ADAPTIVE EQUALIZATION AT HSDPA SYMBOL LEVEL

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ABSTRACT

Although a simple solution, pilot-symbols-trained LMS algorithm is not capable of tracking fast varying channels in UMTS FDD downlink due to insufficient adaptation rate. Pilot-chips-trained LMS adaptation is on the other hand much more prone to noise. These two phenomena manifest themselves in the two components of the adaptation excess mean square (EMSE). A compromise can be found by considering HSDPA symbol-level Griffiths or decision-directed equalization. These two methods enable adapting 16 times more frequently than the pilot-symbols-aided adaptation. They are also attractive for implementation since a modular approach can be adopted by exploiting either one or more of the available HSDPA code domains depending on the instantaneous channel quality and the performance requirements.

I. INTRODUCTION

High Speed Downlink Packet Access (HSDPA) has been standardized in the Release-5 of UMTS FDD standard [1]. In HSDPA, one or more of the High Speed Physical Downlink Shared Channels (HSPDSCHs) at spreading factor level 16 (SF-16), in particular 1,5,10 or 15 of the 16 available codes, are dynamically time multiplexed (scheduled) among users, preferably all allocated to a single user at any time. The goal is to exploit *multiuser diversity*, i.e. the temporal channel quality variance among the users, in order to increase the *sum capacity*, that is the total delivered payload by the BS. The choice for the type of scheduling mechanism is left to the operators. The criterion for the choice is the compromise between throughput and fairness [2, 3].

The initial stages of the UMTS development utilized only the Rake receiver due to its simplicity [4]. Although it is well known that Rake receiver is far from being optimal in multipath-rich environments, mobile vendors and chip manufacturers have been reluctant to switch to a more advanced and hence a more costly solution for the sole benefit of the base station side. In HSDPA, however, such a solution is meaningful due to the fact that the mobile terminal directly benefits from it by not only obtaining more data rate on HSPDSCHs once a connection is established during the scheduling process but also by increasing the probability of getting a connection if fairness is partially sacrificed for throughput in user scheduling. In any case it is inevitable since Rake receiver is not satisfying the 3GPP test performance requirements [5].

Max-SINR receiver which is also known as the chip level LMMSE equalizer followed by descrambling and despreading operations is an approximation of LMMSE receiver obtained from modeling the scrambler as a stationary random sequence [6, 7]. It is considered as the baseline receiver for HSDPA [5]. Max-SINR receiver can be implemented in several different ways. In this paper we are covering Griffiths and NLMS implementations at HSDPA symbol level.

II. TRANSMISSION MODEL

The baseband downlink transmission model of the UMTS-FDD mode system with HSDPA support is given in Figure 1.

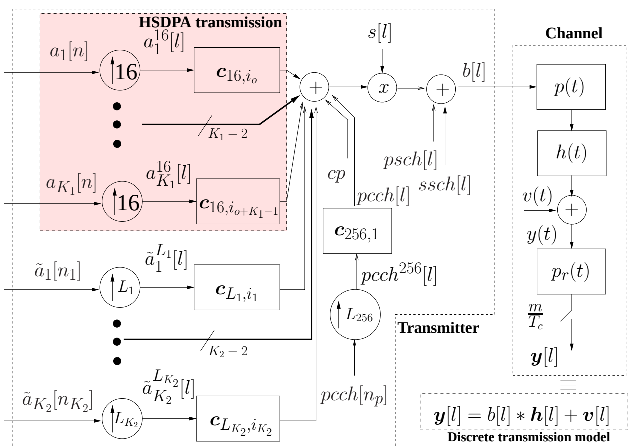

Figure 1: Downlink transmission and channel model

At the transmitter, the first group of K_1 i.i.d QPSK or 16-QAM modulated symbol sequences $\{a_1[n], a_2[n], \ldots, a_{K_1}[n]\}$ which belong to the HSDPA transmission are first upsampled by a factor of 16 and then convolved with their respective unit-amplitude channelization codes $\{\boldsymbol{c}_{16,i_o}, \boldsymbol{c}_{16,i_o+1}, \ldots, \boldsymbol{c}_{16,i_o+K_1-1}\}$ from OVSF code tree [8]. An important property that we will often exploit is that all the HSPDSCH symbols have the same power and the same modulation scheme.

The second group of multi-rate transmissions $\{\tilde{a}_1[n_1], \tilde{a}_2[n_2], \ldots, \tilde{a}_{K_2}[n_{K_2}]\}$[1] representing the dedicated physical channels (DPCHs), HSSCCHs and other control channels are similarly upsampled and convolved with their respective channelization codes $\{\boldsymbol{c}_{L_1,i_1}, \boldsymbol{c}_{L_2,i_2}, \ldots, \boldsymbol{c}_{L_{K_2},i_{K_2}}\}$.

[1] different symbol indices such as $n, n_1, n_2, \ldots, n_{K_2}, n_p$ are used to stress the *multi-rate* property of the transmission scheme

The third group of chip sequences associated with PCCPCH, PCPICH, PSCH and SSCH channels are explicitly demonstrated as $pcch[l]$, cp, $psch[l]$ and $ssch[l]$ respectively. It is particularly important for further discussion to state that cp is a positively scaled form of the unit vector $\frac{1+j}{\sqrt{2}}$ which has 45 degrees phase.

The sum of all the generated chip sequences is multiplied with the unit-energy BS-specific aperiodic scrambling sequence $s[l]$. PSCH and SSCH are the exceptions, multiplexed after the scrambler, since as a first-step task in the receiver they are utilized for determining, i.e. searching, which scrambling sequence is assigned to the BS. The resultant effective BS chip sequence $b[l]$ is transmitted to the channel with three cascade components with the order of a *root-raised-cosine* (rrc) pulse shape $p(t)$ with a roll-off factor of 0.22, the *time-varying multipath propagation* channel $h(t)$ and a receiver front-end filter $p_r(t)$ which is in general chosen to be again an rrc pulse shape with a roll-off factor of 0.22 due to the fact that the *raised cosine* (rc) result of the rrc-rrc cascade is a *Nyquist pulse* whose T_c-spaced disrete time counterpart is a single unit pulse at time instant 0. In this case the only ICI source[2] is $h(t)$. Alternatively a low pass antialiasing filter with a cutoff frequency between $\frac{1.22}{T_c}$ and $\frac{2}{T_c}$ might be considered as $p_r(t)$ in the case of twice chip rate sampling. The latter case is a reasonable choice for fractionally spaced equalizers [6, 7]. The *effective* continuous time channel is hence given as

$$h_{eff}(t) = p(t) * h(t) * p_r(t) \tag{1}$$

We assume that there is no beamforming, so the propagation channel and the effective overall channel are unique for all the transmitted data from the same BS.

III. PCPICH Symbol Level LMS Equalizer

Unlike TDMA systems like GSM, pilot-aided equalizer design for CDMA systems is very problematic [9]. In TDMA systems, common pilot signal is time-multiplexed with payload data. Therefore it is not interfered by any other same BS signal but only the co-channel interference coming from the other cells and the AWGN. However in CDMA systems like UMTS FDD downlink, pilot data, i.e. the PCPICH, is code-multiplexed with all the other existing users and the control channels [8]. Therefore, since affected by a high amount of interference, it cannot be used efficiently for training the equalizer weights at baud rate, i.e. at chip rate [9]. In order to remedy this situation Frank et.al considered first despreading the received signal with the PCPICH code, hence suppressing most of the interference over the PCPICH signal [10]. Later Petre et.al extended it to the fractionally spaced implementations [11].

A basic schematic of this *standard* scheme is shown in Figure 2. The architecture is made up of a tap delay line, descrambler s^* and despreader with the PCPICH channelization code $\boldsymbol{c}_{256,0}$ on each branch originating from the associated tap, a filter of length N equal to the number of channel taps and the standard LMS adaptation scheme. The adaptive filter input $\boldsymbol{x}_{n_p}$ serves as the input regressor, the filter output y_{n_p} is the PCPICH symbol estimate, $d[n_p]$ is the desired signal, i.e. the correct PCPICH symbol, $e[n_p]$ is the error signal. Adaptive filter is the standard NLMS algorithm that trains via $\boldsymbol{x}_{n_p}$, y_{n_p} and $d[n_p]$.

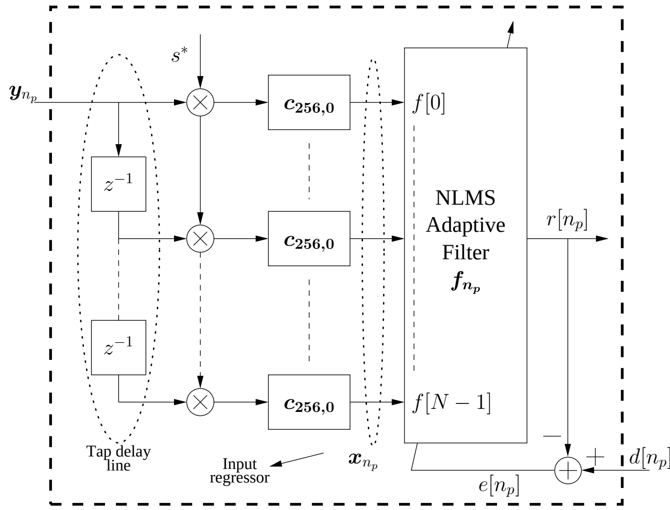

Figure 2: PCPICH Symbol Level NLMS Equalizer

Although this is quite an effective method in cases where the pilot code length is short, due to the long PCPICH code length, adaptation can only be done once every 256 chips in UMTS FDD downlink. This drives the technique to be slow in both converging and tracking the highly time-varying channels [9, 12].

IV. HSDPA Symbol Level Adaptive Equalizers

The proposed HSDPA symbol level adaptation schemes benefit from the knowledge of multiple HSDSCH codes and their identical power and constellation properties.

A. Griffiths Equalization at HSDPA Symbol Level

Griffiths adaptation scheme is shown in Figure 3. The filtering delay is taken to be $N-1$ chips.

When only one HSDSCH code domain, say the first one, is used, Griffiths Equalization at HSDPA symbol level can be derived starting from the Wiener filtering expression as

$$\boldsymbol{f}_{MMSE} = \boldsymbol{R}_{a_1 \boldsymbol{x}_{n,1}} \boldsymbol{R}^{-1}_{\boldsymbol{x}_{n,1}\boldsymbol{x}_{n,1}} \tag{2}$$

By making the connection between the cross-correlation term and the channel matched filter as[3]

$$\boldsymbol{R}_{a_1 \boldsymbol{x}_{n,1}} = \sigma^2_{a_1} \boldsymbol{h}^\dagger_n \tag{3}$$

[2] there is no notion of ISI in UMTS downlink since scrambling distorts the symbol level cyclostationarity

[3] $h^\dagger_n[i] = h^*_n[N-1-i], \quad i \in \{0, 1, \ldots N-1\}$, i.e. $\boldsymbol{h}^\dagger_n$ is the channel matched filter

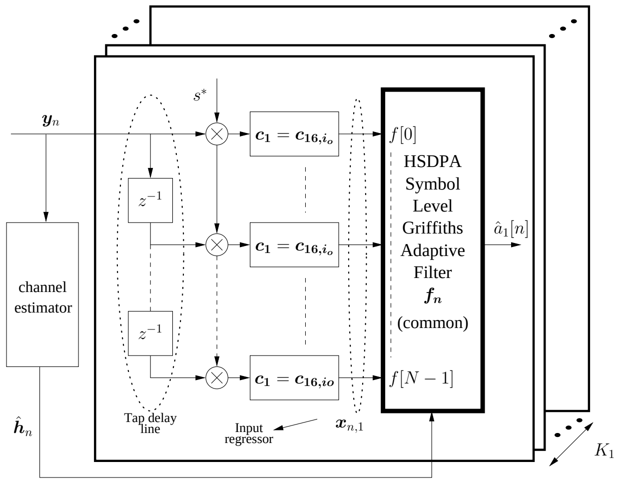

Figure 3: HSDPA Symbol Level N-Griffiths Equalizer

and replacing the regression vector covariance matrix by the instantaneous sample statistics, we reach to the Griffiths adaptation mechanism as

$$\begin{aligned} \boldsymbol{f}_{n+1} &= \boldsymbol{f}_n - \mu_{n,1}\nabla \boldsymbol{f}_{n,1} \\ &= \boldsymbol{f}_n - \mu_{n,1}(\boldsymbol{f}_n\boldsymbol{x}_{n,1}\boldsymbol{x}_{n,1}^H - \sigma_{a_1}^2\boldsymbol{h}_n^\dagger) \end{aligned} \tag{4}$$

where $\boldsymbol{f}_n$, $\mu_{n,1}$, $\nabla\boldsymbol{f}_{n,1}$, $\boldsymbol{x}_{n,1}$, respectively denote the filter weights, the step size, MSE gradient error vector and input regression vector for the first code at HSDPA symbol instant n.

The normalized form of the symbol level Griffiths adaptation scheme can be written as

$$\boldsymbol{f}_{n+1} = \boldsymbol{f}_n - \frac{\mu_{n,1}(\boldsymbol{f}_n\boldsymbol{x}_{n,1}\boldsymbol{x}_{n,1}^H - \sigma_b^2\boldsymbol{h}_n^\dagger)}{\boldsymbol{x}_{n,1}^H\boldsymbol{x}_{n,1}} \tag{5}$$

B. Decision Directed Equalization at HSDPA Symbol Level

Decision-directed adaptation scheme is shown in Figure 4.

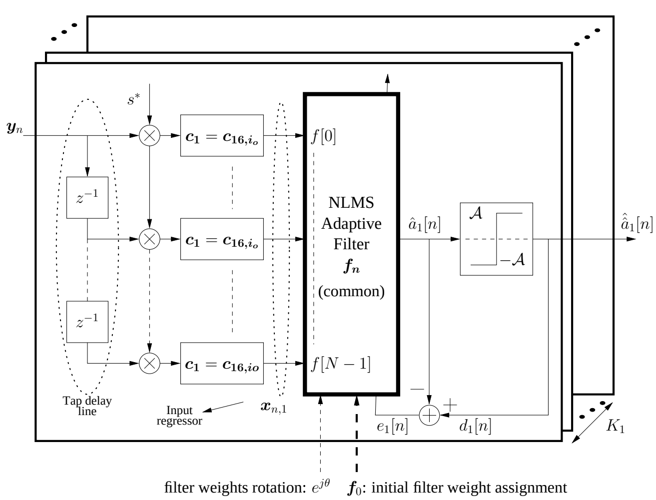

Figure 4: HSDPA Symbol Level DD-NLMS Equalizer

With only the first HSDSCH code domain, decision-directed LMS adaptation can be formulated as

$$\begin{aligned} \boldsymbol{f}_{n+1} &= \boldsymbol{f}_n - \mu_{n,1}\nabla \boldsymbol{f}_{n,1} \\ &= \boldsymbol{f}_n - \mu_{n,1}(\boldsymbol{f}_n\boldsymbol{x}_{n,1} - \hat{\hat{a}}_1[n])\boldsymbol{x}_{n,1}^H \\ &= \boldsymbol{f}_n + \mu_{n,1}e_1[n]\boldsymbol{x}_{n,1}^H \end{aligned} \tag{6}$$

where $\boldsymbol{f}_n$, $\mu_{n,1}$, $\nabla\boldsymbol{f}_{n,1}$, $\boldsymbol{x}_{n,1}$, $\hat{\hat{a}}_1[n]$ and $e_1[n]$ respectively denote the filter weights, the step size, MSE gradient error vector, input regression vector, the hard decided HSDSCH symbol which serves as the desired response and the error signal for the first code at HSDPA symbol instant n.

The normalized form of the symbol level DD-LMS adaptation scheme can be formulated as

$$\boldsymbol{f}_{n+1} = \boldsymbol{f}_n - \frac{\mu_{n,1}(\boldsymbol{f}_n\boldsymbol{x}_{n,1} - \hat{\hat{a}}_1[n])\boldsymbol{x}_{n,1}^H}{\boldsymbol{x}_{n,1}^H\boldsymbol{x}_{n,1}} \tag{7}$$

C. Misconvergence Problem

Any decision directed scheme is prone to misconvergence problem. This is a phenomenon which occurs when the equalizer locks to a rotated constellation state, does systematic errors all the time and cannot recover from there [12]. A common remedy is to have a backup solution such as a pure pilot-aided method or a constant modulus algorithm (CMA) [13, 14] which takes the turn when the equalizer diverges and gives the turn back to the decision-directed scheme when SINR conditions are again above an acceptable level [12, 15, 16, 17]. Since we want to *avoid* misconvergence without a backup solution, we obtain a Super-PCPICH-Symbol, i.e. sum of a block of PCPICH-symbols, every 5 or 10 PCPICH symbol periods and derotate the equalizer filter weights by an angle θ which is equal to the difference between the phase of the estimated Super-PCPICH Symbol and 45 degrees, the correct phase of the pilot signal. This, adds a local zero-forcing (ZF) dimension to the global MMSE equalization problem. The PCPICH tone is a significant element, not only for avoiding misconvergence most of the time but also bringing the filter back to convergence state if misconvergence cannot be avoided in deep fades. Consider the very initialization of the adaptation, for example. We are using the channel matched filter (CMF), i.e. the Rake receiver in its FIR form for the initialization of the filter weights. This is a nice-to-have but not a strictly essential feature. Even if we start with the all-zeros filter weights, by the aid of the CPICH zero-forcing mechanism, the filter passes the transient phase and the filter output locks to the correct constellation. This of course takes longer than starting with the CMF.

D. Extension to Multiple Code Domain Usage

Existence of K_1 equal amplitude and equal power HSDSCH codes is an opportunity to adapt symbol level DD-LMS and symbol level Griffiths schemes by better gradient estimates as

$$\nabla \boldsymbol{f}_n = \sum_{i=1}^{K_1} \nabla \boldsymbol{f}_{n,i} \tag{8}$$

so that

$$\boldsymbol{f}_{n+1} = \boldsymbol{f}_n - \mu_n \nabla \boldsymbol{f}_n \tag{9}$$

This multi-code adaptation diversity comes from the fact that the input regression vectors $\boldsymbol{x}_{n,k},\ k \in \{1, 2, \ldots, K_1\}$ at different code domains are uncorrelated.

The N-DDLMS and N-Griffiths adaptations are formulated as

$$\boldsymbol{f}_{n+1} = \boldsymbol{f}_n - \mu_n \sum_{i=1}^{i=K_1} \frac{\nabla \boldsymbol{f}_{n,i}}{\boldsymbol{x}_{n,i}^H \boldsymbol{x}_{n,i}} \tag{10}$$

The Mean Square Error (MSE) expression for (N)-LMS adaptation has two components: Minimum Mean Square Error (MMSE) and Excess Mean Square Error (EMSE). MMSE is the error floor performance of the Wiener Filter, which cannot be avoided. EMSE is the additional interference due to imperfect adaptation. It also has two ingredients as the stochastic gradient noise due to only one instantaneous sample support for obtaining the required adaptation statistics and the lag noise due to the time variation of the channel. Using K_1 codes decreases the stochastic gradient noise K_1 times if step size is chosen as $\mu_n = \frac{\mu_{n,1}}{K_1}$. The optimal step size compromising the two ingredients of EMSE would be a value $\frac{\mu_{n,1}}{K_1} < \mu_n < \mu_{n,1}$, exact value depending on the mobile speed, i.e. the time variation of the channel and the instantaneous noise level.

V. Simulations and Conclusions

For simulations, we consider 3GPP-RAN4 compatible HSDPA service scenarios in the UMTS FDD downlink for mobile terminals from Category 7 and Category 8 [5]. Table 1 shows the simulation settings.

Table 1: Simulation Settings

Parameters	**Settings**
Chip rate	3.84 Mcps
Number of HSPDSCH codes	10 (All belonging to the user)
Modulation scheme	QPSK
Total HSPDSCHs power (E_c)	50% of the BS power
$\hat{I}_{or}/I_{oc}$	6dB or 10dB
Orthogonal channel noise power	Remaining BS power randomly distributed to the remaining codes
Equalizer tap spacing	1 chip
Number of receive antennas	1
Equalizer length	24
Transmission pulse shape	rrc with roll-off factor 0.22
Channel model	Jakes fading model
Channel power delay profile	ITU Vehicular A
Relative Path Delays [ns]	[0 310 710 1090 1730 2510]
Relative Mean Power [dB]	[0 -1 -9 -10 -15 -20]
Mobile speeds	30km/h and 120km/h
Channel update rate	Once every 16 chips

On Figure 5, Figure 6 and Figure 7 {D, C, SD, SG, M, G, P} respectively denote {DD-NLMS with all codes, CMF, one-code DD-NLMS, one-code N-Griffiths, Max-SINR, N-Griffiths with all codes, PCPICH symbol level NLMS} and $\{\hat{I}_{or}, I_{oc}, E_c\}$ respectively denote {chip level received BS signal power, additive white noise power modeling also the intercell interference, total power assigned to HSPDSCH codes}.

For plotting convenience we sample the instantaneously obtained SINR at every slot of 160 HSDPA symbol periods.

Step size for each scheme was set to the best performing step size among exhaustive number of trials in a dynamic range between 0.005 and 0.4.

CMF serves as the SNR bound and Max-SINR receiver serves as the SINR bound. Therefore, for CMF and Max-SINR we assumed that we have ideal channel information. For N-Griffiths, however, to have a reasonably fair comparison with DD-NLMS scheme, we perturbed the correct channel parameters by adding random Gaussian noise with the normalized MSE values -8dB, -7dB and -5dB for Figure 5, Figure 6 and Figure 7 respectively. These are some judiciously chosen values taking into account the performance of the channel estimation methods in [18]. They can be modified to reflect the performance of any particular channel estimation technique.

Although PCPICH symbol level NLMS scheme performs better than CMF at 30km/h UE speed, it performs much worse at 120km/h. The crossover UE speed when CMF starts performing better is about 55km/h[4].

When all the HSDSCH code domains are exploited for adaptation, DD-NLMS performs better than N-Griffiths in most of the cases, approaching the Max-SINR performance at high SNR regions but performs worse in low SNR regions. When only one code is used, however, N-Griffiths showed better speed of convergence characteristics than DD-NLMS. One can attribute this to the fact that the cross-correlation term in the adaptation of the N-Griffiths scheme is common for all the gradient vectors. Therefore there is more correlation between N-Griffiths filter update vectors and hence less difference between one code and multiple codes adaptation qualities, which in fact makes one code N-Griffiths implementation a very attractive solution due to its low complexity.

Although not shown on plots, we tested the N-Griffiths performance with different channel estimation NMSE values and did not observe much difference. Hence we reached to the conclusion that N-Griffiths is a very robust scheme against channel estimation errors.

An ideal implementation strategy would be to adapt also the number of used codes and even more to switch between N-Griffiths and DD-NLMS schemes depending on the channel conditions, in particular N-Griffiths serving as an eye-opener for N-DDLMS. A more costly solution would be to concurrently run the N-Griffiths and DD-NLMS schemes and update filter weights by a weighted combination of their gradient vectors. This would in particular be useful with 16-QAM modulation. Such methods have been especially considered in literature between constant modulus (CM) and DD-LMS schemes, CM taking the

[4] not shown in simulations

role of the eye-opener [16].

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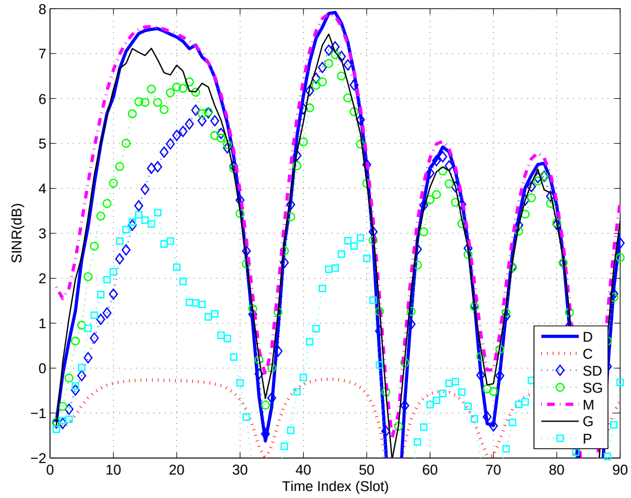

Figure 5: vA30, 10 codes, $E_c/\hat{I}_{or} = -3dB$, $\hat{I}_{or}/I_{oc} = 10dB$

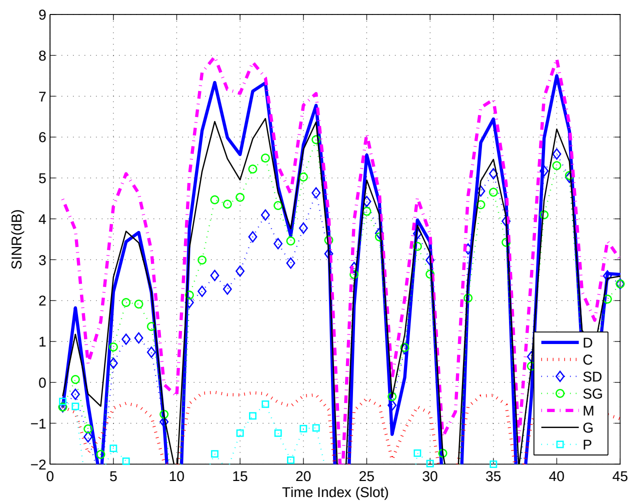

Figure 6: vA120, 10 codes, $E_c/\hat{I}_{or} = -3dB$, $\hat{I}_{or}/I_{oc} = 10dB$

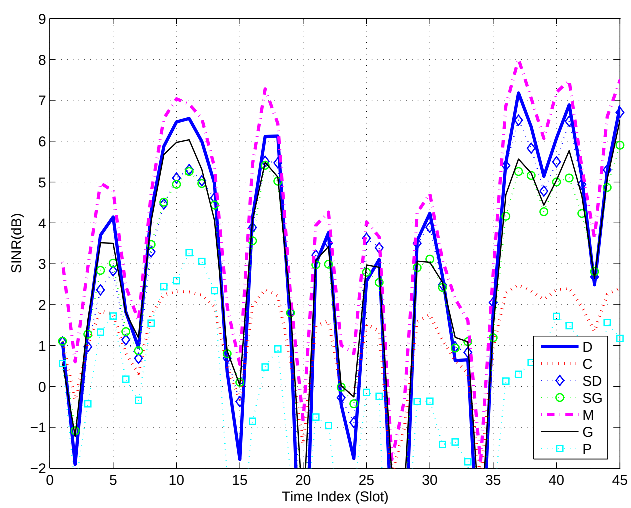

Figure 7: vA120, 5 codes, $E_c/\hat{I}_{or} = -3dB$, $\hat{I}_{or}/I_{oc} = 6dB$