Minutes of the IEEE p802.4L Working Group

Parsippany, New Jersey
January 16-20, 1990

Intermediate Meeting.

Chairman. Vic Hayes

Secretary & Editor. Michael Masleid, Chuck Thurwachter, Tom Phinney.

Attendance

Mr. VICTOR HAYES	NCR Systems Engineering B.V	phone +31 3402 76528
Mr. MICHAEL MASLEID	Inland Steel Co. MS2-465	phone 219 399 2454
Mr. THOMAS L. PHINNEY	Honeywell	phone 602 863 5989
Mr. OREST L. STOROSHCHUK	General Motors of Canada	phone 416 644 6994
Mr. DONALD C. JOHNSON	NCR Corporation WHQ 5E	phone 513 445 1452
Mr. LARRY van der JAGT	Knowledge Implementations Inc	phone 914 986 3492
Mr. ROBERT S. CROWDER	Ship Star Associates Inc	phone 302 738 7782
Mr. NEIL WELLENSTEIN	Motorola Inc	phone 602 952 3436
Mr JAMES. NEELEY	IBM	phone 919 543 3259
Mr. PAUL PIRILLO	NCR E&M Atlanta	phone 404 623 7505
Mr. JONATHAN CHEAH	HUGHES Network Systems	phone 619 453 7007
Mr. STAN KAY	HUGHES Network Systems	phone 301 428 7165

Tuesday 90.01.16

Vic Hayes opened the meeting at 0845. The time until 0900 (the published start time) was devoted to introductions. 11 people were in attendance - 3 from NCR (3 sites), 3 from GM and Hughes/GM (3 sites), and 5 others.

The minutes of the Ft. Lauderdale meetings were reviewed. A number of corrections were made:

- page 7, line 1: "9281" should read "928"
- page 7 paragraph 5 should read "the symbol time must exceed the peak"
- page 7 paragraph 9 should read "b) for fixed energy per symbol, energy per bit ..."
- page 7 paragraph 11 should read "if non-Gaussian noise power density ..."
- pages 10 -15: A number of changes were recorded.

Tom Phinney moved adoption of the minutes as corrected. Larry van der Jagt seconded. Carried 8-0-3.

Orest Storoshchuk stated that the 930 MHz interferer in the Oshawa testing (referred to as "Beethoven's Fifth" in the minutes of the Chicago meeting) turned out to be a 931.6125 MHz paging transmitter for the Oshawa region, located on the plant roof near column L44, with an EIRP (Effective Radiated Power) of 340 W.

Tom Phinney stated that there were no known Barker codes beyond Barker-13, and that any longer Barker codes, if they do exist, are known to have a length so long (> 64 k) as to be unuseful.

The "Running Objectives and Directions Document" was reviewed. Items 2, 3, 5, 6 and 7 from page 5 of the Ft. Lauderdale meeting minutes were added to clause 3.1. The sentence "At least 2 channels will be accommodated in the band." was deleted from clause 3.2.

Bob Crowder, who had stayed after finishing the 802.4J4 meeting, departed for the Wilds of Deleware.

Tom Phinney moved the removal of item 6 from clause 3.3. Larry van der Jagt seconded. After prolonged discussion, Tom Phinney moved to call the question. Mike Masleid seconded. The motion to call carried 9-0-0. The motion to remove then carried 8-0-2.

The words "the modulator" were changed to "this modulator" in the second line of clause 3.4.

The meeting broke for lunch at 1300 and reconvened at 1400.

Table 4 of 802.4L/89-17 was discussed. Mike Masleid pointed out that DQPSK could also be used without FDM in the 2.4 GHz and 5.8 GHz bands, giving 7000 and 12000 k bit/s capacity, respectively.

A long discussion of impulse noise ensued, with the following conclusions:

A) Impulse noise is coherent in phase vs. measurement frequency, while white noise is uncorrelated in phase vs. measurement frequency.

B) For impulse noise the output voltage of a filter goes up linearly with the filter bandwidth.

C) In a system with no clipping, where CV = the voltage at the output of the correlator:

1) $CV_{impulse}$ is proportional to the chip rate

2) $CV_{impulse}$ is invariant with the code length

3) CV_{signal} is proportional to the code length

Therefore

4) S/N after the correlator is directly proportional to the symbol duration (or, alternatively, is inversely proportional to the symbol rate)

(See addendum 1 for supporting material)

D) In a system with a clipping level substantially above the expected receive level, so that the receiver is still linear for signal plus Gaussian noise in the absence of impulse noise, then this clipping can be used to limit the impact of impulse noise on the output of the correlator.

Thus for a constant symbol signaling rate, increasing the processing gain (and thus the required bandwidth), with a non-linear clipping process at the input to the correlator, improves the ability to reject impulse noise of an amplitude greater than the clipping level.

It was pointed out that in this committee the chairman was unlikely to influence the votes of other committee members, and so his personal voting was encouraged.

Larry van der Jagt moved the acceptance of the prior two paragraphs as a conclusion. Tom Phinney seconded. Carried 10-0-0. *(See addendum two for further confusion)*

In Table 5, last column,

- for the 127 row, change 0? to <21
- for the 255 row, change 0? to <27

Tom Phinney moved adjournment for the day at 1825. Jim Neeley seconded. Carried 10-0-0.

Wednesday 90.01.17

The meeting was called to order at 0810. 11 people were in attendance. Mike Masleid summarized the discussion of the prior afternoon, and then addressed **multiple impulses within the period of a single symbol**. A long discussion resulted in the conclusions that increasing the symbol duration resulted in a proportional improvement in S/N, which is just Eb/No, but that for isolated large impulses (with no more than one large impulse per symbol), the improvement is Eb**2/No.

Increasing the chip rate for a constant symbol duration provides no improvement unless a clipping process is introduced at the output of the pre-correlator low-pass filter, in which case there is an improvement in the system's ability to reject impulses; for widely-spaced large impulses in a non-dispersive channel, constant envelope modulation, and optimal clipping, this improvement can approach the ratio of increase in the chip rate. Smaller or more-frequent impulses or less-optimal clipping result in less of an improvement.

The number of resolved paths (N_{RP}) equals the channel delay spread divided by the chip duration, or equivalently the delay spread times the chip rate, both rounded up. N_{RP} cannot exceed the number of taps. Also, for simple system design the symbol duration should not be less than the channel delay spread. The worst-case peak-to-average voltage ratio is sqrt(N_{RP}); a more realistic expected worst-case peak-to-average voltage ratio is

sqrt(ceiling(150 ns/chip duration)).

Thus the clipping threshold must be increased when the chip rate is increased.

The meeting broke for lunch at 1245 and reconvened at 1350. The attendees reintroduced themselves, since a new person (Neil Wellenstein) had joined earlier in the day. The discussion of Table 5 of 802.4L/89-17 resumed.

In Table 5, change last two rows as follows:

- for the 127 row, change `0...10.6` to `10-21` and `0?` to `<10`
- for the 255 row, change `0...13.7` to `13-26` and `0?` to `<13`

In general, the power of a single line interferer is suppressed by a factor equal to the spreading factor. However, for some frequencies of line interference, the interfering power will be suppressed by a lesser amount. *(See addendum 3)*

If the density of line interferers in a band is linear with bandwidth, then spreading does not assist in rejecting such interference. However, in most environments this is not the case.

In Table 5, change last four columns of last two rows to:

10	87	-??..10	60.5	10..21	87	<21	60.5
13	43	-??..13	121	13..27	43	<27	121

Stan Kay undertook to redo Table 5 with the objective to make the entries more generic and to improve the lay-out by changing the meanings of the rows and columns.

Note: *the result is in addendum 11.*

The discussion then turned to Table 4. The conclusions were that increasing the chip rate, when the chip period was less than the channel delay spread, provided more path diversity, which can be used to increase resistance to fading due to path cancellation.

Also, for isotropic transmit and receive antennae, the effective path loss, including losses from both antennae, increases with the square of the channel frequency.

Thursday 90.01.18

After waiting for the note-taker and handling some organizational matters, the technical part of the meeting began at 0835. 11 people were in attendance.

A discussion of other uses of the bands concluded:

> Use of the 900 MHz band may interfere with field-sensing devices such as Sensormatic shoplifting detectors. There are other devices which use the band, such as some 15-year-old GE microwave ovens, which may interfere with our use of it. Amateur radio is also authorized to use the band. Paging systems also appear to be just out of band, between 930 and 932 MHz.
>
> In one case, what appears to have been a base-station to commercial radio transmitter point-to-point link was observed in-band at 904 MHz.
>
> Most microwave ovens use the 2.4 GHz band.

All participants in this effort should look for potential interferers and interferees in these bands, and report the results of any surveys which they conduct.

The issue of whether the FCC will require spreading >= 127, or will treat spreading of 11 of information scrambled by a PN code of period 127 or greater, was discussed again. Jonathan Cheah advocated written communication with the FCC to clarify whether an 11-chip code with a period-127 scrambler would be acceptable.

It was suggested that the eventual 802.4L standard contain a recommendation that the transmit power level be adjustable, so that the interference potential of conformant devices could be reduced where necessary.

The data rate of the radio system was discussed. It is desirable that the modulation technique permit fallback to a lower speed when the environment required. This would be done by keeping the same chip rate and changing the spreading code where required.

The potential need for working at sub-Mb/s rates was discussed. Different modulation techniques were discussed, with consideration of the implied cost of components. CTIA investigated $\pi/4$ QPSK for digital cellular, because it has the potential for an inexpensive discriminator.

It was noted that at multi-GHz frequencies, most components operate class C and that this limits usable modulation techniques. The discussion then turned to forward error correction (FEC), and recessed for lunch at 1235.

The meeting restarted at 1415, and the discussion of FEC resumed. The conclusion was that a system which transmits at a constant chip rate could improve its S/N ratio through a combination of reducing the effective information rate and applying the remaining available signaling capacity to an FEC code. For example, halving the information rate and using a rate 1/2 FEC can provide a gain of 8.5 dB against Gaussian noise, a roughly equivalent gain against impulse noise (more if impulses are infrequent and less if impulses are in dense bursts), and a small gain against single line interferers, improving as the number of line interferers increases. *(See addendum 4)*

Table 1 of 802.4L-89/17 was updated.

After a long break, Larry van der Jagt presented P.4L/4. Use of a similar technique, transmission of exactly one of 2**N orthogonal code words, was discussed. The advantages of limiting the integrate-and-dump process to integrate just for the duration of the delay spread was discussed. Then the group examined whether the output of one correlator, after squaring, could be used to determine the magnitude of the noise in the second correlator.

Jonathan Cheah moved to adjourn for the day at 1755. Tom Phinney seconded. Carried unanimously.

Friday 90.01.19

Vic Hayes called the meeting to order at 0810. 10 people were in attendance. (Neil Wellenstein had departed)

Mike Masleid presented an on-line display of the even and odd periodic and aperiodic correlation of the following Barker and M-sequences:

Barker-11 vs Barker-11

Barker-11 vs the inverse of Barker-11

M-seq 561 preload 376 vs M-seq 561 preload 376

M-seq 561 preload 376 vs inverse of M-seq 561 preload 376

M-seq 561 preload 376 vs M-seq 651 preload 036

M-seq 561 preload 376 vs inverse of M-seq 651 preload 036

(See addendum 5)

Stan Kay presented block diagrams of a conceptual 16-ary CSK (code shift keying) system, showing transmitter and receiver for signaling at 2 Mb/s using a rate 3/4 FEC (forward error-correcting code) followed by a choice of one of sixteen 16-bit Hadamard (orthogonal) code words, followed by a scrambler before modulating the channel. This results in a system with a 667 k symbol/s, 10 2/3 M chip/s signaling rate, with FEC, requiring an uncoded Eb/No of about 12 dB for a BER of 10^{-8} and an additional 3 dB coding gain, yielding a net Eb/No requirement of 9 dB for a BER of 10^{-8}. *(See addendum 6)*

In contrast, the DPQSK discussed in p802.4L/89-16 has a system with 1 M symbol/s, 11 M chip/s, and requires an Eb/No of 15 dB for a BER of 10^{-8}.

Stan noted that the system he had just presented was not really a spread-spectrum system, since there was no processing gain, but that his system has a spectrum like spread-spectrum. The FEC is used to combat Gaussian noise, giving about 6 dB of improvement over a DQPSK system. The FEC also should provide substantially the same improvement against impulse noise. Both systems (DQPSK and 16CSK) occupy about the same spectrum. The two systems are probably about equally resistant to frequency-selective fading. The resistance of the 16CSK system to line interferers needs to be evaluated; the DQPSK system's resistance is 10.4 dB. The overall complexity of the 16CSK system is greater than the DQPSK system.

For the record, the set of sixteen 16-bit Hadamard vectors for 16CSK (in hexadecimal) are:

FFFF	AAAA	CCCC	9999
F0F0	A5A5	C3C3	9696
FF00	AA55	CC33	9966
F00F	A55A	C33C	9669

Upon examination it was realized that these vectors worked in the absence of multipath and with perfect timing recovery, but that the cross-correlation of CCCC and 9999, for instance, gave a full-height correlation spike just one chip off from the nominal time of the expected correct correlation peak. Thus, even with perfect clock recovery, multipath can render this code ineffective.

For an N-ary CSK system like this to work, we need to find a set of orthogonal code words that remain orthogonal in the presence of multipath. The observed delay spread of the factory channels was 500 - 600 ns, so with a chip rate of about 10 M chip/s, this means that the code words should remain orthogonal under shifts (time displacements) of up to six bits.

Mike Masleid sketched a conceptual implementation of a SAW with four correlators, two in series in two in series in quadrature, to support differential detection. *(See addendum 7)*

Tom Phinney sketched two approaches for a distribution system. *(See addendum 8)* In the first, separate modems connected to separate antennae feed independently-received digital "uplink" signals to a central point of storage and assessment, where the "best" signal is chosen for rebroadcast during the "downlink" phase of operation.

In the second approach, each antenna is connected to an LNA (low noise amplifier) which sends an analog signal back to the central assessment point, where all of the analog signals are combined and then processed by a single receiver. The receiver output is delayed and repeated, and sent to separate transmitters associated with each antenna.

In both cases, the modems may either be near the antennae or near the central repeater. However, the LNAs must always be near their respective antennae.

The group broke for lunch at 1238, and agreed to travel to Knowledge Implementations, Inc. (KII) for a demonstration, further discussion, and a fine meal courtesy of KII.

The meeting reconvened after travel (about 1 hour) at KII at 1450. 9 people were in attendance. (Don Johnson, who was feeling ill, had remained behind) Larry van der Jagt was able to show on lab equipment that a microwave oven truly was a line interferer of amazing spectral purity (40 dB in the vicinity of the "carrier") when viewed over the duration of a few symbol times. Larry also demonstrated that physically small interference sources, such as motorized children's toys, could generate impulses of large amplitude with respect to a received signal.

Mike Masleid was able to show that the Hadamard codes discussed earlier in the day gave very high auto-correlation and cross-correlation which renders these specific codes unusable.

The meeting adjourned for the day at 1730. This was followed by a fine meal provided by our host (KII), which was enjoyed by all present.

Saturday 90.01.20

Vic Hayes called the meeting to order at 0810. 10 people were in attendance. The discussion turned again to the use of N-CSK. The proponents of N-CSK were asked to look at codes somewhat longer than the Hadamard codes, which would be almost orthogonal, and which would have a single auto-correlation peak; low cross-correlation in the vicinity of the auto-correlation peak; and low auto-correlation in the vicinity of, but not at, the auto-correlation peak.

It was pointed out that the measurements made by Rappaport were at quarter-wavelength intervals, but gave only power-delay information and not I and Q phase change vs motion information at the symbol times and velocities of interest. The DQPSK approach of p802.4/89-16 relies on short-term and short-distance coherence of the channel impulse response over a number of symbols. The N-CSK approach will work in cases where the channel impulse response is short-term coherent over just a single symbol.

Coherence over a hundred symbols permits coherent detection. Coherence over a few symbols permits differential but not coherent detection. Coherence over a single symbol requires a modulation technique based only on power reception. The velocities of interest in manufacturing are on the order of 10 m/s, which is ten micro-meter (micron) per microsecond, or about 20 miles/hr.

To better measure the effects of microwave oven interference in the 2.4 GHz band, Jonathan Cheah will be making measurements of microwave ovens using a high-quality spectrum analyzer (HP 8566). The group specified the type of measurements that should be made to find out the energy distribution over the band, as well as the short-term frequency stability of the power source. This will establish whether a microwave oven is a broadband or a narrowband emitter, and at what rate the predominant frequency changes.

Don Johnson asked that the measurements also be done separately for the first 26 MHz sub-channels of the 2.4 GHz band, which will provide information on whether or not channel selection (of channels of the bandwidth equal to that available at 900 MHz) would permit avoidance of the interferers. *(See addendum 9)*

It was noted that most microwave ovens have a slow-mode mixer (the "stirring" paddle which sits between the magnetron and the oven cavity). However, "convection" microwave ovens use a motorized fan (probably of 1800 rpm, but slipping slowly) to provide convection as well as mode stirring, and this probably will cause a much faster rate of frequency shift in the magnetron output.

The **next interim meeting** will be in Irvine, CA on 11-12 March, immediately preceding the IEEE 802 plenary. The **following interim meeting** will be held the week of May 14-18 in Atlanta, GA. (Accommodations are still to be arranged)

The group next addressed the subject of **distribution systems**. Tom Phinney claimed that a simplistic solution to the distribution system was to synchronize the transmitters as if they were receiving their transmit clock and data from geosynchronous orbit, and to then view the multiple transmissions as being roughly equivalent to a single system with peculiar attenuation and with a delay spread increased over that of a single transmitter system by the one-way propagation delay of the plant.

Discussion pointed out that with free-space radiation from a dipole, the received power exponent has theoretic exponent zero in a waveguide, one in an expanding cylinder, and two in free space. If the transmitter population is constant per unit area, then the population of transmitters at distance D varies inversely with D (the circumference of the circle), whereas the attenuation from each such transmitter varies with an exponent somewhere between 1.0 and 4.3 (Rappaport's measurements).

Rappaport found that fading was either Rayleigh or Rician. Consideration of the fading which might occur at the center of an equally-spaced grid due to fading of the line-of-sight paths seems to indicate that the transmitters should be staggered (probably in a 1-2-4-3 pattern around the square) by a chip duration or more relative to each other, which will tend to separate the sources into different resolvable paths for the situations where the attenuation from all paths is equal. Such staggering may not be inexpensive, since it requires sub-symbol-duration delays whereas a SAW-based modem may only need a symbol clock for its other functions.

Mike Masleid contended that such staggering was unnecessary, since Rappaport's data showed that line-of-sight cancellation was common, but that real buildings had so much scattering and receivers had such sensitivity that even line-of-sight cancellation was irrelevant; the multipath from other scatterers would be adequate to provide a receivable signal.

Don Johnson pointed out that 4-way staggering may still be needed if the receiver needs to determine and compensate for the apparent carrier frequency offset of the transmitter and receiver. How this will work when receiving signals from multiple different transmitters which have differing rates of phase precession one from the next is difficult to envisage.

Discussion then turned to an **optimal antenna array**. If each antenna is placed on the ceiling, with its E and H fields horizontal, an absorbent floor (e.g, concrete), and a ring of absorbent material on the ceiling around the antenna, then each antenna tends to illuminate only its own cell and the edge of the adjacent cells, and so a receiver should receive line-of-sight from no more than two cells, with all other received energy resulting from scattering from remote transmitters.

It may also be reasonable in a plant to have transmitters at a high density in the center of the plant, and to decrease the density and increase the directionality (away from the center) of antennae placed around the periphery of the plant. *(See addendum 10)*

Tom Phinney contended that there would be times in many real plants, usually observed during system acceptance testing, when a remote source would illuminate a receiver with an attenuation coefficient approaching one. In the case of the Oshawa plant, over a four-year period of system operation, the number of seconds of receiver operation would be 10^{11} (10^3 receivers, 10^8 seconds). A long discussion about whether or not such an unfortuitous situation could occur ensued.

The group broke for lunch at 1300, and reconvened at 1415. During lunch, the group agreed that this problem would not be worth a lot of design effort for the retail or office markets, even if it were likely to be observable in rare cases, since moving the terminal an inch or two would resolve the problem even when it did occur. For the industrial market, it appears that the antenna diversity as supported by the (presumably forthcoming) 802.4K media/modem diversity standard can provide a solution to the problem of local fading while clearly hearing a remote transmitter. However, Mike Masleid contended that this type of diversity would not solve the problem of a receiver positioned such that it had moved into a deep fade-area.

Jonathan Cheah pointed out that the words "raw bit error rate" in 3.9 of 802.4L/89-24 were inappropriate. 3.9 was changed to read "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."

Mike Masleid expressed his concern that the effective delay spread caused by the antenna diversity of the distribution system be considered in the choice of the modulation technique, so as not to preclude use of this proposed standard in large enclosed areas. The implication is that the modulation technique should permit but not mandate receivers designed for peak delay spread of 3 microseconds.

A number of papers (802.4L-89/23A-D) with respect to regulatory issues were included in the last mailing. They are called to everyone's attention.

At 1530, Tom Phinney moved that the meeting be adjourned. Jonathan Cheah seconded. Carried 10-0-0.

List of temporary documents

Temp.	Source	Title	Document number
P.4L/1	Hayes	Document list	
P.4L/2	Hayes	Agenda	
P.4L/3	Hayes	Attendance list	
P.4L/4	Van der Jagt	Receiver block diagrams for illustration	Attachment 1

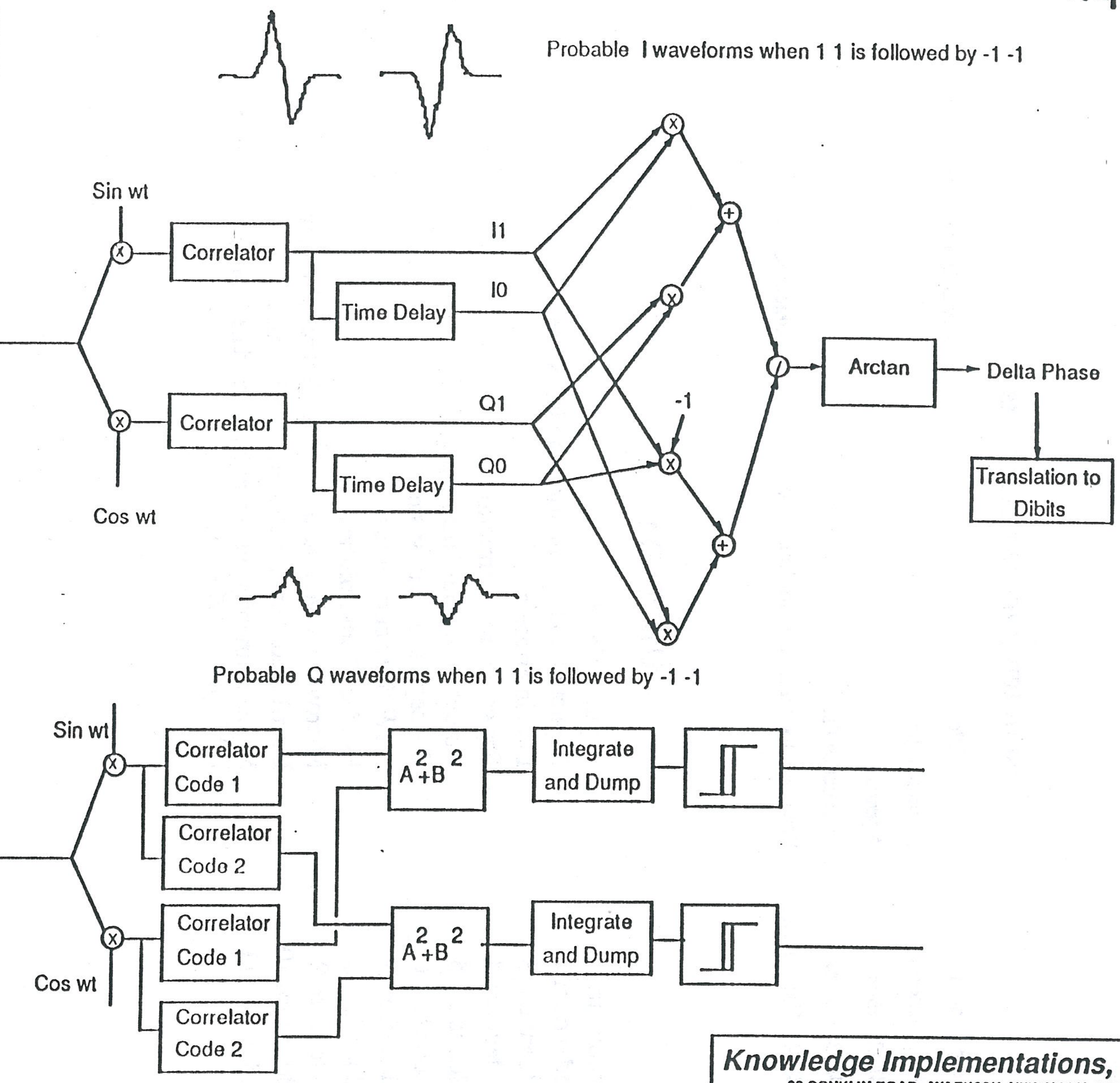
ADDENDA

Addendum 1	Impulse Noise
Addendum 2	Impact of code length on Impulse Noise
Addendum 3	Line interferers
Addendum 4	Forward error Correction
Addendum 5	Output of various correlators
Addendum 6	Conceptual 16-ary system
Addendum 7	SAW filter correlators
Addendum 8	Head-end possibilities
Addendum 9	Microwave Oven measurement, Test requirements
Addendum 10	Distributed Antenna System, possibilities
Addendum 11	Noise immunity vs spreading (updated table 5 of doc.: IEEE p802.4L/89-17)

P.4L/4

Note: These block diagrams are intended as illustrations only of a set of elements which implement possible approaches to the RadioLan problem and are not intended to reflect actual implementations.

QPSK Transmit System as Proposed
(as outlined in IEEE 802.4.L/89/16)



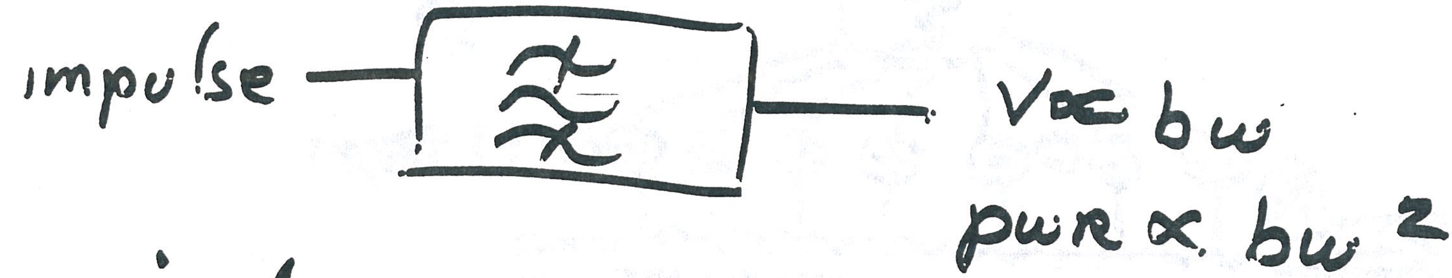
Code1,Code 2 Transmit System
Possible Coding Structures

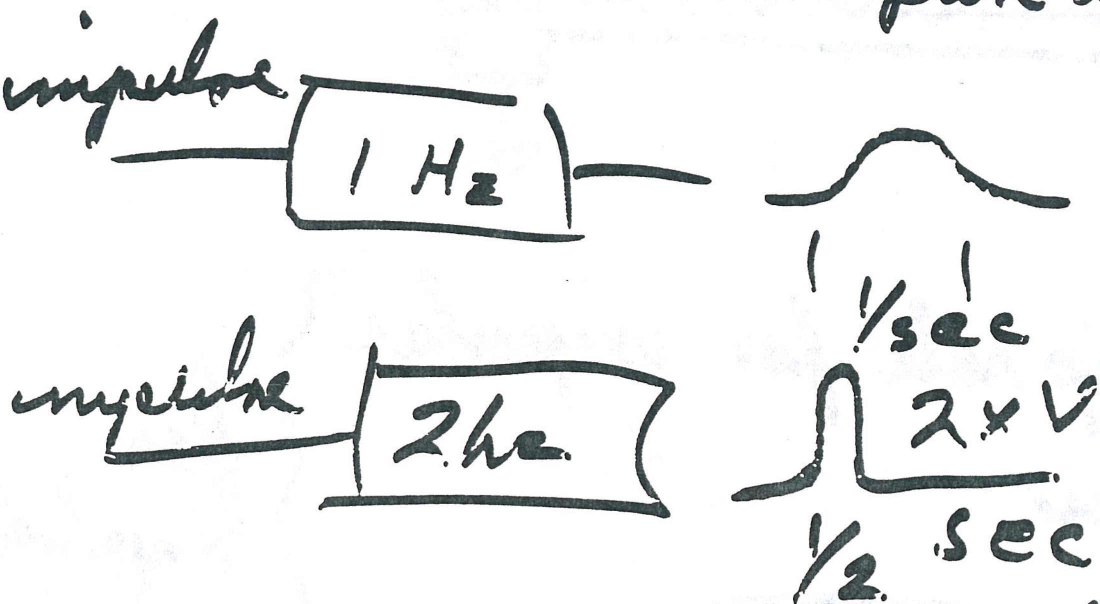
- Binary:Code 1 = 1,Code 2 = 0 (note: correlators for Code 2 optional in this scenario)
- Three Level: Code 1, nothing
  Code 2, nothing
  Code 1, Code 2
- [illegible]

Assumption in both cases - There is a way to find sampling times

Note: Coding Choice is a tradeoff between data rate and error performance

Knowledge Implementations, Inc. 32 CONKLIN ROAD, WARWICK, NEW YORK 10990	
TITLE: RECEIVER BLOCK DIAGRAMS FOR ILLUSTRATION	
PROJECT: RADIOLAN PRODUCT DEVELOPMENT	REV. 0.1
ENGINEER: L. VAN DER JAGT	DATE: 1/15/90 SHEET 1 OF 1

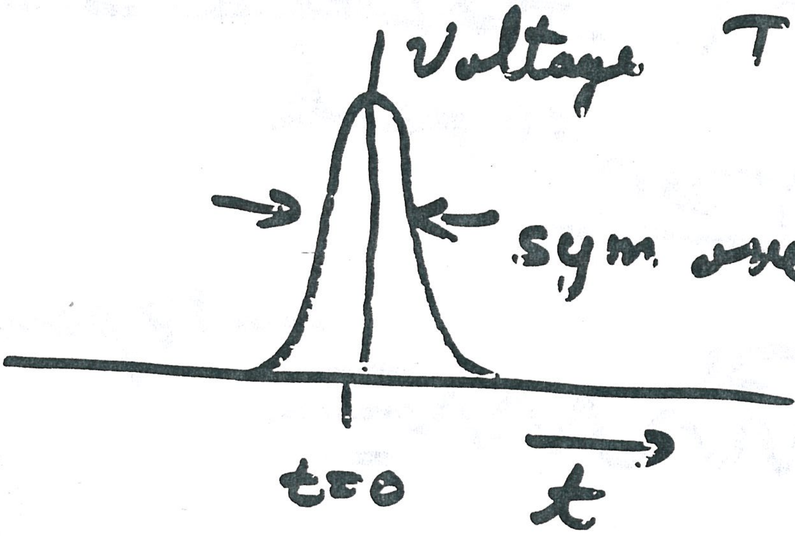




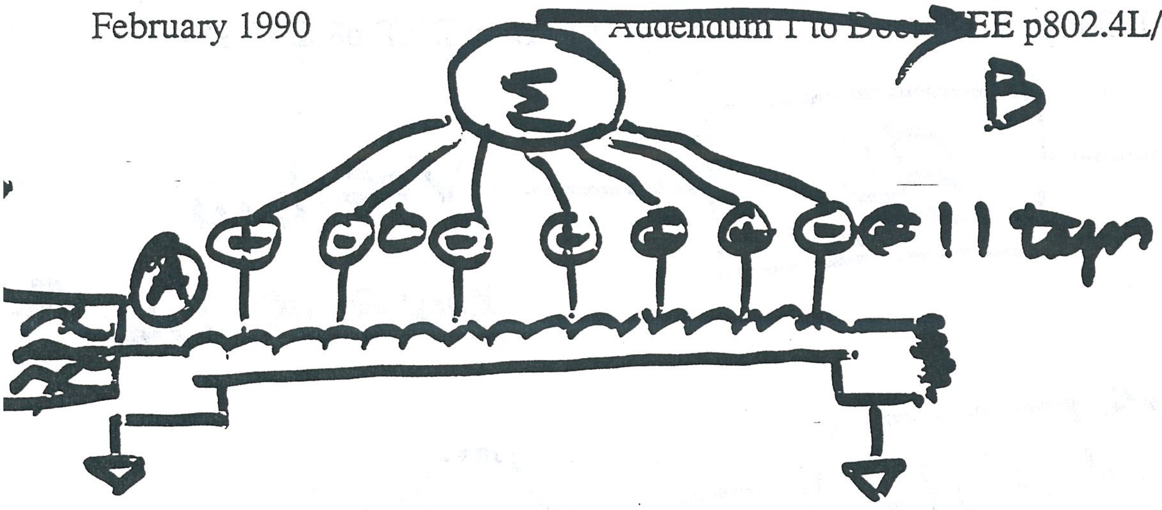
½ of the time

$2^2 \times \frac{1}{2}$ time = 2 energy

...ulse = ∫pwr · dt

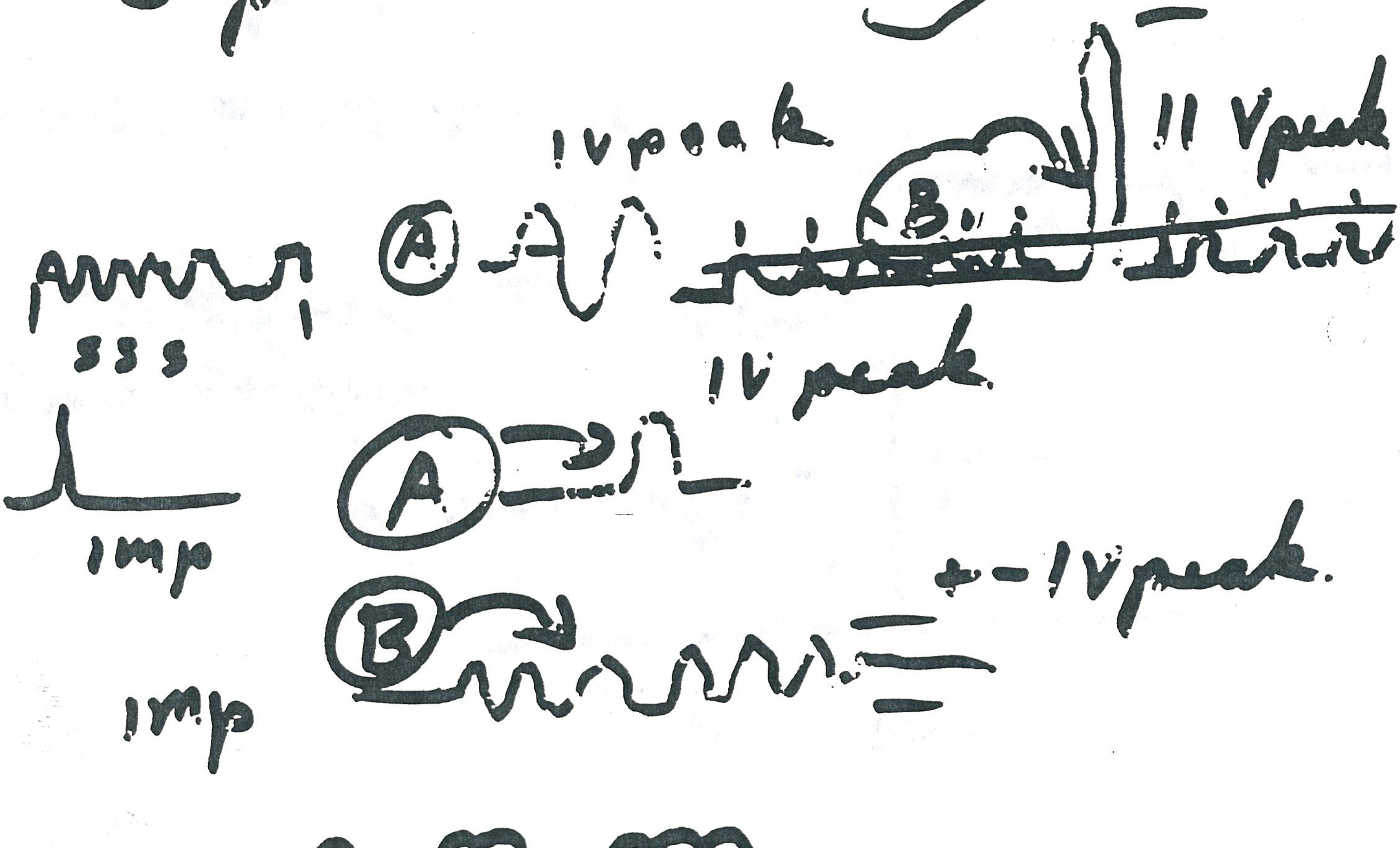


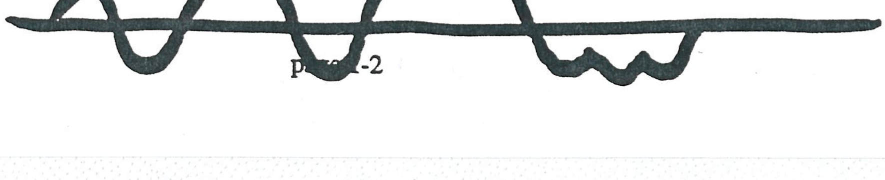
T = length of impulse response

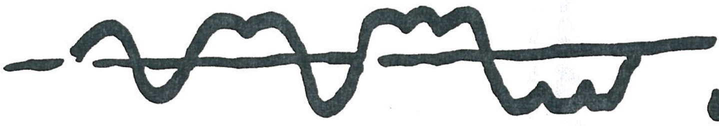


$V \propto$ chip rate for impulse
$V \propto$ bw
$V \propto \frac{1}{\text{chip time}}$

} for impulse






impulse

ADD1/4

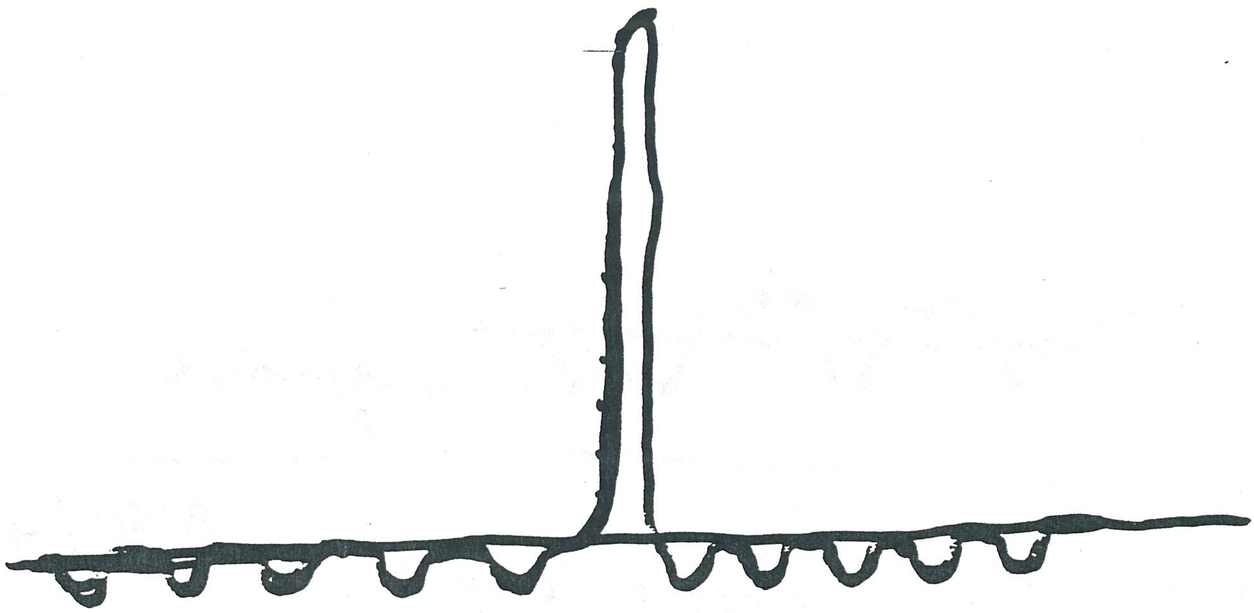
$\sigma_{impulse} \propto$ CHIP RATE

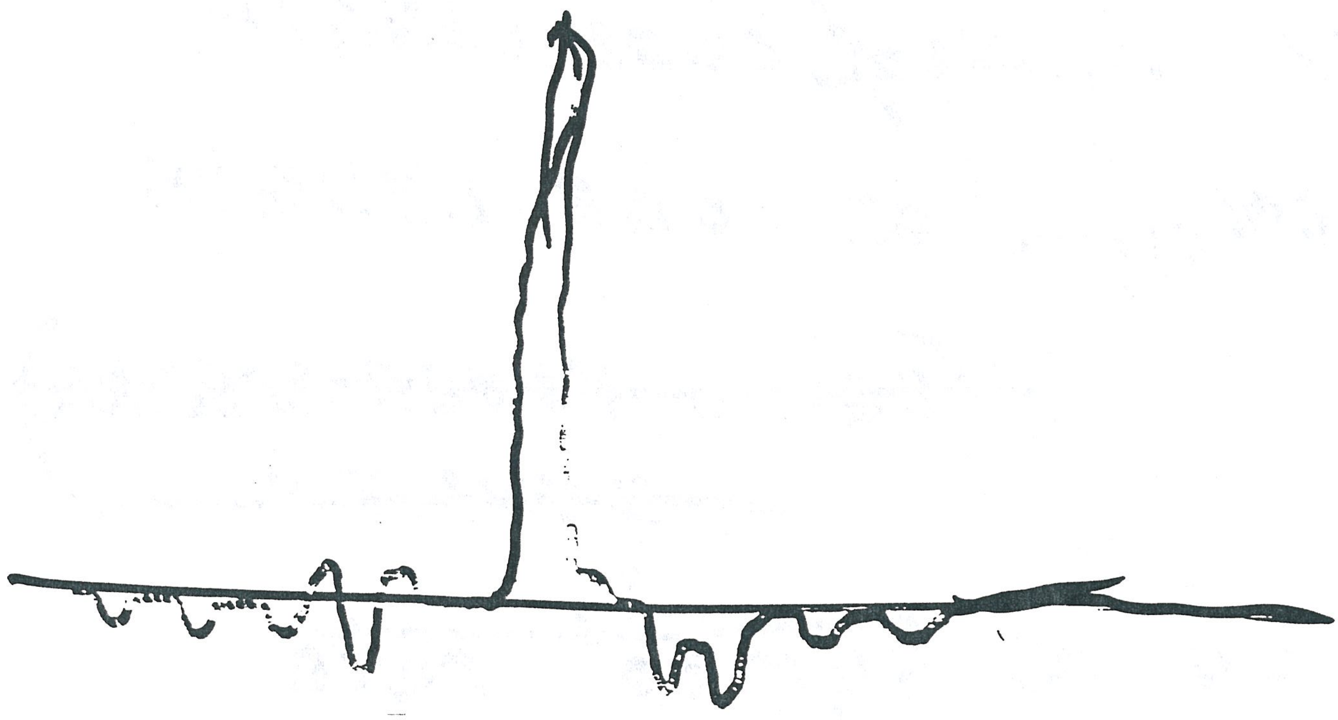
$\sigma_{impulse}$ NOT $\propto$ CODE LENGTH

$\sigma_{signal} \propto$ CODE LENGTH

~~(EQUAL POWER SPECTRAL DENSITY)~~

CSN $\propto$ $\overline{\text{symbol rate}}$





IMPACT OF CODE LENGTH ON SINGLE IMPULSE NOISE COMPONENT VS. SIGNAL COMPONENT AT CORRELATOR OUTPUT

IN BW = 1 SAY IMPULSE = 1 = IMPULSE WEIGHT

CORRELATOR PEAK ≈ CODE LENGTH,

NOISE FROM IMPULSE ≈ 1

$$\text{RATIO} = \frac{\text{CODE LENGTH}}{\text{IMPULSE WEIGHT}} = R_1$$

IF BW = N SAME IMPULSE IS N

SAY BANDWIDTH HAS GONE UP BECAUSE CODE LENGTH IS INCREASED BY N AND SYMBOL RATE IS MAINTAINED

THEN

CORRELATOR PEAK ≈ N × CODE LENGTH,

SO

$$\text{RATIO} = \frac{N \times \text{CODE LENGTH}_1}{N \times \text{IMPULSE WEIGHT}} = R_1$$

IF IMPULSE IS CLIPPED AT 5 IMPULSE WEIGHT AND N IS 10

THEN

$$\text{RATIO} = \frac{10\ \text{CODE LENGTH}_1}{5\ \text{IMPULSE WEIGHT}} = 2R_1$$

⊛ = CORRELATION

* = COMPLEX CONJUGATE

x = MULTIPLICATION

FT = FOURIER TRANSFORM

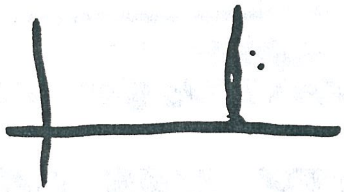
t = TIME

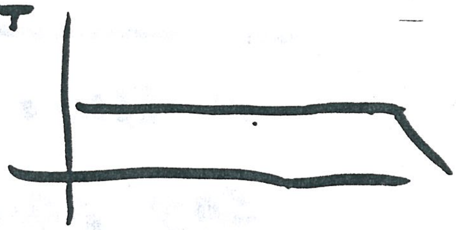
u = FREQ.

$$FT[f_1(t) \circledast f_2(t)] = FT[f_1(t)] \times FT^*[f_2(t)]$$

$f_1(t) = \sin \omega t$

$f_2(t) = $ CODE SEQ

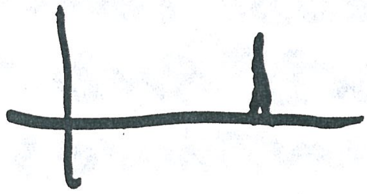


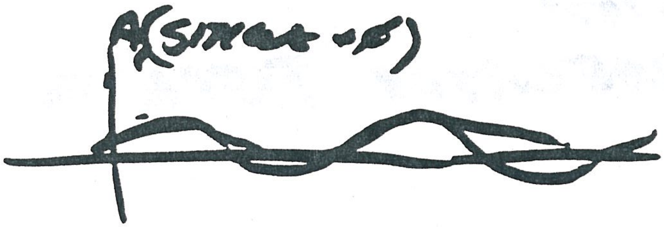


SIN ωt ⊛ CODE

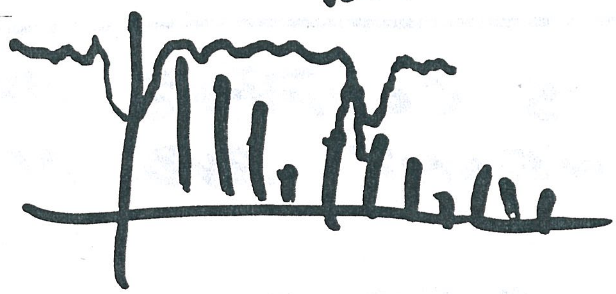
FO CODE

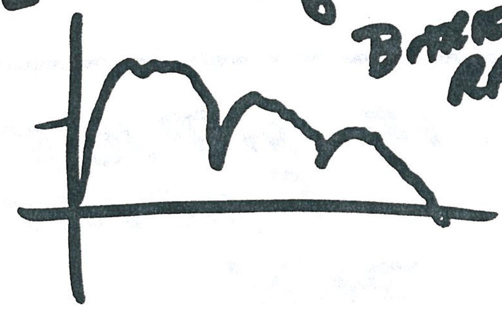
N





BARKER REPETATIVE





BARKER RANDOM

Uncoded	R 1/2 Coded
X I X I	X I X P X I
→| 1µsec |←	→| 1µsec |←
P	P
$E_{bi} = P \cdot 1\mu sec$	$E_{bi} = P \cdot 1\mu sec$
$E_{bt} = P \cdot 1\mu sec = E_{bi}$	$E_{bt} = P \cdot \frac{1}{2} \cdot .5\mu sec = \frac{E_{bi}}{2}$
$BER = Q\left(\sqrt{\frac{2E_{bi}}{N_0}}\right)$	
$\frac{E_{bi}}{N_0} = 10.5\,dB$, $BER = 10^{-6}$	$\frac{E_{bt}}{N_0} = \frac{1}{2}\frac{E_{bi}}{N_0}$ $\frac{E_{bt}}{N_0} = 7.5\,dB$ BERs = 5x10

~~FER corrects~~

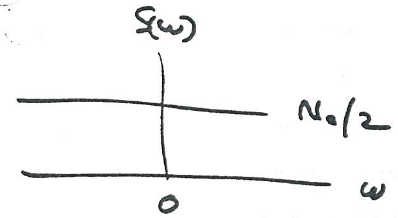
typical System

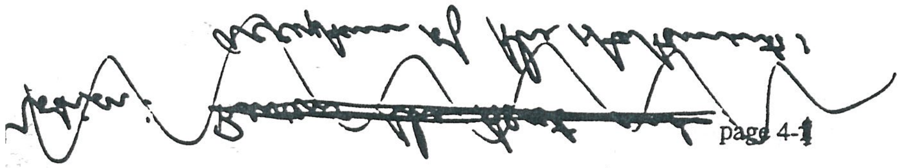
$$\frac{E_{bi}}{N_0} = 5.0\,dB$$

$$\frac{E_{bt}}{N_0} = 2.0\,dB$$

Raw BER = 6×10^{-2}

Corrected → 1×10^{-6}





The graphs on the following pages are submitted by Michael Masleid on his little portable computer. They depict the output of a correlator working in the following ways:

The top graph correlates against "codeword 1", when its input is

all ZEROs - "codeword 1" - "codeword 2" - all ZEROs.

The bottom graph correlates against "codeword 2" with the same input as fdor the top graph.

The codewords are selected from the following table:

M-sequences of length 256

int	octal code
0	0561
1	0651
2	0551
3	0545
4	0765
5	0747
6	0615
7	0607

int	Barker sequence
8	+++---+--+-
9	---+++-++-+
10	-+--+---+++
11	+-++-+++---

int	Hadamard Code	int	Hadamard Code
12	++++++++++++++++	20	++++++++--------
13	+-+-+-+-+-+-+-+-	21	+-+-+-+--+-+-+-+
14	++--++--++--++--	22	++--++----++--++
15	+--++--++--++--+	23	+--++--+-++--++-
16	++++----++++----	24	++++--------++++
17	+-+--+-++-+--+-+	25	+-+--+-+-+-++-+-
18	++----++++----++	26	++----++--++++--
19	+--+-++-+--+-++-	27	+--+-++--++-+--+

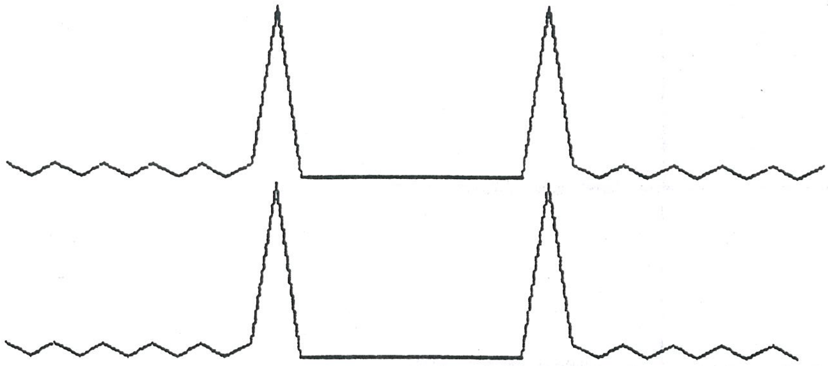


Figure 1.

Barker Sequences
Codeword 1 11
Codeword 2 11

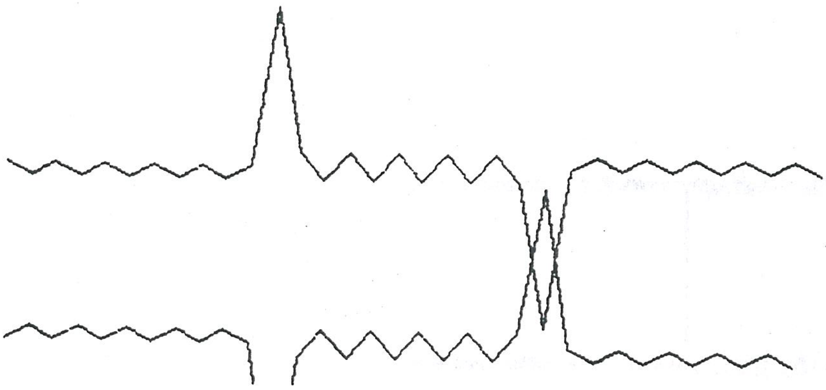


Figure 2.

Barker Sequences
Codeword 1 11
Codeword 2 -11

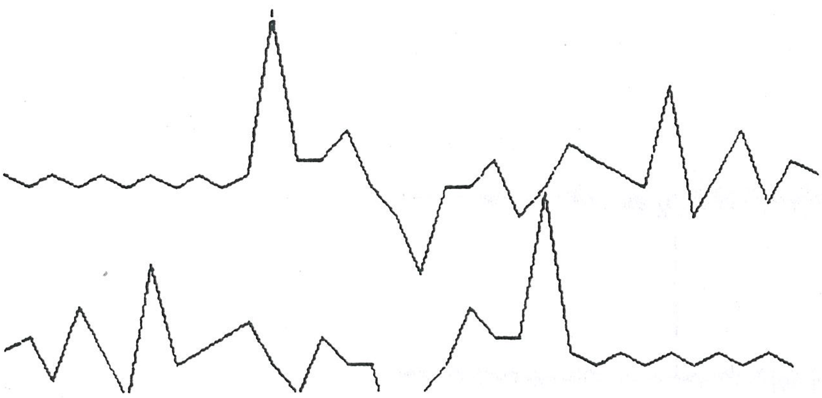


Figure 3.

Barker Sequences
Codeword 1 11
Codeword 2 9

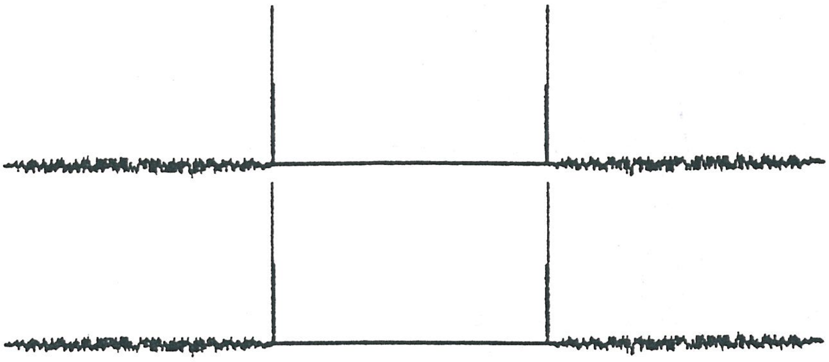


Figure 4. Maximal Pseudo-random
Codeword 1 0
Codeword 2 0

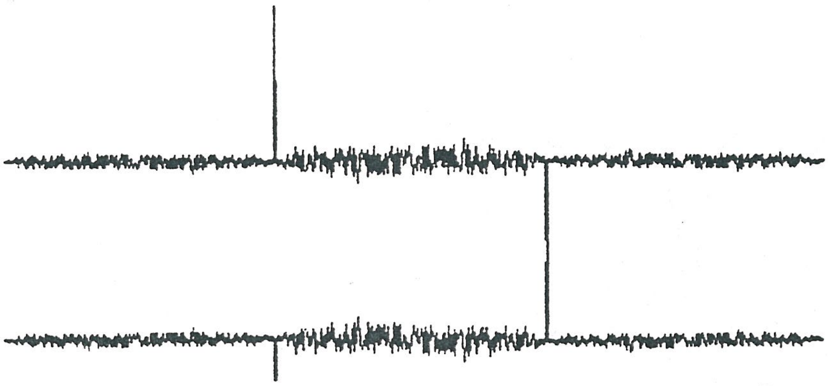


Figure 5. Maximal Pseudo-random
Codeword 1 0
Codeword 2 -0

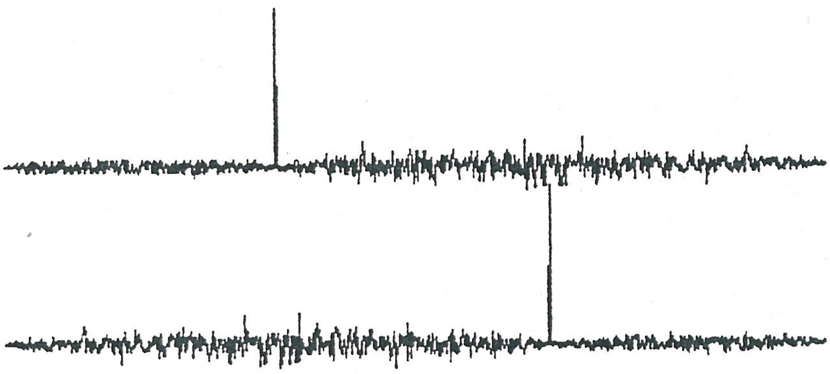


Figure 6. Maximal Pseudo-random
Codeword 1 0
Codeword 2 1

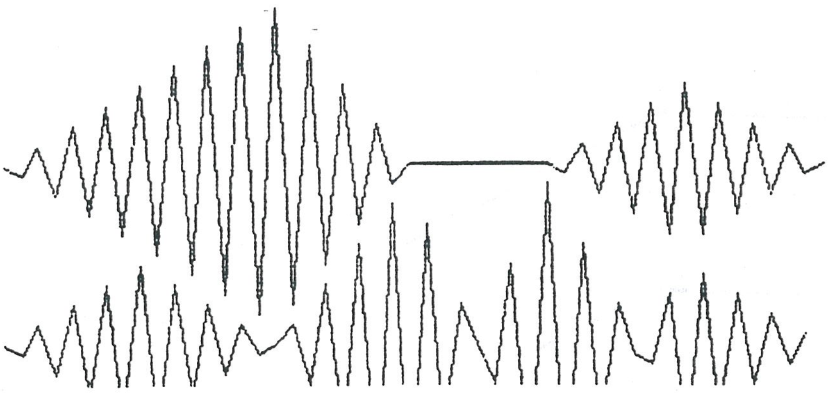


Figure 7. Hadamard
Codeword 1 13
Codeword 2 21

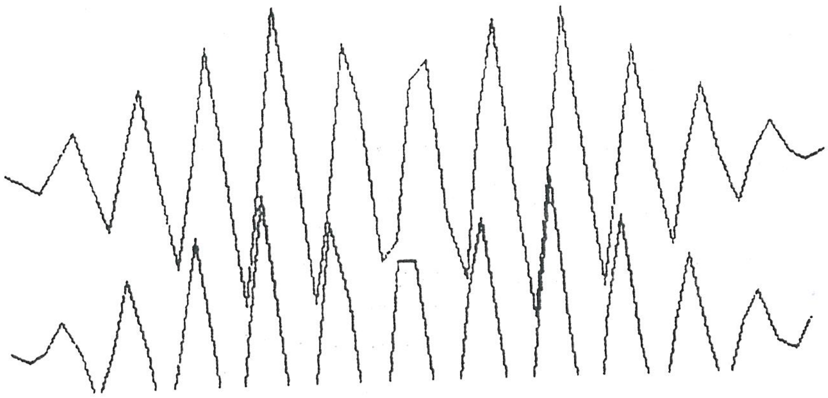


Figure 8. Hadamard
Codeword 1 14
Codeword 2 15

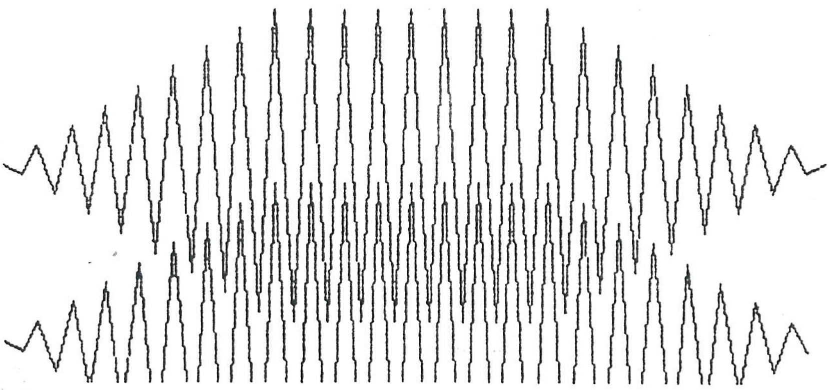


Figure 9. Hadamard
Codeword 1 13
Codeword 2 13

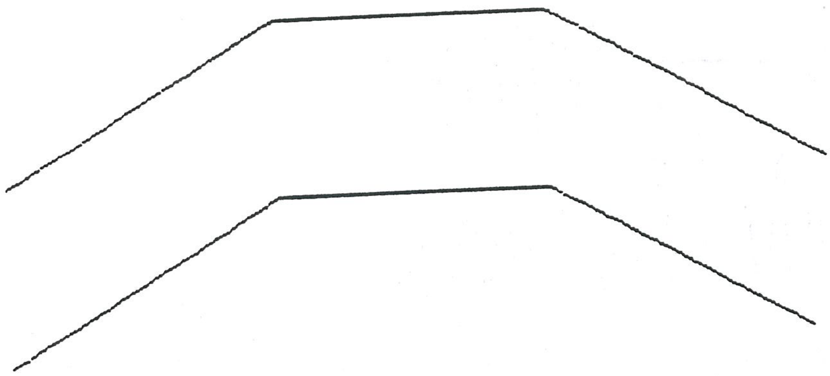
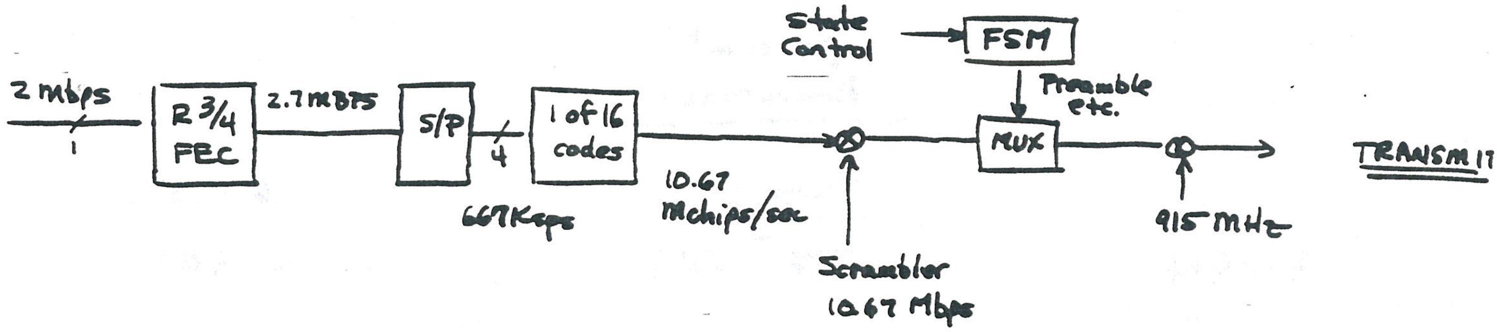
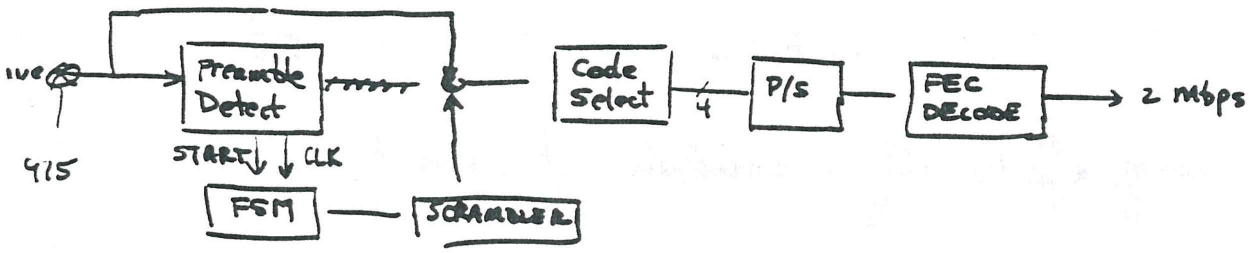


Figure 10.

Hadamard	
Codeword 1	12
Codeword 2	12

A System





A Comparison

Modulation	DQPSK	16-ary orthogonal
Info Bit Rate	2 Mbps	2 Mbps
FEC Code	1	3/4
Bits/Symbol	2	4
Symbol Time	1000 ~~500~~ nsec	1500 nsec
Chip Rate (MCPS)	~~22~~ 11	10 2/3
E_b/N_o for 10^{-8} BER	≈ 15dB ?	12 dB ? ~~Hard Decision~~ ~~Added~~ to coding
Coding Gain	0	≈ 3 dB
E_b/N_o for 10^{-8} with coding	≈ 15dB	≈ 9 dB with coding
Complexity	Minimal	32 corellators FEC Comparators

Impairment Resistance

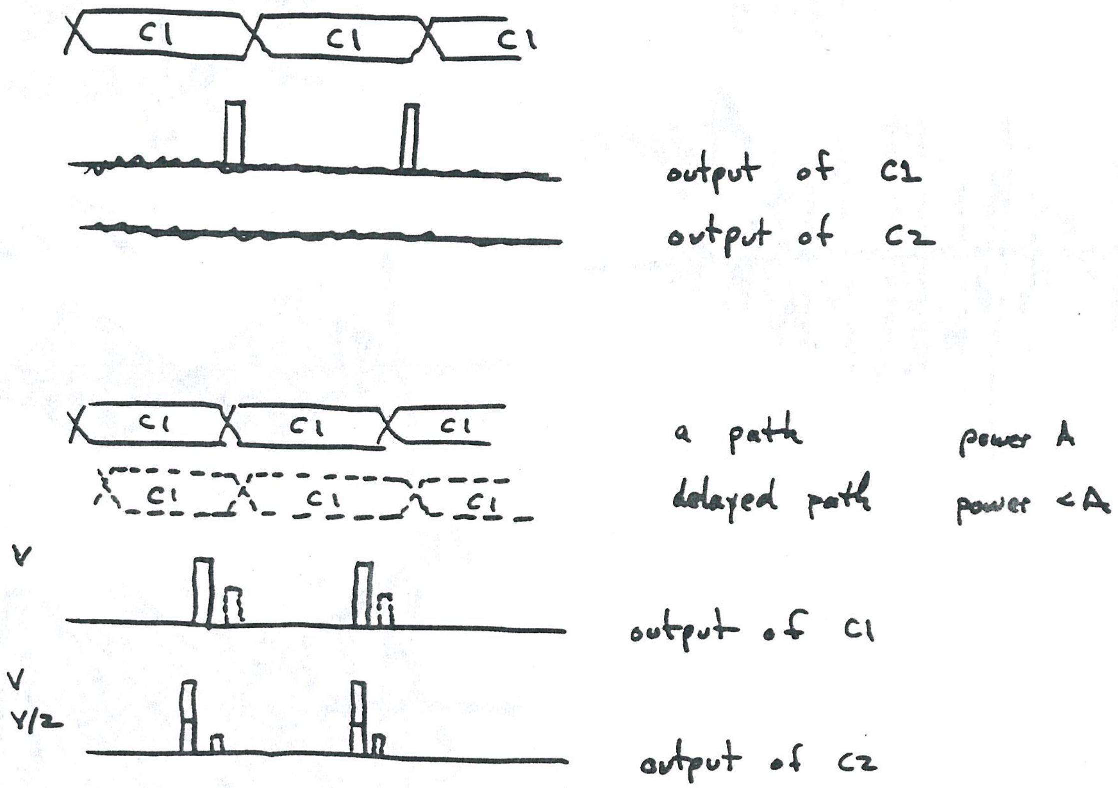
	DQPSK	16-CSK
AWGN	≈ 15 dB	≈ 9 dB
Impulse Noise	Clipping according to B_c	Clipping according to B_c
	no FEC	FEC

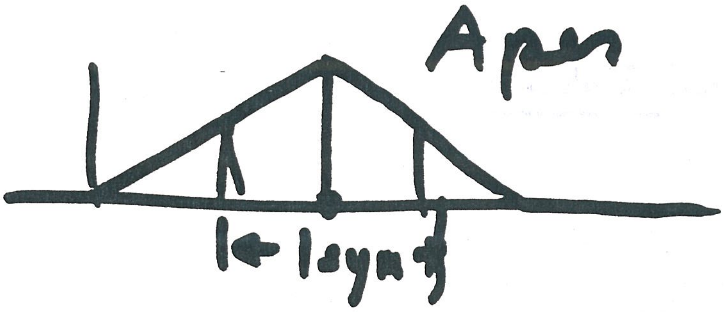
Impulse ~~may~~ affects all 16 constellations at same time.

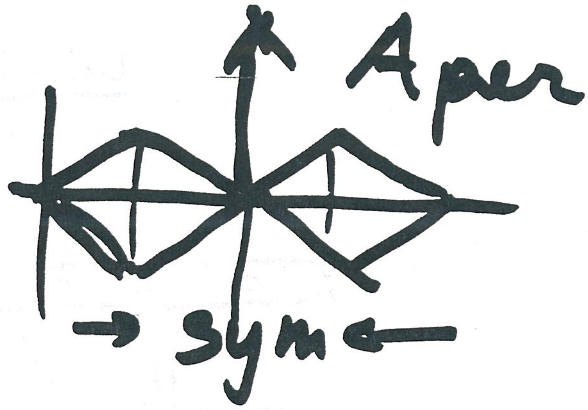
	DQPSK	16-CSK
Tone Jammer	10.4 dB	??
Freq Sel Fades	10.4	~~10.4~~ ??

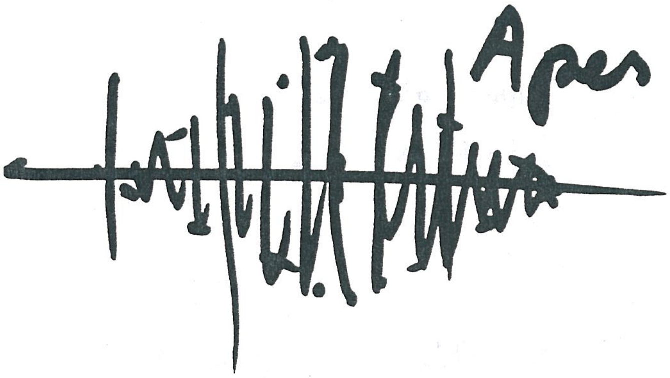
11.3	10^{-7}
10.5	-6
9.6	-5
8.4	10^{-4}

Delay Spread Behavior




Aper


Aper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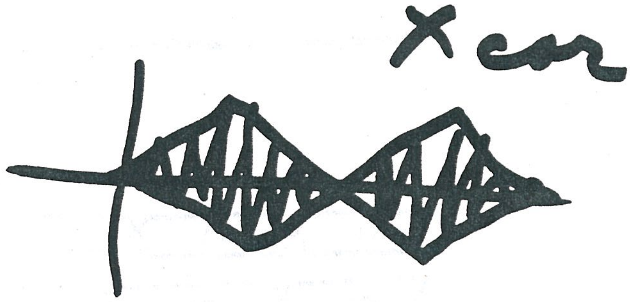
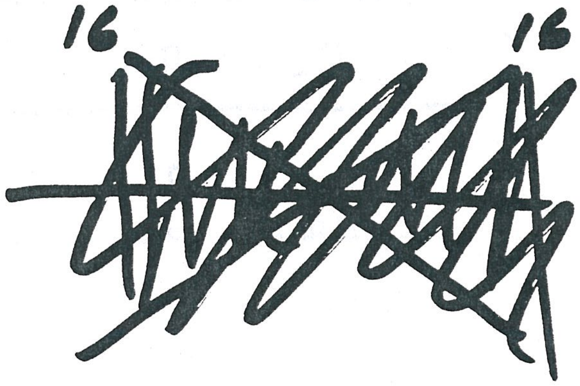
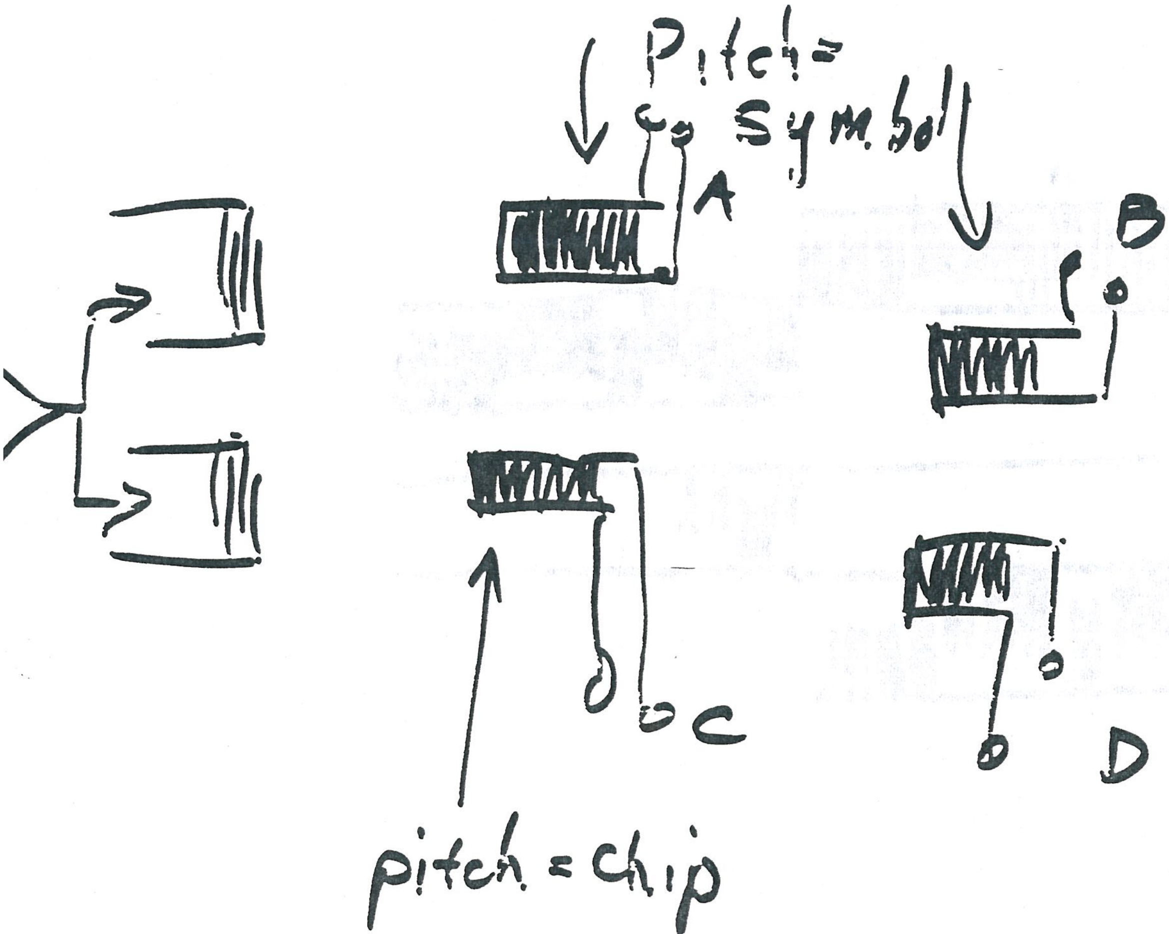
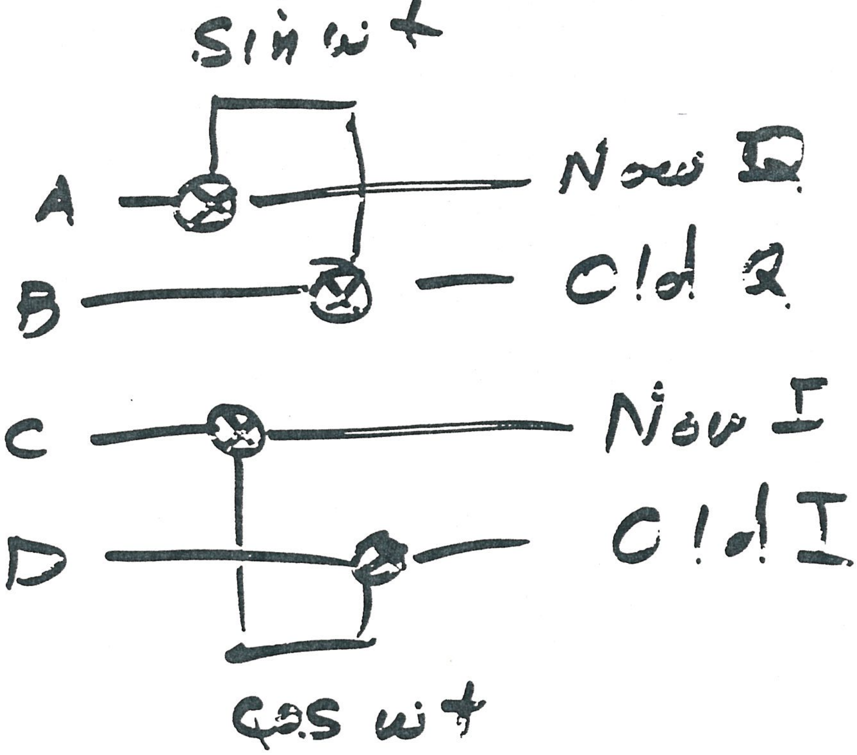
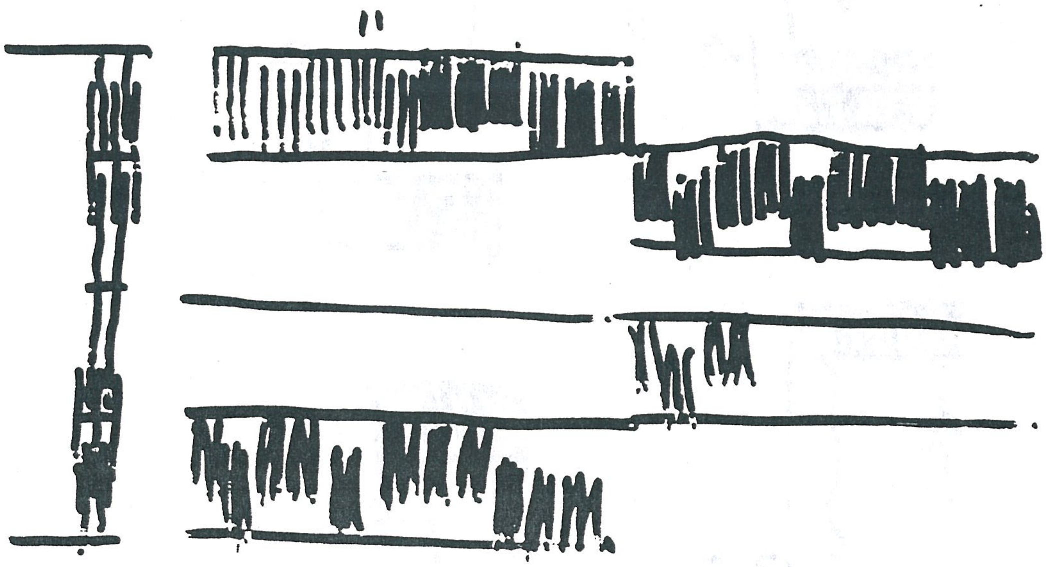
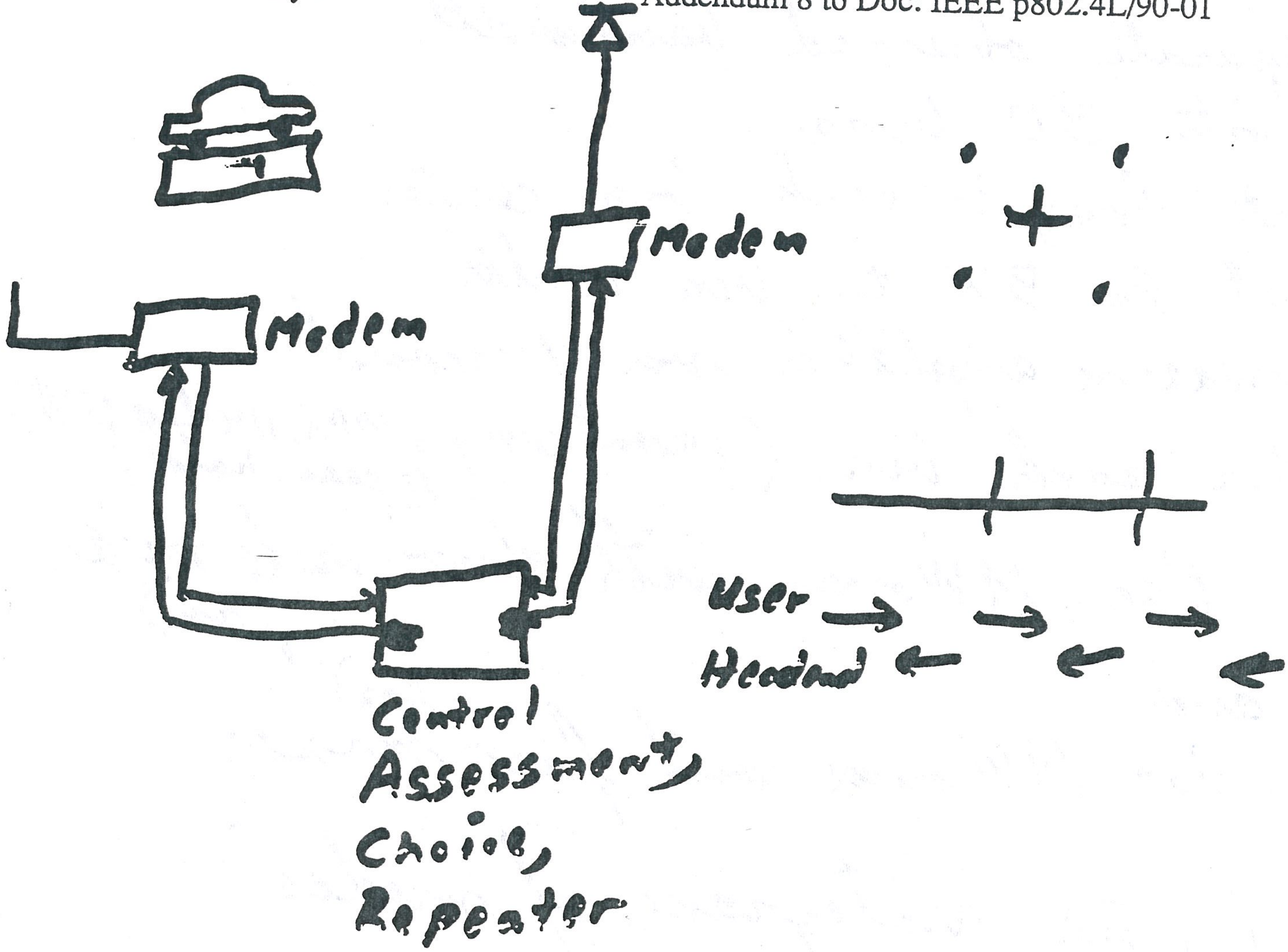

Xcor

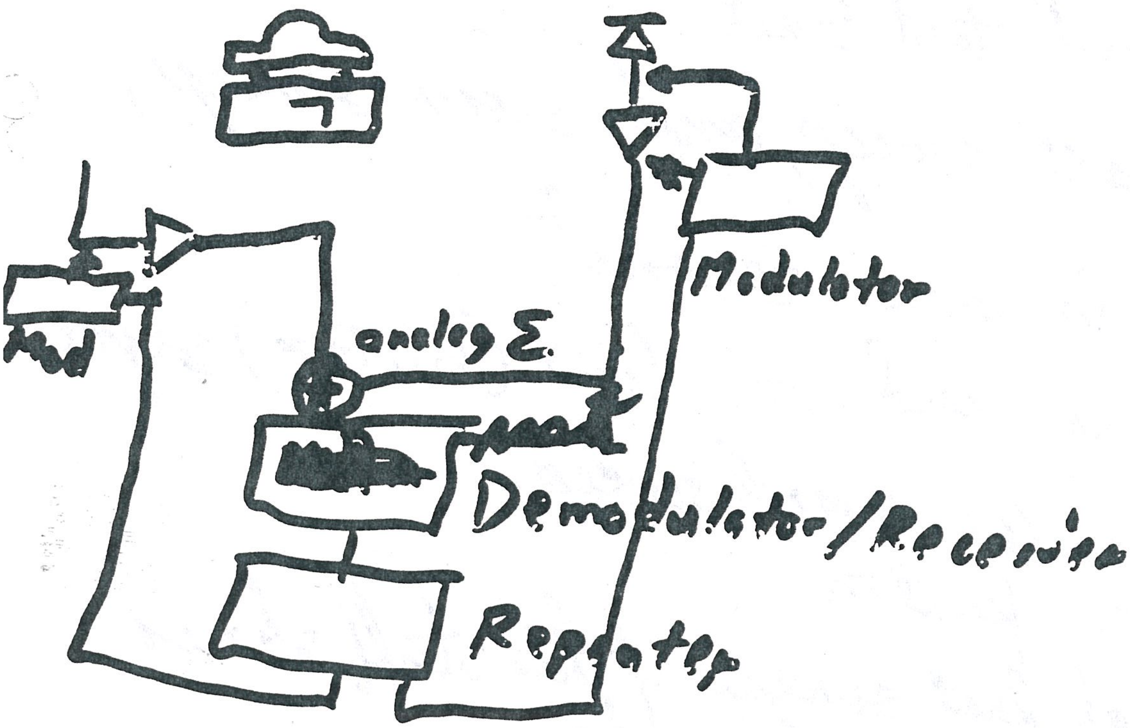

16
16


Pitch = symbol
A
B
C
D
pitch = chip


sin ωt
A
B
New Q
Old Q
C
D
New I
Old I
cos ωt




Modem
Modem
Control
Assessment,
Choice,
Repeater
User
Headend


Mod
analog Σ
Modulator
Demodulator
Repeater

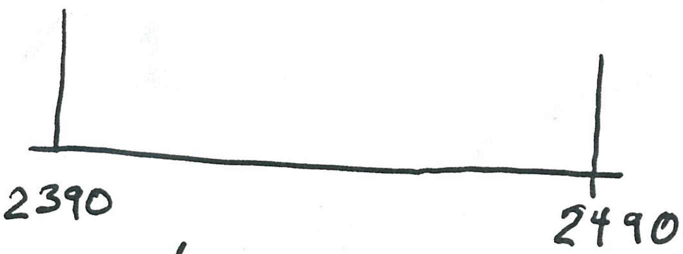
separate observed bandwidth
into 30 bins.
set freq to each bin center
set Res BW to bin width
measure amplitude over 1 second
for each bin. (1 second sweep, MAX video BW)
no peak hold?

Use MW oven with slow mode mixer
(fan)
and
Use MW oven with fast mixer. (fan)

(time histogram of modes)

Second test Suit
try fast sweep, and slow sweep,
slow is one second
fast is used on line (60 Hz) sync
, 1/30, 1/60 second sweep.
or faster.
try real fast sweeps if this stuff shows
high speed structure.

1st 4 tests.



10 min sample time

peak Hold, Res BW 3 MHz, Vid BW MAX

sweep slow
@slowest sweep
addiquate for test suite

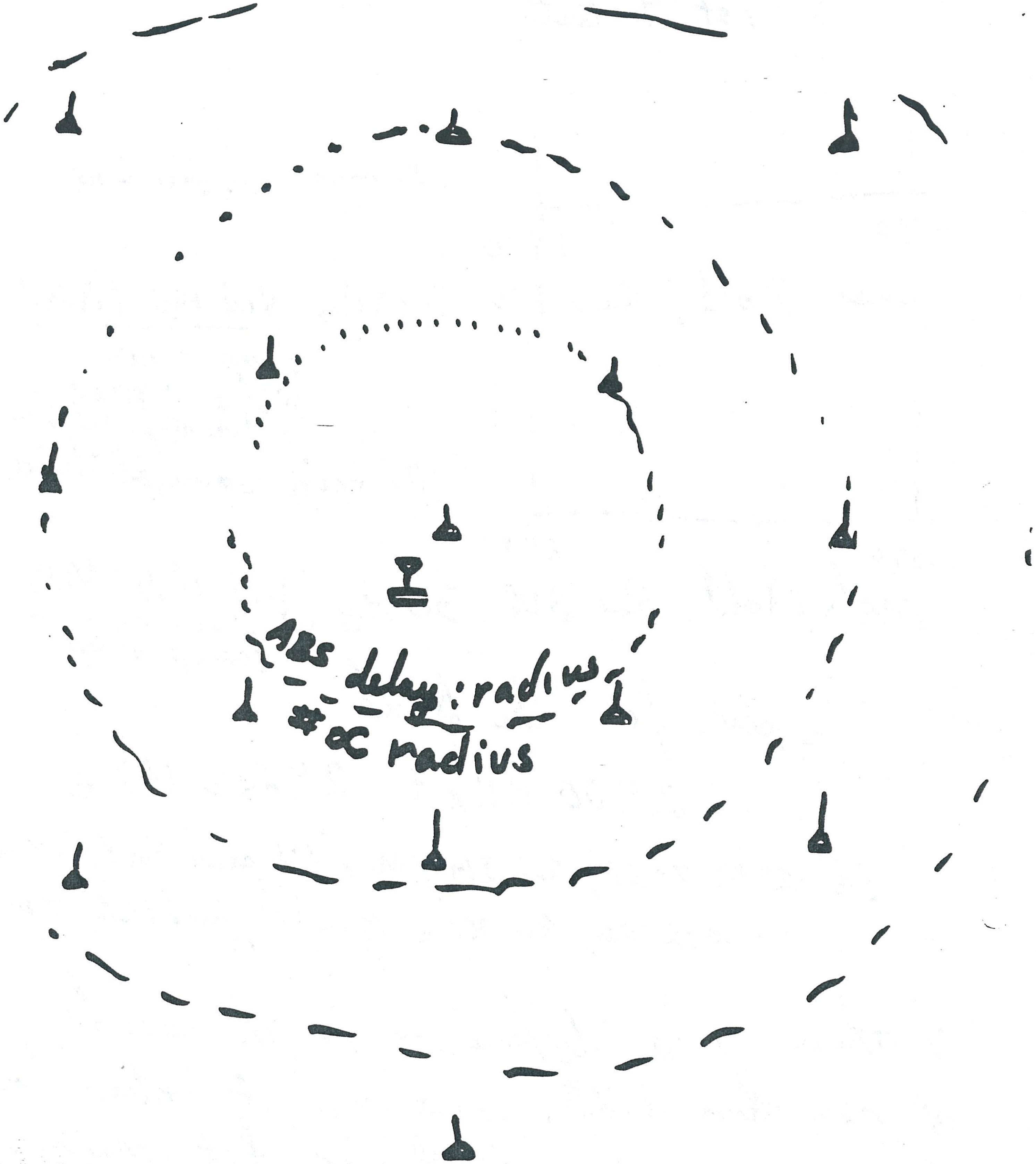
10 min sample time

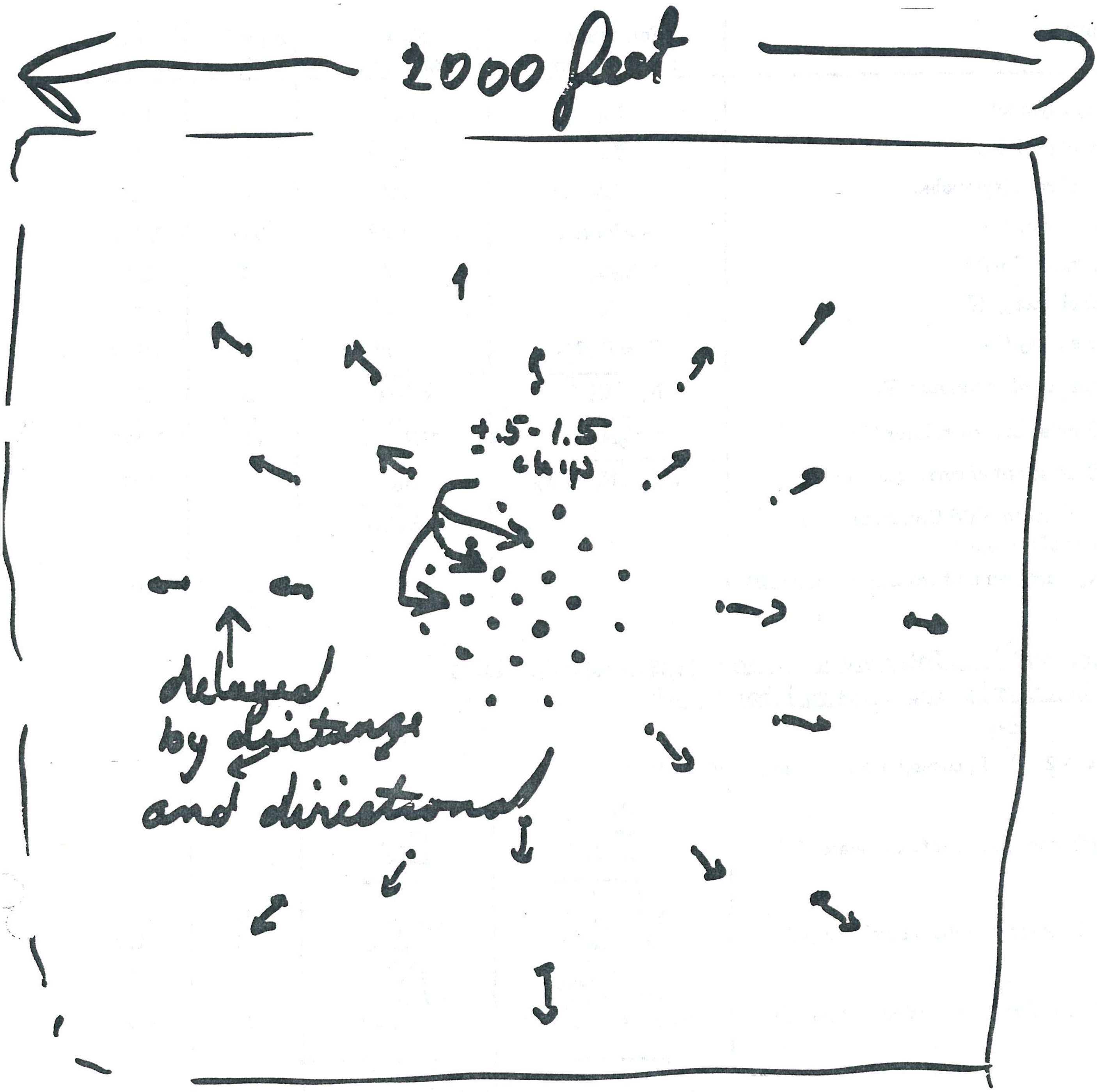
peak Hold, Res BW 30 Hz, Vid BW MAX
sweep slow

repeate for the band
2400 MHz to 2426 MHz

ie 2400-2426, Res 3MH, Max Vid, peak Hold, 10 Min
2400-2426, Res 30Hz, Max Vid, peak Hold, 10 Min

if there is a difference in the result of more than 2 dB, measure at intermediate frequency's to get things that map bw. ie, same at 3MHz, 1MHz, 300kHz, dropping from 100kHz → to 30Hz


ARS delay : radius
∝ radius


2000 feet
±.5-1.5
chips
delayed
by distance
and directional

Constant Power, Varying Chip Rate, Constant Symbol Rate

Quantity	Formula or Nomenclature	$N_c = 1$ Base Case	$N_c = 11$ vs N_c=1	$N_c = 127$ vs N_c=1
# chips/symbol	N_c	1	11	127
Symbol period (s)	T_s	10^{-6}	1	1
Symbol rate (symbol/s)	$1/T_s$	10^6	1	1
Chip period (s)	$T_c = T_s / N_c$	10^{-6}	1/11	1/127
Chip rate (chip/s)	N_c/T_s	10^6	11	127
Symbol energy (J)	E_s	10^{-6}	1	1
Chip energy (J)	$E_c = E_s/N_c$	10^{-6}	1/11	1/127
Signal out of correlator (V)	$N_c \sqrt{E_s/T_s}$	$\sqrt{E_s/T_s}$	11	127
RMS noise into correlator (V)	$\sqrt{N_o N_c/T_s}$	$\sqrt{N_o/T_s}$	$\sqrt{11}$	$\sqrt{127}$
RMS noise out of correlator (V)	$\sqrt{N_c}\sqrt{N_o N_c/T_s}$	$\sqrt{N_o/T_s}$	11	127
Avg. signal to RMS Gaussian noise out of correlator		$\sqrt{E_s/N_o}$	1	1
E_s/N_o improvement from spreading (dB)		0	0	0

Incoherent Line Interferers Uniformly Distributed in Band (i.e., number increases with bandwidth)

$$L(t) = \sqrt{2} \sum^{\kappa N_c} L_i \cos(\omega_i t + \phi_i) \quad \text{where } \omega_i/2\pi < B_c$$

Interference power into correlator (W)	$\sum^{\kappa N_c} L_i^2$	$\sum^{\kappa} L_i^2$	11	127
RMS interference into correlator (V)	$\sqrt{\sum^{\kappa N_c} L_i^2}$	$\sqrt{\sum^{\kappa} L_i^2}$	$\sqrt{11}$	$\sqrt{127}$
RMS interference out of correlator (V)	$\sqrt{N_c}\sqrt{\sum^{\kappa N_c} L_i^2}$	$\sqrt{\sum^{\kappa} L_i^2}$	11	127
Avg. signal to RMS interference out of correlator	$\sqrt{E_s/(T_s \sum^{\kappa} L_i^2)}$	$\sqrt{E_s/(T_s \sum^{\kappa} L_i^2)}$	1	1
E_s/I_o improvement from spreading (dB)		0	0	0

M Incoherent Line Interferers in Band
(i.e., constant number independent of bandwidth)

$L(t) = \sqrt{2} \sum^{M} L_i \cos(\omega_i t + \phi_i)$ where $\omega_i/2\pi < B_c$

Quantity	Formula or Nomenclature	$N_c = 1$ Base Case	$N_c = 11$ vs $N_c=1$	$N_c = 127$ vs $N_c=1$
Interference power into correlator (W)	$\sum^{M} L_i^2$	$\sum^{M} L_i^2$	1	1
RMS interference into correlator (V)	$\sqrt{\sum^{M} L_i^2}$	$\sqrt{\sum^{M} L_i^2}$	1	1
RMS interference out of correlator (V)	$\sqrt{N_c}\sqrt{\sum^{M} L_i^2}$	$\sqrt{\sum^{M} L_i^2}$	$\sqrt{11}$	$\sqrt{127}$
Avg. signal to RMS interference out of correlator	$\sqrt{N_c E_s/(T_s \sum^{M} L_i^2)}$	$\sqrt{E_s/(T_s \sum^{M} L_i^2)}$	$\sqrt{11}$	$\sqrt{127}$
E_s/I_0 improvement from spreading (dB)		0	10.4	21

Single Impulse Interferer
$v(t) = K\,\delta(t)$

Quantity	Formula or Nomenclature	$N_c = 1$ Base Case	$N_c = 11$ vs $N_c=1$	$N_c = 127$ vs $N_c=1$
Energy from filter $2K^2 B_c = 2K^2 N_c / T_s$	$2K^2 N_c / T_s$	$2K^2 / T_s$	11	127
Peak voltage from filter $2 K B_c$	$2K N_c / T_s$	$2K / T_s$	11	127
Peak signal to peak impulse voltage ratio into correlator (V/V)	$\sqrt{E_s T_s} / (2K N_c)$	$\sqrt{E_s T_s} / 2K$	1/11	1/127
Total improvement in clipping potential due to spreading		0	10.4	21
Avg. signal to clipped impulse out of correlator (V/V)				

Constant Power, Constant Chip Rate, Varying Symbol Rate

Quantity	Formula or Nomenclature	$N_c = 1$ Base Case	$N_c = 11$ vs N_c=1	$N_c = 127$ vs N_c=1
# chips/symbol	N_c	1	11	127
Chip period (s)	T_c	10^{-7}	1	1
Chip rate (chip/s)	$1/T_c$	10^7	1	1
Symbol period (s)	$T_s = N_c T_c$	10^{-7}	11	127
Symbol rate (symbol/s)	$N_s = 1/T_s$	10^7	1/11	1/127
Chip energy (J)	E_c	10^{-7}	1	1
Symbol energy (J)	$E_s = N_c E_c$	10^{-7}	11	127
Signal out of correlator (V)	$N_c \sqrt{E_c/T_c}$	$\sqrt{E_c/T_c}$	11	127
RMS noise into correlator (V)	$\sqrt{N_o/T_c}$	$\sqrt{N_o/T_c}$	1	1
RMS noise out of correlator (V)	$\sqrt{N_c}\sqrt{N_o/T_c}$	$\sqrt{N_o/T_c}$	$\sqrt{11}$	$\sqrt{127}$
Avg. signal to RMS Gaussian noise out of correlator	$\sqrt{N_c}\sqrt{E_c/N_o}$	$\sqrt{E_c/N_o}$	$\sqrt{11}$	$\sqrt{127}$
E_s/N_o improvement from spreading (dB)		0	10.4	21

Incoherent Line Interferers in Band
(i.e., constant number independent of bandwidth)

$L(t) = \sqrt{2} \sum L_i \cos(\omega_i t + \phi_i)$ where $\omega_i/2\pi < B_c$

Interference power into correlator (W)	$\sum L_i^2$	$\sum L_i^2$	1	1
RMS interference into correlator (V)	$\sqrt{\sum L_i^2}$	$\sqrt{\sum L_i^2}$	1	1
RMS interference out of correlator (V)	$\sqrt{N_c}\sqrt{\sum L_i^2}$	$\sqrt{\sum L_i^2}$	$\sqrt{11}$	$\sqrt{127}$
Avg. signal to RMS interference out of correlator	$\sqrt{N_c E_s/(T_s \sum L_i^2)}$	$\sqrt{E_s/(T_s \sum L_i^2)}$	$\sqrt{11}$	$\sqrt{127}$
E_s/I_o improvement from spreading (dB)		0	10.4	21

Single Impulse Interferer

Quantity	Formula or Nomenclature	$N_c = 1$ Base Case	$N_c = 11$ vs $N_c=1$	$N_c = 127$ vs $N_c=1$
$v(t) = K\,\delta(t)$				
Energy from filter $2K^2 B_c = 2K^2 / T_c$	$2K^2 / T_c$	$2K^2 / T_c$	1	1
Peak voltage from filter $2 K B_c$	$2K / T_c$	$2K / T_c$	1	1
Peak signal to peak impulse voltage ratio into correlator (V/V)	$\sqrt{E_c T_c} / (2K)$	$\sqrt{E_c T_c} / 2K$	1	1
Total improvement in clipping potential due to spreading		0	0	0
Avg. signal to peak impulse out of correlator (V/V)	$\frac{N_c}{2K}\sqrt{E_c T_c}$	$\frac{1}{2K}\sqrt{E_c T_c}$	11	127
·ovement due to spreading (dB)		0	10.4	21