NORTH ATLANTIC TREATY ORGANIZATION

ORGANISATION DU TRAITE DE L'ATLANTIQUE NORD

MILITARY AGENCY FOR STANDARDIZATION (MAS)
BUREAU MILITAIRE DE STANDARDISATION (BMS)

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16 February 1989

To : See MAS Distribution List No.2

Subject : STANAG 4285 EL (EDITION 1) - CHARACTERISTICS OF 1200/2400/3600 BITS PER SECOND SINGLE TONE MODULATORS/DEMODULATORS FOR HF RADIO LINKS

Reference : AC/302-D/439 dated 1 July 1987

Enclosure : STANAG 4285 (Edition 1)

1. The enclosed NATO Standardization Agreement which has been ratified by nations as reflected in page iii is promulgated herewith.

2. The reference listed above is to be destroyed in accordance with local document destruction procedures.

3. AAP-4 should be amended to reflect the latest status of the STANAG.

ACTION BY NATIONAL STAFFS

4. National staffs are requested to examine page iii of the STANAG and if they have not already done so, to advise the Defence Support Division, IS, through their national delegation as appropriate of their intention regarding its ratification and implementation.

A.J. MELO CORREIA
Major-General, POAF
Chairman, MAS

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STANAG 4285
(Edition 1)

NORTH ATLANTIC TREATY ORGANIZATION

(N A T O)

MILITARY AGENCY FOR STANDARDIZATION

(M A S)

STANDARDIZATION AGREEMENT

SUBJECT : CHARACTERISTICS OF 1200/2400/3600 BITS PER SECOND SINGLE TONE MODULATORS/DEMODULATORS FOR HF RADIO LINKS

Promulgated on 16 February 1989

A.J. MELO CORREIA
Major-General, POAF
Chairman, MAS

[redacted]

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RECORD OF AMENDMENTS

No.	Reference/date of amendment	Date entered	Signature
1-2		10/07/01	n. R

EXPLANATORY NOTES

AGREEMENT

1. This NATO Standardization Agreement (STANAG) is promulgated by the Chairman MAS under the authority vested in him by the NATO Military Committee.

2. No departure may be made from the agreement without consultation with the tasking authority. Nations may propose changes at any time to the tasking authority where they will be processed in the same manner as the original agreement.

3. Ratifying nations have agreed that national orders, manuals and instructions implementing this STANAG will include a reference to the STANAG number for purposes of identification.

DEFINITIONS

4. Ratification is "The declaration by which a nation formally accepts the content of this Standardization Agreement".

5. Implementation is "The fulfilment by a nation of its obligations under this Standardization Agreement".

6. Reservation is "The stated qualification by a nation which describes that part of this Standardization Agreement which it cannot implement or can implement only with limitations".

RATIFICATION, IMPLEMENTATION AND RESERVATIONS

7. Page iii gives the details of ratification and implementation of this agreement. If no details are shown it signifies that the nation has not yet notified the tasking authority of its intentions. Page iv (and subsequent) gives details of reservations and proprietary rights that have been stated.

NAVY/ARMY/AIR

NATO STANDARDIZATION AGREEMENT
(STANAG)

CHARACTERISTICS OF 1200/2400/3600 BITS PER SECOND SINGLE TONE MODULATORS/DEMODULATORS FOR HF RADIO LINKS

ANNEXES:

A. Required Characteristics of 1200/2400/3600 Bits per Second Single Tone Modulators/Demodulators for HF Radio Links
B. Evaluation of Modems Employing the STANAG 4285 Waveform (for information only)
C. First Example of Demodulation Technique (for information only)
D. Second Example of Demodulation Technique (for information only)
E. Error Correction Coding, Interleaving and Message Protocols for use with the Standard Modulation Formats (for information only)
F. Use of the Synchronization Sequence for Signal Detection and Acquisition/Tracking of Doppler, Synchronization and Channel Parameters (for information only)

RELATED DOCUMENTS:

STANAG 4203 Technical Standards for Single Channel HF Radio Equipment

AIM

1. The aim of this agreement is to define the technical interoperability characteristics for communicating over digital data radio channels, the effective bit rate of which can be 1200, 2400 or 3600 bits per second, by means of a single tone modulator/demodulator.

AGREEMENT

2. The participating nations agree to use the characteristics contained in this STANAG (Annex A and Appendix 1 to Annex A) for their 1200/2400/3600 bits per second single tone modems for digital data communications over HF radio equipment

IMPLEMENTATION OF THE AGREEMENT

3. This STANAG is implemented by a nation when single tone modulators/demodulators, for transmission of 1200/2400/3600 bps data over HF radio links, in that nation's forces comply with the characteristics detailed in this agreement and are placed in service.

REQUIRED CHARACTERISTICS OF 1200/2400/3600 BITS PER SECOND SINGLE TONE MODULATORS/DEMODULATORS FOR HF RADIO LINKS

Appendix 1: Required Characteristics
Appendices 2 and 3: Information Only

INTRODUCTION

1. This document describes the modulation and call establishment processes required to ensure interoperability between modems transmitting data over HF radio links where the input interface rates may be 1200, 2400 or 3600 bps.

MODULATION

2. (a) the modulation technique consists of phase shifting of a sub-carrier of 1800 Hz. Modulation speed is 2400 bauds with a minimum accuracy of 1 part in 10^5.

   (b) the accuracy of the clock linked with generation of the 1800 Hz is 1 part in 10^5.

   (c) the phase shift of the modulated signal relative to the unmodulated reference sub-carrier may take one of the following values (see Figure A-1).

Symbol Number	Phase
0	0
1	$\pi/4$
2	$\pi/2$
3	$3\pi/4$
4	π
5	$5\pi/4$
6	$3\pi/2$
7	$7\pi/4$

The complex number $e^{jn\pi/4}$ is linked with the symbol number n.

TRANSCODING

3. Transcoding is an operation involving linking a symbol to be transmitted to a group of bits:

(a) 1200 bps Data Rate

transcoding is achieved by linking one symbol to one bit according to the following rule:

Bit	Symbol
0	0
1	4

(b) 2400 bps Data Rate

transcoding is achieved by linking one symbol to a set of two consecutive bits according to the following rule:

Dibit	Symbol
0 0	0
0 1	2
1 0	6
1 1	4

Oldest bit Most recent bit

(c) 3600 bps Data Rate

transcoding is achieved by linking a symbol to a set of three consecutive bits according to the following rule:

Tribit	Symbol
0 0 0	1
0 0 1	0
0 1 0	2
0 1 1	3
1 0 0	6
1 0 1	7
1 1 0	5
1 1 1	4

Oldest bit Most recent bit

POWER SPECTRAL DENSITY

4. The power spectral density of the modulated signal after filtering and transposition to 1800 Hz shall be as shown in Figure A-2 (see Appendix 3). The total bandwidth occupied is 3000 Hz.

FRAME STRUCTURE

5. (a) the frame structure is shown in Figure A-3;

(b) the symbols to be transmitted are structured in recurrent frames 106.6 ms in length. The number of bits transmitted per frame is 128 at 1200 bps, 256 at 2400 bps and 384 at 3600 bps;

(c) a frame consists of 256 symbols. A frame can be broken down into; 80 symbols for synchronization, 48 reference symbols and 128 data symbols;

(d) these 176 reference and data symbols are scrambled by a scrambling sequence with eight phase states of length 176 (see Annex A, paragraph 7);

(e) the modem makes use of the synchronization sequence to detect the presence of the signal and for correction of the frequency shift resulting from the doppler effect or the difference between the transmit and receive pilots, bit synchronization and either equalizer training in the case of equalization by recursive filtering or HF channel estimation in the case of detection according to the maximum likelihood method;

(f) the reference and data symbols are formed into 4 blocks: the first 3 consist of 32 data symbols followed by 16 reference symbols; the last block consists of 32 data symbols. The reference symbols are all symbol number 0.

SYNCHRONIZATION SEQUENCE GENERATOR

6. (a) the synchronization consists of 80 symbols and is transmitted recurrently every 106.6 ms. This sequence uses 2PSK modulation and the modulation rate is equal to 2400 bauds;

(b) the sequence is identical to a pseudo random sequence of length 31, which is repeated periodically within the 80 symbol window, i.e. the synchronization sequence consists of 2 periods of length 31 plus the first 18 symbols of another period;

(c) a generator for the synchronization sequence is described in Figure A-4. The generator polynomial is:

$$X^5 + X^2 + 1$$

(d) at the beginning of every frame the generator is initially set to the following value:

1 1 0 1 0

(e) the first symbol of the synchronization sequence is identical to the least significant bit of this initial value. The remaining 79 symbols are obtained by applying the clock 79 times.

GENERATION OF THE DATA BLOCK SCRAMBLING SEQUENCE

7. (a) the scrambling sequence is composed of 176 symbols and is repeated every 106.6 ms. This sequence is transmitted in eight-phase-state modulation at the rate of 2400 bauds;

(b) data scrambling by an eight-phase-state sequence makes it possible to create an eight-phase-state frame, whatever the data rate may be (1200, 2400 or 3600 bps);

(c) the scrambling symbol generator is shown in Figure A-5. The symbols are formed by means of a pseudo-random code of length 511, the generator polynomial of which is:

$$X^9 + X^4 + 1$$

(d) the generator is initialized to 1 at the start of each frame;

(e) a symbol is derived from the triplet consisting of the last three bits in the PN register, i.e. $X^0\ X^1\ X^2$ by the following relationship:

Scrambling symbol $B_k = e^{jn\pi/4}$

where:

$n = 4.X^2 + 2.X^1 + X^0$

$X^0 = 0$ or 1

$X^1 = 0$ or 1

$X^2 = 0$ or 1

(f) generation from one symbol to the next is by successive shifting of the PN register by three positions.

DATA BLOCK SCRAMBLING

8. The scrambling operation is carried out on reference and data symbols only, not on the synchronization sequence. This operation consists of modulo 8 addition of the data symbol number to the scrambling symbol number; this amounts to complex multiplication of the data symbol by the scrambling symbol.

TOLERANCE OF ERRORS IN FREQUENCY BETWEEN HF TRANSMISSION AND RECEPTION CARRIERS

9. The modem must be capable of tolerating a frequency error of of $\pm$ 75 Hz between the transmission and reception HF carriers (transmitter/receiver frequency error and Doppler shift included) and rate of frequency change up to 3.5 Hz.

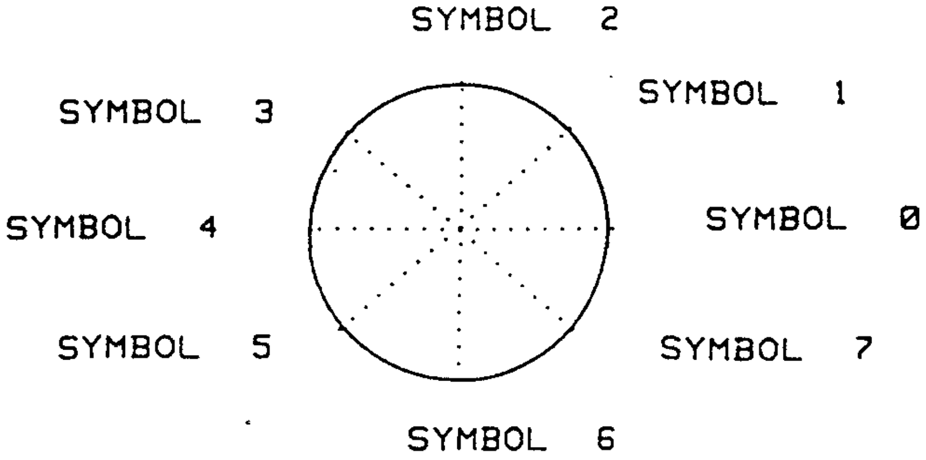


FIGURE A-1: PHASE STATE ENCODING

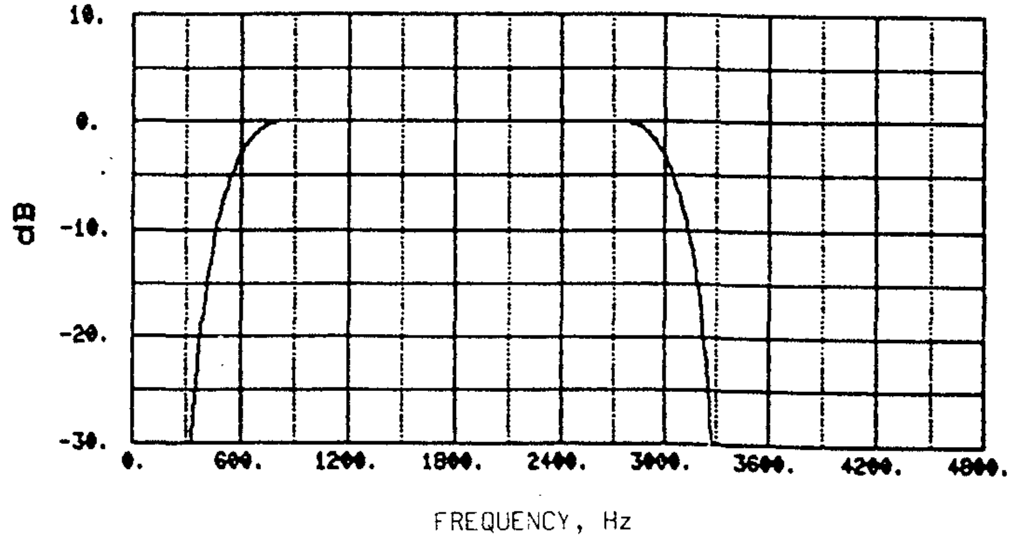
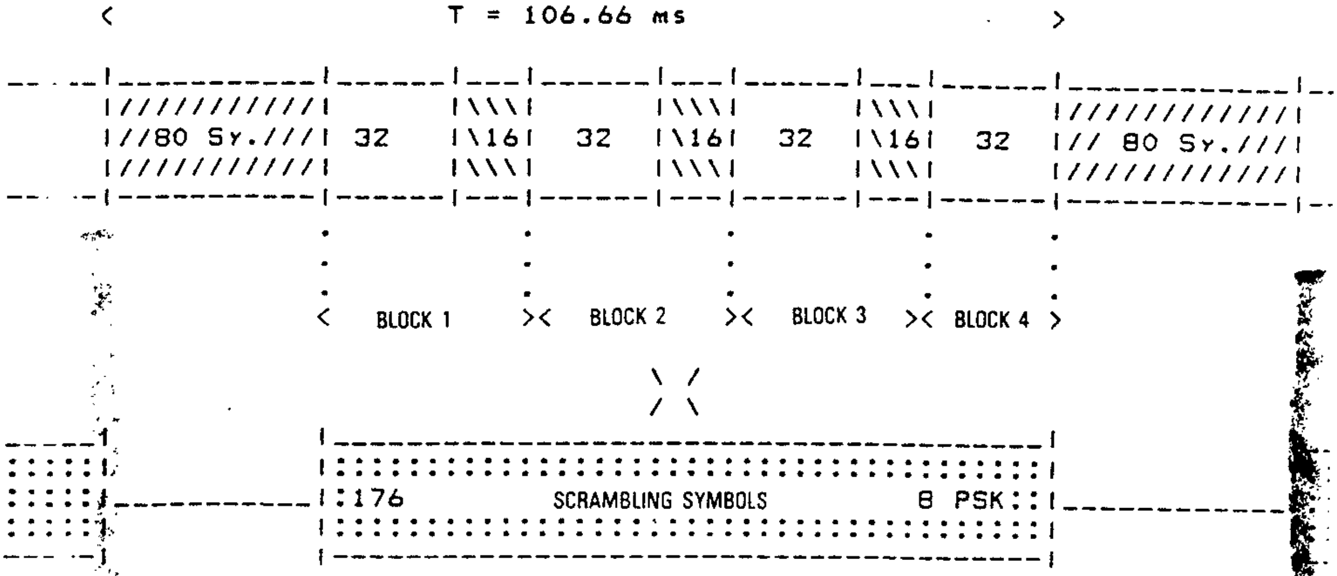


FIGURE A-2: POWER SPECTRAL DENSITY OF MODULATOR OUTPUT SIGNAL



SYNCHRONISATION SYMBOLS

DATA SYMBOLS:

1200 b/s ....> 2 PSK

2400 b/s ....> 4 PSK

3600 b/s ....> 8 PSK

REFERENCE SYMBOLS: FOLLOW-UP OF CHANNEL DEVELOPMENT
(equal to symbol N = 0)

FIGURE A-3: FRAME STRUCTURE

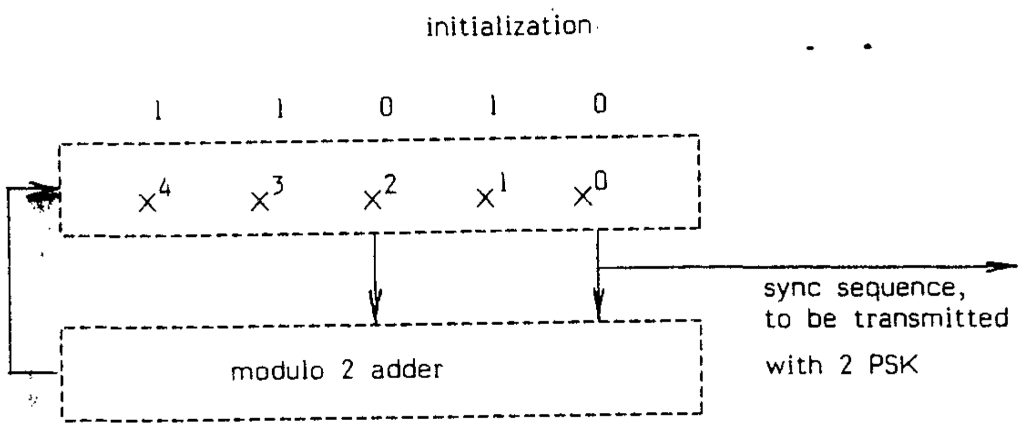


FIGURE A-4: SYNCHRONIZATION SEQUENCE GENERATOR

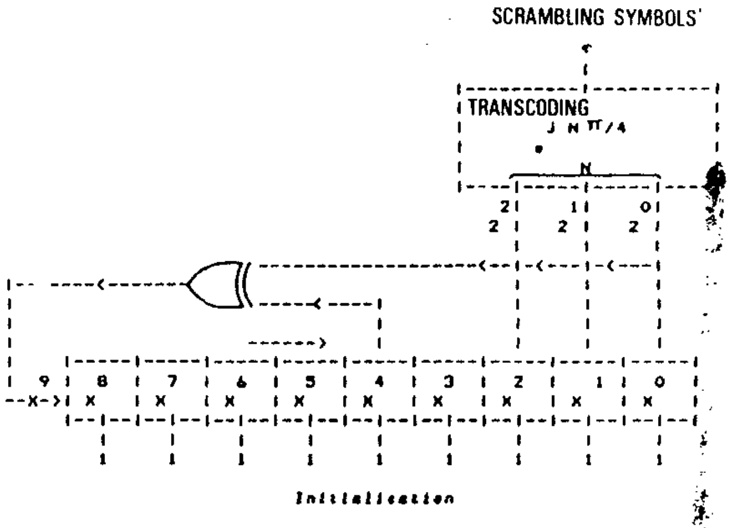


FIGURE A-5: GENERATION OF SCRAMBLING SEQUENCE

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PERFORMANCE OF ASSOCIATED COMMUNICATIONS EQUIPMENT

1. To obtain optimal performance, the following characteristics for transmitters and receivers are required:

(a) they must have a bandpass such that, between 300 and 3,300 Hz, variations in attenuation are at most $\pm$ 2 dB;

(b) over 80% of the passband, the group delay time must not vary by more than 0.5ms;

(c) the accuracy of the transmitter and receiver pilots must be at least 10^{-6}. (The modem is capable of tolerating an offset of $\pm$ 75 Hz, transmitter/receiver frequency variation and Doppler shift included.);

(d) the time constant of the AGC circuit must be less than 10 ms on desensitization and less than 25 ms on resensitization.

MODEM DIALOGUES AND INTERFACES WITH TRANSMITTER, RECEIVER AND DATA TERMINALS
(For Information Only)

MODEM/DATA TERMINAL INTERFACES

1. (a) the modem interface with the data terminals meets the requirements of CCITT Recommendation V24 and interface electrical characteristics conform with Recommendation VII (RS 422);

(b) the wires used on these interfaces are at least as follows:

signal ground or common return (102)
transmit data (103)
received data (104)
request to send (105)
ready for sending (106)
data channel received line detector (109)
transmitter signal element timing (DCE) (114): minimum accuracy 10^{-5}
receiver signal element timing (DCE) (115)
data signal quality detector (110)
frame transmission (supplied by the modulator): 1 bit (not part of V24)
frame reception (supplied by the demodulator): 1 bit (not part of V24)

MODEM/TRANSMITTER AND MODEM/RECEIVER INTERFACES

2. (a) Interface with Transmitter (Audio Frequency). The modulated signal is provided on a balanced output at an impedance of 600 Ω and a level 0 dBM. The modem controls setting of the transmitter to transmission by means of a switching contact, the resistance of which must be less than 300 Ω to control transmission and greater than 30 K Ω to change to reception (if a transmitter-receiver is involved) or to non-transmission;

(b) Interface with Receiver (Audio Frequency). The modem input circuit is balanced with respect to ground and offers an impedance of 600 Ω for a nominal signal level of 0 dBM.

DATA TERMINAL/MODEM DIALOGUE

3. (a) communication is initialized by reception of a request to transmit from the data terminal; the modem replies by closing the switching contact controlling changeover to transmission

after a time equal to T1, which is a function of the transmitter used. The modem replies "ready for sending" and sends the clock (1200, 2400 or 3600 Hz) over wire 114; the synchronization sequence is transmitted simultaneously, while the data are stored in order to form blocks;

(b) provision of the clock is interrupted by changeover to the low level of the "request to send" signal, the transmitted signal being interrupted when the data buffer memory is empty;

(c) the receiving modem carries out a continuous search for the synchronization sequence. When this has been recognized, channel equalization is performed and the data are restored at a rate of 1200, 2400 or 3600 bps with a maximum delay of 212 ms (excluding the propagation time due to the radio channel and to the transit time in the transmitter-receivers);

(d) the data are no longer provided after non-recognition of one or more synchronization codes, which are transmitted periodically during transmission of data;

(e) the "frame transmission" signal is a 106.6 ms period clock. This signal is in the open state during 127, 255 or 383 bit periods and changes to the closed state during one bit period according to the data rate: 1200, 2400 or 3600 bps. The data item in wire 103 when this signal is in the closed state is the first item in the transmitted frame;

(f) the "frame reception" signal is a 106.6 ms period clock. This signal is in the open state during 127, 255 or 383 bit periods and changes to the closed state during one bit period according to the data rate; 1200, 2400 or 3600 bps. The data item in wire 104 when this signal is in the closed state is the first item in the received frame. When wire 109 is in the open state, "frame reception" signal remains in the open state;

(g) the exchange diagram is given in Figure A-2-1.

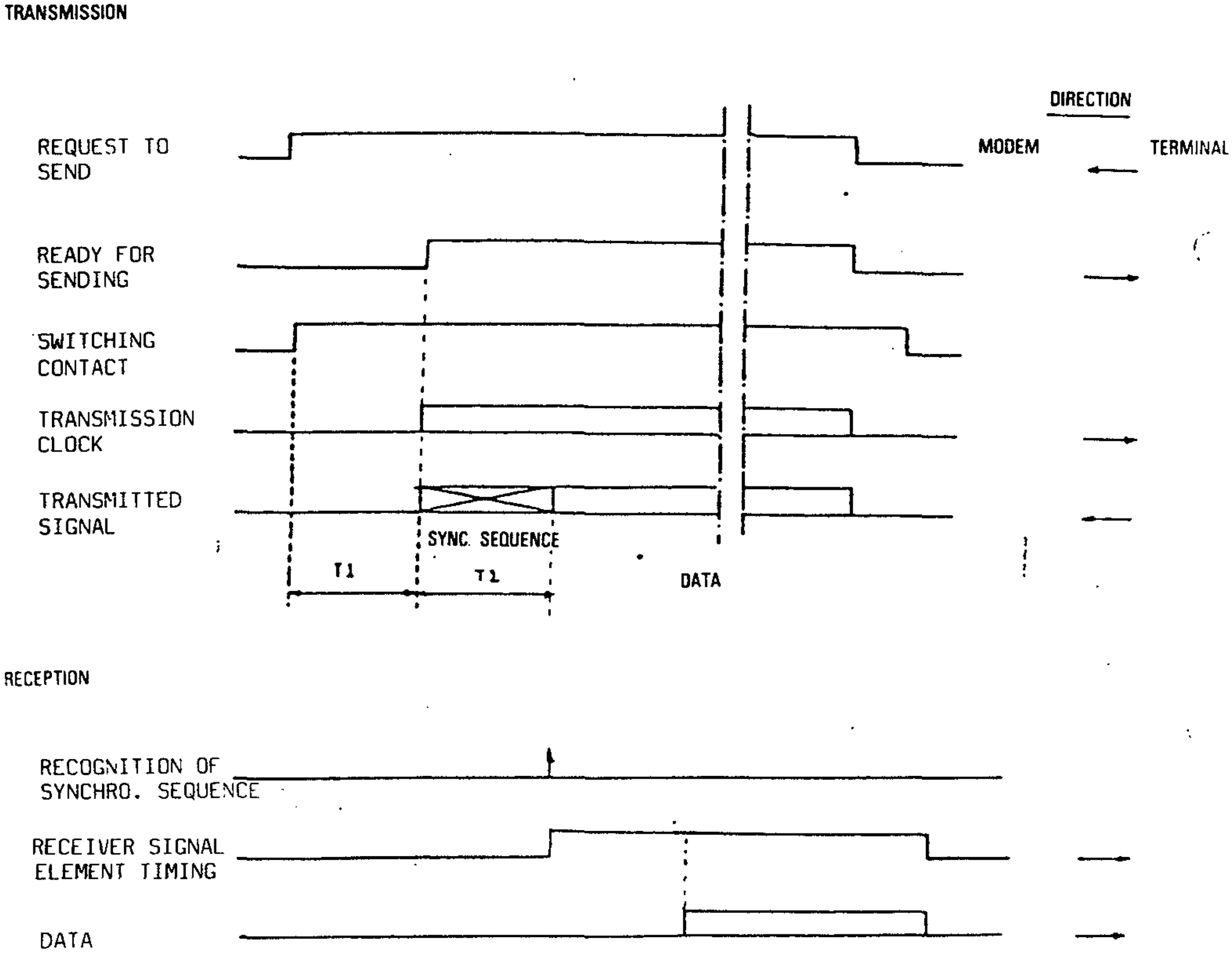


FIGURE A-2-1: DIALOGUE EXCHANGE DIAGRAM

MODULATOR/DEMODULATOR FILTERING
(For Information Only)

1. The forming filters used by the modulator and the demodulator must test the following criteria:

(a) compatibility with 300 to 3300 Hz band pass transmitter/receivers;

(b) reception filter matched to transmission filter (to maximize the signal-to-noise ratio in the demodulator);

(c) putting the transmission and reception filters in series should form a filter minimizing inter-symbol modulation in the demodulator. This criterion is provided by a raised cosine filter, the impulse response of which cancels out all multiples of the bauds T = 1/2400s period. The transfer function of such a filter is as follows:

$$H(f) = 1 \qquad f \leq fn - pfn$$

$$H(f) = 0.5\left(1 - \sin\frac{(f-fn)}{pfn}\frac{\pi}{2}\right) \qquad fn - pfn \leq f \leq fn + pfn$$

$$H(f) = 0 \qquad f \geq fn + pfn$$

where

fn is the Nyquist frequency ($fn = 1/2\,T$);

p is the roll-off factor.

The above criteria are realized by taking:

- transmission filter (f) = reception filter (f) = $\sqrt{H(f)}$;
- roll-off = 0.2.

Such filters are synthesized by Finite Impulse Response (FIR) filters with a sampling rate of 4/T. An example of filtering obtained with a transmission filter with 40 coefficients is given in Figure A-3-1.

[redacted]

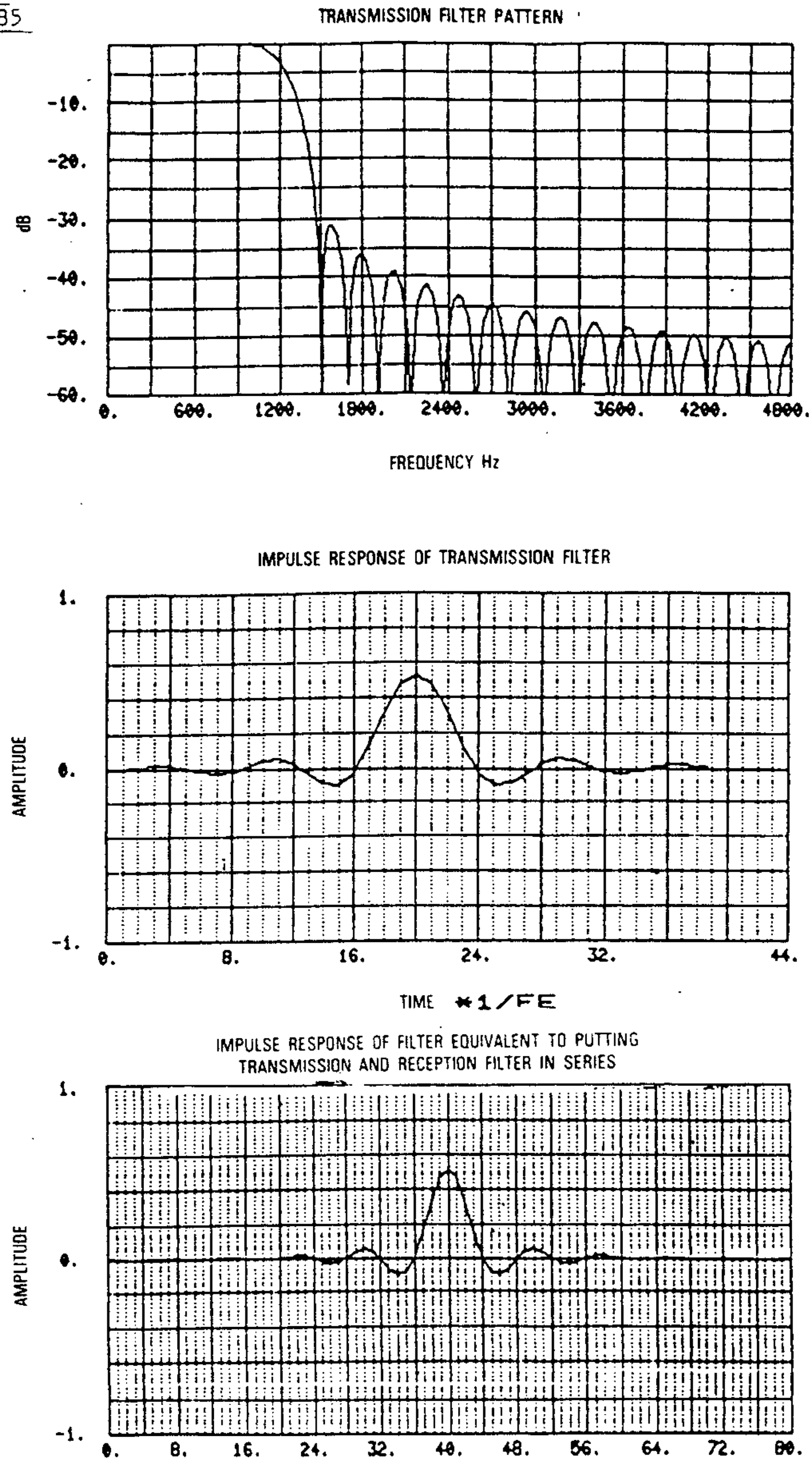


FIGURE A-3-1: TRANSMISSION/RECEPTION FILTERING

[redacted]

EVALUATION OF MODEMS EMPLOYING THE STANAG 4285 WAVEFORM
(For Information Only)

PURPOSE

1. Since the STANAG 4285 waveform is suitable for reception using a variety of different receive algorithms, it is possible to achieve interoperability with a modem with very different link performance characteristics. This section is intended to provide guidance in characterizing the operational performance of a modem with STANAG 4285 capability in a manner that will allow better comparison of modem characterizations that may have been performed at different places.

MATHEMATICAL DEFINITION OF HF CHANNEL SIMULATOR

2. (a) in order to provide comparable simulation results between two modems, it is necessary that the statistical behaviour of the HF channel provided by the HF channel simulator be the same. The model shown in Figure B-1 is one developed by the National Telecommunication Information Agency(1). The input is assumed to be complex obtained by either a Hilbert transform of a baseband signal from the modem or a quadrature demodulation from a higher frequency version of the modem signal (such as that in an exciter IF);

(b) four forms of channel disturbances are included in the simulator. Multi-path is produced by providing selectable delay for several different versions of the transmit signal. For the recommended tests, only one delay line is required, since none of the tests involve more than two paths and one path may have no delay. The "multipath spread" in this case is equal to the difference in delay of the two paths;

(c) a second type of disturbance is amplitude and phase variation which is provided by mutliplying each of the multiple paths by a complex signal obtained as shown in Figure B-2. Each of the paths has an independent complex multiplying factor so that, for two paths, four independent Gaussian noise sources are required. It should be noted that the low pass filters which determine the Doppler spread have a bandwidth of approximately one-half the Doppler spread due to the factor of two in the equation defining Doppler spread (as supplied by Watterson). The shape of the low pass filter is not critical but it should be at least a two-pole filter;

(d) the other two disturbances are additive in nature. The first is additive white Gaussian noise and the second is an inter-

(1) See reference document listed on Page B-8.

ference simulator. The Gaussian noise source should be spectrally flat over the bandwidth required for the modem signal. The interference generator should have the capability of generating the four types of intereference signals specified in tests 8 through 11 detailed in paragraph 4 of this Annex.

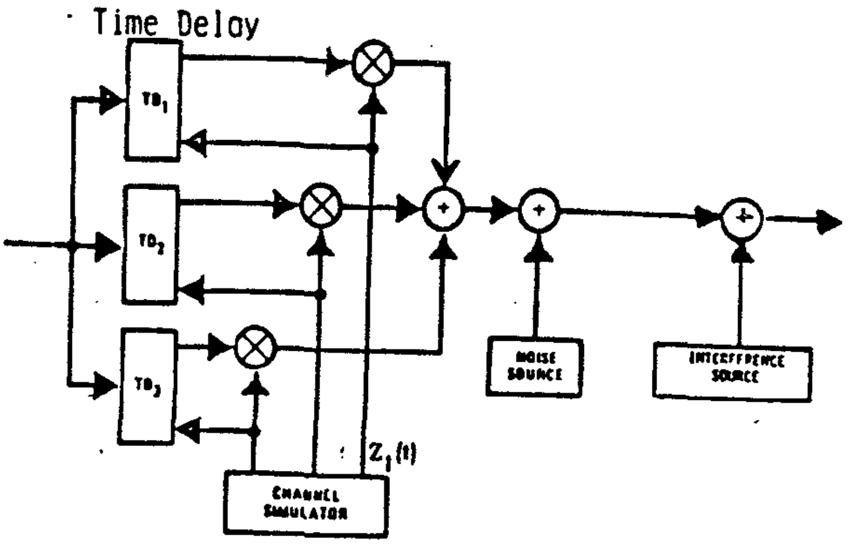
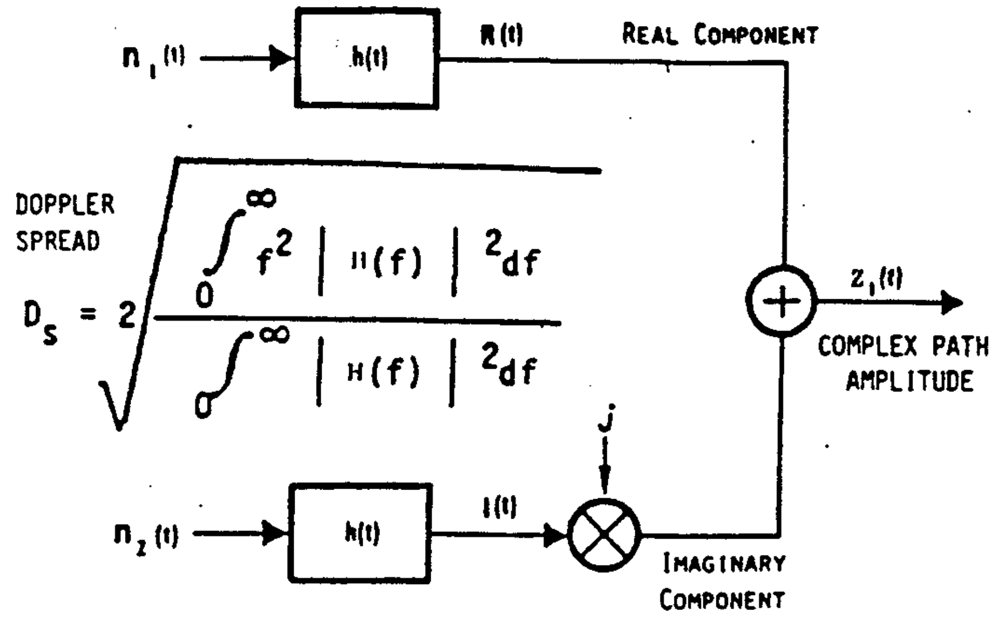


Figure B-1. HF Link Stimulation - Simulation of Fading Multi-Path HF Channel



$n_1(t)$ and $n_2(t)$ are INDEPENDENT WHITE GAUSSIAN NOISE SOURCES.

h(t) IS A LOW PASS FILTER.

Figure B-2. HF Link Stimulation - Generation of a Rayleigh Fading Path

3. Two cautions should be given to the simulator design. The first is that the output of the low pass filters providing the multiplicative factors for the paths must be sampled (if a sampled system is used) at a much higher rate than twice their bandwidth, since the phase variation implied by the complex signal can change much faster than the waveforms of each individual component. It is recommended that the sampling rate be at least 32 times the Doppler spread. The second point concerns simulation of exciter and receiver IF filtering and receiver AGC. Where possible, simulation or use of these system elements is recommended and the test results should specify the types of filters and AGC involved in the simulation. If the simulator does not provide the capability of such simulation, and the

modem does not have a dynamic AGC at its input (the modem can normally depend upon the receiver for AGC) then a separate baseband AGC should be provided to prevent degradation in modem performance due to fades that are not representative of that which will occur when the modem is operated with a receiver.

TEST CONFIGURATIONS

4. In this paragraph, 11 test configurations are specified. The selection of this set represents a compromise between modem characterization and time to complete the tests. The tests involve characterization relative to all four of the types of disturbances discussed in paragraph 2 of this Annex:

(a) Characterization Versus Signal-to-Noise Ratio. Five different signal-to-noise tests are recommended. The first two tests apply to line-of-sight and groundwave operation and the remaining three address skywave operation. The simulator parameters for these tests are given in Table B.1. The last two tests are recommended in CCIR recommendation 520(1). The CCIR "good" channel described in the CCIR recommendation has been omitted since the time required to obtain a representative channel is too large;

Table B-1: Bit Error Rate Versus Signal-to-Noise Ratio for the Following Conditions

TEST #	DOPPLER SPREAD (0 = NONFADING) PATH 1	PATH 2	POWER OF PATH 2 RELATIVE TO PATH 1	DELAY SPREAD MS	TEST NAME
1	0	-	Path 2 Off	-	Gaussian Noise
2	0	1 Hz	6 dB less	0	Ricean Fading
3	1.0 Hz	-	Path 2 Off	-	Flat Fading
4	0.5 Hz	0.5 Hz	Equal	1.0 ms	CCIR Moderate
5	1.0 Hz	1.0 Hz	Equal	2.0 ms	CCIR Poor

(1) See reference document listed on Page B-8.

(b) Characterization Versus Doppler Spread. Test 6 is a test of a bit-error rate versus Doppler spead (0 to 8 Hz) for two independent equal average power fading paths. The multipath spread is 1.0 ms and no additive noise is included;

(c) Characterization of Multipath Spread Performance. Test 7 is a test of bit-error rate versus multipath spread (0 to 8 ms) for two independent equal average power fading paths. Each path has 0.5 Hz Doppler spread and no additive noise;

(d) Characterization of Interference Performance:

Test 8 — CW
Measure bit-error rate versus signal-to-interference ratio.
Interference frequency at centre of signal spectrum.

Test 9 — Impulse CW
Measure bit-error rate versus duty cycle.
Interference frequency at centre of signal spectrum.
Pulsewidth equal to symbol duration.
Pulse amplitude equal to: (a) Signal Amplitude
(b) Signal Amplitude -6 dB

Test 10 — Swept CW
Measure bit-error rate versus signal-to-interference ratio.
Sweep from $fc - fb/_2$ to $fc + fb/_2$.
where fc is centre frequency of signal, fb is symbol rate.
Sweep time (a) 0.5 sec.
(b) 20 sec.

Test 11 — FSK
Measure bit-error rate versus signal-to-interference ratio.
FSK shift 850 Hz. Tone frequencies fc $\pm$425 Hz.
Keyed at 75 bps, 511 bit pseudorandom data sequence.

TEST TIME

5. (a) one of the difficulties in comparing HF modem performance on noisy fading channels is assuming that the channel characteristics for the modems during the test time are close enough from a statistical standpoint to assure that the observed performance difference is due to the modem and not

due to the fact that one modem observed easier conditions than the other. Although HF simulators tend to equalize conditions, it is necessary to run for extended lengths of time on each bit-error measurement to assure that the conditions were indeed the same for each modem. The length of time required per measurement value is dependent upon the Doppler spread of the channel, the number of paths, the modem characteristics, and the bit-error rate resulting from the test. Table B-2 lists the approximate test time to obtain observed bit-error rate values that have a 90 percent confidence level of being within a factor of two of the actual value. This table applies to tests for implementations without coding and interleaving. If the coding/interleaving recommendations (Annex E) are carried out, the recommended test time changes with interleaver delay and are listed in Tables B-3 and B-4;

(b) the test times required for the remaining tests are not as large since the S/N ratio from the channel simulator is not produced by the fading mechanism. In these cases the test time can be determined as the time required to produce a specified number of errors. To produce approximately the same level of confidence as that stated for tests 3, 4 and 5, ten independent error events or more are required.

Table B-2: Test Point Times - No Coding

Test	10^{-2}	10^{-3}	10^{-4}	10^{-5} (Actual Bit Error Rate)
3	2.65 min	27.2 min	4.55 hrs	45.5 hrs
4	2.72 min	28.4 min	4.61 hrs	45.7 hrs
5	1.36 min	14.2 min	2.30 hrs	22.8 hrs

Table B-3: Test Point Times - Coding & Short Interleaver

Test	10^{-2}	10^{-3}	10^{-4}	10^{-5} (Actual Bit Error Rate)
3	56.8 sec	8.19 min	1.24 hrs	11.8 hrs
4	1.35 min	10.4 min	1.44 hrs	13.0 hrs
5	40.5 sec	5.2 min	43.2 min	6.5 hrs

Table B-4: Test Point Times - Coding & Long Interleaver

Test	10^{-2}	10^{-3}	10^{-4}	10^{-5} (Actual Bit Error Rate)
3	4.12 min	26.4 min	3.06 hrs	23.8 hrs
4	3.41 min	20.3 min	2.16 hrs	15.6 hrs
5	3.41 min	20.3 min	2.16 hrs	15.6 hrs

[redacted]

REFERENCES

1. Watterson, C.C., Juposher, J.R., and Bensema, W.D. [1969b], Experimental Verification of an Ionospheric Channel Mode, ESSA Tech. Rept. ERL 112-ITS (US Government Printing Office).

2. CCIR Recommendation 520, Use of High Frequency Ionospheric Channel Simulators.

[redacted]

FIRST EXAMPLE OF DEMODULATION TECHNIQUE
(For Information Only)

Appendix 1: Complexity/Performance Analysis of Equalization by Recursive Filtering

DESCRIPTION OF A TECHNIQUE FOR DEMODULATION BY RECURSIVE FILTERING

Correction of received signal distortion is performed by automatic equalization, the principles and example of which are described below.

A. TRANSMISSION CHANNEL EQUALIZATION

1. Principle:

(a) equalization of a communications channel, the transfer function of which is F (f), consists of inserting a quadripole, characterized by transfer function G (f) in the transmission path, which satisfies the following equations throughout the signal passband:

$$G(f) \cdot F(f) = 1 + E(f),$$

where:

E(f), the equalization residual, gives a measure of the equalization quality, if E(f) is null, equalization is rigorous;

(b) the value of E(f) depends on:

- the level and nature of the noise;
- the structure of the equalizer;
- the criterion retained to determine the equalizer transfer function;
- the evolution of the communications channel;

(c) automatic equalization consists of the association of a variable characteristic filter and computing facilities which enable real time determination of the equalizer transfer function from the received signal;

(d) the equalization study for an ionsopheric channel, which started by simulation of the data transmission over a HF channel and various equalization processes, has led to the definition of an equalizer adapted to the ionsopheric channel.

2. Equalizer Structure:

(a) the filter used is a sampled filter having a hybrid structure, transversal and recursive, with a decision device inserted in the recursive part feedback loop (Decision Feedback Equalizer). This filter, a diagram of which is given in Figure C-1, consists of a transverse part with a double sampling rate of 32 coefficients and a recursive part of 8 coefficients;

(b) in the mathematical sense of the term, to exploit all the energy of the various paths, the filter has a complex structure which also renders it suitable to handle two-, four-, or eight-stage phase keyed signals;

(c) the presence of the decision device in the loop enables the stability range of the recursive filter to be increased and simultaneously suppresses the effect of noise on corrections made by the recursive part when the decision is error free, thus improving equalization quality;

(d) if the post-decision data are equal to the data transmitted and if a Z transform is used, the recursive filter transfer function can be written:

$$G(z) + \frac{A(z)}{1 - B(z)}$$

A (z) and B (z) are polynomial forms:

$$A(z) = \sum_{i=-N}^{+M} A_i z^i \qquad (M + N + 1\text{: transverse part length})$$

$$B(z) = \sum_{j=L}^{L} B_j z^j \qquad (L\text{: recursive part length})$$

(e) using the same conventions, the communications channel transfer function can be written:

$$F(z) = \sum_{j=-n}^{+m} F_j z^j \qquad (n + m + 1 \text{ is the number of samples in the channel percussional reply})$$

and the equalized channel transfer channel function is equal to:

$$F(z)\,G(z) = \frac{A(z)\,F(z)}{1 + B(z)} = 1 + E(z)$$

- the recursive part B (z) affects the coefficients of A (z) F (z), the exponent z of which is positive, i.e. samples of A (z) F (z) are delayed with respect to the maximum amplitude path;

- the linear part A (z) is set so that the coefficients corresponding to exponents of z negatives of product A (z) F (z) are as low as possible;

(f) the criterion used to obtain the equalizer setting consists of rendering the noise affecting the equalized signal minimum, whether this noise is due to equalization residuals or noise superimposed on the received signal. This criterion is equivalent to rendering the signal-to-noise ratio maximum before the decision and leads to an almost optimum error probability. The equalizer coefficients are real time calculated for example iteratively, with the gradient algorithm by means of simple calculations.

3. Equalizer Properties:

(a) the structure of the equalizer enables it to handle fading and time shifts over several paths without loss of data bits. At the instant of synchronization, the coefficient which weighs the sample corresponding to the maximum amplitude path is the central coefficient. When the propagation time varies, the position of this coefficient is altered by the coefficient adjustment algorithm and the clock synchronization to keep the transmission delay between the modulator and the demodulator constant. If the shift exceeds filter compensation capabilities, some data bits may be lost. To prevent this phenomenon, the central coefficient occupies a mid-position, on sychronization, which provides it with the fluctuation margin required for compensation of propogation time variations. Shift compensation is provided in real time by maintaining demodulation optimal;

(b) this equalizer can be used to correct received signals after ionospheric transmission and, in particular, those affected by selective fading, such as those generated by the combination of two paths of equal attenuation, a case whose existence was high-lighted during experiments;

(c) the algorithm used enables fluctuations in the propagation medium to be followed, while conserving a resistance to noise such that it can adapt to the channels for which the signal-to-noise ratio, in the baseband, is equal to 6 dB. The maximum bit error rate which the algorithm can tolerate is between 2 and 10%, depending on the multiple path characteristics.

B. FUNCTIONAL DESCRIPTION

1. Modulator:

(a) the modulator is broken down as follows, as shown in the block diagram of Figure C-2:

- junction with data terminal equipment;
- transcoding;
- storage of data:
  - insertion of synchronization sequences and known data;
  - data scrambling;
- transmission filtering;
- transposition to intermediate frequency at 1800 Hz;
- interface with radio transmitter.

2. Reception and Demodulation:

(a) the demodulator is broken down as follows, as shown in the block diagram of Figure C-3:

- receiver interface;
- analogue/digital conversion;
- transposition to baseband according to two quadrature channels;
- equalizer;
- synchronizations;
- interface with data terminal.

3. Equalizer:

(a) the signal of the two quadrature channels is processed by the equalizer, a description of which is provided in paragraphs 2 and

3 of Section A of this Annex. At the equalizer output the data signal is restored;

(b) on setting up the link, the equalizer must be apprenticed to propagation conditions. To achieve this, the synchronization sequence is used to accelerate equalizer convergence. Once convergence has been obtained, the equalizer is capable of following rapid changes in the propagation channel by virtue of the known data blocks inserted in the frame;

(c) regular repetition of the synchronization sequence is used to reinitialize the equalizer in the event of transmission interruptions;

(d) the unknown information symbols undergo unscrambling by signal multiplication prior to decision with the complex conjugate of the scrambling symbol.

4. <u>Synchronization</u>:

(a) the self-adapting modem has two types of synchronization:

- bit synchronization;

- frame synchronization;

(b) bit synchronization is intended to extract the clock associated with the transmission bit rate, with minimum phase noise, so as to define the optimum demodulated signal sampling moment:

- extraction is performed by a digital phase locked loop, using transitions of the demodulated signal preprocessed so as to render them independent of multiple paths and the frequency difference;

(c) frame synchronization is obtained by recognition of the known sequence. Recognition of the synchronization code is performed by means of a matched filter which synchronizes the modem to the maximum amplitude path.

5. <u>Frequency Correction</u>. The correlator incorporated in the frame synchronization unit delivers information on the frequency difference between transmitter and receiver. This information can be used for frequency corrections to the intermediate frequency generator to reduce this difference to zero.

6. <u>Data Restoration</u>. Once the data are recognized, the message must be restored in serial form at the rate at which it appears at the transmitting terminal.

7. Junction with Data Terminal. The junction with the data terminal restores the data received in serial form, together with the associated clock, and also enables the dialogues necessary to the various modem operating modes.

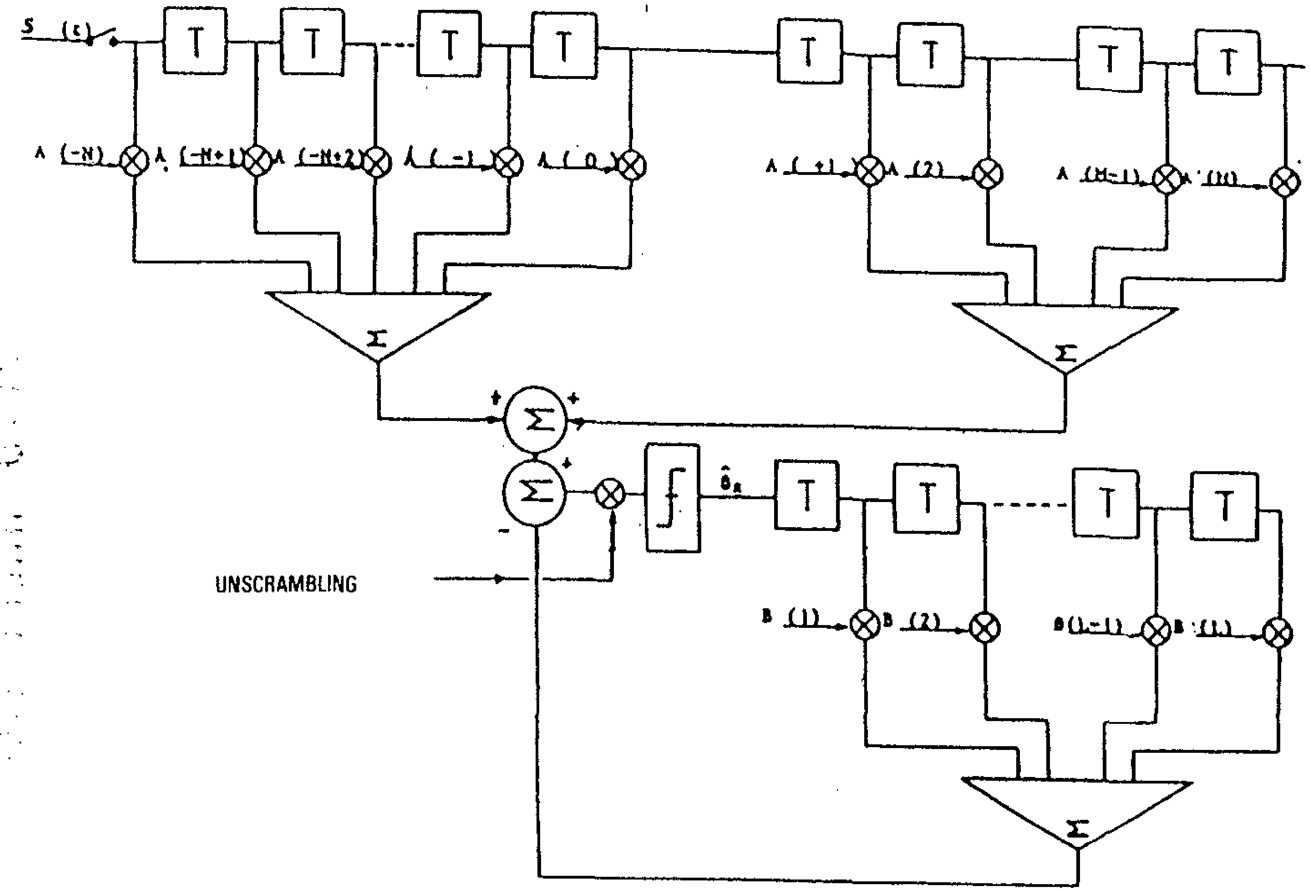


FIGURE C-1 : DECISION FEEDBACK EQUALIZER

FIGURE C-2 : MODULATOR

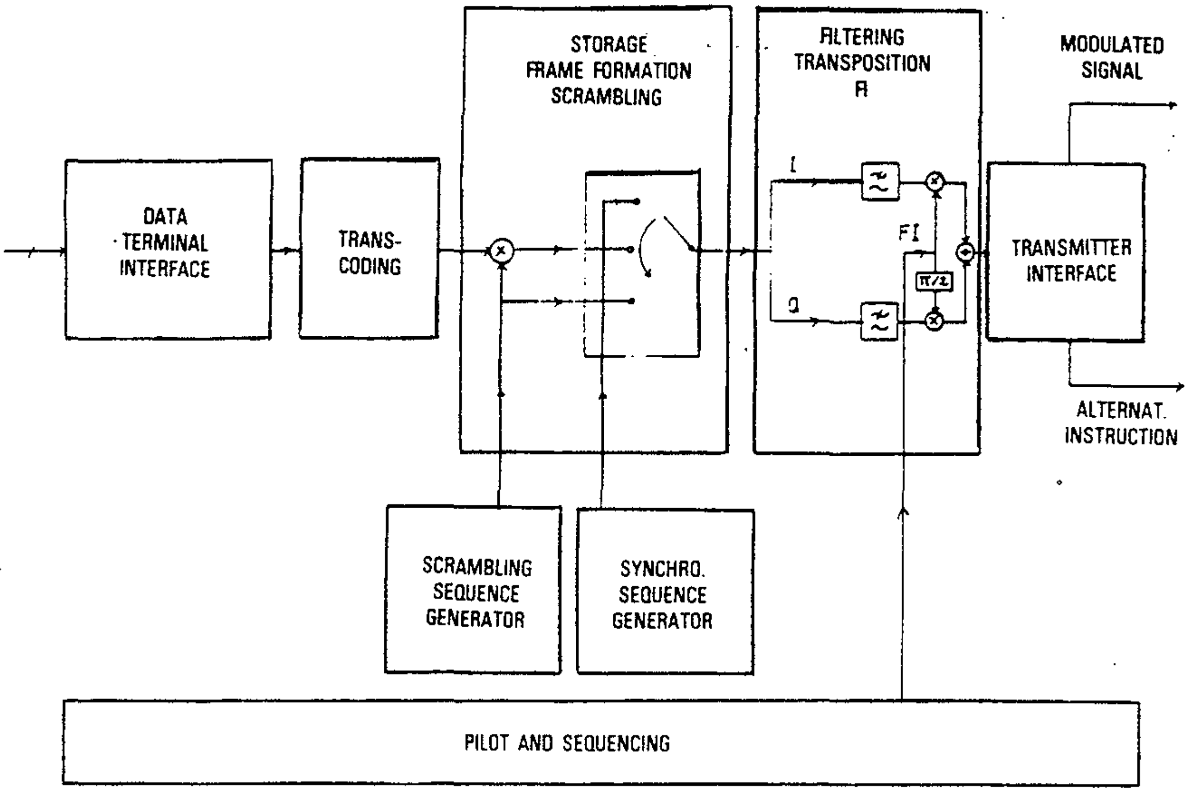
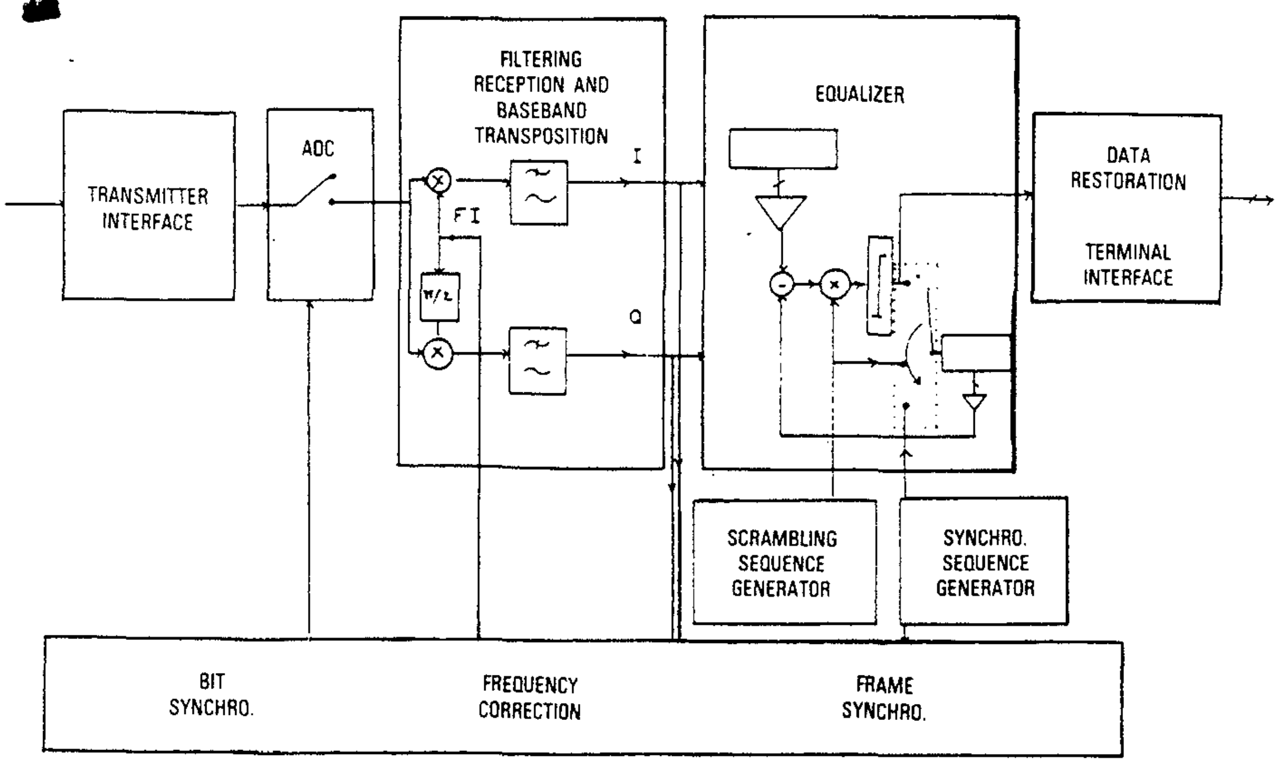


FIGURE C-3 : DEMODULATOR



COMPLEXITY/PERFORMANCE ANALYSIS OF EQUALIZATION BY RECURSIVE FILTERING USING A GRADIENT TAP UPDATE ALGORITHM

(For Information Only)

COMPLEXITY

1. Recursive filtering operation by a gradient algorithm calls for 2N complex multiplications and 2N complex additions, where N is the number of coefficients in the filter.

PERFORMANCE

2. (a) the performance of a modem of this type was measured by simulation for various propagation conditions (see Annex B);

(b) the various error rates are measured without using error correction codes;

(c) the first case corresponds to a single-path channel. The error probability at 3600, 2400 or 1200 bps in the presence of Gaussian white noise is given in Figure C-1-1 as a function of the signal-to-noise ratio (measured in a 3 kHz band);

(d) the second case corresponds to a single-path channel affected by a frequency spread of 1 Hz. The error probability curve as a function of the signal-to-noise ratio is given in Figure C-1-2;

(e) the two other cases correspond to propagation with two paths of equal amplitude shifted by T. Each path is affected by a frequency spread of magnitude . The two configurations used are described in CCIR Report 549;

	T	σ
CCIR channels:		
average	1 ms	0.5 Hz
poor	2 ms	1 Hz

(f) the error probability curves, as a function of the signal-to-noise ratio, are given in Figure C-1-3 for average channels and in Figure C-1-4 for poor channels;

(g) the modem's capability to resist frequency spread is shown in Figure C-1-5;

(h) this Figure shows the error rate as a function of the frequency spread for propagation along two paths of equal amplitude separated by 1 ms. Measurement is made without noise for rates of 3600, 2400 or 1200 bps;

(i) the modem's capability to resist time spread is shown in Figure C-1-6. This Figure shows the error rate as a function of time spread for propagation along two paths of equal amplitude each affected by a frequency spread of 0.5 Hz. Measurement is made without noise for rates of 3600, 2400 or 1200 bps.

[redacted]

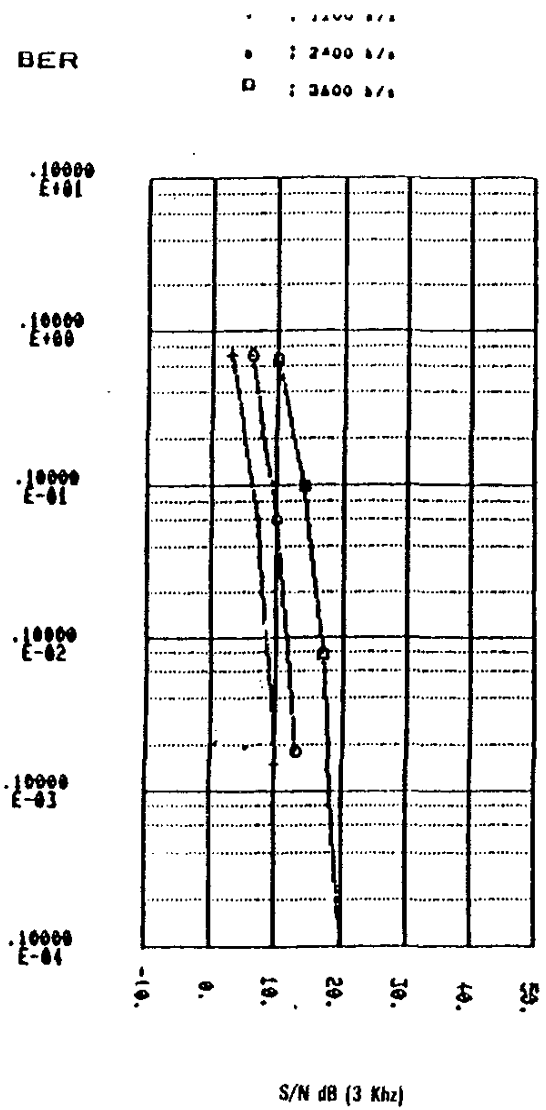


FIGURE C-1-1: ERROR RATE
SINGLE-PATH CHANNEL
GAUSSIAN WHITE NOISE

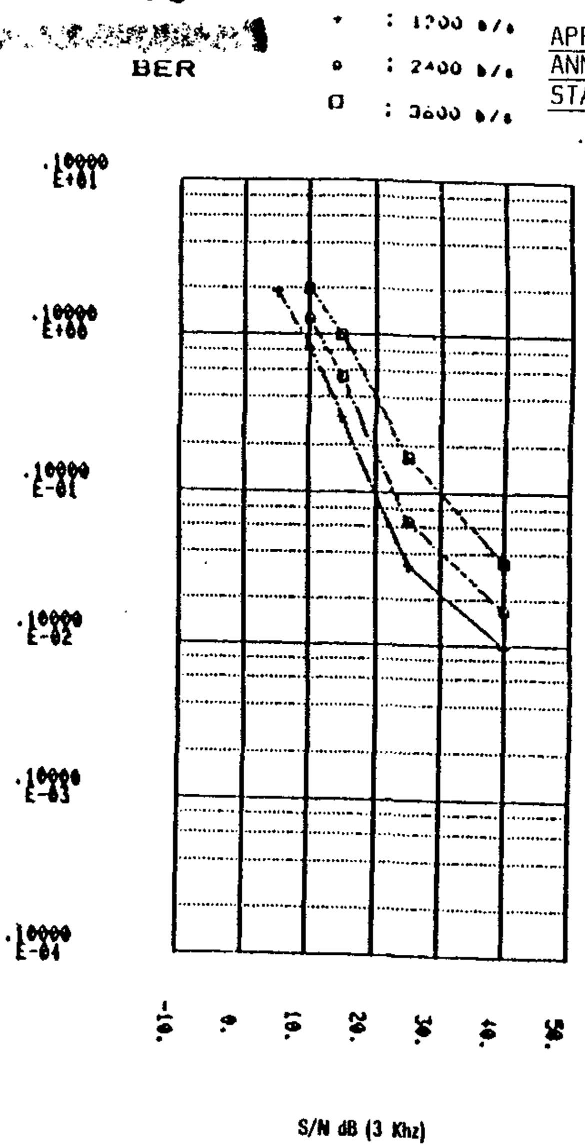


FIGURE C-1-2: ERROR RATE
SINGLE-PATH CHANNEL
FREQUENCY SPREAD : 1 Hz
GAUSSIAN WHITE NOISE

[redacted]

C-1-3

APPENDIX 1 to
ANNEX C to
STANAG 4285

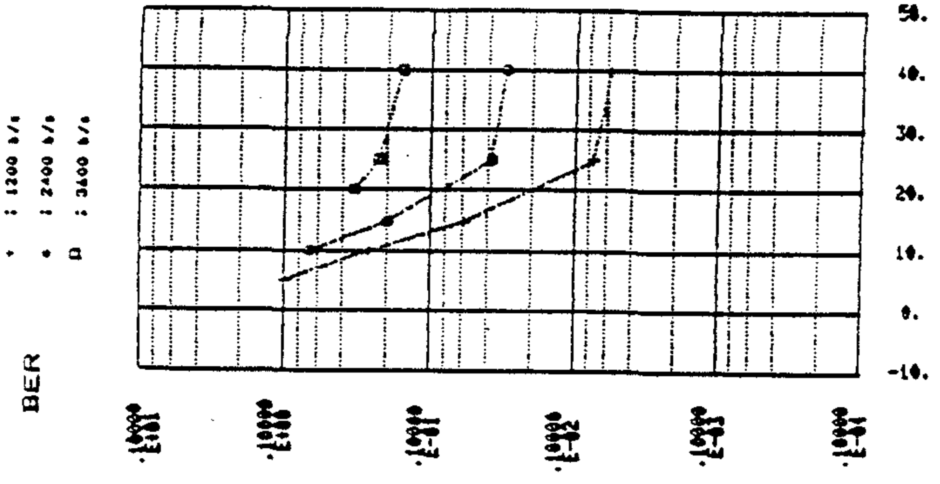


FIGURE C-1-3: ERROR RATE
AVERAGE CCIR CHANNEL
2 PATHS OF EQUAL AMPLITUDE
TIME SPREAD : 1 ms
FREQUENCY SPREAD : .5 Hz
GAUSSIAN WHITE NOISE

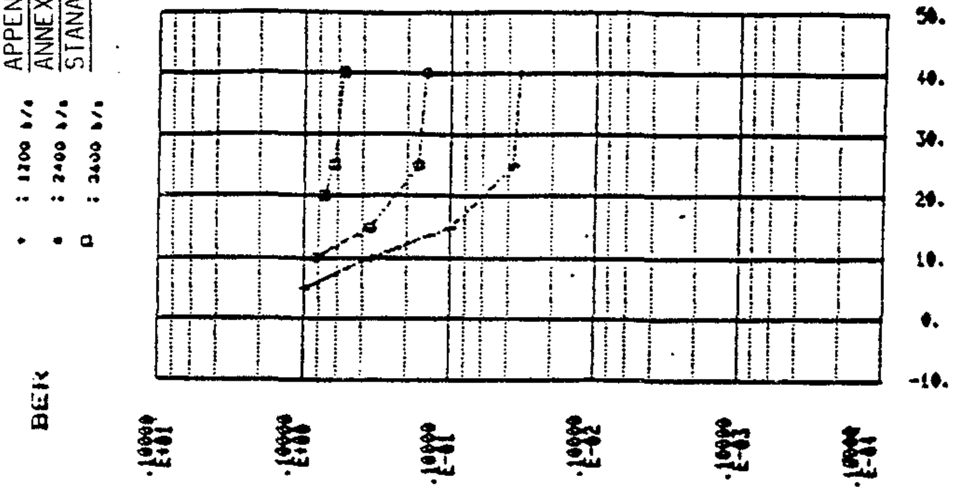


FIGURE C-1-4: ERROR RATE
POOR CCIR CHANNEL
2 PATHS OF EQUAL AMPLITUDE
TIME SPREAD : 2 ms
FREQUENCY SPREAD : 1 Hz
GAUSSIAN WHITE NOISE

[redacted]

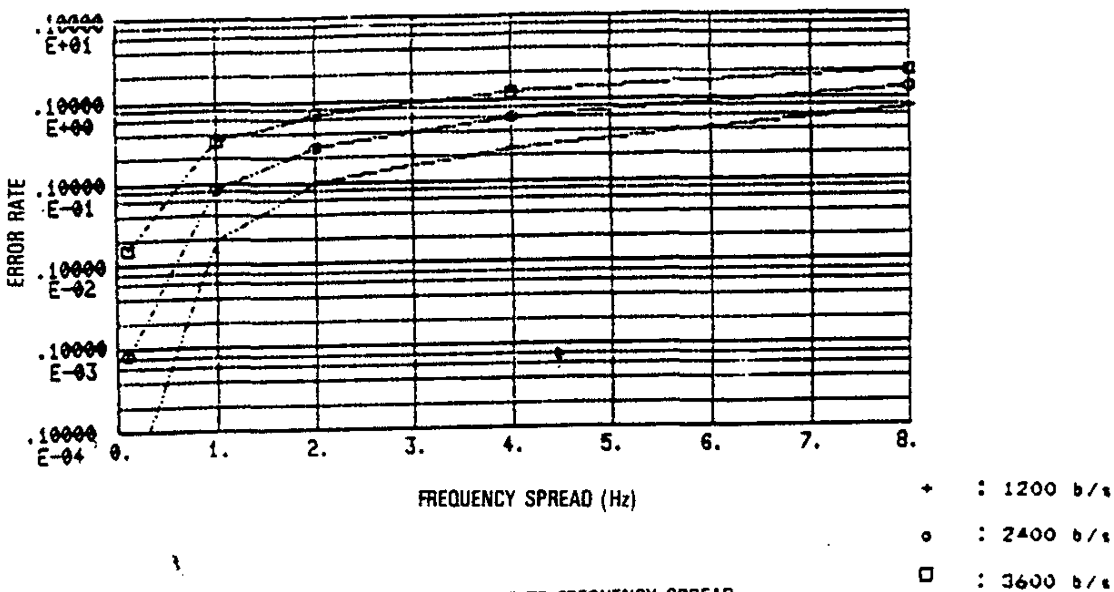


FIGURE C-1-5: RESISTANCE TO FREQUENCY SPREAD
2 PATHS OF EQUAL AMPLITUDE
TIME SPREAD : 1 ms
NO NOISE

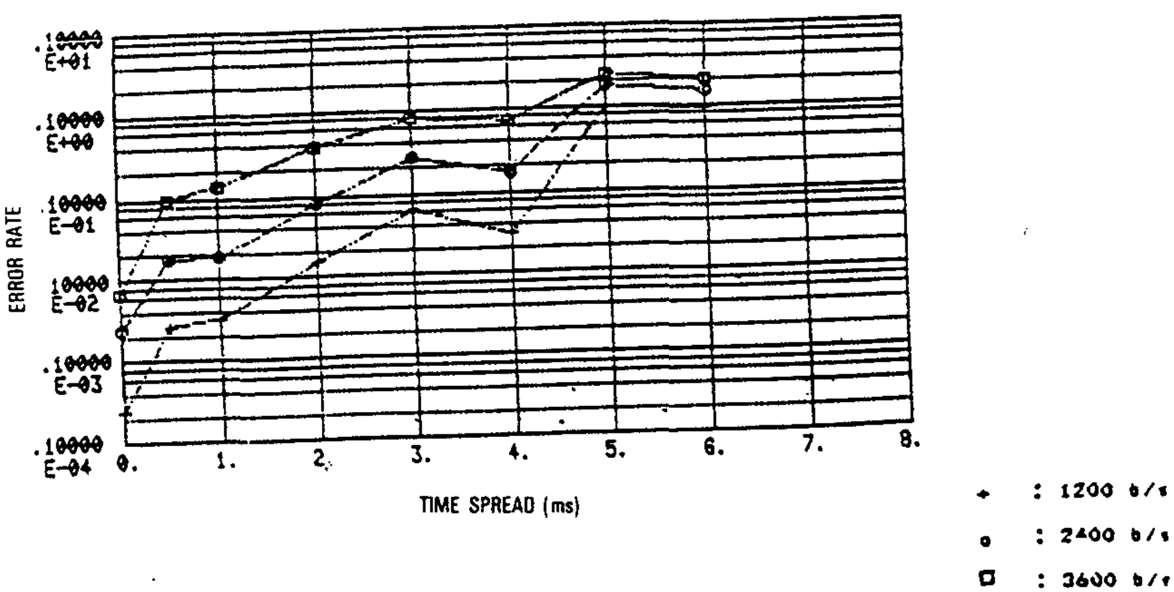


FIGURE C-1-6: RESISTANCE TO TIME SPREAD
2 PATHS OF EQUAL AMPLITUDE
FREQUENCY SPREAD : .5 Hz
NO NOISE

[redacted]

SECOND EXAMPLE OF DEMODULATION TECHNIQUE
(For Information Only)

Appendix 1: Complexity/Performance Analysis

DESCRIPTION OF A TECHNIQUE FOR DEMODULATION BY DATA DIRECTED EQUALIZATION

1. (a) this Annex describes an alternative equalization technique to the technique of Annex C. The data-directed equalization technique has the advantage that it is capable of providing good performance over a wider range of channel conditions and hence provides better availability than that provided by use of decision feeback. It has the disadvantage that it requires a greater computational capability than decision feedback and hence a more complicated processor;

   (b) since the data-directed equalization algorithms must be implemented in some sort of signal processing computer, the description of the suggested implementation is simply a description of the mathematics to be solved by the processor in order to make decisions or in order to provide soft decisions to an error correction decoder;

   (c) data-directed equalization, as opposed to most other equalization techniques, provides decisions or estimates on an entire block of unknown transmitted symbols based upon a block of received samples. The next block of transmitted symbols are estimated based upon the next block of received samples;

   (d) the general process involved is to estimate a block of transmitted signals based upon a hypothesized estimate of the conditions of the dynamic dispersive HF channel. Based upon these data estimates, the estimates of the channel conditions are upgraded. This process must be iterated at least once and preferably four times. The process of estimating transmitted symbol blocks will first be discussed and this will be followed by the procedure for updating the channel estimate.

TRANSMITTED DATA ESTIMATIONS

2. (a) a transmitted data "frame" is defined as a sequence of 64 sequential symbols in the format of Figure A-3 (Annex A). The 64 symbols contain the 32 unknown symbols in the centre with 16 known symbols on each side. In the case where the 32 unknown symbols are adjacent to the sequence of 80 known symbols, the 16 closest of these 80 are used as the known sequence. Successive "transmit frames" overlap in that the known symbols transmitted last in one frame serve as the first transmitted symbols in the next transmit frame. A receive frame consists of receive samples taken during the

reception of 48 symbols, 32 unknown symbols followed by the 16 known symbols (shown in blocks 1, 2, and 3 in Figure A-3, Annex A). The fourth receive frame consists of the samples corresponding to block 4 plus the first 16 known symbols of the 80 symbol synchronization sequence;

(b) as was explained in Annec C, the HF channel can be modelled as a transversal filter where the weighting values in the transversal filter are complex and vary with time. In this case each received value is a weighted sum of the transmitted values, t_k, and is of the form:

$$r_i = \sum_{j=1}^{N+1} W_{i+j-1} \quad T_{1+j} + N_i$$

where the N+1 taps in the tranversal filters are assumed to be spaced by a transmit symbol apart in time and the channel memory is no longer in time than N+1 symbols. The weighting values, W_{i+j-1} are complex as are the noise values N_i. The data directed approach is based upon the assumption that the channel memory (or multipath spread) is no longer in duration than the sequence of 16 known symbols (6.67 milliseconds). In this case, the set of received samples in a receive frame can be expressed as:

$$R = WT + N \qquad \text{(D-1)}$$

where R is a column vector consisting of the received samples in a receive frame, r_1, r_2 -- r_{48}

T is a column vector consisting of the transmit samples in a transmit frame, t_1, t_2 -- t_{64}

N is a column vector consisting of noise samples, n_1, n_2 -- n_{48}

and W is a column vector of channel weights of the form:

$$
W \quad \begin{array}{c} \overbrace{\hphantom{W_1 \quad W_2 \quad - - - - \quad W_{16} \quad W_{17} \quad 0 \quad 0 - - - - 0}}^{64} \\ \left|\begin{array}{ccccccccc} W_1 & W_2 & ---- & W_{16} & W_{17} & 0 & 0 & ---- & 0 \\ 0 & W_1 & ---- & W_{15} & W_{16} & W_{17} & 0 & ---- & 0 \\ - & - & - & - & - & - & - & - & - \\ 0 & 0 & ---- & ---- & W_1 & --- & W_{16} & & W_{17} \end{array}\right| \end{array} \quad 48
$$

(c) if we split the 64 transmit elements of the transmit vector into two components associated with the known and unknown symbols as shown in Table D-1, we can rewrite the previous vector equation as:

$$R = W_1 A + W_2 B + N \tag{D-2}$$

where A is a column vector of known symbols, a_1 --- a_{32}

B is a column vector of unknown symbols, b_1 --- b_{32}

and W_1 and W_2 are the portions of the W matrix which are operated upon by the A and B vectors;

(d) the data directed algorithm first estimates the set of b_i values that minimize the value of $N^{*T}N$ (where N^{*T} is the conjugate of the transpose of the noise vector). This vector estimate, B, is obtained by solving the matrix equation:

$$(W_2^{*T} W_2)\, B = [W_2^{*T}(R - W_1 A)] \tag{D-3}$$

Table D-1

	Components of Transmitted Vector T	Corresponding Components in Vectors A and B
	t_1	a_1
	t_2	a_2
N Known	.	.
Values	.	.
	.	.
	t_{16}	a_{16}
	t_{17}	b_1
	t_{18}	b_1
M Unknown	.	.
Values	.	.
	.	.
	t_{48}	b_{32}
	t_{49}	a_{17}
N Known	.	.
Values	.	.
	.	.
	t_{64}	a_{32}

Since the matrix $(W_2{}^{*T}W_2)$ is a Toplitz matrix, the equation can be solved without the computational load associated with a general matrix inverse.

(e) once the estimates of the unknown transmitted values are obtained, an algorithm called the "Procrastination Algorithm" is employed to further refine the estimates and provides the basis for decisions. This algorithm puts off making the most difficult symbol decisions until last which allows use of the most information in their selection. The approach is to look at the estimates associated with the two symbol values, b_1 and b_{32}, which are adjacent to the known symbols. From a statistical point of view, these symbol estimates are more likely to be close to the actual transmitted values than are those in the middle of the unknown symbols. The complex estimate which is closest to one of the eight possible complex transmit symbol values is selected, and a decision is made which corresponds to that of the closest transmit symbol value. (For use with a soft decision decoder, Annex E, the soft bit decisions are formed from this estimate.) The decided value is then assumed to be a "known" symbol for equation D-3 and the equation is resolved on this basis for the remaining 31 unknown symbols. This leads to two more "end" estimates, a new estimate for the end that was not selected and the new end estimate at b_2 or b_{31}, adjacent to the value previously selected. The process is then repeated with the number of unknowns reducing by one of each step until all 32 values have been decided (and soft decision metrics obtained);

(f) the above discussion is based upon obtaining one receive sample per transmit symbol. In order to reduce the stress associated with the symbol synchronization process, it is desirable to obtain two received samples per symbol. In this case, two separate sets of channel weight estimates must be obtained, one for each of the sampling phases. Equation D-3 can still be used where the matrix and vector elements are the average of the two values obtained during each phase;

TRACKING ALGORITHM

3. (a) as was explained previously, in order to carry out the decision algorithm, some means of determining the channel weights must be provided. In order to accommodate dynamic variations which might be anticipated in the HF channel, the approach taken must have the capability of tracking weight channel variations. Moreover, since the channel signal may fade for relatively long periods during a transmission and the receiver must be able to recover without requiring retraining (since the transmitter site may be unaware of the fade), the selected scheme must be capable of determining channel weights without the use of an initial training mode;

(b) these requirements are satisfied by re-estimating the channel weights each iteration based upon the decisions and the estimated weights of the previous iterations. Assuming initially that the weights for the previous frame have been established, the first step is to make a set of decisions using these weights in the algorithm described above. Based upon these 32 decisions, the two sets of 16 known transmitted values and the weights for the previous frame, the 48 received values can be predicted and 48 errors between the prediction and observed received values can be calculated. These errors are then correlated (multiplied and averaged) with the 64 assumed transmitted values corresponding to each of the channel weights in the predictions. The channel weights are updated by a fraction of the correlation. The value of the fraction is the loop gain and is chosen as a compromise between tracking speed and noise performance;

(c) in order to reduce the stress associated with weight tracking, it is recommended that a separate phase adjustment be provided for all weights based upon the average phase difference between the predicted and observed values.

FUNCTIONAL DESCRIPTION

4. The general functional description of the modem, other than that of the equalizer is the same as that described for the example of Annex C.

COMPLEXITY/PERFORMANCE ANALYSIS
(For Information Only)

COMPLEXITY

1. The number of mathematical operations required to perform the operations associated with data directed equalization, as described in this Annex, will depend strongly upon the specific algorithms employed to solve the matrix equations which may in turn depend upon the architecture of the processor that is to be used. Using very efficient algorithms, the equalization process requires approximately 9500 complex multiplies and adds and 34 real divides per received data frame. This converts to a requirement for 2,375,000 complex multiplies and adds (9,500,000 real multiplies and adds) and 8500 real divides per second if five iterations are used.

PERFORMANCE

2. The performance of a modem employing data-directed equalization has been estimated based upon the simulator measurements for various propagation conditions. These predictions are tabulated below:

BER for a Single Rayleigh Fading Path (1 Hz Doppler Spread)
Versus Gaussian Noise

AVERAGE SIGNAL/NOISE IN A 3 kHz BANDWIDTH	Data Rate (Bits/Second) Standard Uncoded 3600	2400
15 dB	6.2E-2	3.0E-2
25 dB	6.2E-3	3.1E-3
40 dB	6.7E-4	2.9E-4

BER for CCIR Poor Conditions Versus Gaussian Noise
(Two Independent Equal Average Power Rayleigh Fading Paths,
with 1 Hz Doppler Spreads, and Separated by 2 ms)

	Data Rate (Bits/Second)	
	Standard Uncoded	
AVERAGE SIGNAL/NOISE IN A 3 kHz BANDWIDTH	3600	2400
15 dB	3.6E-2	1.04E-2
20 dB	1.11E-2	1.48E-3
25 dB	4.0E-3	2.3E-4
40 dB	1.11E-3	<1E-5

BER for CCIR Moderate Conditions Versus Gaussian Noise
(Two Independent Equal Average Power Rayleigh Fading Paths,
with 0.5 Hz Doppler Spreads, and Separated by 1 ms)

	Data Rate (Bits/Second)	
	Standard Uncoded	
AVERAGE SIGNAL/NOISE IN A 3 kHz BANDWIDTH	3600	2400
15 dB	3.6E-2	1.02E2
20 dB	9.1E-3	1.25E-3
25 dB	2.3E-3	1.5E-4
40 dB	2.5E-4	<1E-5

[redacted]

BER Versus Doppler Spread of Two Independent Equal Average Power Rayleigh Fading Paths Separated by 1 ms with No Noise

	Data Rate (Bits/Second)	
	Standard Uncoded	
DOPPLER SPREAD	3600	2400
0.1 Hz	1.0E-5	0.0
1 Hz	1.25E-3	1.55E-5
2 Hz	8.5E-2	3.3E-4
4 Hz	4.6E-2	4.1E-3
8 Hz	1.5E-1	3.5E-2

BER Versus Multipath Spread of Two Independent Equal Average Power Rayleigh Fading Paths with 0.5 Hz Doppler Spread and No Noise

	Data Rate (Bits/Second)	
	Standard Uncoded	
MULTIPATH SPREAD	3600	2400
0.0 ms	3.5E-4	7.5E-5
0.5 ms	3.8E-4	1E-5
1 ms	2.5E-4	1E-5
2 ms	3.4E-4	2.1E-5
3 ms	5.6E-4	4.2E-5
4 ms	1.7E-3	1.1E-4
5 ms	1.1E-2	4.5E-3
6 ms	2.7E-1	1.2E-1

[redacted]

ERROR CORRECTION CODING, INTERLEAVING AND MESSAGE PROTOCOLS FOR USE WITH THE STANDARD MODULATION FORMATS
(For Information Only)

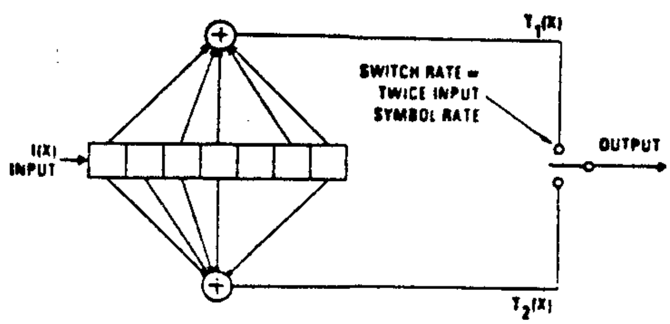
Appendix 1: Deinterleaving and Decoding Techniques for Use with the Error Correction Coding, Interleaving and Message Protocols
Appendix 2: Performance/Decision Feedback Equalization
Appendix 3: Performance Using Data Directed Equalization

INTRODUCTION

1. The use of error correction coding coupled with sufficient interleaving can make substantial improvements in the bit-error rate of data transmitted over HF channels. This document describes one specific implementation of coding and interleaving, which is well suited to the waveform modulation format, for data rates of 75 bps, 150 bps, 300 bps, 600 bps, 1200 bps and 2400 bps. A message protocol format is also specified to facilitate the optimal usage of the robust performance provided by the coding and interleaving.

ERROR CORRECTION CODING

2. (a) a constraint length 7, rate 1/2 convolutional encoder as shown in Figure E-1 is to be used as the basis for error correction coding for all of the data rates;



- RATE 1/2
- CONSTRAINT LENGTH = 7
- GENERATOR POLYNOMIALS:

FOR T1 $X^6 + X^4 + X^3 + X + 1$ (133)

FOR T2 $X^6 + X^5 + X^4 + X^3 + 1$ (171)

Figure E-1. Convolutional Encoder

(b) the two summing nodes in the figure represent modulo 2 addition. For each bit input to the encoder, two bits are taken from the encoder, the upper output bit, T1(x), being taken first. Coded bit streams of 4800 bps, 2400 bps and 1200 bps are thus generated for the input data rates of 2400 bps, 1200 bps and 600 bps respectively. For the lower input data rates, a 1200 bps coded bit stream is generated by repeating the pairs of output bits the appropriate number of times. It must be noted that the bits are repeated in pairs rather than repetitions of the first, T1(x), followed by repetitions of the second, T2(x). The basic formats for error correction coding at each of the data rates may be summarized as follows:

CODED DATA RATE	WAVEFORM FORMAT USED	EFFECTIVE CODE RATE	METHOD FOR ACHIEVING THAT CODE RATE
2400 b/s	8 Phase (3600 bps)	2/3	Rate 1/2 Punctured to Rate 2/3
1200 b/s	4 Phase (2400 bps)	1/2	Unmodified Rate 1/2 Code
600 b/s	2 Phase (1200 bps)	1/2	Unmodified Rate 1/2 Code
300 b/s	2 Phase (1200 bps)	1/4	Rate 1/2 Code Repeated 2 Times
150 b/s	2 Phase (1200 bps)	1/8	Rate 1/2 Code Repeated 4 Times
75 b/s	2 Phase (1200 bps)	1/16	Rate 1/2 Code Repeated 8 Times

(c) the puncturing of the rate 1/2 code to a rate 2/3 code for the 2400 bps data rate is done at output of the interleaver, to produce the 3600 bps rate needed for the 8-phase modulation format of the standard waveform. A 4800 bps encoded bit stream is therefore sent to the interleaver for the 2400 bps data rate.

INTERLEAVER

3. (a) the interleaving technique to be utilized is a slight modification of a normal convolutional interleaver. A conceptual representation of a convolutional interleaver and deinterleaver is shown in Figure E-2. For a normal implementation, coded bits are shifted into interleaver shift-registers (rows) on the left side of the Figure. On each new bit, the commutator switches into the next lower row (shift-register). Each shift-register has j more bits of storage than the preceding one above it. Bits are extracted through the output commutator in a similar fashion and transmitted over the channel. In the receiver, deinterleaving is performed by a similar operation, however, each shift-register in

[redacted]

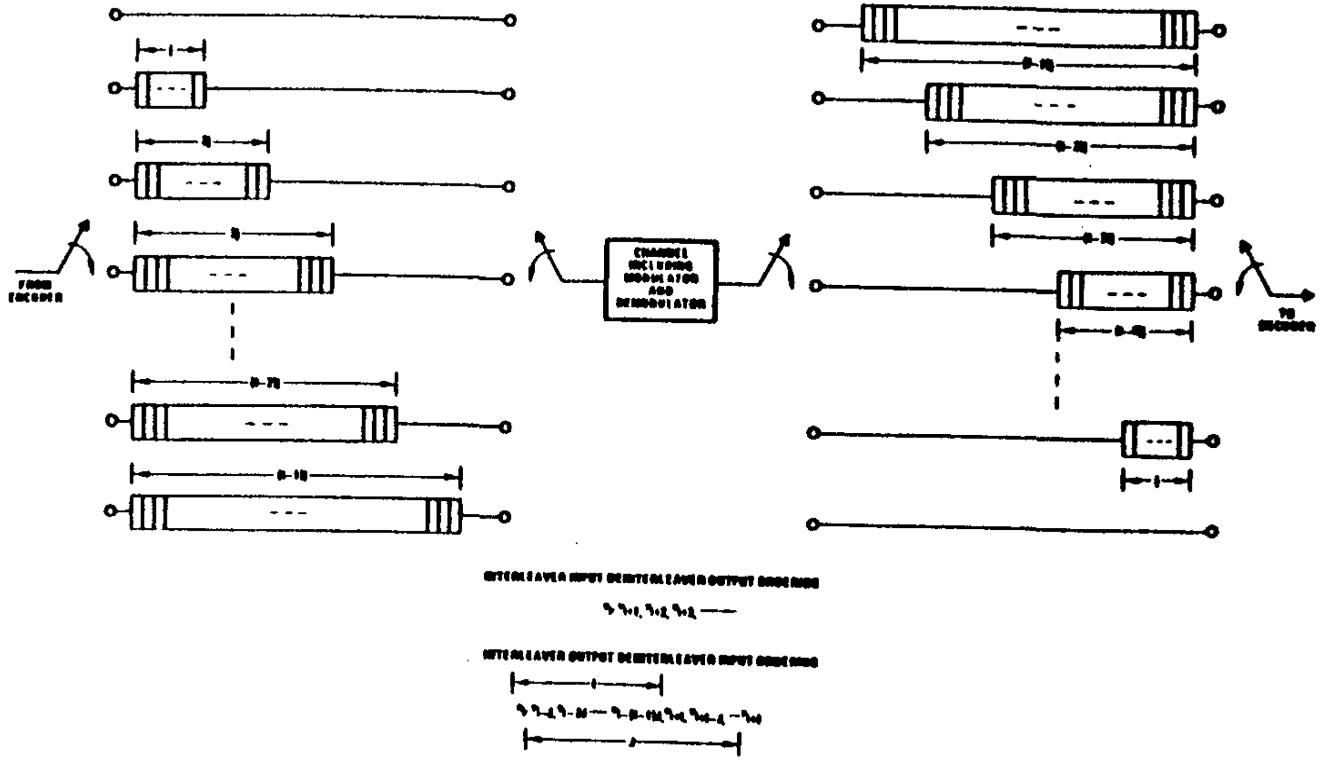


Figure E-2. Convolutional Interleaving

[redacted]

the deinterleaver has j fewer bits of storage than the preceding one above it. This normal interleaving technique is to be modified by making the commutators at the input of the interleaver and the output of the deinterleaver cycle through all of the rows in a nonsequential pattern as specified below. For the 2400 bps data rate, 3600 bps must be taken out of the interleaver instead of 4800 bps encoded rate input to the interleaver. This is accomplished by skipping every fourth row when taking bits out;

(b) the proper timing relationship between input and output bits with these modifications of the row sequencing is maintained if all rows of the shift-registers are shifted simultaneously after each complete cycle of the commutators. The specific parameters of the convolutional interleaver to be utilized in this application are as follows:

Number of rows:

I = 32 for all data rates

Delay increment "j" for each successive row:

	Total Interleaving Delay	
	10.24 seconds	0.853 seconds
Data Rates		
2400 bps	48	4
1200 bps	24	2
600 - 75 bps	12	1

Commutator row sequence for outputting bits from the interleaver at all data rates other than 2400 bps:

This is the normal sequence for one complete cycle.

0, 1, 2, 3, 4, 5, 6, 7, 8, 9, 10, 11, 12, 13, 14, 15, 16, 17, 18, 19, 20, 21, 22, 23, 24, 25, 26, 27, 28, 29, 30, 31

Commutator row sequence for outputting bits from the interleaver for the 8-psk modulation used for the 2400 bps data rate:

This is one complete cycle of 24 output bits and is repeated for every eight successive 8-psk symbols.

0, 1, 2, 4, 5, 6, 8, 9, 10, 12, 13, 14, 16, 17, 18, 20, 21, 22, 24, 25, 26, 28, 29, 30

Commutator row sequence for inputting bits in the interleaver:

This sequence is generated by using the modulo 32 results of the multiplication of each number in the normal sequence by 9.

0, 9, 18, 27, 4, 13, 22, 31, 8, 17, 26, 3, 12, 21, 30, 7, 16, 25, 2, 11, 20, 29, 6, 15, 24, 1, 10, 19, 28, 5, 14, 23

INTERLEAVER SYNCHRONIZATION

4. The interleaver and deinterleaver are synchronized when the two centre commutators shown in Figure E-2 are synchronized, i.e. a bit taken from the i th row of the interleaver is sent to the i th row of the deinterleaver. Synchronization shall thus be maintained by making the top (0th) row always be the starting position for the 1st bit sent or received in every data frame of the starting waveform. This is the only synchronization required since every frame of the transmitted waveform contains an integer multiple of 32 data bits.

INITIALIZATION AND MESSAGE PROTOCOL FOR USE WITH CODING AND INTERLEAVING

5. The following initializations and message formatting are to be used when operating with the coding and interleaving:

(a) the encoder shift-register and the interleaver shift-registers should be set to all zeroes before the start of any message;

(b) a unique 32-bit start of message pattern (SOM) is inserted into the bit stream sent to the encoder before the first bit of a message;

(c) a unique 32-bit end of message pattern (EOM) is appended after the last bit of the message;

(d) a string of zeroes, equal in length to the interleaver delay plus 102, is appended to the EOM in the bit stream. Only after the last of these zeroes is input to the encoder is the transmission terminated. These zeroes are used to flush the interleaver, coder, and allow for the maximum practical traceback delay in the receiver's decoder. Note that the encoder and interleaver will be filled with all zeroes, ready for the next message:

- SOM → ←---- MESSAGE BITS ----→ ←EOM → ←- FLUSH ZEROES -→

←------------ MODULATED WAVEFORM SENT TO TRANSMITTER ------------→

(e) the 32 bit words used for the SOM and EOM are as follows in hexadecimal representation (the left most bit is the first bit sent):

SOM = 03873C3C

EOM = 4B65A5B2

(f) the number of flush zeroes for each of the data rates and the two interleaver lengths are as follows:

DATA RATE	INTERLEAVER DELAY 10.24 SECONDS	INTERLEAVER DELAY 0.853 SECONDS
2400 bps	24678	2150
1200 bps	12390	1126
600 bps	6246	614
300 bps	3174	358
150 bps	1638	230
75 bps	870	166

DEINTERLEAVING AND DECODING TECHNIQUES FOR USE WITH THE ERROR CORRECTION CODING, INTERLEAVING AND MESSAGE PROTOCOLS
(For Information Only)

INTRODUCTION

1. The improvement in performance which can be realized through the use of error correction coding and interleaving as part of an HF modem is highly dependent on the decoding techniques utilized. Excellent performance with this particular coding and interleaving arrangement can be achieved if maximum likelihood decoding (Viterbi algorithm) of soft bit decisions is utilized. This Appendix outlines these procedures but does not describe the details of the Viterbi algorithm.

SOFT DECISION COMPUTATION

2. (a) a complex value (inphase and quadrature components) of each received, demodulated and equalized symbol is obtained. For the equalization techniques described previously, this is the equalizer output;

(b) Soft Bit decisions may be computed from this complex value as follows:

(1) the initial magnitude of each soft bit decision is the linear distance between this complex value and the nearest decision boundary line which would change the hard decision value of this bit. For BPSK modulation this is the magnitude of the real component;

(2) the final magnitude of the soft bit decision is this initial magnitude divided by the average of the squared distances between these symbol values and their corresponding perfect reference values. The time constant on this averaging is typically on the order of 100 ms;

(3) the sign of the soft decision is determined by the hard decision value, positive for a "0" and negative for a "1";

(4) for the 2400 bps data rate, a fourth soft bit decision is made for each 3-bit 8-PSK symbol. The value of this soft decision is zero, a complete erasure. This fourth soft bit decision thus properly represents an erasure decision for the bit that was skipped (punctured) when taking bits from the interleaver at the transmit end. This results in a rate of 4800 soft decisions per second which is appropriate for rate 1/2 decoding.

DEINTERLEAVING OF THE SOFT DECISIONS

3. The deinterleaver has the structure as shown previously in Figure E-2 of Annex E; however, each soft bit decision is typically at least 6 bits, rather than a single bit. This must be provided for in the deinterleaver. The soft decisions are entered into the deinterleaver in the normal commutator sequence and are taken out using the previously described non-sequential commutator sequence. Since soft decision erasures are generated for the punctured encoded bits used for the 2400 bps data rate, no special sequence is required for that case.

SOFT DECISION DECODING

4. For the data rates of 2400 bps, 1200 bps and 600 bps, soft decisions are taken from the deinterleaver two at a time and input to a soft decision Viterbi decoder. For data rates of 300 bps, 150 bps and 75 bps, soft decisions are taken from the deinterleaver four, eight or sixteen at a time respectively. The even numbered and odd numbered soft decision values are averaged separately to form two soft decision values which are input to the decoder.

INITIALIZING AND PROTOCOL

5. (a) the deinterleaver should be filled with all zeroes (erasures) prior to the input of the first soft decision of a message. The deinterleaving and decoding should begin immediately with the first soft decision computed;

(b) the last 26 bits (rightmost) of the 32-bit SOM should be successively correlated with the output of the decoder until a perfect match is made indicating the start of the message. The decoder output should then be output on the receive data line. If no SOM has appeared at the decoder output after a number of bits, equal to the appropriate number of zero flush bits plus 32, the decoder output should by default output on the receive data line;

(c) the decoder output should then be monitored for the 32-bit EOM and reception terminated when it is received. If the modem continuously fails to detect the 80 symbol synchronization preamble for a time interval equal to the interleaver delay, the reception should also be terminated.

PERFORMANCE/DECISION FEEDBACK EQUALIZATION
(For Information Only)

1. In this Appendix, predicted performance is presented when the decision feedback algorithm of Annex C is combined with the Error Correction coding interleaving technique described in Annex E. The performance is tablulated below.

NOTE

The feedforward section spans 6.67 ms. With taps at a half symbol spacing. The feedback section is 8 symbols long, thus spanning 3.33 ms.

BER for a Single Non-Fading Path Versus Gaussian Noise

Data Rate (Bits/Second)

Using Suggested Coding with 10 Sec. Interleaver

SIGNAL/NOISE IN A 3 kHz BANDWIDTH	2400	1200	600
3 dB			0.0
6 dB		6.00E-5	
7 dB			0.0
10 dB	8.45E-3	0.0	0.0
13 dB		0.0	0.0
14 dB	0.0		
16 dB		0.0	
17 dB	0.0		
20 dB	0.0		

BER for a Single Rayleigh Fading Path (1 Hz Doppler Spread) Versus Gaussian Noise

Data Rate (Bits/Second)

Using Suggested Coding with 10 Sec. Interleaver

AVERAGE SIGNAL/NOISE IN A 3 kHz BANDWIDTH	2400	1200	600
5 dB			2.86E-1
10 dB	0.5	3.18E-2	3.4E-4
15 dB	1.25E-1	0.0	0.0
20 dB	4.46E-4		
25 dB	0.0	0.0	0.0
40 dB	0.0	0.0	0.0

BER for CCIR Poor Conditions Versus Gaussian Noise (Two Independent Equal Average Power Rayleigh Fading Paths, with 1 Hz Doppler Spreads, and Separated by 2 ms)

Data Rate (Bits/Second)

Using Suggested Coding with 10 Sec. Interleaver

TOTAL AVERAGE SIGNAL/NOISE IN A 3 kHz BANDWIDTH	2400	1200	600
5 dB			6.63E-3
10 dB		1.33E-3	0.0
15 dB	2.13E-1	0.0	0.0
20 dB	5.65E-2		
25 dB	2.53E-2	0.0	0.0
40 dB	1.14E-2	0.0	0.0

[redacted]

BER for CCIR Moderate Conditions Versus Gaussian Noise
(Two Independent Equal Average Power Rayleigh Fading Paths,
with 0.5 Hz Doppler Spreads, and Separated by 1 ms)

TOTAL AVERAGE SIGNAL/NOISE IN A 3 kHz BANDWIDTH	Data Rate (Bits/Second) Using Suggested Coding with 10 Sec. Interleaver		
	2400	1200	600
5 dB		2.15E-1	6.67E-3
10 dB		1.78E-4	0.0
15 dB	1.49E-2	0.0	0.0
20 dB	9.90E-5		
25 dB	0.0	0.0	0.0
40 dB	0.0	0.0	0.0

BER Versus Doppler Spread of Two Independent Equal Average
Power Rayleigh Fading Paths Separated by 1 ms with No Noise

DOPPLER SPREAD	Data Rate (Bits/Second) Using Suggested Coding with 10 Sec. Interleaver		
	2400	1200	600
0.1 Hz	0.0	0.0	0.0
1 Hz	6.4E-5	0.0	0.0
2 Hz	2.71E-2	0.0	0.0
4 Hz	2.63E-1	5.46E-4	0.0
8 Hz	0.5	1.81E-1	1.68E-3

[redacted]

BER Versus Multipath Spread of Two Independent Equal Average Power Rayleigh Fading Paths with 0.5 Hz Doppler Spread and No Noise

MULTIPATH SPREAD	Data Rate (Bits/Second) Using Suggested Coding with 10 Sec. Interleaver		
	2400	1200	600
0.0 ms	0.0	0.0	0.0
0.5 ms	0.0	0.0	0.0
1 ms	0.0	0.0	0.0
2 ms	2.88E-4	0.0	0.0
3 ms	3.19E-2	0.0	0.0
4 ms	2.36E-2	0.0	0.0
5 ms	0.5	2.27E-1	4.73E-3
6 ms	0.5	2.50E-1	2.24E-3

PERFORMANCE USING DATA DIRECTED EQUALIZATION
(For Information Only)

1. In this Appendix, predicted performance is presented when the data directed algorithm of Annex D is combined with the Error Correction coding/interleaving technique described in Annex E. The performance is tabulated below.

BER for a Single Non-Fading Path Versus Gaussian Noise

Data Rate (Bits/Second)

Using Suggested Coding with 10 Sec. Interleaver

SIGNAL/NOISE IN A 3 kHz BANDWIDTH	2400	1200	600
2 dB	0.5	7.0E.2	6.5E-5
4 dB	0.5	2.0E-3	0.0
6 dB	0.5	5.5E-6	0.0
8 dB	2.2E-2	0.0	0.0
10 dB	2.1E-4	0.0	0.0
12 dB	0.0	0.0	0.0

BER for a Single Rayleigh Fading Path (1 Hz Doppler Spread)
Versus Gaussian Noise

Data Rate (Bits/Second)

Using Suggested Coding
with 10 Sec. Interleaver

AVERAGE SIGNAL/NOISE IN A 3 kHz BANDWIDTH	2400	1200	600
5 dB	0.5	0.5	7.1E-3
10 dB	0.5	2.1E-4	0.0
15 dB	1.3E-3	0.0	0.0
25 dB	0.0	0.0	0.0
40 dB	0.0	0.0	0.0

BER for CCIR Poor Conditions Versus Gaussian Noise
(Two Independent Equal Average Power Rayleigh Fading Paths,
with 1 Hz Doppler Spreads, and Separated by 2ms)

Data Rate (Bits/Second)

Using Suggested Coding
with 10 Sec. Interleaver

TOTAL AVERAGE SIGNAL/NOISE IN A 3 kHz BANDWIDTH	2400	1200	600
5 dB	0.5	0.5	1.0E-4
10 dB	0.5	3.0E-4	0.0
15 dB	2.0E-3	0.0	0.0
20 dB	0.0	0.0	0.0
25 dB	0.0	0.0	0.0
40 dB	0.0	0.0	0.0

BER for CCIR Moderate Conditions Versus Gaussian Noise
(Two Independent Equal Average Power Rayleigh Fading Paths,
with 0.5 Hz Doppler Spreads, and Separated by 1 ms)

	Data Rate (Bits/Second) Using Suggested Coding with 10 Sec. Interleaver		
TOTAL AVERAGE SIGNAL/NOISE IN A 3 kHz BANDWIDTH	2400	1200	600
5 dB	0.5	0.5	1.2E-2
10 dB	0.5	1.8E-3	0.0
15 dB	5.0E-3	0.0	0.0
20 dB	0.0	0.0	0.0
25 dB	0.0	0.0	0.0
40 dB	0.0	0.0	0.0

BER Versus Doppler Spread of Two Independent Equal Average
Power Rayleigh Fading Paths Separated by 1 ms with No Noise

	Data Rate (Bits/Second) Using Suggested Coding with 10 Sec. Interleaver		
DOPPLER SPREAD	2400	1200	600
0.1 Hz	0.0	0.0	0.0
1 Hz	0.0	0.0	0.0
2 Hz	0.0	0.0	0.0
4 Hz	0.0	0.0	0.0
8 Hz	2.5E-2	0.0	0.0

APPENDIX 3 to
ANNEX E to
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BER Versus Multipath Spread of Two Independent Equal Average Power Rayleigh Fading Paths with 0.5 Hz Doppler Spread and No Noise

Data Rate (Bits/Second)

Using Suggested Coding with 10 Sec. Interleaver

MULTIPATH SPREAD	2400	1200	600
0.0 ms	0.0	0.0	0.0
0.5 ms	0.0	0.0	0.0
1 ms	0.0	0.0	0.0
2 ms	0.0	0.0	0.0
3 ms	0.0	0.0	0.0
4 ms	0.0	0.0	0.0
5 ms	0.0	0.0	0.0
6 ms	0.5	0.5	0.5

USE OF THE SYNCHRONIZATION SEQUENCE FOR SIGNAL DETECTION AND ACQUISITION/TRACKING OF DOPPLER, SYNCHRONIZATION AND CHANNEL PARAMETERS (For Information Only)

INTRODUCTION

1. (a) at the start of a receiption, the synchronization sequence can be used for:

   - signal detection;
   - Doppler acquisition;
   - synchronization acquisition;
   - channel acquisition;

   (b) during the reception of data, a periodic initialization of reception algorithms is possible;

   (c) in order to perform these tasks as efficiently as possible, the synchronization sequence should consist of a cut-out of a periodically repeated pseudo noise (PN) sequence transmitted with 2 PSK;

   (d) in the following, a PN sequence of period 31 symbols has been assumed. To get the 80 symbol synchronization sequence, the PN sequence must be repeated 80/31 times.

SIGNAL DETECTION

2. (a) if g(i) is the sequence of received input samples in the low pass domain, then the first step consists in correlating this sequence g(i) with a reference PN sequence $x_{pnref}(i)$:

   $$v(i) = g(i) \circ x_{pnref}(i) \qquad (F-1)$$

   o means continuous correlation (matched filtering). This correlation is performed according to the sampling theorem, i.e. $x_{pnref}(i)$ contains at least $2 \cdot 1500/2400$ samples per PN symbol (bandwidth of low pass signal = 1500 Hz). v(i) is the continuous output signal of the correlator, which is considered in the following;

(b) now the energy E(i) over v(i) is taken within a sliding window, i.e.:

$$E(i) = \sum_{k=o}^{NH} |v(i-k)|^2 \qquad \text{(F-2)}$$

NH is the maximum length expected for the channel impulse response;

(c) if a periodically repeated channel impulse response occurs in the sequence v(i), then E(i) will also be periodic; the maximum is given if the channel impulse response is within the window of length NH and the minimum is given if there is no channel impulse response within the window (i.e. in the middle between the two periods resulting from the two periods of the transmitted PN sequence);

(d) if the actual energy, the energy one period in the past and the energy in the middle between those two points in time, is considered, i.e.:

$$\begin{aligned} E1(i) &= E(i) \qquad \text{(F-3)} \\ E2(i) &= E\,(i-NPN) \\ E3(i) &= E(i-NPN/2) \end{aligned}$$

then it is possible to define a quantity synchronization(i) as follows:

$$\text{(F-4)}$$

$$\text{synchronization}(i) = \begin{cases} 1 \text{ if} & \dfrac{E1(i)}{E3(i)} \geq C \wedge \dfrac{E2(i)}{E3(i)} \geq C \\ 0 \text{ otherwise} & \end{cases}$$

NPN is the length of the PN sequence in samples according to the sampling theorem, C is a threshold and $\wedge$ means logical and;

(e) synchronization(i)=1 is an indication of the presence of a periodic behaviour within the sequence of energy values E(i) (max-min-max) with proper period:

(1) the behaviour of synchronization(i) is independent of the special shape of the actual channel impulse response;

(2) in case of synchronization(i)=1, the acquisition process will be started (see paragraph 3);

(f) to discriminate against unwanted periodic interference, the synchronization sequence symbols are received (like data symbols) and the errors are counted. Only if the number of errors is below a given threshold, will the detection of a valid synchronization sequence be declared;

(g) the reference PN sequence $x_{pnref}(i)$ used for correlation in (F-1) may be a cyclic shifted version of the PN sequence used at the transmitting side. Any cyclic shift results in a <u>non-cyclic</u> shift of the output signal v(i) of the correlator. Therefore, it is possible to adjust the position of the two periods of channel impulse responses without changing the transmitted signal of the modem. The cyclic shift and hence the position should be determined in order to minimize disturbing effects from preceeding noise, AGC and data following the synchronization sequence.

<u>ACQUISITION OF SYNCHRONIZATION, CHANNEL AND DOPPLER</u>

3. (a) synchronization may be determined easily from synchronization(i) (see equ. F-3 and F-4), e.g. it could be the position for which the channel impulse response is at the middle of the window;

(b) if the latest values of v(i) have been stored, the channel impulse response is known too;

(c) the frequency offset between transmitting and receiving side (or Doppler shift) may be calculated from the phase shift between the two channel impulse responses. Because of the range allowed for this offset ($\pm$75 Hz), it is necessary to perform all calculations above, e.g. in 3 separate Doppler channels.

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