to employ a stronger FEC code to protect the Class One ACELP bits transmitted over the 16-QAM C1 subchannel than for the Class Two speech bits conveyed over the lower integrity C2 16-QAM subchannel, while maintaining the same number of BCH-coded bits in both subchannels. However, the increased number of redundancy bits of stronger BCH codecs requires that a higher number of sensitive ACELP bits are directed to the lower integrity C2 16-QAM subchannel, whose coding power must be concurrently reduced in order to accommodate more source bits. This non-linear optimisation problem can only be solved experimentally, assuming a certain sub-division of the source bits, which would match a given pair of BCH codecs.

Based on our previous findings we then decided to increase the C1-C2 16-QAM subchannel BER discrepancy by about a factor of two so that the Class One ACELP bits are guaranteed a BER advantage of about a factor of eight over the more robust Class Two bits. After some experimentation we found that the BCH(127,57,11) and BCH(127,85,6) codes employed in the C1 and C2 16-QAM subchannels provided the required integrity. The SEGSNR degradation caused by a certain fixed BER assuming randomly distributed errors is portrayed in **Figure 5b** for both the full-class and the above twin-class system, where the number of ACELP bits in the protection classes One and Two is 57 and 85, respectively. Note that the coding rate of this system is the same as that of the full-class scheme and each 142-bit ACELP frame is encoded by two BCH codewords, yielding again $2 \cdot 127$=254 encoded bits and curtailing error propagation at the block boundaries. The FEC-coded bit rate became $\approx$8.5 kbps.

With the incentive of perfectly matching the FEC coding power and the number of bits in

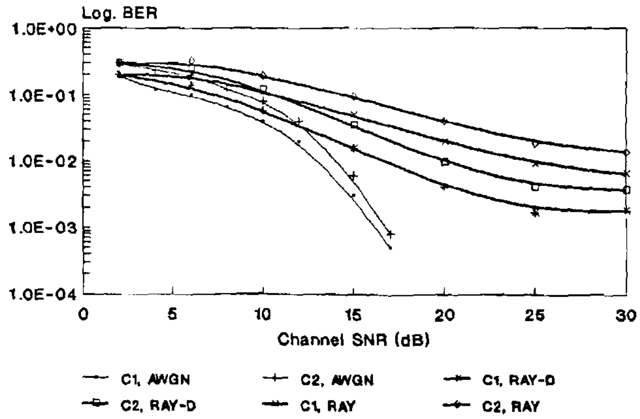

Figure 4: C1 and C2 BER versus channel SNR performance of 16-QAM over Rayleigh and Gaussian channels

5 Re-configurable Modulation and Multiple Access

The different source rates of the 4.7 and 6.5 kbits/s ACELP operating modes will be equalised using a combination of appropriately designed FEC codecs and 4 bits/symbol or 6 bits/symbol modulators. When the channel signal-to-noise ratio (SNR) and signal-to-interference ratio (SIR) are high, 64-level Quadrature Amplitude Modulation (64-QAM) is used to convey the 196 bits of the 6.5 kbps ACELP codec, while for worse channel conditions the 142 bits of the lower quality 4.7 kbps codec are delivered by a more robust 16-QAM modem in the same bandwidth, as the 64-QAM scheme.

Non-coherent QAM modems are less complex to implement than their coherent counterparts, but typically require higher SNR and SIR values. Hence in our proposed scheme second-order switched-diversity assisted coherent Pilot Symbol Assisted Modulation (PSAM) using the maximum-minimum-distance square QAM constellation is preferred [9]. For the 16-QAM scheme we have shown [9] that it has two independent subchannels exhibiting different integrities, depending on the position of the bits in a four-bit symbol. On the same note, our 64-QAM modem posesses three different subchannels [9] having different bit error rates. This property naturally lends itself to un-equal error protection, if the source sensitivity-matched integrity requirements are satisfied by the QAM subchannel integrities.

Therefore initially we have evaluated the bit error rate (BER) versus channel signal-to-noise ratio (SNR) performance of our 16-QAM modem for both subchannels using no forward error correction coding (FEC) with and without second-order diversity. The C1 and C2 BER results are shown in **Figure** 4 for the experimental conditions characterised by a pedestrian speed of 3 mph, propagation frequency of 1.8 GHz, pilot symbol spacing of P=10 and a signaling rate of 100 kBd. As a benchmarker, also the Gaussian performance is shown. In case of this signaling rate the modulated signal will fit in a bandwidth of 200 kHz, when using a unity roll-off factor.

Observe that for channel SNRs around 15-20 dB there is an approximately factor three BER difference between the two subchannels. How this BER difference between the two subchannels can be exploited in order to provide source-matched FEC protection for the 4.7 kbps ACELP codec will be described in the next Section. Following a similar approach for the 6.5 kbps/64-QAM scheme leads to the higher speech quality operating mode of our transceiver, which can be invoked under more favourable channel conditions typically encountered in indoors pico-cells.

In conventional Time Division Multiple Access (TDMA) mobile systems the grade of service (GOS) degrades due to speech impairments caused by call blocking, hand-over failures and speech frame interference engendered by noise, as well as co- and adjacent-channel interference. In PRMA-assisted systems calls are not blocked due to the lack of an idle time-slot. Instead, all users becoming active are allowed to contend for a slot reservation and hence the packet dropping probability is increased gracefully. Hand-overs will be performed in the form of contention for an idle time slot provided by the specific BS offering the highest signal quality amongst the potential target BSs.

The specific physical up-link to the BS offering the best signal quality during decoding the packet header is not likely to substantially degrade during the life-time of an active speech spurt having a typical mean duration of 1s or about thirty consecutive 30ms speech frames. If, however the link degrades before the next active spurt is due for transmission, the subsequent contention phase is likely to establish link with another BS. Hence this process will have a favourable effect on the channel's quality, effectively simulating a diversity system having independent fading channels and limiting the time spent by the MS in deep fades, thereby avoiding channels with high noise or interference.

This advantageous PRMA property can be exploited in order to maintain low interference levels by training a self-adjusting adaptive system using the channel segregation scheme proposed for PRMA systems in reference [11]. The BS will compile a time-slot quality list on the basis of monitoring idle time-slots as well as the packet header decoding success ratio of active slots and time-slots will be classified according to their quality. This channel quality information can be broadcast to the portable stations (PSs). If there are sufficiently high-quality idle slots available, the PS will use its higher speech quality, less robust 64-QAM mode in its attempt to acquire a reservation, but if no high-quality channels are available, the PS will invoke its lower speech quality 16-QAM mode of operation.

6 Sensitivity-matched Forward Error Correction

6.1 Robust Mode

Recent advances in coded modulation and bandwidth efficient transmission published in reference [2] provide a plethora of attractive transceiver schemes for fading mobile channels. However, for our ACELP codec source sensitivity-matched un-equal error protection has to be used. Hagenauer introduced rate compatible punctured convolutional (RCPC) codecs[12] in order to perfectly match the FEC power to the sensitivity of the source bits. In order to provide coded modulation schemes having un-equal source-sensitivity matched error protection similarly to our approach in reference [9], Wei [3] suggested a range of non-uniformly spaced phasor constellations. Another method proposed by Wei [3] was to deploy a number of independent trellis coded modulation (TCM) schemes having different grade of protection and multiplex the sequences for transmission.

In this Section we will exploit the QAM subchannel integrity difference highlighted in the previous Section in **Figure 4** and combine these subchannels with appropriate binary Bose-Chaudhuri-Hocquenghem (BCH) FEC codecs. The error sensitivity of the 4.7 kbps ACELP source bits was evaluated in **Figures 3a** and **3b**, but the number of bit protection classes n still remains to be resolved. Intuitively, one would expect that the more closely the FEC protection power is matched to the source sensitivity, the higher the robustness. In order to limit the system's complexity and the variety of candidate schemes, in case of the 4.7 kbits/s ACELP codec we have experimented with a full-class BCH codec, a twin-class and a quad-class scheme, while maintaining the same coding rate.

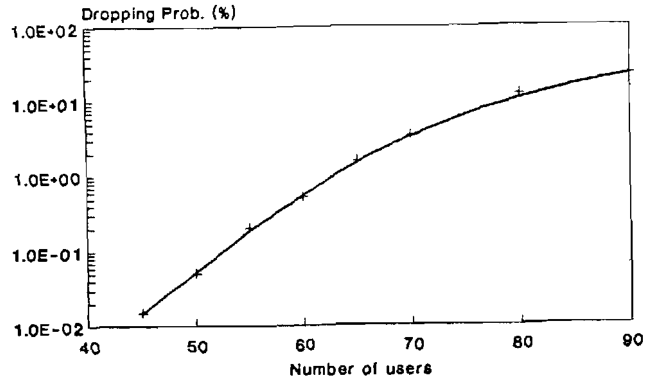

Figure 6: Packet dropping versus number of users performance of the proposed system

the distinct protection classes to the ACELP source sensitivity requirements we also designed a quad-class system, while maintaining the same coding rate. We used the BCH(63,24,7), BCH(63,30,6), BCH(63,36,5) and BCH(63,51,2) codes and transmitted the most sensitive bits over the C1 16-QAM subchannel using the two strongest codes and relegated the rest of them to the C2 subchannel, protected by the two weaker codes.

The PRMA [7] control header was allocated a BCH(63,24,7) code in case of all schemes and hence the total PRMA framelength became 317 bits, representing 30 ms speech and yielding a bit rate of $\approx$ 10.57 kbps. The 317 bits constitute 80 16-QAM symbols and 9 pilot symbols as well as 2+2=4 ramp symbols must be added, resulting in a PRMA framelength of 93 symbols per 30 ms slot. Hence the signaling rate becomes 3.1 kBd. Using a PRMA bandwidth of 200 kHz, similarly to the Pan-European GSM system [5], and a filtering excess bandwidth of 100 % allowed us to accommodate 100 kBd/3.1 kBd $\approx$ 32 PRMA slots. The packet dropping probability versus number of users curve of our system is shown in **Figure 6**. Observe that more than 60 users can be supported at a packet dropping probability of 1%.

6.2 High-quality Mode

Following the approach proposed in the previous Subsection we designed a triple-class source-matched protection scheme for the 6.5 kbps ACELP codec. The reason for using three protection classes this time is that the 6.5 kbps ACELP codec's higher bit rate must be accommodated by a 64-level QAM constellation, which inherently provides three different subchannels [9]. When using second-order switched-diversity and pilot-symbol assisted coherent square-constellation 64-QAM amongst our previously stipulated propagation conditions with a pilot-spacing of P=5 best matched codes for the C1, C2 and C3 subchannels are BCH(126,49,13), BCH(126,63,10) and BCH(126,126,84,6). The packet header was allocated again a BCH(63,24,7) code. The total number of BCH-coded bits becomes $3 \cdot 126 + 63 = 441/30$ ms, yielding a bit rate of 14.7 kbps. The resulting 74 64-QAM symbols are amalgamated with 15 pilot and 4 ramp symbols, giving 93 symbols/30 ms, which is equivalent to a signaling rate of 3.1 kBd, as in case of the low-quality mode of operation. Again, 32 PRMA slots can be created, as for the low-quality system, accommodating more than 60 speech users in a bandwidth of 200 kHz and yielding a speech user bandwidth below 4 kHz, while maintaining a packet dropping probability of about 1 %.

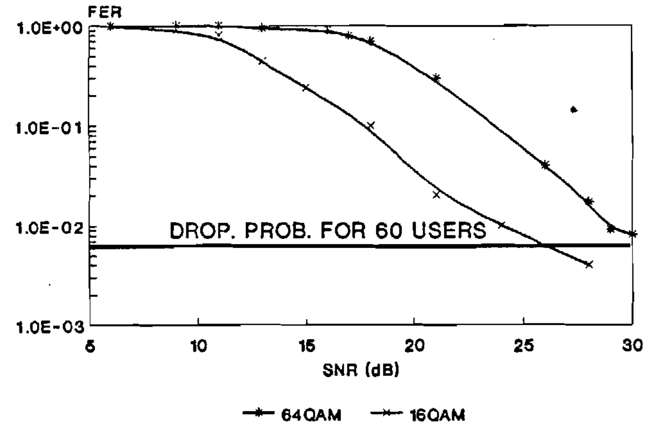

Figure 7: FER versus channel SNR performance of the proposed 3.1 kBd transceiver in both of its operating modes

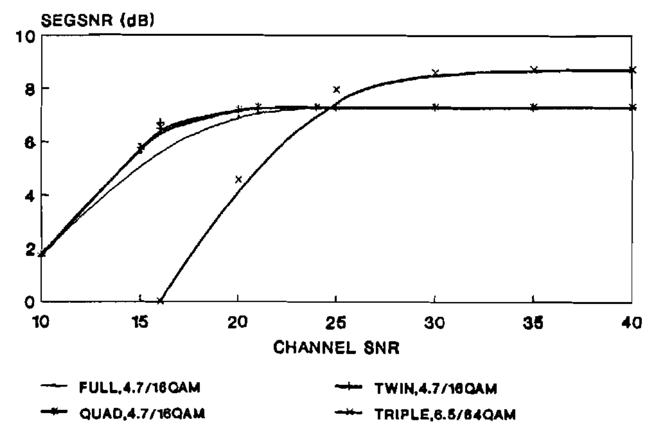

Figure 8: SEGSNR versus channel SNR performance of the proposed 3.1 kBd transceiver

7 System Performance and Conclusions

The PRMA packet dropping probability curves of **Figure 6** can now be contrasted with the channel-induced frame error rate (FER) curves of **Figure 7** in order to study the speech impairments inflicted. The overall SEGSNR versus channel SNR performance of the proposed speech transceiver is displayed in **Figure 8** for the various systems studied under low traffic loading, when the number of users is below 50, implying a packet dropping probability of less than 0.1%. Observe that the source sensitivity-matched twin-class and quad-class 4.7 kbps ACELP-based systems have a virtually identical performance, suggesting that using two appropriately matched protection classes provides adequate system performance, while maintaining a lower complexity than the quad-class scheme. The full-class system was outperformed by both source-matched schemes by about 4 dB in terms of channel SNR, the latters requiring an SNR in excess of about 15 dB for nearly un-impaired speech quality over our pedestrian Rayleigh-fading channel. When the channel SNR was in excess of about 25 dB, the 6.5 kbps/64-QAM system outperformed the 4.7/16-QAM scheme in terms of both objective and subjective speech quality. Non-interfered high-SNR time-slots will be dedicated to the operation of the higher speech quality mode, while the channel segregation algorithm will relegate low-quality slots for use by the lower speech quality, more robust mode of operation. Both systems have a single-user rate of 3.1 kBd, accommodate 32 PRMA slots at a PRMA rate of 100 kBd in a bandwidth of 200 kHz and can support in excess of 50 users. The effective user bandwidth becomes 4 kHz.

8 Acknowledgement

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