Error Control Schemes in Wireless Broadband Networks

Konstantinos Makris, Dimitrios D. Vergados, *Member, IEEE*, Micahel E. Theologou, *Member, IEEE*, Christos Anagnostopoulos *Student Member, IEEE,* and John Anagnostopoulos, *Student Member, IEEE*

Abstract-- **With rapid advances in wireless packet networking technology, it is becoming possible to provide mobile users not only with data, voice, but also with various communication services. In order to provide these services to mobile users /terminals, it is necessary to have not only sufficient bandwidth for support of high speed transmission, but also effective control over wireless channel errors. Forward error correction coding (FEC) is a widely used method to improve transmission quality in wireless broadband communication systems. In this paper, we present and evaluate three Error Control algorithms in broadband wireless networks and we discuss methods for comparing the performance of codes of various convolutional codes and Reed-Solomon codes.**

Index terms-- FEC, Wireless Broadband Networks, Error Control, QoS

I. Introduction

The first problem is that the wireless channel error pattern is different from the one of the wired networks. The wireless channel error is location dependent, and is also characterized as bursty. Bursty channel errors will drive to suffer from back-to-back packet losses, unlike the more random packet losses on a wired link. For example, for VBR traffic [1] i.e. wireless video transmission, not all lost packets should be recovered, as compressed video data are usually encoded as frames of different importance. In that case, error control should be performed in a differentiated manner. The second problem is that we need to achieve both good scalability and transmission efficiency. Automatic Repeat reQuest (ARQ) has been proposed [7] as effective technique for reliable multicast in wired networks, as it is able to reduce the error ratio, increasing at the same time the cell delay by retransmissions. However, FEC can reduce this number of retransmissions [6]. Static FEC algorithms, can degrade the performance, as they are not changeable [4]. Thus, they must be implemented to guarantee a certain Quality of Services (QoS) requirement for the worst case channel conditions. Our scope is to provide an error free channel, increasing the throughput due to missing unnecessary overheads, especially when the channel path error rate fluctuates widely. In our days, the majority of the networking applications require this error free transmission [15]. In wireless broadband networks, i.e. WATM this goal can only be reached by a combination of FEC and Automatic Repeat Request ARQ schemes, applied to the whole broadband network architecture [9], [13].

In this paper, however, we present some schemes on the use of Forward Error Correction (FEC) in wireless broadband networks. The use of advanced coding technology became possible through the ongoing research developments in the wired and wireless transmission environments. There are two main groups of applications for FEC coding: broadband communication systems and storage systems [8]. The main objective in both cases is a capacity improvement. Wireless broadband systems can benefit from the experiences of these implementations by optimizing the error correction strategy with respect to the specific radio channel and service requirements.

The main concept of FEC is to add a certain amount of redundancy to the information, which can be exploited by the receiver to correct transmission errors due to channel noise. In the literature, FEC coding is therefore often described as channel coding [8]. C. Shannon presented in his mathematical theory of communication, that every transmission channel has a theoretical maximum capacity, which depends on the bandwidth and the signal-to-noise ratio. Also, the use of suitable codes will allow further improvement of bandwidth efficiency [4]. On the other hand, FEC codes alone cannot guarantee an error free transmission, a critical requirement in our days. The mapping of the broadband quality of service parameters to the radio link quality will have a strong impact on the FEC/ARQ schemes [5]. For example, the ATM standards define the main QoS parameters Cell Loss Rate (CLR), Cell Delay Variation (CDV) and cell transfer Delay (CTD). There will be certain absolute limitations on the wireless connections, which are caused by the physical transmission environment.

Therefore, our paper is organized as follows: Section II discusses The Reed-Solomon related work. Section III describes the FEC schemes applied in wireless broadband

National Technical University of Athens, Department of Electrical and Computer Engineering, Heroon Polytechneiou 9 Str., GR-157 73, Athens, Greece, e-mail: {vergados; kmakris@telecom.ntua.gr}

networks. Finally, section IV evaluates the presented schemes and suggesting a new adaptive algorithm.

II. Error Control Schemes

A. Reed-Solomon codes

In the following chapter, we will compare the use of block codes and convolutional codes with respect to wireless broadband networks i.e. WATM. For the block codes we selected the Reed-Solomon (RS) codes, because they are superior to binary block codes in most applications [8]. The RS codes are a subclass of the nonbinary BCH cyclic codes. The generator polynomial of a RS code is specified in terms of its roots from the Galois field GF(q), where *q* is any power of a prime number *p*. A t-error correcting Reed – Solomon code with symbols from GF(q) has the following parameters:

- Block length : n = q - 1
- Number of parity-check bits : n – k = 2t
- Minimum distance : dmin = 2t + 1

In order to describe how those codes function, we consider Reed – Solomon codes with code symbols from the Galois field $GF(2^m)\,(i.e., q = 2^m)$. The generator polynomial of a primitive t-error correcting Reed – Solomon code of length $2^m - 1$ is:

$$\begin{aligned} g(x) &= (X+a)(X+a^2)...(X+a^{2t}) \\ &= g_0 + g_1 X + g_2 X^2 + ... + g_{2t-1} X^{2t-1} + X^{2t} \end{aligned} \quad (1)$$

where α is a primitive element in $GF(2^m)$. Obviously $a, a^2, ..., a^{2t}$ are the roots of g(x), which has coefficients from $GF(2^m)$. This polynomial generates a (n, n–2t) cyclic code, which consists of those polynomials of degree n–1 or less with coefficients from $GF(2^m)$ that are multiples of g(x). For the encoding of this code let

$$a(X) = a_0 + a_1 X + a_2 X^2 + ... + a_{k-1} X^{k-1} \quad (2)$$

be the message to be encoded where k = n - 2t. Dividing the message polynomial $X^{2t} a(X)$ by the generator polynomial g(X) we obtain the remainder

$$b(X) = b_0 + b_1 X + ... + b_{2t-1} X^{2t-1} \quad (3)$$

(with degree less than that of g(X)) whose coefficients are the 2t parity – check digits .

Let

$$\begin{aligned} v(X) &= \upsilon_0 + \upsilon_1 X + ... + \upsilon_{n-1} X^{n-1} \\ r(X) &= r_0 + r_1 X + ... + r_{n-1} X^{n-1} \end{aligned} \quad (4)$$

be the transmitted code vector and the corresponding receiver vector respectively. The error pattern added by the channel is:

$$e(X) = r(X) - v(X) = e_0 + e_1 X + ... + e_{n-1} X \quad (5)$$

where $e_i = r_i - \upsilon_i$ is a symbol from $GF(2^m)$. With the error pattern containing v errors (nonzero components) at location $X^{j_1}, X^{j_2}, ..., X^{j_v}$ 0

where $0 \le j_1 \le j_2 \le ... \le j_v \le n-1$, we obtain:

$$e(X) = e_{j_1} X^{j_1} + e_{j_2} X^{j_2} + ... + e_{j_v} X^{j_v} \quad (6)$$

Therefore, in order to determine the error e(X) we need to know the v pairs $(X^{j_i}, e_{j_1})'s$.

For decoding of a t – error correcting Reed-Solomon code the first step is to compute the syndrome from r(X): $S = (S_1, S_2, ..., S_{2t})$. The *i*th component of syndrome is given by:

$$S_i = r(a^i) = b_i \quad (7)$$

where b_i is the remainder when dividing r(X) by $X + a^i$. But since $a, a^2, ..., a^{2t}$ are the roots of each code polynomial, $v(a^i) = 0$ for $1 \le i \le 2t$, then the *i*th component of syndrome is given by:

$$S_i = e(a^1) \quad (8)$$

We define $\beta_l = a^{j_l}$ for l=1,2,…v and thus we have:

$$\begin{aligned} S_1 &= r(a) = e_{j_1}\beta_1 + e_{j_2}\beta_2 + ... + e_{j_v}\beta_v \\ S_2 &= r(a^2) = e_{j_1}\beta_1^2 + e_{j_2}\beta_2^2 + ... + e_{j_v}\beta_v^2 \\ &\vdots \\ S_{2t} &= r(a^{2t}) = e_{j_1}\beta_1^{2t} + e_{j_2}\beta_2^{2t} + ... + e_{j_v}\beta_v^{2t} \end{aligned} \quad (9)$$

The second and most difficult step of the decoding is to find the error-location polynomial:

$$\begin{aligned} \sigma(X) &= (1+\beta_1 X)(1+\beta_2 X)...(1+\beta_v X) \\ &= 1 + \sigma_1 X + ... + \sigma_v X^v \end{aligned} \quad (10)$$

so, that we can determine the error values. The $\sigma_i's$ are related to the syndrome components $S_1's$ by the following Newton's identities:

$$\begin{aligned} &S_1 + \sigma_1 = 0 \\ &S_2 + \sigma_1 S_1 + 2\sigma_2 = 0 \\ &\vdots \\ &S_v + \sigma_1 S_{v-1} + \ldots + \sigma_{v-1} S_1 + v\sigma_v = 0 \\ &S_{v+1} + \sigma_1 S_v + \ldots + \sigma_{v-1} S_2 + \sigma_v S_1 = 0 \\ &\vdots \end{aligned} \quad (11)$$

In order to find the error-location polynomial, we can use the Berlekamp's iterative algorithm. Firstly, we find a minimum degree polynomial $\sigma^{(1)}(X)$ whose coefficients satisfy the first Newton's identity. Then it should be tested, whether those coefficients also satisfy the second Newton's identity. If they do we set:

$$\sigma^{(2)}(X) = \sigma^{(1)}(X) \quad (12)$$

If they do not satisfy this identity, a correction term to $\sigma^{(1)}(X)$ to form $\sigma^{(2)}(X)$ should be added, such that, $\sigma^{(2)}(X)$ has minimum degree and its coefficients satisfy the first two Newton's identities. Let the minimum-degree polynomial, whose coefficients satisfy the first μ Newton's identities, (determined by the μth iteration step) be:

$$\sigma^{(\mu)}(X) = 1 + \sigma_1^{(\mu)} X + \sigma_2^{(\mu)} X^2 + \ldots + \sigma_{l_\mu}^{(\mu)} X^{l_\mu} \qquad (13)$$

To determine $\sigma^{(\mu+1)}(X)$, the following quantity should be computed:

$$d_\mu = S_{\mu+1} + \sigma_1^{(\mu)} S_\mu + \ldots + \sigma_{l_\mu}^{(\mu)} S_{\mu+1-l_m} \qquad (14)$$

which is called the μth discrepancy. If $d_\mu = 0$, then $\sigma^{(\mu+1)}(X) = \sigma^{(\mu)}(X)$, but if $d_\mu \neq 0$, go back to the steps prior to the μth step and a polynomial $\sigma^{(\rho)}(X)$ is determined, such that $d_\rho \neq 0$ and the number $\rho - l_\rho$ (this polynomial's degree) has the largest value. Then the minimum-degree polynomial whose coefficients satisfy the first μ +1 Newton's identities is:

$$\sigma^{(\mu+1)}(X) = \sigma^{(\mu)}(X) + d_\mu d_\rho^{-1} X^{(\mu-\rho)} \sigma^{(\rho)}(X) \qquad (15)$$

Iteration continues until $\sigma^{(2t)}(X)$ is obtained, which is taken to be the error-location polynomial σ(X).

The third step of the decoding is to find the roots of σ(X), which are the inverses of the error location numbers. For the forth and last step of the decoding procedure let:

$$Z(X) = 1 + (S_1 + \sigma_1)X + (S_2 + S_1\sigma_1 + \sigma_2)X^2 \ldots \\ (S_v + \sigma_1 S_{v-1} + \ldots + \sigma_v)X^v \qquad (16)$$

and then the error value at location $\beta_l = a^{j_l}$ is given by :

$$e_{j_l} = \frac{Z(\beta_l^{-1})}{\prod_{\substack{i=1 \\ i\neq l}}^{v} (1 + \beta_i \beta_l^{-1})} \qquad (17)$$

If β is not a primitive element of $GF(2^m)$ then the 2^m - ary code generated by $g(x) = (X+\beta)(X+\beta^2)\ldots(X+\beta^{2t})$ is a nonprimitive t-error correcting Reed - Solomon code, whose decoding is identical to the decoding of a primitive Reed-Solomon code.

III. FEC Schemes in Wireless Broadband Networks

In the Adaptive FEC (AFEC) Scheme for Data Traffic in wireless broadband networks i.e. WATM, a dynamic error control mechanism is provided, applicable for only data traffic, by combining a convolutional inner code (which itself is not sufficient for wireless ATM networks) and a Reed – Solomon outer code with interleaving in order to protect the ATM cell payload, as well as, the payload type indicator/cell loss priority fields in the ATM cell header [6]. The AFEC scheme functions proposed by [11] with the LANET framing [12] and addresses protocol so that the error tolerance is enhanced in terms of framing and correct delivery.

At the cell encoding procedure a group is formed by accumulating ATM cells while payload is extracted and the group is segmented into 6 blocks. For each block space for one piggyback byte as each block's header is allocated. Each block size increases according to the encoding rates and then we use RS (*n*=255,*k*=249) code with 3-B correcting capability. The output of the encoder is in the reverse order (first the parity bytes, last the piggyback bytes). The RS-encoded cell group is prepared for the output, the delineation byte is loaded in the first byte location of each output cell within a group and then we load payload [11]. At the end of the encoding the least-significant nibble between bytes 1 (delineation byte) and 25 is swapped and the cell group is output with cell 1 (delineation byte 0x5A) at the beginning. At the decoding procedure, firstly there is the cell group delineation with a method of interleaving that doubles the tolerance of the delineation bytes. Then the payload is extracted by de-interleaving and decoded by using the RS scheme. The next step is to rate adjustment in dynamic wireless links with the so-called Bucket concept. At the original cells are regenerated. Address FEC is an implementation for header address protection where 1-nibble correcting RS and 2-nibble correcting RS codes are used [9]. The proposed frame size of LANET [12] is 2400bytes, which are subdivided into 45 ATM cells (53 bytes each cell) and 15 bytes overhead (only 0.63%of the bandwidth). It is also subdivided into nine subframes. In LANET, we meet three basic units, the cell extractor, the framer and the history buffer.

The evaluation of the AFEC's performance for data traffic in broadband networks is made by using two simulation models (with and without LANET framing) [11]. From simulation the Cell Sync Error Rate (%) and the BER Performance after FEC Correction are obtained and used as performance measures. The simulation results show that LANET increases the Sinc performance significantly. This performance is higher for big channel BER and is strongly improved for mobile moving. For AFEC with LANET, at the cases of stationary and wireless terminals, the final BER (after FEC correction) is about 10^{-6} even in the case of high channel BER 10^{-2}. This performance is acceptable for TCI/IP traffic. So, even for mobile wireless terminal cases (where the error rates become higher as the mobile speed increases) the performance of AFEC remain at high levels because the interleaving scheme in payload FEC works effectively to take care of the burst error due to the Doppler shift. The conclusion is that in both studied cases of stationery and mobile terminals, the AFEC scheme provides a slightly better performance over the existing concatenated FEC scheme.

The AFEC scheme has RS code rates (1/2, 3/4, 7/8) and the best performance is achieved by using the code rate of ½ The code rate threshold is determined from the maximum tolerable BER, which depends on network's applications. In the case TCP/IP traffic for channel BER the maximum tolerable rate is 10^{-6}, the code rate threshold is between 7/8 and ¾ for channel BER $2 \cdot 10^{-4}$ and between ¾ and ½ for channel BER $2 \cdot 10^{-3}$. For the channel BER range below the point of $2 \cdot 10^{-4}$ the code rate 7/8 can be used for TCP/IP. However, for channel BER beyond the point of $2 \cdot 10^{-2}$ the performance of AFEC with code rate ½ is degraded. So, this point of the channel BER is a limitation of the AFEC performance. From the simulation the code rate threshold is determined as ½ for 10^{-3} BER and as ¾ for 10^{-4} BER and as 7/8 for 10^{-5} BER.

In the Effective Bandwidth in Wireless ATM Networks, proposed by [13], a hybrid ARQ\FEC mechanism is used for error control. The analytical model for a single ATM connection the traffic source is by an on-off fluid model with peak rate τ and exponential distributed on and off periods with means 1/α and 1/β, respectively. The channel alternates between Good and Bad states with BER P_{eg} and P_{eb} (where P_{eb} is much bigger than P_{eg}) and durations exponentially distributed with means 1/δ and 1/γ respectively. The FEC/ARQ mechanism that is used has a cell format of four parts: ATM header, ATM payload, CRC and FEC (whose length varies).

Numerical results are obtained for two cases. In the simulation, the buffer size was varied and the BER during the Bad State and we fix the BER during the good state at 10^{-6}. At the first case [13] only a stop-and-wait ARQ is used for error control for a service rate c = 400cells/sec .The CLR increases as the buffer size decreases and the P_{eb} increases. The case of $P_{eb} = 10^{-2}$ shows a higher deviation than the other cases. As P_{eb} decreases the variance also decreases significantly and leads to a more deterministic cell departure process justifying the success of the deterministic approximation over the range of P_{eb} from 10^{-3} *to* 10^{-5}. This is also justified by the fact that c_T^2 (the squared coefficient of variation of actual service time T for Bad state) drops rapidly and approaches zero as P_{eb} goes to zero. The performance is studied for c=5000 and 550 cells/sec, where for $P_{eb} = 10^{-3}$ *and* 10^{-5} the analytical results are very close to the simulation, while for $P_{eb} = 10^{-2}$ the difference is smaller than the one for c=400 cells/sec. Also, the wireless effective bandwidth based on the dominant eigenvalue overestimates the actual bandwidth requirement.

The second case a hybrid ARQ/FEC is used for error control. For FEC Bose-Chaudhuri-Hocquenghem (BCH) is used. The total bits add up to 422 bits (ATM header & Payload). From the simulation, it is observed that more bandwidth is required as the target CLR increases and that the optimal code rate increases as P_{eb} increases. The last one indicates that there exists an optimal code rate for every BER and thus an efficient adaptive coding strategy is required. Finally, we understand that proper use of FEC is necessary for maximizing the use of the wireless bandwidth, which is much greater than the average source rate (close to peak rate) for cases of severe BER.

In the Adaptive FEC Algorithm (FECA) for Mobile Wireless Network, suggested by [10], there's the FEC-level Adaptation, which adjust the amount of the parity check bits (that FEC uses) on the channel status. FECA takes type-I hybrid approach, which is more effective for W-networks with heavy losses so that packets can't be recognized because of the corrupted preamble. FECA works without explicit feedback information (from the receivers) and hardware support. It also appropriately adapts the FEC strength to the constantly changing channel status. Five FEC levels were assumed, each of which corrects 5, 10, 15, 20 and 25 corrupted bits based on the observation that the maximum bytes corrupted in this experiment is up to 25 bytes. The packet size is fixed to 35-byte.

FECA rapidly joins at the next higher FEC level at each packet loss. To minimize the effect of false predictions, FECA needs to quickly drop from the newly joined level to the lower level when packet loss is no longer reported. FECA has a drop timer in each FEC level whose timeout value is adjusted by an exponential back-off algorithm. Every time FECA joins a new (higher) FEC level at a packet loss, it increases the drop timer if this FEC level up to *Tmax* by multiplying with the multiplication factor, α greater then 1.The result of this operation is that the more FECA visits a FEC level it larger its drop time, so that FECA stays at this FEC level for longer time. FECA also has a global polling timer in order to decay the drop timers of other FECA levels except the drop timer of the current FEC level. The global polling time expires every *Tp* and then FEC shrinks drop timeouts of all other levels up to *Tmin* by multiplying with a decay factor, β less than 1. The reason of this operation is that FECA needs to forget the old status by gradually resetting the timer's value.

The results from the theoretical analysis of FECA [10], [14] over the relation between the average signal power and the T-R (transmitter-receiver) distance based on the long-term shadowing model (large-scale fading model) [5] can be summarized in two observations: SNR attenuates slower in environments with LOS (line-of-sight) path than those with no LOS path and BER increases faster in environments with no LOS path than those with LOS path.

The analysis over the BER variation by the small-scale fading indicates that the average duration varies from a few milliseconds to a few seconds. As the T-R distance increases the duration of heavy BER lasts longer while the average fading duration is in proportion to the mobility speed. Those results show that the shortest fading duration that FECA would adapt would be a few tens of milliseconds.

It is shown [10] that when the average duration of BER is in the middle between a few hundred microseconds and a few milliseconds, FECA cannot monitor the channel status fast enough to follow this degree of fluctuation and suffers the overhead of adaptation.

FECA's performance was evaluated through simulation [3], [10], where FECA to ARQ and three FEC schemes (1/4 BCH, 1/3 BCH and ½ BCH) were compared, while varying the off-state BER. Firstly their performance over a Wireless Channel was evaluated. For ARQ a retransmission mechanism was used, characterized by a stop-and–wait and an exponential back-off algorithm. FECA selects one of the above three FEC levels with its five tunable parameters α, β, Tmax, Tmin, Tp assigned to 2, 0.9, 1sec, 6ms and 6ms respectively. Each packet takes around 2ms for its transmission. The ARQ's performance quickly degrades as the light-error state BER increases. At BER of 0.5% all packets are corrupted. The three FEC schemes provide poor performance when the light-error state's BER is less than 0.1% due to overhead check bits but their throughputs remain constant before their cutoff points for ¼ and 1/3 BCH (1,25% and 2% BER) while ½ BCH provides poor but constant performance for the whole range due to its strong correction capability. FECA succeeds in adapting its strength to the underlying long-term channel state and achieves nearly the best performance among the four other state algorithms over the whole range of error rates even though there is negligible performance discrepancy due to overhead of FECA's dynamic adjustment.

Following those five error algorithms were checked, as they perform over a Mobile Wireless Channel (when a node moves back and forth between two places). This wireless channel was abstracted with two two-state Markov chains each of which is determined as (5% BER, t_v), (0.3% BER, P_{ed}), (5% BER, 5msec) and (3% BER, t_v) respectively, where t_v is the variable time period varying from 10ms to 1 sec. We set Tmax to 100ms.

The four static algorithms provide the constant performance regardless of the fluctuation duration. FECA begins to perform better than the other algorithms, by up to 12% at 1 second comparing to ½ BCH. As the BER oscillates more rapidly like when the variable duration t_v is smaller than 100ms, the adjustment of FECA achieves less improvement than ½ BCH (even though it is better than the other three algorithms). In conclusion, FECA can outperform static FEC and ARQ algorithms over a range of error conditions, provided that error rates do not oscillate too rapidly.

IV. Conclusions

This paper presents a review of FEC schemes applied in the wireless broadband networks, that dynamically trades-off between reliability and efficiency for better performance. Wireless broadband networks suffer from the poor performance comparing to wired ones due to the heavy propagation error. To improve the reliability over the noisy wireless channels, wireless networks can employ error correction techniques. In [13], it is shown that an excessive amount of bandwidth is required when only ARQ is used due to the wireless effective bandwidth, which in this case overestimates the actual bandwidth requirement. So the proper use of FEC can provide a better utilization of the wireless bandwidth. The weakness of the typical static FEC schemes to match their overhead to the channel condition can be solve by faced by a FEC mechanism, which adapt to the channel bit error variations.

Experiments taken from short-range radios to characterize the range of bit-level errors, and then through simulation show that FECA can provided that error rates do not oscillate too rapidly. We are currently implementing simulation experiments to outperform and compare an adaptive algorithm to several alternatives over a range of error conditions. From the evaluation of [10], it is obvious that a main problem of the current adaptive FEC schemes is their sensitivity on the rapid oscillations of the error rates. This is theoretically proved as those schemes improve the performance only in wireless networks whose average short term fluctuation's width is at least larger than a few tens of milliseconds. It is also indicated that they have a poorer performance in environments with no LOS path. In [11] for channel BER beyond the point of $2 \cdot 10^{-2}$ the performance of Adaptive FEC with code rate ½ is degraded and cannot satisfy the demand of a maximum tolerable BER of 10^{-6}, which is required for TCP/IP traffic. We will develop a new adaptive FEC scheme in order to confront these disadvantages of current adaptive FEC schemes and to provide free error channel with no reduce of the throughput even when BER fluctuates widely. To obtain the results we will assume a model that the cell departure process follows the Poison process. Our scheme will be developed to guarantee certain QoS parameters such as Cell Loss Rate (CLR), Cell Delay Variation (CDV) and cell transfer Delay (CTD).

Therefore, recently as wireless networks are integrated into the Internet [15], the impact of burst error in wireless channels has begun to be investigated. Packet-level network simulators now model both wired and wireless networks. Since one analytical model seldom represents all various wireless channels, simulators tend to include several models to select based on the simulation purpose.

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