DOCUMENTATION

The following documents have been selected for document ordering at the IEEE Document ordering center.

11.1	P802.4L/89-14a: Factory RF channel modeling	121
11.2	P802.4L/89-14b: UHF Characterization in factory	450
11.3	P802.4L/90-08a: Microwave Oven Interference Measurement.	234
11.4	P802.11/90-10: Minutes of September 1990 meeting	32
11.5	P802.11/90-15: Archive August 1990	24
11.6	Selection 1 of various P802.4L documents	102.

Selection of documents of the IEEE P802.4L Task Group

The contents of 11.6 has been tentatatively selected . Any suggestions for change are welcome. The following is the current selection:

IEEE P802.4L/88-02	Communications over an Indoor Radio Channel (Bruce Tuch, NCR)
IEEE P802.4L/89-16	DQPSK Spread-Spectrum Modulation/ Demodulation (Bruce T. Tuch, NCR)
IEEE P802.4L/89-19	Statistic analysis of Oshawa analysis (L. van der Jagt, KII)
IEEE P802.4L/89-20	Impulse noise effect on 4 level QAM Spread Spectrum Signal (Donald C. Johnson, NCR)
IEEE P802.4L/89-21	Retail measurement results (Donald C. Johnson, NCR)
IEEE P802.4L/89-22	Microwave Oven Emissions (Donald C. Johnson, NCR)
IEEE P802.4L/90-16	RLAN Standard, Questions (V. Hayes - NCR)
IEEE P802.4L/90-17	RLAN Standard, Positions and Arguments (V. Hayes - NCR)
IEEE P802.4L/90-22	Running objectives and directions document (eighth edition, V. Hayes - NCR)

For your information, the whole list of papers distributed by IEEE P802.4L is provided in the appendix.

Appendix 1
Document list 1987

IEEE P802.4L/87-01	Objectives for 802.4L--Through-the-air Token Bus Physical Layer--Draft Proposal (Rypinski)
IEEE P802.4L/87-02	Memorandum re CCIR IWP 8/13 Plenary meeting
IEEE P802.4L/87-03	Minutes of the November 1986 meeting (Hotel Del Coronado, San Diego, CA)
IEEE P802.4L/87-04	Detail Supplement to objectives for 802.4L-- Proposals for radio only (Rypinski)
IEEE P802.4L/87-05	Charter and objectives-- IEEE 802.4L Through-the-Air Token Bus Physical Layer
IEEE P802.4L/87-06	Minutes of the March 1987 meeting (New Orleans, LA)
IEEE P802.4L/87-07	FCC Rules and possible LAN operating frequencies (Rypinski)
IEEE P802.4L/87-08	System plan, Wireless LAN--Bus mode, 1.6 Mbit/s, Spread Spectrum, 2400-2483.5 MHz (Rypinski)
IEEE P802.4L/87-09	System plan, Wireless LAN--Bus mode, 3.2 Mbit/s, Spread Spectrum, 1850-1990 MHz (Rypinski)
IEEE P802.4L/87-10	Draft of preparation for counsel, petition for rule making, industrial wireless automation service (Rypinski)
IEEE P802.4L/87-11	System plan, Wireless LAN--Bus mode, 2 Mbit/s, 1700-1710 or 2150-2160 MHz (Rypinski)
IEEE P802.4L/87-12	Investigation of possible optical systems Saito, NEC)
IEEE P802.4L/87-13	FCC Docket 81-413, "In the matter of Authorization of Spread Spectrum and other wideband emissions not presently provided for in the FCC rules and regulations
IEEE P802.4L/87-14	Project Authorization for 802.4L--Passed by the IEEE 802 Executive Committee, July 1987
IEEE P802.4L/87-15	Minutes of the July 1987 meeting (Vancouver, B.C. Canada)
IEEE P802.4L/87-16	Optical fixed point network with 4 links of 2x64 kbit/s, C&C NET STAR2800 (Takumi and Saito NEC)
IEEE P802.4L/87-17	Summary of proposed changes to Part 15 of the Regulations, FCC News release
IEEE P802.4L/87-18	Minutes of the November 1987 meeting (Ft Lauderdale, FA)

Document list 1988

IEEE P802.4L/88-01	Minutes of the July 1988 meeting (Danvers MA)
IEEE P802.4L/88-02	Communications over an Indoor Radio Channel (Bruce Tuch, NCR)
IEEE P802.4L/88-03	Minutes of the November 1988 meeting (Phoenix AZ)
IEEE P802.4L/88-04	Running objectives and directives document (V. Hayes, NCR)
IEEE P802.4L/88-05	Venue for the Interim Meeting Jan 16-18, 1989, Chicago.
IEEE P802.4L/88-06	The portable radio propagation environment. US contribution to CCIR, Doc US 8/13-3 Rev 1
IEEE P802.4L/88-07	Numerical Parameters for radio local area network. C. Rypinski, RadioLAN

Appendix 1
Document list 1989

IEEE P802.4L/89-00	Mailing list of the Taskgroup
IEEE P802.4L/89-01	Running objectives and directives document (second issue, V. Hayes, NCR)
IEEE P802.4L/89-02	Minutes of the January 1989 meeting (Chicago IL)
IEEE P802.4L/89-03	Venue for the Interim Meeting May 22-24, 1989, Atlanta GA
IEEE P802.4L/89-04	Minutes of the March 1989 meeting (New Orleans LA)
IEEE P802.4L/89-05	Additional Reports from the New Orleans meeting (M. Masleid, Inland Steel)
IEEE P802.4L/89-06	Third Edition of the Running objectives and directions document (V. Hayes, NCR)
IEEE P802.4L/89-07	Noise Characteristics (M. Masleid)
IEEE P802.4L/89-08	Minutes of the May 1989 meeting (Atlanta GA)
IEEE P802.4L/89-09	Proposed testing at General Motors Oshawa plant (L. van der Jagt, KII)
IEEE P802.4L/89-10	FCC Rules on Direct Sequence Spread Sppectrum (D.C. Johnson, NCR)
IEEE P802.4L/89-11	Minutes of the July 1989 meeting (Vancouver BC)
IEEE P802.4L/89-12	Running objectives and directives document (fourth edition, V. Hayes, NCR)
IEEE P802.4L/89-13	Venue for the Interim Meeting Sep 11-15, 1989, Chicago IL
IEEE P802.4L/89-14a	*Radio Channel Modelling in Manufacturing environments (parts II and III)* by Scott Y. Seidel, Koichiro Takamizawa and Dr. Theodore S. Rappaport, Bradley Department of Electrical Engineering, Virginia Polytechnic Institute and State University - Blacksburg, VA 24061. February 28, 1989
IEEE P802.4L/89-14b	*Characterization of UHF factory multipath channels* by Dr. Theodore S. Rappaport and Clare D. MacGillem, Engineering Research Center for intelligent manufacturing systems, Schools of engineering, Purdue University, West Lafayette, Indiana 47907. TR-ERC 88-12. June 1988
IEEE P802.4L/89-15	Minutes of the September 1989 meeting (Chicago Illinois)
IEEE P802.4L/89-16	DQPSK Spread-Spectrum Modulation/ Demodulation (Bruce T. Tuch, NCR)
IEEE P802.4L/89-17	Minutes of the November 1989 meeting (Fort Lauderdale Florida)
IEEE P802.4L/89-18	Venue for the Interim Meeting Jan 16-20, 1990, Parsippany, NJ
IEEE P802.4L/89-19	Statistic analysis of Oshawa analysis (L. van der Jagt, KII)
IEEE P802.4L/89-20	Impulse noise effect on 4 level QAM Spread Spectrum Signal (Donald C. Johnson, NCR)
IEEE P802.4L/89-21	Retail measurement results (Donald C. Johnson, NCR)
IEEE P802.4L/89-22	Microwave Oven Emissions (Donald C. Johnson, NCR)
IEEE P802.4L/89-23a	UHF Fading in Factories. By Theodore S. Rappaport and Clare D. McGillem. - IEEE Journal on selected Areas in Communications. Vol. 7. No 1. January 1989
IEEE P802.4L/89-23b	Indoor Radio Communications for Factories of the Future. By Theodore S. Rappaport. - IEEE Commmunications Magazine. May 1989.

IEEE P802.4L/89-24	Running objectives and directives document (fifth edition, V. Hayes, NCR)

Document list 1990

IEEE P802.4L/90-01	Minutes of the January 1990 meeting (Parsippany, NJ)
IEEE P802.4L/90-02	Microwave Oven Interference Measurement (Jonathon Cheah - Hughes Network Systems)
IEEE P802.4L/90-03	Wireless Radio LAN Issues for March Meeting (Stan Kay & Jonathon Cheah - Hughes Network Systems)
IEEE P802.4L/90-04	Wireless Radio LAN Modulation and Coding Proposal and Tradeoffs (Stan Kay - Hughes Network Systems)
IEEE P802.4L/90-05	Running objectives and directions document (sixth edition, V. Hayes, NCR)
IEEE P802.4L/90-06	Minutes of the March 1990 meeting (Irvine, CA)
IEEE P802.4L/90-07	Microwave Oven Frequency Content Measurements (Paul Pirillo, NCR)
IEEE P802.4L/90-08a	Submission on Microwave Oven Interference Measurement. By Jonathon Cheah, Hughes Network Systems.
IEEE P802.4L/90-09	Approaches to Indoor RADIOLAN which provide Processing Gain and Coding Gain (Larry van der Jagt, KII)
IEEE P802.4L/90-10	Venue for the Interim Meeting May 14-18, 1990, Atlanta GA
IEEE P802.4L/90-11	Considerations for a "head-end" distribution system for the 802.4L radio LAN (Tom Phinney, Honeywell, Inc)
IEEE P802.4L/90-12	Running objectives and directions document (seventh edition, V. Hayes, NCR)
IEEE P802.4L/90-13	(Kiwi Smit, NCR)
IEEE P802.4L/90-12	Running objectives and directions document (seventh edition, V. Hayes, NCR)
IEEE P802.4L/90-13	FEC Potential in case of Narrow Band Interference (Kiwi Smit - NCR)
IEEE P802.4L/90-14	Clarification of the Propagation Parameters in the Running Objectives and Directions Document (Donald C. Johnson - NCR)
IEEE P802.4L/90-15	The 11 chip Barker sequence and the FCC rules (Kiwi Smit - NCR)
IEEE P802.4L/90-16	RLAN Standard, Questions (V. Hayes - NCR)
IEEE P802.4L/90-17	RLAN Standard, Positions and Arguments (V. Hayes - NCR)
IEEE P802.4L/90-18	Analysis of the Impact of propagation from multiple antennae on complex conjugate demodulators (Larry van der Jagt - KII)
IEEE P802.4L/90-19	Description of a set of codes for CDMA (Larry van der Jagt - KII)
IEEE P802.4L/90-20	Use of 902-928 MHz by field disturbance devices of Sensormatic (Olin Giles - Sensormatic)
IEEE P802.4L/90-21	Minutes of the May 1990 meeting (Atlanta, GA)
IEEE P802.4L/90-22	Running objectives and directions document (eighth edition, V. Hayes - NCR)

Appendix 1
Document list 1990, cont'd

IEEE P802.11/90-1 IEEE P802.4L/90-23	Venue for the Interim Meeting Sep 10-14, 1990, Oshawa, ON, CANADA
IEEE P802.11/90-2 IEEE P802.4L/90-24	Minutes of the July 1990 meeting (Denver CO)
IEEE P802.11/90-3 IEEE P802.4L/90-25	FCC News release on Spread Spectrum rules
IEEE P802.11/90-4 IEEE P802.4L/90-26	FCC Notice of Inquiry, GEN Docket 90-314
IEEE P802.11/90-5 IEEE P802.4L/90-27	FCC News release on GEN Docket 90-314
IEEE P802.11/90-6 IEEE P802.4L/90-28	Comment from IEEE P802 to FCC as submitted to the Executive Committee
IEEE P802.11/90-7 IEEE P802.4L/90-29	Motion, Qualification criteria , Objectives, Reasons and Arguments and proposed PAR for starting a new Working Group as submitted to the Executive Committee
IEEE P802.11/90-8 IEEE P802.4L/90-30	News Release as sent by the Executive Committee
IEEE P802.4L/90-31	Minutes of the September 1990 meeting (General Motors, Oshawa, Ont, Canada)
IEEE P802.4L/90-32	Epilogue (Vic Hayes, Chairman IEEE P802.4L)

NCR Systems Engineering B.V.

SUBMISSION TO IEEE 802.4L

THROUGH-the-air Token Bus Physical Layer

COMMUNICATIONS OVER AN INDOOR RADIO CHANNEL

BRUCE TUCH

November 1988

NCR SYSTEMS ENGINEERING

ZADELSTEDE 1-10

3431 JL NIEUWEGEIN-HOLLAND

(0)3402-76468

5.) A Direct Sequence "Spread Spectrum" Modulation Technique has been used which satisfies current FCC part 15.126 rules.
6.) These calculations and measurements (of our prototype) have been done with only one transmit/receiver pair. To increase the Coverage Area a fixed array of antenna's, coupled to a head-end can be implemented. It is noted that **this is not mandatory** if the wanted coverage area is small enough. (Running a system without a head-end gives the advantage of only one frequency for TX and RX).

7.) Technical Note:

· The reference BER of the radio channel used in system calculations is 10^{-4} which with FEC Coding is improved to the 10^{-8} 802.4 level. The wanted Outage Probability is achieved with the help of two antennas (selection diversity) and Multipath (two resolvable paths) diversity.

B. Coverage Area is assumed circular with random placement of terminals.

C. A BCH (31,21) FEC code has been assumed in the calculations.

2. ROBUST MODULATION METHOD IN SLOW RAYLEIGH FADING CHANNEL

This report deals with the communication between two terminals in a "Hostile Indoor Radio Wave Channel". It is assumed that further expansion of the coverage area will be done by an array of antenna's connected to a head-end, if needed. Also adjacent channel suppression, from other LAN's will be accomplished by using other frequency bands within the FCC rules and regulations of part 15.

2.1 THE TIME VARYING CHANNEL AND DIFFERENTIAL MODULATION

An important parameter in the LAN Protocol is the needed preamble length. This must be as short as possible for efficient communications. Within our radio channel the following conclusions can immediately be drawn:

1. A coherent form of demodulation, which requires an oscillator reference is not optimal in our situation.

a. Phase acquisition, which requires averaging over many bit times, would need large preamble lengths compared to non- coherent techniques.

b. Path Diversity, which is inherent in the Spread- Spectrum Modulation [see section 3] can not use a locked phase reference.

From the above it seems reasonable to give attention to the most optimal form of non-coherent modulation technique. Differential Phase Shift Keying is a modulation form which does not require a phase reference. It also has the advantage of being spectrally efficient, something which is quite important when our spectral resources are limited by the regulatory agencies. In the next section it is shown that DPSK modulation is robust in our channel environment.

2.2.2 PHASE NOISE DUE TO DOPPLER SHIFT

The discussion of section 2.2.1 has been qualitative in nature to give a "feel" of the channel's coherence time. An analysis of the irreducible error rate (only due to Phase Noise due to motion) has been done [ref. 1b]. It is noted that this analysis assumes a "flat fading" channel, when a fade occurs the total signal energy fades. During a fade the phase change is greater as the antenna moves, which will cause "Random FM" Noise. The following should be considered as worst case since in Direct Sequence Spread-Spectrum Modulation [see section 3] the fading will not effect the total bandwidth. What is important is that the amount of Spreading is not important in terms of using DPSK techniques, since the no spreading case gives us enough margin.

r DPSK the BER of such a channel can be estimated by:

$$P = (1/2)\ (\pi f m T)^2 \qquad [4]$$

Where:

P is the BER caused by Phase Noise due to Doppler Shift (From a large number of Scatters).

fm is the Doppler Shift.

T is the Symbol Period .

Substitution of the system parameters into equation [4] gives:

$$P = 4 \times 10^{-8}$$

Since our Performance Goal is a 10^{-4} BER (Before FEC) it can be seen that Differential Phase Shift Keying is a robust modulation technique with respect to Vehicle Motion.

3.1 THE PROCESSING GAIN SIZE

We are Spreading the Spectrum mainly to satisfy FCC regulations. It is proposed to use the minimum amount of Spreading consistent with the FCC requirements. This will give us more Spectral Efficiency opening up more channels for other collocated LAN systems (There is also a cost factor, as matched filter price is related to Processing Gain). It seems the FCC will allow a minimum **Processing Gain of 10X**. Further contact with the FCC must be made to set this number. (Our prototype has a PG of 31X using a minimum sequence code for spreading).

.2 PATH DIVERSITY

One of our advantages in using Spread Spectrum is the inherent resolution of the system. If an ECHO is produced which is greater than a chip time, it can be resolved by our receiver and used. Post Detection Integration [ref. 2] is assumed in the implementation of the power combining of the echo paths. Following the approximations of Kavehrad [ref. 3], a Path Diversity of 2 can be expected for our system. Together with two antennas this gives us a total of M=4 level Diversity!

SYSTEM PERFORMANCE

The question now arises, using Spread-Spectrum with a QDPSK modulation form, how is the performance in our indoor channel? First the channel parameters must be determined.

4.1 THE CHANNEL

The Indoor Radio Channel complicated by the following factors:

1. High Attenuation Levels due to obstructions. (Mean values, hence Fading is averaged out)

2. Scattering of the wave causing Multipath (Echos) and signal cancelation (Fading).

3. Time variant channel due to antenna and object motion. (As was shown in section 2.1, the channel can be considered time-invariant (quasi-static) for the time intervals of interest).

4.1.3 MULITPATH DELAY SPREAD

An important parameter for echo characterization is the square root of the second moment of the Power Delay Profile, trms. [ref. 4]. The actual shape of the power delay function is not vital in determining performance, and Gaussian Shapes are assumed quite often in the literature. Measurements performed by NCR [ref. 6] in our indoor environments show trms values in the 30 ns range. This agrees quite well with measurements done by Saleh & Valenzuela [ref. 8] in an indoor environment (with about 115 meters largest distance dimensions). Due to our Spread Spectrum Modulation technique intersymbol interference can be neglected, assuming a trms of 30 ns and symbol rate of 800 kBaud. (Actually such a system is quite robust in terms of trms variations. Increasing trms can lead to system improvement due to the creation of more Diversity Paths!)
.ing a minimum expected Delay Spread of 100 ns [ref. 3, 8] (this is the time delay in which an Echo is significant) the number of "resolvable paths" is approximated by a discrete model used by Kavehrad [ref. 3].

$$L = \lfloor Tm/Tc \rfloor + 1 \qquad [7]$$

Where:

$\lfloor x \rfloor$	= the largest interger less than x.
L	= the number of resolvable Paths.
Tm	= the Delay Spread of the Channel.
Tc	= the chip time.

For calculation let us assume a processing gain of 15X. With a symbol rate of 800 kBaud (1/T) and the definition of PG = T/Tc we have:

$$Tc = (1/800K)/15 = 83 \text{ ns}$$

Subsitution into [7] gives at least L = 2 resolvable paths.

2 PERFORMANCE CALCULATIONS

ŀ.2.1 SELECTION DIVERSITY

One way that helps to overcome being in a FADE, is to switch the input to another antenna whose response is uncorrelated with the first (this can be achieved by physical separation or different polarization). In other words the antenna which has the best Signal to Noise Ratio at the Demodulator is chosen. The new cumulative distribution, see section 4.1.2, then becomes:

$$P[Y<y] = [1 - \exp(-y)]^M \qquad [7]$$

Where:

M is the Diversity Level.

5. RADIO PROPAGATION BUDGET

Note: Since Spread-Spectrum has no performance effect with White Gaussian Noise, it is not taken into account.

Thermal Noise No(dBm) = -174 + 10LogB (Hz) Using a 1MHz Noise BW		-114 dBm
Noise Figure (Receiver, Switch, Cable and Filter)		8 dB
Receiver Thermal Noise	(a)	-106 dBm
Man-Made Noise (18dB above Thermal Noise)	(b)	- 96 dBm
Receiver's Input Noise Level (Power Sum of (a) and (b)		- 96 dBm
Instantaneous S/N for 10^{-4} BER (No FEC) QDPSK Modulation		14 dB
Detector Margin		2 dB
Required Instantaneous Receive Level	(c)	- 80 dBm
Isotropic Antenna Path Loss . 1 meter)[2] -27.6dB + 20Log F(MHz) Using F=915MHz		- 31.6 dB
Dipole Antenna Gain[3] (RX and TX each 2dB)		+ 4.0 dB
Transmit Power		20 dBm
Power one Meter Away	(d)	- 7.6 dBm
Signal Attenuation at the Receiver's Noise Limit with respect to one meter.	A = [(d) - (c)]	73.4 dB

2 It is assumed that at one meter away the attenuation follows a free space (n=2 or 6 dB/Octave) law. This would be the case when both antenna's can "see" each other.

3 A small antenna has been assumed (low gain) due to ergonomic requirements.

SPREAD SPECTRUM MODULATION

THIS IS A MODULATION TECHNIQUE IN WHICH THE TRANSMITTED BANDWIDTH IS LARGER THAN THE INFORMATION BANDWIDTH.

SPREAD SPECTRUM TECHNIQUES WAS FIRST USED IN MILITARY SYSTEMS DUE TO ITS ANTI-JAM, ANTI-EAVESDROPPING PROPERTIES.

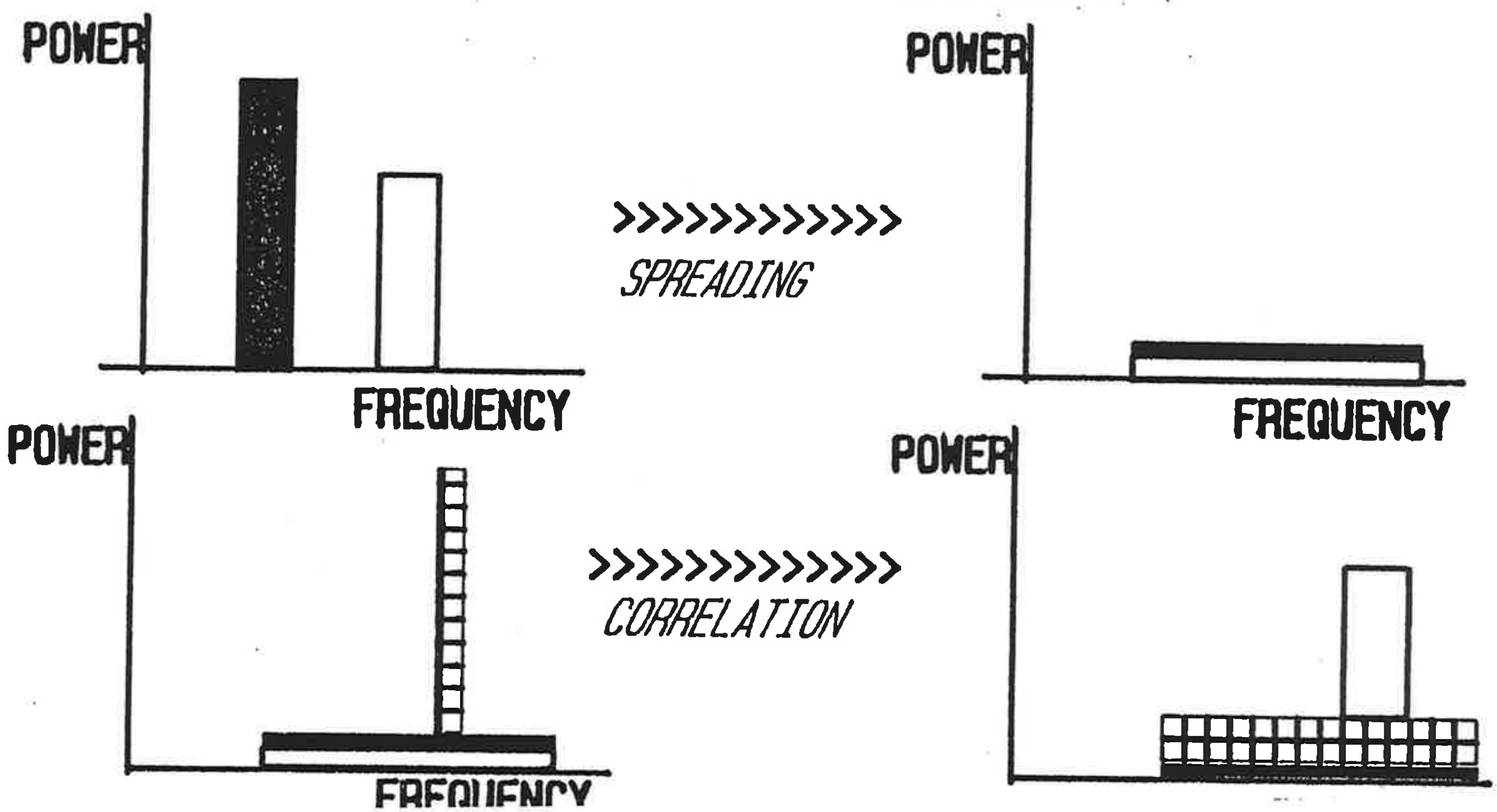
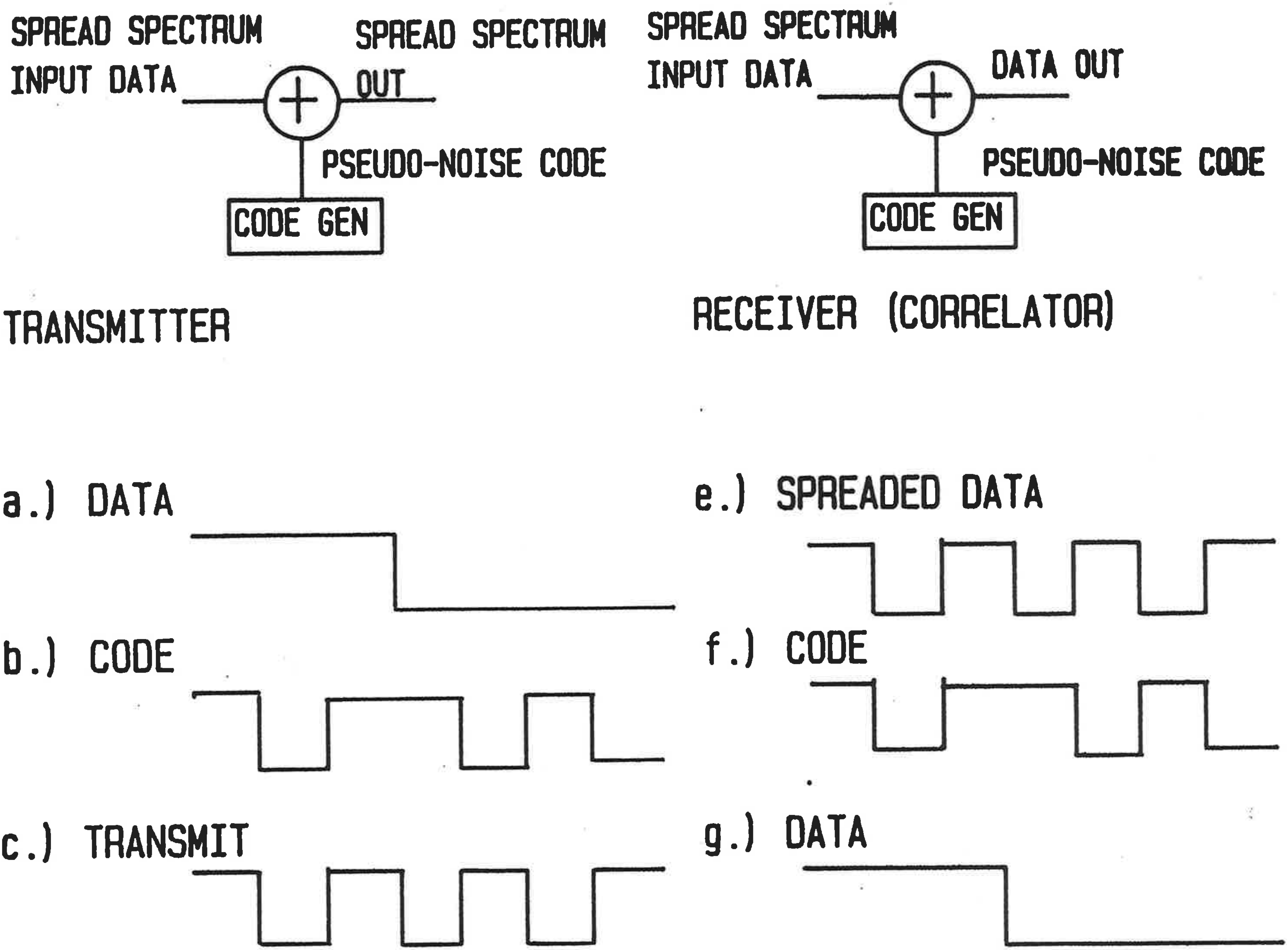


FIG 1


SPREAD SPECTRUM
INPUT DATA
SPREAD SPECTRUM
OUT
PSEUDO-NOISE CODE
CODE GEN
TRANSMITTER
SPREAD SPECTRUM
INPUT DATA
DATA OUT
PSEUDO-NOISE CODE
CODE GEN
RECEIVER (CORRELATOR)
a.) DATA
b.) CODE
c.) TRANSMIT
e.) SPREADED DATA
f.) CODE
g.) DATA

fIG

NCR Systems Engineering B.V.
SUBMISSION TO IEEE 802.4L
<u>Through-The-Air Token Bus Physical Layer</u>

Bruce Tuch
November 5, 1989
NCR SYSTEMS ENGINEERING
ZADELSTEDE 1-10
3431 JL NIEUWEGEIN-HOLLAND

(0)3402-76527

SUBJECT: DQPSK Spread-Spectrum Modulation/Demodulation

1. INTRODUCTION

The mathematical structure of our proposed modulation/demodulation scheme is shown. A general equation for the spectral density of a Direct Sequence Spread Spectrum Signal is derived. The Spectral Density of the 11 chip Barker sequence is calculated and plotted. An expression for the demodulated decision variable is given which can be used for further system performance calculations.

2. DIFFERENTIAL QUADRATURE PHASE MODULATION

2.1. Transmission

Any real valued signal who's frequency content is concentrated around a carrier, f_c, can be expressed:

$$S(t) = a(t) \cdot \mathrm{Cos}(2\pi f_c t + \Omega(t)) \qquad [1]$$

For ease of mathematical manipulation this can be expressed in complex notation:

$$S(t) = \mathrm{RE}[U(t) \cdot \exp(j2\pi f_c t)] \qquad [2]$$

Where:

$U(t) =: a(t)\exp(j\Omega)$ is the Lowpass representation of the signal.

With Digital Phase Modulation the Lowpass function is given by:

$$U(t) = \sum_{n=-\infty}^{\infty} \exp(j\Omega_n) \cdot g(t-nT) \qquad [3]$$

Where:

T is the symbol period of the transmitted information.

n is an index representing the nth transmitted symbol.

g(t) is the baseband modulating pulse with a duration T.

$\exp(j\Omega_n)$ is the Information Vector, I_n. With Quadrature modulation four phase positions are transmitted $\epsilon\{\pi/4, -\pi/4, 3\pi/4, -3\pi/4\}$.

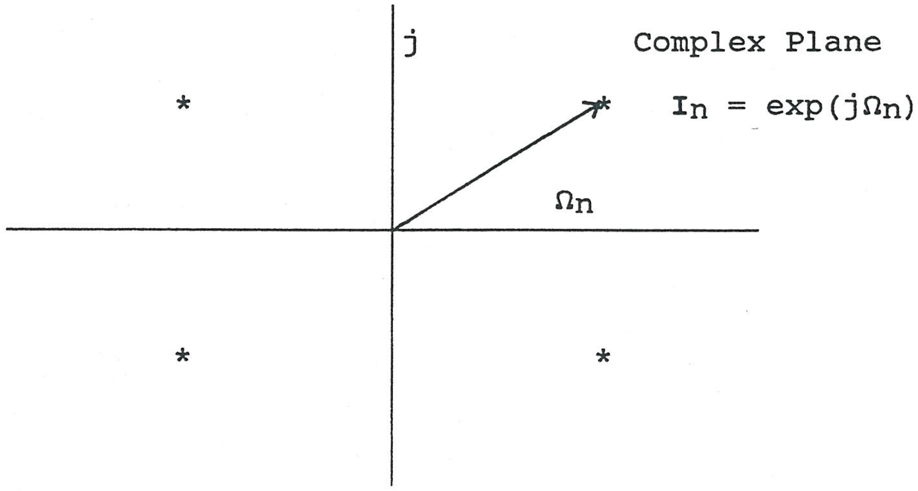


Figure 1 Information Vector in the Complex Plane

To give more insight into the transmitter's functional block diagram, equation [3] can be written as:

$$U(t) = \sum_{n=-\infty}^{\infty} (Ir_n + jIi_n)\cdot g(t-nT) \qquad [4]$$

Where:

$Ir_n \in \{1,-1\}$ is the real axis projection of the Vector $\mathbf{I_n}$.
$Ij_n \in \{1,-1\}$ is the imaginary axis projection of the Vector $\mathbf{I_n}$.

Substitution of [4] into [2] gives the following:

$$S(t) = \sum_{n=-\infty}^{\infty} [Ir_n\cdot g(t-nT)\mathrm{Cos}(2\pi f_c t) - Ii_n\cdot g(t-nT)\mathrm{Sin}(2\pi f_c t)] \qquad [5]$$

Equation [5] lends itself to an easy interpretation. The transmitted phase modulated digital signal can be seen as the sum of two carriers, with a 90 degree offset. The amplitude of each carrier is multiplied by the "wave shaping pulse" g(t), which determines the signal's spectral characteristics, and the Information symbol, Ir_n and $Ii_n \in \{1,-1\}$, which determines the symbol phase state of the transmitted Information Vector. It is noted that due to g(t) the phase of the transmitted signal could change often during a symbol period, for example using Spread-Spectrum Modulation, but this is not dependent upon the information vector and therefore does not determine the symbol phase state.

In order to ease the receiver implementation, within a quasi-stationary channel, Differential Phase Modulation is implemented. In this case the absolute transmitted symbol phase is a function of the previous symbol phase state. The symbol phase encoding is as follows:

$$\Omega_n = d\Omega_n + \Omega_{n-1} \qquad [6]$$

Where:

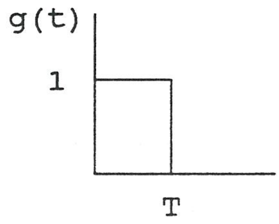
$d\Omega_n$ is the differential symbol phase shift, defined by table I.
Ω_n is the transmitted symbol phase.
Ω_{n-1} is the previous symbol (T delayed) transmitted symbol phase.

The differential symbol phase shift transmitted is derived from the data source following the data mapping of table I.

TABLE I: Data dibit to Differential Phase Mapping

dibit pattern (Left Sent First)	$d\Omega$
0 0	0
0 1	$\pi/2$
1 1	π
1 0	$-\pi/2$

For clarity an example of the various signals for the transmission of Differential Quadrature Phase Modulation is shown. It is noted, in this example, that the "wave shaping pulse" g(t) is given as:



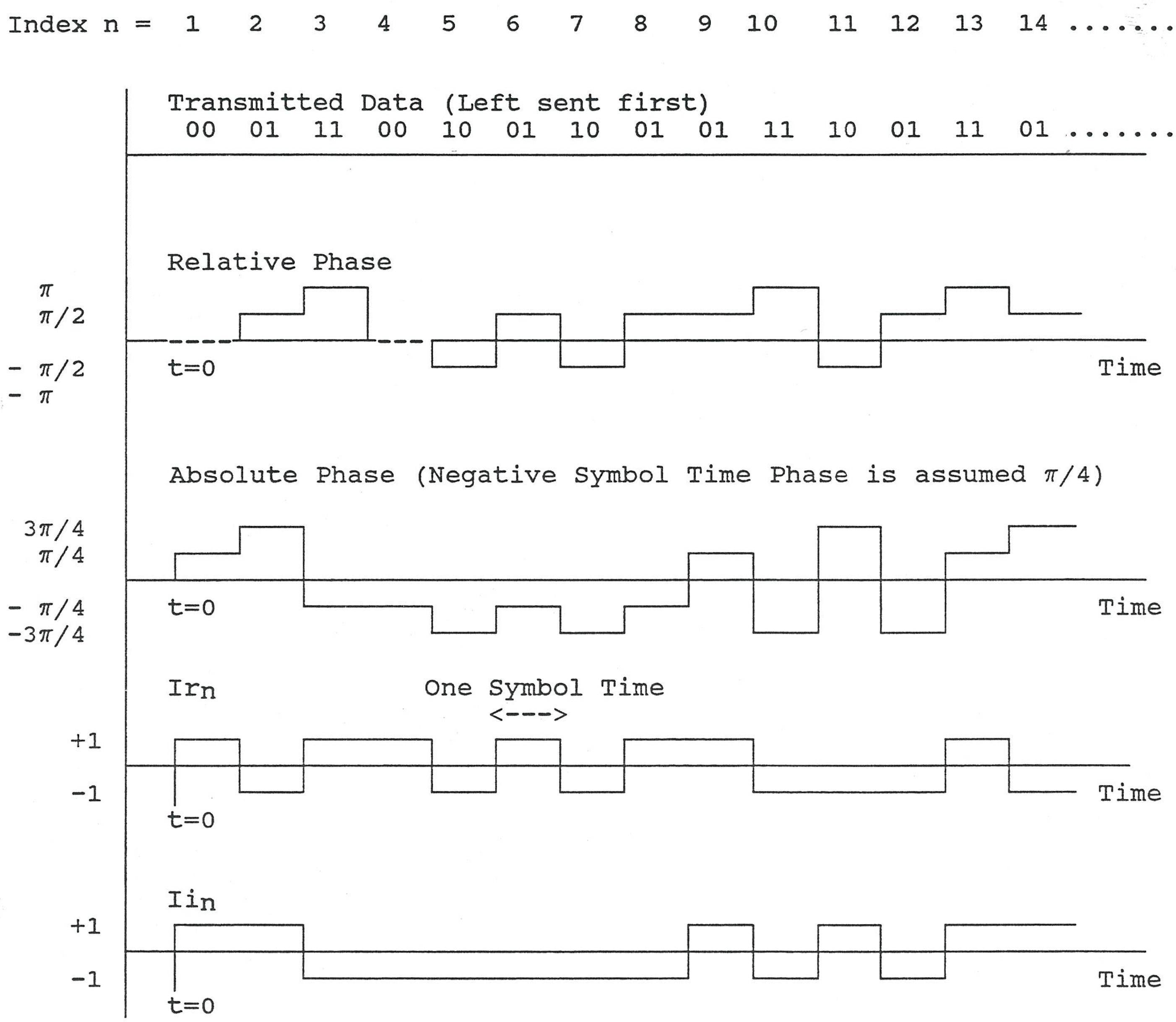


Figure 2 Differential Quadrature Modulation Waveforms

Of course the output signal will be band-limited, which is not shown in figure 2. In figure 3 a functional block diagram can be seen. Note that eventual band limiting of the output signal is shown as a low pass filter F(f) in each real and imaginary axis.

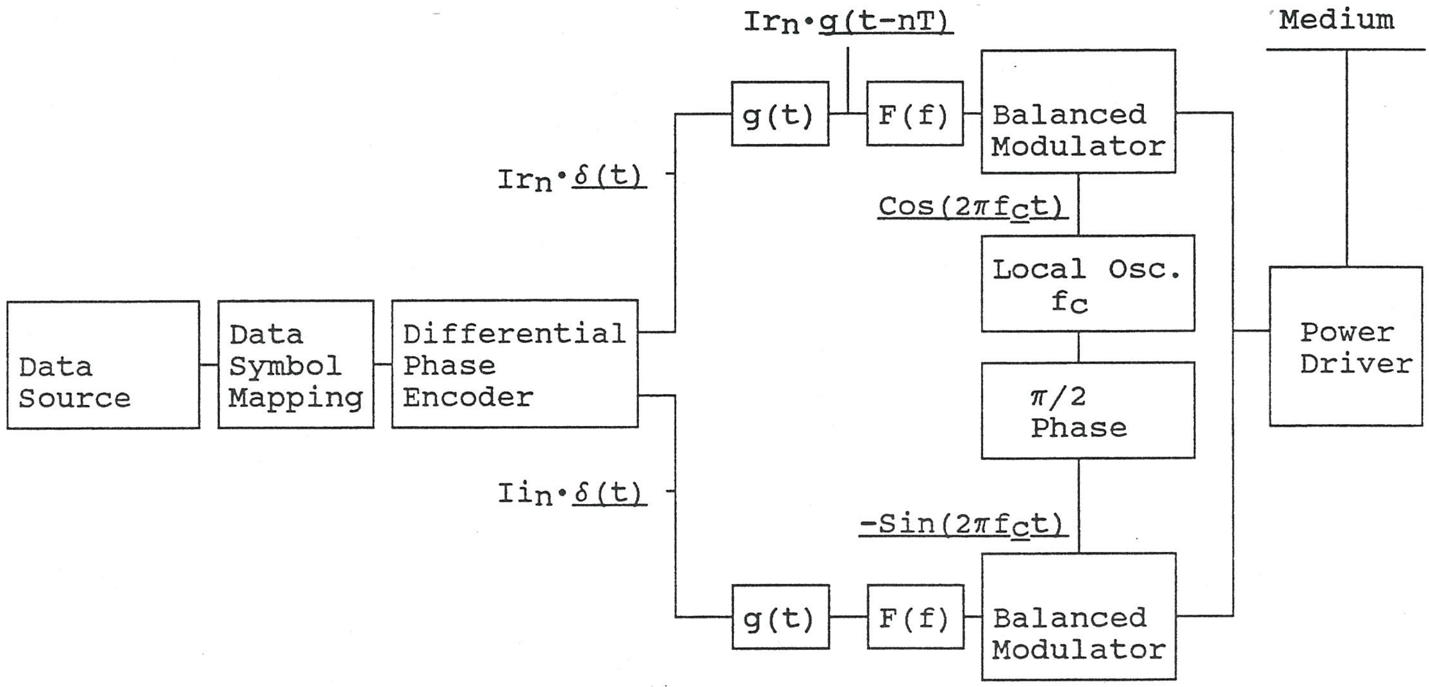


Figure 3 The DQPSK Block Diagram

Two bits from the Data Source, dibit n, are mapped by the Data Symbol Mapping Block according to Table I. The differential phase, $d\Omega_n$, and the previous absolute phase, Ω_{n-1}, are used to calculate the present transmitted phase, Ω_n, by the Differential Phase Encoder following equation [6]. The output of the Differential Phase Encoder are two dirac impulses, $\delta(t)$, weighted by the real and imaginary components of the information Vector I_n.

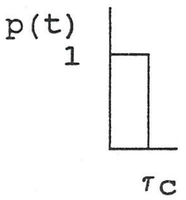
2.1.1. Direct Sequence Spread Spectrum

At this point we have DQPSK transmission in which the spectral bandwidth is determined by g(t) and the spectral limiting function F(f). In the example of figure 2, g(t) is a unit pulse with a duration of one dibit period T. This would be the case with "normal" DQPSK modulation. In Direct Sequence Spread Spectrum, DSSS, the transmitted bandwidth is larger than the "information bandwidth". This is accomplished by having transitions in the output variable, the phase in this case, which are shorter in duration than the information symbol time. This can be accomplished, without any changes with the block diagram of figure 3, by defining g(t) as the spread spectrum sequence:

$$g(t) = \sum_{k=1}^{N} X_k \cdot p(t-k\tau_c) \qquad [7]$$

Where:

g(t) is the spread spectrum sequence.
p(t) is a "chip" pulse:



τ_c is the chip duration such that $N = T/\tau_c$ is an integer.
$X \in \{1,-1\}$ and is the kth chip's coefficient.

The spread spectrum sequence, g(t), is defined by the coefficient vector X_k. In our case the N = 11 chip Barker sequence is used:

Index k =	1	2	3	4	5	6	7	8	9	10	11
X_k =	[1	-1	1	1	-1	1	1	1	-1	-1	-1]

The spectrum spreading is characterized by the parameter N. This gives the number of chips within one symbol period. This is also how much the spectrum bandwidth is increased relative to the no spreading case N=1 (chip = symbol).

2.1.2. Output Spectral Density

The output spectral density of a process with a lowpass representation, U(t), is given by the Fourier Transformation of the autocorrelation function:

$$\phi uu(\tau) \longleftrightarrow \Phi uu(f) = \int_{-\infty}^{\infty} \phi uu(\tau)\cdot\exp(-j2\pi\tau)d\tau \qquad [8]$$

Where:

$\phi uu(\tau) = \frac{1}{2}E[U(t)\cdot U^*(t)]$ is the complex autocorrelation of the process U(t) and * signifies the complex conjugate.

U(t) is a cyclostationary process, its statistics are periodic in T, and therefore, as shown by Proakis [ref 1], the average autocorrelation function, $\phi uu(\tau)$, of the process U(t) given by equation [3] is:

$$\phi uu(\tau) = 1/T\cdot\sum_{m=-\infty}^{\infty} \phi ss(m)\cdot\phi gg(\tau-mT) \qquad [9]$$

Where:

$\phi ss(m)$ is the complex autocorrelation of the Information Symbol Vector, $\mathbf{I}_n$, defined as:

$$\phi ss(m) = \frac{1}{2}E[\mathbf{I}^*_n\cdot\mathbf{I}_{n+m}] \qquad [10]$$

(* signifies complex conjugation)

$\phi gg(\tau)$ is the time autocorrelation of the function g(t).

Substitution of [9] into [8] gives the output spectral density as:

$$\Phi uu(f) = 1/T\cdot|G(f)|^2\cdot\Phi ss(f) \qquad [11]$$

Where:

$|G(f)|$ is the magnitude of the Fourier Transform of g(t).
$\Phi ss(f)$ is the Spectral Density of the Information Vector given by:

$$\Phi ss(f) = \sum_{m=-\infty}^{\infty} \phi ss(m)\cdot\exp(-j2\pi f\cdot mT) \qquad [12]$$

Expressing In in real and imaginary components and substitution into [10] gives:

$$\phi ss(m) = \tfrac{1}{2}[\phi rr(m) + \phi ii(m) - j\phi ri(m) + j\phi ir(m)] \qquad [13]$$

Since the data source is random, as are the dibits which determine the sign of the real and imaginary Information Vector components, these crosscorrelation products are zero:

$$\phi ri(m) = 0 \quad \text{for all } m$$
$$\phi ir(m) = 0$$

Also real and imaginary axis autocorrelation functions are equal, since the statistics are the same for each axis:

$$\phi rr(m) = \phi ii(m) \text{ for all } m$$

Since the data sources random binary bits, previous and present dibits are random. This means that the autocorrelation is:

$$\phi rr(m) = \delta_{mo}$$

Where δ_{jk} is the Kronecker delta such that:

$$\delta_{jk} = \begin{array}{ll} 1 & j = k \\ 0 & j \neq k \end{array}$$

Substitution of the correlation statistics given into [13] gives:

$$\phi ss(m) = \phi rr(m) = \delta_{mo} \qquad [14]$$

It is noted that due to [14] the autocorrelation of the output given in [9], $\phi uu(\tau)$, is only a function of the real (or imaginary) component of the Information Vector and the pulse g(t).

Substitution of [14] into [12] gives the Spectral Density of the Information Vector as:

$$\Phi ss(f) = 1$$

From [11] and $\Phi ss(f) = 1$ the Spectral Density of the Output Process is found:

$$\Phi uu(f) = 1/T \cdot |G(f)|^2 \qquad [15]$$

The next step is to find the function $|G(f)|$ which uniquely determines the Output Spectral Density. The function g(t) given in [7] can be written in another form:

$$g(t) = \sum_{k=1}^{N} X_k \cdot p(t-k\tau_C) = p(t) * \sum_{k=1}^{N} X_k \cdot \delta(t-k\tau_C) \qquad [16]$$

Where:

$*$ denotes the convolution operation:

$$z(t) * h(t) = \int_{-\infty}^{\infty} h(\tau) \cdot z(t-\tau)\, d\tau$$

Looking at [16], imposing the relationship between time domain convolution and frequency domain multiplication, $G|(f)|^2$ can be expressed as the multiplication of two terms:

$$|G(f)|^2 = |P(f)|^2 \cdot |X(f)|^2 \qquad [17]$$

Where:

$|P(f)|$ is the magnitude of the Fourier Transformation of the chip pulse p(t).

$|X(f)|$ is the magnitude of the Fourier Transformation of x(t) which is given by:

$$x(t) = \sum_{k=1}^{N} X_k \cdot \delta(t-k\tau_C) \qquad [18]$$

Noting from the Fourier Transformation that the relationship holds:

$$x(-t) \longleftrightarrow X^*(f) \qquad [20]$$

Since $|X(f)|^2 = X(f) \cdot X^*(f)$, using [20] we have:

$$x(t) * x(-t) \longleftrightarrow X|(f)|^2 \qquad [21]$$

Substitution of [18] into [21] gives:

$$x(t) * x(-t) = \sum_{k=1}^{K} \sum_{i=1}^{K} X_k \cdot X_i \cdot \int_{-\infty}^{\infty} \delta(t+\alpha - k\tau_C) \cdot \delta(\alpha - i\tau_C)\, d\alpha \qquad [22]$$

Using the fact that $\delta(\alpha - i\tau_C)$ has unity weight only for $\alpha = i\tau_C$, else zero, equation [22] can be simplified:

$$x(t) * x(-t) = \sum_{k=1}^{K} \sum_{i=1}^{K} X_k \cdot X_i \cdot \delta(t-(k-i)\tau_C) \qquad [23]$$

Taking the Fourier Transformation of [23] gives the Spectral Density of the Spread-Spectrum Code Sequence:

$$|X(f)|^2 = \sum_{k=1}^{K} \sum_{i=1}^{K} X_k \cdot X_i \cdot \exp(-j2\pi f(k-i)\tau_C) \qquad [24]$$

The first term in the double addition of [24], k=i=1, is a DC component which can be separated, combining negative and positive exponential terms gives:

$$|X(f)|^2 = \sum_{k=1}^{K} X^2{}_k + 2 \cdot \sum_{k=2}^{K} \sum_{i=1}^{k-1} X_k \cdot X_i \cdot \mathrm{Cos}(4\pi f(k-i)\tau_C) \qquad [25]$$

With the 11 chip Barker Code, X_k = [1 -1 1 1 -1 1 1 1 -1 -1 -1], substitution into [25] gives a further simplification:

$$|X(f)|^2 = 11 - 2 \cdot \sum_{k=1}^{5} \mathrm{Cos}(4\pi f \times k\tau_C) \qquad [26]$$

Using the fact the chip pulse, given in [7], p(t) <----> P(f) is given by:

$$|P(f)| = T \cdot \mathrm{Sinc}(\pi f \tau_C)$$

The Output Spectral Density, from [15], is:

$$\Phi uu(f) = T \cdot \mathrm{Sinc}^2(\pi f \tau_C) \cdot [11 - 2 \cdot \sum_{k=1}^{5} \mathrm{Cos}(4\pi f \cdot k\tau_C)] \qquad [27]$$

In figure 4 a plot of the Output Spectral Density is shown. It is noted that the Spectral "Side-Lobes" of the Sinc function are quite significant. Spectral side-lobe suppression will be necessary using the Low Pass filters F(f) shown in the figure 3. It is interesting to note that the spectrum is "smooth". This is due to the Barker sequence and is coupled with its unity bounded non-periodic and odd-periodic autocorrelation sidelobe properties.

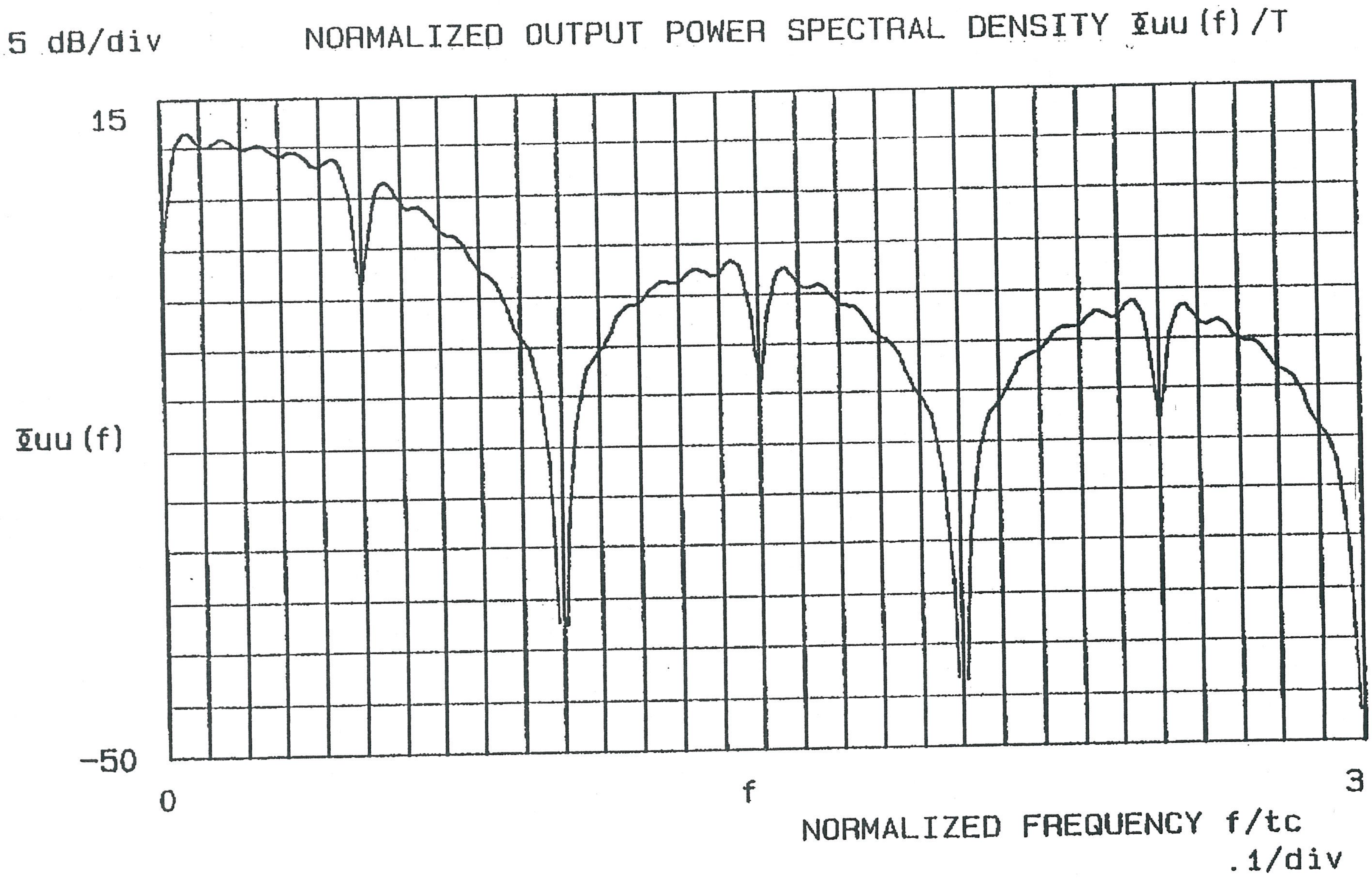

5 dB/div
NORMALIZED OUTPUT POWER SPECTRAL DENSITY Φuu(f)/T
15
Φuu(f)
-50
0
f
3
NORMALIZED FREQUENCY f/tc
.1/div

FIGURE 4

2.2. Reception

The transmitted signal has been given in [3] and is repeated here:

$$U(t) = \sum_{n=-\infty}^{\infty} I_n \cdot g(t-nT) \qquad [28]$$

The time-invariant channel (between two symbol intervals) is modeled as:

$$h(\tau) = \sum_{i=1}^{L} \beta_i \cdot \delta(\tau-\tau_i) \cdot \exp(-j\alpha_i) \qquad [29]$$

Where:

β_i is the ith path gain, with probability density function, pdf, which is Rayleigh distributed.
L is the number of paths.
α_i is the ith path phase, with a unit pdf within $[0,2\pi]$.

The received signal is the convolution of [28] with [29] giving:

$$r(t) = \sum_{n=-\infty}^{\infty} \sum_{i=1}^{L} I_n \cdot \beta_i \cdot \exp(-j\alpha_i) \cdot g(t-nT-\tau_i) \qquad [30]$$

Let the receiver consist of a "Matched Filter" and a symbol delay T. The impulse response of the Matched Filter is given by:

$$hfilter(t) = g(T-t) \qquad [31]$$

From the received signal r(t) given in [30] the present, n=0, and previous symbol, n=-1, shall be used in the calculations for intersymbol interference effects. (It is assumed that the delay of the channel is much less than two symbol periods and multiple symbol interference can be neglected).

For convenience the first path, which is chosen for detection by the receiver, of the channel is set with $\tau_c = 0$. The complex signal under investigation is:

$$r'(t) = \sum_{i=1}^{L} I_0 \cdot \beta_i \cdot \exp(-j\alpha_i) \cdot g(t-\tau_i) + \sum_{i=1}^{L} I_{-1} \cdot \beta_i \cdot \exp(-j\alpha_i) \cdot g(t+T-\tau_i) \qquad [32]$$

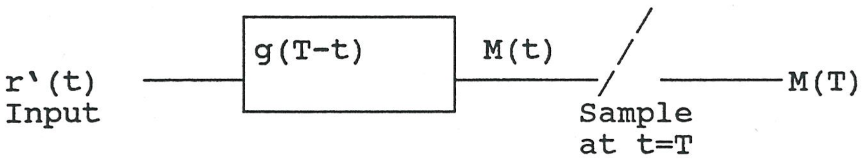


Figure 5 Matched Filter (Correlation) Receiver

In order to calculate M(t) it is noted that the receiver forms a linear system and superposition applies. Therefore each summation term in equation [32] can be applied to the system separately, adding each resulting output to give the total response. Applying the terms of r'(t) containing the I_o complex Information Vector gives:

$$M_o(t) = g(T-t) * g(t)\cdot I_o\cdot\beta_1\cdot\exp(-j\alpha_1) + \sum_{i=2}^{L} I_o\cdot\beta_i\cdot\exp(-j\alpha_i)\cdot g(t-\tau_i) * g(T-t)$$

Taking the value of Mo(t) at the sampling moment T gives:

$$M_o(T) = T\beta_1\cdot\exp(-j\alpha_1) + \sum_{k=2}^{L} I_o\cdot\beta_i\cdot\exp(-j\alpha_i)\cdot\int_{-\infty}^{\infty} g(\alpha-\tau_i)\cdot g(\alpha)\,d\alpha \qquad [33]$$

Where the first chosen path, with zero delay, has been separated from the expression. The integral in [33] is the time autocorrelation of the spread spectrum signal. From the expression for g(t) in [7] for $\tau_1 = 0$ the autocorrelation equals T which is the factor of the first term.
Since g(t) has a duration of T, as shown in figure, 6 the autocorrelation function can be expressed as:

$$R''(\tau) = \int_{-\infty}^{\infty} g(\alpha-\tau)\cdot g(\alpha)\,d\alpha = \int_{\tau}^{T} g(\alpha-\tau)\cdot g(\alpha)\,d\alpha \qquad [34]$$

Only the cross areas contribute to the integral.

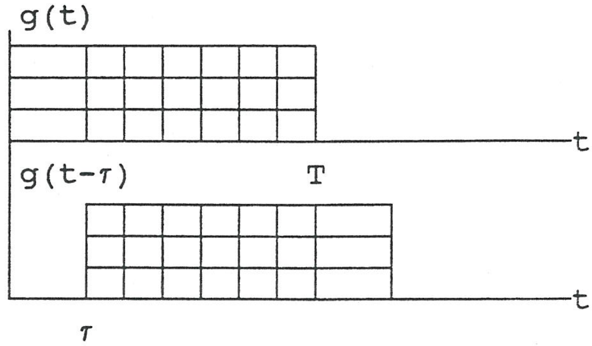


Figure 6 Spread-Spectrum Sequence Autocorrelation

$M_0(T)$ can now be expressed as:

$$M_0(T) = T \cdot \beta_1 \cdot I_0 \cdot \exp(-j\alpha_1) + I_0 \cdot \sum_{i=2}^{L} \beta_i \cdot \exp(-j\alpha_i) \cdot R''(\tau_i) \qquad [35]$$

Using the same arguments as given above, the sampled output due to the terms in r'(t) containing the Complex Information Vector I_{-1} gives:

$$M_1(T) = I_{-1} \cdot \sum_{i=1}^{L} \beta_i \cdot R(\tau_i) \qquad [36]$$

Where:

$$R(\tau) = \int_{-\infty}^{\infty} g(\alpha+T-\tau) \cdot g(\alpha)\,d\alpha = \int_{0}^{\tau} g(\alpha-\tau) \cdot g(\alpha)\,d\alpha$$

Summation of [35] and [36] gives the total output as sample moment T:

$$M(T) = I_0 \cdot [T \cdot \beta_1 \cdot \exp(-j\alpha_1) + \sum_{i=2}^{L} \beta_i \cdot \exp(-j\alpha_i) \cdot R''(\tau_i)] + I_{-1} \cdot \sum_{i=2}^{L} \beta_i \cdot \exp(-j\alpha_i) \cdot R(\tau_i) \qquad [37]$$

Due to the choice of the 11 chip Barker sequence and its unity bounded odd-periodic correlation function, the autocorrelation terms in [37] are bounded by:

$$R''(\tau) = R(\tau) < \tau_c \qquad [38]$$

Which is the area under one chip pulse.

Substitution of the upper bound of [38] into [37] gives:

$$M(T) = \mathbf{I}_o \cdot [T \cdot \beta_1 \cdot \exp(-j\alpha_1) + \mathbf{B} \cdot \tau_c] + \mathbf{I}_{-1} \cdot \mathbf{B} \cdot \tau_c \qquad [38]$$

Where:

$$\mathbf{B} = |\mathbf{B}| \exp(-j\Phi) = \sum_{i=2}^{L} \beta_i \cdot \exp(-j\alpha_i)$$

It is noted that the second term of equation [38] is the intersymbol interference term of the previous ($\mathbf{I}_{-1}$) information symbol with the present ($\mathbf{I}_o$). One can see how Spread-Spectrum Modulation is robust with respect to Multipath Delay. The received path's strength is multiplied by a factor T while the interference terms by τ_c. The ratio, $T/\tau = N$, being the "Processing Gain" of the system.

A symbol delay, T, is now added at the output of the matched filter receiver. At the present sample moment M(T) the previous sample output, M(0), is available. Note that M(0) has is the same as [38] except that the Information Vectors are delayed one symbol:

$$M(0) = \mathbf{I}_{-1} \cdot [T \cdot \beta_1 \cdot \exp(-j\alpha_1) + \mathbf{B} \cdot \tau_c] + \mathbf{I}_{-2} \cdot \mathbf{B} \cdot \tau_c \qquad [39]$$

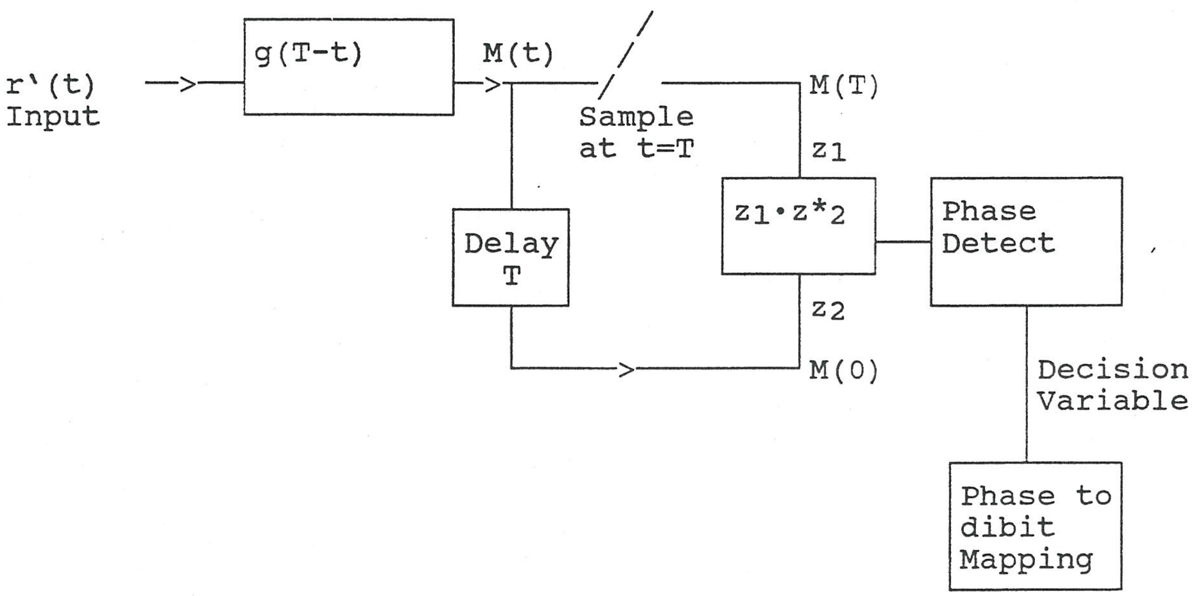


Figure 7 Differential Phase Demodulation Receiver

Performing the complex conjugate multiplication shown in Figure 7 gives:

$$M(T) \cdot M^*(0) = (\beta_1 \cdot T)^2 \cdot \exp[j(\Omega_0 - \Omega_{-1})] + \text{Intersymbol Interference} \quad [40]$$

Taking the phase of [40] gives the decision variable, the differential phase output:

$$d\Omega_0 = \Omega_0 - \Omega_{-1} + \text{Intersymbol Phase Interference}$$

Which gives the output variable needed in order to determine, following the $d\Omega$ to dibit mapping of Table I, the transmitted data dibit.

3. REFERENCES

1. Digital Communications, McGraw-Hill ISBN 0-07-Y66490-0
   John G. Proakis

2. Power Spectrum, Internal NCR Report
   Kiwi Smit

IEEE p802.4L

Through-the Air Physical Media, Radio

Statistic Analysis of Oshawa Measurements

L. van der Jagt, KII

Attached is a submission to the IEEE p802.4L Task group.[1]

Please note that the attached diagrams are excerpts from the full set of analysis results.

1These charts were submitted to the IEEE p802.4L meeting held 5-7 November 1989 in Fort Lauderdale, FL. The temporary number was F.4L/6L.

Knowledge Implementations, Inc.
Fiber Optic and Communications Systems Engineering/Marketing

F.4L/6

Summary of Attached Analysis

The attached drawings and calculations represent a more in depth analysis of the data obtained at the General Motors Car Body Assembly Plant in Oshawa, Ontario. The original analysis which consisted primarily of cataloging the data, and performing correlation on the data to obtain basic delay spread information has been enhanced with the addition of the following calculations:

◇ Instantaneous and average power

◇ Large scale fading characteristics

◇ Delay spread characteristics

◇ Temporal variance of the channel

◇ Analysis of noise samples

◇ Analysis of the impact of impulse noise on barker sequence correlation peaks

The data contained herein is basically self explanatory. It was produced using a combination of custom generated C language programs and the MATHEMATICA software package.

The large scale fading calculations were done by taking the I and Q voltage samples which were captured, scaling them according to the oscilloscope settings when they were captured, and forming a complex number I+iQ. This number was then multiplied by its complex conjugate to produce a value for instantaneous power. The instantaneous power was then summed over a number of samples and divided by the number of samples to provide an indication of average power. The average power from the measurments taken at a 10 wavelength distance at location J47 was used as a base for the generation of path loss numbers for the various locations. This data was also adjusted based on the level of transmitter modulation to a base level of 50 mv modulation. A least squares fit was done on the path loss versus distance data to provide a value for n. The square of the deviations of the actual values from the estimated values were then calculated to produce a deviation value. These were respectively, n = 2.204 and sigma = 9.53 dB. These values are in good agreement with values previously published.

In order to observe the temporal variance of the channel, the correlated data for M-sequences from each location was taken. Again, I and Q were combined through multipilcation of I+iQ by its complex conjugate and the results plotted. Successive peaks were located and displayed as 3 dimensional surface plots. Of particular interest were the plots of MSEQ1 through MSEQ6 which illustrate not only time variance but also the variance of the channel with small (approximately 1-2") micromovements. In

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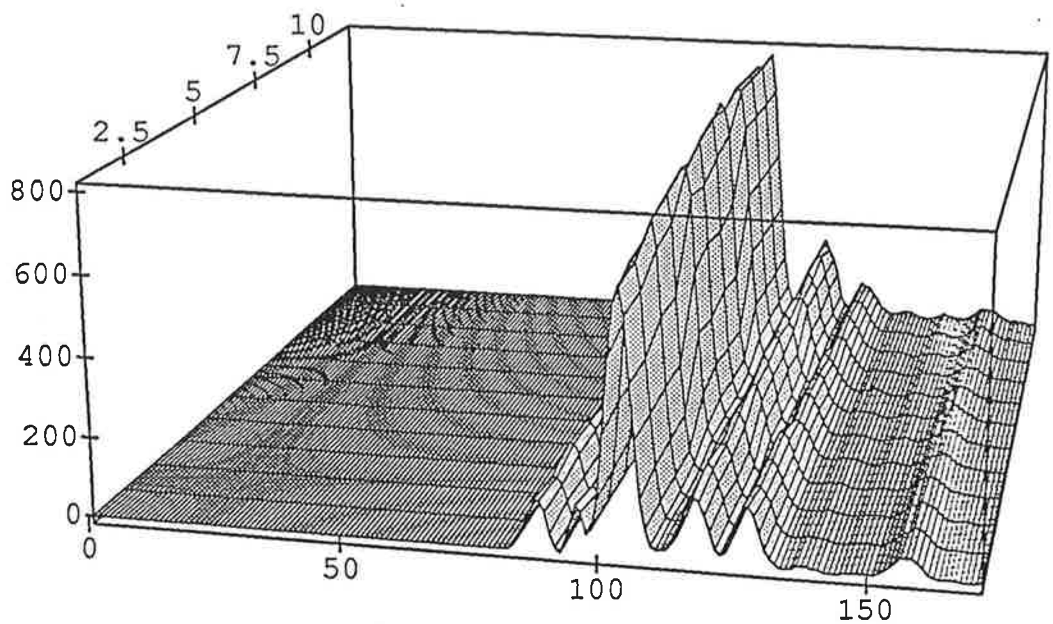
general the duration of time which is represented in the plots is about 180 µs.

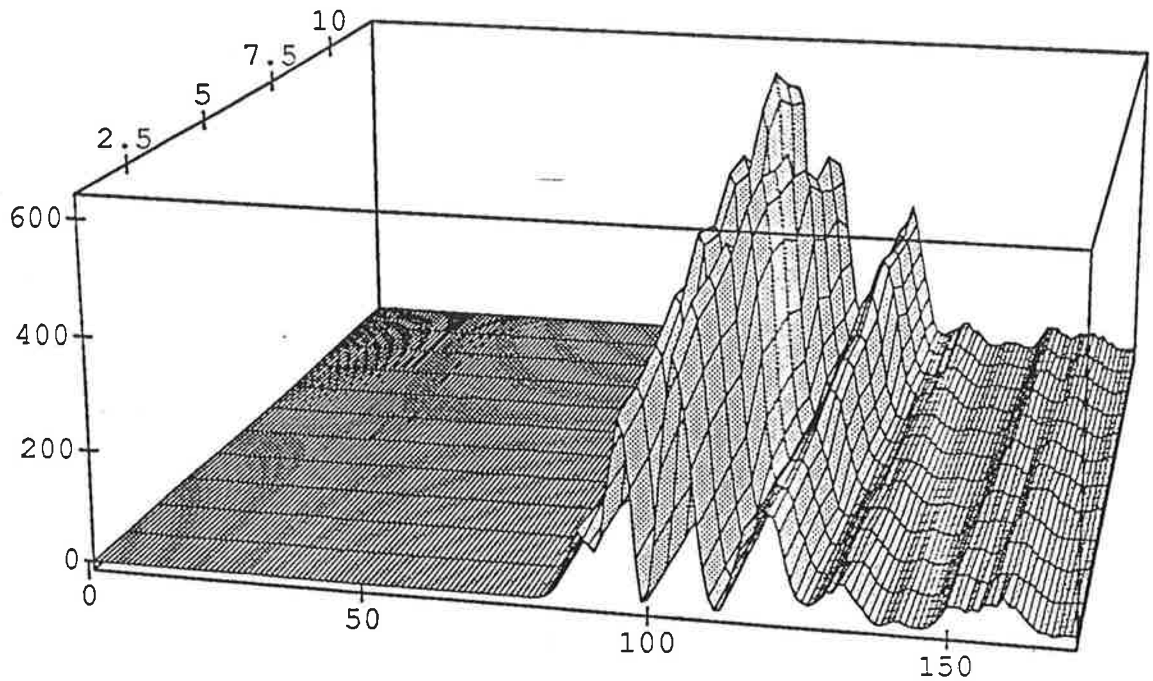
The delay spread was estimated by looking at each temporal variance plot and assigning a value of delay to it. These values were then processed to determine mean and standard deviation. In this calculation the mean of the delay spread was estimated to be 286.84 ns and the Standard deviation of this value was 99.72 ns. As has been previously reported there did not seem to be a correlation between delay spread and distance of transmission.

Finally, noise samples were taken and the amount of noise present in bins of 2 µs and 20 µs was examined. This examination indicated that impulse noise could cause the amount of power present in a bin to vary substantially from the mean value. In the sample BARK17, on an arbitrary scale the received power was 800 per bin under non impulse situations, and this increased to 1600 under impulse situations. Also provided is data regarding the impact of this impulse noise on the correlation peaks derived from the sample.

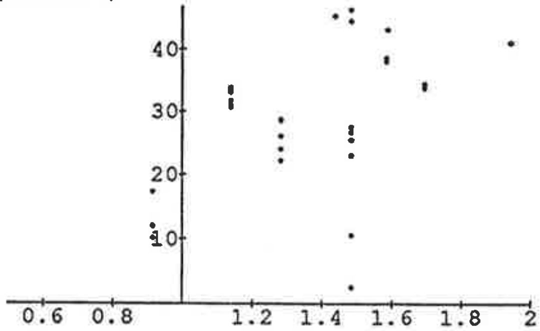
From a noise point of view the most serious problem seems to be with intentional radiators operating within the band we are dealing with. Although the magnitude of the noise from intentional radiators will vary from site to site, it appears from our collection of data at multiple sites that it is reasonable to expect that RADIOLAN's will be expected to operate in many situations in an environment characterized by negative signal to interference ratios and that existance of adequate jamming margin with the technique chosen to implement the system will be essential. This is true in both the 915 MHz and the 2.45 GHz band. In the higher frequency band the presence of substantial interference from microwave ovens is to be expected. In the lower frequency band, high levels of interference can be expected from emergency services radio, theft detection devices, and older microwave ovens.



Y56 Temporal Variance MSEQ1



Y56 Temporal Variance MSEQ3



In[20]:=

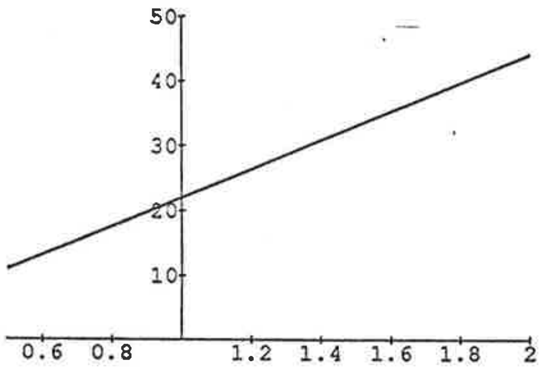
```
Fit[data1,{x},x]
```

Out[20]=

```
22.0439 x
```

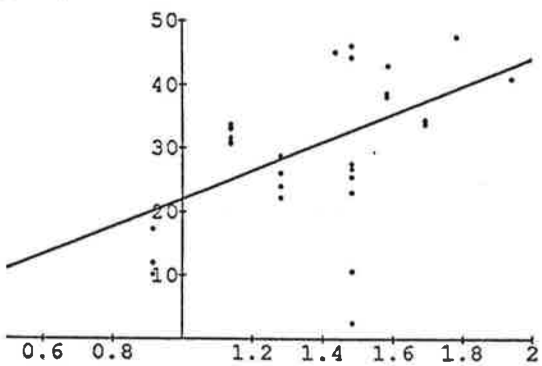
In[21]:=

```
Plot[%,{x,.5,2},PlotRange ->{{.5,2},{0,50}}]
```



In[22]:=

```
Show[%,%%%]
```




```
For[i=0;sigma=0,i<Length[data],sigma=sigma+
((data1[[i]][[2]]-First[Out[20]] data1[[i]][[1]])^2),i++]
```

In[28]:=

```
sigma/(Length[data]-2)
```

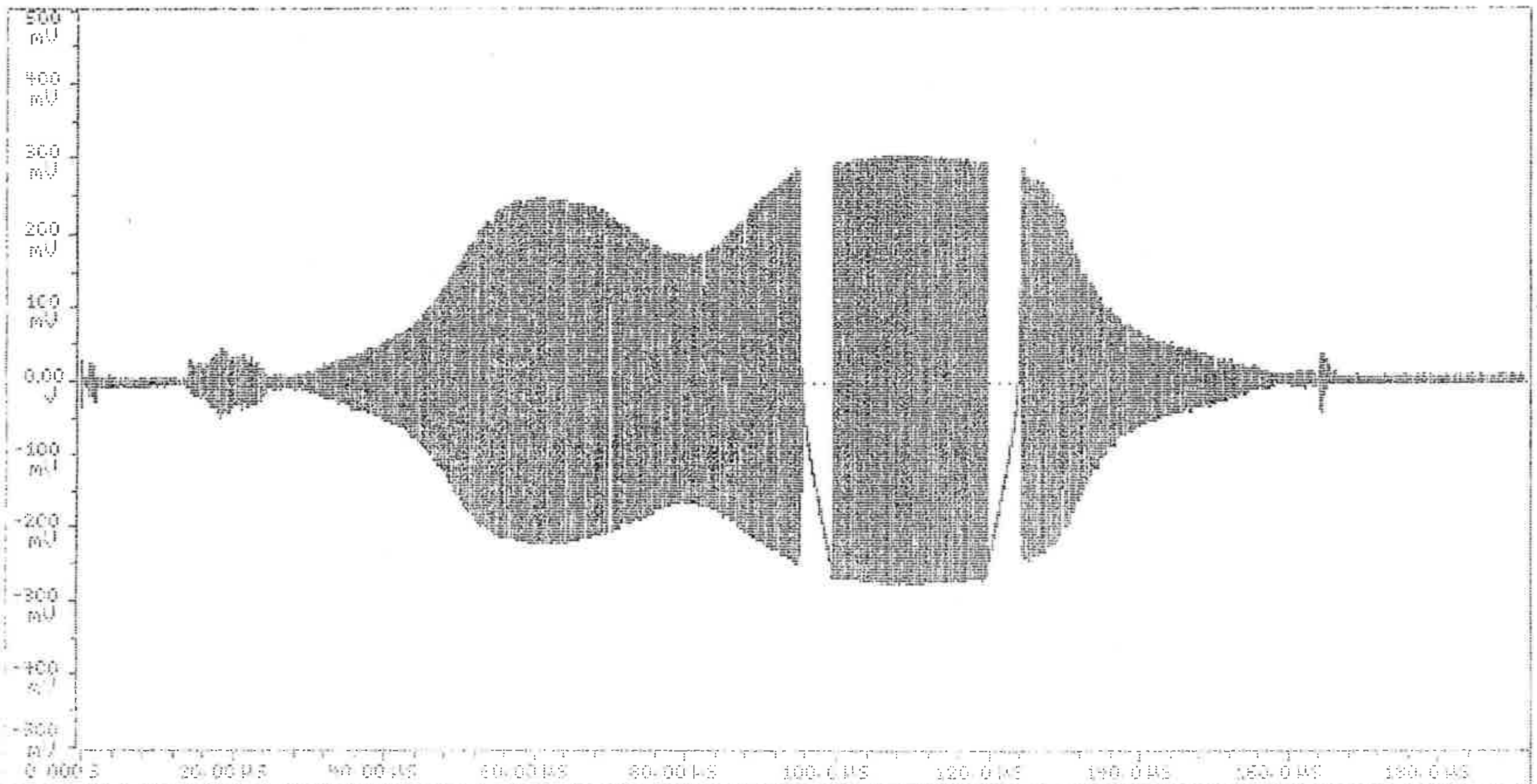
Out[28]=

```
95.3267
```

1 ms/DIV

1 VOLT/DIV.

File LeCroy EASYWAVE 18:34:57

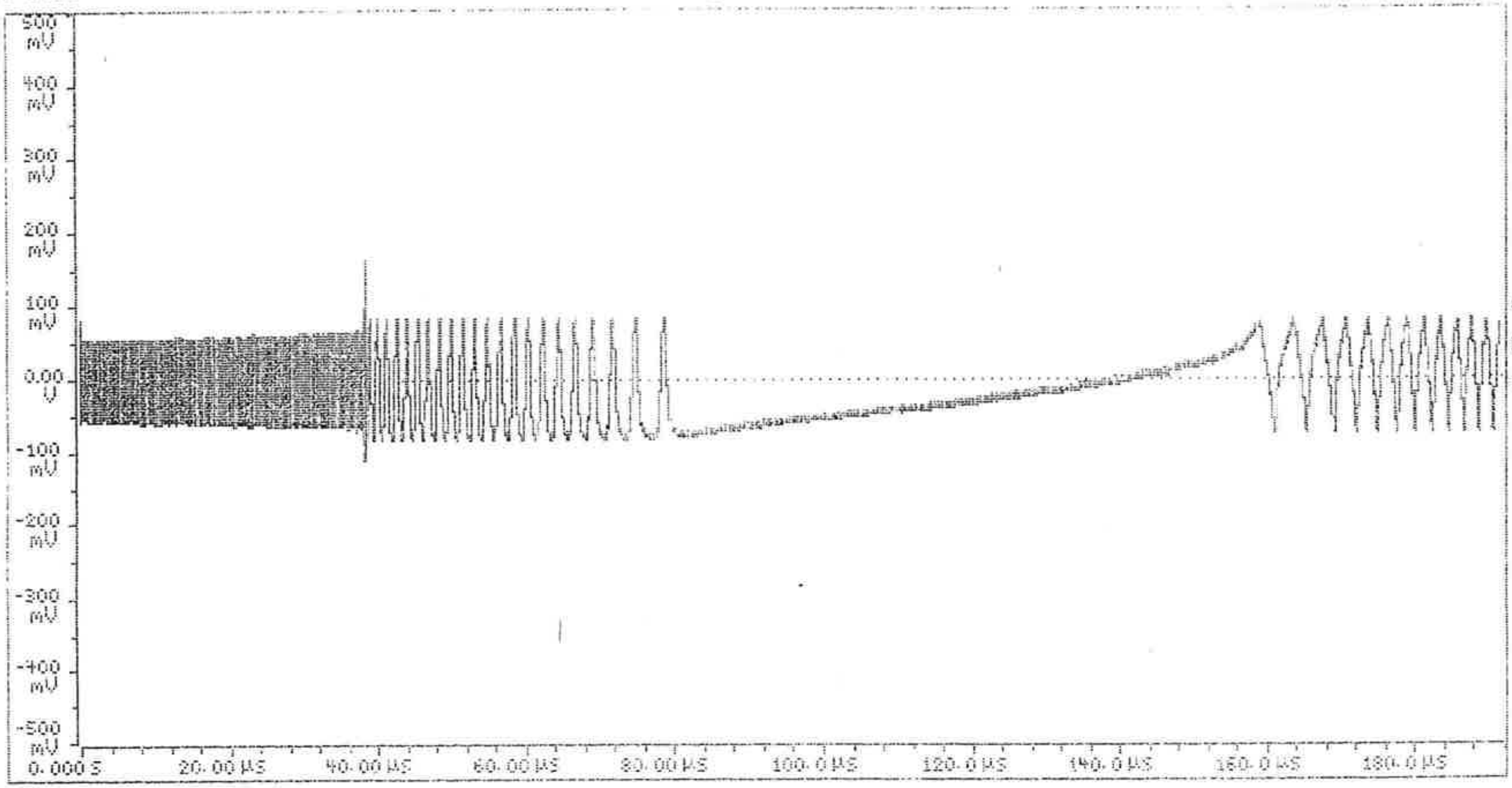


File transfer: Disk to ew | disk to afg and ew | Ew to disk | eW to afg | Afg to disk and ew | afg to ew | More

MICR19.WAV from disk to EW

5ns/pt 1V/DIV

File LeCroy EASYWAVE 18:32:44



File transfer: Disk to ew | dIsk to afg and ew | Ew to disk | eW to afg | Afg to disk and ew | aFg to ew | More

Cursor: 0 (0.000 s), 14 (2.9 mV)

IEEE p802.4L

Through-the Air Physical Media, Radio

Impulse noise effect on 4 level QAM Spread Spectrum Signal

Donald C. Johnson, NCR

Attached is submission to the IEEE p802.4L Task group.[1]

[1]This document was submitted to the IEEE p802.4L meeting held 5-7 November 1989 in Fort Lauderdale, FL. The temporary number was F.4L/7.

IMPULSE NOISE EFFECT ON 4 LEVEL QAM SPREAD SPECTRUM SIGNAL

Donald C. Johnson

1.0 Summary and Purpose

The intent here is to show that the effect of impulse noise on the error probability for 4 level QAM spread spectrum signals can be predicted from the power level of the noise over the spread symbol time. That is, regardless of the pulse shape or energy concentration of the noise, the error probability is dependent principally only on the energy of the noise.

The measurements taken in both the NCR Department Store tests and the GM Oshawa tests show that impulses occur which raise the noise power level considerably above the background noise for durations on the order one or a few chip times. The incidence of occurrence of such impulses is on the order of fractions of a second. At this occurrence rate, each impulse can have an error probability in the order of 0.1 and still achieve a respectable error rate with 1 to 2 Mb/s signaling rates. The example error probability per impulse is thus 0.1 in the following.

It is shown that, in the specific case of an 11 chip spreading code and a typical receiver filter, the signal power to noise power ratio over a symbol time is somewhere between -4.4 dB and -7.4 dB for an impulse probability of error of 0.1. For other spreading code lengths, the values are different but the 3 dB spread is the same. The significance of this is that it makes the analysis of the impulse noise affect on the these data signals considerably easier. It is only necessary to determine the rms voltage envelope value of the impulse over a spreading symbol time in order to predict it's effect within 1.5 dB at most. Once the rms value is known, the effective level of the impulse concerning error production is known within the above spread of +/- 1.5 dB.

The higher S/N ratio (-4.4 dB) is the upper bound requirement for an impulse with energy concentrated in 1 chip time and the lower value is for gaussian noise. Gaussian noise is the limiting case where the noise power is created by many small impulses during the symbol time and the energy is spread over the multiple chips of the symbol.

2.0 Bandpass Impulse Noise and SS 4-Level QAM Signals

Bandpass noise waveforms can be represented in general by their In Phase (I) and Quadrature (Q) components.

$$N(t) = I(t)\text{Cos}\omega_c t + Q(t)\text{Sin}\omega_c t.$$

The I and Q waveforms represent the envelope of the in-phase and quadrature components.

The frequency ($\omega_c/2\pi$) is any frequency in the passband. It is the demodulation frequency in captured waveforms. If ω_c is in the middle of the passband, then I and Q will maintain nearly the same ratio throughout short impulses of noise. If it is not in the mid-band, then Q+jI will tend to form a rotating vector.

The magnitude of N(t) is

$$N_m(t) = [I^2(t)+Q^2(t)]^{1/2}.$$

. This is the envelope of the noise waveform.

A receiver consists of a bandpass filter followed by a demodulator which derives the I and Q components of the received signal. This is followed by a low pass filter which filters the I and Q coefficients. If the bandpass filter has certain symmetry (which it usually does), then the effect of the bandpass filter is the same as that of a low pass filter of 1/2 the bandwidth. The overall filtering effect can be represented by an equivalent baseband low pass filter. The equivalent baseband filter of the I and Q waveforms captured by KII is the actual 25 MHz post-demodulator low pass filter, since the bandpass filter was much wider than 50 MHz.

These KII captured impulse noise signals were further processed through a 3-pole Butterworth low pass filter of 8.25 MHz 3 dB bandwidth. Some of the resultant waveforms are shown in the attached figures. This filter was chosen because it will permit demodulation of the 802.4L planned 11 Mchip/second SS signals. It has a little wider bandwidth than necessary and is not a totally optimum filter. But it is sufficiently accurate for analyzing impulse noise.

See descriptions of the attached figures 2 through 5 in a later section.

Figure 1 illustrates the effect of typical noise impulses on a 4 level QAM spread spectrum signal with 11 chips of spreading.

The upper waveform is that of the optimum 11 chip encoding sequence. In the modulation process, this chip code is used to change the phase (invert the polarity) of the encoded 2-bit data symbol at each change of state of the code. The same code is used to decorrelate the data waveform and achieve the processing gain.

The first and fourth waveforms from the top (labeled noise I and noise Q waveforms) show an example case of the I and Q waveforms where 3 impulsive disturbances occur during the symbol time. The third and fifth waveforms (with the numeric labels) show the effect the decorrelation process has on the noise waveforms. Finally, the I-Q axis at the lower right shows the effect of the noise and signal on the amplitude-phase of the decorrelated signal.

The signal waveform is a filtered replica of the spreading code. The system will sample the composite signal at approximately the midpoint of the vertical lines. For an ideal receiver, the signal will be at optimum amplitude at this time. The phase-amplitude shows the case where the signal phase is 45 degrees. The magnitude of this phasor is 11 X the chip signal amplitude.

The effect of the noise on the composite signal + noise phasor can be examined by looking at the signal and noise separately. Noise impulse 1 has negative I and Q components at the sample time with Q smaller than I. It is represented by phasor number 1 in the phase plane. Impulse 2 is chopped by the decorrelator and has very little effect. It is represented by the small phasor 2 on the phase plane. The effect of impulse 3 is formed in a like manner. In sum, the phasor marked "composite noise" is the vector sum of the 3 random phase noise vectors as they occurred at the sample time.

If the tip of the composite noise phasor in the diagram falls outside the first quadrant, there will be a symbol error. It must be at least 11/sqrt(2) times the spread signal amplitude to do this. This ignores the DPSK effect. If impulses affect 2 consecutive symbols, then the sum of the composite noise vectors must be less than the above to guarantee no error.

3.0 Impulse Probability of Error

The probability of error due to impulse noise can be investigated by looking at two extreme cases.

1. The disturbance is due to a large number of relatively weak impulses and

2. The disturbance is due to just 1 strong impulse.

The extreme of case 1 is steady gaussian noise.

The necessary signal to rms noise envelope power ratio for achieving an error probability of 0.1 will be investigated for the extreme cases.

The steady noise case. The signal to steady noise power necessary to maintain an error probability of 0.1 for 4 level QAM without spreading is 3 dB. This can verified by classical texts. Thus:

S/N power ratio necessary w/o spreading gain	3 dB
S/N gain due to spreading	10.4 dB
S/N necessary with spreading (3-10.4)	-7.4 dB

The necessary signal power (before decorrelation) at the sample instant to gaussian noise power ratio for achieving an error rate of 0.1 errors/symbol is -7.4 dB.

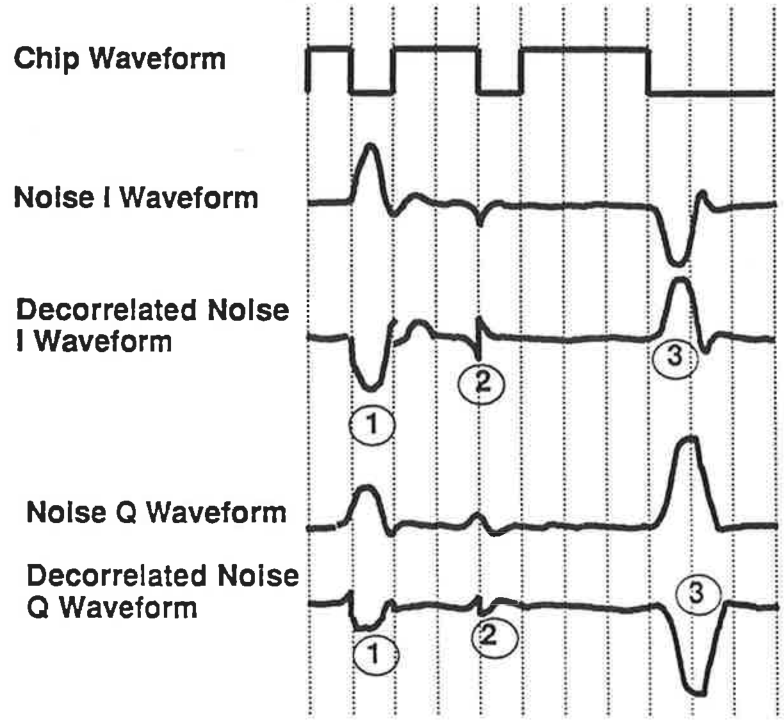
The signal envelope can be 7.4 dB below the noise envelope.

The Single Impulse Case. This will be investigated with the 8.25 MHz 3-pole Butterworth filter described above. A single impulse will create an impulse response magnitude with a shape that is the absolute value of the equivalent low pass filter impulse response. This impulse response shape is shown in figure 2. The signal level necessary to cause this impulse to have a probability of error less than 0.1 will be derived.

This impulse response has an rms value of 40.8 mv. It's magnitude is above 190 mv during 20% of a chip interval. Thus, the impulse will create a composite noise vector as shown in figure 1 at the sample time of 190 mv with no more than 20% probability. If this noise vector is within 3 dB of the composite decorrelated signal amplitude it is just strong enough to cause an error when it has exactly the right phase. The noise will aid the signal on at least 1/2 of it's occurrences so long as it's amplitude is less than that of the composite decorrelated signal. At this level, it will cause an error for 1/2 of the possible phases. But, from the impulse response, the magnitude never exceeds the 80 percentile point by as much as 3 dB.

EFFECT OF IMPULSE NOISE ON SS SIGNAL PHASE

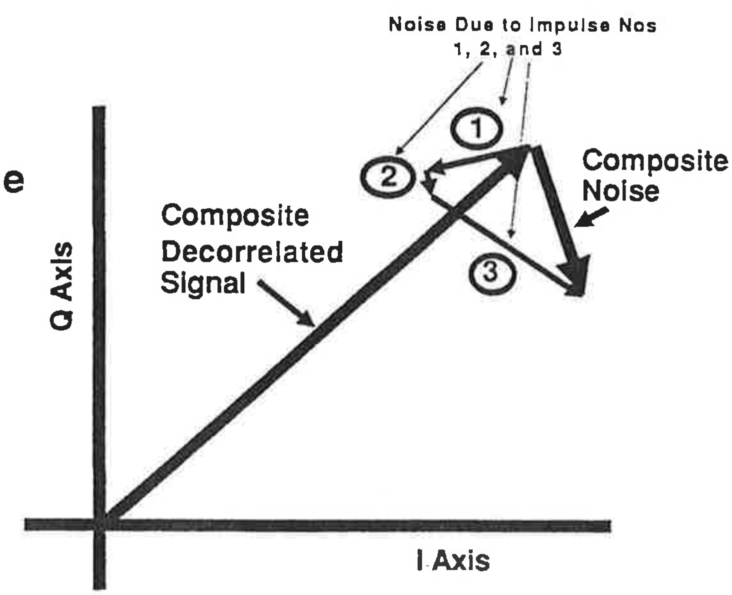
Figure 1



Effect of Impulse Noise on Signal Phase

- 11 Chip SS Correlation/Decorrlation Sequence
- Composite Signal is 11x Voltage-Time Integral of the Spread Signal Chips
- Composite Noise is Random Phasor Sum of the Individual Impulse V-T Integrals

Noise will cause an error if either the I or Q composite component exceds sqrt(2) times the composite decorrelated signal amplitude.



The conditions and result are summarized below:

Impulse 80 percentile level	190 mv
Impulse envelope rms level over the 1 us symbol time	40.8 mv
Envelope rms to 80 percentile ratio	-13.4 dB
Signal decorrelation gain (11x)	20.8 dB
N/S(after decorrelator) ratio for just causing error	-3 dB
N/S(after decorrelator) ratio for 50% Pe	0 dB
Most Pessimistic S/N(rms) (-20.8+13.4+3) (Upper Bound)	-4.4
Lower Bound S/N(rms) (-20.8+13+0)	-7.8

The late impulse response undershoot was ignored in the above. This shouldn't affect the results very much.

The best estimate for the necessary rms S/N for achieving an error rate of 0.1 is about -6 dB.

Further Comments. The above comparison only holds for an error probability of 0.1. If it were repeated for very low error probabilities, the necessary S/N for steady noise would increase considerable (4 dB for 0.01 for example). In the single impulse case, only about 2 dB would be needed to achieve 0 error probability.

If very low error probabilities per impulse are needed, then equating the impulse rms value to gaussian noise over the symbol time will over estimate the error rate for single isolated impulses. Thus, the picture would need another look in this case. Thus far, it appears that in the real world, most impulses occur in short bursts. For this reason the rms analysis is probably appropriate for error probabilities as low as 0.01 at least. It is safe for lower probabilities (it overestimates the signal power required) but may be inaccurate.

Another factor not considered is that the receiver signal processing is likely to ignore very high impulse peaks. There is a good likelihood that 5 to 6 dB S/N improvement can be achieved this way. This should be kept in mind when analyzing impulse noise.

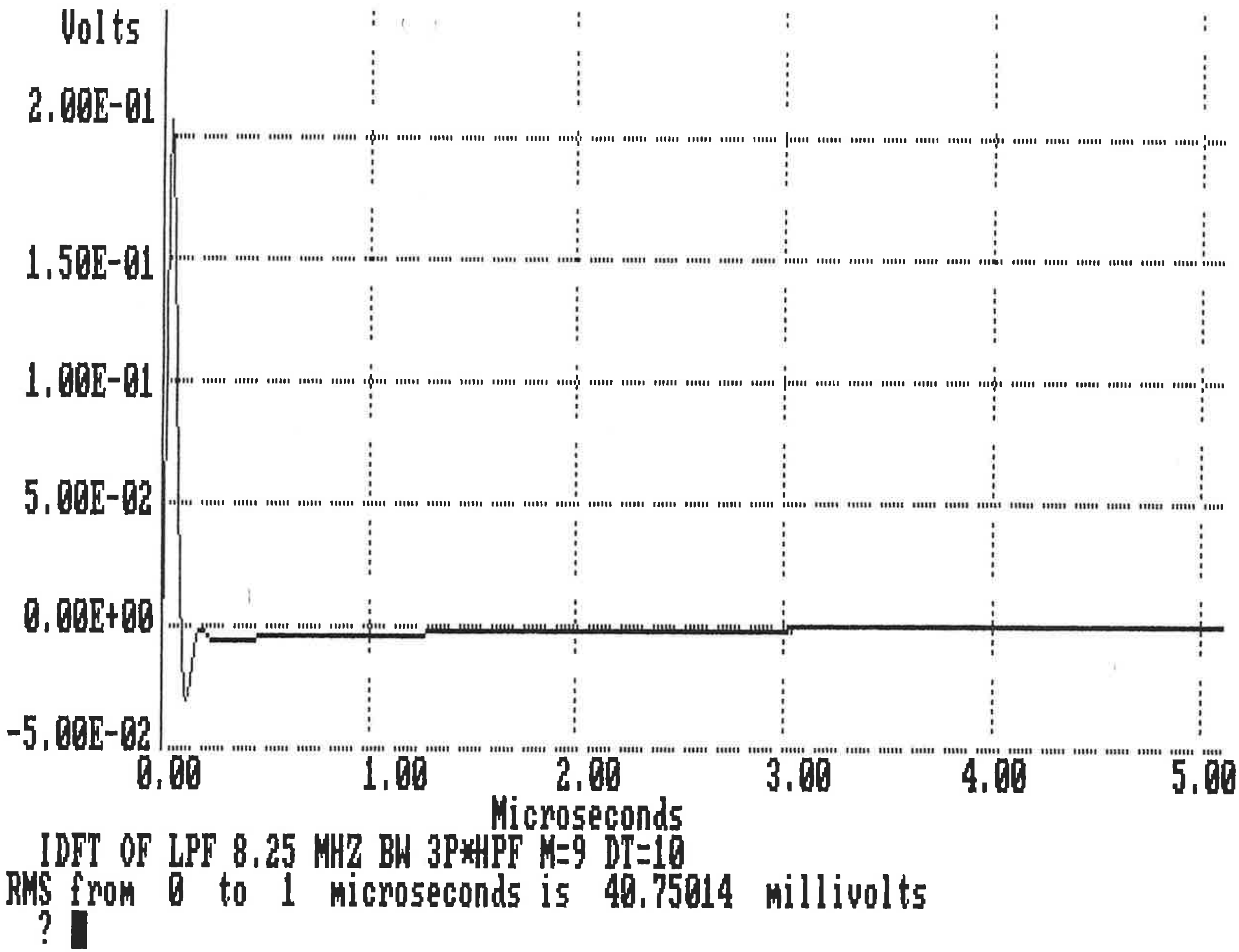

Volts
2.00E-01
1.50E-01
1.00E-01
5.00E-02
0.00E+00
-5.00E-02
0.00
1.00
2.00
3.00
4.00
5.00
Microseconds
IDFT OF LPF 8.25 MHZ BW 3P*HPF M=9 DT=10
RMS from 0 to 1 microseconds is 40.75014 millivolts
?

Figure 2A

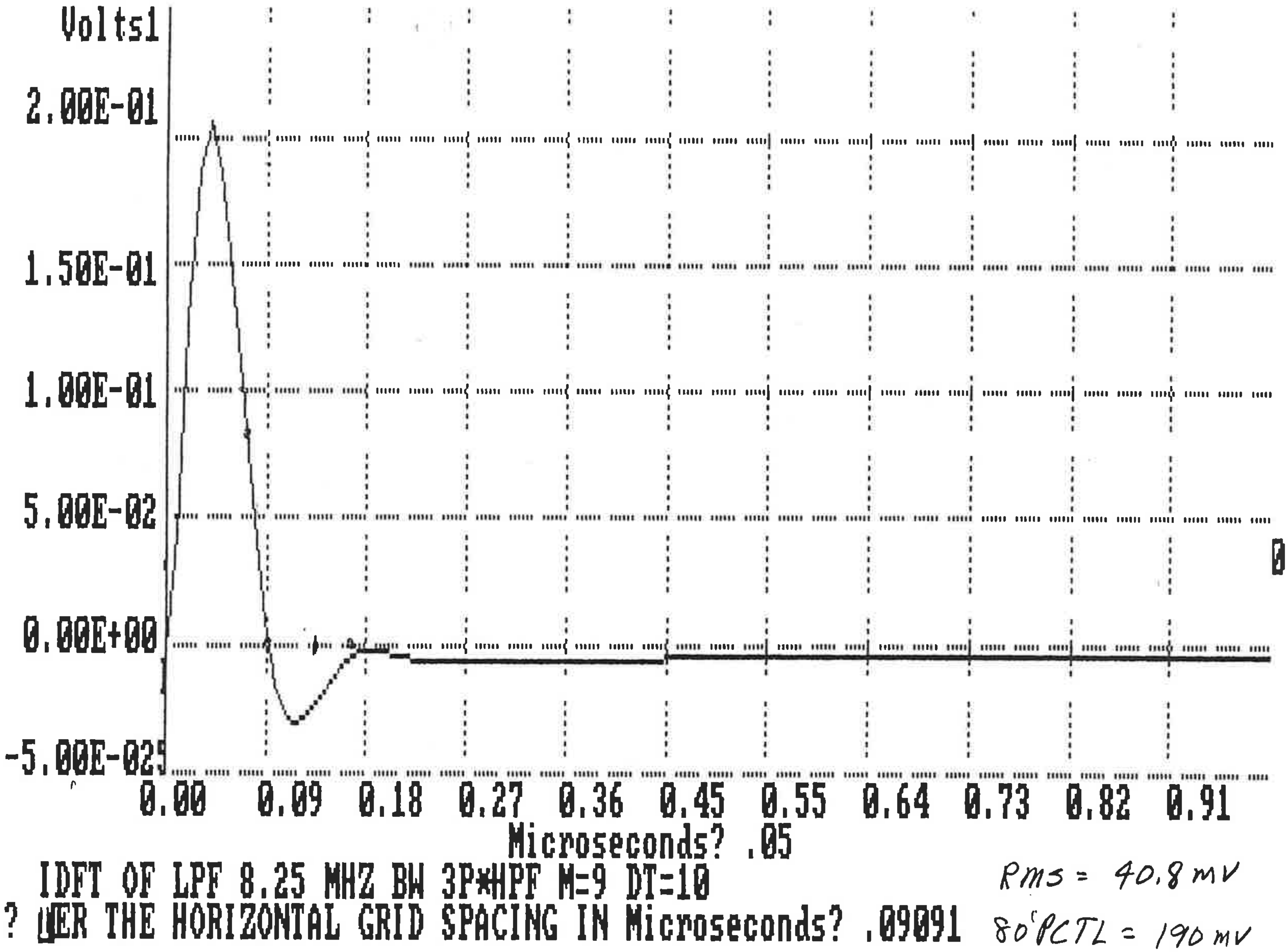
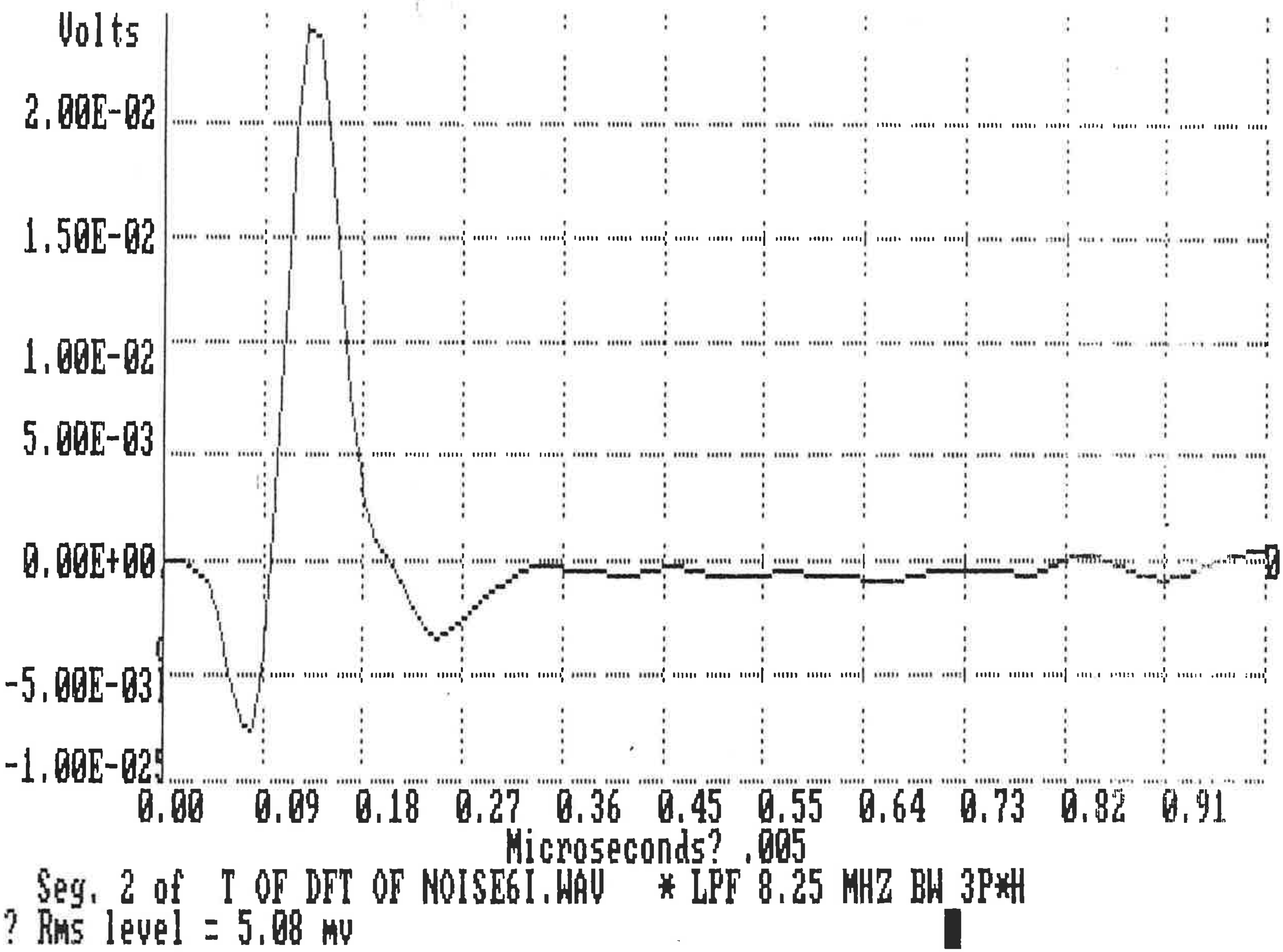

Volts1
2.00E-01
1.50E-01
1.00E-01
5.00E-02
0.00E+00
-5.00E-02
0.00 0.09 0.18 0.27 0.36 0.45 0.55 0.64 0.73 0.82 0.91
Microseconds? .05
IDFT OF LPF 8.25 MHZ BW 3P*HPF M=9 DT=10
? ER THE HORIZONTAL GRID SPACING IN Microseconds? .09091
RMS = 40.8 mV
80% CTL = 190 mV

Figure 2B



Volts
2.00E-02
1.50E-02
1.00E-02
5.00E-03
0.00E+00
-5.00E-03
-1.00E-02
0.00
0.09
0.18
0.27
0.36
0.45
0.55
0.64
0.73
0.82
0.91
Microseconds? .005
Seg. 2 of T OF DFT OF NOISE6I.WAV * LPF 8.25 MHZ BW 3P*H
? Rms level = 5.08 mv

Figure 3

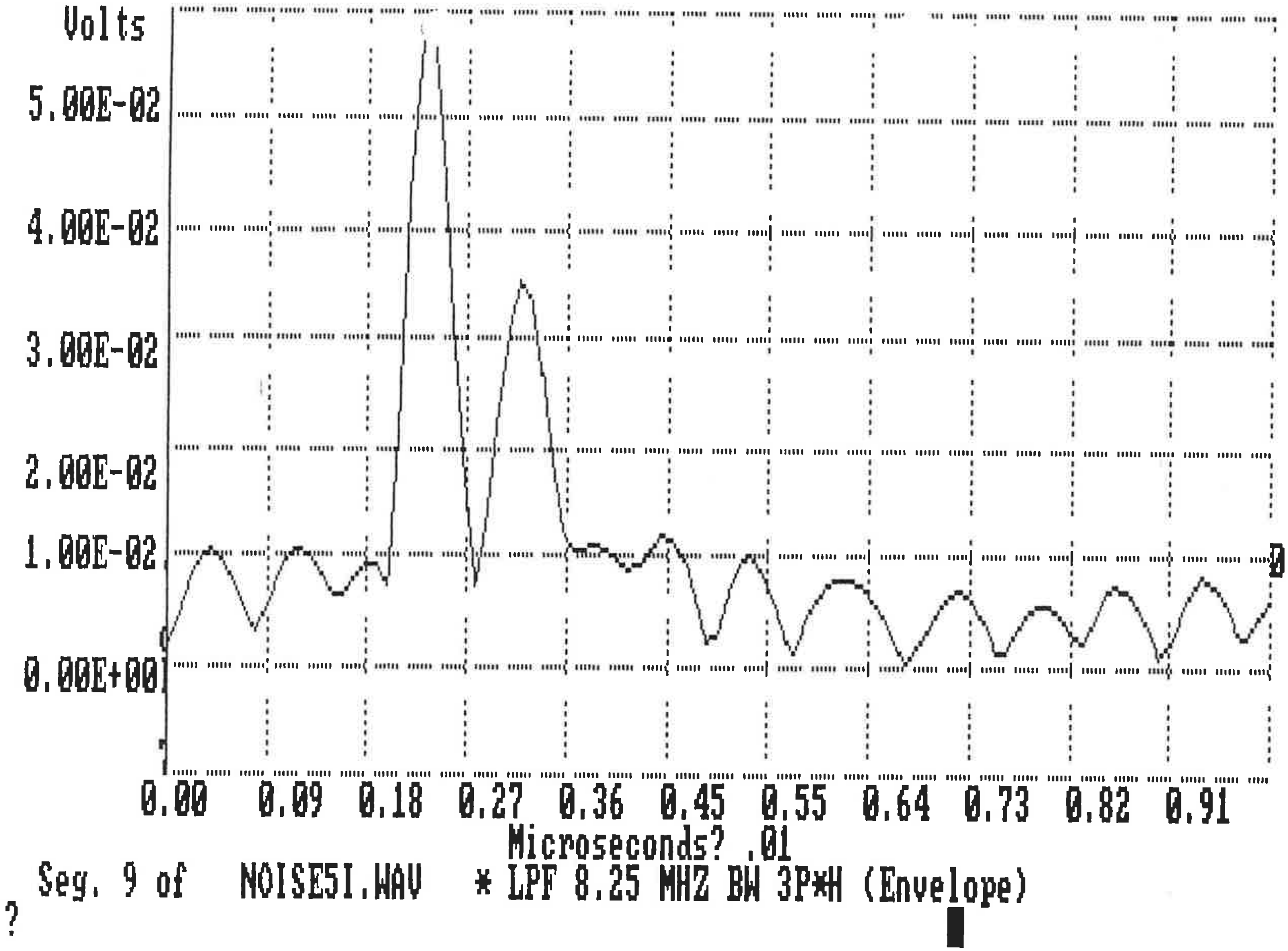

Volts
5.00E-02
4.00E-02
3.00E-02
2.00E-02
1.00E-02
0.00E+00
0.00
0.09
0.18
0.27
0.36
0.45
0.55
0.64
0.73
0.82
0.91
Microseconds? .01
Seg. 9 of NOISE5I.WAV * LPF 8.25 MHZ BW 3P*H (Envelope)
?

Figure 4A

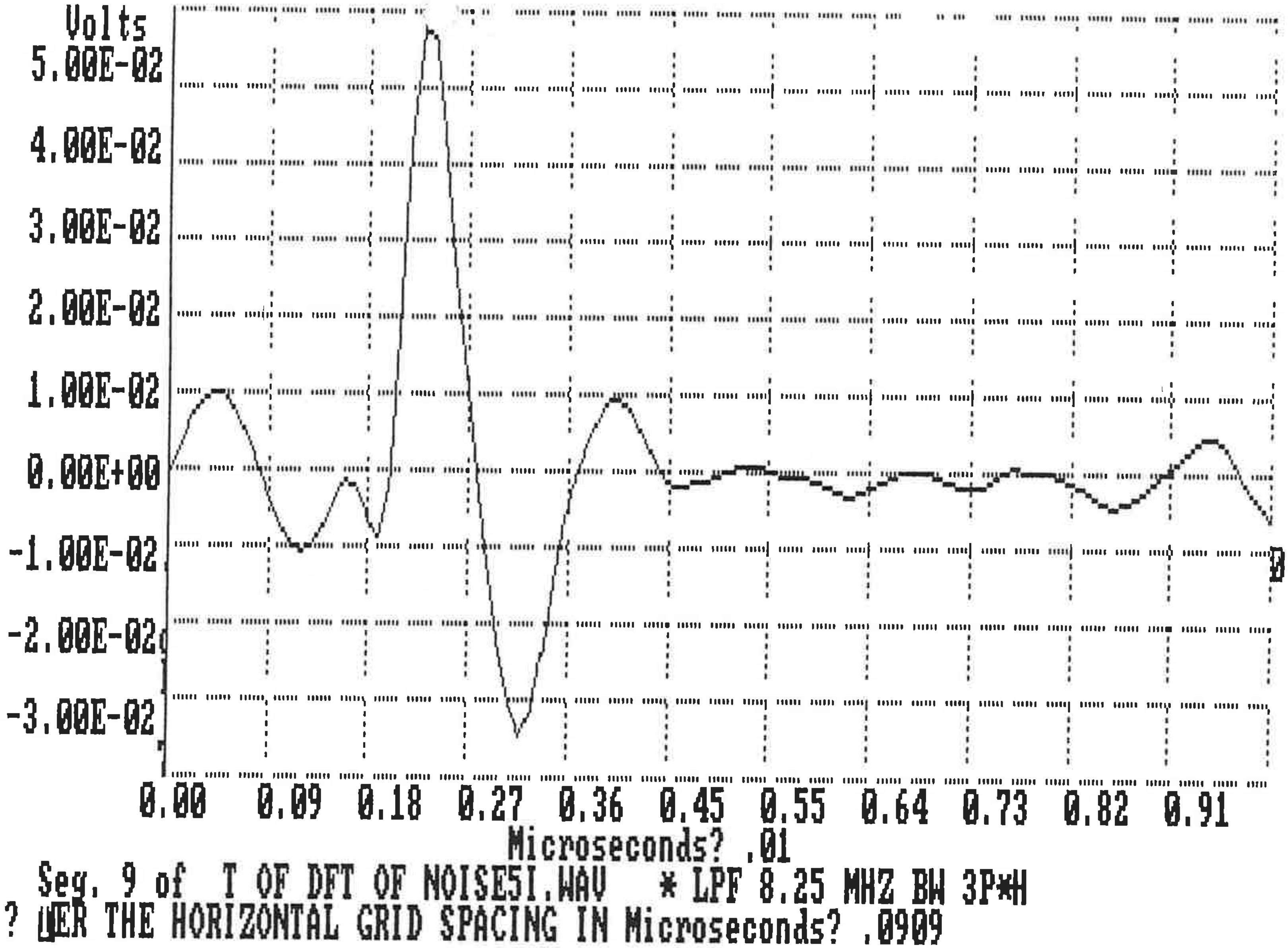

Volts
5.00E-02
4.00E-02
3.00E-02
2.00E-02
1.00E-02
0.00E+00
-1.00E-02
-2.00E-02
-3.00E-02
0.00 0.09 0.18 0.27 0.36 0.45 0.55 0.64 0.73 0.82 0.91
Microseconds? .01
Seg. 9 of T OF DFT OF NOISE51.WAV * LPF 8.25 MHZ BW 3P*H
? ER THE HORIZONTAL GRID SPACING IN Microseconds? .0909

Figure 4B

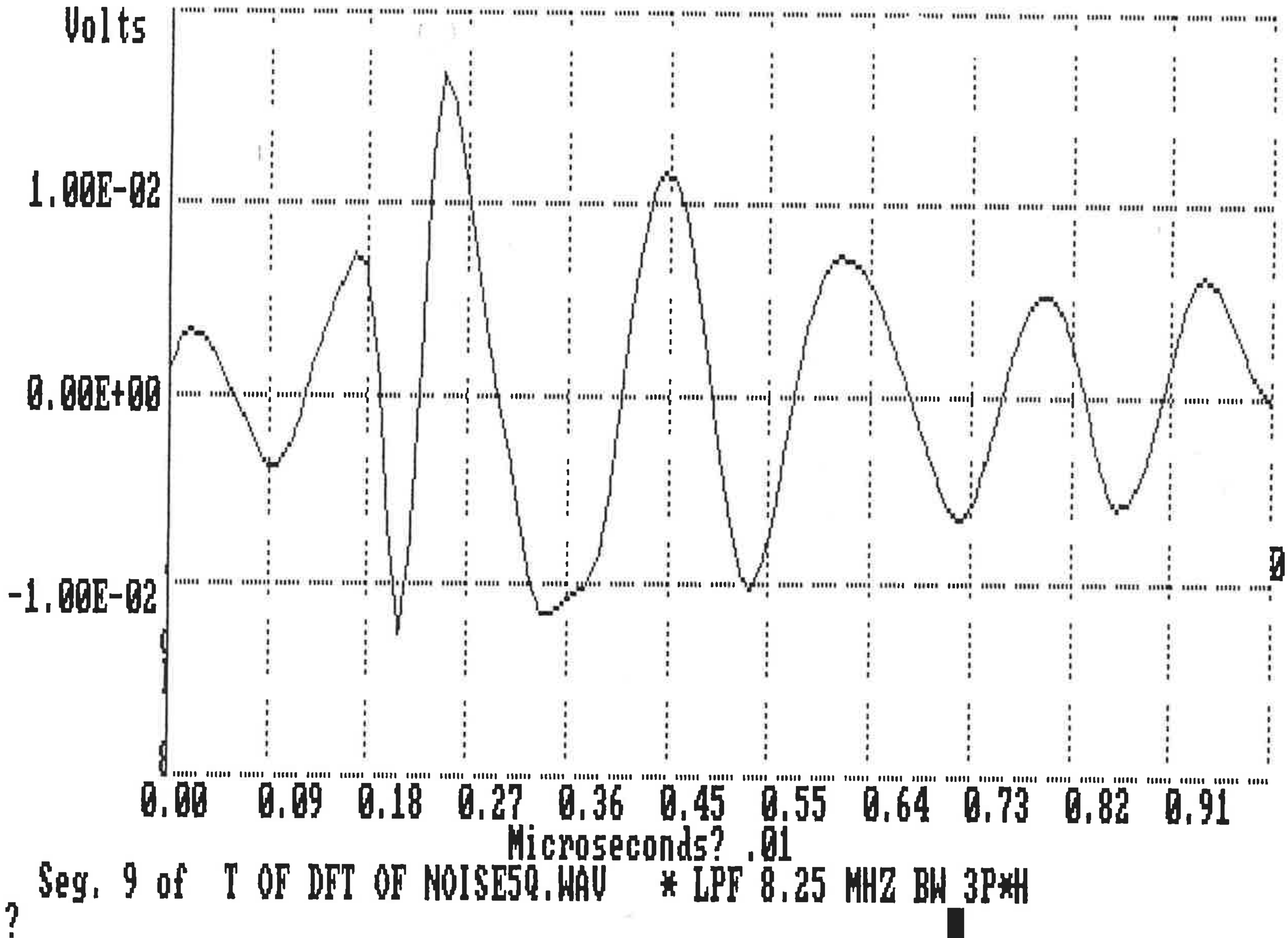

Volts
1.00E-02
0.00E+00
-1.00E-02
0.00
0.09
0.18
0.27
0.36
0.45
0.55
0.64
0.73
0.82
0.91
Microseconds? .01
Seg. 9 of T OF DFT OF NOISE5Q.WAV * LPF 8.25 MHZ BW 3P*H
?

Figure 4C.

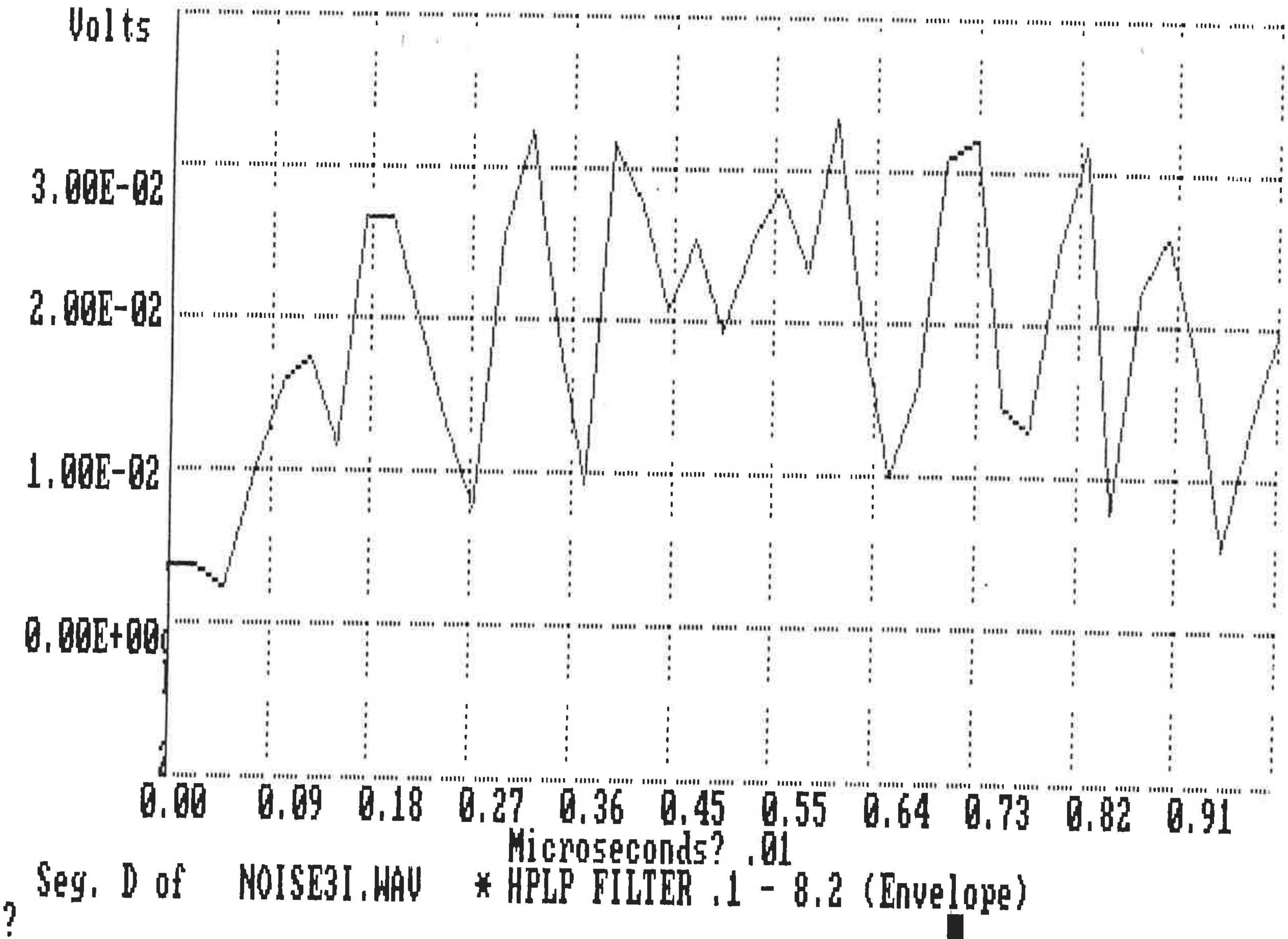

Volts
3.00E-02
2.00E-02
1.00E-02
0.00E+00
0.00
0.09
0.18
0.27
0.36
0.45
0.55
0.64
0.73
0.82
0.91
Microseconds? .01
Seg. D of NOISE3I.WAV * HPLP FILTER .1 - 8.2 (Envelope)
?

Figure 5

4.0 Description of Example Waveforms

The impulse response of the 8.25 MHz 3-pole Butterworth filter is shown in figures 2A and 2B.

Figure 3 shows the I component of a noise waveform captured by KII at a department store near Atlanta at 2.4 GHz. This is an example of a noise waveform which probably resulted from 1 isolated impulse.

Typical noise waveforms are created by multiple impulses within the 8.25 MHz resolution time and are represented by the waveforms of figures 4A through 4C. This shows the envelope and the 2 components of a waveform captured at 905 Mhz in the department store near Atlanta. The time scale of figure 4 is divided into chip times at 11 Mchips/second. From the envelope trace, it is likely that this waveform was created by impulsive events about 1 chip time separated.

Figure 5 shows an extreme case where many equivalent excitation impulses must have created the noise.

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Through-the Air Physical Media, Radio

Retail measurement results

Donald C. Johnson, NCR

The attached charts[1] are impulse responses measured at a department store in a mall near Atlanta.

The charts are the decorrelated waveforms of a 255 bit M-sequence. The chip time is 40 ns and the equivalent baseband bandwidth of the receiver was 25 MHz. This gives an approximately 40 ns delay resolution.

Each chart is marked with a distance and attenuation value relative to the attenuation at 1 meter distance. The reference chart (not shown) was taken at 10 feet (3.3 meters) and the 1 meter attenuation is taken as 10.4 dB less than that at the measured distance of 3.3 meters.

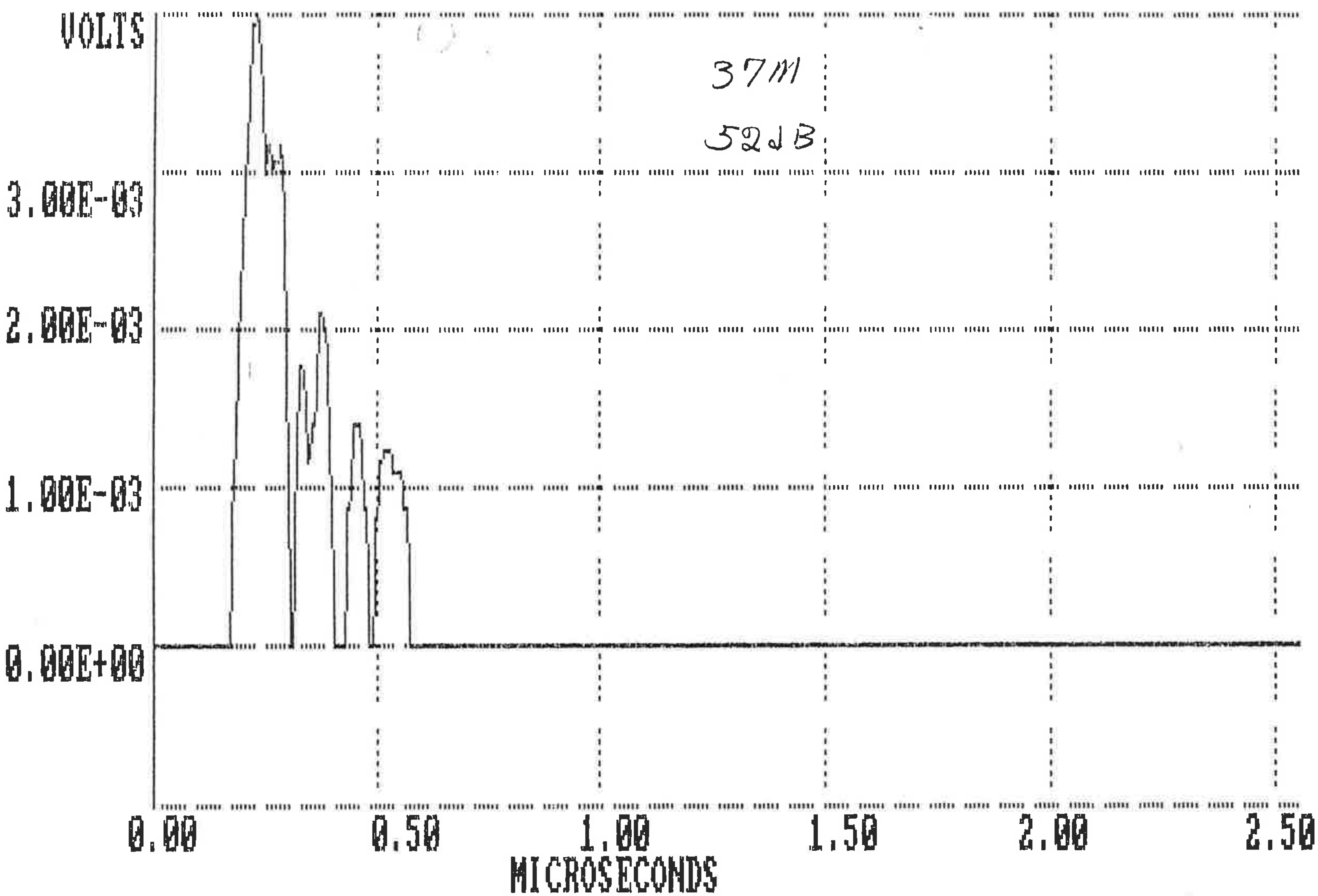
The computed rms delay spread does not have any correction for the 40 ns resolution. Actual delay spreads are less than that measured.

MSEQ08 is marked "between floors". One end of the path was near an elevator opening. All of the other paths were confined to a single floor.

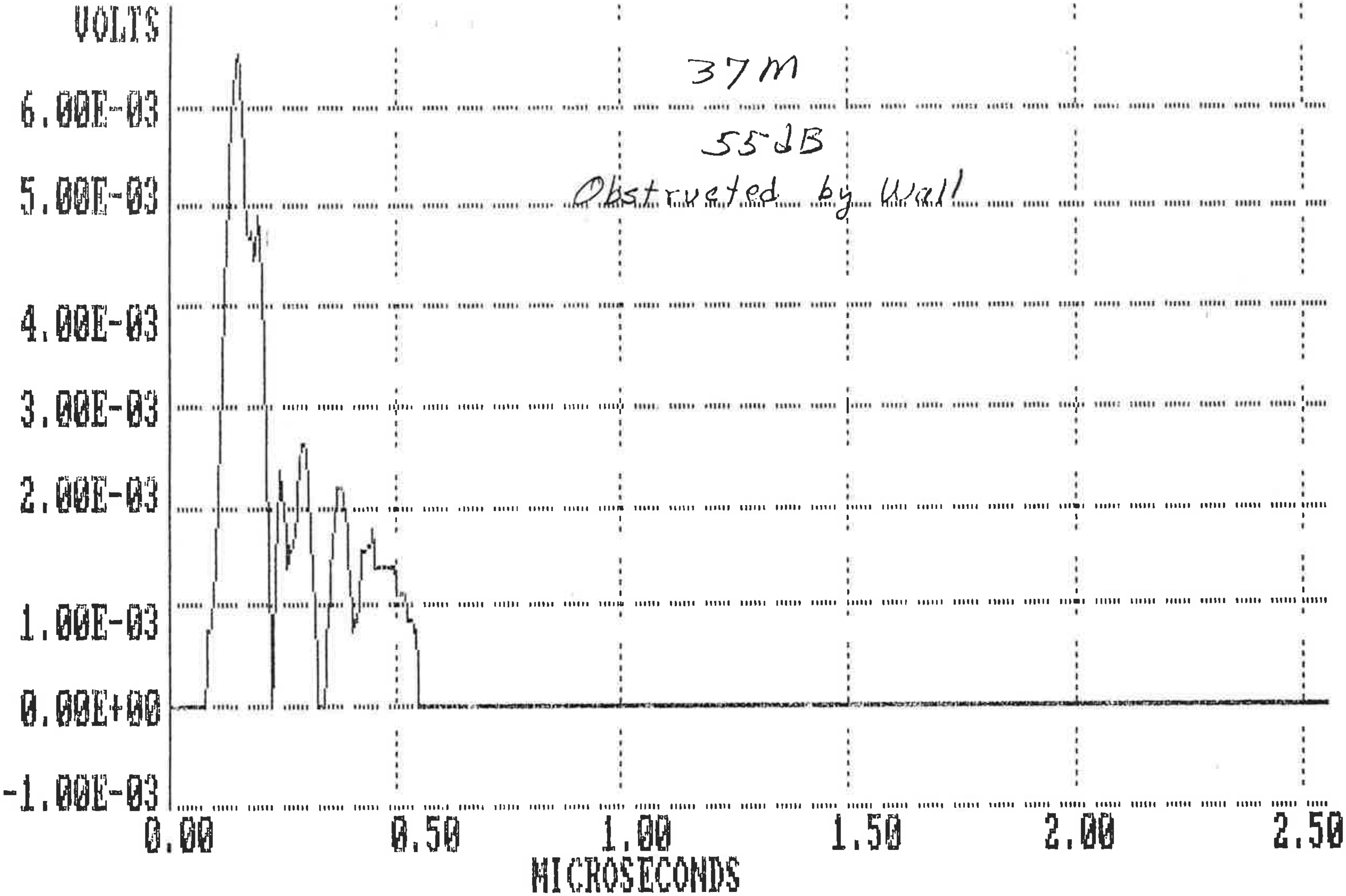
Refer to the charts for further attributes.

The result of an analysis is provided as follows

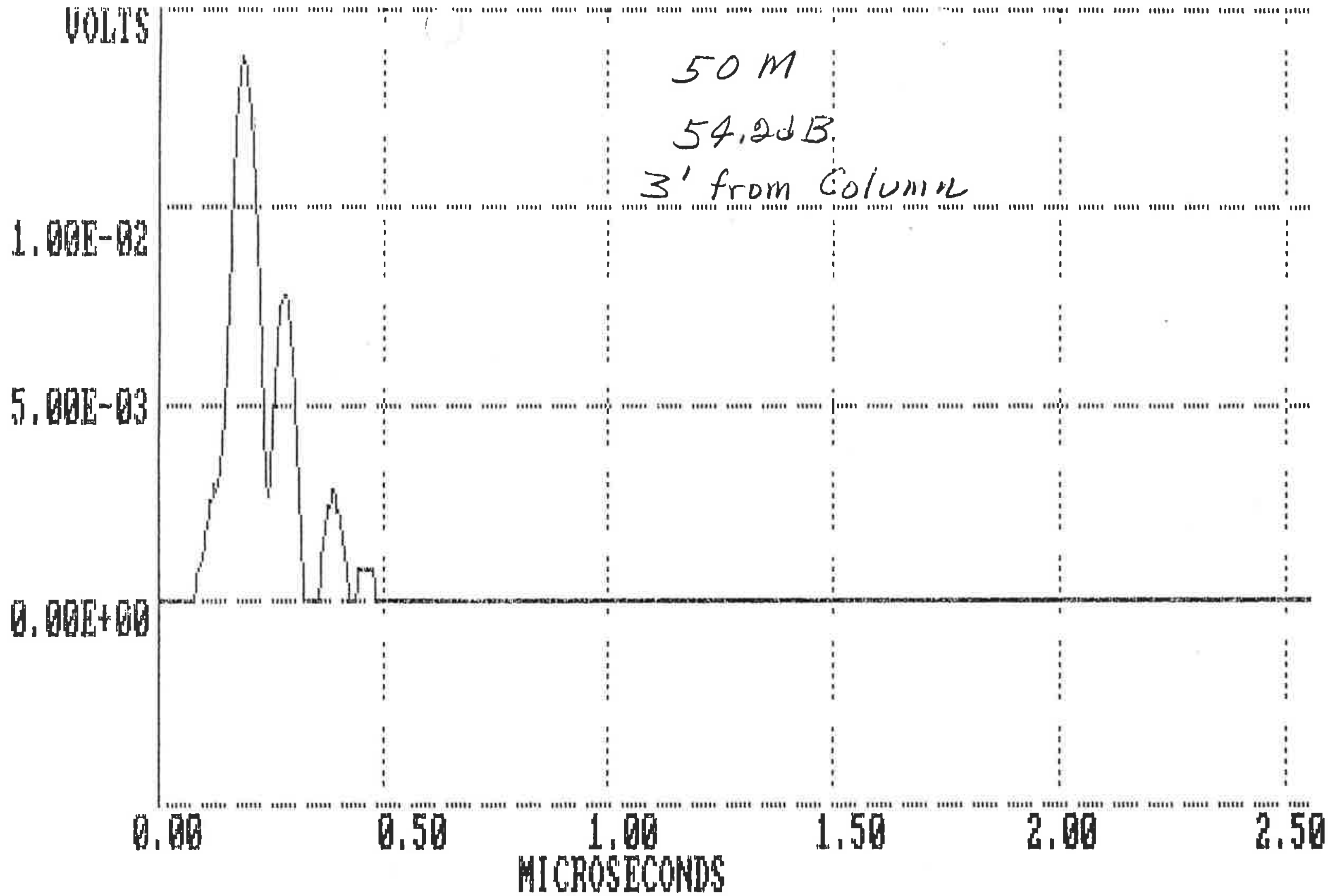
[1]These charts were submitted to the IEEE p802.4L meeting held 5-7 November 1989 in Fort Lauderdale, FL. The temporary number was F.4L/8.



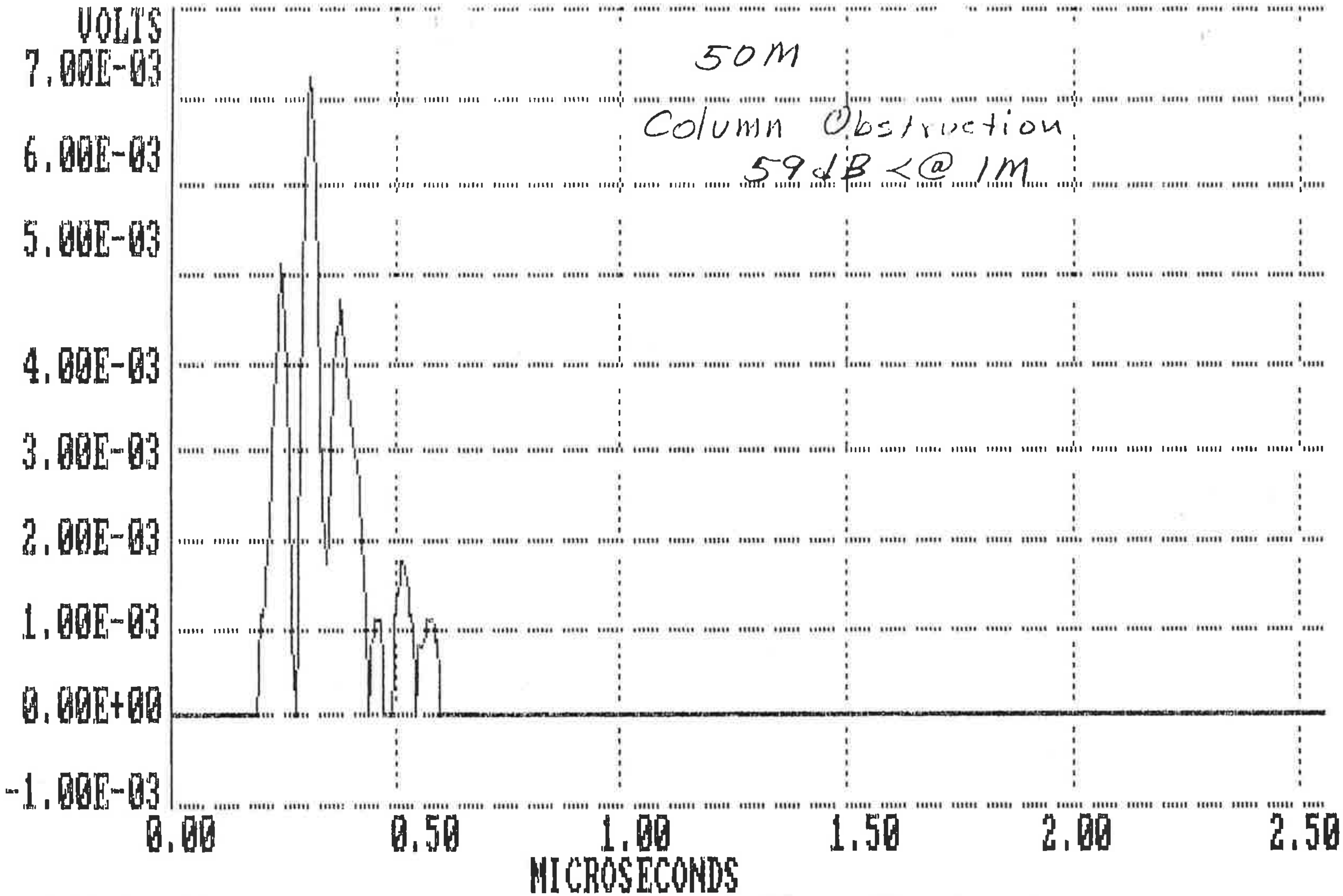
MSEQ19.WV1 - Dlt= 5ns - M= 9 - Offs= 64.35 us (Envelope)
Rms delay spread (Trms) = 89.27528 ns
?



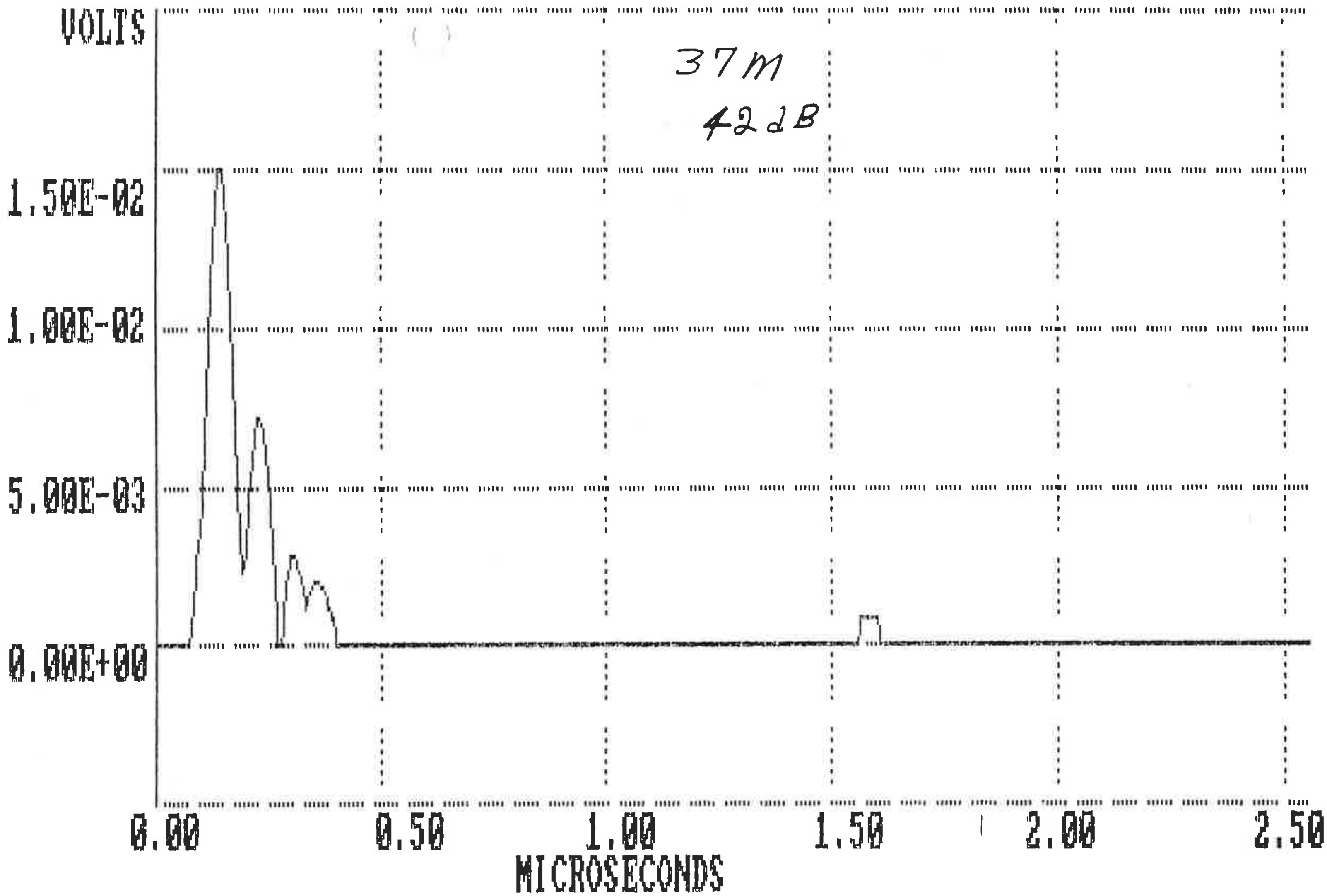
MSEQ18.WV1 - Dlt= 5ns - M= 9 - Offs= 18.15 us (Envelope)
Rms delay spread (Trms) = 95.14538 ns
?



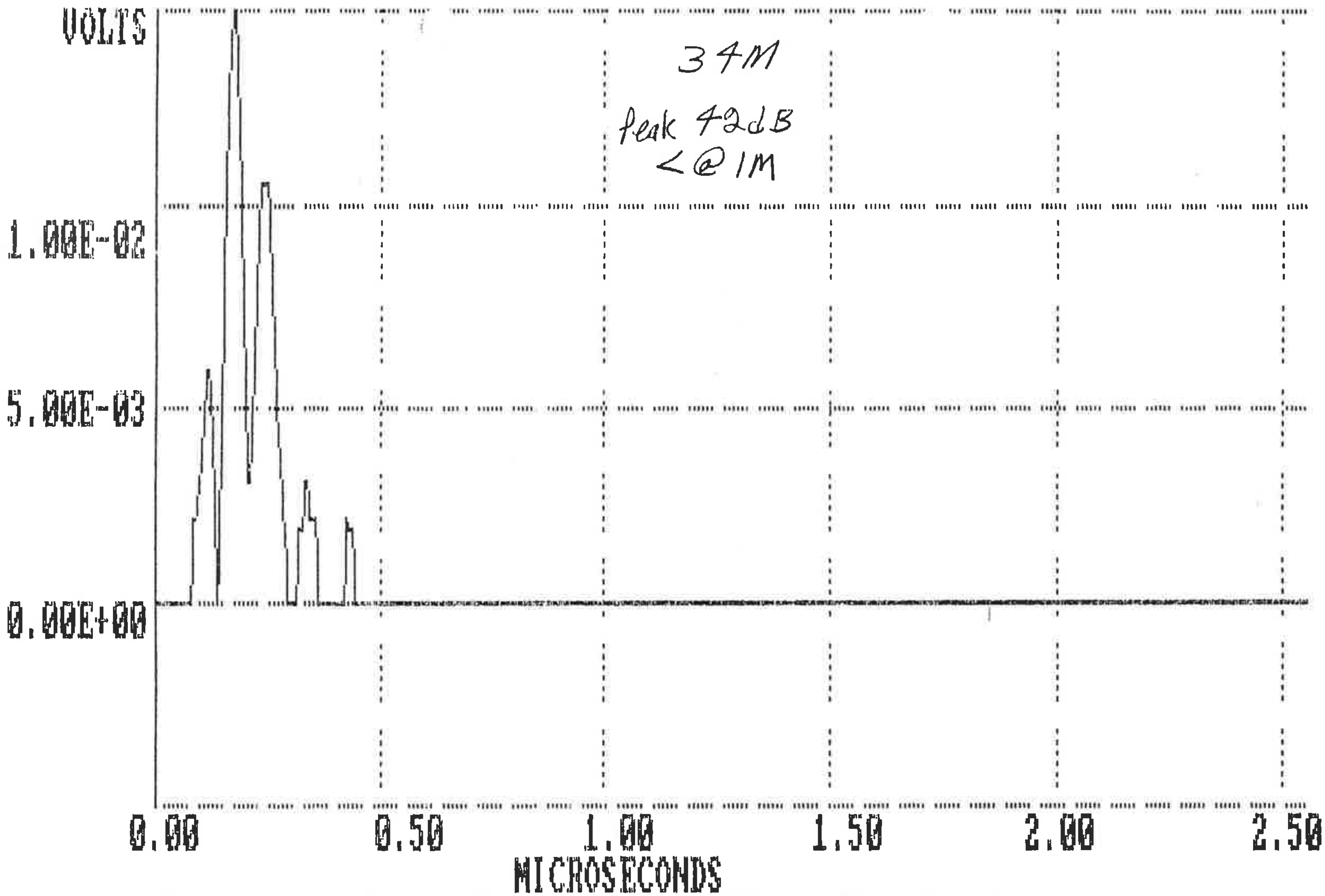
MSEQ14.WV1 - Dlt= 5ns - M= 9 - Offs= 4.65 us (Envelope)
Rms delay spread (Trms) = 49.98744 ns
?



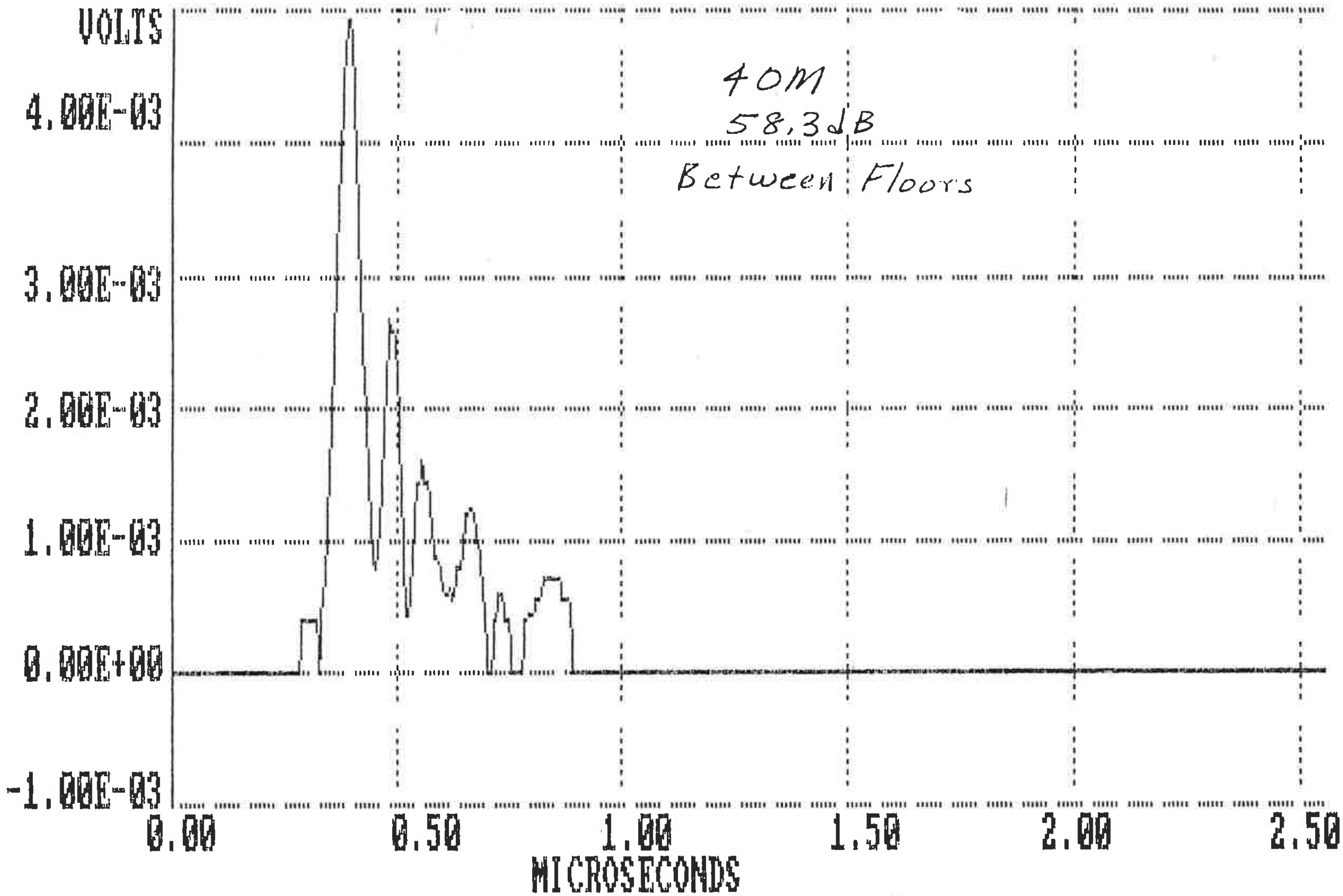
MSEQ13.WV1 - Dlt= 5ns - M= 9 - Offs= 1.62 us (Envelope)
Rms delay spread (Trms) = 68.22855 ns
?



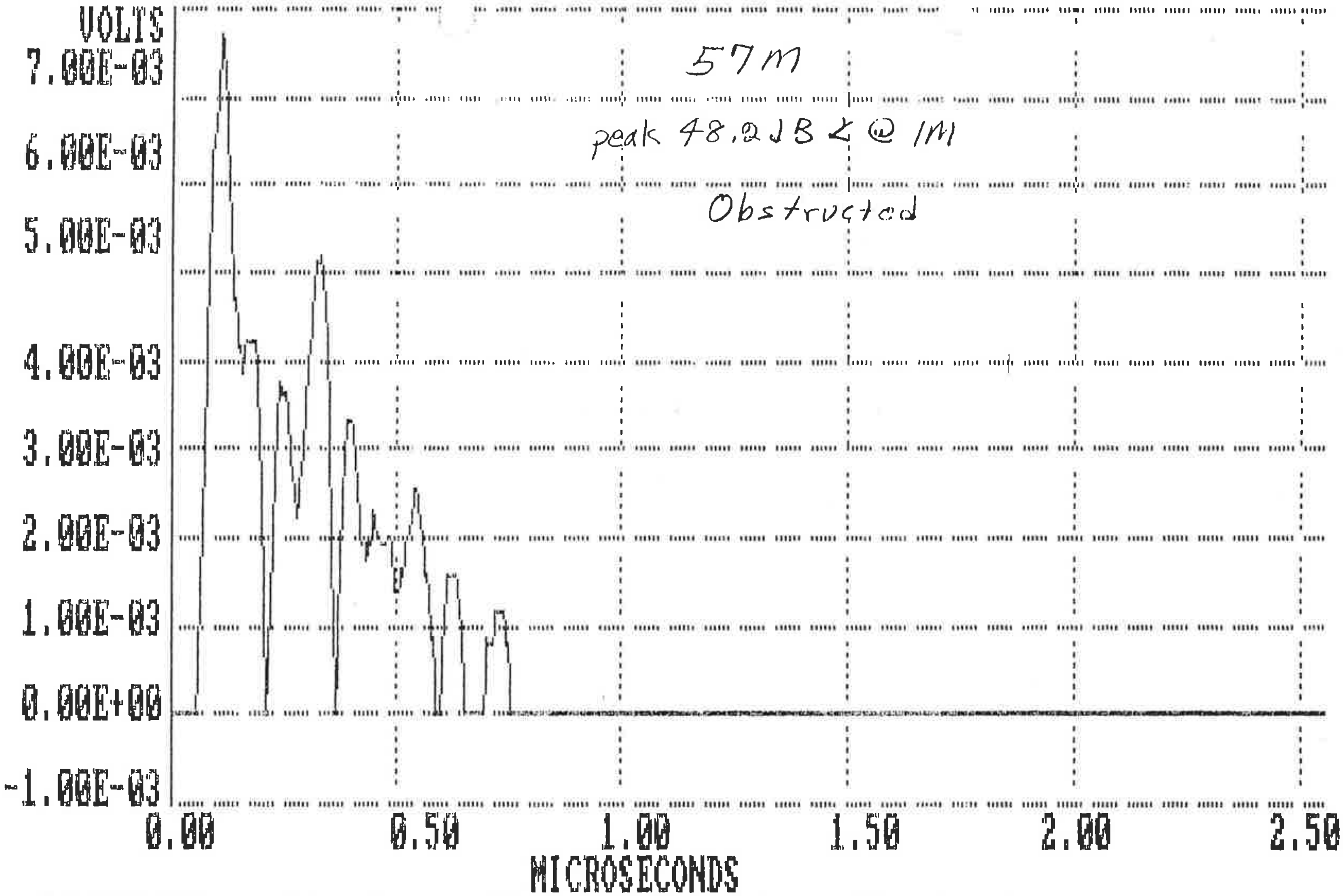
MSEQ10.WV1 - Dlt= 5ns - M= 9 - Offs= .81 us (Envelope)
Rms delay spread (Trms) = 86.16861 ns
?



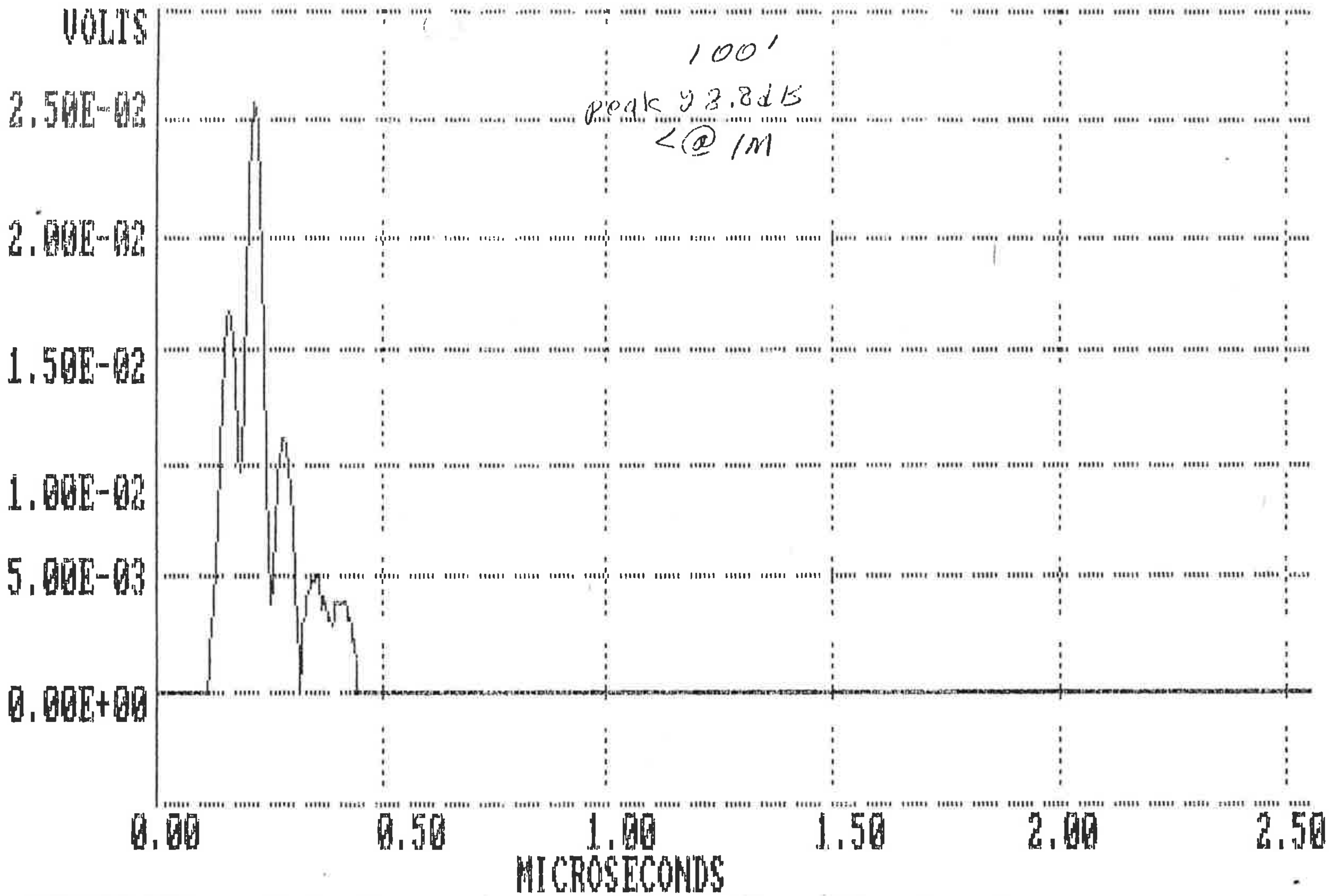
MSEQ09.WV1 - Dlt= 5ns - M= 9 - Offs= .88 us (Envelope)
Rms delay spread (Trms) = 49.36026 ns
?



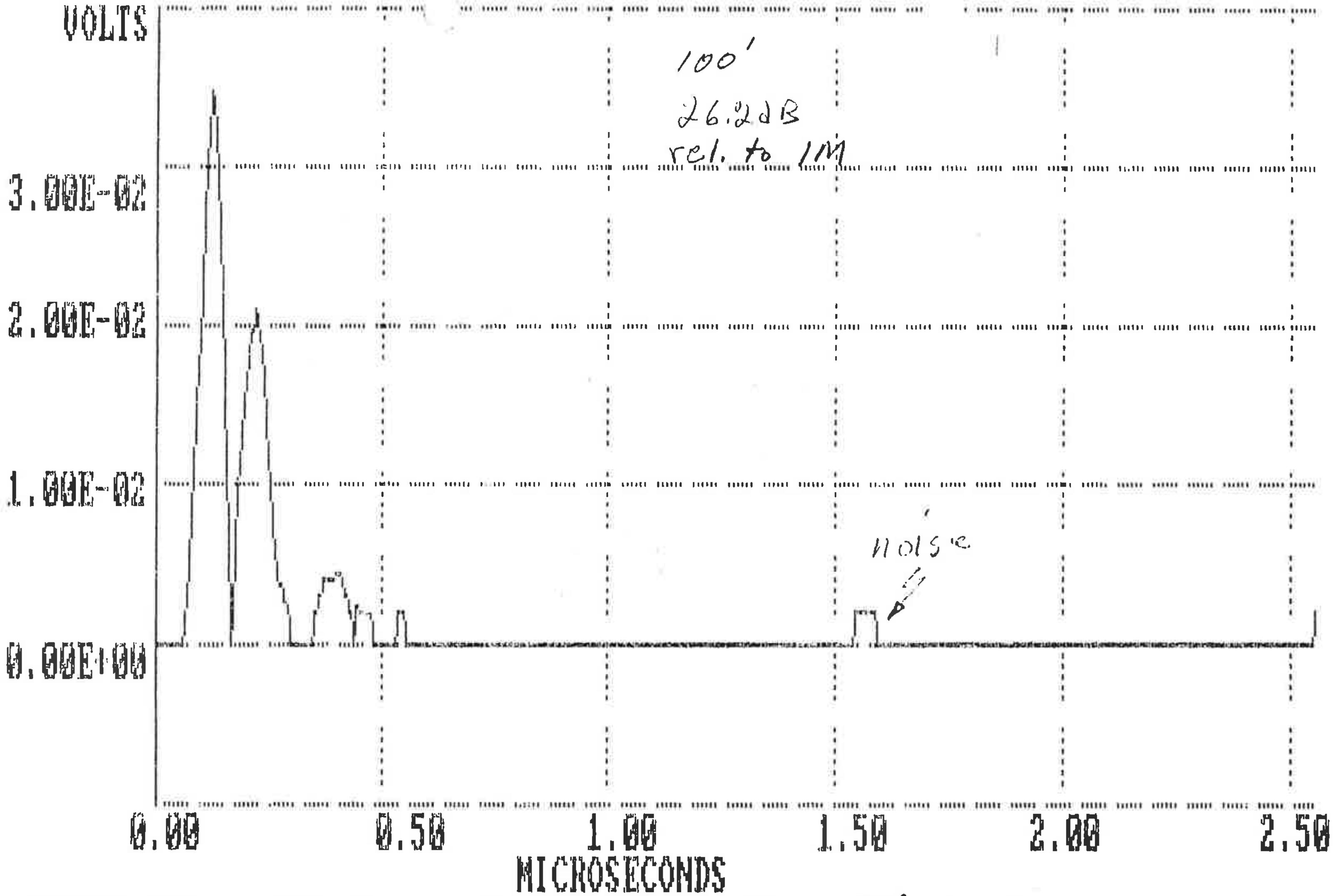
MSEQ08.WV1 - Dlt= 5ns - M= 9 - Offs= 10.5 us (Envelope)
Rms delay spread (Trms) = 96.5174 ns
?



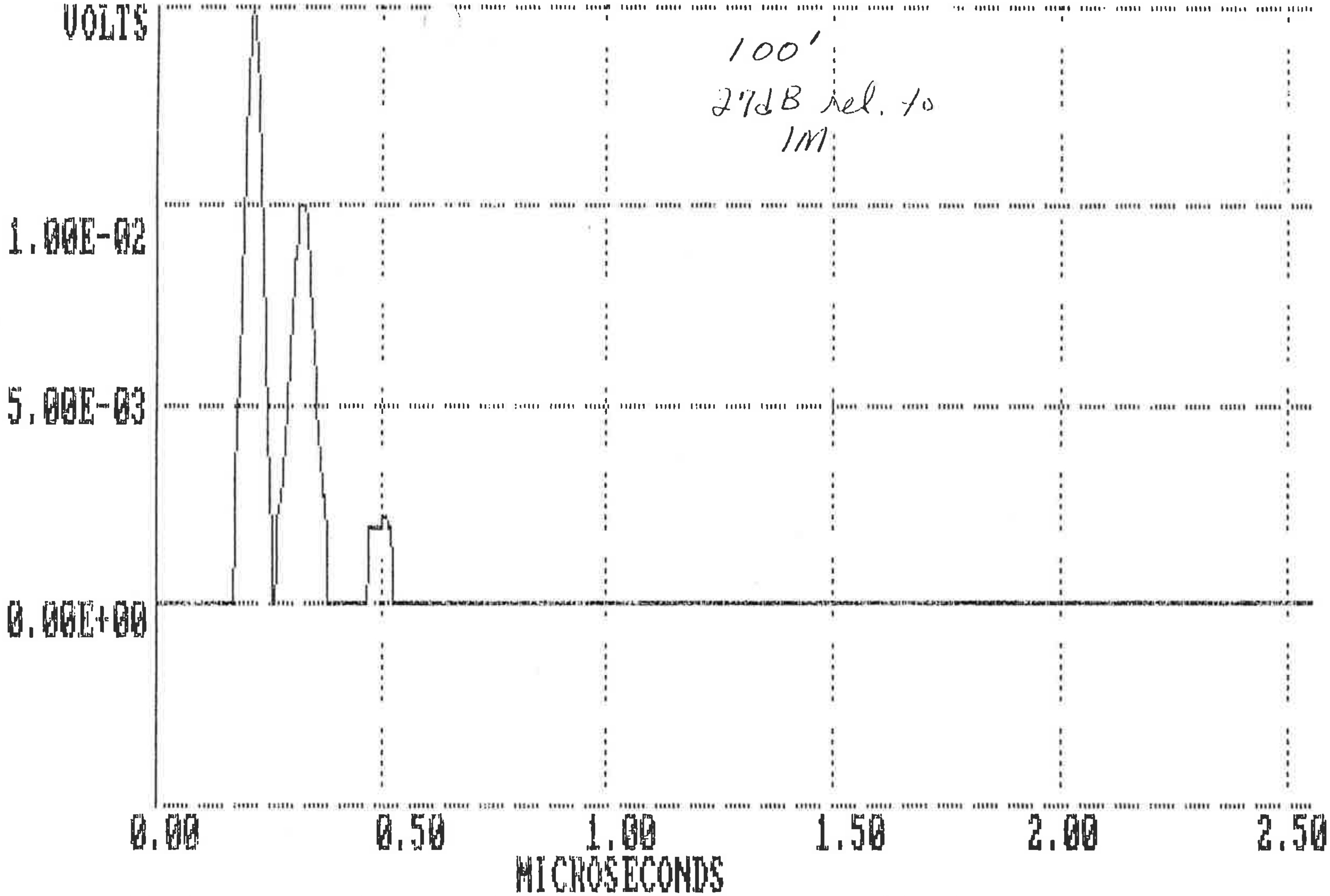
MSEQ07.WV1 - Dlt= 5ns - M= 9 - Offs= 2.1 us (Envelope)
Rms delay spread (Trms) = 144.1652 ns
?



MSEQ04.WV1 - Dlt= 5ns - M= 9 - Offs= .15 us (Envelope)
Rms delay spread (Trms) = 50.86082 ns
?



MSEQ03.WV1 - Dlt= 5ns - M= 9 - Offs= 1.03 us (Envelope)
Rms delay spread (Trms) = 91.14362 ns
?



MSEQ02.WV1 - Dlt= 5ns - M= 9 - Offs= 3 us (Envelope)
Rms delay spread (Trms) = 61.95154 ns
?

Attenuation profiles

Retail department store

Location	Frequency	6 db/octave to knee	Final slope	Deviation from regression line σ
	GHz	m	dB/octave	dB
Second floor	.915	10	10	4.5
Second floor	2.44	8	10.8	6.7
First floor	.915	13	12.6	5.1
First floor	2.44	7	10.5	6.3

Distances measured: 1 - 70 m

915 MHz Impulse Noise

Retail Terminal source (10 ft away)

Trigger Threshold	Triggers/sec	Total number of triggers
dBm	note 1	
-57 (50 mV-KII)	3.15	31
-51 (100 mV)	0.24	62

note 1 5.5 dB per decade count change

Necessary signal level (See document 89-20)

-51	dBm	wideband peak @ 0.24 / s (25 MHz LPF)
-18	(+5.2 σ)	rms (8MHz BW)/peak (25 MHz BW) 13 random samples
-6		S/N @ 10 impulses/error
+2		receiver allowance
-73	dBm	

-73 - - 68 (-73+σ) dBm for $P_e <= 2.4 \times 10^{-8}$

-79 - - 74 dBm for $P_e <= 3.15 \times 10^{-7}$

(74 to 79 dB excess loss over 1 m at 600 mW)

Note: The above does not take into account any S/N gain against impulse noise due to peak clipping. This clipping effect will give about a 6 dB gain resulting in 80 to 85 dB for the excess loss of the last line above.

Other sources - Less frequent than retail terminal.

2.44 GHz Impulse noise

Retail Terminal source (10 ft away)

Trigger Threshold	Triggers/sec	Total number of triggers
dBm		
-44 (10 mV-KII) 25 MHz LPBw	0.0284	47

Necessary signal level (See document 89-20)

-44	dBm	wideband peak @ 0.0284 / s (25 MHz LPF)
-21.8	(+5.2 dB σ)	rms (8MHz BW)/peak (25 MHz BW)
-6		S/N @ 10 impulses/error
+2		receiver allowance
-69.8	dBm	

-69.8 - - 64.7 (-69.8+σ) dBm for P_e <= 2.8×10^{-9}

56.2 to 61.3 dB excess loss over 1 m @ 600 mW

Note: The above does not take into account any S/N gain against impulse noise due to peak clipping. This clipping effect will give about a 6 dB gain resulting in 62.3 to 67.3 dB for the excess loss of the last line above.

902 - 928 MHz Interefrence levels

Department Store

Atlanta, GA

Location	Frequency	Level (into dipole antenna)
	MHz	dBm
Entrance, Both Levels	905.7 907.4 902.0 908.0 930.0	-50 -54 -70 -70 -63
Upper Level, Under Skylight	930 Others	-50 <-72
Lower Level, Sporting Goods	930 All Others	-50 <-72
Mobile Voice Transmitter	929.1 (second harmonic)	-56 @ 1 m

IEEE p802.4L

Through-the Air Physical Media, Radio

Microwave Oven Emissions

Donald C. Johnson, NCR

Attached is a spectrum tracing showing the rf emissions of a microwave oven. The measurements were taken using a spectrum analyzer with a dipole antenna 3 m from the oven. The chart is a 1 to 2 minute spectrum accumulation using a 30 kHz resolution bandwidth.

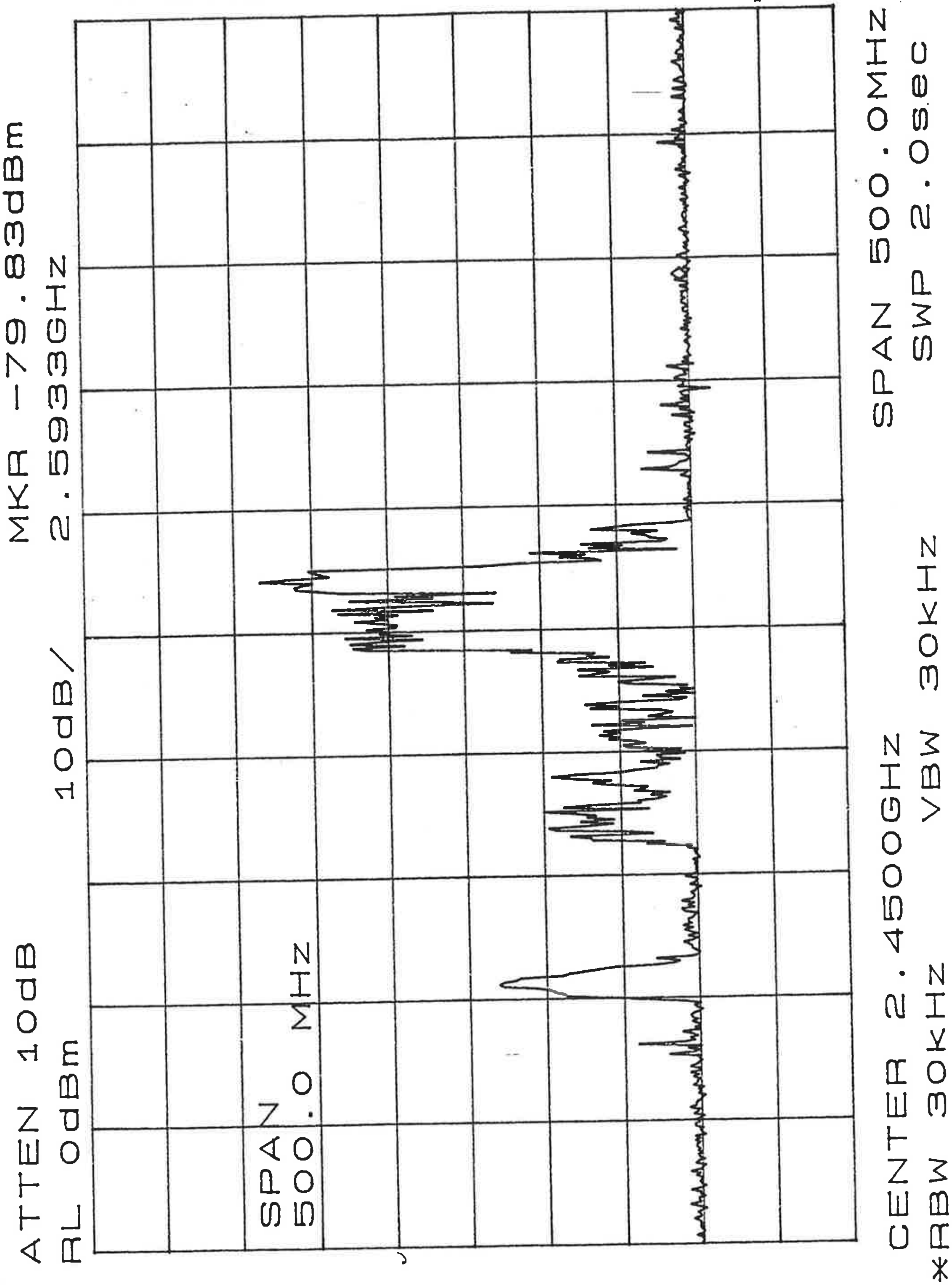

MKR -79.83dBm
2.5933GHz
10dB/
ATTEN 10dB
RL 0dBm
SPAN
500.0 MHz
SPAN 500.0MHz
SWP 2.0sec
CENTER 2.4500GHz
*RBW 30kHz
VBW 30kHz

Chart 1

IEEE p802.4L
Through-the-Air Physical Media, Radio
Questions
First Version

INTRODUCTION

This document summarizes the main questions to be resolved for obtaining a standard for the radio physical medium.

The questions given in this document are the same as the issues stated in the "Positions and Arguments" document currently published as Doc: IEEE p802.4L/90-17 and they carry the same number.

Submissions

Members of the Task-group are requested to make submissions just addressing a single question and to state the number of the question in the heading of the submission in the following way:

Question: 2.1.3.

However, if the subject of the submission nevertheless encroaches more then one question, then all question numbers are to be stated in the heading in the following way:

Question: 3.1 / 4 / 8.

1 *I: What are the Product Requirements of the Radio LAN PHY?

1.0.1 *I:What is the acceptable BER?

2 *I: What is the definition of the transmitted signal?

2.1 *I: What type of modulation (Carrier) should be used?
2.2 *I: What power should be transmitted
2.3 *I: What type of modulation (spreading) should be used?
2.4 *I: Should Forward Error Correction (FEC) be used?
2.5 *I: Is scrambling required?
2.6 *I: How should the MAC symbols be encoded?
2.7 *I: What data rate?
2.8 *I: How is the requirement for limiting phase noise specified?

Questions

*I: What should be specified for the receiver?

*I: What Network Management objects are to be defined?

*I: What are the characteristics of the medium?

.1 *I: What is the characteristic of the RF Propagation?
.2 *I: What can be done to control the RF propagation?

*I: What is the definition of the Distribution System?

.1 *I: How are the forward and reverse channels separated?
.2 *I: What is the format of the PHY PDU?
.3 *I: How is the best reverse radio access point selected?
.4 *I: What is to be specified for the antennae deployment?
.5 *I: What is the transmitted Power?
.6 *I: How is the RF frequency of the forward radio access points controlled?

*I: What should be defined in environmental Specifications?

3 *I: What are the characteristics of noise and likely interferers?

3.1 ?I: How can Field Disturbance Devices (FD devices), in particular shoplifting devcies, co-exist with IEEE p802.4L (RLAN) systems?

9 *I: What is required on labelling?

10 *I: What is to be specified about Coverage Area?

11 *I: What guidelines should be given in the standard?

12 *I: What steps are required to obtain a frequency band allocation dedicated to RLAN?

13 *I: What features are required of p802.4 RLAN to work together with other p802.x RLANs?

IEEE p802.4L

Through-the-Air Physical Media, Radio

Positions and Arguments

First version

INTRODUCTION

This document represents the results of the discussions held in the Radio-LAN task-force. The way of representation is acording to the IBIS method.

OBJECTIVE

The objective of this document is to keep a record of the decisions taken on the many questions that had to be taken along with the arguments on which the task-group based its decision. This way, even rejected positions are kept on file so that when new ideas come up it is easy to recupe our thought process at the time the decision was taken.

Members of the taskgroup are encouraged to submit contributions in the same format.

ABOUT THIS ISSUE

This issue is based on a document prepared by Vic Hayes by regenerating the resukts from the some of the minutes of meeting and the Running Objectives and Directions document. It has partially been reviewed at the May 1990 (Atlanta) meeting. In addition the discussions held at that meeting have been added.

Future contributions are planned for capturing more results from prior meetings.

What is IBIS:

IBIS means Issue Based Information Systems, it presents a method of documenting discussions.

IBIS consists of 3 key elements (nodes): Issues, Positions and Arguments.

Issues :	An Issue articulates a key question.
Position :	A position makes a single point that directly addresses its parent issue.
Argument :	An argument supports or objects to a Position.

IBIS notation (text indentation method):

The textual format of IBIS uses indentation to represent the hierarchical relationship among nodes. The labels used for different types of nodes are:

Issues, Positions and Argument nodes are labeled:

I:	issues
P:	positions
A+:	supporting arguments
A-:	objecting arguments

Each node label is preceded by a status flag.

?	open node (no decision made)
*	Issue resolved, Position or Argument accepted.
-	rejected node.

Issues have been numbered for easy reference

The format is thus:

*I: What is the layout for text indentation IBIS?

  *P: The base Issue at the left margin.

  *P: The Position(s) 1 TAB indented under the Issue.

  *P: The Argument(s) 1 TAB indented under the Position.

.1 *I: Can a node have more than one line of text

    *P: Yes. This position is an example of a position being printed on more than one line.

.2 *I: Can a new issue be raised at all types of nodes.

    *P: Yes it can.
      *A+: Questions can raise everywhere.
.2.0.0.1   ?I: Is this a valid argument?

1 *I: What are the Product Requirements of the Radio LAN PHY?

*P: Meet FCC part 15 rules for spreading, scrambling and Power
- *A+: Permits non-licensed use
- *A-: Spectrum has to be shared with other non-licensed users
- *A-: danger of being forced to stop because Spread Spectrum has the lowest secondary user status.

*P: Economy

*P: Permit system diameter of 300 m
- A+: Feasable under power level permitted by FCC Part 15 rules
  - ?I: Is this only limited to Peer-to-Peer systems?

*P: Minimal acceptable diameter is 100 m

*P: Permit adjacent 802.4-conformant RadioLANs

*P: Support office, retail and industrial environments

*P: Robust with respect to multipath

*P: High data rate at required BER and outage

1.0.1 *I:What is the acceptable BER?

- *P: The BER at the MAC/PHY interface shall be 10^{-8} or less achievable in al but 10^{-3}or less of the area of spatial coverage of the system in a minimally conformant system, and where additional antenna and receiver diversity can be used to reduce the area of outage as required

*P: Accommodate relative motion between Transmitter and Receiver

*P: Meet ISO IS 8802-4, sections 1-10
- A+: a requirement for the IEEE p802.4 group

I: What is the definition of the transmitted signal?

.1 *I: What type of modulation (Carrier) should be used?

-P: Coherent Phase Modulation
*A-: the anticipated coherence time of the channel is insufficient
*A-: the acquisition time is too long compared to the anticipated packet length

*P: Differential (Phase shift) keying
*A+: fast acquisition
*A+ robust for the time varying channel
*A+ simplifies receiver
*A-: theoretical loss in S/N of 2.3 dB

.1.1 ?I: What type of DPSK?

*P: DQPSK
*A+: information is vested in phase change
*A+: allows wave shape form that tolerates non-linear amplifiers
+A+: allows coherent demodulation
*A+: decreases signalling rate (increases symbol time) without excessively compromising S/N (see QPSK)

*P: Quaternary Phase Shift Keying
*A+: decreases signalling rate (increases symbol time) without excessively compromising S/N

?P: $\pi/4$QPSK
*A+: is almost constant envelope
*A+: allows an FM discriminator to be used as demodulator
*A+: decreases signalling rate (increases symbol time) without excessively compromising S/N (see QPSK)
*A-: does not allow Direct sequence spread spectrum before the non-linear elements of the transmitter
*A-: the constant envelope property is lost when the signal is Nyquist filtered
*A+: does not increase cost in complexity compared with other QPSK type modulation techniques

-P: Frequency Modulation
*A+: is 100 % constant envelope
*A+: demodulation is simple
*A-: is not spectrum efficient except for M-ary FSK with M>2
*A-: can not be passed through a filter and maintain constant envelope
*A-: has a clicking phenomena below 10 dB S/N ratio

-P: Amplitude Modulation
*A+: works with minimal channel coherence time
*A-: requires linear amplifiers (expensive)
*A+: economy
*A-: technical risk
*A-: spectrum efficiency
*A-: not robust against fading environments

2.2 *I: What power should be transmitted

*P: less than one Watt

*A+: Required by FCC

2.2.1 *I: Should the transmitter power level be adjustable?

?P: Yes
*A+: permits adjustment to prevent interference with an adjacent LANs
*A-: higher cost
?I: how should the adjustment be controlled?
?I: is this a managed object?

*P: NO
*A+: cost
*A-: may prevent operation with an adjacent LAN

2.3 *I: What type of modulation (spreading) should be used?

*P: Direct Sequence Spread Spectrum (DSSS)
*A+: allows use of FCC 15.126
*A+: limits the effect of interferers
*A+: increases the symboltime to a value beyond multipath
*A+: reuse of code and frequency possible in head-end systems where antennae are separated more than one chiptime
2.3.0.0.1 ?I: does not interchip interference due to multipath increase the necessary separation?
2.3.0.0.2 ?I: would not a fraction of the symboltime offset reuse of the code render unnecessary antenna-space time separation?

*A-: loss of bandwidth

2.3.0.1 *I: What is the best spreading size?
*P: Minimal within FCC rules
*A+: most efficient use of bandwidth

2.3.0.2 *I: How to use the spreading?

*P: Encode the data on a single code
*A+: simple and satisfies the requirement
*A+: allows for the use of short codes
*A+: short codes allow efficient use of bandwidth

?P: Use a long code for the clock and use CDMA with shorter codes to encode data; e.g. 3 data symbols per clock symbol, 8 bits per data symbol 89-
*A+: single complex correlator for clock recovery, data recovery with simple correlators
*A+: allows clock recovery as delay spread approaches or exceeds data code length
*A-: reduces power devoted to data
*A-: requires a linear amplifier over a wide range because it is a high peak-to-rms wave shape

2.3.0.2.0.1 ?I: Is this method feasible in the light of cross correlation components?

2.3.0.2.0.2 ?I: What is the transmitted spectrum?

3.0.3 *I: What spreading code to use?

*P: Use the Barker 11 code: +-++-+++---
*A+: it has bounded auto-correlation
*A+: It has bounded periodic auto-correlation
*A+: It has bounded odd periodic auto-correlation
*A+: It has good spectral properties

-P: Frequency Hopping
*A+: allows use of FCC 15.126
*A+: limits the effect of interferers
*A-: Can not meet data rate requirement because current regulation limits bandwidth to 25 kHz/hop

.3.0.3.1 ?I: Has the FCC changed the rule yet?

*A-: loss of bandwidth

.4 *I: Should Forward Error Correction (FEC) be used?

?P: Should be avoided if possible 89-15

?P: No
*A+: does not help against narrow band interferers 90-13 /11

*A+: Higher effective data rate
*A+: better delay spread tolerance
*A-: should be avoided if possible

?P: Yes
*A+: provides protection against impulse noise
*A+: provides protection against white Gaussian noise
*A-: Lowers the effective data rate because it takes part of the bits transferred unless the bitrate is increased
*A+: improves performance of inexpensive receivers in time variant channels
*A-: increases slot time
*A+: if used correctly there is a net gain against impulse noise and white Gaussian noise
*A+: allows operation in high impulse noise (energy/impulse/symboltime >> -24 dB of E_B) (puncture) environments

?P: FEC support is required in Distribution system and an installation option for the DTEs (mobile stations)
*A+: Higher data rates in benign environments
*A+: fall back available
*A+: application is possible in all bands

:.4.0.1 ?I: At what level is interoperability required?

P: Make FEC transparent to Distribution system, i.e. Header and trailer have a separate FEC in all cases, MAC PDU part has another FEC if used.
A+: single distribution system for all markets
A+: Header and trailer are longer for market not using FEC.

2.4.0.2 ?I: Should the header encoding be the same or nearly the same for 802.3 and 802.4?

2.4.0.3 ?I: Should the information bits in the header and trailer be "in the clear", i.e. systematic code?

2.4.0.4 ?I: What is a suitable coding rate and type?

2.5 *I: Is scrambling required?

*P: Yes
*A+: to smooth the output spectrum

2.5.1 *I: What size polynomial should be used?
*P: 7-bit polynomial (length 127) when the adopted spreading code is < 127 chips **89-11**
*A+: to suit FCC requirements

2.5.1.0.1 *I: What polynomial is to be used?

*P $1+X^{-4}+X^{-7}$ **89-11**
*A+: to suit FCC requirements
*A+ Economy

2.6 *I: How should the MAC symbols be encoded?

*P: Preamble and frame delimiters should be encoded without increasing the signal constellation **89-15**

*P: The non-data symbol in the frame delimiter should be encoded by a different chip sequence: e.g. Barker-11 backwards
*A-: no need to be encoded

2.7 *I: What data rate?

*P: Between 1 Mbit/s and 20 Mbit/s
*A+: Is within the charter of IEEE p802

2.8 *I: How is the requirement for limiting phase noise specified?

*I: What should be specified for the receiver?

*I: What Network Management objects are to be defined?

5 *I: What are the characteristics of the medium?

5.1 *I: What is the characteristic of the RF Propagation?

5.1.1 ?I: What is the coherence time?
?P: the reciprocal of the Doppler frequency spread (effectively the maximum Doppler frequency componant in the channel)
?P: the Doppler spread is proportianal to the square root of the amplitude-Doppler frequency function
A-: is not consistent with textbook
?P: Doppler spread is reciprocal of the Δt span for the non-zero values of the spaced-frequency spaced-time correlation function defined in previous minutes

5.1.2 ?I: What is the importance of coherence time to our subject?
?P: If coherence time is not significantly longer than the symbol time, then a coherent QPSK detection will be difficult
?P: It would be nice if coherence time was >300 symbol times for coherent QPSK. However, it may be possible to work with 50.
?P: It would be nice if coherence time was > 50 symbol times for differential QPSK. However, it may be possible to work with 5.
?P: It would be nice if coherence time was > 5 symbol times for on/off keying. However, it may be possible to work with 2.
?I: Are these figures correct?

5.1.3 ?I: Do we have to decide on a coherent or a differential demodulator?
?P: No, provided the transmitter provides information for differential and coherent demodulators
5.1.3.0.1 ?I: How long preamble is needed for coherent demodulators?
?P: 300 symbols would be really nice, but at least 50 symbols

5.2 *I: What can be done to control the RF propagation?

*I: What is the definition of the Distribution System?

1 *I: How are the forward and reverse channels separated?

-P: Frequency division

*P: Time division

?P: Code division

.2 *I: What is the format of the PHY PDU?

.2.1 ?I: What is to be conveyed in the header of the PHY chunk?
*P: the NWID
*A+: to be able to reject PDUs of other LANs
.2.1.0.1 ?I: how long?
*P: 5 octets
*A+: to prevent the commitment to a registration authority for NWID by adopting the p(international) telephone number
*A-: this is longer than necessary for an administrative mapping
*P: string length of user data field in octets (1 octet long)
*A+: to identify the MAC End delimeter
.2.1.0.0.1 ?I: should this parameter refer to clear stringlength?
*P: Yes
*P: last chunk indication (1 bit)
*A+ to set AGC in strategic position
*P: NAK (3 bits)
*A+: allows management of ditribution system respond to NAK from stations
*P: FEC applied (1 bit)
*A+: to signal the receiver to apply FEC decoding or not
*P: Start delimeter (1 octet)
*A+: to mark the beginning of the header fields
*P: Slice number (4 bits)
*A+: to identify the channel (or channel type) or slot
*P: FEC Parity bits
*A+: to carry FEC information

.2.2 ?I: What is to be conveyed in the trailer of the PHY chunk?
*P: End delimeter (1 octet)
*A-: chunk length is not variable
.2.2.1 ?I: why is it needed?
*P: CRC
*A+: to support the distributed assessment algorithm
*A+: to support the MAC protocol

6.3 *I: How is the best reverse radio access point selected?

6.4 *I: What is to be specified for the antennae deployment?

6.5 *I: What is the transmitted Power?

6.6 *I: How is the RF frequency of the forward radio access points controlled?

6.6.1 *I: How is the coherence time controlled?

6.6.2 *I: How is the emission pattern controlled?

6.6.3 *I: How is the requirement for limiting phase noise specified?

*I: What should be defined in environmental Specifications?

*I: What are the characteristics of noise and likely interferers?

1 ?I: How can Field Disturbance Devices (FD devices), in particular shoplifting devcies, co-exist with IEEE p802.4L (RLAN) systems?

?P: FD restricts to 902-905 MHz, RLAN restricts to 905-928 MHz
A+: Allows mutual co-existance
A+: is consistent with Sensormatic's positon relative to consumer devices
A+: installed base of FD to be equipped with new crystal
A-: RLAN has to give-up 3 MHz
A-: the remainder of the users of the band still utilize 902 - 928 MHz

?P: FD use the RLAN RF signal
A+: RLAN has an additional radiator
A+: is a synergistic solution
A-: FD has to swap pedestals of installed base
A-: logistically unmanageable

?P: FD restricts to 902-905 MHz, RLAN uses 902-928 MHz
A+: RLAN keeps full bandwidth
A+: In the 4 kHz band of the FD, RLAN merely transmits 0.16 mW(/4 kHz). This may even be lower due to the spectral curve
1.0.1 ?I: What is the saturation power level for the FD device?
1.0.2 ?I: What is the bandwidth of the front-end stage of the FD device?
1.0.3 ?I: If RLAN saturates the FD device, what is the duty cycle / rest-time required from RLAN to co-exist?

?P: Use FD device in the notch at Carrier frequency
A+: Allows co-existance
A-: RLAN has to commit to the notch at Carrier frequency, i.e. at 915 MHz
1.0.0.1 ?I: Does the carrier frequency move?
?P: Carrier of the distribution system will not move
A+: it is a broadcast internal to the system
?P: Carrier of the remote stations will move
?I: Does the moving carrier hurt the performance?
A+: High-pass filter in the receiver is required anyway

?P: FD restricts to 902-905 MHz, RLAN filters the 902-928 MHz asymetrically such that the cut-off at the lower side is less steep and starts higher than required (Vestigial sideband)
A-: the correlator needs some or extensive equalization
A-: extensive filtering required at RF

*I: What is required on labelling?

10 *I: What is to be specified about Coverage Area?

11 *I: What guidelines should be given in the standard?

*I: What steps are required to obtain a frequency band allocation dedicated to RLAN?

3 *I: What features are required of p802.4 RLAN to work together with other p802.x RLANs?

IEEE 802.4L

Through-the-Air Physical Media, Radio

Running Objectives and Directions Document

Eighth Version

The IEEE 802.4L Task Group maintains a group of documents to record the discussions held to date.
This document provides the PAR contents, points to the other documents and documents factual information submitted to the group..
Another document provides the questions to be answered to guide the group to the end-result.
The third document is a record of the discussions in IBIS format. IBIS provides the features: a) to capture the arguments that lead to certain decissions and b) keeps rejected positions in the information system.
~~Each decision will be marked in this document along with the reference to the motion on which the decision has been based (column Base) and with the reference of the document on which the present decision is based (Doc no).~~
After each meeting new documents will be prepared to reflect the decisions made at the meeting.
Current document:

IEEE p802.4L/90-16	RLAN Standard, Questions
IEEE p802.4L/90-17	RLAN Standard, Positions and Arguments

Table of Contents

Subject	Base	Doc no

1. Scope

To define an alternative Physical Layer for Through-the-air communication, which is part of a local area network using 802.4 media access techniques and which is primarily for mobile environments.	PAR	4L/87-014

2. Purpose

To provide LAN access to moving automatic machines and other stations for which wireless attachment is appropriate.	PAR	4L/87-014
To add description of standards criteria for through-the-air transmission parameters to support Physical Layer Service.		
To prepare, if necessary, a petition to the FCC for rule making which authorizes use of radio spectrum for wireless LAN.		

3. Directions

The entries in this section have been replaced by the Doc.: IEEE p802.4L/90-17 section 1. Please review the completeness of Doc 17.

~~3.1 Design Principles~~

~~1. Meet FCC rules - spreading, scrambling, power, etc.~~	~~Jul 89~~	~~4L/89-11~~
~~2. Meet 802.4 requirements implicit in ISO DIS 8802-4-1-10~~	~~Jul 89~~	~~4L/89-11~~
~~3. Economy~~	~~Jul 89~~	~~4L/89-11~~
~~4. Permit adjacent 802.4L conformant radio LANs~~	~~Jul 89~~	~~4L/89-11~~
~~5. Provide for both single channel (direct peer-to-peer) and dual channel (head-ended) operation~~	~~Jul 89~~	~~4L/89-11~~
~~6. Single-channel system size: The objective is to permit a system diameter of 300 m. The minimum acceptable system diameter is 100 m.~~	~~Jul 89~~	~~4L/89-11~~
~~7. Modulation technique must support office, retail and industrial environments.~~	~~Jul 89~~	~~4L/89-11~~
~~8. Want high data rate at required BER and outage.~~	~~Nov 89~~	~~4L/89-17~~
~~9. Robust with respect to multipath~~	~~Nov 89~~	~~4L/89-17~~
~~10. Want to accommodate relative motion between Transmitter and Receiver~~	~~Nov 89~~	~~4L/89-17~~
~~11. For a given operating band (902-928 MHz, 2400-2483.5 MHz, 5725-5875 MHz), want the interoperability relationship of differing modems to form a direct inclusion relationship (full and not partial ordering).~~	~~Nov 89~~	~~4L/89-17~~

~~points of interoperability.~~

~~finish definition of the primary air interface before considering any other interfaces.~~

~~3.2 System plan~~

~~The radio system plan for one community of users is proposed to be a single frequency bus mode with head end, but will accomodate single frequency station to station operation for small systems. The physical layer including the head end and radio system shall support the existing 802.4 MAC. (Among other things, this implies that when any station is transmitting, all stations must hear something.)~~	~~Jan 89~~ ~~Jul 89~~	~~4L/89-02~~ ~~4L/89-11~~
~~In the single frequency bus mode with head end normal token rotation shall be used, only for stations in the outskirt, immediate response mode will be considered. (see issue 5)~~	~~May 89~~ ~~Jul 89~~	
~~Whatever plan is evolved, it shall be suitable for use under current FCC part 15 regulations, in particular the three bands, 902-928 MHz, 2400-2483.5 MHz, and 5725-5875 MHz.~~	~~Jul 88~~	

Subject	Base	Doc no

3.2 Directions (cont..d)

Subject	Base	Doc no
~~The 902-928 MHz band will be used in the first standard. At least 2 channels will be accomocated in the band.~~	Jan 90	

~~3.2 System plan (cont..d)~~

Subject	Base	Doc no
~~To separate transmissions of stations of nearby networks.the preamble will contain a Network Identification.~~	~~May 89~~ ~~May 89~~	

~~3.3 System Design Parameters~~

Subject	Base	Doc no
~~Relation to the Objective List in [3.1]~~	~~Jul 89~~	~~4L/89-11~~
~~1. Use a 7-bit (length 127) scrambler if the adopted chip rate is < 127. [1] The preferred polynomial is~~ $\cancel{1 + X^{-4} + X^{-7}}$ ~~.[1+3]~~		
~~2. Choose a modulation technique that does not include an amplitude modulation component, for [3] and to lower technical risk.~~	~~Jul 89~~	~~4L/89-11~~
~~3. Permit differential demodulation for fast acquisition, to provide robustness for the time-varying (fading) radio channel, and to simplify the receiver [3]. The primary disadvantage of this approach is a 2.3 dB (theoretical) loss in S/N.~~	~~Jul 89~~	~~4L/89-11~~
~~4. Use some form of quaternary PSK as a reasonable means of decreasing signaling rate (for multipath) without excessively compromising S/N or [3,7].~~	~~Jul 89~~	~~4L/89-11~~
~~5. Spread the minimum amount practical [1,3]. The preferred spreading code is + - ++ - + + + - - - . This is a known Barker code, with bounded auto-correlation, bounded periodic auto-correlation, and bounded odd-periodic auto-correlation, and good spectral properties.~~	~~Jul 89~~	~~4L/89-11~~
~~6. Filtering should consider adjacent-channel single frequency (single channel) and simultaneous dual frequency (dual-channel) operation. [4,5]~~	~~Jul 89~~ ~~Jan 90~~	~~4L/89-11~~ ~~4L/90-01~~
~~7. Initial focus should be on 902-928 MHz band. [3]~~	~~Jul 89~~	~~4L/89-11~~
~~8. The design goal for the overhead of each **Ph-PDU be 25 octets** or less. This includes synchronization pattern, network id, CRC on the Ph-PDU content, and FEC flush. Note that the overhead can be different for the forward and reverse channel.~~	~~Mar 90~~	~~4L/90-5~~

~~3.4 Modulation~~

Subject	Base	Doc no
~~Differential Phase Modulation shall be used.~~	~~Nov 88/1~~	~~4L/88-02~~
~~Doc: IEEE p802.4L/89-16 is adopted as the basis for the description of the modulator.~~	~~Nov 89~~	~~4L/89-17~~
~~For the spreading sequence at least 10 and not more than 15 chips shall be used. This provides a processing gain of between 10 and 15 allowing frequency division multiplexing of co-located LANs~~	~~Nov 88/3~~	~~4L/88-02~~

~~3.5 Encoding~~

Subject	Base	Doc no
~~The goal is to encode the preamble and the frame delimiters without increassing the signal constellation.~~	~~Sep 89~~	~~4L/89-15~~
~~It is suggested to encode the MAC non-data symbol by a different chip sequence (e.g. Barker 11 backwards).~~	~~Sep 89~~	~~4L/89-15~~

Subject	Base	Doc no
Directions (cont..d)		
~~3.6 Data Rate~~		
~~The data rate for comparison purposes shall be 1 Mbit/s. We can only consider the IEEE data rates of 1 to 20 Mbit/s.~~	~~Jan 89~~	
~~3.7 Distribution System~~		
~~The design model shall assume a 16 antenna array in a square grid. For purpose of analysis, it will be assumed that the antenna array is driven by one power splitter with equal length loss less cable from the splitter to each antenna.~~		
~~3.8 Performance definition~~		
~~The performance of the Token Bus standard will be expressed in the number of MAC Service Data Units with undetected errors per time unit, at 0 frame overhead.~~	~~May 89~~	
~~The performance requirement is: less then one MSDU with undetected errors per year at 200 bit data units.~~		
~~The frame loss rate shall be less then 1 per 10^{-8} frames transmitted.~~		
~~3.9 Bit Error Ratio~~		
~~The Bit Error Ratio (BER) at the MAC/PHY interface shall be 10^{-8} or less achievable in all but 10^{-3} or less of the area of spatial coverage of the system in a minimally-conformant system, and where additional antenna and receiver diversity can be used to reduce the area of outage as required.~~	~~Sep 89~~ ~~Jan 90~~	~~4L/89-15~~ ~~4L/90-01~~
~~3.10 Outage~~		
~~MAC protocol assumes the communication channel is always available. Since the radio medium is known to have an outage rate on the order of 10E-2, a method is required to reduce outage rate to less than 10E-5.~~	~~Jul 88~~	
~~3.11 Velocity ranges~~		
~~The following are the ranges for the velocity of the stations:~~	~~Jan 89~~	
~~902-928 MHz — 0 - 53.7 miles/h~~		
~~2400-2483.5 MHz — 0 - 20.0 miles/h~~		
~~5725-5875 MHz — 0 - 8.3 miles/h~~		
~~3.12 Transmission Power~~		
~~XMTR power output: — 1 W max~~	~~Jan 89~~	
~~Station antenna gain: — TBD~~	~~Jan 89~~	
~~Station antenna directivity: — TBD~~	~~Jan 89~~	
~~Receiver noise figure: — 6 dB at 902-928 MHz~~	~~Jan 89~~	
~~8 dB at 2400-2483.5 MHz~~	~~Jan 89~~	
~~10 dB at 5725-5875 MHz~~	~~Jan 89~~	
~~For a distributed antenna system, we assume that each transmitter should be measured separately (for complying with the regulation). The transmit carriers should not be phase locked but should be approximately the same frequency.~~	~~Nov 89~~	~~4L/89-15~~

Subject	Base	Doc no

Directions (cont..d)

~~3.13 Error correction codes~~

~~The goal is to avoid the use of Forward Error Correction code, if possible.~~	~~Sep 89~~	~~4L/89-15~~
~~Allowable overhead: 1.2x~~	~~Jan 89~~	
~~Type: TBD~~	~~Jan 89~~	
~~Spectral efficiency: TBD~~		

3.14 Propagation

Office/retail environment: **6 dB/octave under 10 meters**

Local Spatial Correlation (LSC) is defined as follows:

Let,

$A(\tau,x)$ = The signal attenuation of the impulse response amplitude in dB at excess time τ and position x.

$M_A(\tau,x)$= The local spatial average of the signal attenuation at excess time τ in the vicinity of location x.

Rappaport reports that $A(\tau,x)$ was found to be approximately normally distributed with a mean of $M_A(\tau,x)$. The local spatial correlation (LSC) is,

$$LSC(\tau,\Delta x) = \frac{E[(A(\tau,x)-M_A(\tau,x))(A(\tau,x+\Delta x)-M_A(\tau,x))]}{E[(A(\tau,x)-M_A(\tau,x))^2]}$$

Much the same as for the coherence time and the spaced-time correlation function, coherence distance could be defined as the value of Δx at which LSC becomes = 0. The local spatial correlation is about 0.2 at $\lambda/4$ and effectively 0 at $\lambda/2$ at nearly all values of excess delay. Thus coherence distance is approximately $\lambda/2$ in the Rappaport measurements.

environment	20 meter attenuation relative to 1 meter (dB)	slope (dB/octave)	standard deviation (dB)	RMS delay spread (within 20 dB from max peak) (ns)	notes	Local Spatial Correlation	Coherence Time
open retail	29-35	10-13.8	2.1-5.3	10-150	1		
obstructed retail	40	19.4	4.5	not measured	2		
factory	25-32	5.7-7.3	4.8-10.2	30 min 160 (95%) 280 max	3		
office	39 1 location	11.7 1 location	2.2 1 location	10-50	4		

Note 1: The open retail environment consists of a typical department store or supermarket with no more than 1 floor-to-ceiling wall in any path. Some otherwise shaded paths are included. These include paths shaded by elevator shafts and by concrete columns as well as merchandise and displays in the line-of-sight paths. The size varies from 21 meter maximum linear dimension to 110 meters maximum linear dimension.

The lowest delay spreads were measured in a small supermarket. These delay spreads were measured indirectly using the coherence bandwidth method. The variation of 4 measurements was 8 to 20 ns (coherence bandwidth of 8 to 20 MHz). The larger delay spreads were measured using the direct impulse response power delay profile. Values in large department stores are 50 to 150 ns.

The attenuation statistics (first 3 columns) were taken with CW measurements and were recorded separately for each location. The first 2 column parameters were computed by finding the set of values which minimized the standard deviation (third column). The standard deviation is that of the deviation from a regression line of 0 dB at 1 meter and 6 dB/octave (straight line against log distance) from 1 meter to the point where the low slope line intersects the higher slope line. An iterative procedure was used which varied the slope and 20 meter attenuation of the higher slope segment for minimum RMS deviation.

Note 2: The obstructed retail location was a department store with multiple floor-to-ceiling walls. Wall attenuation was measured at approximately 6 dB/wall. The maximum linear dimension of this store was 100 meters. There were approximately 10 walls in the longest paths.

Note 3: The factory information is from the report *Characterization of UHF Factory Multipath Channels* by Theodore S. Rapport and Claire D. McGillem, School of Engineering, Purdue University, West Lafayette, Indiana 47907, TR-ERC-88-12.

5 Light to heavy manufacturing locations were measured.

The attenuation statistics (first 3 columns) differ from the retail and office statistics in the manner in which the large scale loss curve fit was computed. The 10λ distance is the reference. The curve (regression line) was forced to 0 dB at the reference point and there is only one curve segment. The slope (second column) of the regression line is the value which minimizes the standard deviation (third column). The principal difference is that the regression line for the retail and office statistics was not forced to a particular point, but was allowed to vary in the vertical dimension to further minimize the standard deviation. Thus, the standard deviation of the factory measurements can be expected to be higher than that which would be determined by the retail and office environment method and the slope can be expected to be different. Rappaport reported that the techniques differ by no more than 0.2 dB in standard deviation and 1.5 dB/octave in slope.

Rappaport also computed attenuation values from the 50 wideband measurements made over the 5 sites. The attenuation values were computed from the impulse response power-delay profiles. The result for all sites was:

20 meter attenuation relative to 1 meter (dB)	slope (dB/octave)	standard deviation (dB)
26	6.5	4.9

The delay spreads were from the 50 wideband measurements. They were further broken down into those for obstructed paths (OBS) and line-of-sight (LOS) paths:

	Min.	50 Pctl.	95 Pctl.	Max.
OBS	30 ns	110 ns	140 ns	155 ns
LOS	30 ns	90 ns	150 ns	280 ns

The second largest RMS delay spread was 155 ns.

Note 4: The office environment information is from measurements of an engineering office location in The Netherlands and from the article *A Statistical Model for Indoor Multipath Propagation* by Adel A. M. Saleh and Renaldo A. Valenzuela, IEEE Journal on Selected Areas in Communications, Vol. SAC-5, No. 2, Feb., 1987.

The attenuation measurements are from the office location. *(Note: I do not have a copy of the article. Further attenuation information is probably included. Perhaps this can be added later)*

The maximum office delay spread (50 ns) is the maximum reported by Saleh and Valenzuela. In addition, the coherence bandwidth was measured for two paths in the office location. Coherence bandwidths were 8 and 16 MHz, corresponding approximately to RMS delay spreads of 20 and 10 ns.

Subject	Base	Doc no

Directions (cont..d)

Coherence time is defined as follows:
Given a time-variant (wide-sense stationary) channel impulse response of

$$c(\tau;t) = \alpha(\tau;t)\ e^{-j2\pi f_c \tau},$$

where τ is the delay and $\alpha(\tau;t)$ is the attenuation of the signal components at delay τ at time instant t.

Let $C(f;t) = \int_{-\infty}^{+\infty} c(\tau;t)\ e^{-j2\pi f \tau d \tau}\ dt$ be the Fourier transform of this impulse response.

$$\phi_c(f_1,f_2;\Delta t) = 1/2\ E\ [(C^*(f_1;t)\ C(f_2;t+\Delta t)] = \phi_c(\Delta f;\Delta t),$$

where E is expectation, is called the spaced-frequency spaced-time correlation function.

If you hold Δf to 0 you have the spaced-time correlation function. The period of time over which the magnitude of this function is essentially non-zero is the coherence time of the channel.

Table prepared	Nov 89	4L/89-17
Table updated	Jan 90	

Noise: Jan 89

at 902-928 MHz	10 dB above thermal	
at 2400-2483.5 MHz	thermal	Jan 89

Contributions on noise are requested in the following format:

Device	Band	distance from source	Power *) level	Number of hits per second Threshold			
				-10 dB	-20 dB	-30 dB	-40 dB
		m	dBm				

Table 2. Characteristics of impulsive noise generators

Table prepared Nov 89

Directions (cont..d)

Device	Freq	Power		Bandwidth	Duty cycle
		EIRP	Receive level		
	MHz	W	dBm	kHz	
(1)	(2)	(3)	(4)	(5)	(6)
Pager	931.6125	340		15	5 sec/call 1 call/5 min
Radio Channel	904			30	continuous
Pager	930.0		- 50 indoor	15	5 se/call 1 call/min
Field disturbance sensors	902-928	0.075		<1	continuous
Part 15 devices	902-928 2400-2483.5 5725-5875	.00075			
Digital oscillators					
Digital devices					

Table 3. Characteristics of Constant Wave Interferers

Subject	Base	Doc no
Directions (cont..d)		
	Nov 89	4L/89-17
NOTES: * reference antenna : dipole for the appropriate band distance from source > 1 m vary measurements over a sphere with at least 10 measurements * for impulsive noise measurements: make the measurements in the time domain * for CW measurements: include a graph of frequency versus time behavior for sweeping devices, e.g. microwave ovens. It appears that the magnetron has a negative resistance on turn-on and turn-off, and this causes relaxation oscillations at the beginning and end of each power cycle, which cause an apparent broadband emission. In reality, during the beginning and end of each power cycle, the magnetron produces a series of very short bursts of carrier (<< 300 ns each) with decaying power and a frequency which changes slightly during the burst, and with more substantial changes in frequency from one burst to the next. In the middle of each power cycle the magnetron just stays on, with occasional instantaneous frequency changes due to shifts in mode-locking caused by the changing magnetron plate voltage and the motion of the stirrer in the oven cavity. (See addendum L1, and IEEE 802.4L-89/19 for time domain pictures of this phenomenon.) These instantaneous changes may be accompanied by additional bursts. (See IEEE 802.4L/90-8a figure 4-46.)	Jan 90	4L/90-01
<u>3.15 Antenna</u>		
	Jan 89	
<u>3.16 Higher Layer concerns</u>		
When considering the use of the immediate response mode for stations in the outskirts of the coverage area, thus avoiding the higher probability of losing the Token, the implication is that a station can use only the responder services of LLC type 3. Use of LLC types 1 or 2, or the initiator services of LLC type 3, will cause the station to try to get and later pass the token.	Sep 89	4L/89-15

4. Meeting Plan

Type	Dates	Place	Objective
Interim	May 14-18, 90	Atlanta, GA	Prepare second 802.4 draft
Plenary	Jul 8-13, 90	Denver, CO	Second 802.4 draft
Interim	Sep 10-14, 90	Oshawa(Ont) Canada	Prepare 802.4 Voting draft
Plenary	Nov 11-16, 90	Kauai, HI	802.4 Ballot
Interim	Jan ..-.., 1990	?	prepare TCCC voting draft
Plenary	Mar 11-15, 1991	East coast	TCCC Ballot
Interim	May ..-.., 1991	?	Prepare Final draft
Plenary	Jul 8-12, 1991	West Coast	Final Draft
Plenary	Nov 11-15, 1991	Ft Lauderdale, FL	PM

5. Possible Document Outline

20. Radio Bus Physical Layer

    20.1 Nomenclature

    20.2 Object

    20.3 Compatibility Considerations

    20.4 Operational Overview

    20.5 General Overview

    20.6 Application of Network Management

    20.7 Functional, Electrical and Mechanical Specifications

    20.8 Environmental Specifications

21. Radio Bus Medium

    21.1 Nomenclature

    21.2 Object

    21.3 Compatibility Considerations

    21.4 General Overview

    21.5 Functional, Electrical and Mechanical Specifications

    21.6 Environmental Specifications

    21.7 Transmission Path Delay Considerations

    21.8 Documentation

    21.9 Network Sizing

    21.10 Guidelines

6. Issues

The entries in this section have been replaced by the Doc.: IEEE p802.4L/90-17 and -16 . Please review the completeness of these documents.

~~1 Is a Bit Error Ratio (BER) of 10**-8 detected and 10**-9 achievable with operation with a dual frequency head-end distribution system.~~
~~2 Is the BER described in issue 1 achievable for direct station-to-station operation and what is the condition to achieve this BER.~~
~~3 What Forward Error Correcting Code (FEC) is suited for channels with burst errors characteristics.~~
~~4 Considering the agreement that non-data will not be encoded as a PHY symbol: Find a method of start and end delimeter encoding, e.g. use a combination of an alternative constellation and corrolation.~~
~~4a What is the characteristic of the impulse noise in the various media.~~
~~5 What are the implicatios on the LLC when the immediate response mode is required to communicate with stations in the outskirt?~~
~~6 How should a distributed antenna system be represented for ruling measurements.~~
~~7 What are the trade-offs in data rate vs noise immunity (long vs short codes) [refer to doc: IEEE p802.4L/89-17, pages 6-8]~~
~~8 What are the trade-offs of long codes vs short codes at higher frequencies (wider bands) and multiple channels (FDM vs CDM) [refer to doc: IEEE p802.4L/89-17, pages 6-8]~~
~~9 What are the noise characteristics for various devices [refer to tables 2 and 3 above]~~
~~10 Is table 1 above accurate?~~
~~11 Data on coherence time is needed. Part of the data could be recovered from Oshawa measurements and from Rappaport's report. More measurements are to be made when the results prove some parameters have been missed.~~

7. Referenced papers.

The following papers are of interest to the taskgroup members:

Environmental Monitoring for Human Safety Part 1: Compliance with ANSI Standards. By John Coppola and David Krautheimer, Narda Microwave Corporation. - RF Design--.
RF Radiation Hazards: An update on Standards and Regulations. By Mark Gomez, Assistant Editor, and Gary A. Breed, Editor. - RF Design, October 1987
RF Radiation Hazards: Power Density Predeiction for Communications Systems. By Gary A. Breed, Editor. - RF Design, December 1987
Microprocessor Interference to VHF Radios. By Daryl Gerke, PE Kimmel Gerke & Associates, LTD. - RF Design, March 1988
Distributed Antennas for Indoor Radio Communications. By Adel A.M. Saleh, A.J. Rustako, Jr and R.S. Roman. - IEEE Transactions on Communications, Vol. Com-35, No12, December 1987
UHF Fading in Factories. By Theodore S. Rappaport and Clare D. McGillem. - IEEE Journal on selected Areas in Communications. Vol. 7. No 1. January 1989
Indoor Radio Communications for Factories of the Future. By Theodore S. Rappaport. - IEEE Commmunications Magazine. May 1989.
A differential offset QPSK modulation/demodulation technique for point-to-multipoint radio systems. By Tho Le-Ngoe. GLOBECOM 87.
Highly Efficient Digital Mobile Communications with a Linear Modulation Method. (p/4 QPSK) By Yoshihito Akaiwa and Yoshinori Nagata. - IEEE Journal on Selected Areas in Communications. Vol. SAC-5, No. 5, June 1987, pp.890-895.