論文 96-21-4-14

TDMA Jammer Suppression on CDMA Overlay

Dong-ku Kim*, Hyung-Il Park** *Regular Members*

正會員 김 동 구*, 박 형 일**

본 연구는 연세대학교 학술연구비 지원에 의하여 수행되었음.

ABSTRACT

The effect of inband TDMA narrow band jammers to DS-CDMA system performance and the suppression techniques are investigated using Monte Carlo simulations.

TIA standard North American Digital Cellular wea used as jammer. Levinson Dubin and conventional recursive least square algorithm were emphasized since these techniques can be implemented with a few DSPs for CDMA application. Two filter structures, i.e., complex suppression filter and real suppression filter in each inphase and quadrature channels, are investigated and their performances are compared. Complex suppression filter with Levinson Durbin algorithm of 20msec update rate is the most promising with respect to implementation and performance point of view. Implemenation feasibility is discussed and the channel capacity lost by suppression is computed.

요 약

주파수 사용효율을 높이기 위해, TDMA 시스템과 CDMA 시스템을 같은 주파수대역에서 사용하는 경우 (TDMA overlaid by CDMA) 또는 unlicensed 주파수대역에서 DS-cDMA를 이용해 통신하는 경우, DS-CDMA 시스템 성능에 대한 inband 협대역 Jammer를 제거하는 여러 방법에 대해서 시뮬레이션을 하였다.

본 연구에서는 TDMA USDC 신호를 Jammer신호로 사용하였다. 여러 방법중 적은 수의 DSP로 실현할 수 있는 Levinson Durbin과 recursive least square 알고리듬에 연구의 중점을 두었다. 두개의 필터구조, 다시말해서 real suppression filter와 complex suppression filter성능에 대해 연구했으며, filter update rate 등이 연구 비교되었다. 20msec마다 suppression filter의 파라미터를 적응시키는 Levinson Durbin complex suppression filter가 가장 적절한 방법임을 시뮬레이션을 통해 보였다. 또한 Jammer를 제거한 후 잃어버린 채널 용량도 연구되었다.

*연세대학교 전파공학과
**연세대학교 전자공학과
論文番號:94244
接受日字:1994年

Ⅰ. INTRODUCTION

Various jammer suppression techniques have been studied[1][2][3]. In this paper, feasible jammer suppression technique using a few DSPs for Code Division Multiple Access(CDMA) system which are designed to share common spectrum with non-CDMA system (narrow band waveforms), is presented. Specially TDMA jammer such as TIA standard North American Digital Celluar system is interested. It is assumed that these narrowband signals are being overlaid by a CDMA system in order to increase the overall spectral efficiency of the frequency band and therefore they do not represent intentional jamming. Rather, the interference that they impose on the CDMA waveforms is the unavoidable result of attempting this type of spectral sharing, and it is the purpose of the interference suppression techniques to minimize the resulting degradation to system performence. In some applications, frequency spectrum sharing is unavoidable and suppression might be needed all the time.

Specially, TIA/North American Digital Cellular (called USDC) system overlaid by IS-95 CDMA system is considered in this paper. The performances of jammer suppression are compared for each jammer suppression strategy, filter structure, and adaptive rate.

In Section 2, the communication system of interest is described. In Section 3, the adaptive algorithms used in the paper are explained and justified. In Section 4, suppression filter structures are described. In Section 5, jammer model is specified. In Section 6, simulation results are listed and compared. Section 7 is conclusions.

Ⅱ. SYSTEM DESCRIPTION

For the non-CDMA users with strong power in which 21 dB processing gain of IS-95 is not enough to combat, notching those non-CDMA users before despreading is inevitable. Figure 1 shows mobile transmitter. Speech signal is digitally encoded with QCELP and convolutionally encoded in which code,

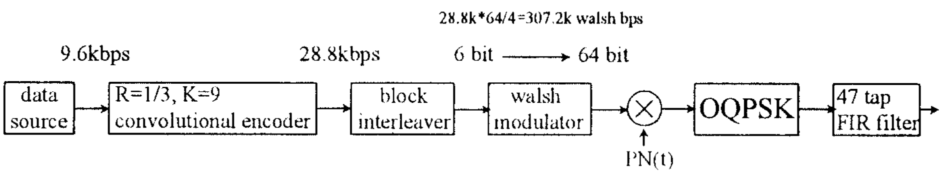

Fig 1. Mobile transmitter

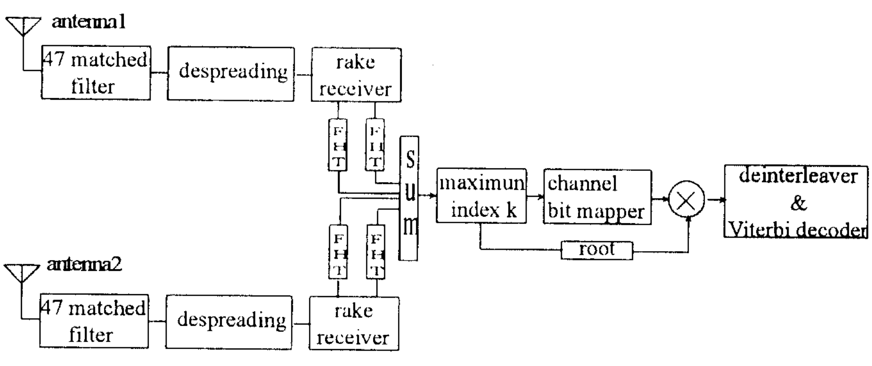

Fig 2. Base station receiver

rate is 1/3 and constraint length is 9. Encoded bit stream is interleaved for 20 msec. 6 coded bits are transformed to Walsh code 64 binary Walsh bits, then the binary data stream is spread by PN code, and modulated by OQPSK(Offset QPSK) and filtered by digital FIR filter which is specified in IS-95. Figure 2 shows a base station receiver. It uses antenna diversity. Rake receiver at each antenna has 2 fingers. The received signal at each fingers are despread and go through FHT(Fast Hadamard Transform). Each antenna is connected to each Rake receiver which has two branches. FHT is performed on each 4 branches and produces a 64 by 1 vector. 4 FHT outputs from each branches are summed and produces one 64 by 1 vector. The index of the 64 by 1 vector having the largest power is selected where the largest is defined as soft information of the vector. A Walsh code is transformed to channel bits by selecting 6 channel bits corresponding to selected index. 6 channel bits are multipied by the soft information. One frame coded bits are stored, deinterleaved and decoded by Viterbi Algorithm. Figure 3 shows additive white Gaussian noise channel model. There is one path from a transmitter to each antennas of base station. AWGN 1 and 2 are statistically independent and the same TDMA jammer signal is added to both paths.

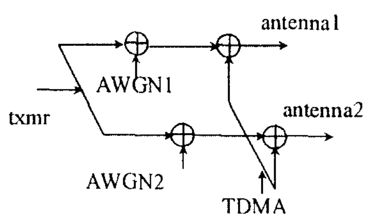

Fig 3. Reverse channel model

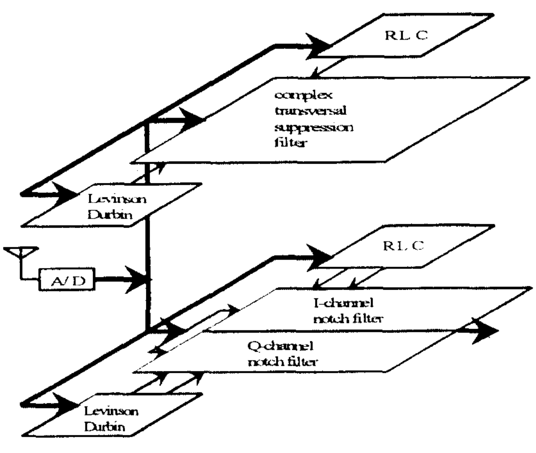

Fig 4. Complex and real suppression filter

To suppress the TDMA signal, suppression filter is placed between A/D and despreading. Figure 4 shows complex suppression filter and real suppression filter. Complex suppression filter notches I-channel and Q-channel components of jammer jointly and real suppression filter notches I-channel and Q-channel components of jammer separately.

Ⅲ. ADAPTIVE ALGORITHM

In this section, the adaptive jammer suppression algorithms applied in our application are shown.

3.1 Recursive Least-Squares Estimation(RLS)

Given the least squares estimate of the tap-weight vector of the filter at time n-1, the updated estimate of this vector at time n upon the arrival of new data is computed. The RLS is explained by exploiting a relation in matrix algebra known as the matrix inversion lemma. It utilizes information contained in the input data, extending back to the instant of time when the algorithm is initiated. In the method of exponentially weighted least squares, we minimize the cost function

$$g(n)=\sum_{i=1}^{n} \lambda^{n-i} \mid e(i)\mid^2 \tag{1}$$

where $e(i)$ is the difference detween desired response and output.

The optimum value of the tap-weight vector, $\hat{\mathbf{w}}(n)$, of the filter for which the cost fucntion g(n) attains its minimum value is defined by the normal equation written in matrix form;

$$\Phi(n)\,\hat{\mathbf{w}}(n)=\theta(n) \tag{2}$$

where the M-by-M correlation matrix $\Phi(n)$ is defined by

$$\Phi(n)=\sum_{i=1}^{n}\lambda^{n-i}\,\boldsymbol{u}(i)\,\boldsymbol{u}^{H}(i). \tag{3}$$

The M-by-1 cross-correlation $\theta(n)$ between the tap inputs of the transversal filter and the previous one $u(k-1),\cdots,u(k-n)$ is defined by

$$\theta(n)=\sum_{i=1}^{n}\lambda^{n-i}\,\boldsymbol{u}(i)\,d^{*}(i) \tag{4}$$

where the asterisk denotes complex conjugatation and $d(i)$ is desired response. By doing some calculation, the resulting RLS algorithm is expressed as follows.

$$\mathbf{k}(n)=\frac{\lambda^{-1}\,\mathbf{P}(n-1)\,\boldsymbol{u}(n)}{1+\lambda^{-1}\,\boldsymbol{u}^{H}(n)\,\mathbf{P}(n-1)\,\boldsymbol{u}(n)} \tag{5}$$

$$\alpha(n)=d(n)-\hat{\mathbf{w}}^{H}(n-1)\,\boldsymbol{u}(n) \tag{6}$$

$$\hat{\mathbf{w}}(n)=\hat{\mathbf{w}}(n-1)+\boldsymbol{k}(n)\,\alpha^{*}(n) \tag{7}$$

$$\mathbf{P}(n)=\lambda^{-1}\,\mathbf{P}(n-1)-\lambda^{-1}\,\boldsymbol{k}(n)\,\boldsymbol{u}^{H}(n)\,\mathbf{P}(n-1) \tag{8}$$

$\mathbf{k}(n)$: gain vector
$\alpha(n)$: priori estimation error
$\mathbf{P}(n)=\Phi^{-1}(n)$.

3.2 Levinson Durbin Algorithm

Basically, the procedure utilizes the solution of the augmented Wiener-Hopf equations for a prediction-error filter of order m-1 to compute the corresponding solution for a prediction-error filter of order m(i.e., one order higher). The order $m=1, 2,\cdots,M$, where M is the final order of the filter. The important virtues of the Levinson Durbin algorithm is its computaional efficiency. To derive the Levinson Durbin recursive procesure, we will use the matrix formulation of both forward and backward predictions.

Let the $(m+1)$-by-1 vector $\boldsymbol{a}_m$ denotes the tap-weight vector of a forward prediction-error filter of order m. The $(m+1)$-by-1 tap-weight vector of the corresponding backward prediction-error filter is obtained by backward rearrangement of the elements of vector $\boldsymbol{a}_m$ and their complex conjugation. We denote the combined effect of these two operations by $\boldsymbol{a}_m^{B^*}$. Let the m-by-1 vectors $\boldsymbol{a}_m$ and $\boldsymbol{a}_m^{B^*}$ denote the tap-weight vectors of the corresponding forward and backward prediciton-error filters of order m-1, repectively.

The tap-weight vector of a forward prediction-error filter may be order-updated as follows.

$$\boldsymbol{a}_m=\begin{bmatrix}\boldsymbol{a}_{m-1}\\0\end{bmatrix}+\Gamma_m\begin{bmatrix}0\\\boldsymbol{a}_{m-1}^{B^*}\end{bmatrix} \tag{9}$$

where Γ_m is a constant. The scalar version of this order update is

$$a_{m,k}=a_{m-1,k}+\Gamma_m\,a^{*}_{m-1,m-k},\ k=0,\ 1,\cdots,m \tag{10}$$

where $a_{m,k}$ is the kth tap weight of a forward prediction-error filter of order m and $a_{m-1,0}$, $a_{m-1,m}=0$.

The tap-weight vector of a backward prediction-error filter may be order-updated as follows.

$$\boldsymbol{a}_m^{B^*}=\begin{bmatrix}0\\\boldsymbol{a}_{m-1}^{B^*}\end{bmatrix}+\Gamma_m^{*}\begin{bmatrix}\boldsymbol{a}_{m-1}\\0\end{bmatrix} \tag{11}$$

The scaler version of this order update is

$$a^{*}_{m,m-k}=a^{*}_{m-1,m-k}+\Gamma^{*}_{m}a_{m-1,k},\ k=0,\ 1,\cdots,m \tag{12}$$

To establish the condition that the constant Γ_m has to satisfy in order to justify the validity of the Levinson Durbin algorithm, we proceed some stages as follows.

Premultiplying both sides of Eq.(9) by $\mathbf{R}_{m+1}$, the $(m+1)$-by-$(m+1)$ correlation matrix of the tap inputs $u(n)$, $u(n-1),\cdots,u(n-m)$ in the forward prediction-error filter of order m, we get

$$\mathbf{R}_{m+1}\,\mathbf{a}_m=\begin{bmatrix}P_m\\0_m\end{bmatrix} \tag{13}$$

where P_m is the forward prediction-error power. Then the forward prediction-error filter of order m-1 is

$$\mathbf{R}_m \mathbf{a}_{m-1} = \begin{bmatrix} P_{m-1} \\ 0_{m-1} \end{bmatrix} \tag{14}$$

and define the scalar

$$\begin{aligned} \Delta_{m-1} &= \boldsymbol{r}^{BT}{}_m a_{m-1} \\ &= \sum_{k=0}^{m-1} a_{m-1,k}\, r(k-m) \end{aligned} \tag{15}$$

where $\boldsymbol{r}_m^{B*}$ is the cross-correlation vector between tap inputs and desired response $u(n\text{-}m)$.

Following similar precedure, the backward prediction-error filter of order m-1 is

$$\mathbf{R}_m \mathbf{a}_{m-1}^{B*} = \begin{bmatrix} 0_{m-1} \\ P_{m-1} \end{bmatrix} \tag{16}$$

and define the scalar

$$\begin{aligned} \Delta^{*}_{m-1} &= \boldsymbol{r}_m^H\, \boldsymbol{a}_{m-1}^{B*} \\ &= \sum_{l=0}^{m} a^{*}_{m-1,m-1}\, r(l) \end{aligned} \tag{17}$$

where $\boldsymbol{r}_m$ is the m-by-1 cross-correlation vector between tap inputs and desired response $u(n)$.

Finaly we get

$$\begin{bmatrix} P_m \\ 0_m \end{bmatrix} = \begin{bmatrix} P_{m-1} \\ 0_{m-1} \\ \Delta_{m-1} \end{bmatrix} + \Gamma_m \begin{bmatrix} \Delta^{*}_{m-1} \\ 0_{m-1} \\ P_{m-1} \end{bmatrix} \tag{18}$$

If the conditions desribed by Eq.(18) apply, the tap-weight vector of a forward prediction-error filter may be order-updated as in Eq.(9).

$$P_m = P_{m-1} + \Gamma_m \Delta^{*}_{m-1} \tag{19}$$

$$\Gamma_m = -\frac{\Delta_{m-1}}{P_{m-1}} \tag{20}$$

$$P_m = P_{m-1}(1 - |\Gamma_m|^2) \tag{21}$$

$P_0 = r(0)$ where $r(0)$ is the autocorrelation function of the input process for lag zero.

The prediction-error filter of of final order M equals

$$P_M = P_0 \prod_{m=1}^{M} (1 - |\Gamma_m|^2). \tag{22}$$

Basically, these two algorithms all estimate correlation function of received signals, then compare optimum coefficients of prediction filter based on the estimated correlation functin using batch method or recursive method. All these adaptive techniques compute $\{a_i\}$ which minimizes

$$\sum_{n=0}^{t} \lambda^{t-n} \, | x(n) - \sum_{i=1}^{m} a_i(n) x(n-i) |^2 \tag{23}$$

where n implies $n \times \mathrm{T_c}$ where $\mathrm{T_c}$ means chip duration or chip time, and $x(n)$ is a received signal at time n. The optimum filter[1] is presented as follows,

$$a_m(n) = \begin{bmatrix} a_1(n) \\ a_2(n) \\ \vdots \\ a_m(n) \end{bmatrix} = R_X^{-1} R_{x(n)X} \tag{24}$$

where $R_X = E\{X(n)X^{*t}(n)\}$, $X(n) = \begin{bmatrix} x(n-1) \\ x(n-2) \\ \vdots \\ x(n-m) \end{bmatrix}$,

$R_{x(n)X} = E\{x(n)X^{*t}\}$.

Since CDMA signal samples are independent over a chip time, the correlation function of CDMA signal can be considered as 0 after $\mathrm{T_c}$ delay, i.e., correlation function of psuedorandom sequence $R_{pn}(\tau) = 0 \,|\tau| > \mathrm{T_c}$. Therefore $\{a_i(n)\}$ predicts only the correlated jammer, i.e., narrow band jammer components in the way of minimizing Eq.(23). That is, FIR linear predictor in which each tap has chip delay does not depend on desired CDMA signals.

It was thought that suppression filter at the base station has to be realized with digital signal processing chips rather than ASIC because of the limited number of base stations. In this paper, the adaptive algorithms which can be implemented with a few DSPs and slow updating rate are of interest. If a receiver can accommodate the computational complexity required to update filter coefficients for each chip time, fast algorithms such as fast recursive least square and FAEST[5] may be the best choices and their computational complexity is proportional to filter length N. However, chip time update rate cannot be accomplished with limited number of DSPs at the the 2.4567MHz which is two times over sampling rate of IS-95. Besides that, updating slower than chip rate does not hold shifting property that fast algorithm has to satisfy, so that fast algorithms cannot be used with lower update rate than chip rate when suppression algorithm is implemented with a limited number of DSPs. Even though the computation complexity required for updating conventional RLS is proportional to N^2 while that of fast algorithm is proportional to N, RLS can be updated under the chip rate, i.e., slow updating rate, and can be implemented by off-the-shelve DSPs and ICs.

Ⅳ. SUPPRESSION FILTER

Two different suppression filter structures, i.e., complex and real suppression were investigated. Suppression filter is tapped delay line structure in which delay is fixed to a chip time.

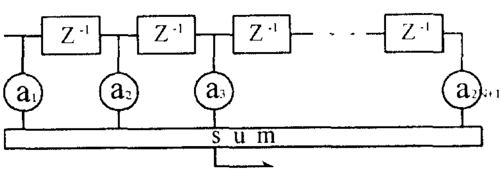

Fig 5. Double sided symmetric suppression filter

There are two key reasons of fixing a tap delay to a chip time. One is that if tap delay is less than a chip time, suppression filter would suppress signal as well as jammer because CDMA signal is correlated within a chip time. The other is that if tap delay is greater than a chip time, frequency response of suppression filter has more than one notch in CDMA band. Therefore tapped delay line is fixed to a chip time. Complex suppression filter is implemented with 4 real filters which is twice as many as suppression filters. Adaptive algorithm with complex filter requires as twice as many computations that adaptive algorithm with real suppression filters require.

$2N+1$ filter coefficients can be obtained by computations proportional to N^2 in the following way. Adaptive algorithm computes length N linear single sided prediction filter coefficients and convolves it with is matched filter to make $2N+1$ double sided filter coefficients. For example, the 31 tap transversal double-sided suppression filter is obtained in the way that 15 tap transversal linear prediction filter is formed and adjusted adaptively by Levinson Durbin or RLS algorithm and then convolved with its matched filter.

Ⅴ. JAMMER MODEL

TIA North American standard Digital Cellular (USDC) signals are used as TDMA jammer model throughout the simulations. USDC adapts 48.8Kbps, 24.4K symbol/sec, square root raised cosine pulse with 0.35 roll-off factor, 1/3 duty factor out of 20 msec. In the simulation, square root raised cosine filter with 6 symbol duration was used.

Ⅵ. SIMULATION RESULTS

While PCM(pulse code modulation) requires 10^{-4} information bit error rate, QCELP requires 10^{-3} information bit error rate(BER) for speaker intelligibilty. The information bit energy to noise energy ratio, E_b/N_o, to achieve 10^{-3} information bit error rate(BER) is computed for each suppression algorithm, filter structures, update rates, and multiple number of

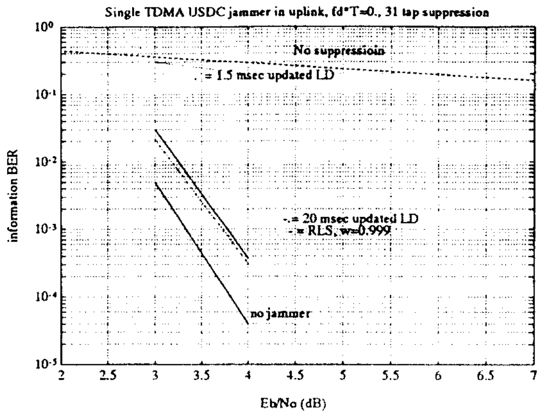

Fig 6.

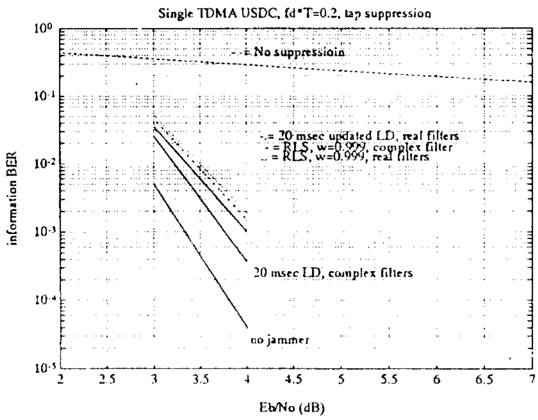

Fig 7.

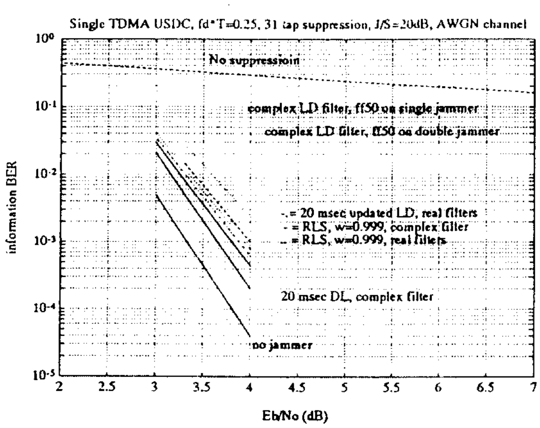

Fig 8.

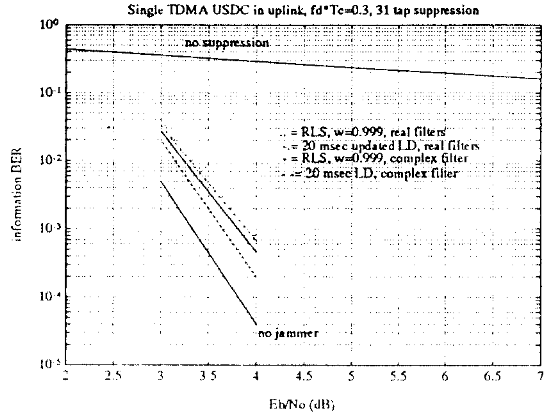

Fig 9.

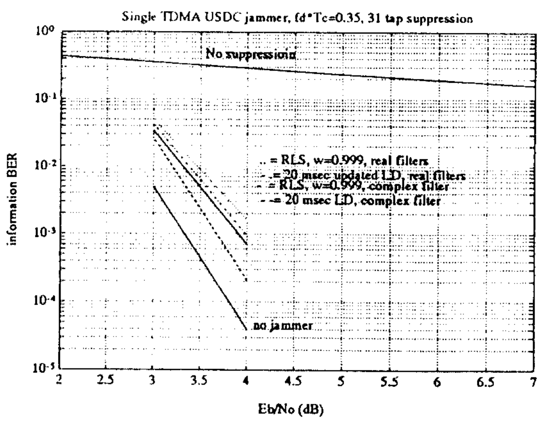

Fig 10.

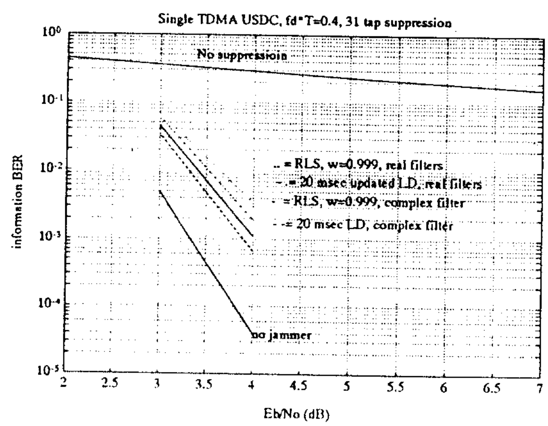

Fig 11.

jammers.

Figure 6 shows tha the E_b/N_o for 0.1% information bit error rate is 3.3 dB in additive white Gaussian noise(AWGN) channel. Let's denote J as a jammer power and S as a CDMA signal power. Let's define f_d as the frequency offset between CDMA carrier frequency and jammer's center frequency. The range of $T_c \times f_d$ is over 0 to 0.5 in the sumulation. Levinson Durbin algorithm updates suppression filter tap coefficients every 20 msec(1 frame), and RLS updates the filter every $110T_c$ and $75T_c$ for complex and real suppression each. These update rate can be realized with 2 DSPs(30MMACS/DSP for Levison Durbin, 20MMACS/DSP for conventional RLS). 2,000 to 4, 000 frames are simulated to get BER for each E_b/N_o value. The first frame that a jammer starts attacking CDMA signals, is always in error for Levinson Durbin case, but the first frame may not be in error for most values of $T_c \times f_d$ with proper initialization for RLS case. Levinson Durbing can be implemented with fixed point DSP while RLS requires floating point DSP. Figure 6 shows information BER when $T_c \times f_d$ 0, i.e., TDMA USDC is at the center of CDMA signal band J/S i 20 dB and 31 tap symmetric suppression filter. The suppression filter shows outstanding performance improvement for TDMA jammer compared to the performance without suppression. However, E_b/N_o required for 0.1% BER after suppression is half dB worse than no jammer and no suppression filter case. Note that the performance curve of 1.5 msec update Levinson Durbin shows no improvement. It implies that when jammer power is very strong, continuous suppression is a better strategy than tracking on-off characteristics of TDMA jammer. Figure 6~13 shows the information BER for various $T_c \times f_d$, complex and real suppression filters, and Levinson Durbin and RLS with memory factor 0.999. Figure 7 compares the performance of complex and real suppression filters with RLS algorithm. In the case that $T_c \times f_d$ is 0.2, the complex filter updated by Levinson Durbin algorithm with 20 update rate shows the best performance and 0.5 dB performance degradation compared to that of no jammer and no suppression. Figure 8 is the case that $T_c \times f_d$ is 0.25. The result is similar to Figure 7. The performance degradation is less than 0.5 dB. Figure 9 and 10 are the case that $T_c \times f_d$ is 0.3 and 0.35. Symmetric complex filter using Levinson Durbin algorithm with 20 msec update rate shows the best performance and degradation due to suppression filter is less than 0.5 dB. Figure 11 is the case that $T_c \times f_d$ is 0.4 and shows the worst performance among the simulation results. The degradation due to suppression is around 0.6 dB. Figure 12 is the case that $T_c \times f_d$ is 0.45 and degradation due to suppression is less than 0.5 dB. Figure 13 is the case that $T_c \times f_d$ is 0.5, in which f_d is $1/2T_c$. That is, jammer is located at the edge of main lobe of CDMA power spectrum. Therefore degradation due to suppression is around 01. dB.

It can be concluded tat 20 msec update Levinson Durbin with complex suppression shows the best performance among them. E_b/N_o difference for 0.1% information BER between complex suppression filter and real suppression filter is around 0.3 dB.

6.1 SUPPRESSION FILTER WITH SEVERAL LENGTHS

Complex transversal suppression filter updated at every 20 msec by Levinson Durbin algorithm is investigated on AWGN channel for filter length 9, 17, 31 with 20 dB of J/S and $T_c \times f_d = 0$. Figure 14 shows that E_b/N_o to get 0.1% information BER was 0.7, 0. 36, 0.4 dB higher for filter length 9, 17, 31 than no jammer and no suppression case. The best performance was obtained when filter of length 9 was used. This means thap upon determining the estimation of present jammer, the received signal over 17 samples is statistically independent and so they acts as noises. The determining of optimum filter length depends on jammer characteristics.

6.2 SUPPRESSION OF MULTIPLE NUMBER OF JAMMER

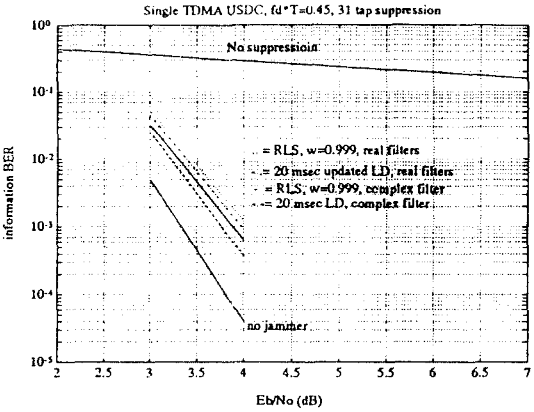

Fig 12.

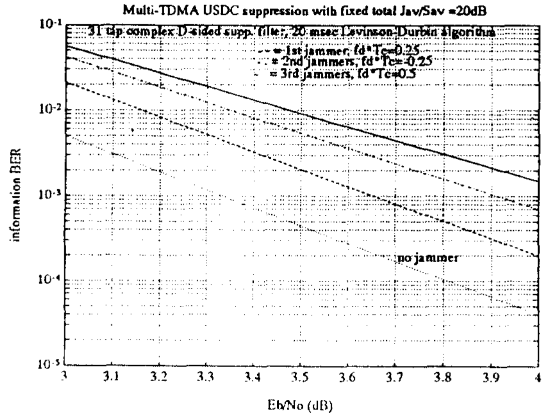

Fig 15.

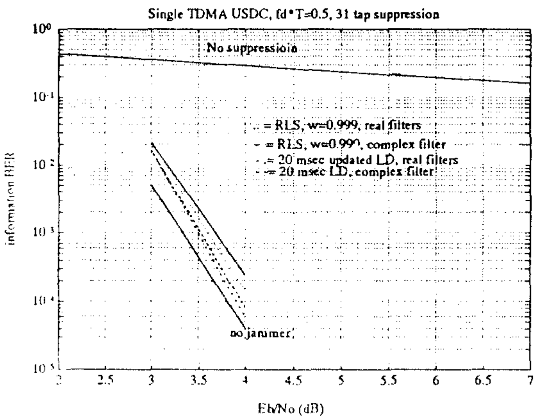

Fig 13.

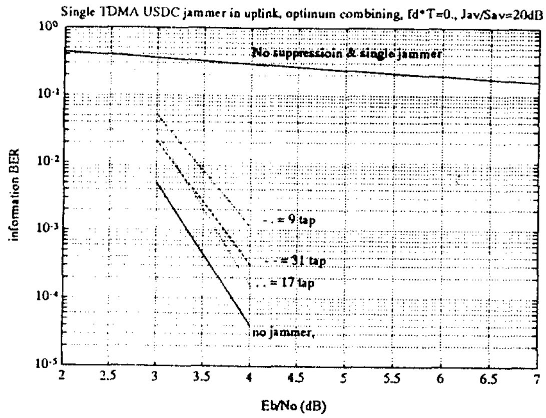

Fig 16.

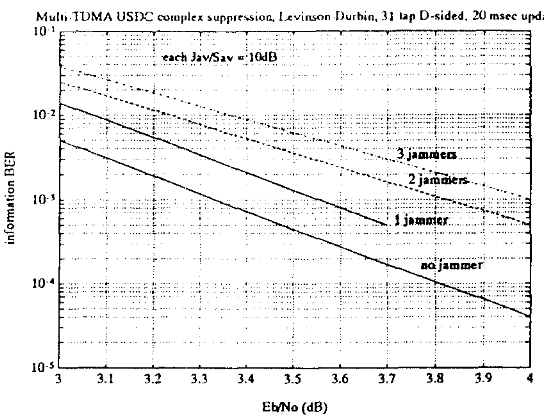

Fig 14.

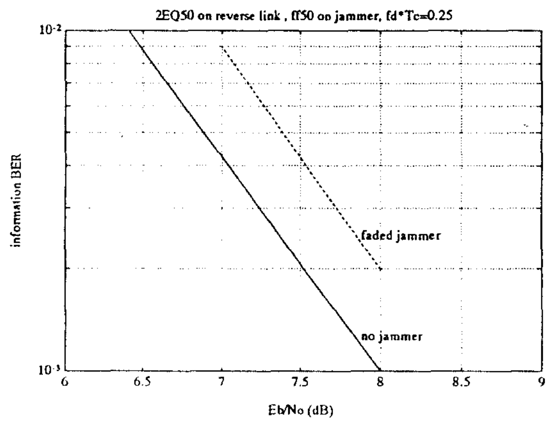

Fig 17.

Complex suppression filter updated every 20 msec by Levinson Durbin algorithm for suppressing multiple number of TDMA USDC jammer was investigated. Total jammer power to desired signal power ratio was fixed to be 20 dB. Figure 15 shows that E_b/N_o to get 0.1% information BER degrades 0.2 dB per jammer. Figure 16 shows the jammer suppression filter performance for multiple jammers, in which each jammer to desired signal power ratio is 10 dB and complex 31 tap double sided symmetric filter with Levinson Durbin algorithm of 20 msec update rate is used and channel is assumed to be AWGN. Additional E_b/N_o degradatuib was 0.2 dB per jammer.

6.3 FADING ON JAMMERS

The performance of jammer suppression filter at the base staion was investigated for marrowband jammer faded on frequency flat fading channel. The total jammer power to desired CDMA power ratio is set to 20 dB and suppression filter is 31 tap double sided symmetric. Figure 8 shows that for single TDMA USDC jammer on flat fading 50km/h, E_b/N_o for 0.1% BER was 0.6 dB worse than no jammer case and 0.4 dB worse than unfaded jammer with suppression filter case For double TDMA USDC jammers on independent flat fading 50km/h channel, E_b/N_o for 0.1% BER was 0.2 dB worse than single jammer case, which corresponds to E_b/N_o difference between single and double unfaded TDMA USDC jammers case. In Figure 17, the channel from mobile to base is 2 equal ray model, in which each path has equal power and the delay between two paths is 25 chips. The channel between jammer and base station is frequency flat fading with 50km/h velocity. It is shown that E_b/N_o for 0.1% BER was 0.5 dB worse than no jammer case. Suppression filter is though to be better performed with power control on the lower vehicle speed.

The performance degradation can be thought as loss of channel capacity. For simplicity, the lost channel capacity can be computed for single cell case. For single cell case, the relation between the number of users and E_b/N_o can be described as follows,

$$N-1=\frac{W}{R}\frac{E_b}{N_0} \tag{25}$$

where N is the number of users, W is chip rate, and R is information data rate. 0.5 dB E_b/N_o degradation reduces the number of users, say N', then N' can be computed as follows,

$$N'-1=\frac{W}{R}\frac{E_b}{N_0}\frac{1}{1.122} \tag{26}$$

Using Eq.(25) and (26), we can easily find the relation between N and N', i.e., $N'=0.9N$. Therefore 10% channel capacity corresponding to 0.5 dB E_b/N_o degradation is expected due to jammer suppression.

Ⅶ. CONCLUSION

The effects of inband TDMA USDC narrowband jammers to DS-CDMA system(TDMA overlaid by CDMA) performance and the suppression techniques are investigated using Monte Carlo simulations. Levinson Durbin and conventional recursive least square algorithm are emphasized and two filter structures, i.e., complex suppression filter and real suppression filter on each inphase and quadrature channels, were investigated and their performance was compared. Complex suppression filter with Levinson Durbin algorithm of 20 msec update rate is the most promising with respect to implementation and performance point of view. 31 tap double sided suppression filter is used throughout the simulation. However 17 tap suppression filter is enough long for TDMA USDC jammer. Continuous suppression strategy is much better than tracking on-off charateristics of jammer for high power TDMA jammer. The suppression filter shows outstanding performance improvement and 0.5 dB degradation compared to no jammer and no filter case. As the number of jammer increases with fixed total jammer power, the additional degradation is 0.2 dB per jammer for non-fading and frequency flat fading with 50km/h vehicle speed. Complex double sided suppression filter of length 17 is implementable with four ICs and two DSPs.

References

1. L. B. Milstein, "Interference rejection techniques in spread specturm communications", *Proceedings of the IEEE*, vol. 76, no. 6, pp. 657-571, June 1988.
2. J. Wang and L. B. Milstein, "Application of suppression filters for CDMA overlay situation", *ICC*, pp. 310, June 1992.
3. H. Vincent Poor and Leslie A. Rusch, "Narrowband Interference Suppression in CDMA Spread Spectum Communications", *IEEE Trans. on Commun.*, vol. 42, no. 4, pp. 1969-1979, April 1994.
4. John G. Proakis, *Advanced Digital Signal Processing*, McGraw-Hill, 1989.
6. 준마이크로파대(1~3GHz) CDMA 전파특성에 관한 연구, 한국이동통신주식회사 중앙연구소 최종보고서, 1994.
7. CDMA 이동통신용 순방향 역방향 모뎀 시뮬레이터 제작, 한국전자통신연구소 최종보고서, 1994.

김 동 구(Dong Ku Kim) 정회원
1960년 4월 20일생
1983년 2월 : 한국항공대학교 통신공학과 졸업(공학사)
1985년 5월 : University of Southern California 전기공학과 졸업(공학석사)
1992년 5월 : Univeristy of Southern California 전기공학과 졸업(공학박사)
1992년 5월~1992년 9월 : 미국 University of Southern California Post Doc.(연구원)
1992년 9월~1994년 2월 : 미국 Motorola cellular infrastructure group 연구소 연구원
1994년 3월~현재 : 연세대학교 공과대학 전파공학과 조교수

박 형 일(Hyung-Ill Park) 정회원
1971년 12월 7일
1994년 2월 : 연세대학교 전자공학과 졸업
1996년 2월 : 연세대학교 전자공학과 대학원 졸업
1996년~현재 : 삼성전자 근무