QoS-Oriented Solutions for Satellite Broadcasting Systems

Aharon Vargas, Wolfgang H. Gerstacker, and Marco Breiling

Abstract: **In this paper, we analyze the capability of satellite broadcasting systems to offer different levels of quality of service (QoS). We focus on the European telecommunications standards institute satellite digital radio and digital video broadcasting satellite handheld (DVB–SH) standards, which have recently been proposed for satellite broadcasting communications. We propose a strategy to provide different levels of QoS for the DVB–SH standard on the basis of an extension of the interleaving scheme, referred to as molded interleaver, which supports low latency service requirements for interactive services. An extensive analysis based on laboratory measurements shows the benefits of this solution. We also present a multilevel coding (MLC) scheme with multistage decoding designed for broadcasting communications as an alternative to the existing standards, where services with different levels of QoS are provided. We present a graphical method based on mutual information for the design and evaluation of MLC systems used for broadcasting communications. Extensive simulations for a typical satellite channel show the viability of the proposed MLC scheme. Finally, we introduce multidimensional constellations in the proposed MLC scheme in order to increase the number of different protection levels.**

Index Terms: **Hierarchical modulation, multilevel coding (MLC), quality–of–service (QoS), satellite broadcasting.**

I. INTRODUCTION

Satellite broadcasting communication systems with high throughput have increased in popularity in recent years. In particular, some standards have been recently released, e.g., the European telecommunications standards institute (ETSI) satellite digital radio (ESDR) standard [5], [6], and the digital video broadcasting satellite handheld (DVB–SH) standard [7]. A special requirement is becoming very important for these systems: Configurable content–dependent quality of service (QoS). It is desirable to have different levels of QoS for different content when multimedia information is broadcast. The configuration of the QoS is normally a task of the transport layer. However, the physical layer can also provide different levels of QoS in terms of different protection levels, e.g., by splitting the services in different data pipes so that each service has a different protection level. The protection can be provided by three different mechanisms:

a) A configurable code rate in the forward error correction (FEC) coding block,
b) the duration of the channel interleaver, and
c) hierarchical modulation which provides implicitly different levels of protection for each bit fed into the mapper.

The ESDR and the DVB–SH standards exploit some of the mechanisms described above to provide a configurable level of QoS in terms of robustness of transmission. Besides, the ESDR standard uses different data streams in a single transmission multiplex, which can be individually interleaved and protected. This automatically guarantees different levels of QoS by configuring the data streams in different ways. On the other hand, the DVB–SH standard supports a unique code rate and interleaver configuration for all kinds of data, unless hierarchical modulation is employed. In this case, only two different levels of protections are provided, and still a unique interleaver profile is supported.

Alternatively, the required QoS can be defined in terms of tolerable transmission delay. For non–real–time (NRT) applications, the QoS demands might be still fulfilled for a high transmission delay, but for real–time (RT) communications, a high transmission delay has a significantly adverse impact on the QoS. The DVB–SH standard does not provide a configurable transmission delay for information streams, which prohibits the simultaneous transmission of NRT and RT content.

We investigate the problem of providing different levels of QoS in a DVB–SH environment in terms of robustness and transmission delay. A molded interleaver is introduced as a solution which provides a second level of QoS with high flexibility, and it can be easily integrated with the current DVB–SH standard. This approach is validated through extensive laboratory measurements. We also propose a system concept where multilevel codes are used in combination with high performance FEC blocks. A previous analysis of multilevel coding (MLC) schemes can be found in [8], where each level of MLC is evaluated independently by considering equivalent channels and the FEC block is assumed to be ideal. We propose a novel way to characterize the decoding capability of practical FEC blocks, making use of the concept of extrinsic information transfer (EXIT) charts [9], and determine the optimal operation point of each level. In addition, a graphical method involving the use of multistage decoding (MSD) charts is presented for designing the levels, where the available and maximum decoding capacity of the receiver are shown. Finally, multidimensional constellations are introduced in an MLC scheme in order to increase the number of different protection levels.

The remainder of this paper is structured as follows. Section II presents a system model for ESDR and DVB–SH and an analysis of mechanisms that can be used to provide different levels of QoS. Section III focuses on the DVB–SH standard. A backward compatibility solution is introduced which provides an extra level of QoS. An MLC scheme suitable for broadcasting communications is discussed in Section IV, where a method to design and evaluate the MSD receiver is presented. Finally, some conclusions are stated in Section V.

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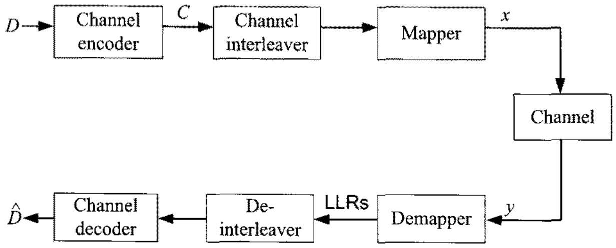

Fig. 1. Typical transmission model for satellite broadcasting communications, involving an encoder, a channel, and a decoder.

II. SYSTEM MODEL

Fig. 1 shows a simplified model of the physical layer of a typical satellite broadcasting system, including the transmitter, the channel, and the receiver. It is based on the DVB–SH implementation guidelines [10] and the ESDR standard [5], where only the blocks most relevant for our analysis are depicted: A channel encoder, a convolutional interleaver, and a mapper. Both standards employ a turbo encoder [11], which supports code rates from 1/5 to 2/3 (DVB–SH) or 3/4 (ESDR). A data stream D is encoded and the output codewords C are interleaved using a convolutional interleaver which provides durations from a few milliseconds to tens of seconds. Coded bits are grouped and mapped onto a unique point in a constellation, which is determined by the modulation format. At the receiver, the symbols are demapped, and the a–posteriori probabilities (APPs) of the codeword symbols based on the received signal y can be obtained. The APPs are represented as conditional log–likelihood ratios (LLRs), which are de–interleaved and fed to the FEC block. This scheme corresponds to a bit–interleaved coded modulation (BICM) [12] structure with no iterations between the demapper and the FEC block, as proposed in [10].

The three blocks presented above (channel encoder, channel interleaver, and mapper) can provide different levels of QoS in terms of robustness and delay for different code rates and interleaver lengths if they are configured properly. The rest of this section analyzes how the DVB–SH and the ESDR standard exploit the capability of each block to offer different levels of QoS.

A. Channel Encoder and Code Rate

The FEC block adds some redundancy to the information for protection. The protection level is determined by the amount of redundancy, which is in turn determined by the code rate of the encoder. FEC codes with a configurable code rate, for example, codes generated by puncturing a mother code (third generation partnership project 2 (3GPP2) turbo code [13]), lead to different levels of protection. A drawback of increasing the redundancy is that the amount of useful information that can be transmitted decreases.

The ESDR standard splits the data in different pipes, which are protected using different code rates. In this case, different levels of QoS can be achieved for each pipe, thanks to the wide range of code rates offered by the FEC block.

The DVB–SH standard also offers a configurable code rate, but all data must be protected using the same code rate. The structure of different pipes for each information type is not defined, which means that the FEC block cannot provide different levels of QoS for this standard.

B. Channel Interleaver and Duration

A typical drawback in satellite systems is the frequent absence of line of sight (LOS) because of obstructions such as buildings or bridges. In this case, the signal might be lost, resulting in long dropouts, even if a low code rate is used. A long channel interleaver can mitigate this effect by spreading the codewords over the time in order to restrict the loss during a dropout to only some parts of the codewords. The missing information could be recovered by the FEC block for a certain erasure probability if an appropriate code rate is employed. Therefore, the channel interleaver helps avoid long periods of erroneous decoding, thus increasing the reliability and QoS of the system. The duration of channel interleaving determines the protection level of the data, leading to different levels of QoS. However, an increase of the duration entails two major drawbacks: An increase in the complexity of the receiver in terms of memory, and a higher latency due to the decoding of the information, which can be unacceptable in RT applications.

The ESDR standard allows each pipe to be interleaved with a different interleaver profile, with durations ranging from a few milliseconds to tens of seconds. For example, important information can be transmitted in a pipe with an interleaver duration of 20 seconds and a code rate of 1/5, while secondary information can be protected using short interleaving and a code rate of 3/4.

A configurable interleaver is also supported by the DVB–SH standard, but all the information has to be interleaved in the same way. There is no distinction between different types of information, implying that only one level of QoS is possible.

C. Mapper and Hierarchical Modulation

The use of hierarchical modulation provides inherent unequal protection to the different bits fed into the mapper. For example, if anti–Gray mapping (unequal bit protection) is used in quadrature phase shift keying (QPSK) modulation, one bit is better protected than the other one [14], which leads to different levels of QoS. In principle, the number of different protection levels depends on the modulation scheme, for example, M–ary amplitude phase shift keying (M–APSK) modulation provides $\log_2(M)$ different protection levels. However, some modulated bits could exhibit the same protection level, as shown in [4]. Note that the hierarchical modulation technique can be combined with different code rates or interleaver durations, adding one extra configuration parameter when setting the QoS.

The ESDR standard does not specify any hierarchical modulation, and therefore, the adjustment of the protection level is solely through the selection of an appropriate code rate and interleaver duration.

The DVB–SH standard specifies a hierarchical modulation for a 16–ary quadrature amplitude modulation (16–QAM), where only two different levels of protection are possible [2] referred to as "hi–prio" and "low–prio" respectively. Independent code rates for each stream can be used, but a unique interleaver profile is supported. Besides, three different versions of the constellation are defined depending on a factor α, which indicates the spread of the signal points. Higher values of α lead

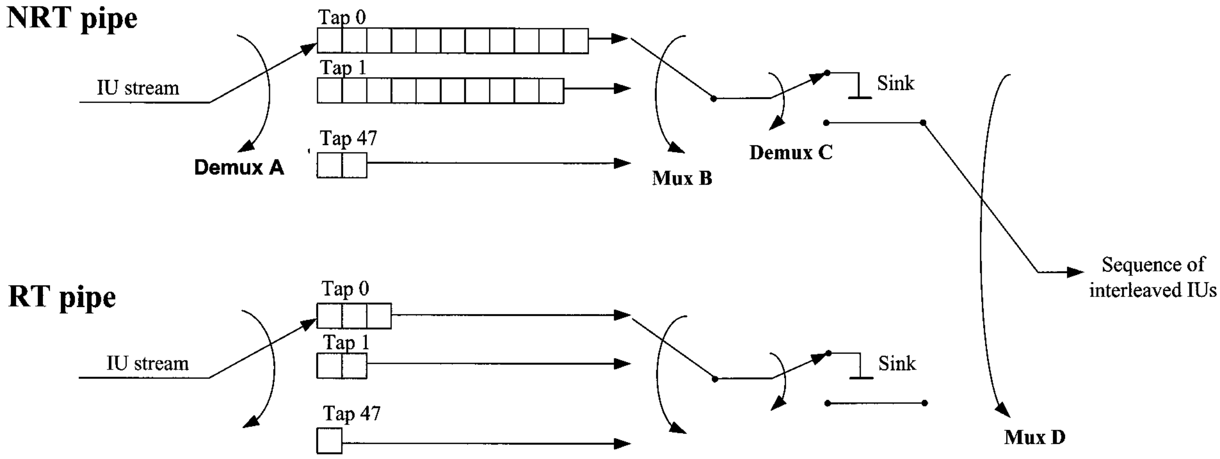

Fig. 2. Multiplexing in the RT pipe and the NRT pipe.

to a higher separation in terms of protection between the hi–prio and low–prio streams. However, if the spread is too large, the channel capacity decreases [1].

III. ADDING LEVELS OF QOS TO THE DVB–SH STANDARD

In this section, we focus on DVB–SH transmission systems, where new types of services are becoming attractive, for example, traffic congestion management, or the inclusion of a return link for the transmission of short messages in order to interact with the forward link broadcast services. This kind of information requires RT transmission (low latency), which is in conflict with the requirement of high robustness for conventional NRT broadcast data, which is achieved by a long interleaver. The same requirements appear for other types of RT services such as bidirectional voice services. As stated in subsection II-B, data transmission with two different interleaver profiles is not possible in the current DVB–SH standard.

We propose a solution based on the concept of different pipes, which was introduced in [5]. We define a new pipe that carries the RT information and is embedded in the NRT information stream. Fig. 2 shows the principle underlying this concept, where the DVB–SH multiplex signal is split into two sub–multiplexes (pipes), which are combined in a time–multiplexed manner by multiplexer D. Each pipe has its own code rate and interleaver profile. The NRT data can be protected using a long interleaver, while a sufficiently short interleaver is employed for the RT data. The codewords are cut into 126–bit parts, named interleaver units (IUs), which are delayed by a convolutional interleaver using a structure of 48 taps [7]. The interleaver process is performed by the de–multiplexer A and the multiplexer B. In order to maintain backward compatibility with the current DVB–SH standard, the positions of the IUs of the NRT pipe inside the DVB–SH multiplex must be the same as for the legacy waveform, and the RT pipe is only allowed to fill the gaps between those, i.e., it must be embedded into the stream defined by the NRT pipe. To this end, dummy codewords are inserted as placeholders for the RT information, and they are removed by de–multiplexer C. The legacy receivers cannot distinguish between the two pipes. This means that they use the configuration of the NRT pipe to decode the RT information, which obviously leads to erroneous decoding. The DVB–SH standard provides a mechanism to handle these codewords properly and to discard them.

Due to the backward compatibility requirement, the interleaver profile of the RT pipe cannot be chosen arbitrarily. A very short interleaver profile is derived from the long interleaver profile of the NRT pipe by considering its tap delays modulo the number of interleaver cycles per frame. Using this strategy, the IUs of the RT pipe are molded between the positions of the IUs of the NRT pipe, filling the gaps left for this pipe. The modulo operation also guarantees that the length of the RT pipe is less than one DVB–SH frame, which matches with the requirement of a low latency because a typical DVB–SH frame has a duration around a few hundreds of milliseconds. Therefore, the resulting QoS of the RT pipe in terms of delay is fixed. Note that the RT pipe has almost no protection against long dropouts, which reduces its robustness compared to the NRT information. This decrease in robustness can be mitigated by using a very low code rate, which would increase the protection provided to the RT information.

The proposed solution provides a mechanism to combine two different types of data using different code rates, where the interleaver profile of the RT pipe (very short interleaver) is derived from the long interleaver of the NRT pipe. Obviously, the absence of a long interleaver leads to a deterioration in the performance of the RT pipe. In order to quantify this degradation, we describe the satellite channel via the land mobile satellite (LMS) channel model proposed in [15] and [16]. The received signal is given by

$$y(t) = a(t)x(t) + n(t) \tag{1}$$

where $x(t)$ is the continuous–time transmit signal, $a(t)$ corresponds to the complex attenuation factor that depends on the channel, and $n(t)$ is a complex–valued white Gaussian noise process with double–sided power spectral density N_0 in equivalent complex baseband. The variations of $a(t)$ are modeled using states for the different transmission conditions, which can be grouped in line of sight, shadowing, and blockage. The long channel interleaver is necessary to overcome long dropouts pro-

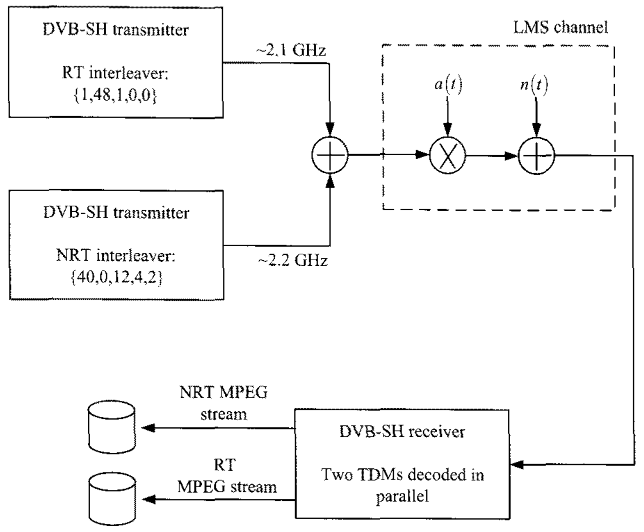

Fig. 3. Laboratory setup for transmission of the RT and NRT pipes.

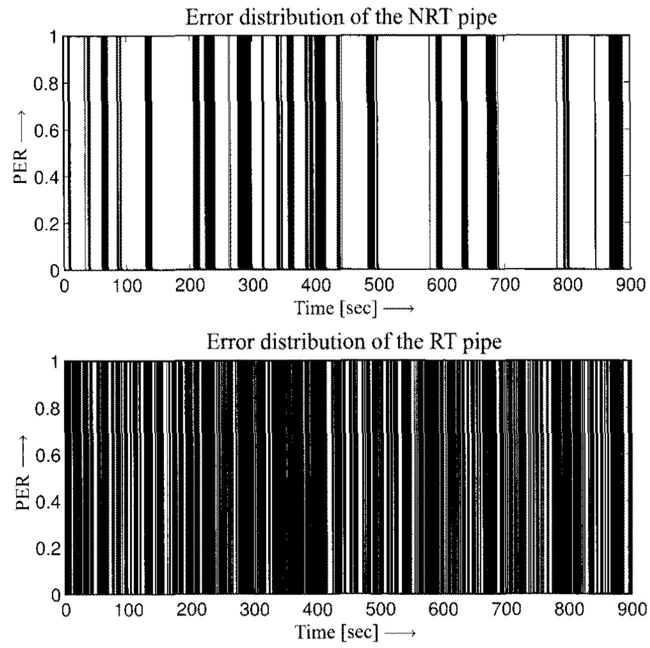

Fig. 4. Error distribution of the RT and NRT pipes.

duced by the blockage state, in which the direct path from the satellite is blocked.

Fig. 3 shows the laboratory setup used to evaluate the RT pipe. It consists of two standard DVB–SH transmitters, an LMS channel generator, and a DVB–SH prototype receiver. The RT pipe cannot be tested by using only one standard DVB–SH transmitter because the simultaneous transmission of NRT and RT data is not supported. However, the receiver supports the parallel reception of two independent DVB–SH streams, which allows us to use the two DVB–SH transmitters that are configured independently.

One of the transmitters outputs the NRT pipe by employing a uniform interleaver profile of a duration of 10 seconds in one of the satellite multiplexes at a central frequency of about 2.2 GHz. The second transmitter emulates the RT pipe by using a very short interleaver profile (with duration less than one DVB–SH frame) at a central frequency of about 2.1 GHz. Both TDM signals are added and a broadband LMS channel is applied to the composite signal. We use the sub–urban scenario proposed in [10] with a velocity of the terminal of 50 km/h, where the channel coefficients $a(t)$ are generated according to a Loo distribution [17] during the shadowing state, and additive white Gaussian noise (AWGN) $n(t)$ is added. A QPSK modulation and a code rate of 1/4 are used in all measurements. The DVB–SH prototype receiver has two independent demodulators in order to receive both satellite multiplexes and decode them in parallel. The output packets are saved and post–processed in Matlab.

Restricting the data carried by the RT pipe to short message communications, we analyze the error distribution over the time. This is shown in Fig. 4, where the error probability of both pipes is shown for 15 minutes for $E_s/N_0 = 1$ dB (E_s: Average received energy per symbol). The NRT pipe is characterized by a few long sequences of error bursts due to the long interleaver, whereas the RT pipe presents many short error bursts. Re–transmission of the same message is required in order to guarantee successful reception, even for good channel conditions. Fig. 5 shows the probability of successful transmission of a codeword for the RT pipe for different numbers of re–transmissions. The relation between the re–transmission period and the probability of success can be also observed. Note that the re–transmissions added additional delay to the RT pipe, which can be unacceptable in certain applications. Increasing the time between re–transmissions leads to a higher probability of success. For a re–transmission period of around one second, the measured probability is close to the theoretical curve, where the re–transmissions are assumed to be uncorrelated.

Note that the introduction of an RT pipe in the DVB–SH standard provides a second level of QoS to the transmitted data and offers backward compatibility. However, the level of protection is low especially for LMS channels, and multiple re–transmissions are required for increasing the reliability of the system causing additional delay. In the following section, we propose an alternative transmission scheme to solve this problem.

IV. MULTILEVEL CODES

Fig. 6 shows a multilevel coding system for satellite broadcasting communications; the system was originally presented in [4]. Typically, a different level of protection is necessary for each service. In the proposed multilevel encoder, a set of symbols of data services $\mathcal{D} = \{D_1, \cdots, D_n\}$ is encoded in parallel, where different QoS is provided by the three mechanisms described in Section II:

a) Independent code rates $R_1, \cdots, R_n$ for each level.
b) Independent interleaver profiles $\pi_1, \cdots, \pi_n$ for each level.
c) Unequal error protection for each input bit to the mapper by using hierarchical modulations.

In this section, we focus on the capability of MLC schemes to provide different protection levels by using appropriate mappings, where the input bits to the mapper are unequally protected. In order to generate a channel symbol x, n bits are read in parallel, one from each codeword C_i, and mapped onto a unique

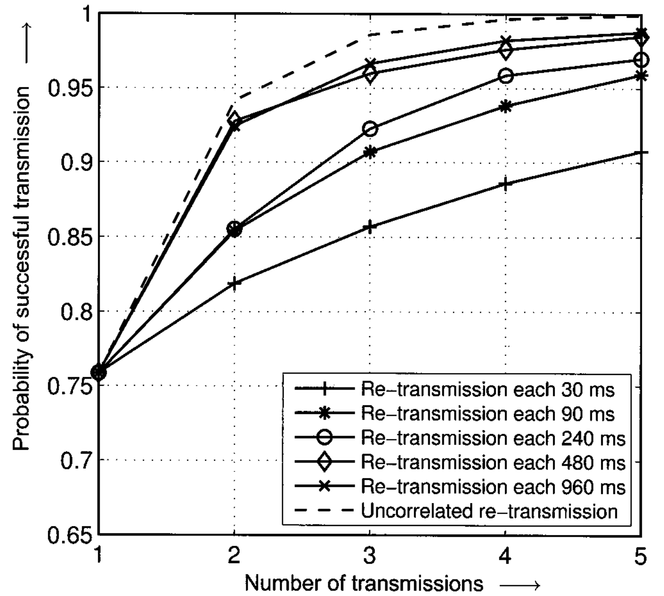

Fig. 5. Rate of successful transmission for the RT pipe. Sub–urban scenario, 50 km/h, $E_s/N_0 = 1$ dB.

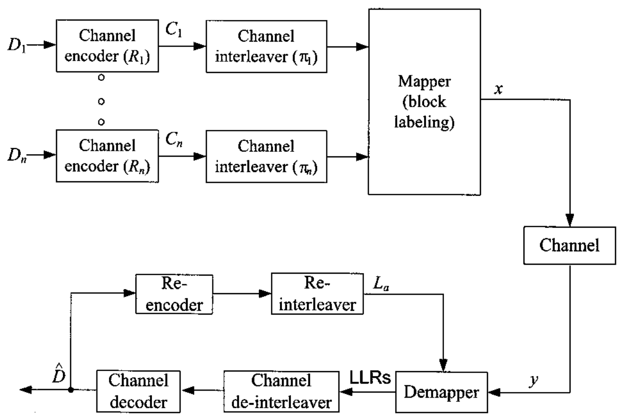

Fig. 6. Multilevel coding system, composed by a multilevel encoder, a channel and a multistage decoder.

constellation point determined by the block labeling (BL) partitioning, cf. also [8]. BL divides the constellation into two subsets that are grouped as clusters; the clusters are well separated from each other. Each cluster is again divided into two sub–clusters, and this process continues iteratively.

An MLC scheme using multistage decoding [18] and BL as the partitioning strategy presents the possibility to reach the channel capacity [8]. In contrast, a classical bit-interleaved coded modulation (BICM) approach with Gray mapping, as proposed in the DVB–SH standard [10], shows a suboptimal behavior [19] because the statistical dependencies between the mapped bits are not considered in the demapping process. Obviously, a properly BL partitioning must be found for a given constellation. Finding this BL partitioning can involve a highly complex computer search, especially when many levels are desirable, because the number of signal points M increases exponentially with the number of levels. A given constellation $\mathcal{A} \in \mathbb{C}^{D/2}$ with D real-valued dimensions and M signal points can be partitioned into two subsets $\mathcal{A}^0$ and $\mathcal{A}^1$, each comprising $M/2$ symbols, in $\binom{M}{M/2}$ different ways, where $\binom{\cdot}{\cdot}$ represents the binomial coefficient. Similarly, each of the subsets $\mathcal{A}^0$ and $\mathcal{A}^1$ can be partitioned into 2 subsets in $\binom{M/2}{M/4}$ possible ways. Iterating $i-1$ times, we can partition a given constellation into subsets on i levels. Note that a brute–force search is infeasible even for moderate values of M due to the large number of combinations. The complexity can be reduced drastically using efficient search algorithms such as the binary switching algorithm (BSA) [20]. The BSA starts with the random partitioning of $\mathcal{A}$ into $\mathcal{A}^0$ and $\mathcal{A}^1$ and iteratively switches the elements of the subsets to reduce a given cost function after every switching.

In BL partitioning, the subsets have to be separated from each other as much as possible so that previous levels have better mutual distances leading to a better protection. We define a cost function on the basis of the minimization of the intra–subset variance [21] as

$$J_a = \frac{2}{MD}\left(\sum_{\forall a_m \in \mathcal{A}^0}\sum_{d}\left(a_{m,d} - r_d^0\right)^2 + \sum_{\forall a_m \in \mathcal{A}^1}\sum_{d}\left(a_{m,d} - r_d^1\right)^2\right) \quad (2)$$

where $d \in \{0, 1, \cdots, D-1\}$ represents the dimension index, and r_d^0 and r_d^1 correspond to the dth real-valued component of the center of mass of the subsets $\mathcal{A}^0$ and $\mathcal{A}^1$, respectively. We propose a second similar criterion basis on the idea that the level protection increases when the separation of the clusters also increases. This criterion can be obtained by maximizing the squared distance between the centers of mass of the subsets r_d^0 and r_d^1 and can be expressed as

$$J_b = \frac{4}{M^2 D}\sum_{d}\left(\sum_{\forall a_m \in \mathcal{A}^0} a_{m,d} - \sum_{\forall a_m \in \mathcal{A}^1} a_{m,d}\right)^2. \quad (3)$$

On the receiver side of the scheme presented in Fig. 6, the symbols are de–interleaved and decoded using an (iterative) MSD process. The levels are decoded sequentially using the results of the previous levels as a–priori information L_a. First, the channel symbols y are demapped, resulting in APPs of the codeword symbols c_i for each level i, based on the received signal y. The APPs are represented as conditional LLRs, expressed as

$$L\left(c_i|y\right) = \ln\frac{P\left(c_i = 1|y\right)}{P\left(c_i = 0|y\right)}. \quad (4)$$

The FEC block outputs the estimated data sequence $\hat{D}_i$ based on the LLRs. Subsequently, $\hat{D}_i$ is re–encoded and re–interleaved and used as a–priori information for the demapping of the next levels. It is shown in [4] that error propagation between the levels can be avoided if the operation points of the levels are well separated and the following rule holds

$$\mathrm{SNR}_1 < \mathrm{SNR}_2 < \cdots < \mathrm{SNR}_n, \quad (5)$$

where SNR_i represents the minimum signal–to–noise ratio (SNR) for the error–free decoding of level i.

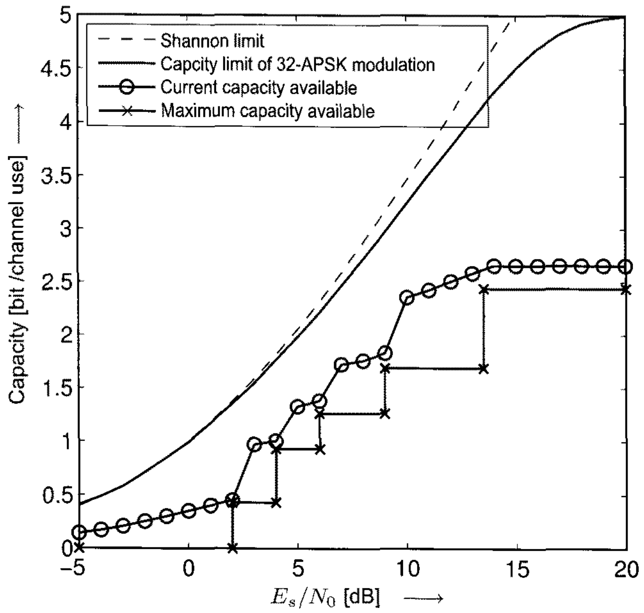

Fig. 7. MSD chart for 32–APSK with BL partitioning.

The decoder complexity is reduced by using the same FEC block for all levels, which is possible because only one decoding process is performed in each stage. The decoding algorithm is as follows:

Loop from $i = 1$ to n
- Demap level i using a–priori information from levels $1, \cdots, i{-}1$
- Channel de–interleaver
- Channel decoder (check if the decoded level i is error–free)
- If error–free $\rightarrow$ re–encode and re–interleave level i
- If not error–free $\rightarrow$ stop decoding

end

The process continues until all levels have been decoded or until the channel decoder detects that one level was not successfully decoded. In this case, the decoding process stops, saving the effort of decoding the higher layers. This leads to a significant reduction in the decoding time in bad channel situations, in which only a few layers are decodable.

A specific code rate can be selected for each protection level, which implies that each level carries a different amount of information. In order to design an efficient MLC scheme, the code rates must be chosen properly according to the desirable QoS of each level. We propose a convenient and practical way to design the levels. It is based on the calculation of the channel capacity using the mutual information (MI) I [8] between the output LLRs l of the demapper and the original input data d to the mapper of a level. It can be expressed as

$$I = \sum_{d\in\{0,1\}} \int_{-\infty}^{+\infty} p(l,d) \log\left(\frac{p(l,d)}{p_1(l)p_2(d)}\right) \mathrm{d}l \qquad (6)$$

where $p(l,d)$ denotes the joint probability density function (pdf) of l and d, and $p_1(l)$ and $p_2(d)$ are the corresponding marginal pdfs. The Monte Carlo method is used to calculate the MI efficiently. Besides, the channel decoder is also characterized by calculating the MI at its input and output. Using these calculations, an MLC scheme can be designed in two ways:

a) A certain transmission rate is assured for each level i by setting the code rates R_i, for example, a rate of 0.8 bit/channel use is required for a level that transmits music. In this case, a set of SNR operation points (in dB) $\mathcal{S}$ are calculated by using the MI of the demapper and the channel decoder [4]. This provides the minimum required SNR for each level for error–free decoding.
b) A certain level i must be error–free for SNRs higher than SNR_i. In this case, the designer wants to assure error–free transmission for a certain level for SNRs higher than a given threshold. The set of required code rates $\mathcal{R}$ is obtained using the MI calculations.

We present a graphical method involving the used of MSD charts [4] for designing the levels; the current capacity available (*cca*) and the maximum decoding capacity (*mca*) functions are defined. The *cca* function corresponds to a stair function; it is created using $\mathcal{S}$ and $\mathcal{R}$ and is defined as

$$cca(x) = \begin{cases} 0, & x < \mathrm{SNR}_1 \\ R_1, & \mathrm{SNR}_1 \le x < \mathrm{SNR}_2 \\ \cdots & \\ \sum_{j=1}^{i} R_j, & \mathrm{SNR}_i \le x < \mathrm{SNR}_{i+1} \\ \cdots & \\ \sum_{j=1}^{n} R_j, & \mathrm{SNR}_n \le x < \infty. \end{cases}$$

The first representation of this function was introduced in [22]. The *mca* function is defined as

$$mca(x) = \begin{cases} I_1(x), & x \le \mathrm{SNR}_1 \\ R_1 + I_2(x), & \mathrm{SNR}_1 < x \le \mathrm{SNR}_2 \\ \cdots & \\ \sum_{j=1}^{i} R_j + I_{i+1}(x), & \mathrm{SNR}_i < x \le \mathrm{SNR}_{i+1} \\ \cdots & \\ \sum_{j=1}^{n-1} R_j + I_n(x), & \mathrm{SNR}_{n-1} < x < \infty. \end{cases}$$

These functions also allow the designer to graphically evaluate the loss due to the FEC block. The loss is represented as the distance between the *mca* and *cca* functions at the operation point SNR_i of each level. Fig. 7 gives examples of plots of these functions vs. E_s/N_0, obtained by simulating a 32–APSK transmission with BL partitioning of the constellation and a 3GPP2 turbo code as the FEC code. Additionally, we show the Shannon limit and the modulation–constrained capacity limit of the 32–APSK constellation. Levels are defined according to $\mathcal{S} = \{2, 4, 6, 9, 13.5\}$, resulting in $\mathcal{R} = \{2/5, 1/2, 1/3, 3/8, 3/4\}$. It is to be noted that the distance between *mca* and *cca* is high when high code rates are used, implying that the 3GPP2 turbo code results in high loss at high code rates. This is in agreement with theory because this turbo code is based on a mother code rate of 1/5, and higher code rates are achieved by puncturing, which is not optimal.

MSD charts are a powerful method to design MLC schemes for different requirements. Besides, they can be used to compare constellations with different mapping strategies. However, an AWGN channel is assumed, which is not valid for satellite broadcast systems. Only if the channel interleaver is extremely long can the fluctuations of the satellite channel be reduced to

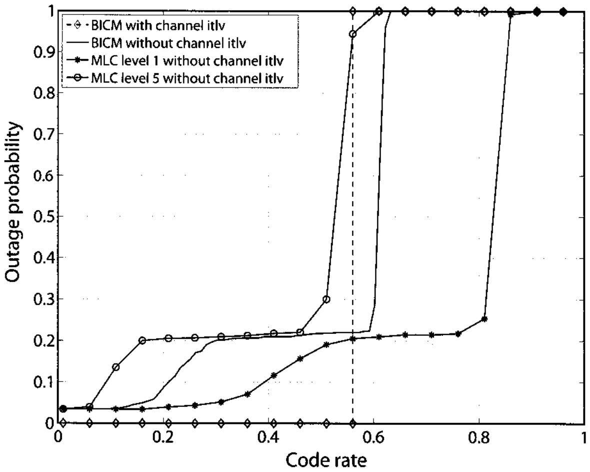

Fig. 8. Outage probability of BICM and MLC for the suburban scenario.

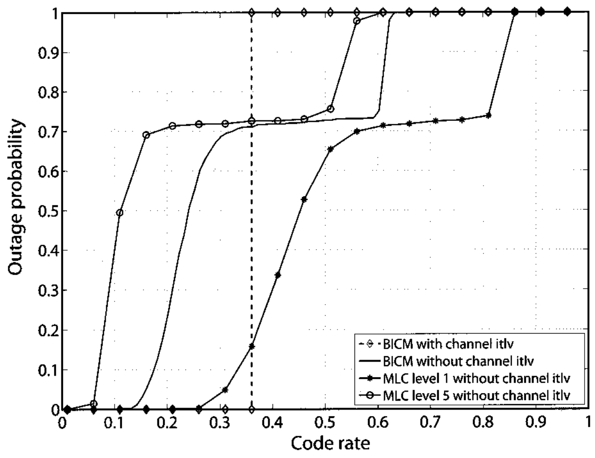

Fig. 9. Outage probability of BICM and MLC for the forest scenario.

almost zero, resulting in a constant capacity. In order to analyze the effect of satellite channels in the proposed MLC scheme, we again use the LMS channel model and remove the channel interleaver from the system. Simulations are performed for the BICM scheme (Fig. 1) and the proposed MLC scheme. Besides, a suburban scenario where almost no shadowing occurs and a forest scenario with almost no blocking are analyzed.

For very low code rates, the outage probability tends to zero because almost all errors can be corrected. On the other hand, high code rates imply bad decoding capabilities and therefore the loss of many codewords. Fig. 8 shows the outage probability for the same scenario as above, i.e., a 32–APSK constellation with BL partitioning and a 3GPP2 turbo code as the FEC code; the solid line without any mark represents the outage probability for BICM. For code rates below 0.6, the LOS region is decodable. For successgul decoding under shadowing conditions, the code rate must be below 0.2. Error–free decoding of the first level of MLC is possible in the regions of LOS and shadowing by using code rates below 0.4, where almost no outage occurs. The worst protected level requires extremely low code rates (< 0.1) to overcome shadowing.

The same analysis is performed for the forest scenario, whose outage probability is depicted in Fig. 9. Assuming a code rate of 0.2 for the BICM, we find that an outage of about 20% occurs, while the same code rate for the first level of MLC results in error–free decoding all the time. On the other hand, the fifth level of MLC is very weak. This level could be used for non–critical information, where a high outage probability is acceptable.

In urban scenarios with a high probability of blockage, the use of a very low code rate for the first levels could not be sufficient to overcome the dropouts, even if scattered and reflected components are received. In this case, these levels should be interleaved or terrestrial repeaters should be used, as proposed in [6] and [7].

Table 1 shows a comparison between MLC and BICM. For the suburban scenario and BICM, the code rate was set to 0.3, resulting in a transmission rate of 1.5 bit/channel use. The code rates for the different levels of MLC are selected in such a way

		Suburban		Forest	
		Tx. rate	Outage	Tx. rate	Outage
Level	1	0.20	4%	0.16	0%
	2	0.20	4%	0.29	5%
	3	0.43	24%	0.10	33%
	4	0.35	24%	0.12	74%
	5	0.36	29%	0.12	75%
MLC		1.54	–	0.79	–
BICM		1.5	21%	0.80	12%

Table 1. Comparison between BICM and MLC for the suburban and forest scenarios for a transmission rate of 1.5 bit/channel use.

that their sum approximately equals the transmission rate of BICM in order to guarantee a fair comparison. BICM has an outage probability of 21% while MLC presents a range from 4% to 29% for the first and fifth level, respectively. Again, it is shown that MLC provides more flexibility when setting the QoS for an LMS channel. The same calculations were made for the forest scenario, where a low code rate is used for the first level in order to avoid outages. Note that the MLC scheme allows the choice of different code rates and interleaver profiles for each level and that the hierarchical modulation provides a wide range of SNRs when choosing the level of protection. However, the number of different levels of protection is limited in principle by the modulation order. For satellite broadcasting systems, modulation orders above 32 (e.g., corresponding to 32–APSK) are normally not employed due to the limited link budget of these systems, thus restricting the number of protection levels to five or less.

Multidimensional MLC

We introduce the use of multidimensional (MD) constellations $\mathcal{A} \in \mathbb{C}^{D/2}$ with $D \geq 2$ in the MLC scheme proposed in Fig. 6 in order to increase the number of protection levels. Obviously, more dimensions means more signal points M, and therefore more levels. However, these new levels could overlap with the existing levels and new protection levels do not arise. An MD constellation $\mathcal{A}$ can be defined as the Cartesian product of some constituent constellations $\mathcal{B}_k$. For illustration, we assume a unidimensional binary phase shift keying (BPSK) modulation of capacity C^{BPSK} as the constituent constellation. The Cartesian product of two BPSK constellations produces a QPSK constellation ($D=2$) with two levels of capacity, C_0^{QPSK} and C_1^{QPSK}. The capacity of a BPSK constellation should be equal to two times the capacity of the two dimensional BPSK constellation giving

$$\sum_{i=0}^{1} C_i^{\mathrm{QPSK}} = 2C^{\mathrm{BPSK}}. \tag{7}$$

Besides, a decrease in the protection of the levels is desirable in order to assure (5), as BL does, which can be expressed as

$$C_0 > \cdots > C_i > \cdots > C_n. \tag{8}$$

If Gray mapping is used as partitioning strategy for the QPSK constellation, both levels are equally protected because the distances between the subsets are the same. Therefore, $C_0^{\mathrm{QPSK}} = C_1^{\mathrm{QPSK}}$, which is in conflict with (8). On the other hand, if an anti–Gray mapping is selected, the signal points of each of the subsets $\mathcal{A}^0$ and $\mathcal{A}^1$ of the low protection level are located at the corners of the square formed by the QPSK constellation. Therefore, the distance between these points is $\sqrt{2}$ times higher than the distance between the signal points of a BPSK constellation, leading to $C_1^{\mathrm{QPSK}} > C^{\mathrm{BPSK}}$; applying (7), we conclude that $C_0^{\mathrm{QPSK}} < C_1^{\mathrm{QPSK}}$, which is also in conflict with (8). Both mappings of the QPSK constellation do not produce the desirable behavior for our MLC scheme. The same calculations can be done for 3xBPSK. The lowest protected level always has a higher distance than that between the two points of a BPSK constellation for an anti–Gray mapping, leading to a conflict with (8). If Gray mapping is employed, all levels have the same protection. Note that this result can be extended to MD constellations with any number of dimensions.

Thus, we have shown that MD constellations comprising identical constituent constellations are not suitable for our MLC scheme. Instead of using exactly the same constellation for $\mathcal{B}_k$, constellation points of different constituent constellations might be slightly shifted towards each other. In the following, we restrict ourselves to the use of two real–valued or complex–valued constituent constellations $\mathcal{B}_0$ and $\mathcal{B}_1$, creating a D–dimensional ($D=2$ or $D=4$) constellation, $\mathcal{A}=\mathcal{B}_0\times\mathcal{B}_1$.

An increase in the number of protection levels can be also achieved by a straightforward scheme where $\mathcal{B}_0$ and $\mathcal{B}_1$ are considered independently. The codewords are mapped parallel to $\mathcal{B}_0$ and $\mathcal{B}_1$, respectively, and the results are output in a time–multiplexing manner. This scheme and the MD approach are investigated below. As an example, we use the 16–QAM constellations of [7], where three different versions of the constellation are defined depending on a factor α, which indicates the spread of the constellation. We use the versions with $\alpha=1$ and $\alpha=4$, representing a uniform 16–QAM constellation $\mathcal{B}_0$ and a non–uniform 16–QAM constellation $\mathcal{B}_1$, respectively. Fig. 10 shows the MI of each level for $\mathcal{B}_0$ and $\mathcal{B}_1$, which corresponds to the straightforward approach. Note that $\mathcal{B}_1$ results in a loss in the overall channel capacity due to the extreme non–uniformity of the signal constellation. Besides, $\mathcal{B}_1$ provides the best and

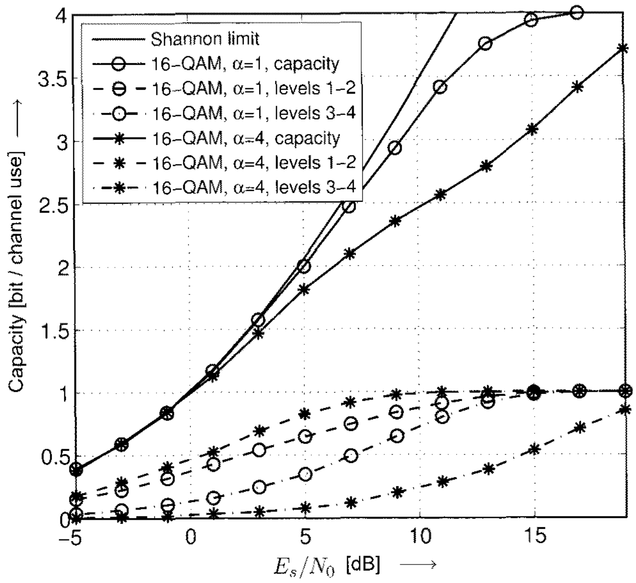

Fig. 10. Mutual information of the levels of 16–QAM constellation with $\alpha=1$ and $\alpha=4$.

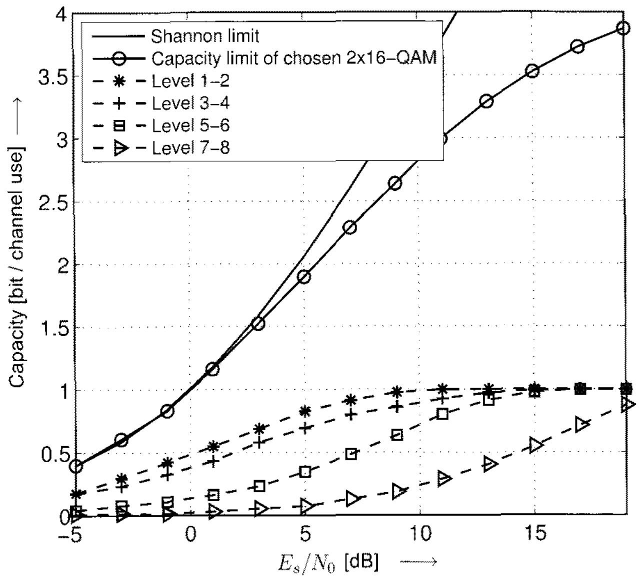

Fig. 11. Mutual information of the levels of a 2x16–QAM MD constellation with BL.

the worst protection levels, while the levels of $\mathcal{B}_0$ are in between, which means that the levels must be ordered properly in the encoding stage to match the different QoS requirements. An MD constellation with random partitionings does not provide the desirable hierarchical distribution of the levels for MLC using MSD, cf. [4], where it is required that the protection level decrease with an increase in the level index. Thus, an MD constellation with BL partitioning is considered now. To this end, we use a 2x16–QAM constellation $\mathcal{A}_r$ defined as the Cartesian product of the uniform and non–uniform 16–QAM, leading to a four–dimensional constellation expressed as $\mathcal{A}_r=\mathcal{B}_0\times\mathcal{B}_1$.

In this case, a theoretical analysis is too complex due to the high number of combinations of the different partitioning strategies of the constellation. Therefore, we propose the use of the

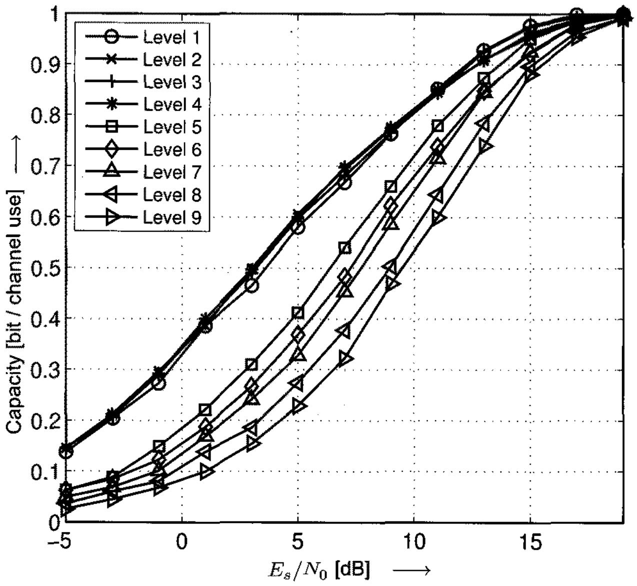

Fig. 12. Mutual information of levels of a 16–APSKx32–APSK MD constellation with BL.

BSA presented in this section by applying the cost functions J_a and J_b. The BSA is applied to A_r with an initial random partitioning in order to find a proper BL partitioning. Fig. 11 shows the MI of the levels of A_r with BL partitioning. Now the protection levels of the MD constellation present the desirable hierarchical behavior, ranging from the best protection for the first level to the worst protection for the last level. Both cost functions J_a and J_b provide the same result, showing that they are in principle equivalent. The straightforward approach and the MD scheme raise the number of different protection levels when non–uniform constituent constellations are used, while the MD scheme provides an additional inherent ordering of the levels according to a decrease in protection.

A more general approach uses completely different constellations for $\mathcal{B}_0$ and $\mathcal{B}_1$. For example, one might combine a 16–APSK and a 32–APSK, both extracted from [23], to create a four–dimensional constellation. In this case, a brute–force search for the optimum BL partitioning is definitely infeasible. Applying the BSA again leads to the MI results of Fig. 12. We obtain a maximum range of 8 dB with a fine granularity of the protection of levels arranged in a hierarchical manner. Thus, BL partitioning is also possible for multidimensional constellations comprising different constituent constellations.

V. CONCLUSION

In this paper, we have analyzed different methods for constructing a satellite broadcasting system with different levels of QoS. The possibility of incorporating different levels of QoS in the ESDR and the DVB–SH standards has been analyzed, showing that the DVB–SH standard provides almost no flexibility when a transmission of data with different requirements of protection is required. A solution to this problem has been proposed for the case where a second data stream is required for low latency communications. The RT data stream is molded to the standard DVB–SH signal format by using a very short special interleaver profile derived from the standard DVB–SH stream parameters. It has been proved that this solution is backward compatible to legacy DVB–SH receivers, and the performance has been measured using an appropriate laboratory setup. On the basis of multilevel codes with multistage decoding, we also propose a more general solution for including different levels of QoS in a satellite broadcasting system. A convenient and practical way to design the levels is presented, where either the code rates or the SNR operation points are given as input parameters. Multistage decoder charts have been introduced and serve as a useful tool to evaluate MLC systems with practical codes and decoding blocks. The viability of this system concept under a typical satellite channel has been discussed. Finally, the proposed MLC solution has been extended to the case of multidimensional constellations, which increases the number of different protection levels.

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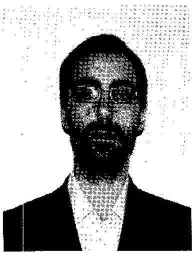

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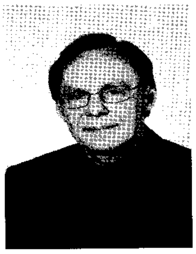

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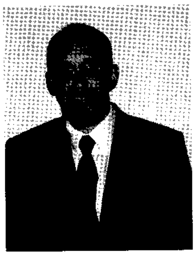

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