Iterative Interference Cancellation for High Spectral Efficiency Satellite Communications

Bassel F. Beidas, Hesham El Gamal, and Stan Kay

***Abstract*—The problem of efficient utilization of the frequency spectrum for satellite systems is investigated; one which results as a consequence of highly crowding adjacent channels. An analytical characterization of the resulting interference channel is introduced and then exploited for interference cancellation. Two classes of cancelers are investigated. The first approach does not benefit from the forward error control (FEC) coding information which limits the performance gain. This motivates the second approach where a joint implementation of interference cancellation and decoding is developed using soft-input–soft-output (SISO) modules along with the iterative structure. It is shown that iterative interference cancellation techniques can achieve significant gains compared with the single-user matched filter receiver.**

***Index Terms*—Forward error control (FEC) decoding, minimum mean square error (mmse) with decision feedback, multiuser detection, satellite communications.**

I. Introduction

THE increasingly rapid growth in wireless communications has made the frequency spectrum an extremely precious resource. One method of achieving efficient utilization of the available bandwidth for satellite applications is by highly packing adjacent channels into a satellite transponder. This increases the system usage capacity but at the expense of generating strong interference. As a consequence, severe degradation in performance is experienced unless efficient interference cancellation schemes are implemented.

In this paper, we investigate interference cancellation techniques for high spectral efficiency satellite communication systems. We consider two design approaches. The first implements the interference cancellation algorithm prior to, and hence does not utilize, the channel decoding procedure. This class of algorithms enjoys relatively lower complexity and demands less synchronization requirements at the network level, as will be shown later. In the second approach, we consider the joint implementation of interference cancellation and forward error control (FEC) decoding and examine the segmented receiver where soft-input–soft-output (SISO) modules are used along with the iterative structure, i.e., Turbo Interference Cancellation. In general, it is shown here that satellite systems that implement combined iterative SISO interference cancellation and decoding can operate at very high spectral efficiency levels with a reasonable receiver complexity. Throughout this paper, we recognize and exploit the similarities that exist between the problem of highly bandwidth-efficient satellite communication and multiuser detection of spread-spectrum code division multiple access (CDMA) signals. In particular, we benefit from the iterative minimum mean square error (mmse) receiver proposed by the second author for CDMA systems in [1] (this algorithm was independently proposed by Wang and Poor in [2]). We also utilize the soft interference cancellation principle proposed for CDMA systems in [3], [4][1]. The iterative algorithms investigated in this paper avoid the exponential complexity exhibited by the maximum *a posteriori* (MAP) approach in [5], [6].

The remainder of the paper is organized as follows. Section II contains the mathematical description of the signal model that represents the situation of highly crowded adjacent channels. In Section III, we examine an iterative interference cancellation rule which does not utilize the decoding information. Section IV investigates two iterative SISO strategies for joint interference cancellation and decoding that trade complexity for performance. Extensive numerical studies illustrating the various design concepts and trade-offs are included in Section V. Finally, in Section VI we offer some concluding remarks.

II. Signal Model

Consider the received waveform $r(t)$ which consists of the signal in noise as

$$r(t) = s(t) + n(t). \tag{1}$$

The noise $n(t)$ is the standard additive white Gaussian noise (AWGN) with single-sided power spectral density (PSD) level of N_0 (W/Hz). The signal $s(t)$ models the situation of satellite adjacent channel interference (ACI) in which there are M binary data sources that are identical and independent. The FEC encoding is then applied, followed by independent interleaving and Gray mapping, onto quadrature phase shift-keying (QPSK) symbols at the rate of T_s^{-1}. The random interleaving is necessary to justify the independence assumption used to derive the Turbo interference cancellation receivers. This data is transformed into waveforms using an arbitrary unit-energy transmit filter $p(t)$, and frequency-translated to its respective slot or center frequency.

The signal can be described in complex form as

$$s(t) = \mathrm{Re}\left\{\tilde{s}(t)e^{j2\pi f_c t}\right\} \tag{2}$$

[1]This recent work by Boutros and Caire has shown that all Turbo multi-user algorithms can be derived in a unified framework using factor graph ideas.

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where f_c is the carrier frequency and $\tilde{s}(t)$ is the baseband complex envelope of the signal and is mathematically expressed as

$$\tilde{s}(t) = \sum_{m=1}^{M} \sum_{k=-\infty}^{\infty} \left[\sqrt{E_{b,m}} a_{m,k} p(t - kT_s - \epsilon_m T_s) e^{j(2\pi \Delta f_m t + \theta_m)} \right] \tag{3}$$

where the symbols $\{a_{m,k}; m = 1, 2, \ldots, M\}$ are taken from the QPSK constellation with values $\sqrt{2} e^{j\pi i/4}; i = 1, 3, 5, 7$. In (3), $\{\epsilon_m, \theta_m\}$ represents the normalized difference in signal arrival time and carrier phase between the channels[2], respectively, $E_{b,m}$ is the energy per encoded bit in the m-th channel, and Δf_m is the m-th center frequency.

Assume that M is an odd integer and that the center, or $(M+1)/2$th, channel conveys the desired data and that the other signals are viewed as being adjacent-channel interferers ($(M-1)/2$ ones on either side). Furthermore, in practical systems these channels are equally spaced in frequency, say by Δf. In terms of the above, then

$$\Delta f_m = \left(m - \frac{M+1}{2} \right) \Delta f, \quad m = 1, 2, \ldots, M. \tag{4}$$

Now, let $\{x_m((k+\epsilon_m)T_s); m = 1, \ldots, M\}$ be a set of sufficient statistics which consists of a bank of filters matched to the modulating signal in each channel, then sampled at the symbol rate of T_s^{-1} or,

$$x_m(t) \triangleq \int_{-\infty}^{\infty} \tilde{r}(\alpha) e^{-j(2\pi \Delta f_m \alpha + \theta_m)} p^*(\alpha - t) dt. \tag{5}$$

We also define

$$\mathcal{C}_{l,j}(t_1, t_2) \triangleq \left[\int p^*(\alpha) p(\alpha + t_2 - t_1) e^{-j2\pi(\Delta f_j - \Delta f_l)\alpha} d\alpha \right] \times \exp(-j(2\pi(\Delta f_j - \Delta f_l) t_2 + (\theta_j - \theta_l)). \tag{6}$$

We now attempt to characterize the effective channel that appears at the output of the matched filters bank: $\{x_m(t); m = 1, 2, \ldots, M\}$ of (5) in terms of the analytical impulse response, for which we substitute (1)–(3) into (5) to yield

$$x_m(t) = \left[\sum_{n=1}^{M} \sum_{i=-\infty}^{\infty} \sqrt{E_{b,n}} a_{n,i} \mathcal{C}_{n,m}((i + \epsilon_n)T_s, t) \right] + n_m(t). \tag{7}$$

From (7) it is clear that the equivalent lowpass interference channel is described by the impulse response $\mathcal{C}_{l,j}(t_1, t_2)$, defined in (6)[3]

[2]These parameters are assumed to be known *a priori* at the receiver.

[3]In (7), the $\{n_m(t); m = 1, 2, \ldots, M\}$ is a set of zero-mean complex Gaussian random processes with covariance

$$\mathbf{E}\{n_l^*(t) n_j(\acute{t})\} = N_0\, \mathcal{C}_{lj}(t, \acute{t}). \tag{8}$$

Focusing on Nyquist pulses, the matched filter statistic (7) can then be described as

$$\begin{aligned} x_m((k+\epsilon_m)T_s) = &\sqrt{E_{b,m}} a_{m,k} \\ &+ \sum_{i=-L}^{L} \sqrt{E_{b,m-1}} a_{m-1,k-i} \mathcal{C}_{m-1,m} \\ &\quad \times ((k - i + \epsilon_{m-1})T_s, (k + \epsilon_m)T_s) \\ &+ \sum_{i=-L}^{L} \sqrt{E_{b,m+1}} a_{m+1,k-i} \mathcal{C}_{m+1,m} \\ &\quad \times ((k - i + \epsilon_{m+1})T_s, (k + \epsilon_m)T_s) \\ &+ n_m((k + \epsilon_m)T_s). \end{aligned} \tag{9}$$

The value L denotes the memory associated with the interference and is defined as the effective single-sided length of the impulse response of the adjacent interfering channels measured at the desired matched filter output (i.e., the total length is $(2L+1)$). From basic principles of Fourier transforms, the value of L is larger for smaller overlap.[4] Finally, note that the ACI in (9) was assumed to result only from one adjacent interferer on either side. This is an acceptable approximation given the spectral overlap values considered in this paper.

III. Iterative Interference Cancellation Prior to FEC Decoding

In this scenario, the interference cancellation modules assumes uncoded data streams. This approach is less complex and imposes less constraints on the network-level synchronization when compared with the one that considers joint detection and decoding.

In the proposed iterative interference canceler[5], the zeroth-iteration data estimate, $\hat{a}_{m,k}^{(0)}$, is

$$\hat{a}_{m,k}^{(0)} = \text{Decision}\left(x_m((k+\epsilon_m)T_s); Th_0 \right) \tag{10}$$

where Decision $(x; Th)$ is a function that provides hard decisions on the real and imaginary parts in parallel. Namely, let $z = x + jy$ then

$$\text{Decision}(z; Th) = \text{Decision}\,(x; Th) + j\ \text{Decision}(y; Th) \tag{11}$$

and

$$\text{Decision}(x; Th) \triangleq \begin{cases} +1, & x > Th \\ 0, & |x| \leq Th \\ -1, & x < -Th. \end{cases} \tag{12}$$

In the lth iteration the data estimates from $(l-1)$th iteration are used to reconstruct the interference which is then subtracted from the matched filter bank outputs. Hence, the decision in the lth iteration is given by

$$\hat{a}_{m,k}^{(l)} = \text{Decision}\left(x_m((k+\epsilon_m)T_s) - \hat{I}_m^{(l)}((k+\epsilon_m)T_s); Th_l \right) \tag{13}$$

[4]As the channel spacing is increased, the amount of frequency overlap is smaller in magnitude and shorter in duration, so the time-domain counterpart is still smaller in magnitude but is longer in duration.

[5]This part of the paper is based on earlier work by the first author [7]. A similar approach was independently proposed later for CDMA signals in [8].

where

$$\hat{I}_m^{(l)}(kT_s) = \sum_{i=-L}^{L} \sqrt{E_{b,m-1}} \hat{a}_{m-1,k-i}^{(l-1)} \mathcal{C}_{m-1,m} \times ((k-i+\epsilon_{m-1})T_s, (k+\epsilon_m)T_s) + \sum_{i=-L}^{L} \sqrt{E_{b,m+1}} \hat{a}_{m+1,k-i}^{(l-1)} \mathcal{C}_{m+1,m} \times ((k-i+\epsilon_{m+1})T_s, (k+\epsilon_m)T_s). \quad (14)$$

This process is repeated $(K-1)$ times for a Kth iteration canceler, until finally the soft values passed to the decoder(s) are given by [6]

$$\hat{a}_{(m,k)}^{(K)} = x_m((k+\epsilon_m)T_s) - \hat{I}_m^{(K)}((k+\epsilon_m)T_s). \quad (15)$$

The decision device used here allows for a "comfort" zone of one-sided length of (Th) so that no decision is made in the region when the decision variable is of little reliability. The threshold value (Th) has been coarsely optimized for different iterations via simulation. This threshold device with three levels represents an "*ad-hoc*" attempt at making soft feedback decisions that would reduce the error propagation.

IV. Turbo Interference Cancellation

In this section, we consider the low complexity iterative linear mmse and soft interference cancellation algorithms. The iterative mmse approach uses the soft information supplied by the M single-user decoders to calculate the optimum, feed-forward and feedback, filter weights after each iteration. For this we extend the algorithm proposed in [1] for CDMA signals to the current application. Let $\underline{x}$ be a $[M(2L+1)\times 1]$ complex vector of the matched filter bank outputs from the $(k-L)$th to the $(k+L)$th samples. Then $\underline{x}$ can be written as

$$\underline{x} = \sum_{n=1}^{M} \sum_{l=k-2L}^{k+2L} \underline{\mathcal{C}}_{n,l} a_{n,l} + \underline{n} \quad (16)$$

where, for example, $\underline{\mathcal{C}}_{n,l} = \sqrt{E_{b,n}}[\mathcal{C}_{n,1}((l+\epsilon_n)T_s, (k-L+\epsilon_1)T_s), \ldots, \mathcal{C}_{n,M}((l+\epsilon_j)T_s, (k+L+\epsilon_M)T_s)]^T$ is the equivalent channel response associated with $a_{n,l}$,[7] and $\underline{n}$ is the Gaussian noise vector. Let, R^I, the interference correlation matrix, be defined as

$$R^I \triangleq [\underline{\mathcal{C}}_{1,k-2L} \ldots \underline{\mathcal{C}}_{m,k-1} \underline{\mathcal{C}}_{m,k+1} \ldots \underline{\mathcal{C}}_{M,k+2L}]^T \quad (17)$$

and

$$\underline{a} \triangleq [a_{1,k-2L}, \ldots, a_{m,k-1}, a_{m,k+1}, \ldots, a_{M,k+2L}]^T \quad (18)$$

then

$$\underline{x} = \underline{\mathcal{C}}_{m,k} a_{m,k} + R^I \underline{a} + \underline{n}. \quad (19)$$

[6]For uncoded systems, hard decisions are enforced at this stage.

[7]This corresponds to the elements of the correlation matrix in the CDMA literature.

The linear mmse receiver attempts to minimize the conditional mean square error (mse) between the output $y_{m,k}$ and the transmitted symbol $a_{m,k}$ given by

$$\begin{aligned} \text{mse} =& E\left[|y_{m,k} - a_{m,k}|^2 \,|\{\lambda\}_{m,k}\right] \\ =& E\left[\left|\underline{c}_f{}^T \underline{x} + c_b - a_{m,k}\right|^2 |\{\lambda\}_{m,k}\right] \\ =& E\left[\left|\underline{c}_f{}^T \left(\underline{\mathcal{C}}_{m,k} a_{m,k} + R^I \underline{a} + \underline{n}\right)\right.\right. \\ & \left.\left. + c_b - a_{m,k}\right|^2 |\{\lambda\}_{m,k}\right] \end{aligned} \quad (20)$$

where $\underline{c}_f$ is the $[M(2L+1)\times 1]$ feed-forward coefficients vector, c_b is the feedback coefficient, $\{\lambda\}$ is the set containing the extrinsic log-likelihood ratios provided by the M SISO decoders in the previous iteration [1], [2], $\lambda_{m,k}^I$ and $\lambda_{m,k}^Q$ are the extrinsic log-likelihoods for the in-phase and quadrature-phase bits of the mth user at the kth time interval, and $\{\lambda\}_{m,k} = \{\lambda\} \setminus \left\{\lambda_{m,k}^I, \lambda_{m,k}^Q\right\}$. It is straightforward to obtain the following solutions for the feed-forward and feedback filter coefficients

$$c_b = -\underline{c}_f^T R^I E\left[\underline{a}|\{\lambda\}_{m,k}\right] \quad (21)$$

$$\underline{c}_f^T = \underline{\mathcal{C}}_{m,k}^H \left(A + B + R^n - CC^H\right)^{-1} \quad (22)$$

where, by definition, we have

$$A \triangleq \underline{\mathcal{C}}_{m,k} \underline{\mathcal{C}}_{m,k}^H \quad (23)$$

$$B \triangleq R^I E\left[\underline{a}\underline{a}^H |\{\lambda\}_{m,k}\right] R^{I^H} \quad (24)$$

$$C \triangleq R^I E\left[\underline{a}|\{\lambda\}_{m,k}\right] \quad (25)$$

and R^n is the $[M(2L+1)\times M(2L+1)]$ noise covariance matrix which can be easily constructed using the component-wise relation (8). In (21)–(25), the $E[\underline{a}|\{\lambda\}_{m,k}]$ and $E\left[\underline{a}\underline{a}^H|\{\lambda\}_{m,k}\right]$ values are obtained from the following component-wise relations

$$E(a_{n,l}|\{\lambda\}_{m,k}) \approx E\left(a_{n,l}^I|\lambda_{n,l}^I\right) + jE\left(a_{n,l}^Q|\lambda_{n,l}^Q\right), \quad (26)$$

$$E\left(a_{n,l}^I|\lambda_{n,l}^I\right) = \frac{e^{\lambda_{n,l}^I} - 1}{e^{\lambda_{n,l}^I} + 1}, \quad (27)$$

$$E(a_{n,l}a_{n,l}^*|\{\lambda\}_{m,k}) = 1, \quad (28)$$

$$E(a_{n,l}a_{i,k}|\{\lambda\}_{m,k}) \approx E\left(a_{n,l}|\left\{\lambda_{n,l}^I, \lambda_{n,l}^Q\right\}\right) \times E\left(a_{i,k}|\left\{\lambda_{i,k}^I, \lambda_{i,k}^Q\right\}\right) \quad (29)$$

where (26) and (29) follow from the independence assumption justified by the random interleaving employed at the transmitters, (28) follows from the constant modulus feature of QPSK modulation, and (27) follows from [1], [2].

The soft interference cancellation algorithm can now be obtained as a lower complexity approximation of the linear mmse solution. Based on (21) and (22), $y_{m,k}$ can be written as

$$y_{m,k} = \underline{c}_f^T \left(\underline{x} - R^I E\left[\underline{a}|\{\lambda\}_{m,k}\right]\right). \quad (30)$$

By observing that the matrix inversion operation is only required to compute $\underline{c}_f^T$, the following approximation is proposed

$$\underline{c}_f^T \approx \underline{c}_{(m-1)(2L+1)+L+1}^T \quad (31)$$

where $e^T_{(m-1)(2L+1)+L+1} = [0, \ldots, 0, 1_{(m-1)(2L+1)+L+1}, 0, \ldots, 0]$. Hence,

$$y_{m,k} \approx x_{m,k} - \left(R^I E\left[\underline{a}|\{\lambda\}_{m,k}\right]\right)_{m,k}. \tag{32}$$

The complexity of this algorithm is a linear function of the product of the number of interfering users and the interference memory. It is interesting to note that the soft cancellation technique is essentially the same algorithm as that proposed under a different derivation for CDMA signals in [3].

Finally, we make some remarks on the restrictions imposed by the iterative approach for joint detection and decoding on the system design. First, this approach requires more complex receivers since SISO decoder(s) must be used and possibly iterated for more than one time. Second, since the soft information is only available after decoding, some form of network synchronization of the various channels is required at the frame level. Due to these restrictions, the interference cancellation approach that does not utilize decoder outputs, as outlined previously in Section III, may be more attractive for some practical applications.

V. Numerical Studies

Monte-Carlo simulations are implemented to evaluate the bit error rate performance and demonstrate the effectiveness of the proposed solutions. For the the coded scenario, we use the four-state, rate $1/2$ convolutional code with optimal free distance. The SISO decoders use the soft output Viterbi algorithm (SOVA) and, unless otherwise stated, the number of iterations is four. The pulse used is the root-raised cosine (RRC) pulse and all channels are transmitting at the same energy level. It is assumed that the frame length, including tail bits, is 100 symbols.

As a reference for the current state-of-the-art bandwidth efficiency for systems with single-user receivers, we choose the one that causes a degradation of about 0.1 dB in input E_b/N_0 at uncoded bit error rate of 10^{-3}. This corresponds to bandwidth efficiency level η [8] of about 0.86 transmitted QPSK symbols-per-second/Hz or channel spacing of $\Delta f = 1.1625 T_s^{-1}$ (Hz).

A. Iterative Interference Cancellation for Uncoded Systems

Fig. 1 includes the uncoded bit error rate performance of the different interference cancellation receivers. As a reference, we include the interference-free performance as the upper-bound to quantify the ability of the proposed receivers to suppress interference. Also, the performance of the conventional receiver is included for comparison. The spectral efficiency level used is in this figure is 1.163 symbols-per-second/Hz, or channel spacing of $\Delta f = 0.86 T_s^{-1}$ (Hz), which represents an improvement of about 35% in spectral efficiency. It is clear that while the

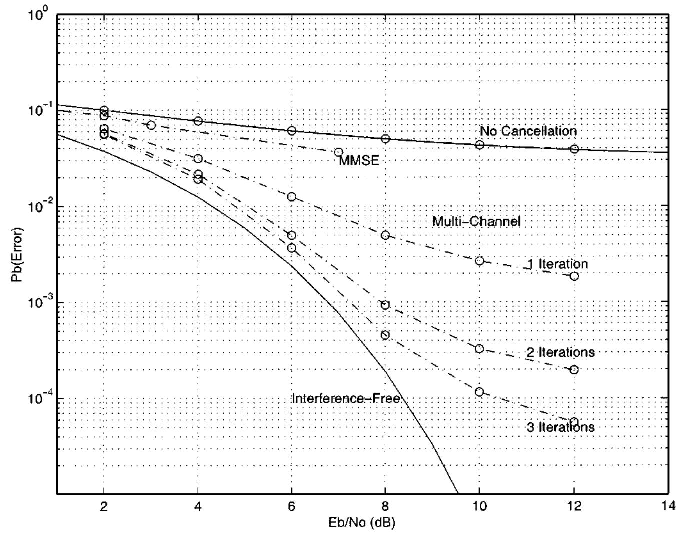

Fig. 1. Performance of the proposed methods at channel spacing of $0.86T_s^{-1}$ when interference cancellation is done prior to FEC decoding. (Solid line: analysis; "o": simulation).

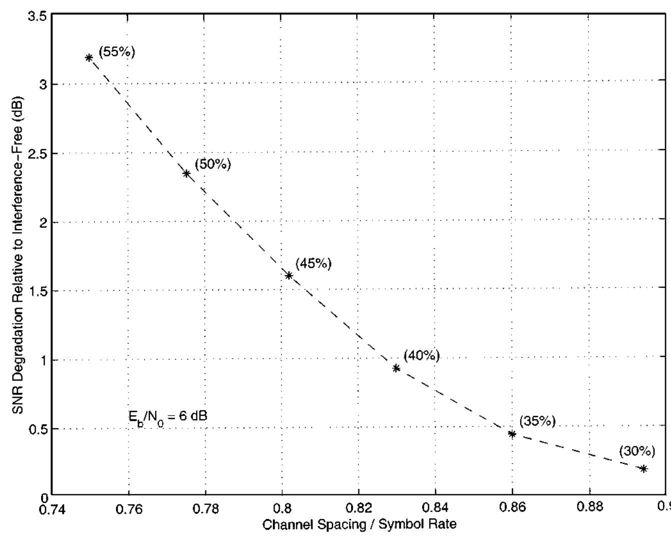

Fig. 2. Performance of the proposed subtractive interference canceler, when interference cancellation is done prior to FEC decoding, with varying levels of channel spacing.

mmse receiver[9] is ineffective, the additional energy requirement needed by the proposed iterative canceler is less than 0.5 dB compared with the interference-free bound for the SNR range of interest in the satellite applications of 3–6 dB.

Even though the proposed iterative canceler is highly effective compared with the linear solutions as shown in Fig. 1, there exists a limitation on the amount of improvement in the level of spectral efficiency when FEC decoding is *not* utilized in the interference cancellation. This is demonstrated in Fig. 2 which displays the performance of the three-iteration subtractive canceler in terms of the SNR degradation compared with the interference-free case as a function of the channel spacing employed. (The corresponding improvement in spectral efficiency relative to the current state-of-the-art system are shown in parentheses in the figure.) For example, the SNR degradation is about

[8]The spectral efficiency, η, in symbols/s/Hz is defined as

$$\eta \triangleq \frac{1}{\Delta f T_s} \quad \text{(symbols/s/Hz)}. \tag{33}$$

This definition is intended to represent the spectral efficiency of the channel independent of the channel code rate.

[9]The MMSE receiver here refers to the linear receiver without iterations.

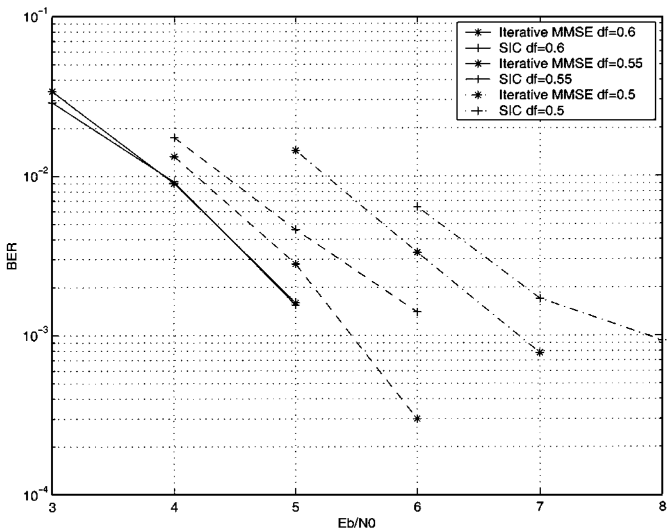

Fig. 3. Performance of the iterative mmse and SIC receiver at different channel spacings with the four-state code.

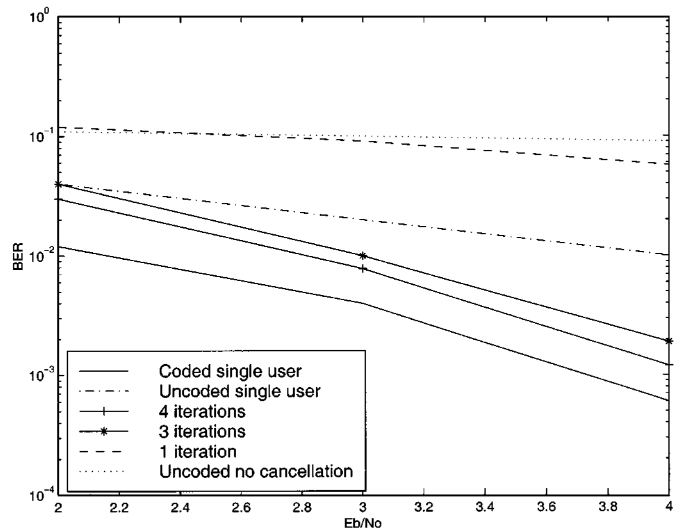

Fig. 4. Performance of the SIC receiver at channel spacing of $0.75T_s^{-1}$ with the four-state code.

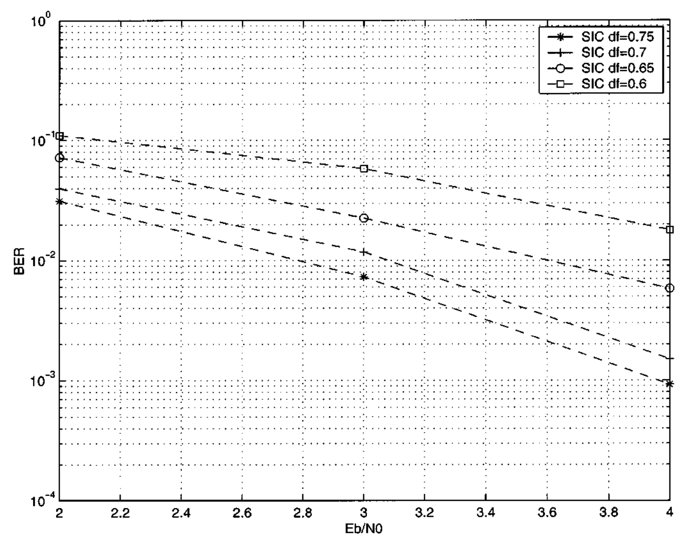

Fig. 5. Performance of the SIC receiver at different channel spacings with the four-state code.

3.2 dB with the highly-crowded channel packing situation of $\Delta f = 0.75T_s^{-1}$ (Hz). This gap is greatly reduced when implementing joint cancellation with FEC decoding.

B. Turbo Interference Cancellation

In Fig. 3, we compare the performance of the iterative mmse receiver with that of the soft interference cancellation (SIC) algorithm for different channel spacings. The roll-off factor of the RRC pulse is $\beta = 0.35$. In the figure, "df" refers to the normalized channel spacing $\Delta f T_s$. We assume that the number of channels transmitting simultaneously is three. As shown in the figure, the gain offered by the iterative mmse is more evident for aggressive channel packing scenarios. In the more relaxed scenario of $df = 0.6$, the two algorithms yield the same performance. Since in this paper we are mainly interested in low complexity receivers suited for practical applications, we will focus in the remainder of this section on investigating the performance of the SIC algorithm with $\Delta f \geq 0.6T_s^{-1}$.

Fig. 4 compares the performance of the SIC scheme with different numbers of iterations, the conventional receiver, and the interference-free system. The channel spacing considered here is $\Delta f = 0.75T_s^{-1}$ (Hz) which corresponds to an improvement of 55% compared to the current state-of-the-art. The receiver processes seven channels jointly in the presence of a total of nine QPSK sources. It is clear that the performance of the SIC algorithm is significantly better than the conventional receiver and is very close to the interference-free system, with a difference of less than 0.5 dB using four iterations at an input $E_b/N_0 = 4$ dB. It is also noted that as the E_b/N_0 increases, the performance gain offered by the iterative algorithm over the conventional receiver increases while the gap to the interference-free scenario diminishes. In Fig. 5, we report the performance of the SIC algorithm versus channel spacing. The roll-off factor of the RRC is $\beta = 0.6$. By limiting the additional E_b/N_0 needed in the crowded system to be less than 1 dB at $P_b(E) = 10^{-3}$, we can conclude that the minimum acceptable spacing is $\Delta f = 0.7T_s^{-1}$ in this scenario. This spacing corresponds to a 66% gain in throughput.

VI. Conclusion

In this paper, we investigated the design of iterative interference cancellation algorithms for high spectral efficiency satellite communications. We considered two classes of algorithms based on whether or not the FEC decoding information is utilized in the interference cancellation algorithm. The SISO iterative approach for joint interference cancellation and FEC decoding was shown to yield excellent results in practical scenarios with a reasonable receiver complexity.

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