COMPUTATION BOOK

NAME Jack Welch

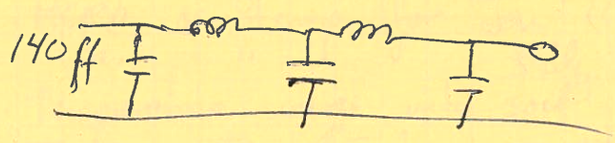
Course ATA #3

AMPAD Dallas, TX 75252

22-157

47-18-80 slu. Labs.

Hat Creek Computing and Control: II (140), IV (62)



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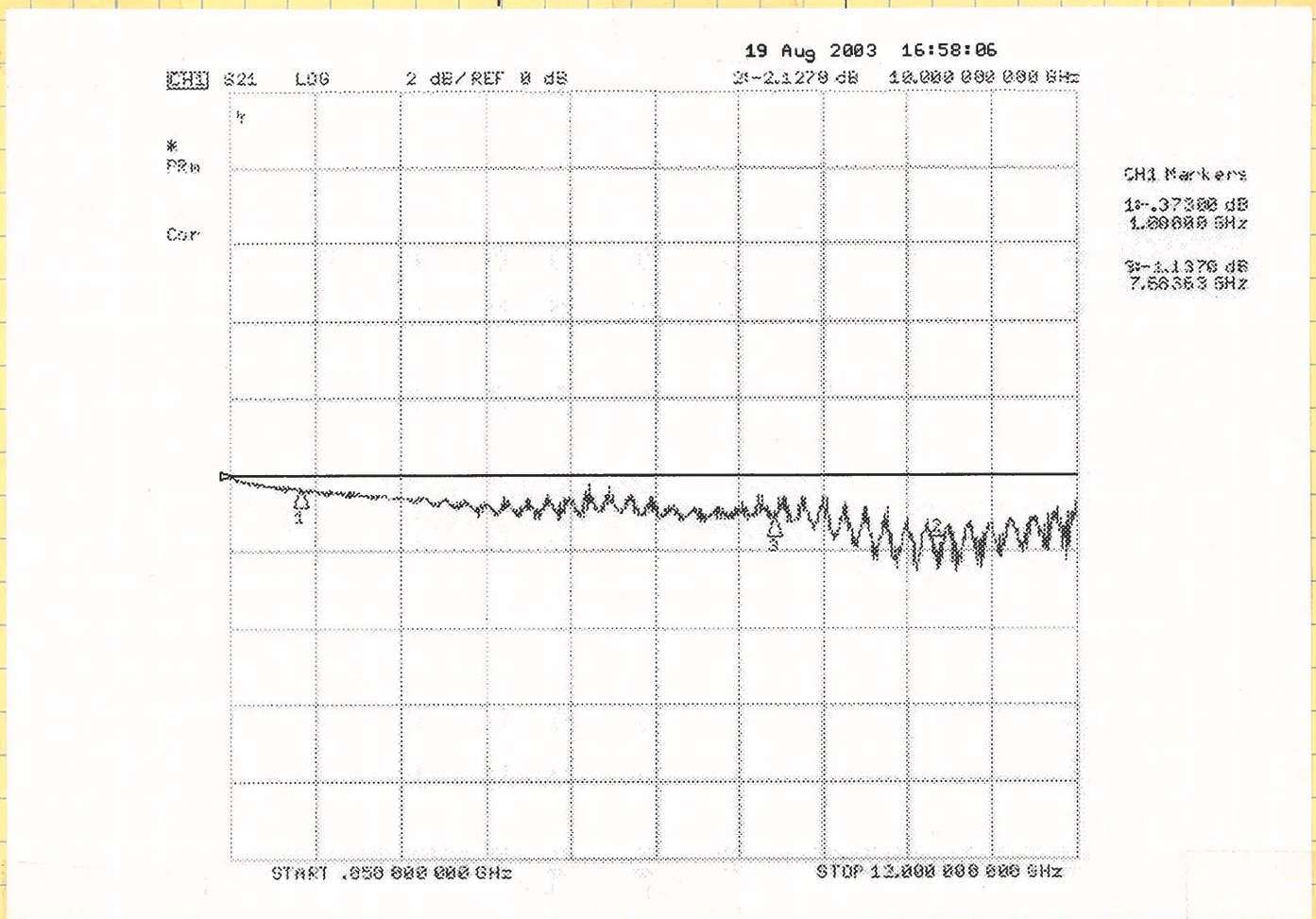
Some measurements of short pieces of coax cable which will be used on the Testfeed.

~ 13" piece of .047 coax with SMA male ends.



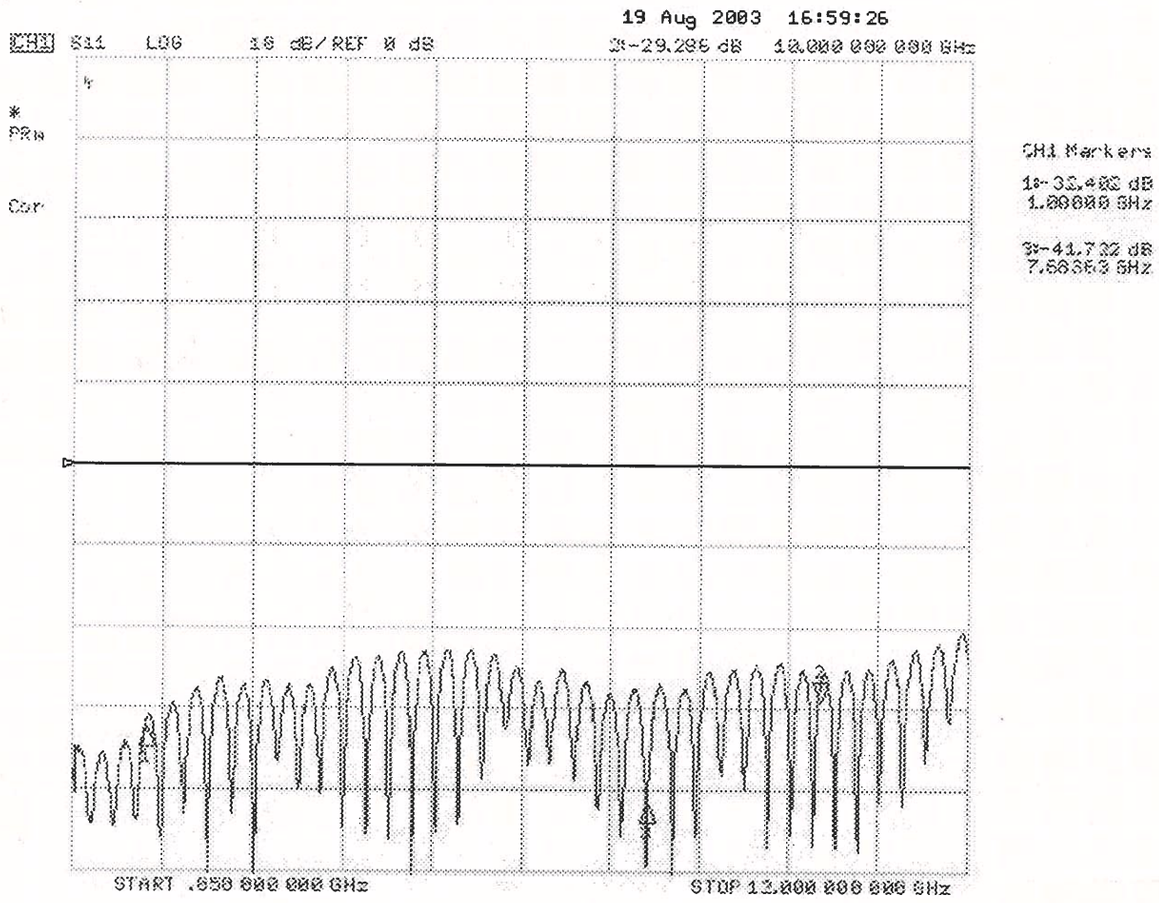
sweep time 1s 801 points

The piece was then cut in two, and each half was used for a polarization. A six foot length work horse was connected bringing the signal out the back.

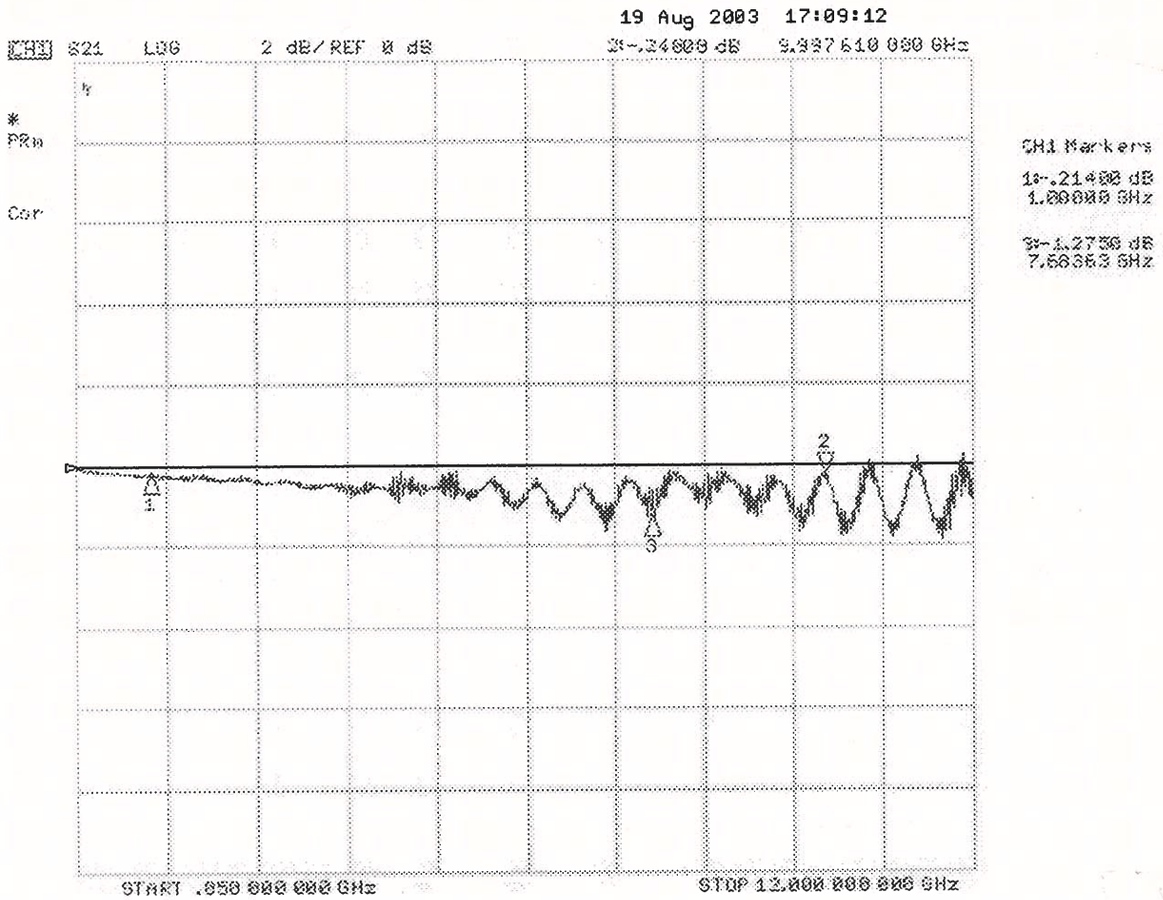
(147) gain and NF for RFCB

(146) Temp. stability measurements on the PAX

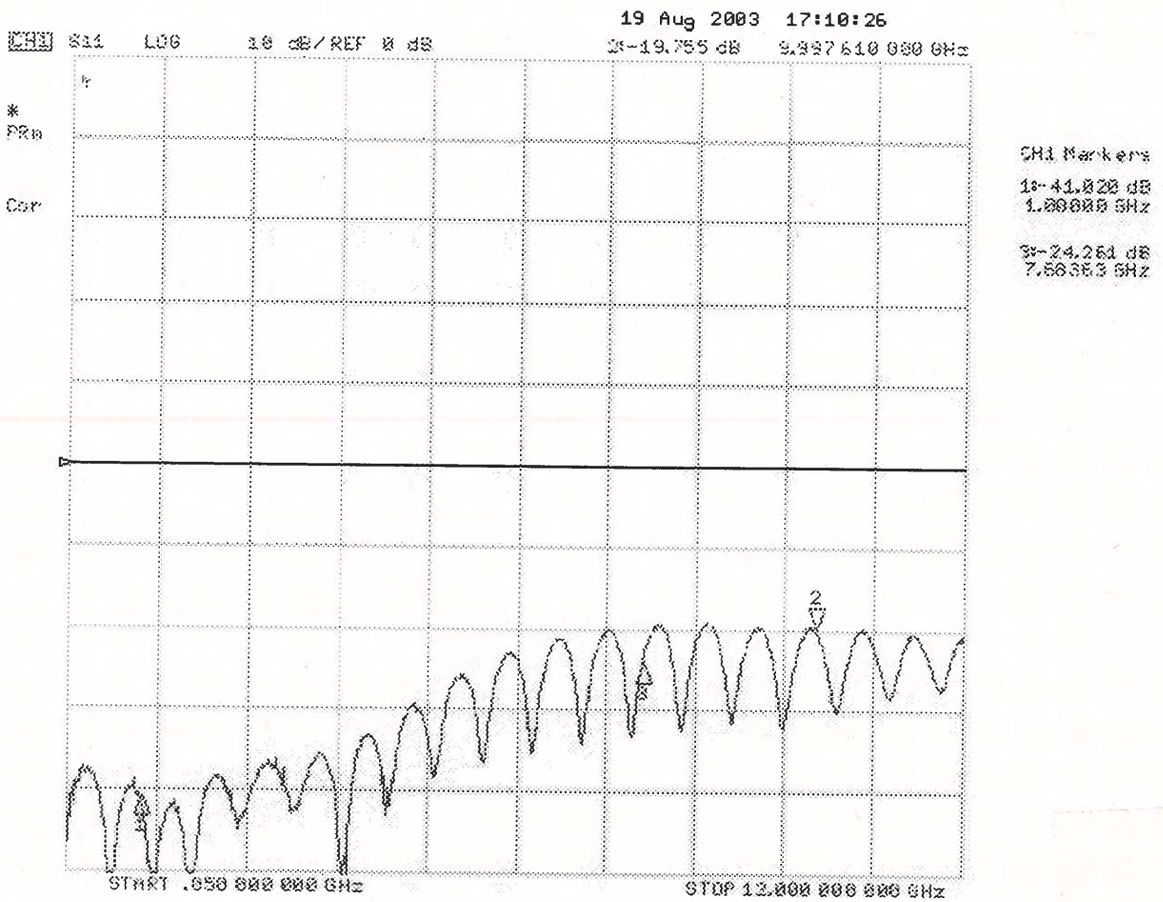
(148) Temp. measurments on 2K



Same 13" piece of 047 with male SMA connectors



6" length of .085 with male ends and stainless steel outer conductor



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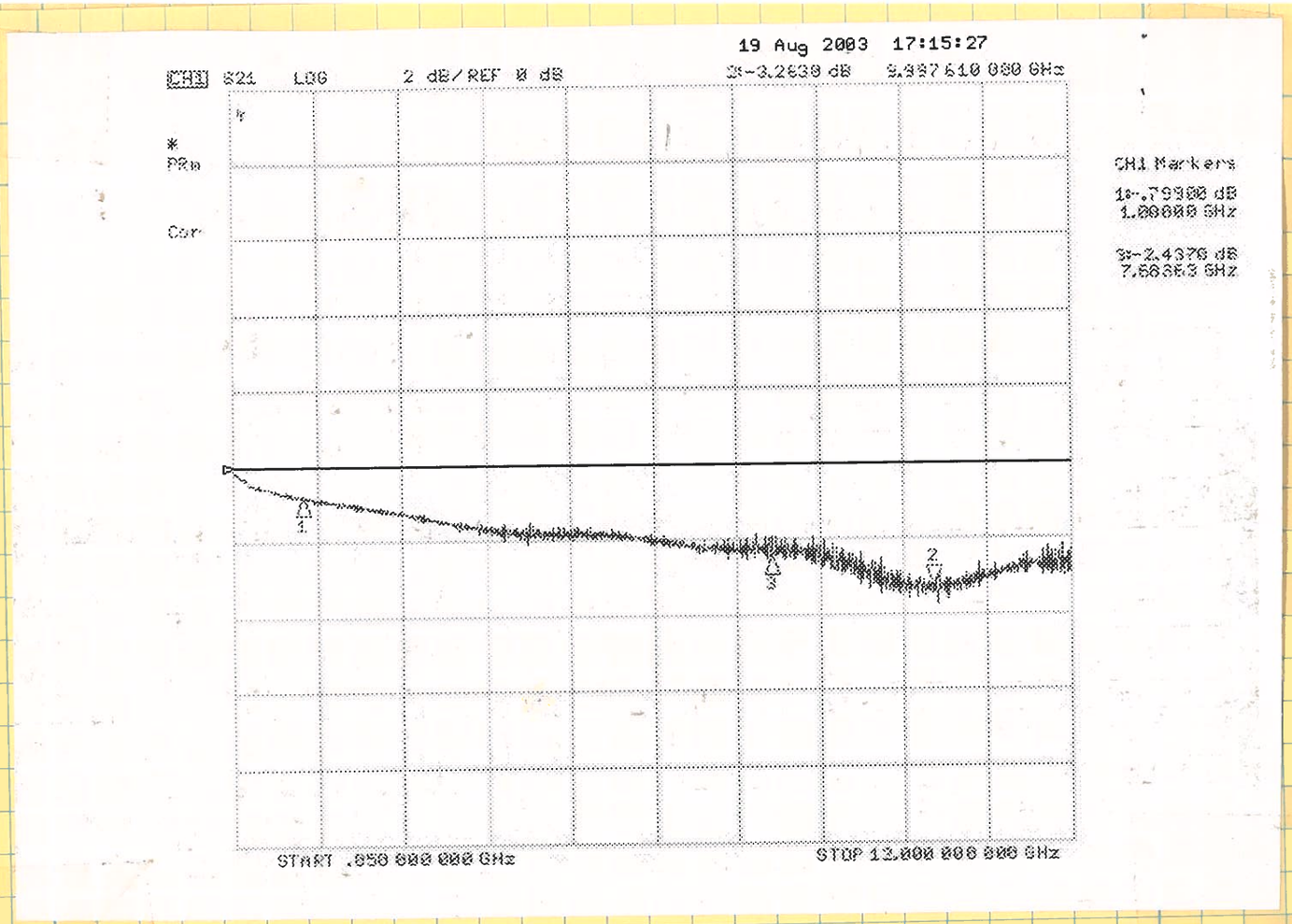
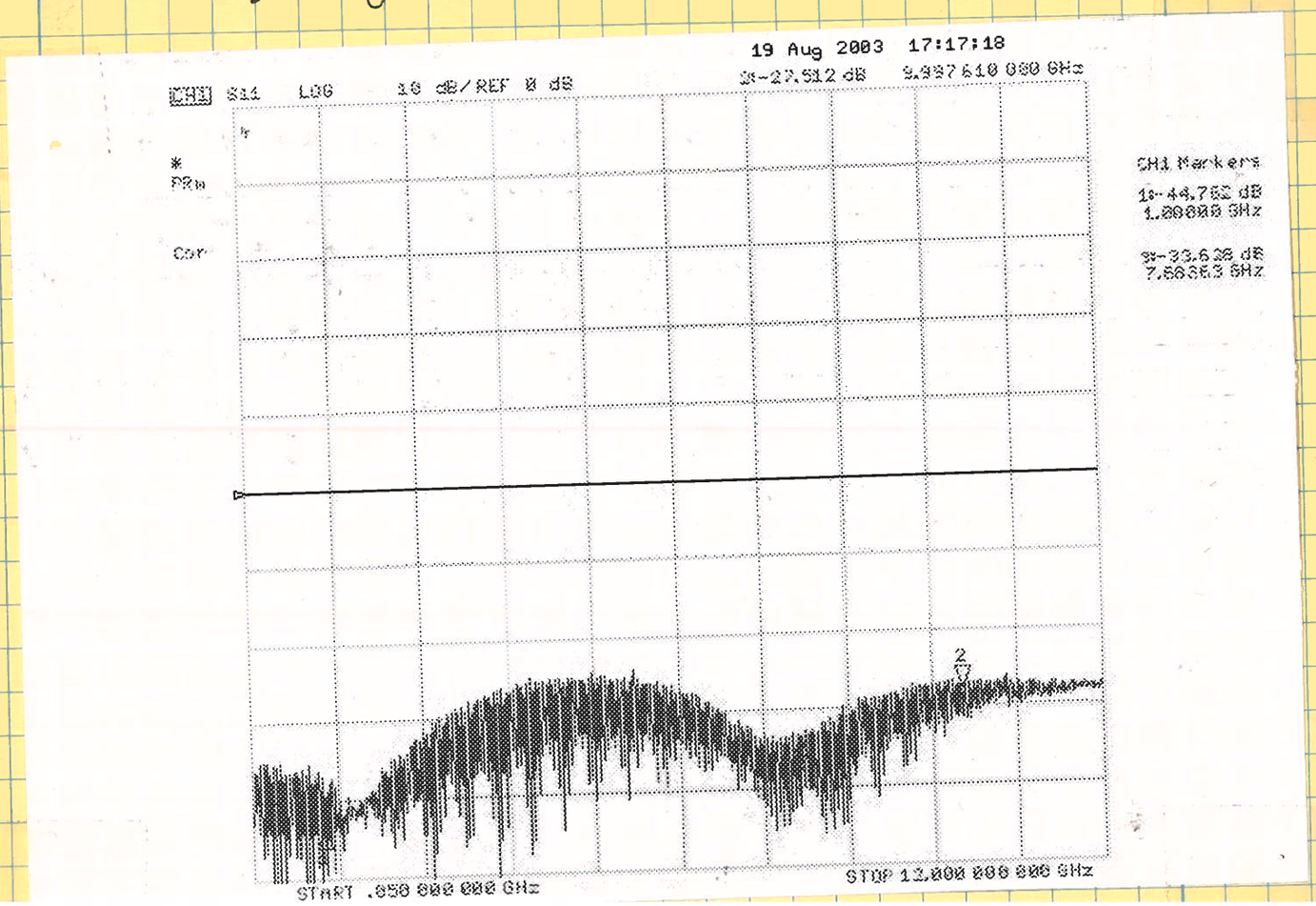


6' length of work house cable



Candidate LNA output Cables

Cryogenic (Stainless Steel Outer Jacket)

?

	PART NUMBER (un-plated)	CS50047	AN50085	BN50085	AS50085	BS50085	JN50085
	Silver Plated (QQ-S-365)	–	–	–	–	–	–
	Tin Plated (MIL-T-10727)	–	–	–	–	–	–
MECHANICAL DATA	**Category***	**Special**	**Special**	**Special**	**Special**	**Special**	**Special**
	Inner Conductor O.D. in. (mm) Tolerance (±)	.0113 (.287) .0005 (.0127)	.0201 (.510) .0005 (.0127)	.0201 (.510) .0005 (.0127)	.0201 (.510) .0005 (.0127)	.0201 (.510) .0005 (.0127)	.0201 (.510) .0005 (.0127)
	Dielectric O.D. in. (mm) Tolerance (±)	.037 (0.94) .001 (.0254)	.066 (1.676) .001 (.0254)	.066 (1.676) .001 (.0254)	.066 (1.676) .001 (.0254)	.066 (1.676) .001 (.0254)	.066 (1.676) .001 (.0254)
	Inner Conductor Material	Silver Plated solid Copper-clad Steel	Silver Plated solid Copper-clad Steel	Silver Plated solid Copper	Silver Plated solid Copper-clad Steel	Silver Plated solid Copper	Silver Plated solid Beryllium-Copper
	Dielectric Material	Solid PTFE	Solid PTFE	Solid PTFE	Solid PTFE	Solid PTFE	Solid PTFE
	Outer Conductor O.D. in. (mm) ** Tolerance (±)	.047 (1.194) .002 (.051)	.085 (2.159) .001 (.0254)	.085 (2.159) .001 (.0254)	.085 (2.159) .001 (.0254)	.085 (2.159) .001 (.0254)	.085 (2.159) .001 (.0254)
	Outer Conductor Material	304SS	304SS/Cu	304SS/Cu	304SS	304SS	304SS/Cu
	Safe Bend Radius in. (mm) (Cable center line)	.150 (3.81)	.250 (6.350)	.250 (6.350)	.250 (6.350)	.250 (6.350)	.250 (6.350)
	Weight lbs./100 ft. (Kg/100M) **	.36 (0.53)	1.215 (1.807)	1.25 (1.85)	1.191 (1.771)	1.191 (1.771)	1.215 (1.807)
	Max. Operating Temp. (°C.)	200	200	200	200	200	200
	Typical Length Made, ft. (M)	15 (4.6)	15 (4.6)	15 (4.6)	15 (4.6)	15 (4.6)	15 (4.6)
ELECTRICAL CHARACTERISTICS	**Impedance** (Ohms)	50 ± 3	50 ± 1	50 ± 1	50 ± 1	50 ± 1	50 ± 1
	Dielectric Strength (KVRMS, 60 Hz)	3.0	5.0	5.0	5.0	5.0	5.0
	Corona Extinction Voltage (KVRMS, 60 Hz)	1.5	1.5	1.5	1.5	1.5	1.5
	Higher Mode Cutoff (GHz)	109	60	60	60	60	60
	Inner Conductor D.C. res. (Ohms) (@ 20 deg. C.)	203/1000 ft. 67/100 M	64/1000 ft. 21/100 M	25.2/1000 ft. 8.26/100 M	64/1000 ft. 21/100 M	25.2/1000 ft. 8.26/100 M	84/1000 ft. 27/100 M
	Nominal Capacitance (pF/ft.) (pF/M)	29.3 96.1	29.3 96.1	29.3 96.1	29.3 96.1	29.3 96.1	29.3 96.1
	Nominal Velocity of Propagation (%)	70.0	70.0	70.0	70.0	70.0	70.0
	Safe Operating Voltage (KVRMS, 60 Hz)	1.05	1.05	1.05	1.05	1.05	1.05

A (Attenuation, dB/100 ft (dB/100M) at 20 °C., and P (Average Power Rating, Watts, unity VSWR. 40 °C ambient still air)

Frequency, GHz	A	P	A	P	A	P	A	P	A	P	A	P
0.100	23.7	161	6.0	740	6.0	740	13.9	315	13.9	315	8.0	525
0.500	53.4	65	13.5	345	13.5	345	32.5	148	32.5	148	17.2	250
1.000	75.5	44	20.0	245	20.0	245	45.8	105	45.8	105	25.0	195
5.000	171	18	47.0	105	47.0	105	105	47	105	47	50.5	105
10.000	244	12	71.0	70	71.0	70	156	32	156	32	72.0	69
15.000	301	9.7	90.0	56	90.0	56	190	26	190	26	90.0	56
18.000	331	8.7	101	49	101	49	216	23	216	23	101	49
20.000	350	8.2	106	46	106	46	225	20	225	20	106	46
30.000	428	6.6	133	37	133	37	277	16	277	16	133	37
40.000	495	5.6	155	32	155	32	323	14	323	14	155	32

*** "Special" may require longer lead times, minimum quantities and/or a setup charge.**
All plated cables are subject to longer lead times and minimum quantities.

**** Specifications for un-plated cable. Add .001" to O.D. and multiply weight by 1.02 for plated cables.**

PRECISION TUBE
COAXITUBE DIVISION
A Mueller Industries Company
Salisbury, MD 21801 • Tel: 410-546-3911 • Fax: 410-546-3913
www.precisiontube.com • coaxsale@voicenet.com

The CS50047 (047) stainless steel cable is apparently available. Using this for the output cables should lower the dewar heat loss without significantly increasing the electrical loss.

Repeat the heat flow calculation of II (pp 121-122) for CS50047. The properties are on the previous page.

(a) <u>The outer conductor</u> $r_o = .024'' = .00061\,m$
$r_i = .019'' = .00048\,m$

$$A = \pi\{(.00061)^2 - (.00048)^2\} = 4.45\times10^{-7}\,m^2$$

$$H_a = A K_0 \left(\frac{T_0}{T_u}\right)^{\gamma} \frac{\alpha T_0}{X_2(m)} = (4.45\times10^{-7})(15)\left(\frac{266.86}{300}\right)^{.46} \frac{(.656)(266.86)}{.15}$$

$= \underline{7.4\text{ mwatts}}$ for a 6" (15cm) length

(b) <u>Teflon dielectric</u> $A = \frac{\pi}{4}\left(.038\times\frac{2.54}{100}\right)^2 = 7.3\times10^{-7}\,m^2$

$$H_b = K A \frac{\Delta T}{\Delta X} = (0.2)(7.3\times10^{-7})\frac{220}{.15} = \underline{0.21\text{ mw}}$$

(c) <u>inner conductor</u> silver plated solid copper clad steel

On the average, Silver has about 80% of the Copper thermal conductivity at the low temperatures: $K_{Cu} \sim 450$ watts/m-K
$K_{Si} \sim 400$ watts/m-K

Plain steel is lower, 60 watts/m-K (although stainless is another order smaller). The area of the steel is much larger than the areas of the plating. For a radius r and a layer thickness Δr, the relative area is $2\Delta r/r$

Assume that $\Delta r = .000100''$ for each layer. Then the total is $.000200'' = .000005\,m = 5$ microns
With the radius $= .000565'' = 1.44\times10^{-4}\,m$,

$$\frac{2\Delta r}{r} = \frac{2(5\times10^{-6})}{1.44\times10^{-4}} = .0694 = \frac{1}{14.4}$$

Lump the silver and copper together.

$$A K_{eff} = K_{ST} A_{ST} + K_{Cu} A_{Cu} \quad ; \quad \frac{425}{60} = 7$$

$k_{ST} = \frac{k_{cu}}{7}$; $A_{ST} = 14.4 A_{cu}$

$(Ak)_{eff} = k_{ST} A_{ST} + 7 k_{ST} \cdot \frac{A_{ST}}{14.4} = 1.49\, k_{ST} A_{ST}$

So the total heat flow is $H_c = (Ak_{eff}) \frac{\Delta T}{\Delta x}$

$H_c = (1.49)(60)(6.5 \times 10^{-8}) \frac{20}{.15} = \underline{8.5 \text{ mwatts}}$

The total heat flow is $7.4 + .21 + 8.5 = \underline{16.1 \text{ mwatts}}$

This is just about half the heat flow in the 085 SS cable.

If the length is increased to 8", the total will fall to $\sim \frac{6}{8} \times 16.1 = \underline{\underline{12 \text{ mwatts}}}$

The copper cladding may be as thick as 60μ, however, which will transmit a lot of heat.

Can we punch holes in the pyramid near the end of the dewar to let out air?

Jackson (2nd edition p 444) gives the transmission coefficient of a small round hole as

$T = \frac{1}{3}(ka)^2 \cos\alpha$, where α is the angle of incidence. For $2a = 1/40$, $ka \sim 1/10$.

This gives $T = \frac{1}{3}\left(\frac{1}{10}\right)^2 = \frac{1}{300}$ for $\alpha = 0°$. Maybe we should take $\langle \cos\alpha \rangle = \frac{2}{\pi}\int_0^{\pi/2} \cos\alpha \, d\alpha = \frac{2}{\pi}$ and $T = \frac{1}{450}$

That makes it a little smaller. To keep the loss less than one percent, there should be no more than ~ 5 holes per unit area.

The expanded PTFE cable is stable for T ≳ 78°F.

Appendix C- Special Performance Considerations

Phase Change vs. Temperature for Solid vs. Expanded PTFE Cable
Figure 8

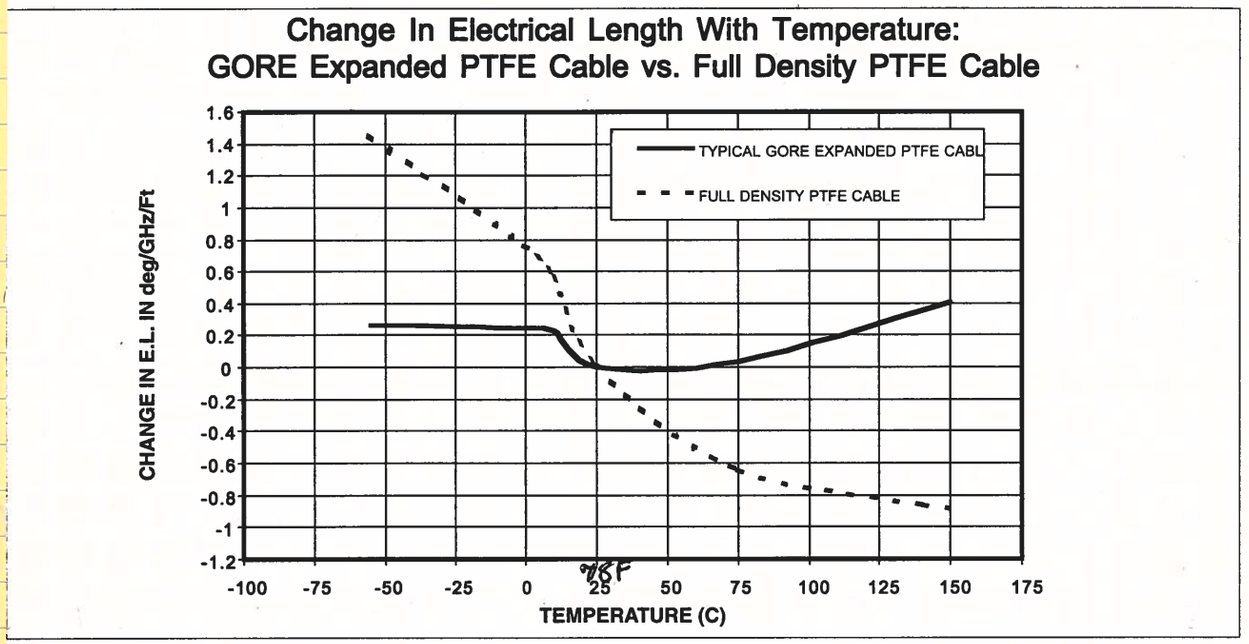


FIGURE 8 PHASE CHANGE VS. TEMPERATURE FOR SOLID VS. EXPANDED PTFE CABLE

Estimate change in electrical length in degrees due to temperature:

First determine the approximate electrical length of the cable in degrees.

$$EL^\circ = \frac{LM(in) \times 360^\circ}{\lambda} \qquad \text{Where: } \lambda(in) = \frac{11.8}{f(\text{GHz})\sqrt{\varepsilon_r}}$$

LM is the mechanical length of the cable in inches.

For Gore cables, ε_r is 1.4 typically. This is a reasonable approximation for GMCAs longer than 2 feet. (This minimizes the effect of connectors.)
Next, determine the change in electrical length (in ppm) due to temperature from the curve of Figure 9.

Change in electrical length in degrees is then determined as follows:

$$\Delta EL^\circ = (EL^\circ)(\Delta ELppm) \times 10^{-6}$$

Example:

A 10ft GMCA operating at 4.2 GHz over the temperature change of +25° C to -55° C:

$$EL^\circ = \frac{120'' \times 360^\circ \times 4.2 \times \sqrt{1.4}}{11.8} = 18{,}193^\circ$$

From Figure 9, ΔELppm=590

$$\Delta EL^\circ = 18{,}193 \times 590 \times 10^{-6} = 10.7^\circ$$

The cable will be approximately 10.7° longer at -55° C than it is at +25° C.

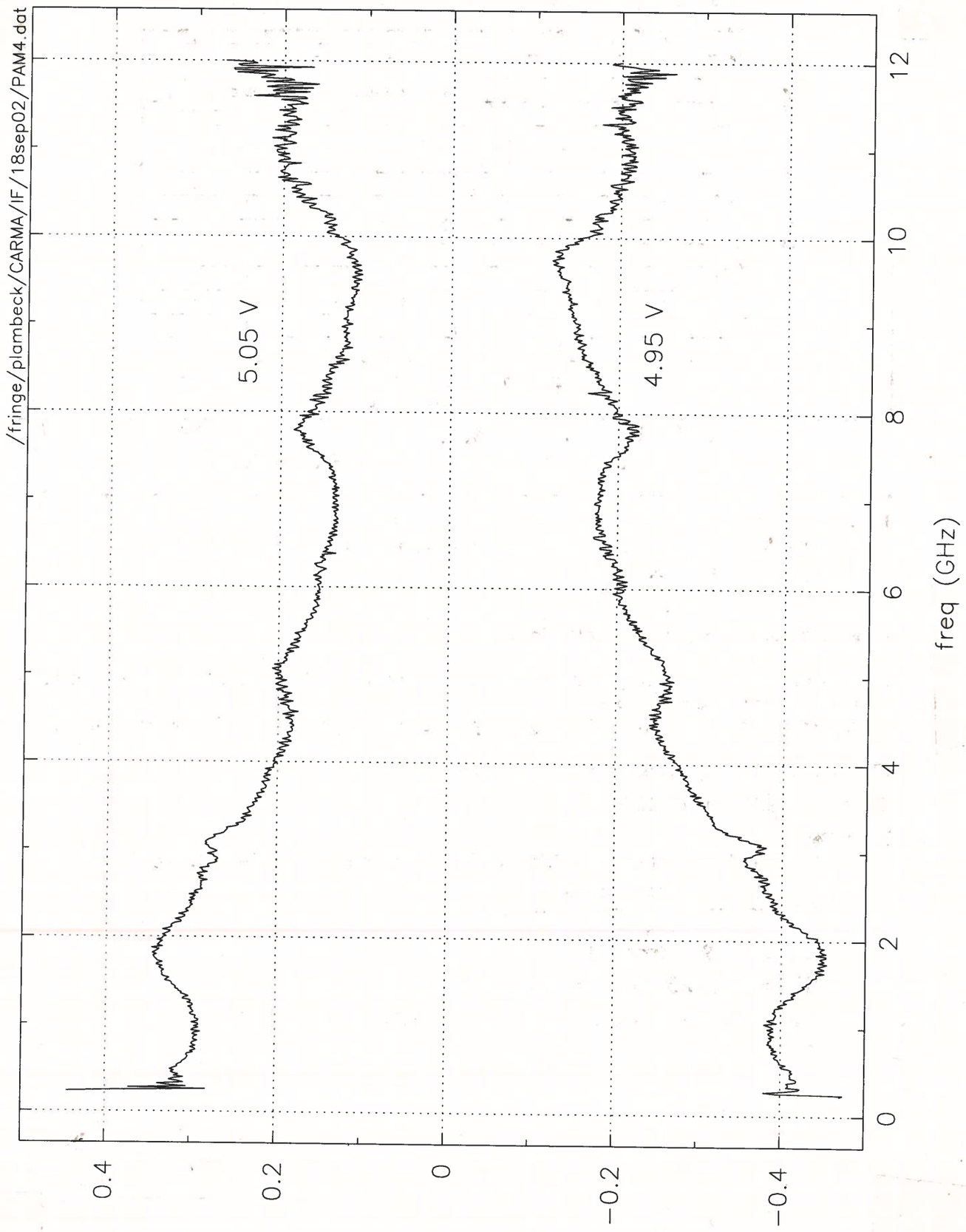
Free space: $\Delta\phi(\text{deg}) = 360^\circ f(\text{GHz})\, d(\text{feet}) = 360^\circ \frac{f}{c} d$

Gain/supply stability for PAM4. $\Delta G \sim 0.3$db for 50mv change

0.3db/50mv ⇒ .006db/mv ⇒ .14% per mv

For temperature at midband .02db/c° ⇒ .002db/0.1°c ⇒ .05% per 0.1°

0.5%/c°



Tests of the New Feed with the 8722ES,

Nov 1, 2003

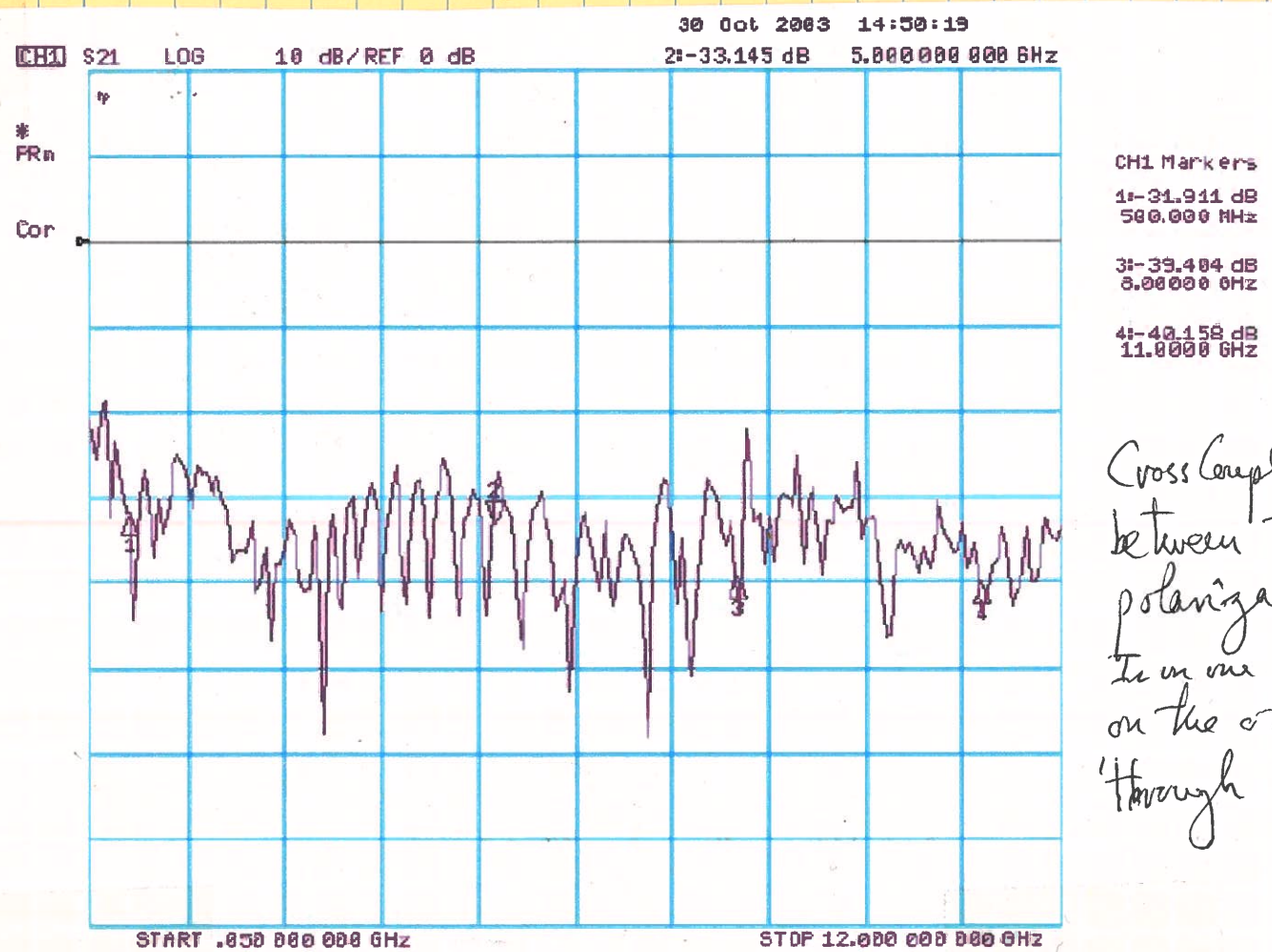
Check of the 8722ES transform mode. (11/10/03)

Set the 8722 for 0.050 GHz – 12.00 GHz. then "transform" and Set Freq. Low Pass 12.00 → 11.999. Then S_{21} Cal on the end of a 2ft piece of workhorse cable. Then transform on, cable open: Peak at t = 0.0 scale is −15 < t < 16 nsec. @ 10 db FWHM is .075 nsec @ 3db FWHM is .024 nsec. [20 pts]

Adding an 8 inch length of semi-rigid cable moved the peak to 2.05 nsec. for the open. 8 inch length of teflon should be close to 1 nsec. so 2 nsec for the round trip makes sense. Also, it moved to the right on the scale.

On the first test, 3 out of the four arms weren't soldered to the tip. That was fixed.

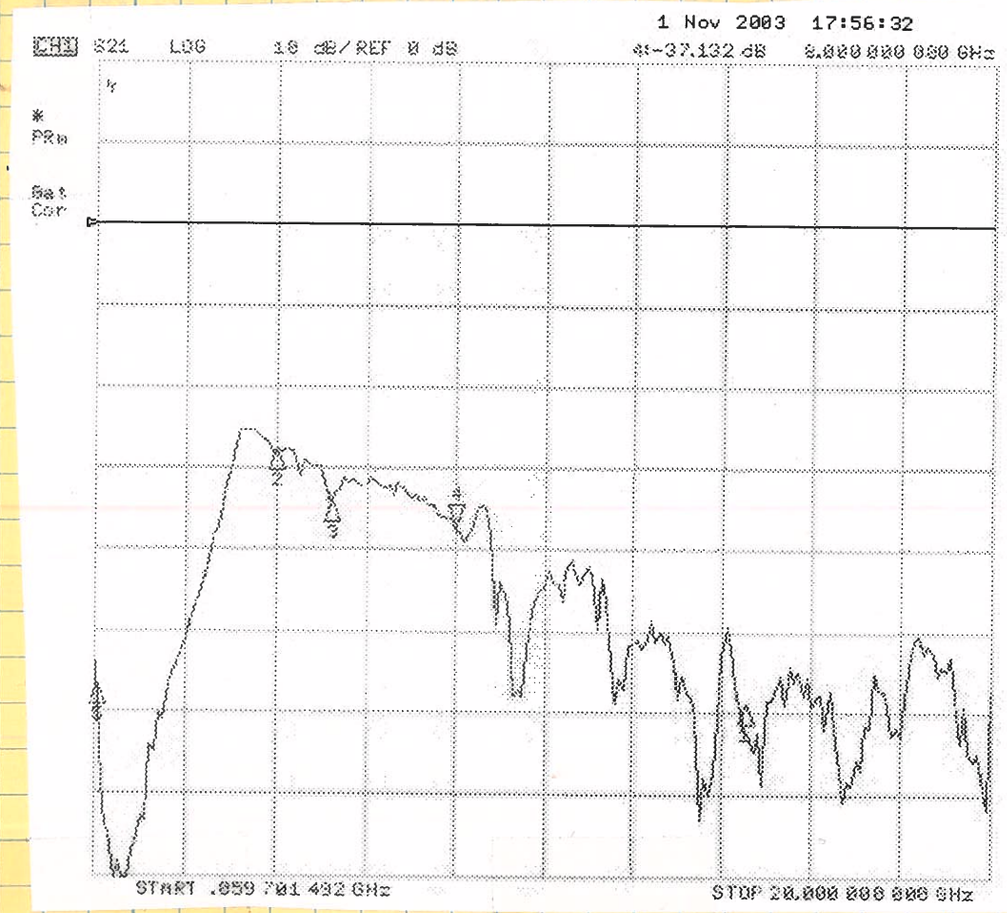
Two six foot lengths of WH cable connected to 8722ES



Cross Coupling between the two polarizations In on one out on the other. 'through cal'

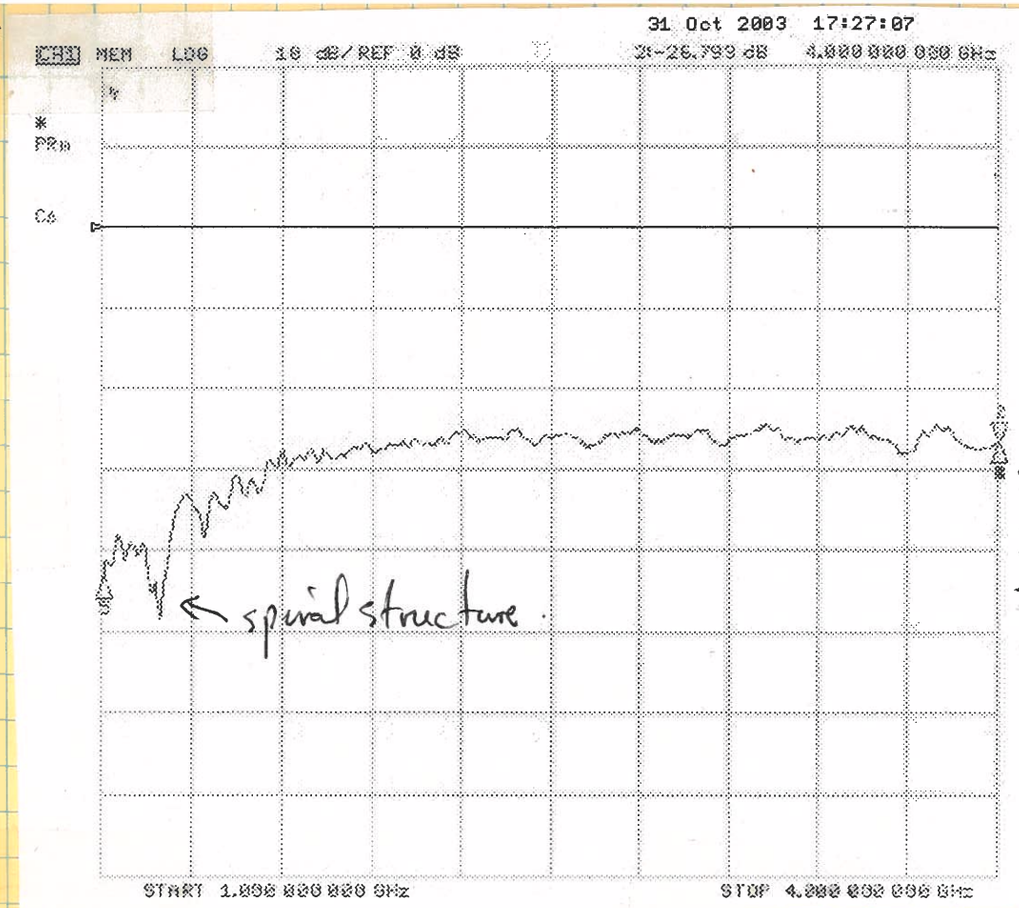
Overall transmission to the small spiral antenna. 0.05–20 GHz. Follows expected 6db/octave out to 14.5 GHz

SCRN12.JPG



SCRN14.JPG

Smooth transmission to the 4–8 GHz coax to waveguide transition used as an antenna. Note the dip at 5.3 GHz associated with the shortest standoffs.



SCRN01.JPG

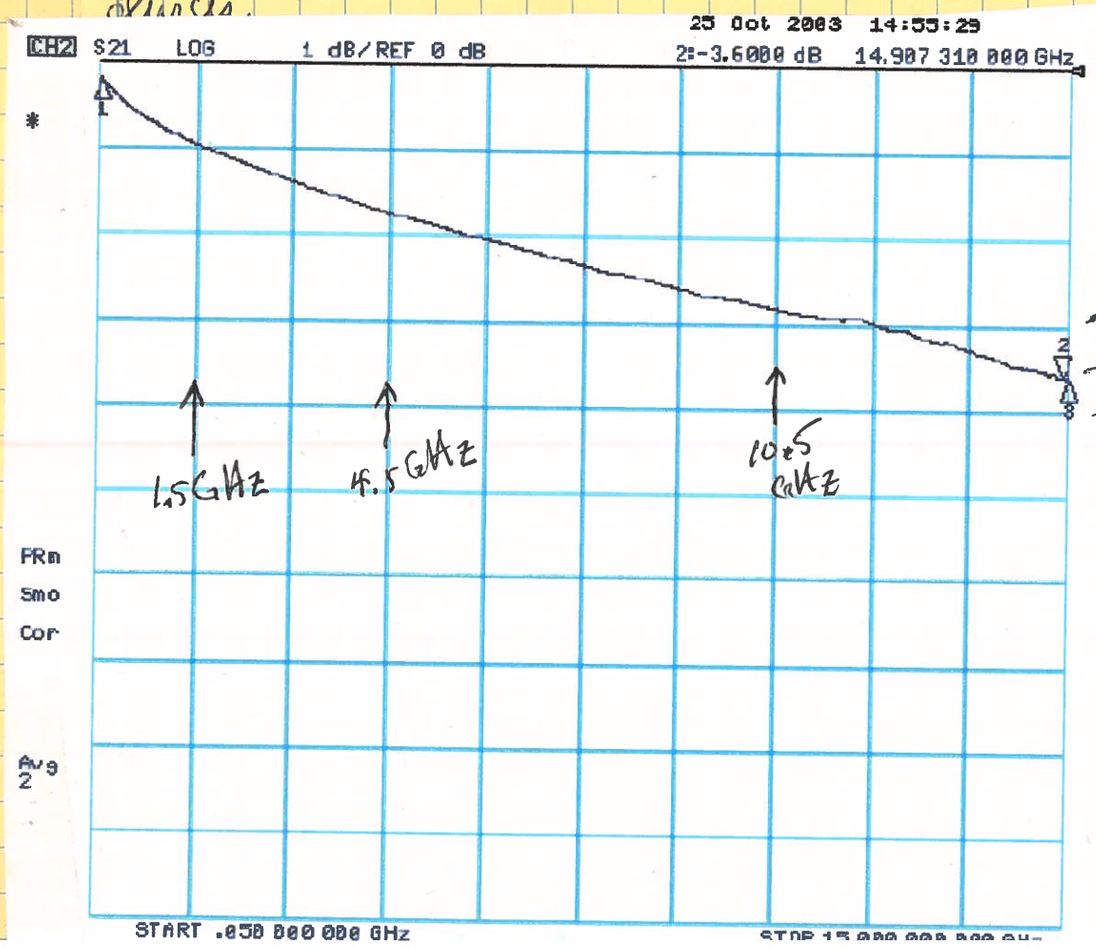
CH1 Markers
1:-26.733 dB 4.00000 GHz
2:-26.793 dB 4.00000 GHz
3:-26.793 dB 4.00000 GHz
4:-43.567 dB 1.00000 GHz

801 pts

← spiral structure

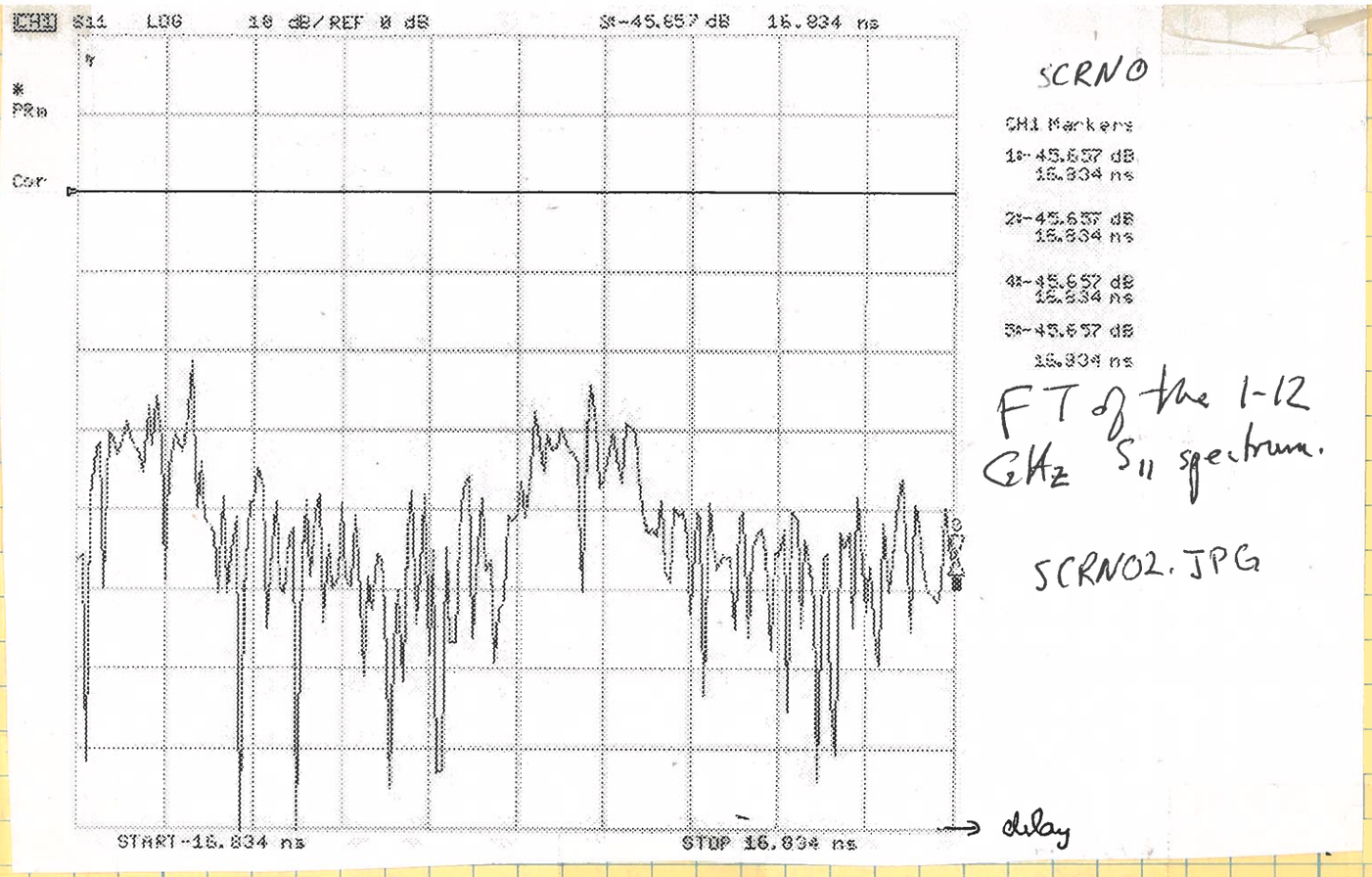
Narrow band scan to small equiangular spiral antenna to search for drop outs. This one: 1-4 GHz. Scans like this to 12 GHz show no dropouts except one at 9.3 GHz in one pol. Only 30db dep.

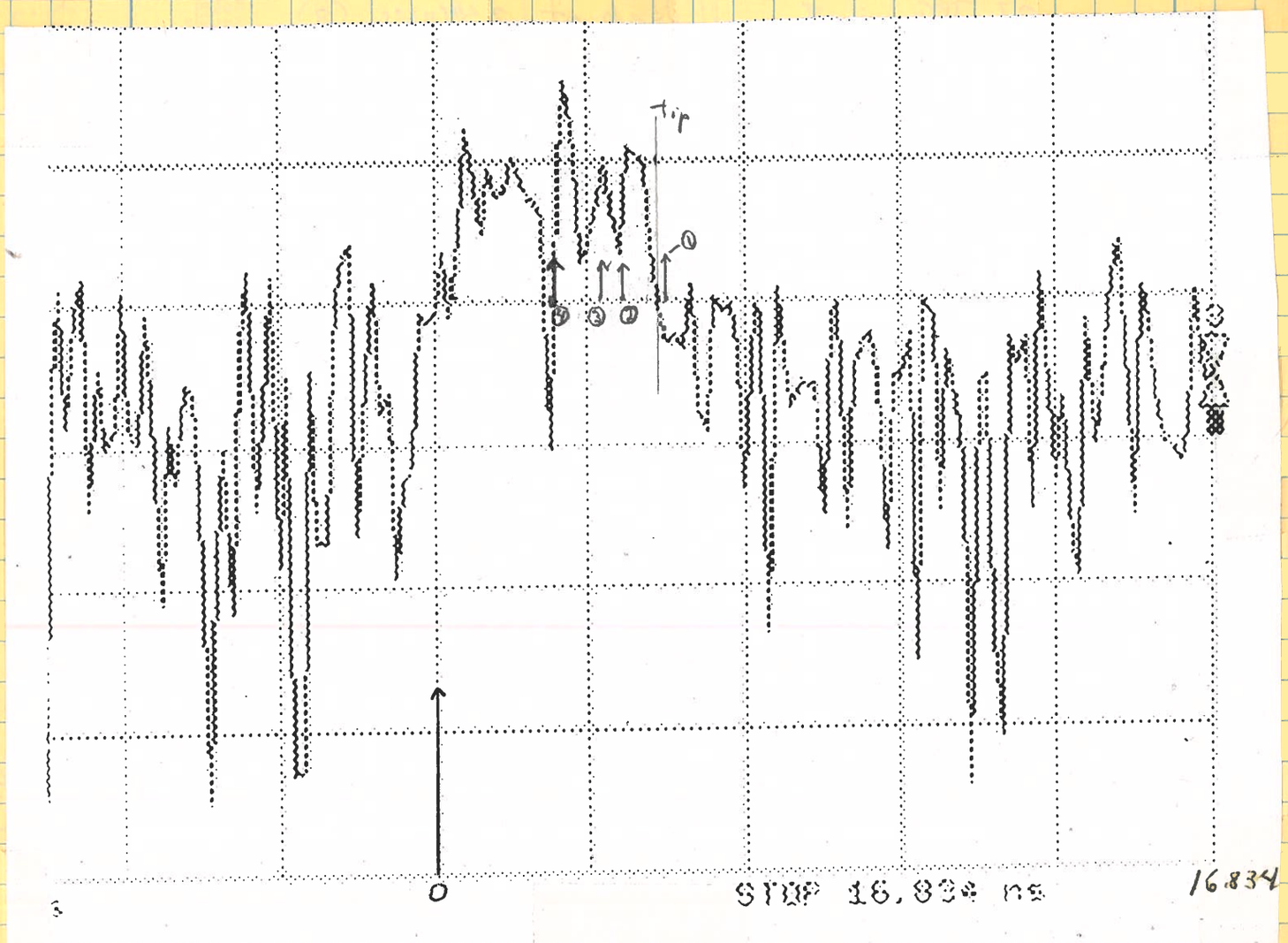
The search for dropouts associated with either the LPent or the dewar interior revealed only one small one at 9.3 GHz in one polarization. Many coax/waveguide transitions and two spiral antennas were used to separate out dropouts in those devices.



CH2 Markers
1:-.21100 dB 50.0000 MHz
3:-3.6240 dB 15.0000 GHz

Transmission through the input cable to the feed.





An expanded version of the above plot showing the gate points. The long line locates the feed tip; a vazer produced a spike.

Time gates were used on the Network Analyzer to show S_{11} beyond various pts along the interior circuit. A razor blade accross the tip circuit board located the tip point along the delay plot. It's shown by the thin line on the previous plot. The measured time resolution on a sharp discontinuity is .075 nsec FWHM at the 10db points. This corresponds to 0.9 inches freespace along the line. It's about half that along the balun where $\epsilon = 4.5$ The .075nsec resolution is about the distance between gate 1 and the tip time.

S_{11} spectra were taken ~~at each~~ with each indicated gate turned on. The tip is at 4.648 nsec.

(also SCRN00.JPG)

SCRN03.JPG is S_{11} for 0-12 GHz scan, 1 pol. no gate
04.JPG is S_{11} same but for the other polarization
05.JPG is S_{11} with the gate at 4.859 nsec. in front of the tip (1)
06.JPG is S_{11} gate at 3.942 (2)
07.JPG is S_{11} gate at 3.44 nsec. (3)
08.JPG is S_{11} gate at 2.393 ns (4)
09.JPG is S_{11} gate at 1.231
10.JPG is S_{11} gate at 30ps

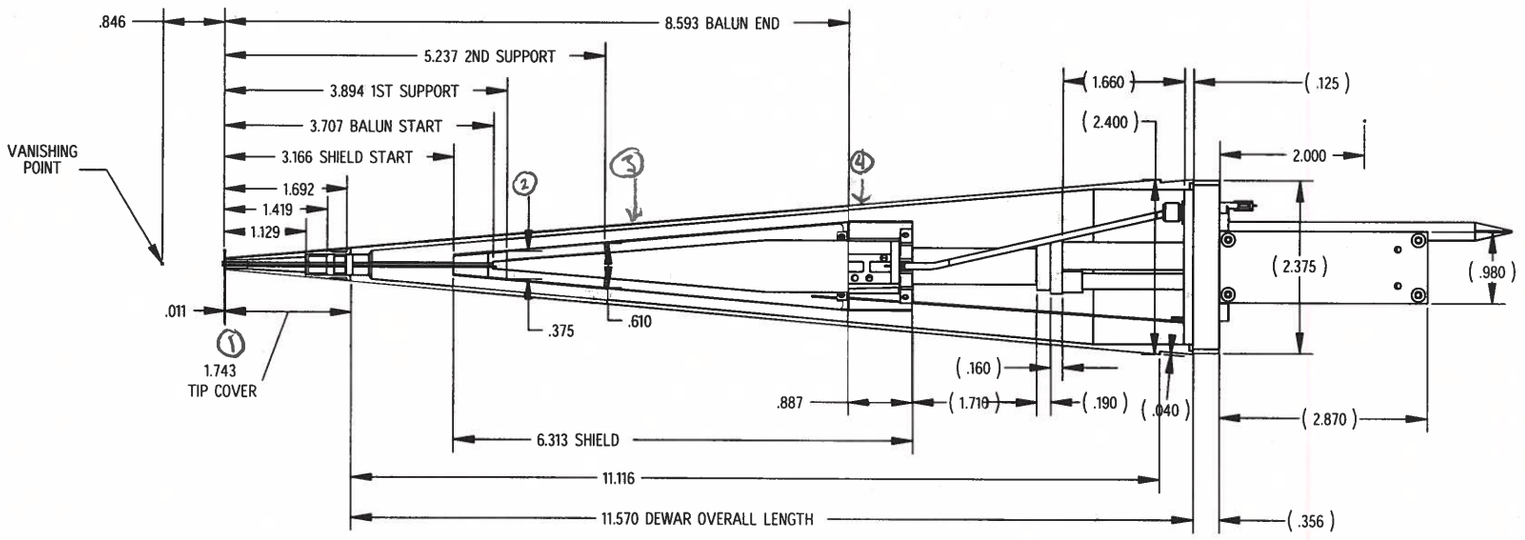
The figure to the right shows the layout of the components inside the pyramid. The effective locations of the four gates are shown. With the gate set at (4) one sees the S_{11} that the LNA will see.

In the following figures S_{11} is shown as seen from each of the gate locations. Only one polarization is shown. The ~~two~~ are essentially identical.

With everything but the antenna gated out at (1) the reflections lie below the -17.5db line. The penciled in line is the correction from p. 14 (double) for the 6 foot WR cable loss. The echos are mostly below the corrected -17.5db line corresponding to VSWR ≤ 1.3 as predicted by the simulations. The bump at 5.3 GHz, seen in both pols corresponds to the first set of standoffs, as can be found from touching the antenna. The corresponding dips can be seen in the figures of p. 13.

17

REVISIONS				
REV	DESCRIPTION	ZONE	NAME	DATE
A	RELEASED		M.C.F.	2002-00-00



PRELIMINARY
2003-11-06

	NAME	DATE
DRAWN	M. C. FLEMING	2002-00-00
CHECKED		
MATERIAL		
FINISH		

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TOLERANCES:	LINEAR	ANGLE
WHOLE & FRACTION	± 1/32	± 1°
DECIMAL X.X	± 0.1	± 0.1°
DECIMAL X.XX	± 0.01	± 0.01°
DECIMAL X.XXX	± 0.005	± 0.001°

BREAK SHARP EDGES 0.010 RAD ± 0.003
MACHINED SURF FINISH < 125 MICROINCH

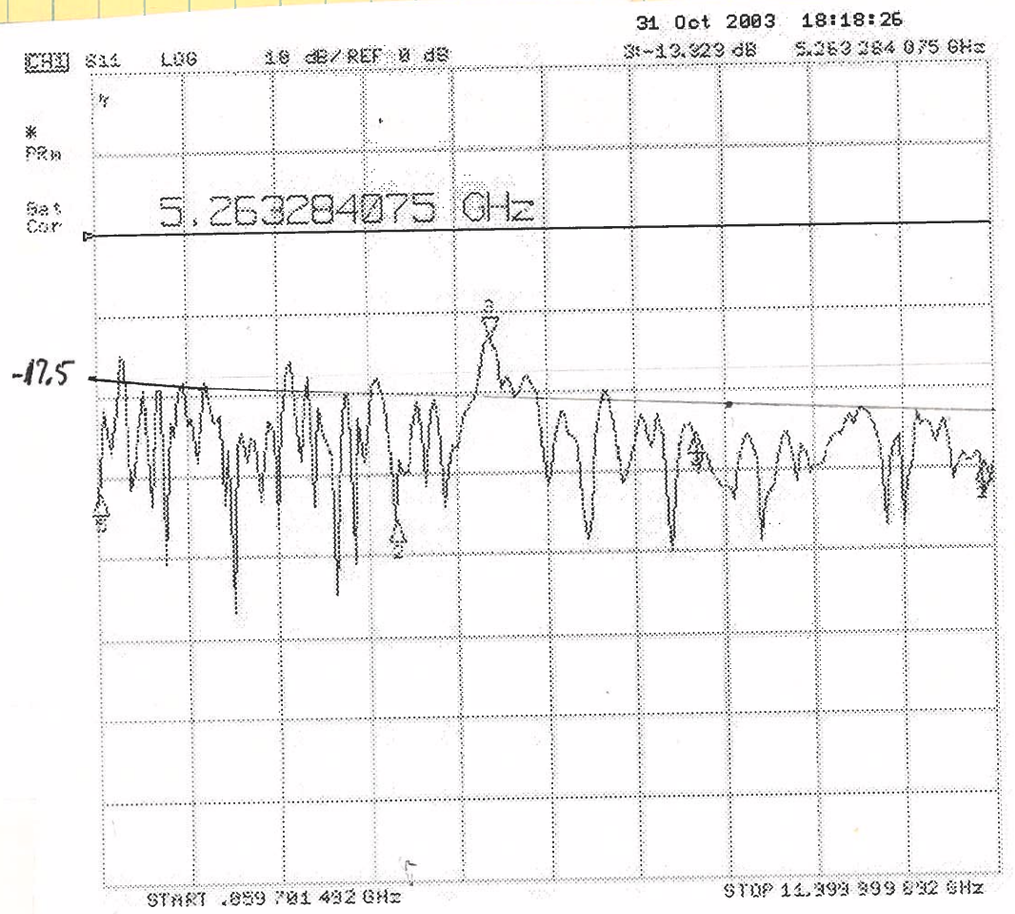
WEIGHT:
SCALE:

RADIO ASTRONLOMY LAB
University of California · Berkeley
Tel: 510-642-5040 Fax: 510-642-3411

NAME

SIZE	DWG NUM:	SHEET	OF	REV.
B	30-28-120	1 of 1		B

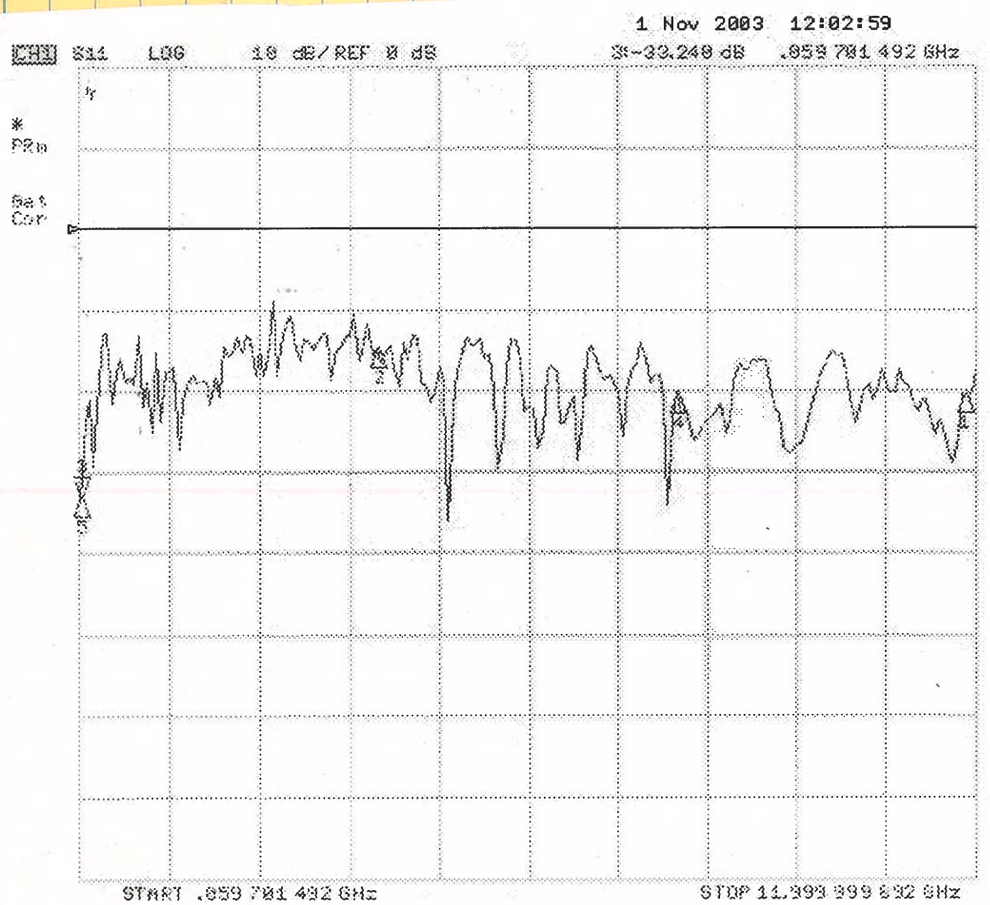
balun length is 4.886". total wire length is 3.707
balun to glass window is 2.015



SCRN00.JPG

CH1 Markers
1:-29.646 dB
11.8852 GHz
2:-35.645 dB
4.00000 GHz
4:-35.697 dB
8.00000 GHz
5:-33.253 dB
53.7814 MHz

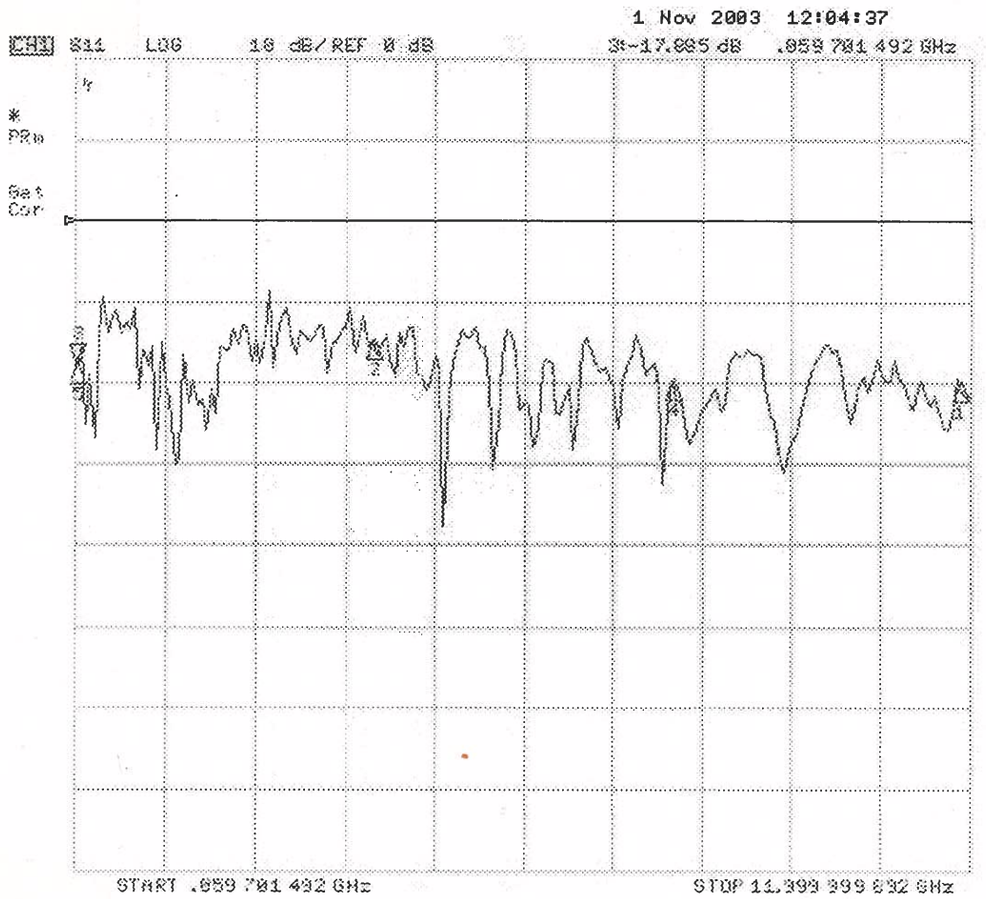
Gate at ①



SCRN06.JPG

CH1 Markers
1:-20.440 dB
11.8852 GHz
2:-14.615 dB
4.00000 GHz
4:-30.394 dB
8.00000 GHz
5:-33.240 dB
53.7814 MHz

Gate at ②



SCRN06.JPG

CH1 Markers

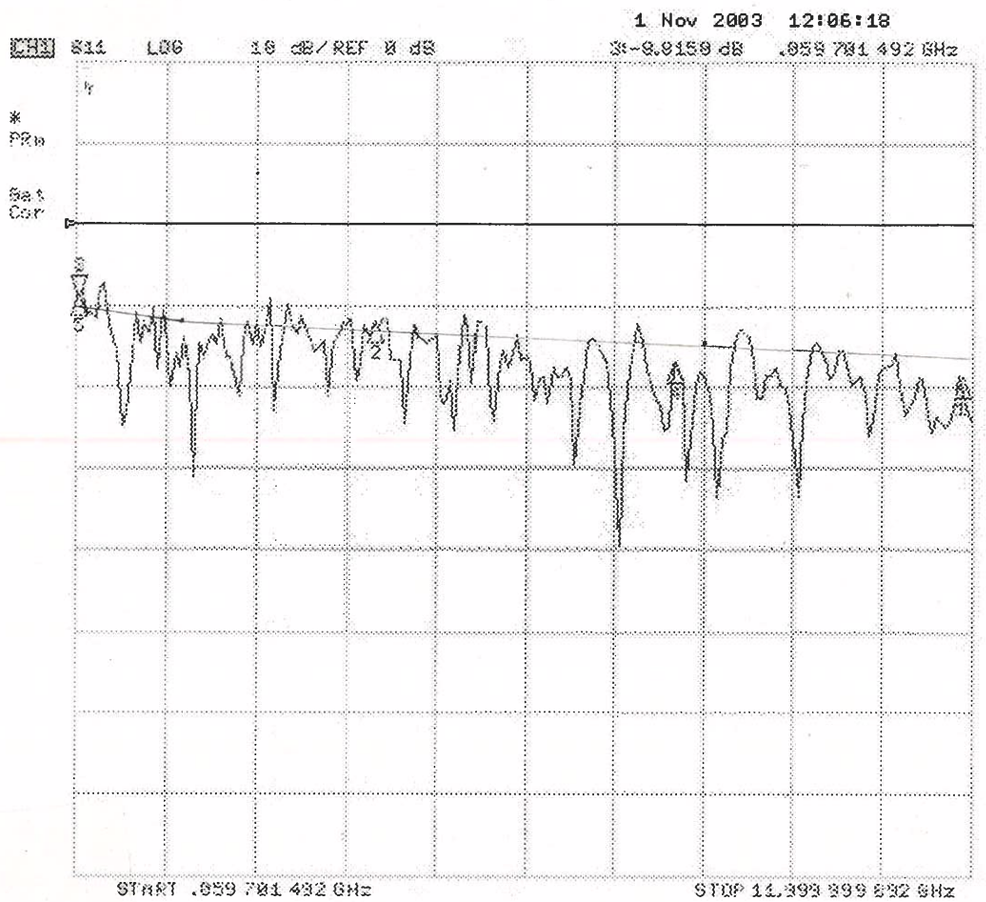
1:-15.870 dB
11.0352 GHz

2:-14.783 dB
4.00000 GHz

4:-19.887 dB
8.00000 GHz

3:-17.885 dB
53.7014 MHz

Gate at 3



SCRN08.JPG

CH1 Markers

1:-15.005 dB
11.0352 GHz

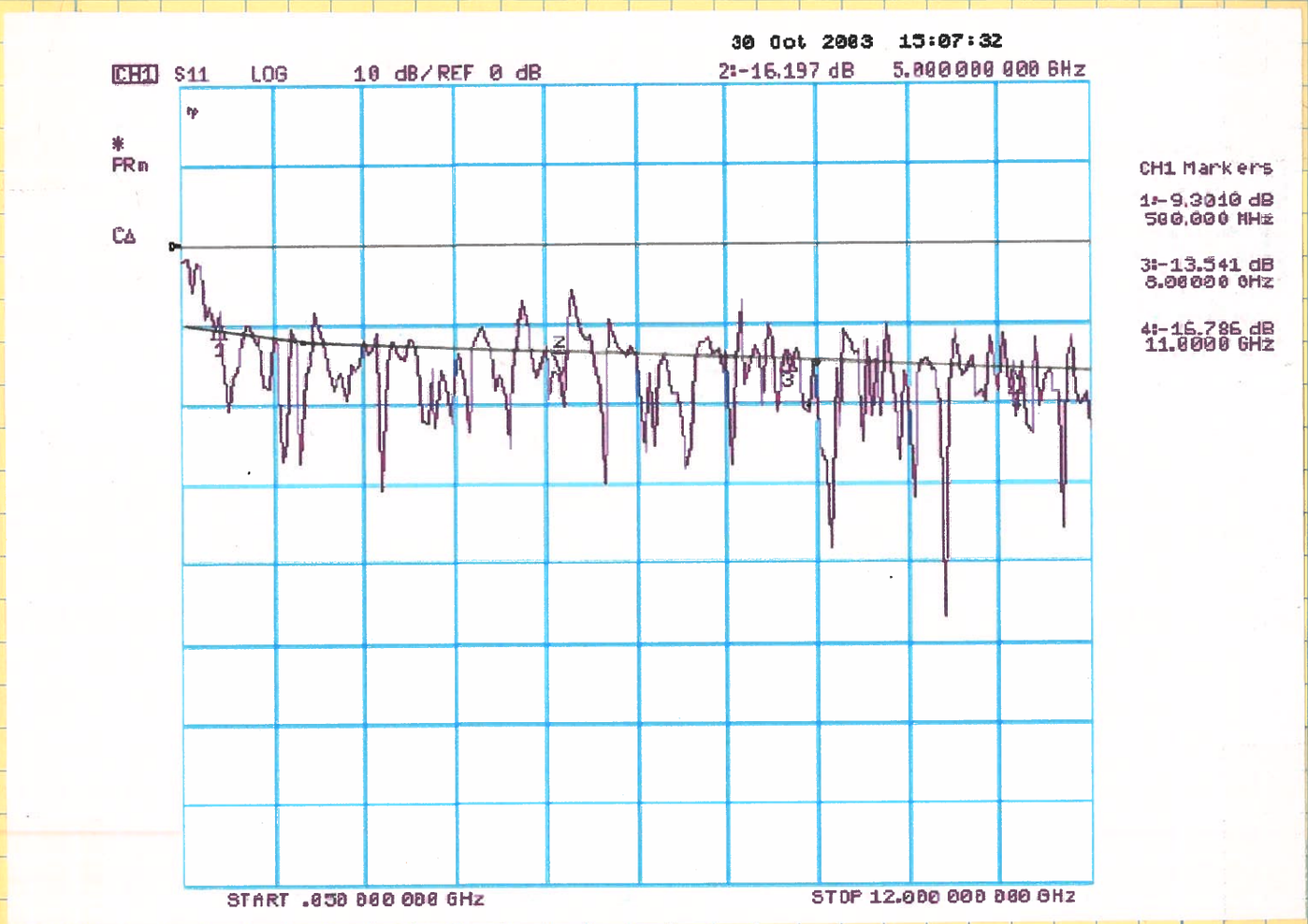
2:-12.241 dB
4.00000 GHz

4:-17.059 dB
8.00000 GHz

3:-8.8159 dB
53.7014 MHz

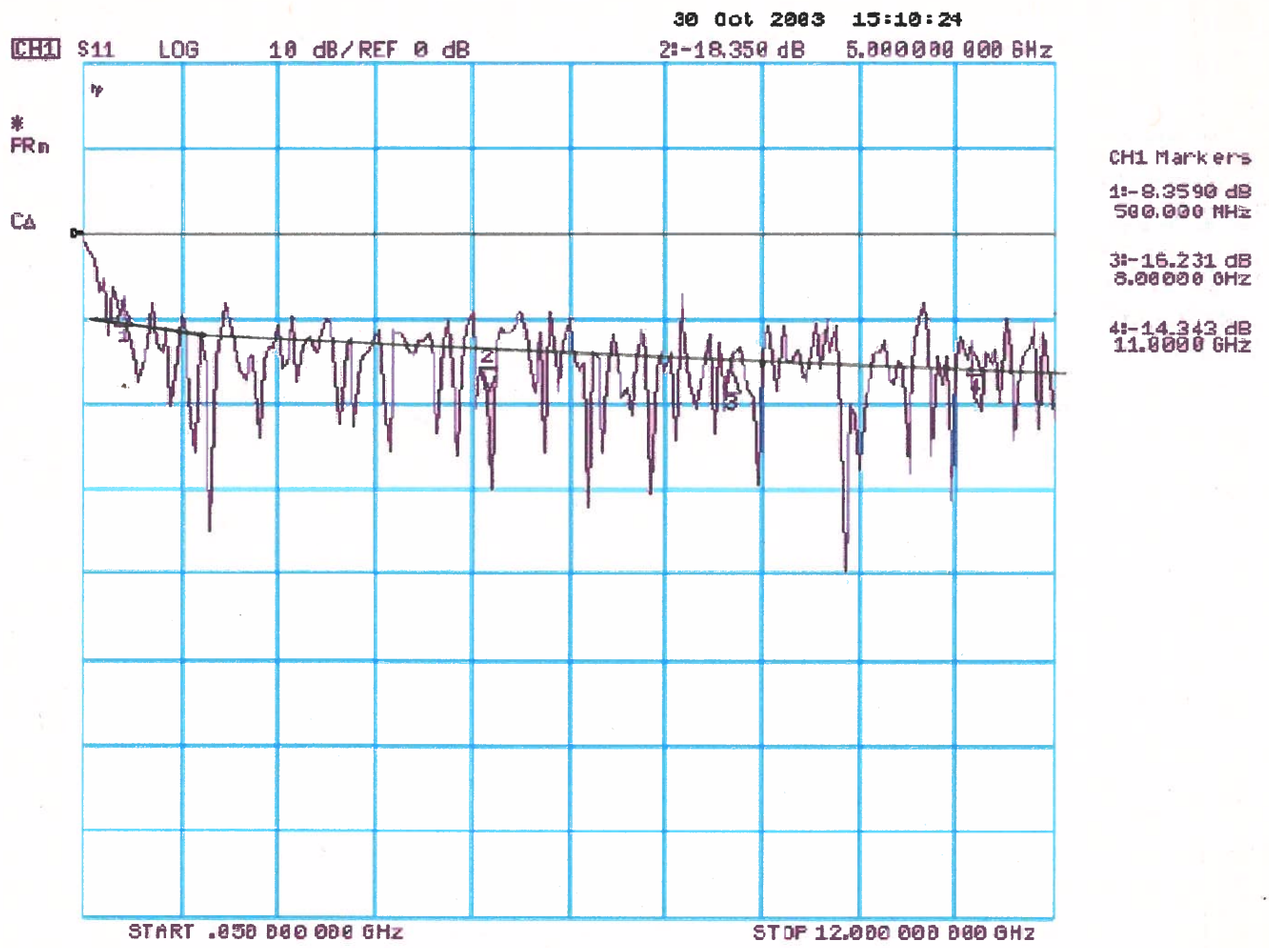
Gate at 4

With the gate set for (4), the S_{11} curve shows the impedance seen from where the LNA will be located. With the loss correction drawn in the biggest echos are at the −10db level. The average is ~ −14 to −15db. The peak reflections of −10db will produce a reflection gain loss of ~10% or ~½ db. These may account for the wiggles in the transmission gain curves.



S_{11} with no gates: one polarization

The pencil line shows the corrected −10db level.



S_{11} no gater: the other polarization

They are very similar. The pencil line shows the corrected -10 db level.

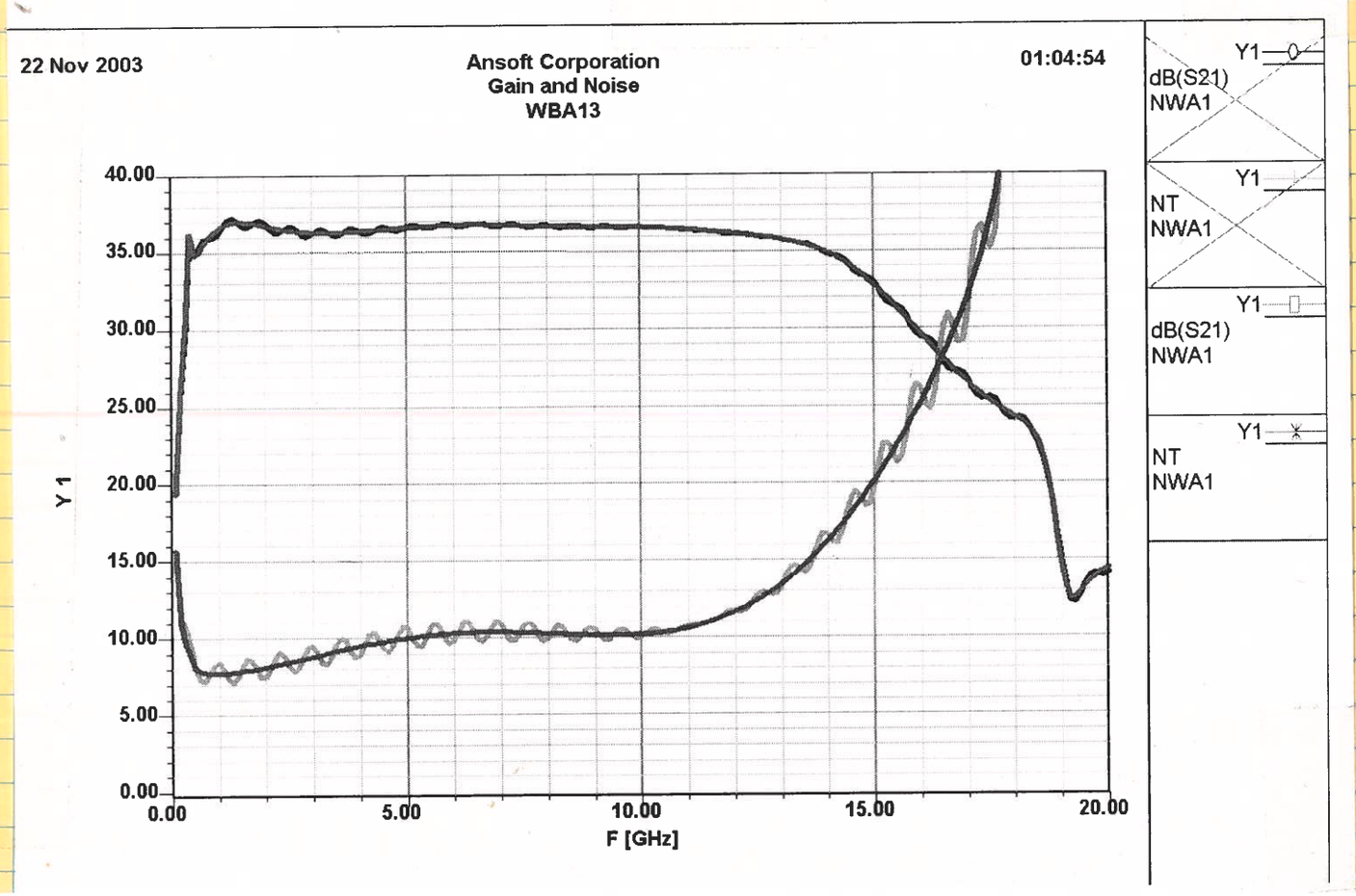
Revisit the expected T_{sys}. with Niklas new numbers, ~8K @ 1.4 GHz or 2.5 GHz and 10K at 10 GHz.

From I p.132 Gnd spill @ 10 GHz from diffraction is .6%
1.25 GHz 1.7%
+1% for geometrical optics spillover

atmospheric opacity (a) 10 GHz 3×10^{-2}, $\times 280 =$ ~~8.4~~ 7.0
(b) 1.4 GHz 2.3×10^{-2}, $\times 280 =$ ~~6.44~~ 4.0

		10 GHz	1.4 GHz
Feed	3%	9.0	9.0
Triplexer		3.0	3.0
~~interconnecting~~ transition line		~~7.0~~ 1.0	1.4
cold line		0.4	
Spillover (1.6%)		4.4	8.1
T_{bb}		2.5	2.6
atmosphere		~~8.4~~ 7.0	4.0
LNA		10	8.0
Total		~~48.8~~ 34.9	36.0 K

Niklas LNA noise and gain model for the WBA13 using the measured ≤ -10db for S_{11} assuming that the mismatch is at the tip of the balun



Recalculate the expected gain at 327 MHz. $\lambda = .92$ m. from [II 151]

The feed will be extended about 19 inches so that it will work at 327 MHz. It is also pushed forward as far as possible, which is 30". The feed is 1.2 m = 47" long. That leaves 17.0" + 19.0" from the focal point = 36" or 91.4 cm just one wavelength.

① From curves of I p 79 a one wavelength phase error gives a gain factor of 0.88 (for F/.65)

② The angle subtended by the secondary is now 57° instead of 84° (see I p 123). From p 77 I, this corresponds to F ≅ .95 and the spillover factor has gone from .835 to .670 total

③ At F = .95 the edge illumination has gone from .211 up to .465 The illumination efficiency is now up to .954

④ The diffraction spillover is now gone up.

$$1 - 2 l_d = 1 - 2 C_d \sqrt{\frac{\lambda}{d}} \sqrt{1 - \frac{d}{D}} A_0 = 1 - 2(.45)\sqrt{\frac{.92}{2.4}}\sqrt{1 - \frac{2.4}{6.1}}\,(.465)$$

$$= 0.81$$

$$\text{Total } \eta = (.88)(.67)(.954)(.81) = .455 = \boxed{.46}$$

Compare this with what we would get if it were completely in focus.

① focus efficiency: 1.0
② Illumination efficiency: .83
③ total spillover: .835
④ diffraction with $A_0 = .211$

$$1 - 2(.45)\sqrt{\frac{.92}{2.4}}\sqrt{1 - \frac{2.4}{6.1}}\,(.211) = 0.92$$

$$\text{Total } \boxed{\eta = 0.63} \qquad .63/.46 = 1.38 \cong \sqrt{2}$$

A difference in speed of a factor of 2; the same as the number of antennas.

Possible? f = 61.4", total feed length is 47" + 19" = 66"
It's not possible to go all the way

Radome Cover: The blue acrilic sail-bag cover material is 20 mils thick. It shows less than .02db attenuation at 8 GHz. It may be less. This was measured by placing it between two C band coax to waveguide transitions spaced about 6" apart. Alternately put between and moved out. It's a scraps from Hogin Sails. It's called "sunbrella". It's supposed to last ~10 years out in the sun. Ordinary sail cloth will only last about 1 year in the sun.

Birdair Flexible Plastic It's 30 mils thick.

It's teflon and fiberglass. 12/14/04

Transmission between two C-band coax to waveguide transitions showed very large loss effects when a sheet of the material was inserted. Signals went up at some frequencies and down at others by as much as 1 db.

Tuesday at 1 PM: Integration + Testing

Control Computer (running Linux) is MAINCNTL.~~ata.org~~ (lc) atarrey.org

user name ptaneer password bzrk2lzy

cd pta/jython ; Type jython ; <>

))) "from startup import *" <>

to get methods of the listed objects, type, e.g., ~~print. list~~ print ~~"~~.list()

For better formatting print track.checkSource( ) e.g. "CasA"

Optical Pointing Issues

With excellent pointing, single dish data can be supplied by individual interferometer dishes. Good pointing for this purpose means to 1/30 beamwidth. (ALMA/ASAC Results)

Encoder resolution ~ 20" arc seconds (Matt will check)

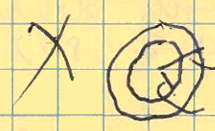
Antenna resolution, ~ 3° (180') at 1 GHz is (1/11.5)180 = 15.7' @ 11.5 GHz

1/30 of 15.7' = 0.5' = 30" ; So 20" accuracy encoders should work. 30" RMS

BIMA pointing is ~ 5" at night 5" RMS.

Get Rick's Memo

Cross Section of the needles



$r_o = \frac{11}{2}$ mils = .011/2

$r_i = \frac{9}{2}$ mils = $\frac{.009}{2}$

$A = \pi \{ (.00014)^2 - (.000144)^2 \}$

$= 6.16 \times 10^{-8} - 4.08 \times 10^{-8}$ m²

The inner area is 0.66 i.e. 2/3

$r_o = .0055'' = .00014$ m

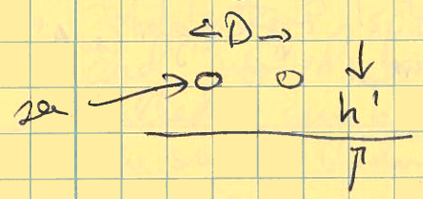
$r_i = .0045'' = .000114$ m

i.e. Making it solid triples the cross-section.

The input lines are 2a = .014" and D = .040"

The possibility of single turn loops in the transmission lines between the shield and the glass wall.

How close to the wall can the lines be?

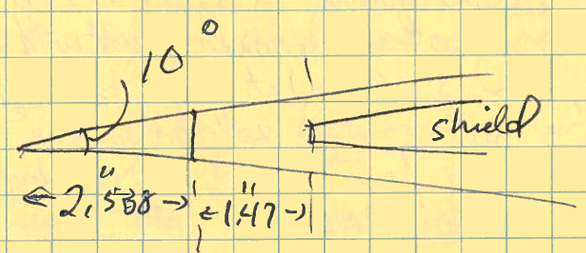


$D = .040$, $2a = .014$

For $h' = \infty$ $\quad Z_0 = 276 \log_{10} D/a = 209\,\Omega$

For $h' = .034$, $Z_0 = \{276 \log_{10}(D/a)\}\sqrt{1+(D/2h)^2} = \underline{190\,\Omega}$

So the lines should be no closer than .034"



Diameter at the shield is 4.01 @ 10° => .700"
For 20 mil wall thickness and 40 mil offset from the wall
$.700 - 2\times(.04+.02) = .580$ diameter for the loop.

One loop @ .580 diameter: $\pi\, .580 = 1.822$".
This roughly doubles the 1.47" length:
$1.47 + 1.822 = 3.29$" or 8.35 cm.

From II p 107 2" at room temperature adds $2.6\sqrt{f_{10}}$ K.
This piece will have the temperature gradient of 80→300 across it. The additional temperature will be ~ 1.2K.

From I p 119 the transmission line with gradient (and with 14 mils wire and 34 mil separation)

$$T_B(L) = 0.5\sqrt{\tfrac{f}{10}L^0}\,K \quad \text{with } L \text{ in cm.}$$

This assumes copper with $T(z) = 300 - z/L\,(220)$.

This is the "transition line" and will be 4.2 °K

(The number we've been using is 2.5 $\boxed{\Delta = 1.7°}$)

Could we use the Micro Lambda SMA inerds?

Expected differential contraction: $10^{-5}/°K = \frac{\Delta L}{L}$

For $\Delta T = 300 - 80 = 220K$

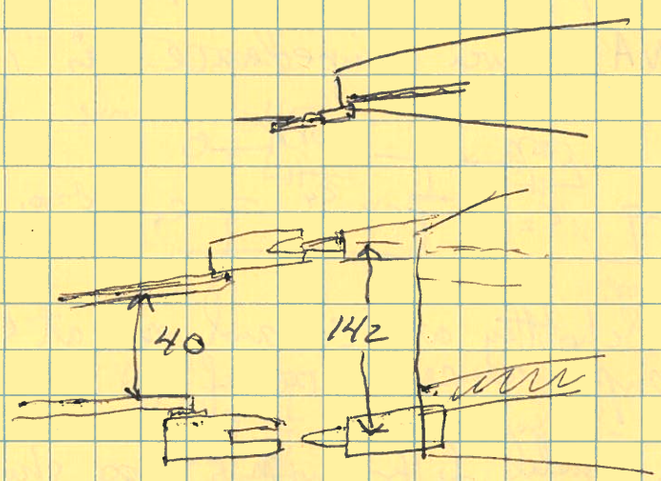
and $L \sim 10''$ get $10{,}000 \times 220 \times 10^{-5} \cong$ <u>20 mils</u>

The lines are $\overset{0}{\leftarrow} \underset{D}{40} \rightarrow \overset{0}{\uparrow} 14$

It looks like the overall pin diameters are 50 mils. Two of these parallell and touching are spaced by 50 mils, longer than 40.

For $2a = 50$, $D = \frac{50}{14} \times 40 = 142$ mils

This might work at the end of the balun.



It's cold here.

Tap uses this sort of connection and reports zero loss

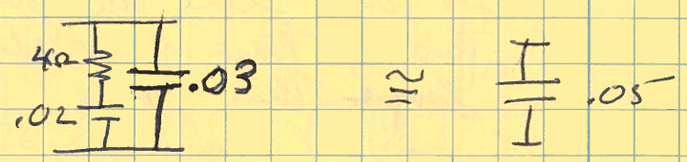
☆ { N.B. Tap is not working now and is getting bored. Maybe we could get him to work on stuff?

Possible Use of Back-to-Back Schottky diodes in front of the LNA for surge suppression.

GaAs Beam Lead from MA Com: model MA4E2037

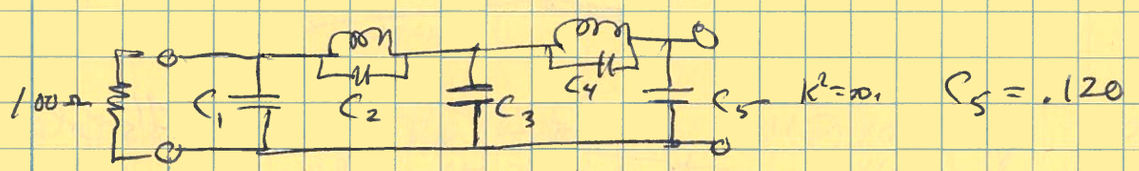
Junction Capacitance : .02 pf
Total Capacitance : .05 pf
Series Resistance (@ 10mA): 4 Ω

Model Diagram:



Two such diodes could be put back-to-back in a low pass filter to separate them out from the input capacitance, .120 pf.

The LNA source impedance is 100 Ω



Put one Schottky at C_1 and one at C_3. C_5 is the LNA input capacitance, .120 pf.

A single ended filter with $k^2 = \infty$ should work; $n=5$, $\rho = 4\%$

	C_1	C_2	L_2	C_3	C_4	L_4	C_5
Norm:	.3137	.1045	.8589	1.1567	.0863	1.233	1.206
Unnorm:	.032	~	.86nh	.115	~	1.23nh	.120

Select $C_r = 10^{-13}$ to get .120 pf for C_5

With $R = 100\,\Omega$, $\omega_r = 1/C_r R = [10^{-13} \times 100]^{-1} = 10^{11} = 2\pi f_r$; $f_r = 15.9\,GHz$

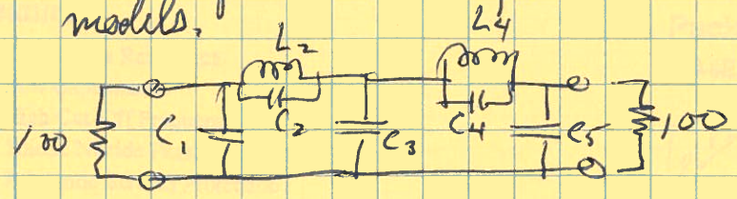
Then $L_r = R/\omega_r = 100/10^{11} = 10^{-9}$ henry $= 1$ nh.

The unnormalized values are above.

C_1 is a little small and C_3 a little large. Adding capacitance is easy. Neglect the series caps.

Zverev. p 209

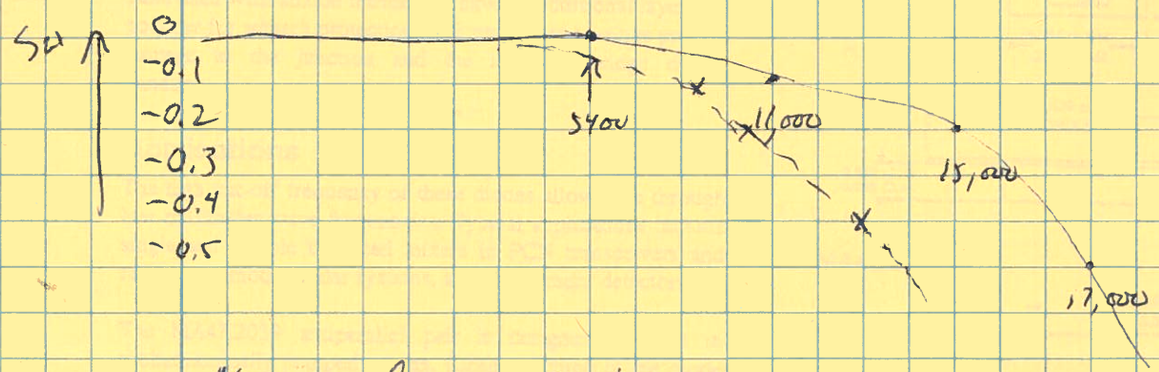
To see what the effect of the resistors is make a symmetric filter that is close to the unsymmetrical form. First make up the ideal filter and then put in the Schottky models.

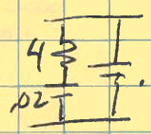


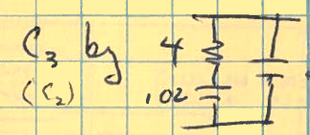
keep $\omega_r = 10^{11}$, $C_r = 10^{-13}$
$L_r = 10^{-9}$

	C_1	C_2	L_2	C_3	C_4	L_4	C_5
Norm:	.689	.034	1.243	1.46	.092	1.544	.635
Unnorm	.069	~	1.24 nh	.146	~	1.54	.064

This filter is pretty flat.



Now replace C_1 by 

and C_3 (C_2) by 

This had no noticeable effect on the response.
Using 8 Ω instead of 4 Ω had a barely perceptible effect.
Using 100 Ω showed the noted difference in the X's.

Conclusion: This way of adding the Schottky diodes should work well enough. If this is put on the LNA board at 80 K, the 0.1 db at ~ 11 GHz should be an extra 6.9 K, not much, ~~and that is only if R = 100 Ω.~~
For R = 4-8 K?, there is ~~no~~ little effect of loss.

Electrical Specifications @ T_A = +25°C

Symbol	Parameters and Test Conditions	Units	MA4E2037 Single			MA4E2039 Anti-parallel			MA4E2040 Series Tee		
			Min.	Typ.	Max.	Min.	Typ.	Max.	Min.	Typ.	Max.
Cj	Junction Capacitance at 0V at 1 MHz	pF	-	.020	-	-	.020[3]	-	-	.020[3]	
Ct	Total Capacitance at 0V at 1 MHz[1]	pF	.040	.050	.060	0.40[3]	.050[3]	.060[3]	.040[3]	.050[3]	.060[3]
ΔCt	Total Capacitance Difference	pF	-	-	-	-	-	-	-	.005	.010
Rs	Series Resistance at +10mA[2]	Ohms	-	4	7	-	4	7	-	4	7
Vf1	Forward Voltage at +1mA	Volts	.60	.70	.80	.60	.70	.80	.80	.70	.80
ΔVf	Forward Voltage Difference at 1mA	Volts	-	-	-	-	.005	.010	-	.005	.010
Vbr	Reverse Breakdown Voltage at -10uA	Volts	4.5	7	-	-	-	-	4.5	7	-

1. Total capacitance is equivalent to the sum of junction capacitance Cj and parasitic capacitance Cp.
2. Series resistance is determined by measuring the dynamic resistance and subtracting the junction resistance of 2.6 ohms at +10 mA.
3. Capacitance for the MA4E2039 and MA4E2040 is per Schottky diode.

Handling Procedures

The following precautions should be observed to avoid damaging these chips:

Cleanliness

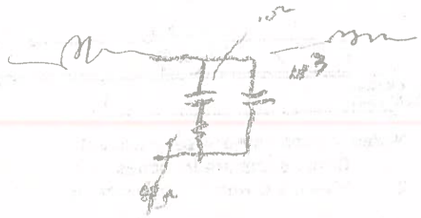
These devices should be handled in a clean environment. Do not attempt to clean die after installation.

Static Sensitivity

Schottky barrier diodes are ESD sensitive and can be damaged by static electricity. Proper ESD techniques should be used when handling these devices.

General Handling

These devices have a polymer layer which provides scratch protection for the junction area and the anode air bridge. Beam lead devices must, however, be handled with care since the leads may easily be distorted or broken by the normal pressures exerted when handled by tweezers. A vacuum pencil with a #27 tip is recommended for picking and placing. A sharpened wooden stick which has been dipped in isopropyl alcohol may also be used as a pick and place tool.

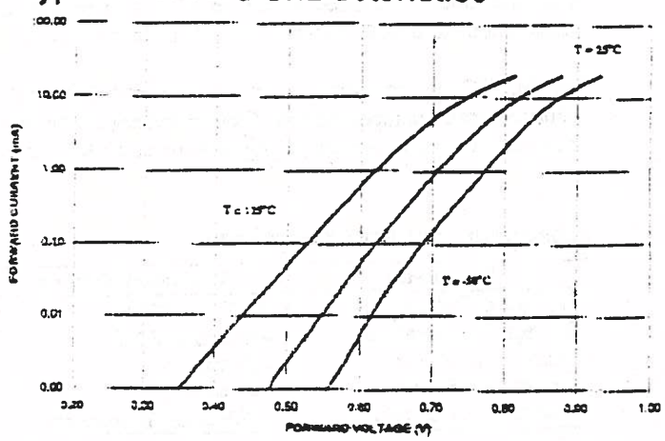


Mounting Techniques

These devices are designed to be inserted onto hard or soft substrates. Recommended methods of attachment include thermocompression bonding, parallel-gap welding, solder reflow, and conductive epoxy.

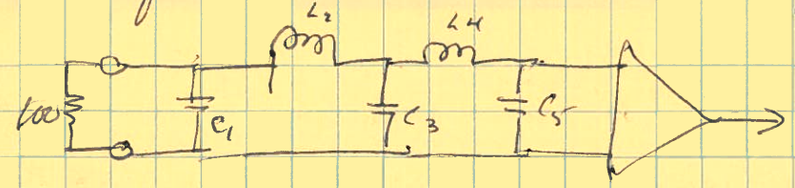
See "Application Note M541, "*Bonding and Handling Procedures for Chip Diode Devices*" for Detailed Instructions.

Typical Forward Characteristics



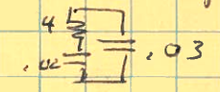
North America: Tel. (800) 366-2266 Fax (800) 618-8883 ■ Asia/Pacific: Tel. +85 2 2375 0618 Fax +85 2 2375 0350 ■ Europe: Tel. +44 (1344) 869 595 Fax +44 (1344) 300 020

Single ended filter enabled with ideal amplifier on output

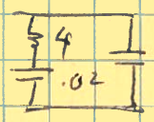


amp has $R_{in} = 10^6\,\Omega$ output $= 50\,\Omega$ DC open loop gain $= 1$

$C_5 = .120$ $C_1 = .05$ or



$C_3 = .115$ or



No observable difference whether the complete diode equivalent circuit is used or just the total capacitance. Also true for 8 Ω instead of 4 Ω for the "bulk resistance".
The full frequency response is shown on the next page. The scale was expanded for the substitution test so that variations in transmission of less than 0.1 db could easily be seen.

A_e/T_{sys} measurement

Allen (1963) gives Cas A as 3100 Jy @ 1 GHz and $\alpha = -.8$

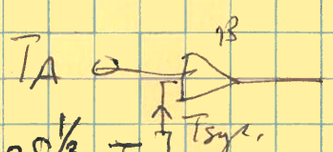
Over the intervening 40 years @ ~1% decrease/year, we expect about 1800 Jy now @ 1 GHz $= S_0'$

$$S_0, \quad \frac{S_0'}{2k} = \frac{1.86\times10^{-23}}{2(1.38\times10^{-23})} = 0.67; \quad .58 \text{ for } 1600\,Jy \text{ @ } 1.4$$

$$KT_A = A_e S_0^{1/2}$$

$$T_A = \left(\frac{A_e}{2k}\right) 2S_0^{1/2}$$

T_{sys} measurement



$$O_1 = B(T_A + T_{sys}) = B\left(\left[\frac{A_e}{2k}\right]2S_0^{1/2} + T_{sys}\right)$$

$$O_2 = B\,T_{sys}$$

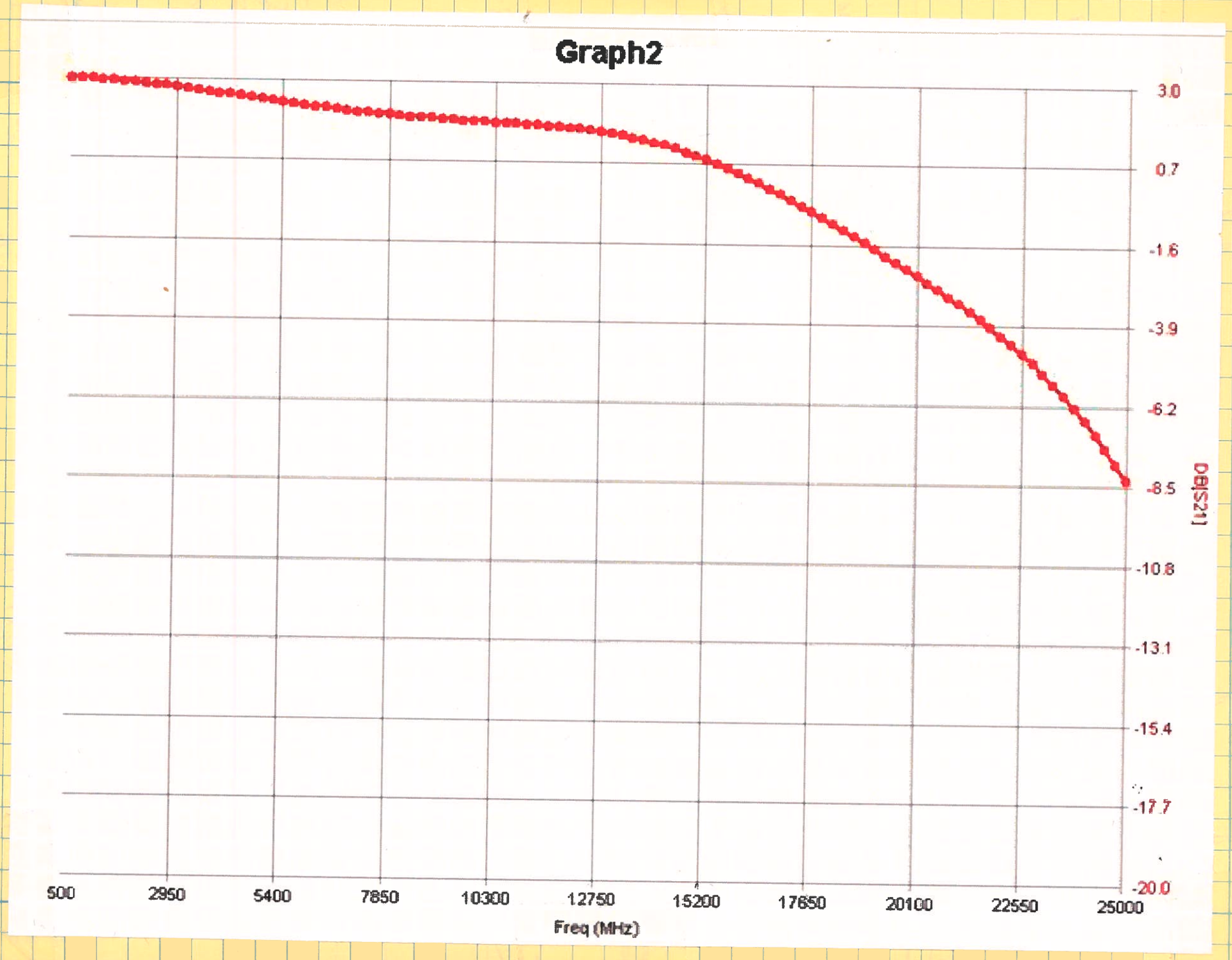
$$\frac{O_1}{O_2} = 1 + \left(\frac{A_e}{T_{sys}}\right)\frac{S_0'}{2k} \qquad S_0' = 2S_0^{1/2}$$

$$1 + \left(\frac{A_e}{T_{sys}}\right) .58$$

A_e/T_{sys}	Y (db)
14.6/50 = .29	.067
14.6/75 = .19	.46
14.6/100 = .146	

for (.57), $A_e = 16.6$
(.50), $A_e = 14.6$

T_{sys}	A_e/T_{sys}	Y
300	.055	.14
150	.111	.27
100	.166	.4
50	.332	.76


Graph2
3.0
0.7
-1.6
-3.9
-6.2
-8.5
-10.8
-13.1
-15.4
-17.7
-20.0
DB[S21]
500
2950
5400
7850
10300
12750
15200
17650
20100
22550
25000
Freq (MHz)

RFI levels from aircraft transponders

KT-76A (unit in N3886P) $f = 1090\,MHz$ FAA standard unit

Peak pulse power: 200 watts (average $\lesssim 1$ watt)

[Dale @ Bay Avionics 250-275 watts peak with new tube.
Units in commercial airliners provide 500-1000 watts peak]

Dale doesn't think military uses anything stronger

In the ATA main beam

① At 1km for the KT-76A $\frac{1}{4\pi(10^3)^2} \simeq 10^{-7}\,m^{-2}$; $\times 200{,}000$ mw $\Rightarrow .02$ mwatts/m^2

$A_e \cong 20\,m^2$ $P_{in} = 0.4$ mwatts peak pulse

② At 300m $P_{in} = 4$ mwatts, peak

③ At 100m $P_{in} = 40$ mwatts peak ok

④ A military jet making a low pass at 300m would give us 20 mwatts peak. At 100m it would be 200 watts peak

Amateur radio transmitter 1kw at 30MHz (λ 10 m)

At 1km distance $S = \frac{10^6 \text{ mwatts}}{4\pi(10^3)} = 0.1$ mwatts/m^2

Into the sidelobes where $g \approx 1$ $A_e = \frac{g\lambda^2}{4\pi} \simeq \frac{\lambda^2}{4\pi} = \frac{(10)^2}{4\pi} \sim 10$

$P \rightarrow 1$ mwatt.

At $\lambda = 10\,m$ (30MHz) $g << 1$, so this is a worse case.

There's also the danger of somebody bringing a transmitter on site, either unwittingly or on purpose (in the case of Bob Krekorian). The above numbers assume that the transmitter is in the main beam.

The low frequencies near 500 MHz may be the most dangerous. In a side lobe at $g \sim 1$ $A_e = \frac{g\lambda^2}{4\pi} \simeq \frac{.6^2}{4\pi} \simeq \frac{0.3}{10}\,m^2$
$A_e \sim .03\,m^2$

For P_{in} to be 100 mwatts with $A_e = .03\,m^2$ and $P_t = 1$kw.

$\frac{A_e P_t}{4\pi D^2} = .1$ watt, $D^2 = \frac{(.03)(1000)}{4\pi \cdot .1} = 2.4$. $D = 5\,m$

In the main beam at 400 MHz (e.g. the GPS transmitter) with $A_e \cong 20\,m^2$

$$\frac{A_e P_t}{4\pi D^2} = 0.1 \text{ watt} \qquad D^2 = \frac{A_e P_t}{4\pi\, 0.1} = \frac{20\ 20}{1.2}, \quad \underline{\underline{D \cong 20\,m}}$$

If we do the radar tests, we will need to know the dishes down to the horizon, and there could be an accident with the GPS transmitter.

Realizing the input filter

The chip dimensions are ~10x10 mils, so 10 mil diameter wire is appropriate. The formula from Montgomery, Dicke and Purcell p. 168 is for the inductance of a short, isolated wire:

$$L = \frac{\mu}{2\pi} \ln\left(\frac{2l}{er}\right) \times l \qquad \text{Henrys}$$

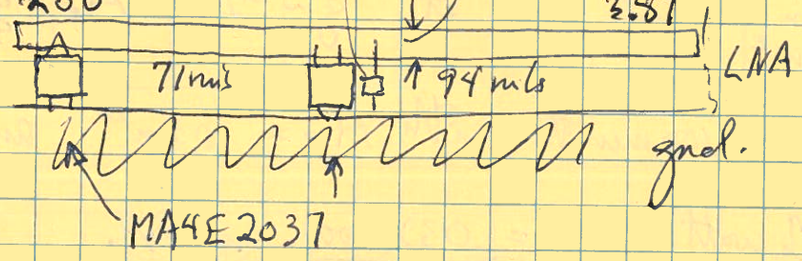
l is the length, r is its radius, $e = 2.7$, μ is the permeability $\mu = 4\pi \times 10^{-7}$ henrys/m. Lengths are in meters in x .0254 → meters

$$L = 2l \times 10^{-7} \ln\left(\frac{2l}{er}\right) \qquad \text{use } r = 5 \text{ mil wire}$$

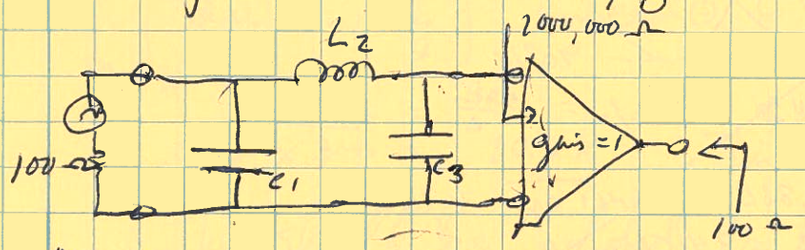
l (mils)	l (m)	L (nH)
50	.00127	.51
60		.67
70		.83
80		1.00
90		1.18
95		1.28
100		1.37
110		1.56
200		3.89

For r = 1.5 mils: .798 nH; for r = 2 mils: 1.8 nH

2.3 nH

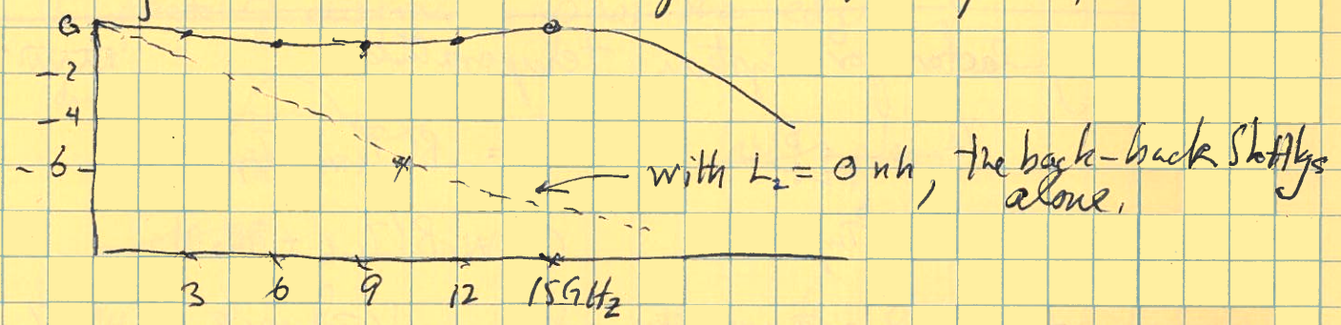


☆ A simpler solution is to keep the Schottky's together, using MA4E0239 ; get ~ .100 pf for the combination.



C_1 = 100 ff b-b Schottkys
C_3 = 120 ff

☆ Tuning L_2 to 1.438 nH. gives the following response.



To get about 1.438 nh (1.50) use 110 mils of 5 mil radius bond wire. Expect no ohmic losses.

Case A ~~pol X~~ with 1515/230 filter
on-off .1 ± .02 db 2 times repeat
~~poly~~ .06 ± .02 2 times repeat.

cal signal was turned on

elevations: { down - went to -.04 db - one degree
up went to -.025 db - one degree

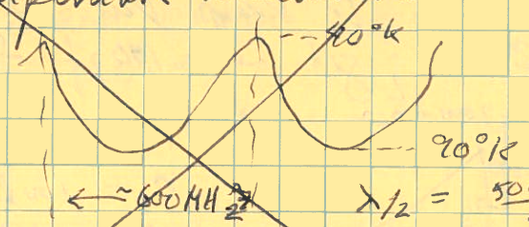
azimuth left - went to -.02 1.4 degrees
right - went to -.02 1.4 degree

Y { Total power from LNA ~~through~~ through the filter is -52.5 dbm two lengths of blue cable included

X { -56.3 dbm as above otherwise

Data miscalculation

~~Noise temperature measurements~~

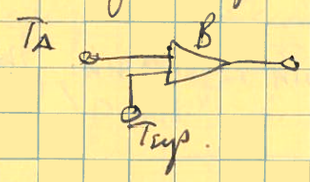


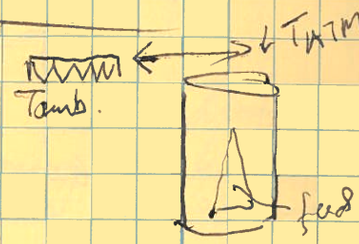
$\lambda/2 = \frac{50\,cm}{2} = \underline{25\,cm}$

~~balun length is 4.886" = 12.41 cm (glass)~~
~~total wire length is 3.707" = 9.415 cm~~
~~distance from balun to glass window is 2.015" = 5.118 cm~~

---

y-factor for system temperature





$$O_1 = B(T_{ATM} + T_{sys})$$

$$O_2 = B(T_{amb} + T_{sys})$$

$$y = \frac{O_2}{O_1} = \frac{T_{amb} + T_{sys}}{T_{ATM} + T_{sy}} \qquad y(T_{ATM} + T_{sys}) = T_{amb} + T_{sys}$$

$$y\,T_{ATM} + y\,T_{sys} = T_{amb} + T_{sys}$$

$$T_{sys}(y-1) = T_{amb} - y\,T_{ATM}$$

$$T_{sys} = \frac{T_{amb} - y\,T_{ATM}}{y-1} = \frac{T_{amb}}{y-1} - \left(\frac{y}{y-1}\right) T_{ATM}$$

Contributions to T_{ATM}

so far we have been using 6K for the O_2 emission plus the 2.7K background. This is ok for the frequencies around 1-2 GHz

1. The O_2 emission: 3K at 1-2 GHz

For $T_B = 3K$ at zenith $\quad 3 = 280(1-e^{-\tau}) \Rightarrow \tau = .0107$

So $T(\theta) = 280(1 - e^{-.0107/\cos\theta})$

$$\langle T(\theta) \rangle = \frac{1}{2}\int_0^{86^\circ = 1.501} 280(1 - e^{-.0107/\cos\theta}) \sin\theta \, d\theta$$

$$= 280(.0139) = 3.9K, + 3K(BB) = \underline{6.9K}$$

2. <u>The horizon on the average is at $E = 4°$</u>

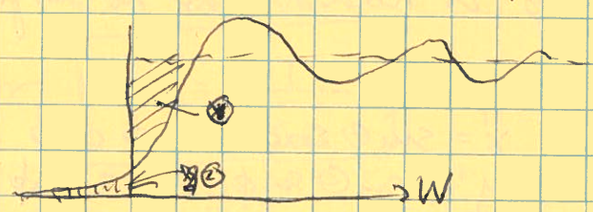
From $0 < E < 4°$, the brightness of the horizon is 300 K.

$$\int_0^{2\pi}\int_{90°-4\pi}^{86°} \frac{d\Omega}{4\pi} = \frac{1}{2}\int_{90°}^{86} \sin\theta \, d\theta = \frac{1}{2}\cos\theta \Big|_{90°}^{86°} = \frac{1}{2}\cos 86° = 0.035$$

$.035 \times 300 = \underline{10.5\,K}$ This is geometric optics leakage into the cylinder

This should be added to the $\underline{6.9K}$ from the gas: $\underline{\underline{17.4}}$

3. Effects of diffraction past the edge.



half-plane

We already have the geometrical optics strip above contributing to the input noise signal. The diffraction correction is given by the above diagram. The part from region ① is removed from the geometric optics contribution and put into region ②. These two regions have about the same area, so the net effect is <u>approximately no change</u>.

Altogether $T_{ATMO} = \underline{17.4\,K}$ at $\sim 1\text{-}3\,GHz$

For 6/3/04 T_{amb} at Hat Creek was ~ 80F which is <u>300 K</u>
It's plausible to assume the the coverage of the absorbing load was ~ 97%. This makes $T_{amb} \sim 290K$

Then $T_{sys} = \dfrac{290 - Y\,17.4}{Y - 1}$

Y	T_{sys}.
8 db	34 K
7 db	50.8
6.5	60.5
6.0	73.4

Holographic sampling of the surface errors

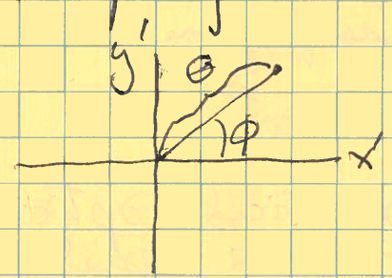
From his map of the antenna at 2.3 GHz, Gerry found an RMS error of 8° in the aperture.
$\sigma_\phi = 8° \Rightarrow 0.14$ radians.

$$\sigma_\phi = \frac{4\pi}{\lambda}\sigma_s \qquad \text{so} \qquad \sigma_s = \frac{\lambda}{4\pi}\sigma_\phi$$

$2.3\,GHz \rightarrow 13$ cm.

$$\sigma_s = \frac{13}{4\pi}(0.14) = .145 \Rightarrow \underline{\underline{1.5\text{ mm}}}$$

How extended a map is needed for the surface sampling?



$$x' = \sin\theta\cos\phi \cong \theta\cos\phi \qquad \text{(radians)}$$
$$y' = \sin\theta\sin\phi \cong \theta\sin\phi$$

$$E_y(R,\theta,\phi) = \frac{j}{\lambda R}e^{-jkR}f(\theta,\phi) \Rightarrow \frac{j}{\lambda R}e^{-jk}f(x',y')$$

$$f(x',y') = \iint_{-\infty}^{\infty} E_{oy}(x,y)\,e^{-j2\pi\left(\frac{xx'}{\lambda}+\frac{yy'}{\lambda}\right)}\,dx\,dy \quad ; \quad u=\frac{x}{\lambda},\ v=\frac{y}{\lambda}$$

$$f(x',y') \propto \iint_{-\infty}^{\infty} E_{oy}(u,v)\,e^{-j2\pi(ux'+vy')}\,du\,dv$$

and
$$E_{oy}(u,v) \propto \iint_{-\infty}^{\infty} f(x',y')\,e^{j2\pi(ux'+vy')}\,dx'\,dy'$$

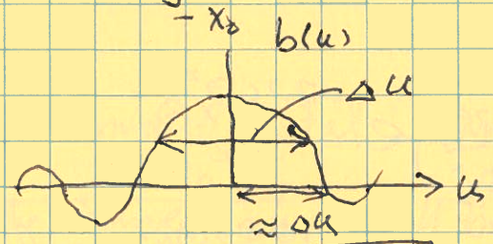
Suppose we get the pattern over the range $-x_0 < x' < x_0$, $-y_0 < y' < y_0$

This defines the sample range function B, which is zero outside that interval.

Taking just the one dimensional span $B(x')$

$b = \mathcal{F}\{B\}$ gives the resolution in u, Δu.

$$b(u) = \int_{-x_0}^{x_0} e^{j2\pi u x'}\,dx' = \frac{\sin 2\pi u x_0}{\pi u} = 2x_0 \cdot \frac{\sin 2\pi u x_0}{2\pi u x_0}$$



$$2\pi \Delta u x_0 \cong \pi, \quad \Delta u = \frac{1}{2x_0}$$

or $$\boxed{2x_0 = \frac{1}{\Delta u}}$$

So the spatial resolution in wavelengths equals the <u>total</u> span in angle (radians)

For a map at 3cm. with a resolution of 30cm.

$$\Delta u = \frac{\Delta x}{\lambda} = \frac{30}{3} = 10.$$

and $2x_0 = \frac{1}{10}$ radian $\sim \underline{6^\circ}$; the beamwidth is $\sim \frac{\lambda}{D}$

or $\frac{3}{600} = \frac{1}{200} = \frac{1}{20}\cdot\frac{1}{10} = \lambda/D = 0.3^\circ$

30cm resolution is 1/20 aperture diameters, so scan 20. beamwidths total.

Observing the moon: correction for resolution

Assume a perfectly symmetric Gaussian Beam.

$$e^{-.69\left(\frac{2\theta}{\Theta_{1/2}}\right)^2} \qquad \Theta_{1/2} \text{ is the } \underline{\text{FWHM}} \text{ beam.} \qquad e^{-.69} = 1/2$$

$$= e^{-2.76\theta^2/\Theta_{1/2}^2}$$

Normalization: $$\frac{G(\theta)}{4\pi} = \frac{.879}{\Theta_{1/2}^2}\, e^{-2.76\theta^2/\Theta_{1/2}^2}$$

Check the normalization: $$\int_0^{2\pi}\int_0^{\infty} \frac{.879}{\Theta_{1/2}^2}\, e^{-2.76\theta^2/\Theta_{1/2}^2}\, \theta\, d\theta\, d\phi \overset{?}{=} 1$$

let $u^2 = \dfrac{2.76\theta^2}{\Theta_{1/2}^2}$, $u = \dfrac{1.66\theta}{\Theta_{1/2}}$, $du = \dfrac{1.66\, d\theta}{\Theta_{1/2}}$; write $\underline{\alpha \equiv \Theta_{1/2}}$

$$2\pi \int_0^{\infty} \frac{.879}{\alpha^2}\, e^{-u^2} \left(\frac{\alpha}{1.66}\right) u \left(\frac{\alpha}{1.66}\right) du = \frac{2\pi}{2.76}\int_0^{\infty} e^{-u^2} u\, du\; .879$$

$$= \frac{2\pi}{2.76}\left(-\frac{1}{2}\right) e^{-u^2}\Big|_0^{\infty} .879 = 1.0005$$

Put the moon at the center of the beam with it's diameter Θ_m and it's brightness temperature T_M.

Then $$T_{AM} = \int_0^{2\pi}\int_0^{\Theta_m/2} T_M \frac{G(\theta)}{4\pi}\, \theta\, d\theta\, d\phi \quad ; \text{ i.e. } T_A = \int T_B(\theta,\phi)\frac{G(\theta\phi)}{4\pi}\, d\Omega$$

$$= T_M\, 2\pi \int_0^{\Theta_m/2} \frac{.879}{\alpha^2}\, e^{-2.76\theta^2/\alpha^2}\, \theta\, d\theta \qquad \text{again write } u^2 = \frac{2.76\theta^2}{\alpha^2},\ u = \frac{1.66\theta}{\alpha}$$

$$T_{AM} = T_M 2\pi .879 \frac{.879}{\alpha^2}\int_0^{\frac{1.66(\Theta_m/2)}{\alpha}} \left(\frac{\alpha}{1.66}\right) u \left(\frac{\alpha}{1.66}\right) du = T_m \frac{2\pi(.879)}{2.76}\int_0^{.83(\Theta_m/\alpha)} e^{-u^2} u\, du$$

$$= T_m \frac{2\pi(.879)}{2.76}\left(-\frac{1}{2}\right)\left[e^{-\left(.83\frac{\Theta_m}{\alpha}\right)^2} - 1\right] = T_m\left[1 - e^{-.69\left(\frac{\Theta_m}{\Theta_{1/2}}\right)^2}\right]$$

$$= \frac{T_m}{2} \text{ when } \Theta_m = \Theta_{1/2}, \quad = .94\, T_m \text{ when } \Theta_m = 2\,\Theta_{1/2}$$

For Monday June 14 (JPL)

The Moon has $\alpha = 02^h 32^m$ $\delta = +14°42'$
The Sun has $\alpha = 05^h 30^m$ $\delta = +23°15'$

So the moon is 3 hours ahead (west) of the sun.

In terms of total flux, for the moon $F = \frac{2kT_m}{\lambda^2}\Omega_m$
The observed flux will be $\frac{2k}{\lambda^2}\int BG\,d\Omega$

The power received (in one polarization) is always kT_A, and in the case of the moon $T_A = T_m\left(1 - e^{-(\theta_m/\theta_{1/2})^2}\right)$

This is the simplest way to think about it. The input power is kT_A and it can be compared to the system temperature.

→ NB. We are assuming that the beam efficiency is 1 here. To be more correct we should multiply by it. e.g. ~0.9 (No, it's smaller) 18% of the feed pattern spills over

Troitsky's summary of many measurements is that $\langle T_m \rangle = 230$ K for radio measurements with, for $\lambda = 3.2$ cm:

$$230° + 14.5° \cos(\Omega t - 50°)$$

One important factor that must be added to the A_e T_{sys} evaluation is the total RCVR input loss. It has been used to estimate the T_{sys} but not included in the effective A_e evaluation. We must include a factor η_ℓ to multiply A_e along with η_B.

Review the input losses

1. The diffractive losses (Kildal) are already included in η_B

2. Feed attenuation - 3% is the estimate of p. 120 (I).

3. tip circuit board 1% $\sqrt{f/10}$ is the estimate of p 120 (I)

4. ambient line 1.7" @ 300 K 1.7" = 4.32 cm.

from p 119 (I) $\alpha_p = .063\, R_s(273) = .063\,(.0262\sqrt{f/10}) = .0017\sqrt{f/10}$ (?)

Repeat the calculation: 14 mils. = d = .036 cm.

→ 40 ←

From R+W $R_s(273)$ for Silver $= 2.52\times10^{-7}\sqrt{f}$

$$\alpha_p = \frac{2R_s}{\pi d Z_0}\frac{(s/d)}{\sqrt{(s/d)^2-1}} = \frac{2(2.52\times10^{-7}\sqrt{f})}{\pi(.036)\,200}\frac{2.86}{\sqrt{(2.86)^2-1}}$$

$$= 2.38\times10^{-8}\sqrt{f} = .0024\sqrt{f/10}\ \text{cm}^{-1}$$ f in GHz finally.

$\alpha_p L = .0024\sqrt{f/10}\ 4.32 = .010\sqrt{f/10}$ loss

emission @ 300 K is 3.1 K

5. The line with the gradient 2.0" = 5.08 cm.

From p 119 (I) $T_B(L) = .50\sqrt{f/10}\,L\,°K$ for the emission

or ~~.94 K for the emission~~.

2.54 K emission.

The corresponding attenuation at a mean temperature of 190 K is .013 $\sqrt{f/10}$ 1.3 %

6. The balun at 80 K 4.9" = 12.4 cm

The calculation of p 119 (for a 25 cm balun) is 2.2 $\sqrt{f/10}$ K°

for an absorption of .027 $\sqrt{f/10}$ ✓

.63 mm = δ

7. Ruze losses are $e^{-\left(\frac{.798}{\lambda(mm)}\right)^2}$ p. 124 (I)

Adding them all up gives: $\left\{3 + \sqrt{f/10}\,(1.0 + 1.0 + 1.3 + \overset{\text{balun}}{2.7})\right\}\% = 3 + 6.0\sqrt{f/10}\ \%$

$= (3.0 + 3.3)\sqrt{f/10}\ \%$ with no balun

Then multiply this by the Ruze Loss:

$$\eta_\ell = \left\{1 - (.03 + .06)\sqrt{f/10}\right\} e^{-\left(\frac{.798}{\lambda(\text{cm})}\right)^2} \quad ; \quad e^{-\left(\frac{1.08}{\lambda(\text{cm})}\right)^2} \text{ for } \sigma = .86\text{ mm}$$

$= \{1 - (.030 + .033)\sqrt{f/10}\}\, e^{-(\ )}$ with no balun

$\eta_\ell(10\text{ GHz}) = (.91)(.93) = 0.85$

$\eta_\ell(7\text{ GHz}) = (.92)(.97) = 0.89$

$\eta_\ell(3\text{ GHz}) = (.94)(.99) = 0.93$

$\eta_\ell(1.5\text{ GHz}) = (.95)(1.0) = .95$

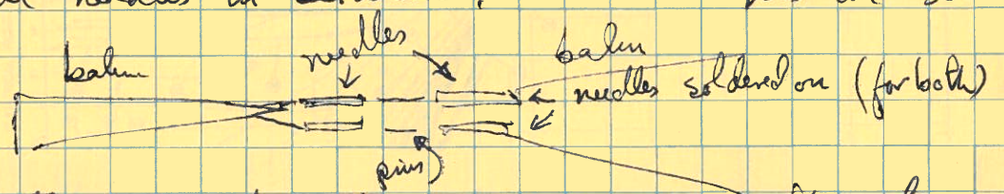
Note: The balun losses should be removed (9/31/08)

see below

Connecting $A_e(0)$ and $G(0)$ $\quad G(0) = \dfrac{4\pi A_e(0)}{\lambda^2}$

6/26/04 Some tests on the needles at Matt's place

2 black colored [teflon? cuflon?] baluns back to back with three needles in between. These are 50Ω baluns.



Pins ~~Needles~~ were taken from a glass feed through. Each pin is just slide into the needles at both ends.

1. So there are 2 sliding joints in series here. We see a very large attenuation: ~ 8db @ 7 GHz. This is recorded on disk. SCRN08.JPG
2. Soldering the pins to the needles made no noticeable difference?! (except after swipe with methanol that reduced transmission!?)
3. Putting in one early double balun reproduced the original low loss this is recorded on disk. SCRN09.JPG

The setup of the black baluns and/or needles and/or sma launchers must be at fault.

(4) Removed the needles and connected the 2 baluns with short pieces of copper wire. Got about the same overall transmission as before this change. SCRN00.IPG
The scale for the comparison is 10db/cm. There may be probably differences at the few tenths of a db level.

Low temperature conductivity of Gold.
R.A. Matula, J. Phys. Chem. Ref. Data, 8, 1147, 1979

ELECTRICAL RESISTIVITY OF COPPER, GOLD, PALLADIUM, AND SILVER 1209

TABLE 5. RECOMMENDED VALUES FOR THE ELECTRICAL RESISTIVITY OF GOLD

[Temperature, T, K; Total Resistivity, ρ, 10^{-8} Ω m; Intrinsic Resistivity, ρ_i, 10^{-8} Ω m]

Solid

T	ρ_i [a, b]	ρ [a, c]	T	ρ_i [a, b]	ρ [a, c]
1		0.0220	175	1.240	1.262
4		0.0220	200	1.440	1.462
7		0.0221	225	1.640	1.662
10		0.0226	250	1.842	1.864
15	0.00376*	0.0258	273.15	2.029	2.051
20	0.0126*	0.0346*	293	2.192	2.214
25	0.0282*	0.0502*	300	2.249	2.271
30	0.0505*	0.0725*	350	2.663	2.685
35	0.0798*	0.1018*	400	3.085	3.107
40	0.119*	0.141*	500	3.952	3.974
45	0.159	0.181	600	4.853	4.875
50	0.199	0.221	700	5.794	5.816
55	0.248	0.270	800	6.786	6.808
60	0.286	0.308	900	7.840	7.862
70	0.373	0.395	1000	8.964	8.986
80	0.459	0.481	1100	10.169	10.191
90	0.544	0.566	1200	11.464	11.486
100	0.628	0.650	1300	12.832	12.854
125	0.835	0.857	1337.58	13.366	13.388
150	1.039	1.061			

At 273K $\rho = 2.051 \times 10^{-8}$ Ω m $= 2.05 \times 10^{-6}$ Ω cm
80K $\rho = .48 \times 10^{-8}$ Ω m
60K $\rho = .308 \times 10^{-8}$ Ω m

Compare with copper (I. p. 118) At 273 1.54×10^{-6} ohm cm.
80 $.18 \times 10^{-6}$ ohm-cm.
66.9 $.13 \times 10^{-6}$

At 80K $\rho_{Cu}/\rho_{Gold} = \frac{.18}{.48} = 0.375$, $\sqrt{} = .61$

6/29/04

Integration and Test meeting

1. The small pointing camera is returned and re-installed. ABIMA camera will be used eventually. Matt has some ideas about the mounting.

   a pointing model (~10 terms) is in place.

2. Downconverter not tested (from SETI)

3. software correlator is available.

4. Monitor for the compressor going in. inc. temps and accelerations.

Fitting the gain pattern to a Gaussian at the FWHM angle:

From an aperture field of $E(r) \propto .2M + .789[1-(r/a)^2]^{1.9}$, the pattern is ~~(for $\lambda = 0.2$ m)~~ (for $a = 3$).

$$E(\theta) \propto \frac{J_1\left[\frac{6\pi}{\lambda}\sin\theta\right]}{\frac{6\pi}{\lambda}\sin\theta} + 25.4\,\frac{J_{2.9}\left[\frac{6\pi}{\lambda}\sin\theta\right]}{\left(\frac{6\pi}{\lambda}\sin\theta\right)^{2.9}}$$

edge field is −13.5 dB

$G(\theta) \propto |E(\theta)|^2$ has FWHM of $2.33^\circ\left(\lambda/0.2\,\text{m}\right)$

① The Gaussian that fits the FWHM has 7% more integrated solid angle than the true $G(\theta)$, the Gaussian

② The spillover from the feed is 18%

③ The edge diffraction loss is $(1-.095\sqrt{\lambda_{cm}})$

These factors define the beam efficiency $\eta_B = (1.07)(.82)(1-.095\sqrt{\lambda_{cm}})$ (feed spillover, diffraction)

~~$A_e(0) = 17.9\,\text{m}^2\,\eta_B\,\eta_\ell = 0.61$ at $\lambda = 0.2$ m, (?)~~

$A_e(0) = \eta_B \eta_\ell\, 21.6\,\text{m}^2$ ~~no! the 1.07 factor only goes into the beam eff.~~

~~Effective beam efficiency: $\eta_\ell \eta_B$~~ ✓ $\boxed{\text{yes, it's ok}}$ ✓

A lot of common factors, but not all. η_ℓ is common.

There's also beam spill to ground due to diffraction which adds to the system temperature

The η_B is correct for the total beam efficiency which goes with the Gaussian of FWHM of 2.33° at $\lambda = .2$ m.

For the aperture efficiency there is the η_{ill} factor and no factor of 1.07.

Take a mean $\eta_{ill} = .82$ (see II p 40)

Then for the in-focus case $\eta_{app} = \eta_{ill}\,(.82)(1-.095\sqrt{\lambda})\,\eta_e$ (η_{spill})

$$\eta_{app} = (.82)(.82)(1-.095\sqrt{\lambda(m)})(\eta_e)$$
$$\eta_{app} = .67(1-.095\sqrt{\lambda(m)})[1-.03-.06\sqrt{f/10}]\,e^{-\left(\frac{.798}{\lambda(m)}\right)^2}$$

Total: $\eta_B' = \eta_B\,\eta_e = (1.07)(.82)(1-.095\sqrt{\lambda(m)})[1-.03-.06\sqrt{f/10}]\,e^{-\left(\frac{.798}{\lambda(m)}\right)^2}$

Convert λ to frequency, f in GHz.

$\lambda(cm) = 3\,cm/f/10 = 30\,cm/f(GHz)$

$\lambda(m) = .3\,m/f(GHz)$ $\qquad \sqrt{\lambda(m)} = .55/\sqrt{f(GHz)}$;

$\frac{.798}{\lambda(m)} = \frac{.798}{30cm} f(GHz) = .0266 f$

$$\longrightarrow \eta_{app} = .67(1-.052/\sqrt{f_{GHz}})[1-.03-.06\sqrt{f/10}]\,e^{-(.0266 f_{GHz})^2}$$

$$\longrightarrow \eta_B' = .88(1-.052/\sqrt{f_{GHz}})([.97-.06\sqrt{f/10}]\,e^{-(.0266 f_{GHz})^2}$$

see below for final

The expected T_{sys}. see p 42 above

1. Half the diffractive losses go to the ground: $\frac{1}{2}\,.095\sqrt{\lambda(m)} = (.048)(.55/\sqrt{f_{GHz}}) = .026/\sqrt{f_{GHz}}$, $\times 300 = 7.8/\sqrt{f_{GHz}}$ °K

2. Ohmic losses @ 300K
   feed 3% ← all at 10 GHz
   tip board 1%
   comb. line 1%
   5% : $(.05)(300)\sqrt{f/10} = 15\,\sqrt{f/10}$ °K

3. The line with the gradient $2.5\sqrt{f/10}$ K°

4. The cold balun $2.2\sqrt{f/10}$ °K $\cong 2.5\sqrt{f/10}$ see below

17.5 with no balun

cooling the tip could reduce this by a factor of $\frac{17.5}{300} = \frac{1}{4}$

The sum of 2, 3, and 4 is $20.0\sqrt{f/10}$ °K

5. The atmosphere is 6°K, approximately constant for 1–10 GHz (except for zenith effect)

6. The LNA is $\lesssim 10$k° for 0.5 – 12 GHz.

The total is: $\underbrace{\frac{7.8}{\sqrt{f}} + \underset{(17.5)}{20}\sqrt{f/10}}_{} + 16 = T_{sys} = \frac{7.8}{\sqrt{f}} + 6.3\sqrt{f} + 16$

Diffractive Spillover

$= 37.5$K at 10 GHz

$= 30$ K at 1 GHz $\longrightarrow$ 32K with geom spillover + emiss @ 45°

There's also the extra 1% ground pick-up at zenith $\sim 3(1-\frac{z}{90°})$

One further change: the absorption in the cold balun.

It's properties: 0.1μ of Ti on the crystal quartz
2.5μ of Cu next : .64 mil
0.1 Ni
10. Gold

balun not used

The top two layers should have little effect.
The Ti should have little effect

We use the loss of the input microstrip dimensions.

$\epsilon = 4$ Surface roughness = 0.1μm, very small (~~or .0003~~)

[Copper at 10 GHz and T = 290k has $\delta = .66\mu$] (skin depth)

Using the TLine program in Cersys we find

For $Z_0 = 100\,\Omega$ $f = 10$ GHz $\epsilon = 4$
$SR = .1\mu m = 4\times10^{-6}$ mils
TanD $= 3\times10^{-4}$ h = 20 mils w = 9 mils

Loss = .11 db/inch $\times$ 4.9 = 0.54 db total at 290k
At 50K : $.54/\sqrt{10} = \underline{0.17\text{db}}$ or $\underline{4.0\%}$

More than the 2.7% used above.
This raises the frequency dependent input loss from 6.0 to 7.3%

11/31/08

// N.B. (~~[illegible]~~) The balun losses here should be removed. //

With this change

$$\eta_\ell \Rightarrow \left\{1 - .03 - .073\sqrt{\frac{f}{10}}\right\} e^{-(.0226 f)^2}$$

$.042\sqrt{f}$

$$\rightarrow \eta_{app} = 0.67\left(1 - .052/\sqrt{f}\right)\left(.97 - .073\sqrt{\frac{f}{10}}\right) e^{-(.0226 f)^2}$$

$$\rightarrow \eta_B = .88\left(1 - .052/\sqrt{f}\right)\left(.97 - .073\sqrt{\frac{f}{10}}\right) e^{-(.0226 f)^2}$$

(.064 instead of .073 with no balun)

The T_{sp} is also changed. The balun emission goes to $80 \times .04 = \underline{3.2}$ instead of 2.2 (leave it out)

$$\rightarrow T_{sp} = \frac{7.8}{\sqrt{f}} + 6.6\sqrt{f} + \underbrace{16}_{\text{sky+LNA}}$$

$= 39.3$ at 10 GHz
$= 30.4$ at 1 GHz
30.8 at 2 GHz

Below 10 GHz for T_{sp} add $1.7 f^{-.8}$. It looks like the diffraction is greater. It hardly matters down to 500 MHz

---

One problem above is that the 3% feed loss scales with frequency but it shouldn't. It should be ~ constant.

Get the x-band horns

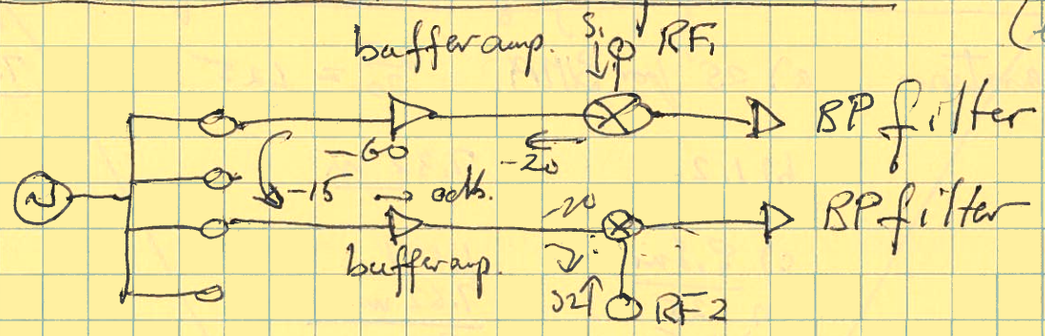
No, the losses in the radiating regions increase with frequency

In T_{sp} the middle term is mostly from input losses at 280 K. If these losses were cooled to 80 K the emission would be reduced by a factor of 4 from $20\sqrt{\frac{f}{10}}$ to $5\sqrt{\frac{f}{10}}$, about 30% overall at 10 GHz

15–20% at the lower frequencies

Possible cross-talk in the first LO

(John D.)
(+ Doug)



mixer: RF ↔ LO is −20db

splitter isolation: ~ −15 db.

bufferamp: reverse isolation: −60 db
gain at ≲ 10 GHz ≲ 0 db.

S1: −20 − 60 − 15 − 0 − 20 − 5 (reflection) = −120db

For the radar case the signal is +60 db above the noise

+60 − 120 = −60 db w.r.t. the S2 noise.

Integrate for 100 seconds. The radar BW is ~ 1 MHz.

$\sqrt{BT} = \sqrt{10^8} = 10^4$ 40db.

The cross talk is −20db w.r.t. this level. Also there will be different delays.

Conclusion: Phase switching of the first LO is un-necessary

The second LO might still be important to switch.

Integration and Test next Tuesday 2PM

Possible close spacing of the ATA for survey

Separations a) 25′ for BIMA $\frac{25}{20} = 1.25$ 7.62 m

b) 1.2 7.32 m

c) 8.8 m 1.44

d) 1.25 7.62 m

e) 1.15 7.0 m

Issues (a) closeness for good uv plane overlap

(b) pointing accuracy: better than 1/30 of a beam. this should be easy for HI.

(c) safety system: the wires proved reliable for BIMA

(d) Use of a single dish. (Bracewell's?)

i) stiffness

ii) surface accuracy $\sigma \leq 0.8$ mm

iii) integration time to compare with ATA

For $\theta_{1/2}$ = FWHM $g(\theta) \propto e^{-2.76(\theta/\theta_{1/2})^2}$

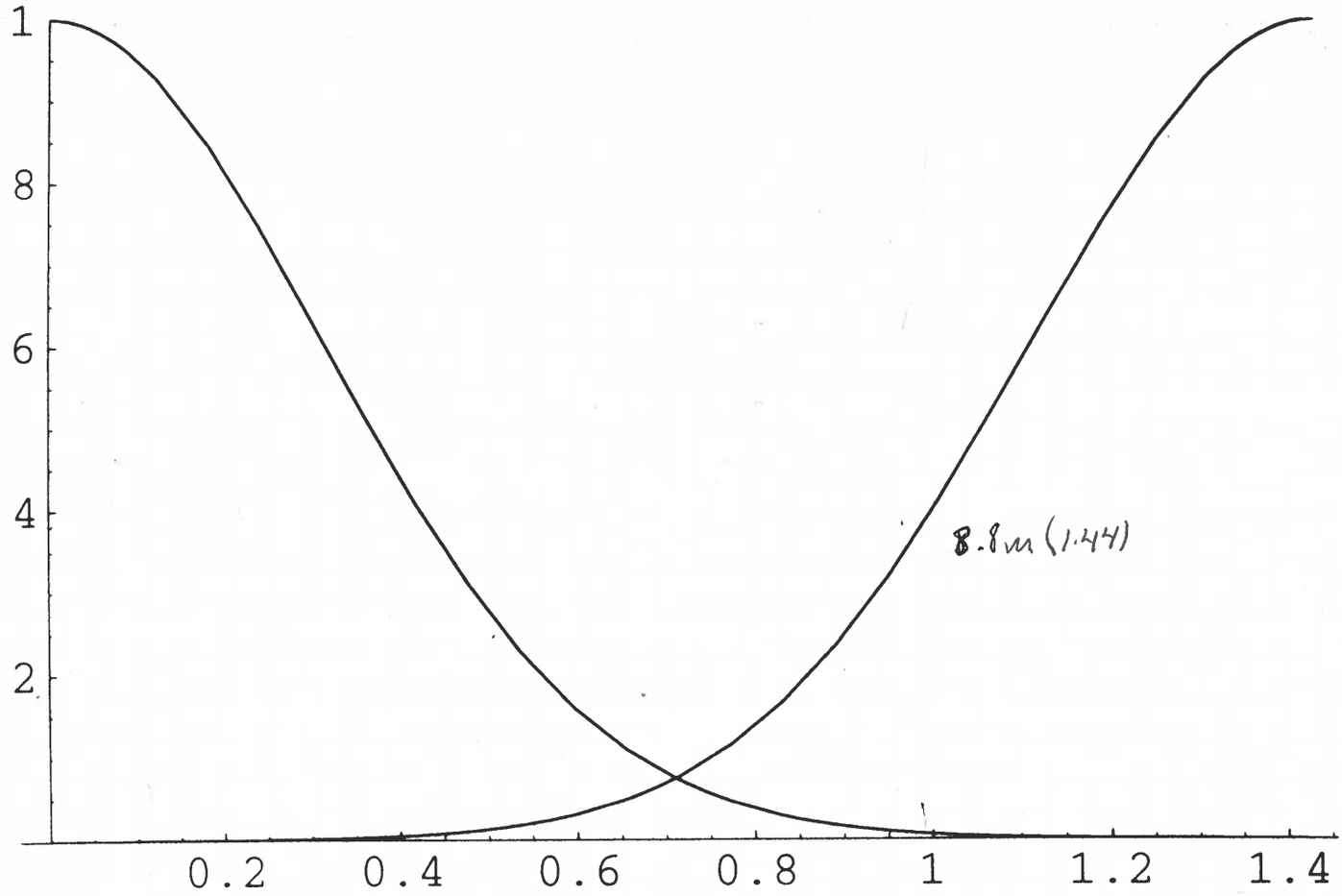
and $G(\beta) \propto e^{-3.58\,\theta_{1/2}^2\beta^2}$; $\theta_{1/2} = \frac{1.2\lambda}{D}$

$$3.58\,\theta_{1/2}^2\beta^2 = 3.58\left(\frac{1.2\lambda}{D}\right)^2\beta^2 = 5.155\,\frac{\beta^2}{D^2} = 5.155\,x^2$$

where $x = \beta/D$ $e^{-5.155(x)^2}$ and $e^{-5.155(x-x')^2}$

Try a range of x' : 1.15, 1.2, 1.25, 1.31, 1.415

i.e. 1.15 (7.0 m); 1.2 (7.32 m), 1.25 (7.62 m); 1.31 (8.0 m)



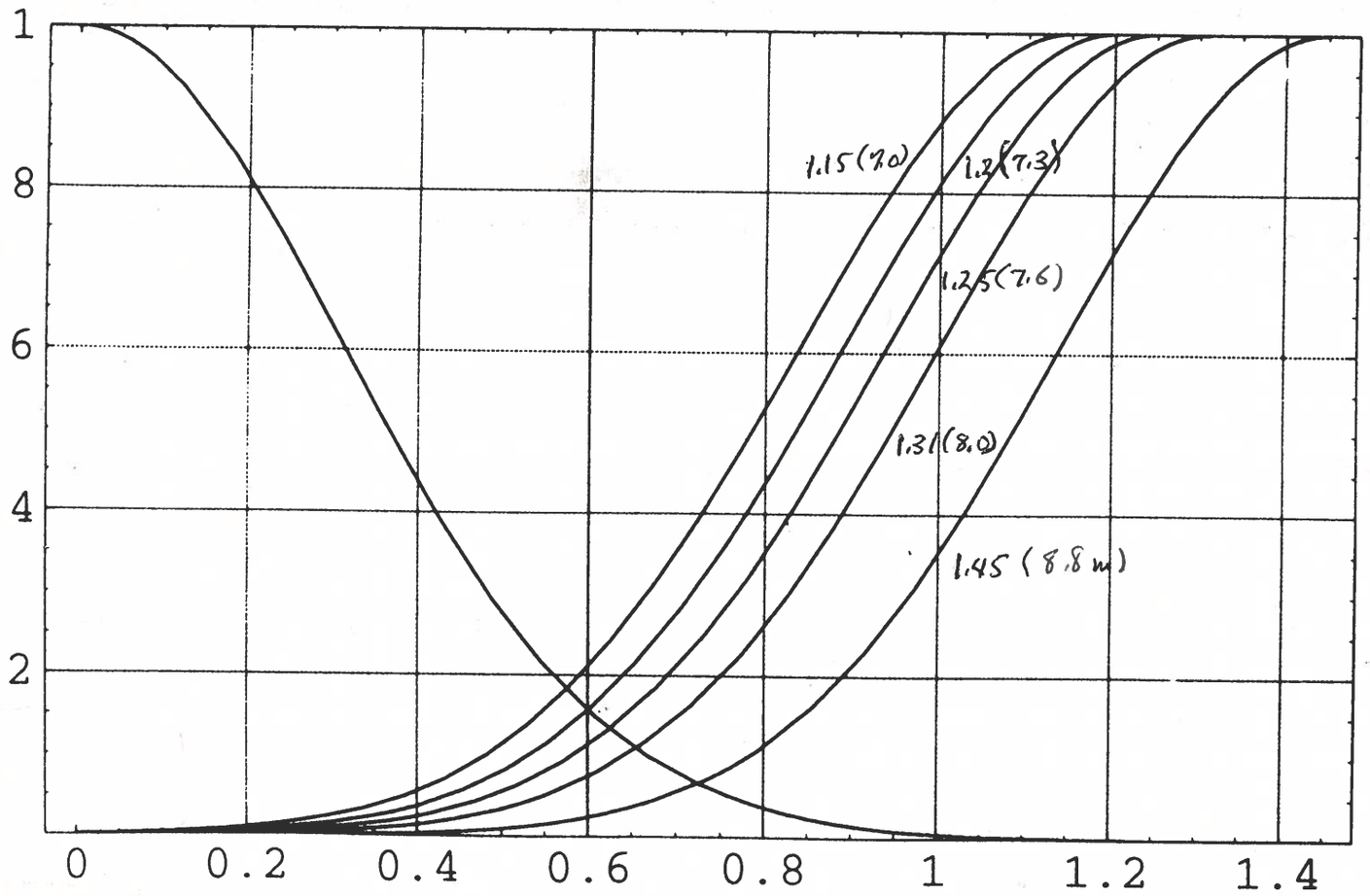
The combination of the transfer function of the single dish and that of the mosaiced array.

For a new feed test silo at Matt's place

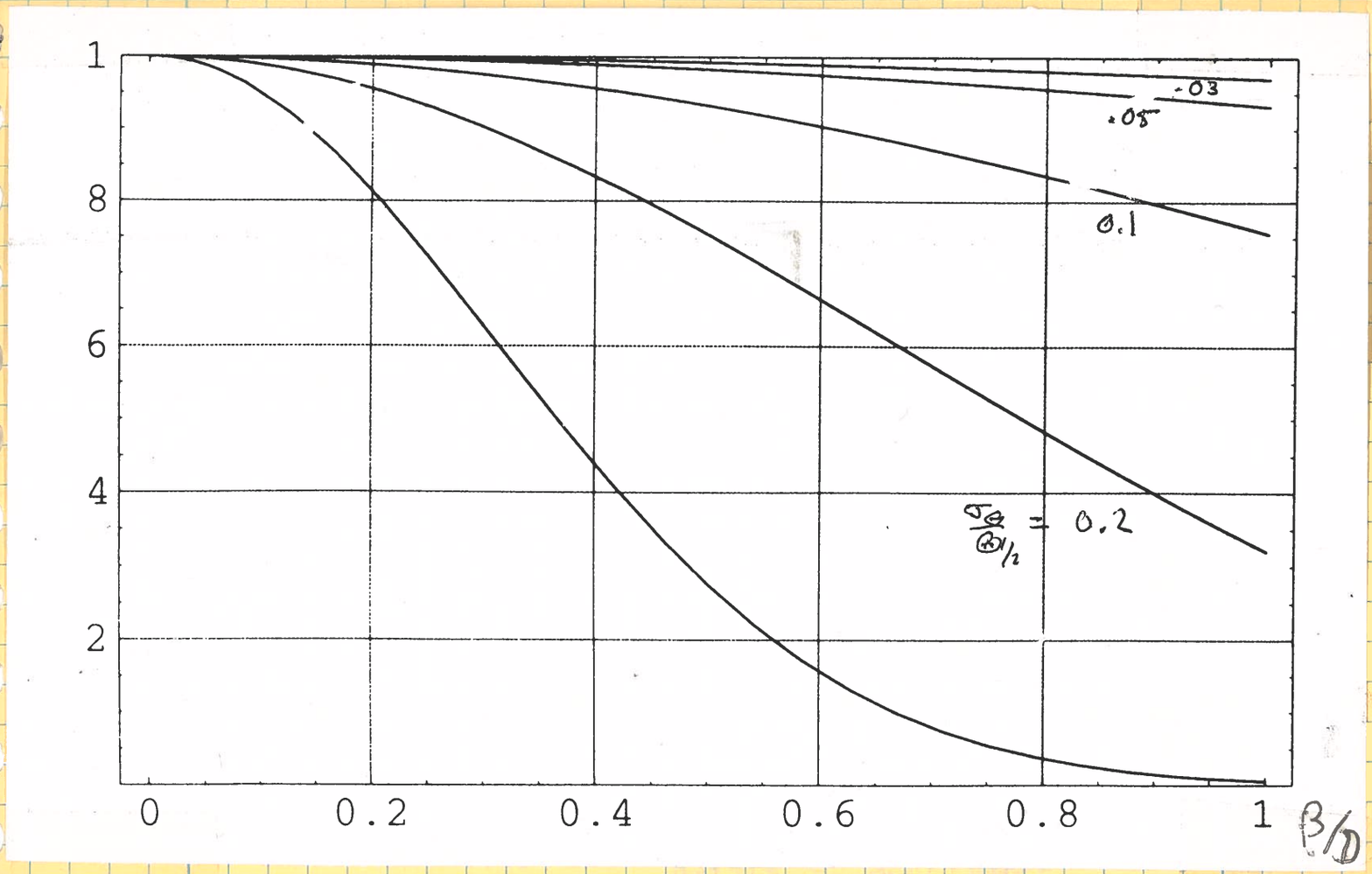
What is the minimum silo diameter?
From II p.18 the echo from the cylindrical shroud was found to be: $P_r/P_i \equiv [G(90°)]^2/10 \; \lambda/a$. a is the radius
$G(90°) \sim .035$

For the longest wavelength $\lambda = .60$ m and $a = 1.2$ m (as for the secondary) $\frac{P_r}{P_i} = \frac{(.035)^2}{10}\frac{0.6}{1.2} = 6\times10^{-5}$ −42 db

Since it's an asymptotic expansion, using $\lambda/a < 1$ probably is inaccurate. Maybe we could go down to $a = 0.9$ m
It's so small anyway maybe $a = \lambda = 0.6$ would be ok.
It's still −34 db at $\lambda = 0.6$ m, and much smaller at shorter wavelengths. That would be 1.2 m diameter 4 feet in diameter
[ok for Dave's silo.]



Overlap of single dish and array transfer functions with different minimum spacings shown as ratios to the dish diameter



Effect of pointing Errors on the Errors of the Transfer Function

The lowest curve is the single antenna transfer function.

The errors on the array Transfer function are the Same.

To recover images with errors of the order of 1-2% (dynamic range in fidelity of 50-100) requires pointing errors smaller than 1/30 of a beam.

For the single dish $F\{T_A(x,y)\} = V(u,v)\, G(u,v)$

with pointing errors δx and δy

$$G_{eff} = G(u,v)\, e^{i 2\pi (u\delta x + v\delta y)}$$

$$\Delta\phi = 2.4\pi\, s\, \sigma_\theta / \Theta_{1/2} \qquad \text{where} \quad s = \frac{\beta\lambda}{D}$$

$s = 1/2$ at the overlap

For $\sigma_\theta = 0.05\, \Theta_{1/2}$ $\qquad \Delta\phi = 1^\circ$ at $s = 1/2$

A 200 MHz filter for the RF converter board of the RPA.

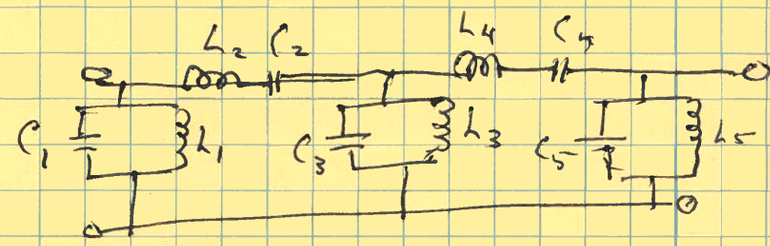
From p.318 of Zverev this is an $n=5$ BP filter with 0.1db ripple. $B = 50$ MHz @ 200 MHz center freq

$f_0 = 200\ MHz = f_r$ $\quad a = \frac{f_0}{f_2 - f_1} = \frac{200}{225-175} = 4.0$

$n = 5 \quad R_0 = 1$ $\quad L_r = R_r/\omega_r = 39.8\ nh$, $C_r = \frac{1}{\omega_r R_r} = 15.9\ pf$

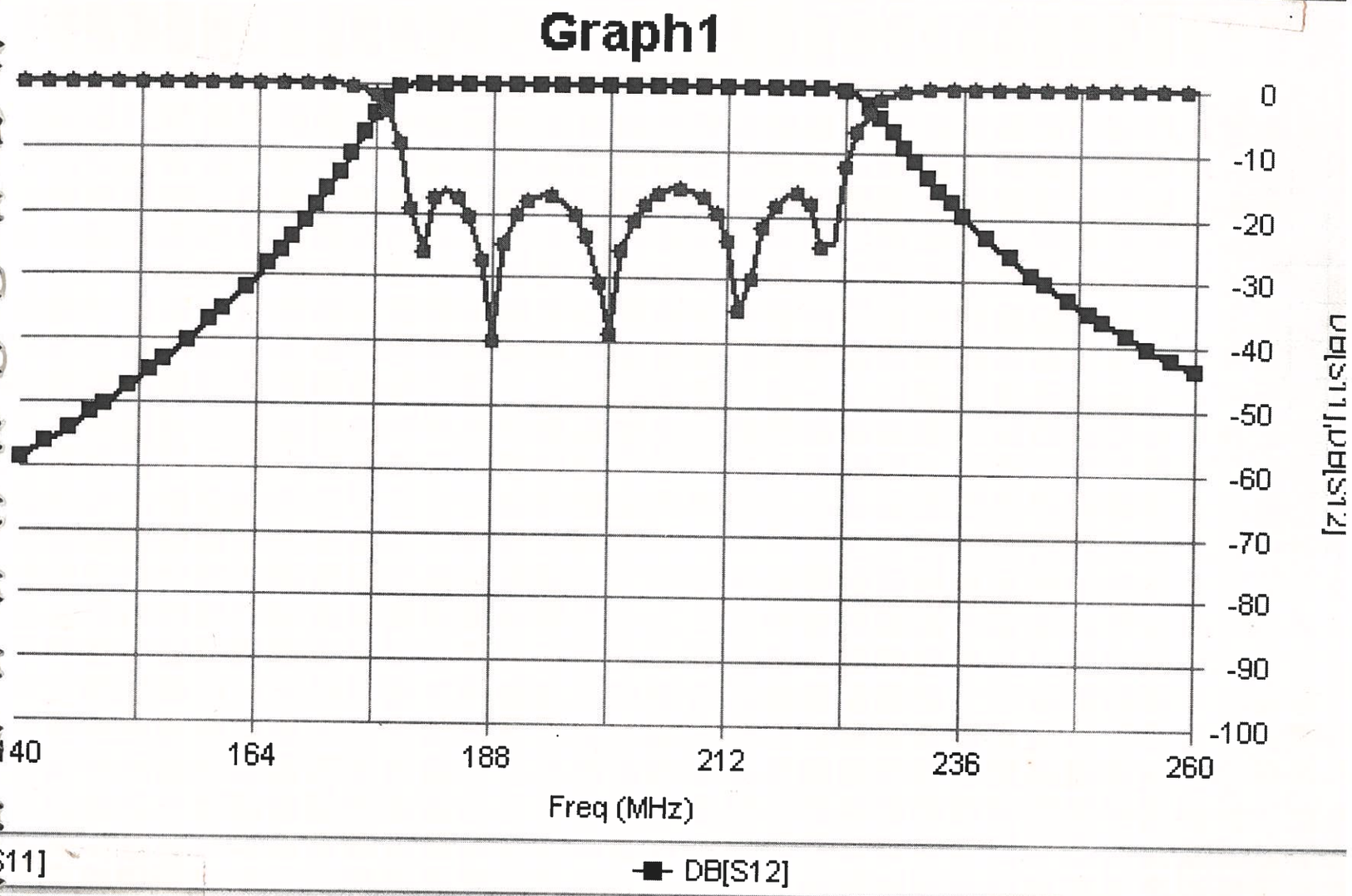
LP circuit

$C_1' = 1.3013$
$L_2' = 1.5559$
$C_3' = 2.2411$
$L_4' = 1.5559$
$C_5' = 1.3013$

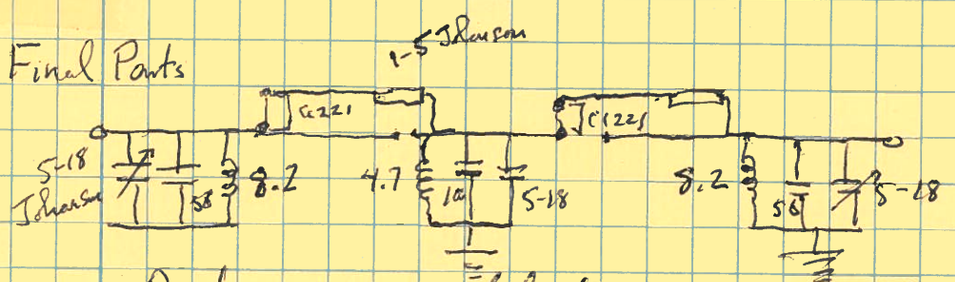


$C_1 = aC_1' C_r = 82.76\ pf$
$L_1 = [aC_1']^{-1} L_r = 7.65\ nh$
$C_2 = [aL_2']^{-1} C_r = 2.55\ pf$
$L_2 = [aL_2'] L_r = 247.7\ nh$
$C_3 = [aC_3'] C_r = 142.5\ pf$
$L_3 = [aC_3']^{-1} L_r = 4.44\ nh$
$C_4 = 2.55\ pf$
$L_4 = 247.7\ nh$
$C_5 = 82.76\ pf$
$L_5 = 7.65\ nh$

The frequency Response ~~and~~ for S_{11} and S_{21} are shown for these part values on the next page. This plot is from the program.



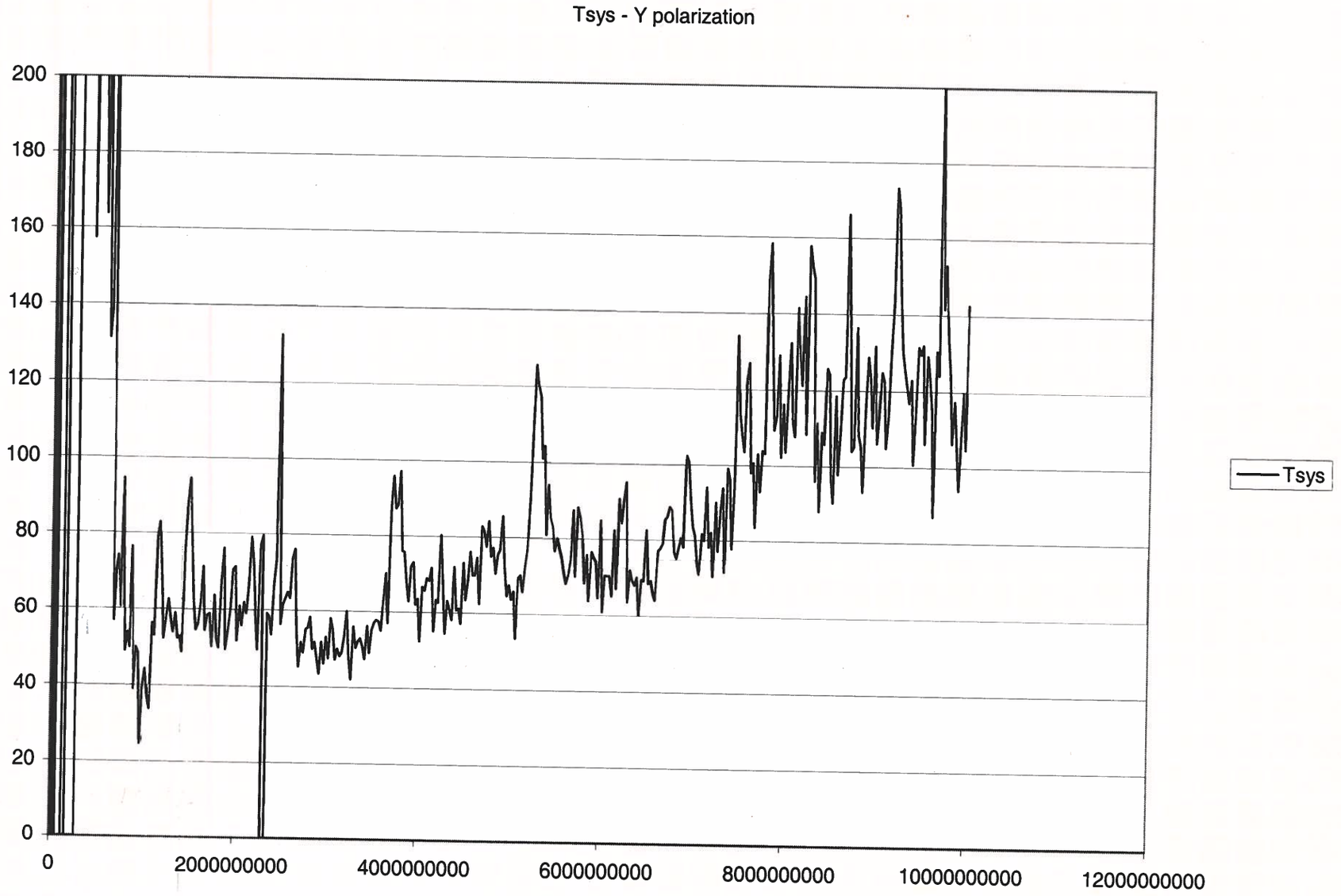
Final Parts

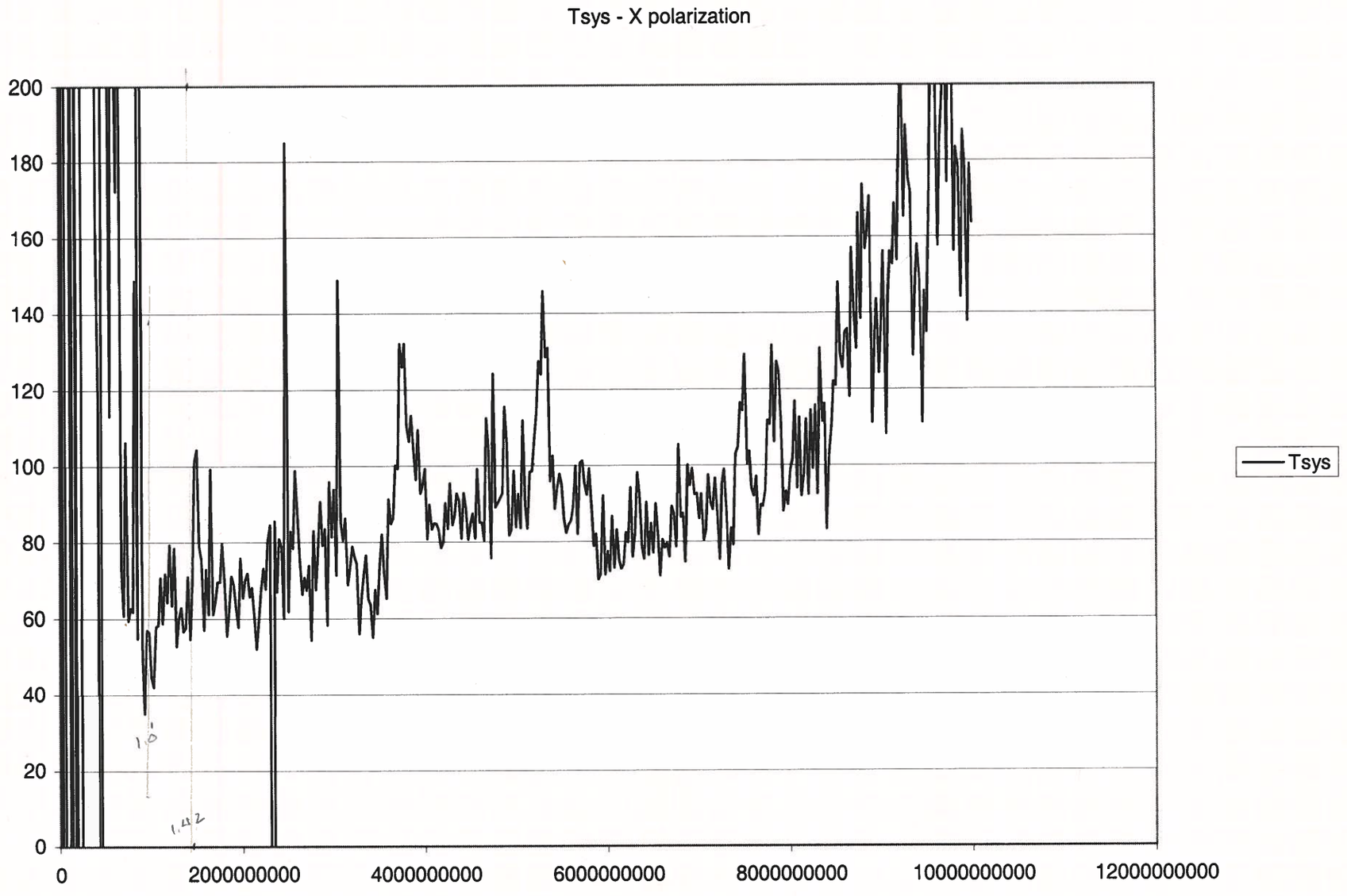


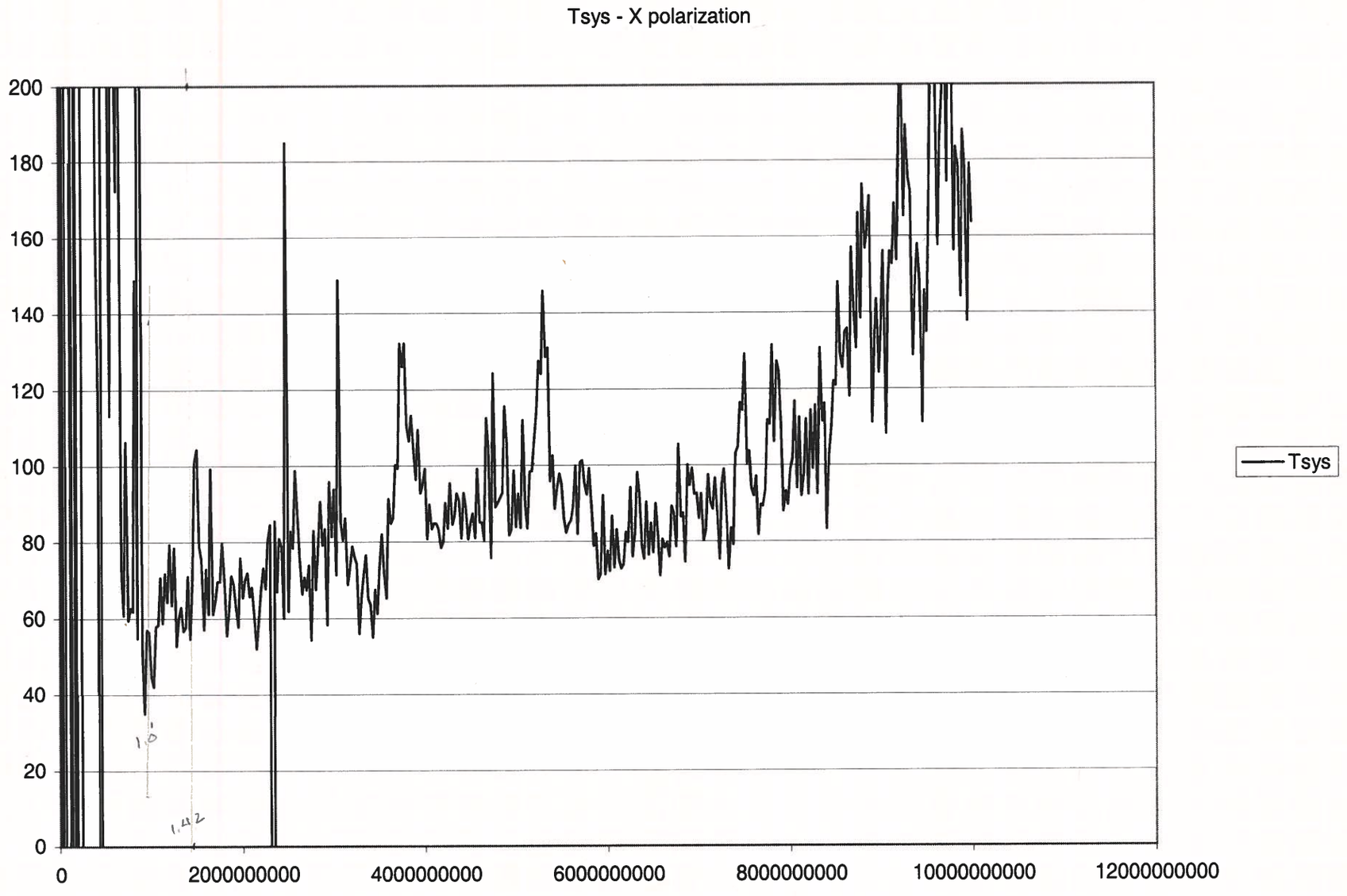
Parts are connected to 50Ω microstrip line with narrow gaps for the series resonant circuits. The series resonant circuit is suspended above microstrip to be isolated from ground.

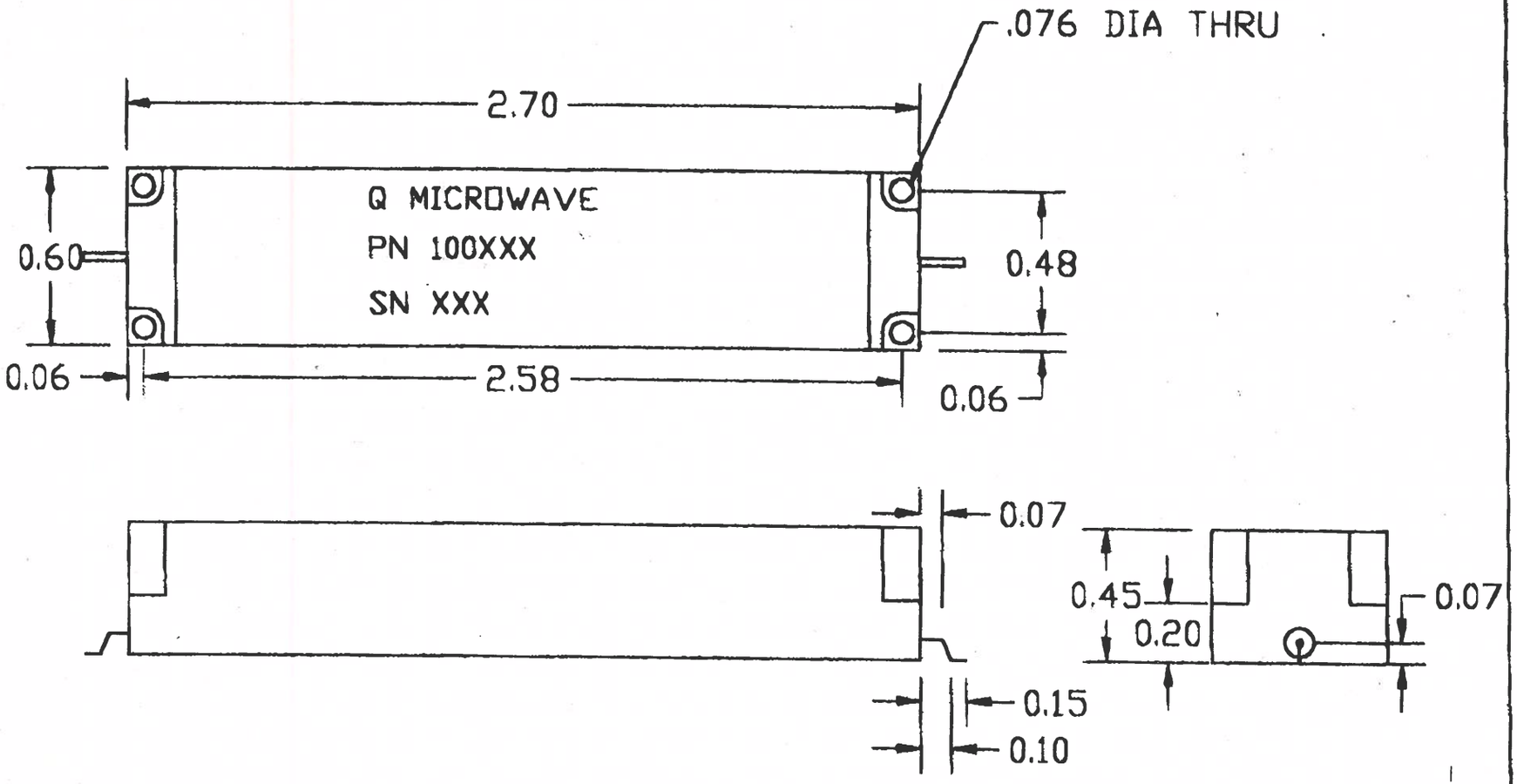
The tuned response follows closely the calculated response shown above, including max S_{21} ≤ -16 dB.

Tsys measurements with the feed in the Silo. References are cold sky and an ambient absorber over the top.




Tsys - X polarization
200
180
160
140
120
100
80
60
40
20
0
0
2000000000
4000000000
6000000000
8000000000
10000000000
12000000000
Tsys
1.0
1.42


Tsys - X polarization
200
180
160
140
120
100
80
60
40
20
0
0
2000000000
4000000000
6000000000
8000000000
10000000000
12000000000
Tsys
1.0
1.42



Filter foot print for the RPA board.

Q MICROWAVE	CAGE CODE 1H177	DRAWING NO 100 -MD	SHEET 2 OF 2	REV —

10/29/04 Jeff Kaufmann reported tests of the antenna 1 feed. He verifies that one polarization (not clear which) has ~20db more loss than the other. (Call Monday morning for any further updates). This was the latest feed with LNA to be installed, and recollections of several pple are that the 'bad pol.' turned up shortly after installation.

11/30/04 Test delayed to 12/13/04

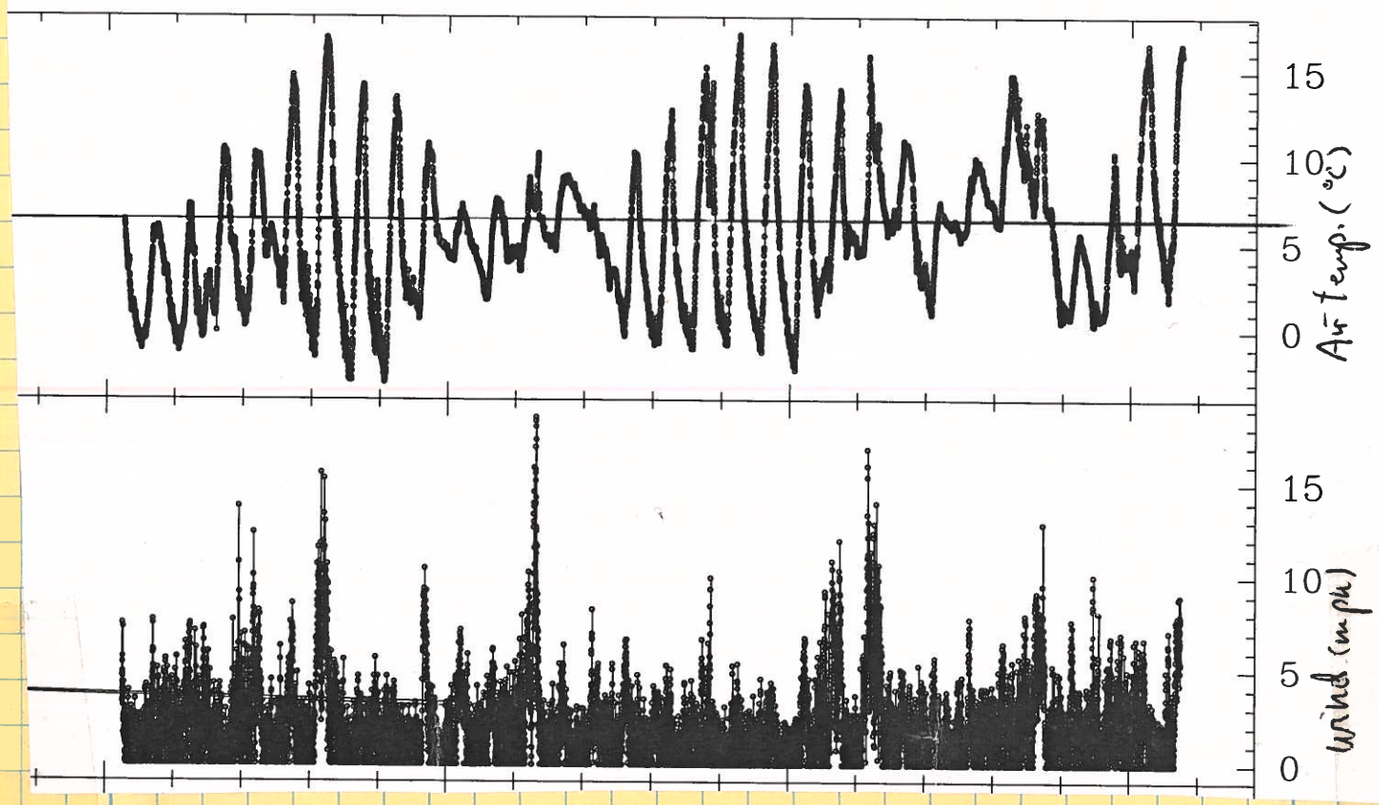
Integration + Test Dec. 14 @ Hat Creek

WJW

Hat Creek air Temperatures for Jan '03 and July '04.

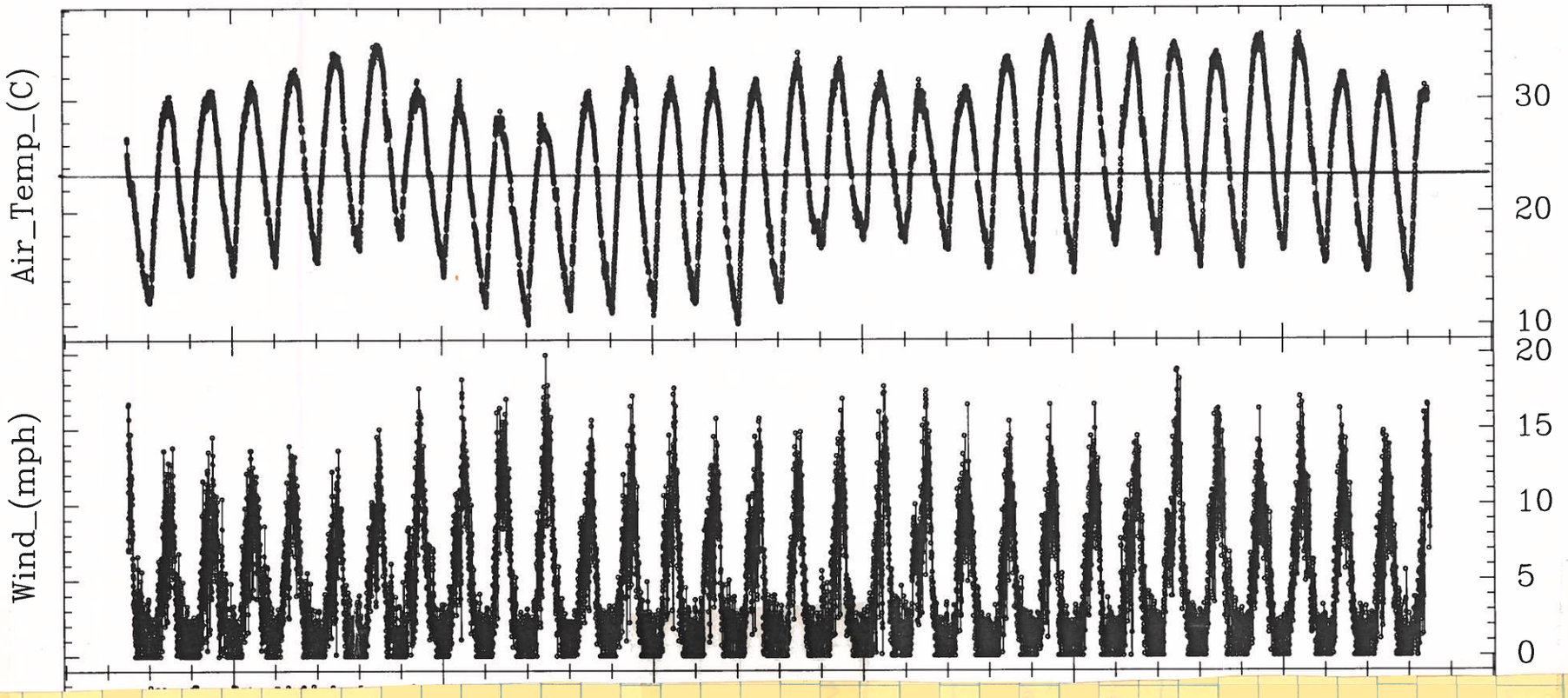
JAN '03

HCRO Weather Data – tmp



July '04

HCRO Weather Data – tmp



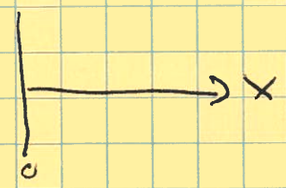
January '03 average = 6.5°C ⇒ 43.7°F

July '04 average = 23° ⇒ 73.4°F

} surface values.

Estimate the daily and annual temperature variations in the soil due to surface temperature as shown in the preceeding plots.

Diffusion into the ground.



$$\frac{\partial^2 T}{\partial x^2} = \frac{c_v}{k}\frac{\partial T}{\partial t} = \frac{1}{D}\frac{\partial T}{\partial t}$$

D (diffusivity) $= k/c_v$

For $T(0,t) = T_s \cos \Omega t = T_s R_e\{e^{i\Omega t}\}$, Ω daily or annual

$$\frac{d^2 T}{dx^2} = \frac{i\Omega}{D} T \quad ; \quad T \propto e^{\lambda x} \quad ; \quad \lambda^2 e^{\lambda x} = \frac{i\Omega}{D} e^{\lambda x}$$

and $\lambda^2 = i\Omega/D$ $\qquad \lambda = \pm\sqrt{\frac{i\Omega}{D}} = \pm\frac{(1+i)}{\sqrt{2}}\sqrt{\frac{\Omega}{D}}$

For $0 \le x \le \infty$ take the $(-)$ sign. $\quad T(x) = T_s e^{-(1+i)\sqrt{\frac{\Omega}{2D}}x}$

$$T(x,t) = R_e\left\{T_s e^{-(1+i)\sqrt{\frac{\Omega}{2D}}x} e^{i\Omega t}\right\}$$

$$T(x,t) = T_s e^{-\sqrt{\frac{\Omega}{2D}}x} \cos\left(\Omega t - \sqrt{\frac{\Omega}{2D}}x\right)$$

Let $\delta_T = \sqrt{\frac{2D}{\Omega}} = \sqrt{\frac{2k}{\Omega c_v}}$ $\qquad c_v = \rho c$

For minerals generally, use $c = 0.2 \frac{cal}{°C\text{-}gm}$; for solids.

$c_v = 6$ cal/deg-mole $= 3NK = \frac{3\rho k}{\nu_0 m_H}$; so $c = \frac{3k}{\nu_0 m_H} \sim 0.2$

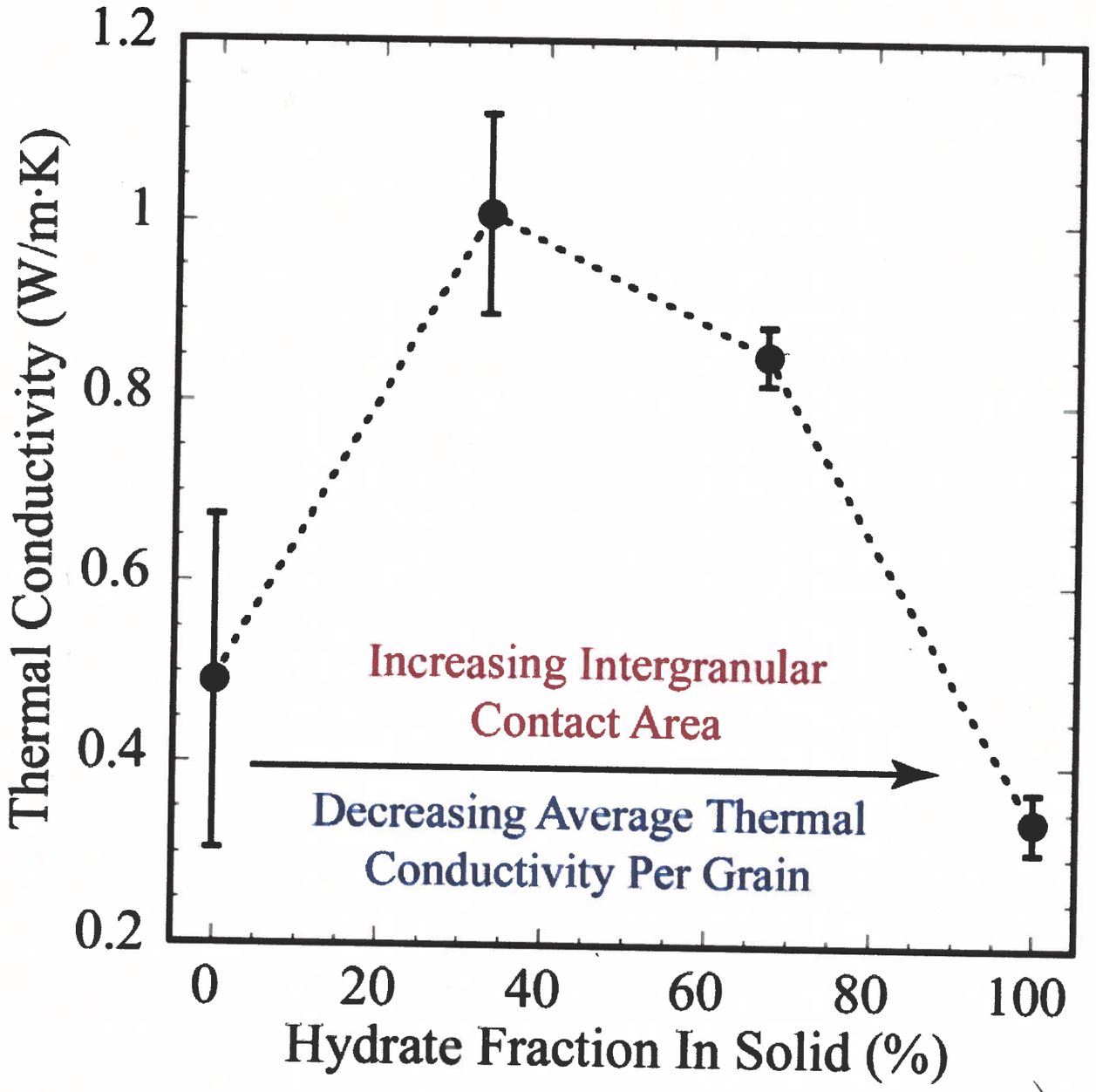
The Hat Creek soil is sandy and very dry in the summer. From the WES one example shows $\rho = 1.8$ gm/cm^3 for sandy soil.
Hence: $c_v = (1.8)(0.2) = 0.38$ cal/deg-cm^3

$\Rightarrow$ We use $k = 0.7$ watts/m-K from the plot on the next page[36]

As for common brick, 20 cm thick cinder agregate, or cement plaster.

$$k = 0.7 \text{ watts/m-K} = \frac{0.7\left(\frac{1}{4.2}\right) \text{cal/sec}}{100 \text{ cm-}°C} = \frac{.0017}{\text{sec}} \frac{\text{cal}}{\text{cm-}°C}$$

Hydrates From Ice (HyFI): Figure 3



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http://woodshole.er.usgs.gov/operations/hi_fi/methane_sand_results/Figure3.html

WGW

12/31/2004

Add the effects of two waves and plot the temperature sum at 3' and 4' depth.

LABOR READY
Dependable Temporary Staffing

RECB is ok

Filter 15.2 – 15.8

$IF_1 \sim 15.6 = LO_1 - RF$, $LO_1 = 15.6 + RF$

image for LO_2 is at 14.06, now ~ 14.8

for RF = .5, $LO_1 = 16.1$

RF = 11.5, $LO_1 = 15.6 + 11.5 = 27.1$

RF = 12.5, $LO_1 = 15.6 + 12.5 = 28.1$ ok

Image of 2nd LO is at (now) ~

top edge of the RF band must be ≤ ~~14.8~~ 14.6

IF power splitter ≤ 10.5 $LO_1 = 15.6 + 14.6 = 30.2$ ok

like the 17GHz

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$$\delta_T = \sqrt{\frac{2k}{\Omega_d C_v}} = \sqrt{\frac{(2)(.0017)}{(.38)(7.3\times10^{-5})}} = 11 \text{ cm.}$$

for the one day period ; one day = 86,400 secs. ; $\Omega_D = \frac{2\pi}{86,400} = 7.3\times10^{-5}$

For a depth of 24" = 61 cm , $e^{-\frac{61}{11}} = .0039$

So at 24" (61 cm) depth the daily periodic variation is reduced to 0.4%.

For the annual period, 365 days, $\delta_T = 11\sqrt{365} = \underline{2.1 \text{ m}}$

The corresponding variation of 0.3% occurs at a depth of 11.65 m

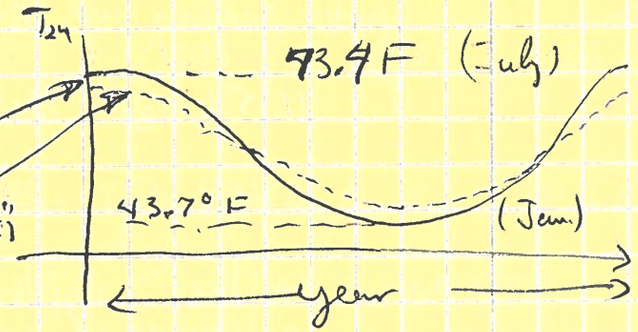
The annual variation is not at all attenuated, and we get the full surface temperature variation at 24" (61 cm). The temperature ~~with~~ corresponding to the annual variation will remain at 24" depth. The annual variation is shown by the averages of the January '03 and July '04 measurements

$$\langle T_{Jan} \rangle = 6.5\,^\circ C = 43.7\,^\circ F$$
$$\langle T_{July} \rangle = 23\,^\circ C = 73.4\,^\circ F$$
$$\Big\} \; \Delta T \cong 30\,^\circ F$$

~~At 24" we will get this~~

Surface :

At 61 cm (24")



Actually, at 61 cm (24") the annual variation is reduced by the factor : $e^{-.61/2.1} = .75$

Annual average $\frac{6.5+23.0}{2} = 14.75\,^\circ C \Rightarrow 58.6\,^\circ F = \langle T_{ann} \rangle$ @ depth of ~4 m

~~Ave. Summer peak at the surface: $\Delta T = 73.4 - 58.6 = 14.8\,^\circ F$~~

Ave. Summer peak at 61 cm (24") = .75 x 14.8 = 11.1 = ΔT

Summer peak at 61 cm : 11.1 + 58.6 = $\underline{69.7\,^\circ F}$ ≅ 70°F

Winter minimum at 61 cm : should be 58.6 − 11.1 = 47.5°

Expected noise with $T_{dewar} = 50°K$.

$\alpha_p = R/Z_0$ $\qquad Z_0 = 220\,\Omega$, $S/D = 3.2$

$\alpha_p \Rightarrow .060 R_s$ (II p 68)

$R_s = .0020\sqrt{f\rho}$ $\qquad$ take $f = 10\,GHz = 10^{10}$

$\rho \cong .05 \times 10^{-6}\ \Omega\,cm$ @ $T = 50K$

$R_s = .002\,[10^{10} \cdot .05 \times 10^{-6}]^{1/2} = \cancel{\ } = .0447$

$\alpha_p = .060 R_s = \cancel{.060(12.4)} = (.060)(.0447) = .0027\ cm^{-1}$

emissivity: $T \cong T\alpha_p = [50][.0027] = .134K/cm$

For $L = 18\,cm$ $\boxed{T = 2.4°K}$ at $f = 10\,GHz$

At $f = 1.4\,GHz$, $T = 2.4\sqrt{\frac{1.4}{10}} = 0.9K$

The actual cold balun length is $12.4\,cm = L$
so this gives $T = [50][.0027][12.4] = 1.7°$ @ $10\,GHz$

Note the expected noise at higher temperatures, just scaling up the uniform loss @ $8\,GHz$

16.5 K @ 130K temperature

68.3K @ 297K Temperature $\quad$ {.9db measured}

so the 1.7K contribution at 50K is striking as compared with 5.3K @ 80K temperature

Ambient components: 2% feed, 1% tip board, 1.2% amp line $= 4.2\%$, @ 297K $\Rightarrow$ 12.5K

Get the gold + silver conductivities, compare with Copper

What are the expected tracking errors if the $\sigma_\theta = 10''$ and we are at 1/10 beam width off, 120'' @ 3cm

pattern $\propto e^{-.69\left(\frac{2\theta}{\theta_{1/2}}\right)^2} = e^{-2.76\left(\frac{\theta}{\theta_{1/2}}\right)^2}$

$\theta_{1/2} = 20' = 1200''$ @ $\lambda 3$ cm.

			Δ
$e^{-2.76\left(\frac{\theta-120}{1200}\right)^2}$:	$\theta_e = 10''$	$e^{-2.76\left(\frac{10-120}{1200}\right)^2} = .9771$	$-.0043$
	$\theta_e = -10''$	$e^{-2.76\left(\frac{-10-120}{1200}\right)^2} = .9681$	$.0047$
	$\theta_e = 0$	$e^{-2.76\left(\frac{0-120}{1200}\right)^2} = .9728$	

This is $\pm 0.5\%$ at 1/10 beam pointing error. ← absolute error

This would tend to average out if the jitter is fast.

At the half power point the jitter is $\pm 2\%$. This will also average out if the jitter is rapid.

Jitter ~~only~~ @ $\frac{\theta_{1/2}}{10}$)

Heat/Air flow Calculations for the Node Air Conditioning

Air @ STP : One mole occupies 22.4 liters $= 22.4 \times 10^3\ cm^3$

One mole volume in $(ft)^3$: $22.4 \times 10^3 \left(\frac{ft}{(12)(2.54)cm}\right)^3 = (.79)\ ft^3$; $(ft)^3 = 1.26$ mole

$1\ ft = 12\ in = 12(2.54)\ cm$

Heat capacities for diatomic molecules : $C_p = 7\ cal/°C/mole$ (O_2 and N_2)

Heat capacity of one CF $= 8.8\ cal/°C$ for air at STP

Heat Air flow @ 1 CFM of air $= 8.8\ cal/min/°C$; $\boxed{1\ watt = 3.41\ BTUH}$

$1\ cal = 4.18\ joules$ $\quad 1\ CFM\ flow = \frac{(8.8)(4.18)\ joules}{60\ sec\ 1.8°F} = \frac{.341\ watts}{°F}$ correct

$1\ CFM = \frac{(.341)\ BTUH/(3.41)}{°F} = 1.16\ BTUH/°F$

$Q(BTUH) = 0.87(CFM)\ [\Delta T(°F)]$ @ STP

For cooling at the antennas The maximum ave gnd temp (in July) will be about 70°F, worst case.

~~[illegible]~~ = 1 CFM => .336 watts/°F => 1.15 BTUH/°F ✓

~~P(watts) = 2.9 F ℓ (CFM) ΔT(°F)~~

~~For a temperature rise of 30°F (70-100) and Fℓ = 50 CFM~~
~~P = (2.9)(50)(30) = 4400 watts~~

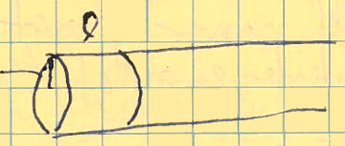
$$\boxed{P(\text{watts}) = .336\,(\text{CFM})\,\Delta T\,°F}$$ ✓ or ~~$P(BTUH) = 1.15\,(CFM)\,\Delta T\,(°F)$~~ no

For $\Delta T = 30°F$ and 50 CFM P = 540 watts 540
This is for a temperature rise of 30°F at 50 CFM

Air and radial heat flow in the 3" PVC Tube.

3" = .25 ft = 2a

velocity and flow rate



volume = $\pi a^2 \ell$

For one cu ft. $\pi a^2 \ell = 1$ $\ell = \frac{1}{\pi a^2} = \frac{1}{.049} = 20.4$ ft (long)

$\bar{V} = \frac{\pi}{4} d^2 \ell$ $\dot{V} = \frac{\pi}{4} d^2 \dot{\ell}$ = area × $\dot{\ell}$ = area × v

For one CFM $v = \dot{\ell} = \frac{20'}{60\,sec.} = \frac{1}{3}$ ft/sec. = $\frac{1}{3}$ ft/sec.

For 50 CFM $\dot{\ell} = \frac{50}{3}\,\frac{ft}{sec.} = \frac{50}{3}\,\frac{(.3\,m)}{sec} = 5\,m/sec$

From the prob. book p. 20 $\rho = \frac{P}{RT} = \frac{(1)\,1.01\times10^5}{(287)(300)} = 1.17\,Kgm/m^3$

$\mu = 2.57\times10^{-5}$ Kgm/ms
$K = .0386$ W/m°C
$C_p = 1.025$ KJ/Kg°C

d = 3" = .0762 m

$R_e = \frac{\rho v d}{\mu} = \frac{(1.17)(5)(.0762)}{2.57\times10^{-5}} = 17{,}345$ (turbulent)

$P_R = \frac{\mu C_p}{K} = \frac{(2.57\times10^{-5})(1.025\times10^3)}{.0386} = .68$

Nusselt no. $Nu = .023 (R_e)^{0.8} (P_r)^{0.4} = 48.54$ ✓ W/W

$h = \frac{K}{d} Nu = \frac{(.0386)}{(.0762)} 48.54 = 24.6 \ W/m^2 °C$ ✓ $\div 4.2$

Heat Flow $= hA\Delta T$ to the wall

Thermal diffusivity of the soil. $\alpha = \frac{K}{\rho c}$

$K = .0017 \frac{cal}{cm\,°C\,sec}$ ✓ $\rho c = .38 \frac{cal}{deg\,C\ cm^3} = c_v$ ✓

$\alpha = \frac{k}{\rho c} = \frac{.0017}{.38} \frac{cm^2}{sec} = \frac{(.0017)}{.38}\left(\frac{1}{30} ft\right)^2 \frac{3600}{hr} = .018 \frac{ft^2}{hr}$ ✓

$K = \frac{.0017\,(4.2)\,watts}{(1/30)\,ft\,(1.8)°F} = \frac{(30)(.0017)(4.2)}{ft\ 1.8°F} 3.41\ BTU/H = .406 \frac{BTU/H}{ft\ °F}$ ✓

Maximum required cooling for 12 stations @ 50CFM each
Outside air (at 105°F) is cooled to 70°F. midday

$P(watts) = .336\,(CFM)\,\Delta T = .336(12 \times 50)(35) = 7056$ watts. for the

This would also be the maximum heating at night

$12 \times 50 = 600$ CFM for the blower

If we double the air flow, so, we double the Reynold's numb
and raise Nu by $2^{0.8} = 1.7$
This will ~~change~~ increase h by 1.7 and lower Rair by 1.7, but not Rgrd or Re

use BTUH, ft, °F units

Temperature Drop (steady state) along the 3" PVC Tube with 50 CFM air flowing through.

Use the same 3' leg model as the report.

For the earth $K_e = \frac{.0017 \text{ cal}}{\text{cm °C sec}} = \frac{0.408 \text{ BTUH}}{\text{ft °F}}$

$R_e = \frac{\ln\left(\frac{OD}{ID}\right)}{2\pi K} = \frac{\ln\left(\frac{3.6''}{3''}\right)}{2\pi(.41)} = 0.96$ (for unit length)

= .91 for 3.5" diameter

= .85 for 4" diameter

Heat flow to the wall $H = hA\Delta T$, so $R_{air} = \frac{1}{Ah}$

$R_{air\,wall} = \frac{1}{Ah} = \frac{1}{\pi d h} = \frac{1}{\pi(.25)\,4.28} = .30$ (for unit length)

$R_{pvc} = .22$ from the report.

Heat flow radially from the warm air for a length dx; $A \to \pi d\,dx$

$H(x) = \frac{(T - T_w)}{R_T}\,dx$ where $R_T = R_{air} + R_{pvc} + R_{earth}$

total unit length thermal resistance to the outside

no. should be R_{air}

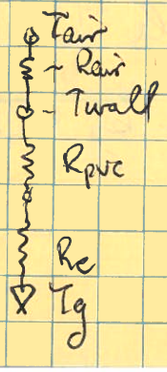
Associated with this heat flow to the wall, there will be a drop in temperature of the air dT, given by

$H = -(.87)(CFM)\,dT$

Equate these: $\frac{(T(x) - T_w)}{R_T\ (R_{air})}\,dx = -.87(CFM)\,dT$

x

Both T(x) and the wall temperature vary along the pipe. We can use the resistive divider to change this difference to the air-outside difference.



$\frac{T_w - T_g}{T_{air} - T_g} = \frac{R_{pvc} + R_e}{R_{air} + R_{pvc} + R_e} = \frac{R_{pvc} + R_e}{R_T} = K$

$T_w - T_g = (T_{air} - T_g)K$

$T_w = (T_{air} - T_g)K + T_g$

$$T_w = K T_{air} - K T_g + T_g$$

$$T_{air} - T_w = T_{air} - K T_{air} + K T_g - T_g = T_{air}(1-K) - T_g(1-K) = (T_{air} - T_g)(1-K)$$

$$1-K = 1 - \frac{(R_{pvc} + R_e)}{R_T} = \frac{R_{air} + R_{pvc} + R_e - (R_{pvc} + R_e)}{R_T} = \frac{R_{air}}{R_T}$$

$$T_{air} - T_w = (T_{air} - T_g)\frac{R_{air}}{R_T} \qquad \frac{T_{air} - T_w}{R_{air}} = \frac{T_{air} - T_g}{R_T}$$

$$(T_{air}(x) - T_g)\frac{R_{air}}{R_T}\,dx = -.87\,(CFM)\,dT$$

put in R_{air} $\rightarrow$ R_{air}

$$\frac{dT}{dx} = -\frac{(T_{air}(x) - T_g)\,R_{air}}{.87\,(CFM)\,R_T}$$

let $\tau = T_{air}(x) - T_g$

$d\tau = dT$

$$\frac{d\tau}{dx} = -\frac{\tau\,R_{air}}{.87\,(CFM)\,R_T} \qquad \tau(x) = \tau(0)\,e^{\lambda x}$$

where $\lambda = \dfrac{-1}{.87\,(CFM)\,R_T}$

Above $R_{air} = .30$, $R_{pvc} = .22$, $R_e = .96$; so $R_T = 1.44$

$$\lambda = \frac{-1}{.87\,(50)\,(1.44)} = .016 \qquad \tau(x) = \tau(0)\,e^{-.016x}$$

x	$e^{-.0033x}$	.009	.011	.02	.016x	.025
50	.84	.64	.58	.37	.45	.30
100	.72	.41	.33	.14	.20	.09
200	.52	.16	.11		.04	.00
250	.44					
300	.37					

With $R_e = .371$ $\frac{°F\text{-}ft}{BTU/h}$ instead of the above .96, $R_T = .89$

$$\lambda = \frac{-1}{.87\,(50)\,(.89)^2} = .009$$

For $R_T = 0.6$ the DAGO R_T

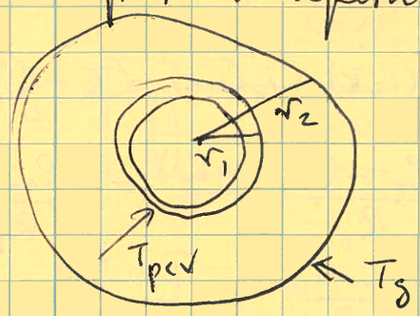
$$\lambda = \frac{1}{(.87)(CFM)\,R_T}; \text{ For } R_e = .96,\ \lambda = \frac{1}{(.87)(CFM)(1.44)} = .016$$

50

$\delta R = .0067$

@ 120 CFM

Temperature drop along the pipe for a periodic input temperature at the node.



In general $\nabla^2 T = \frac{1}{\alpha}\frac{\partial T}{\partial t}$ will describe the temperature distribution in the earth surrounding the PVC pipe. To get the frequency dependence, assume $e^{-i\omega t}$ time dependence.

i.e. $T(r,t) = \mathrm{Re}\left\{ T(r,\omega) e^{-i\omega t} \right\}$

$\nabla^2 T = \frac{1}{\alpha}(-i\omega) T$ Assume $\frac{\partial T}{\partial \theta} = 0$, only radial spatial variations

Also there's essentially no z-dependence, although this is the coordinate along the pipe. We assume that the radial heat flow has no local z-dependence.

Bessel's equation is: $\frac{d^2 T}{dr^2} + \frac{1}{r}\frac{dT}{dr} + \beta^2 T = 0$

$\beta^2 = \frac{i\omega}{\alpha}$ The region is $r_1 < r < r_2$; $\beta = \frac{(1+i)}{\sqrt{2}}\sqrt{\frac{\omega}{\alpha}} = (1+i)\sqrt{\frac{\omega}{2\alpha}}$

Let T_{pvc} = given at r_1 and T_g = given at r_2 and let's calculate the flow of heat at $z=0$, $r=r_1$ the outer radius of the PVC pipe. This is not quite correctly posed, but we find an asymptotic solution as $r_2 \to \infty$. We wish to find the ~~logarithmical~~ the time periodic distribution of temperature, and in particular, the ratio of the temperature, at $r_1 = .14$ ft to ~~its value at~~ the value of the heat flow at $r_1 = .14$ ft. This defines the effective input impedance of the path to $r=\infty$ at $r_1 = 0.14$ ft.

The formal solution of Bessel's equation is

$T(r) = A J_0(\beta r) + B Y_0(\beta r)$. Apply the boundary conditions

$$\begin{aligned} T_{pvc} &= A J_0(\beta r_1) + B Y_0(\beta r_1) \\ T_g &= A J_0(\beta r_2) + B Y_0(\beta r_2) \end{aligned} = \underbrace{\begin{bmatrix} J_0(\beta r_1) & Y_0(\beta r_1) \\ J_0(\beta r_2) & Y_0(\beta r_2) \end{bmatrix}}_{M} = \begin{bmatrix} A \\ B \end{bmatrix}$$

Then $\begin{bmatrix} A \\ B \end{bmatrix} = M^{-1} \begin{bmatrix} T_{pvc} \\ T_g \end{bmatrix}$

For $A = \begin{bmatrix} a & b \\ c & d \end{bmatrix}$; $A^{-1} = \begin{bmatrix} d & -b \\ -c & a \end{bmatrix} \frac{1}{(da-bc)}$

$M = \begin{bmatrix} J_0(\beta r_1) & Y_0(\beta r_1) \\ J_0(\beta r_2) & Y_0(\beta r_2) \end{bmatrix}$ has the following inverse

$$M^{-1} = \begin{bmatrix} Y_0(\beta r_2) & -Y_0(\beta r_1) \\ -J_0(\beta r_2) & +J_0(\beta r_1) \end{bmatrix} \frac{1}{\Delta} \quad ; \quad \Delta = J_0(\beta r_1) Y_0(\beta r_2) - J_0(\beta r_2) Y_0(\beta r_1)$$

The heat flow through the outer PVC pipe wall is:

$$H_0 = -k \, \text{Area} \left.\frac{\partial T}{\partial r}\right|_{r=r_1} = k\beta \left\{ A J_0'(\beta r_1) + B Y_0'(\beta r_1) \right\} \text{Area}$$

$$\begin{bmatrix} A \\ B \end{bmatrix} = \frac{1}{\Delta} \begin{bmatrix} Y_0(\beta r_2) & -Y_0(\beta r_1) \\ -J_0(\beta r_2) & J_0(\beta r_1) \end{bmatrix} \begin{bmatrix} T_{pvc} \\ T_g \end{bmatrix}$$

Finally

$$T(r) = \frac{\left[Y_0(\beta r_2) J_0(\beta r) - J_0(\beta r_2) Y_0(\beta r) \right] T_{pvc} + \left[-Y_0(\beta r_1) J_0(\beta r) + J_0(\beta r_1) Y_0(\beta r) \right] T_g}{\{\Delta\}}$$

$J_0'() = -J_1()$, $Y_0'() = -Y_1()$

$$H_0 = \frac{k\beta \, \text{Area}}{\Delta} \left[\left\{ Y_0(\beta r_2) T_{pvc} - Y_0(\beta r_1) T_g \right\} J_1(\beta r_1) + \left\{ -J_0(\beta r_2) T_{pvc} + J_0(\beta r_1) T_g \right\} Y_1(\beta r_1) \right]$$

$$H_0 = k\beta \, \text{Area} \left[\frac{\left\{ \overbrace{Y_0(\beta r_2) J_1(\beta r_1) - J_0(\beta r_2) Y_1(\beta r_1)}^{Q} \right\} T_{pvc} + \left\{ \overbrace{-Y_0(\beta r_1) J_1(\beta r_1) + J_0(\beta r_1) Y_1(\beta r_1)}^{P} \right\} T_g}{\Delta} \right]$$

Some numerical values: $\beta = (1+i)\sqrt{\frac{\omega c_v}{2k}}$

In cgs: $c_v = \rho c_p = .38 \frac{\text{cal}}{\text{cm}^3 \, {}^\circ\text{C}}$; $k = .0017 \frac{\text{cal}}{\text{cm} \, {}^\circ\text{C} \, \text{sec.}}$

For 1 day period. $\omega = \frac{2\pi}{T} = \frac{2\pi}{24 \cdot 3600} = 7.3 \times 10^{-5}$

$$\beta = (1+i) \sqrt{\frac{(.38)(7.3 \times 10^{-5})}{2(.0017)}} = \frac{(1+i)}{11.07 \text{ cm}} = (1+i) \, 2.753 \ \text{ft}^{-1}$$

$r_1 = 1.7'' = 4.32 \text{ cm} = 0.1417 \text{ ft}$

$\beta r_1 = \frac{(1+i) \, 4.318}{11.07} = 0.390(1+i)$; for $r_2 = 1.5' = 45.72 \text{ cm}$, $\beta r_2 = 4.130$

Let $x = r_2$, then $\beta r_2 = 2.753(1+i)x$

$P = -Y_0(\beta r_1)J_1(\beta r_1) + J_0(\beta r_1)Y_1(\beta r_1) = -0.816179 + I0.816179$, with $\beta r_1 = .39(1+i)$

$f_1(x) = \Delta = J_0[.39(1+i)]Y_0[2.7533x(1+i)] - J_0[2.7533x(1+i)]Y_0[.39(1+i)]$

$f_2(x) = Q = Y_0[2.7533x(1+i)]J_1[.39(1+i)] - J_0[2.7533x(1+i)]Y_1[.39(1+i)]$

$K\beta \text{Area} = 0.89(.406)(1+i)2.7533 = .9949(1+i)$

$$H(r_1) = K\beta \text{Area}\left\{\frac{Q}{\Delta}T_{pvc} + \frac{P}{\Delta}T_g\right\} = K\beta \text{Area}\left\{\frac{f_2(x)}{f_1(x)}T_{pvc} + \frac{P}{f_1(x)}T_g\right\}$$

For $\underline{x = 1.5'}$

$$H = (2.02745 - I1.09917)T_{pvc} - (.2408 + I.03068)T_g$$

For $\underline{x = 2.9065'}$

$$H = (2.02574 - I\,1.1001)T_{pvc} + (.00406 - i.00582)T_g$$

The term multiplying $T_g \to 0$ with large x. The term multiplying T_{pvc} remains the same. Dropping the second term gives us the assymptotic solution for $r_2 \to \infty$.

We should write $\Delta H = \dfrac{\Delta T_{pvc}}{Z_e}$

With these numbers $Z_e = (2.02574 - I1.1001)^{-1} = .3812 + 0.2070I$.

This is substantially less than the steady state value of $R_e = .96$.

The frequency dependence of $|Z_e|$

r_2 is large so the second term drops out.

Let x = period in days $\qquad \beta \propto \sqrt{\Omega} \propto \dfrac{1}{\sqrt{p}}$

So $\beta = \dfrac{2.7533(1+i)}{\sqrt{p}}$; $\beta r_1 = \dfrac{.39(1+i)}{\sqrt{p}}$; $\beta r_2 = \dfrac{8.00(1+i)}{\sqrt{p}}$

$$\beta K A_{rea} = \frac{.9945(1+i)}{\sqrt{p}} \qquad \Delta H = \left\{\frac{Q}{\Delta}\right\} \frac{.9945(1+i)}{\sqrt{p}}$$

$$|Z_e| = \left| \frac{\sqrt{p}}{.9945(1+i)} \left\{ \frac{\Delta}{Q} \right\} \right|$$

$$\Delta = f_1[p] = J_0\left[\frac{.39(1+i)}{\sqrt{p}}\right] Y_0\left[\frac{8.0(1+i)}{\sqrt{p}}\right] - J_0\left[\frac{8.0(1+i)}{\sqrt{p}}\right] Y_0\left[\frac{.39(1+i)}{\sqrt{p}}\right]$$

$$Q = f_2[p] = Y_0\left[\frac{8.0(1+i)}{\sqrt{p}}\right] J_1\left[\frac{.39(1+i)}{\sqrt{p}}\right] - J_0\left[\frac{8.0(1+i)}{\sqrt{p}}\right] Y_1\left[\frac{.39(1+i)}{\sqrt{p}}\right]$$

$$|Z_e| = \mathrm{Abs}\left\{\left[\frac{\sqrt{p}}{.9945(1+i)}\right]\left[\frac{f_1[p]}{f_2[p]}\right]\right\}$$ <u>see the plot</u> next page.

<u>For $p=1$</u>, $Z_e = .3812 + i.2070)$, $R_{air} = 0.30$ and $R_{pvc} = .22$

$|Z_T| = |.30 + .22 + .3812 + i.2070| = 0.9247$; $R_{air} + R_{pvc} \overset{?}{=} .52$

$Z_T = .90 + i.20$

(Take the real part here) $\rightarrow |\lambda| = \dfrac{1}{(.87)(50)(.9247)} = .025$

$$\left|\frac{Z_e}{Z_T}\right| = \frac{.434}{.9247} = .47$$

$$(T_{air} - T_g)(t) = (T_{air} - T_g)(0)\, e^{-\lambda x}$$

different x here

x	$e^{-.025x}$	x 35°F (peak ~~max~~ $\Delta T = T_{air} - T_g$)
50	0.29	10.2°
100	.082	2.9°
150	.024	.84°
200	.007	.24°
250	.0019	

$Re\{\lambda\} = .024$

<u>Rate of change</u> $T = f(t) = A \cos \frac{2\pi t}{P}$; $f'(t) = -\frac{A 2\pi}{P} \sin \frac{2\pi t}{P}$

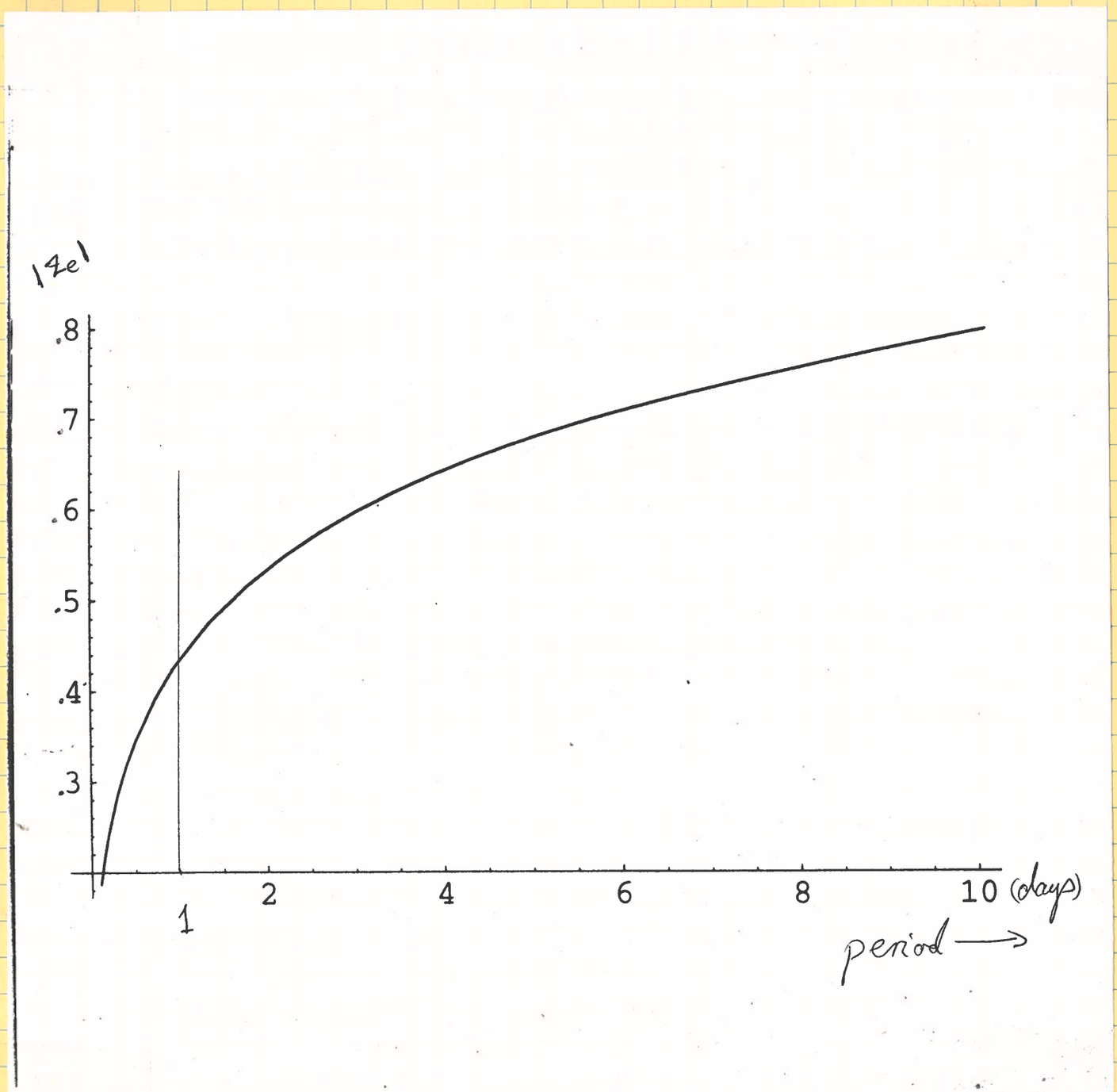
$\max f'(t) = \frac{A 2\pi}{P}$ $\qquad \max \Delta T = \frac{A 2\pi}{P} \Delta t = \frac{A 2\pi}{1} \frac{10\,\text{min}}{24 \cdot 60} = .04A$

The gain change in 10 minutes. For the shortest run, 80'; A = 10.
A = 10° ; .04A = <u>0.4°</u> If the T&C regulator has a gain of 10, we get .04°F fluctuation at the PAM

Cooling power $P(\text{watts}) = \cancel{.336}\ .26\,(CFM)\,\Delta T\,(°F)$? no $.34\,(CFM)\,\Delta T$

$= \cancel{420}\ 325$ watts for $\Delta T = 25\,°F$, $\cancel{390}$ 390 watts for $\Delta T = 30°F$

$= 588$ watts for $\Delta T = 35\,°F$



The $|Z_e|$ as a function of the period of the excitation. Variations with periods less than a day are shorted out. (mostly)

Heat Loads in the Alidade WJW

The azimuth drive

Input to the driver @ 120 volts, using a clip on ammeter

a) 3 amps slewing against a big load: 900 lbs [corr. to ≈35 mph wind]
b) .3 amps holding against the 900 lb load; ~~corr. to tracking in ≈35 mph wind.~~
c) 0.4-0.5 amps for tracking
d) ~1 amp for normal slew

powers for the loads.

a) 360 watts
b) 36 watts
c) 60 watts
d) 120 watts

Assume an overall efficiency for the drives of 20%
The elevation should be about the same: 60 watts for normal tracking
[No there is just one driver]

Totals

1. Normal Tracking: 120 watts x 0.8 = 100 watts heat
2. Normal Slewing: 240 watts x 0.8 = 190 watts heat
3. Heavy slew: 720 watts x 0.8 = 575 watts heat

Summary Normal tracking, 100 watts, goes to 200 watts for normal slewing. Expect up to 575 watts slewing ~~(at 35 mph~~ against a heavy wind (≈35 mph)

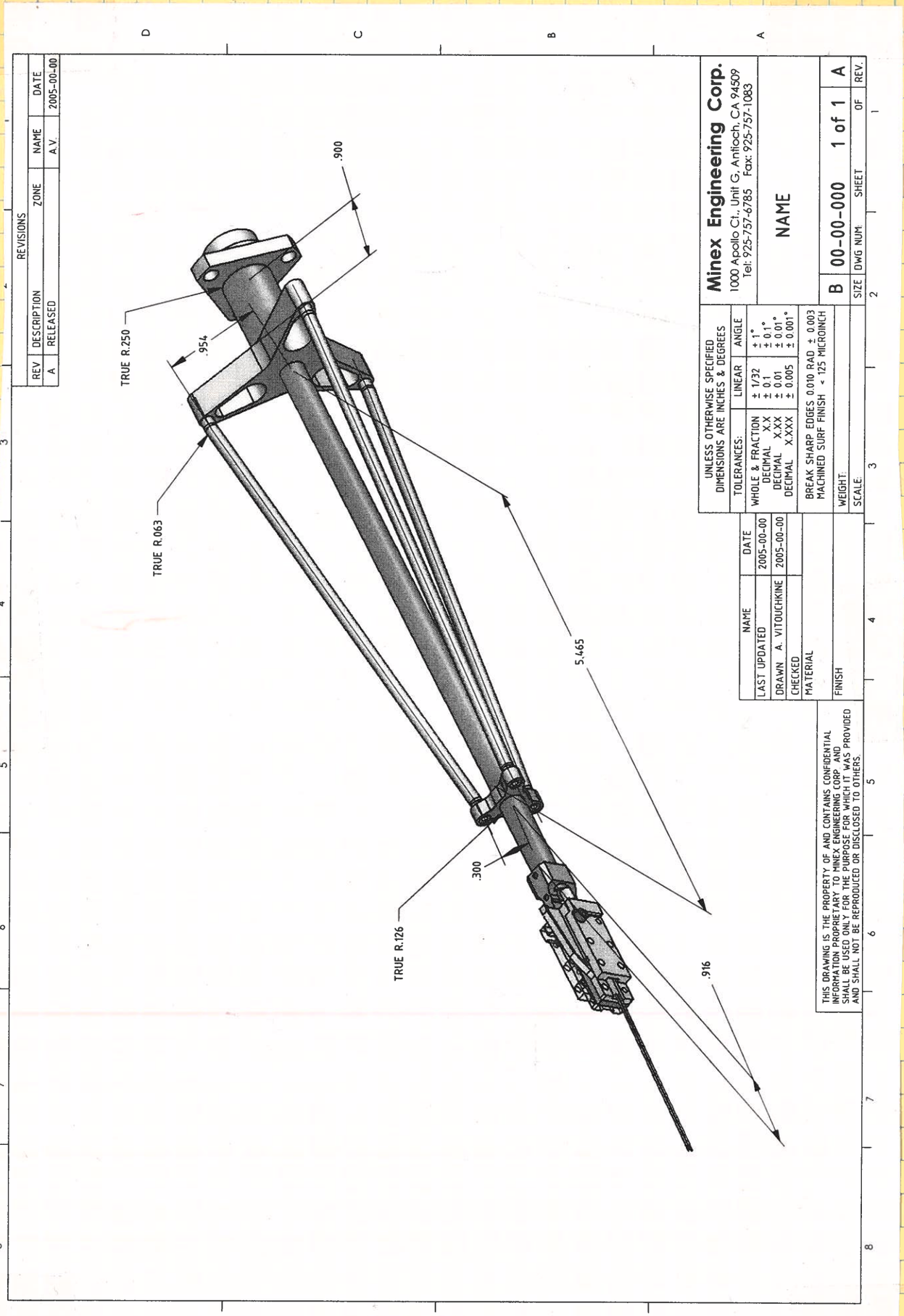
Pipe sizes in the alidade

Input 3" ID
to the front end 2.0"
exhaust: 0.5"
to elevation: 1.0"
to bar back of dish: 1.0"

Station	Total from top of pedestal to node entry
1B	205
1C	297
1D	218
1G	122
1H	179
2A	175
2C	163
2D	310
2F	167
2G	267
2H	249
2K	213
2L	132
3C	157
3D	65
3E	162
3F	229
3G	261
3H	252
3J	97
3L	214
4E	167
4G	235
4H	282
4J	182
5B	61
5C	49
5E	95
5G	104
5H	60
4F	191
4K	102
4L	61

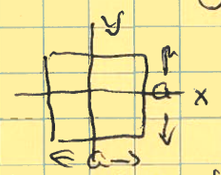
Pipe lengths from Node to Antenna.

Stiffness of the extended heat exchanger in the Dewar



Artom's Design

An approximate calculation of the gravitational Deflection of the tip of this cantelever beam can be made by assuming a beam of square crossection 1"x1" and either 5.5" or 8.5" length. This gives the deflection due to a 1 g acceleration.



$a = 1'' = 2.54\ cm = .0254\ m$

$Area = a^2 = (2.54)^2 = 6.45\ cm^2$

Weight of beam / unit length $\quad A \times 1 = 6.45\ cm^3$

Copper has $\rho = 8.96\ gm/cm^3$

Mass / unit length $= 8.96 \times 6.45 = 57.8\ gms/cm = W$

$gms \times 980 = dynes = mg$

$W = 57.8\ gms/cm \times 980 = 56{,}644\ dynes/cm \times 10^{-5} = 0.566\ Newtons/cm$
$= .000566\ KN/cm!$

Copper E modulus $\cong 20{,}000\ kip/in^2 = \frac{(20{,}000)(4.442)}{(2.54)^2} = 13{,}894\ kN/cm^2$

$L = 5.5'' = 13.97\ cm \quad$ or $\quad 8.5\ in = 22\ cm.$

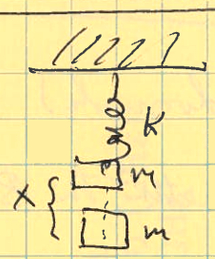
$$\boxed{\Delta_{max} = \frac{WL^4}{EI}} \qquad \propto \frac{W}{E} \cdot \rho \text{ cancels out}$$

$I_x = I_y = \frac{a^4}{12} = \frac{(2.54\ cm)^4}{12} = 3.47\ cm^4$; For $L = 5.5$ inches

$$\Delta_{max} = \frac{(.000566)(13.97)^4}{(13{,}894)(3.47)} = .00045\ cm = 4.5\ \mu m = \underline{.18\ mils}$$

For 22cm : $\Delta_{max} = \left(\frac{8.5}{5.5}\right)^4 (0.18) = \underline{\underline{1\ mil}}$

Direct Measurement using the resonant frequency



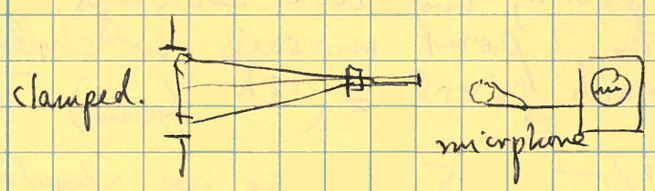
$$kx = mg \qquad \omega^2 = \frac{k}{m}$$

$$x = \frac{m}{k}g = \frac{g}{\omega^2} = \frac{(890)}{(2\pi)^2 f^2} = \frac{24.8}{f^2} \cong \left(\frac{5}{f}\right)^2$$

So $x = 1\,cm$ for $f = 5\,Hz$
$1\,mm$ for $f = 15\,Hz$

Generally $\quad x_{mm} = \left(\frac{15}{f}\right)^2$

The test of fixture at Minex



1" = 25 mm

Striking the end produced oscillations that the microphone detected. One problem: the side bar resonances drop close to that of the overall piece. Holding the bars damped their oscillation and an overall resonance of 300 Hz was found.

$$x_{mm} = \left(\frac{15}{300}\right)^2 = \frac{1}{400}\,mm = \frac{1}{(400)(2.5)} = 0.1 \text{ mils } 10^{-4}\,in$$

(The side bar resonances were close to 500 Hz)

0.1 mil deflection with one g acceleration

Try some observing for the Proposal Due in June.

One interesting possibility is M31 (Andromeda) in HI.

There is a map by Cram et al made with the Bonn 100m.
(AA Sup. 1980, 40, 215)
BW = 5MH, $T_{sys} \cong 60K$, $\theta = 8.8'$

We could do a 4 antenna synthesis over 12 hours and get about the same resolution. The Cram et al map is actually $1.2° \times 4°$, bigger than the ATA field, so we could do a mosaic: a 7 pt. hexagon will leave us some off space.

Relative sensitivity Area: $(100m/12m)^2 = 69$

They spent 4m on each pointing and took 200 days.
If we do 3 days with the 7 point mosaic, we should have much more time on each pixel: $\cong (12\ hour/4min) = 180$

times more time $69/\sqrt{180} = 5$ so we should be

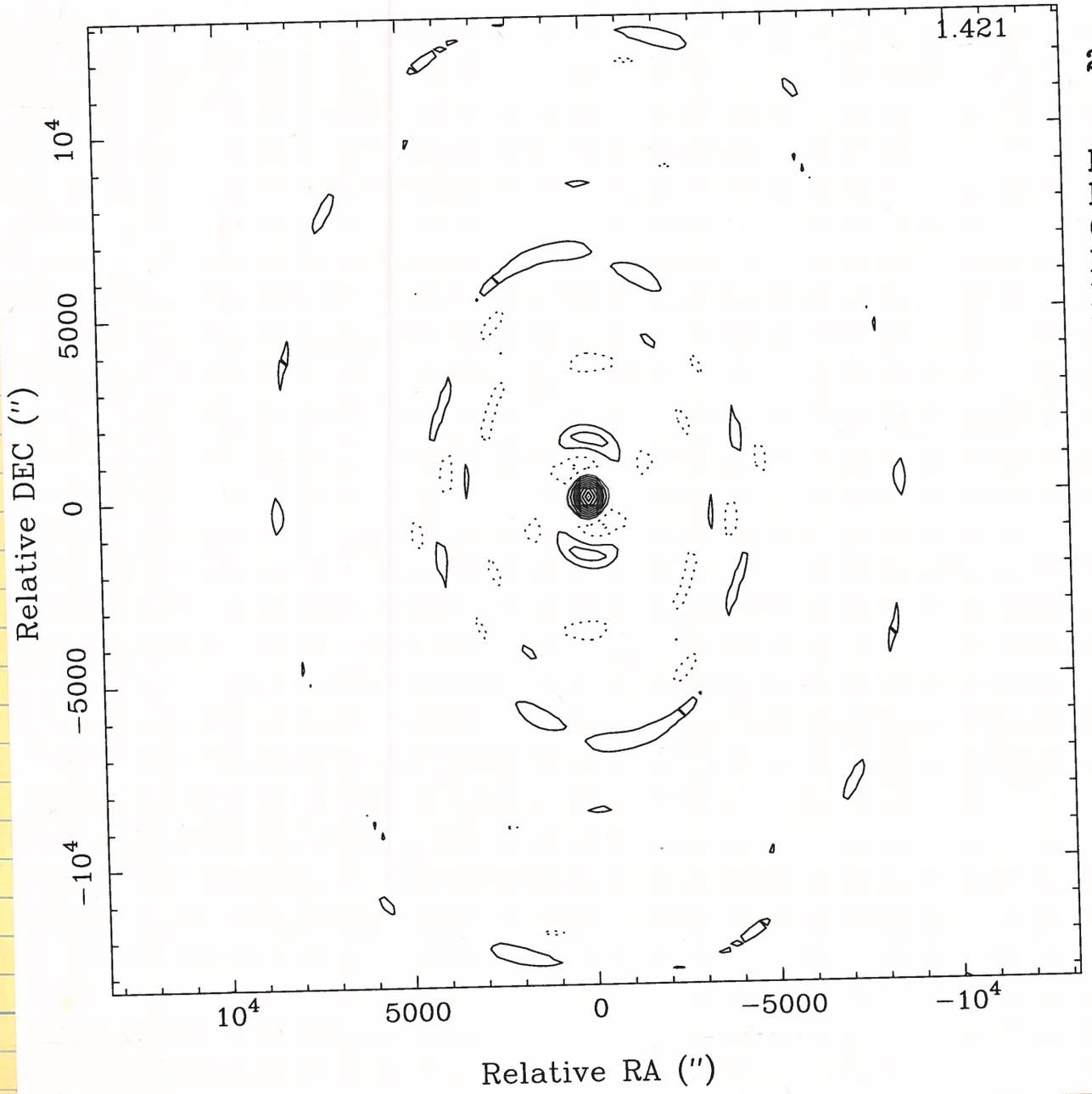
about 5 times noisier. Their $\Delta T_{rms} = .05K$ for each pixel
Ours should be $\sim 5 \times .05 = .25K$/pixel. Mel got the same result in his simulations.

Location of M31: $\begin{cases} \alpha = 00^h 40^m \\ \delta = +40° \end{cases}$

The sun on May 20 will be at $\alpha = 3^h 46^m$, $\delta = 20°$

Mel's beams for three different choices of the fourth antenna are on the next three pages. These are 12 hour syntheses. at $\alpha - 4h.400$ looks the best with lowest side lobes and best resolution: $9.5' \times 8.2'$ This is close to the 100m beam.

ata−4g.40



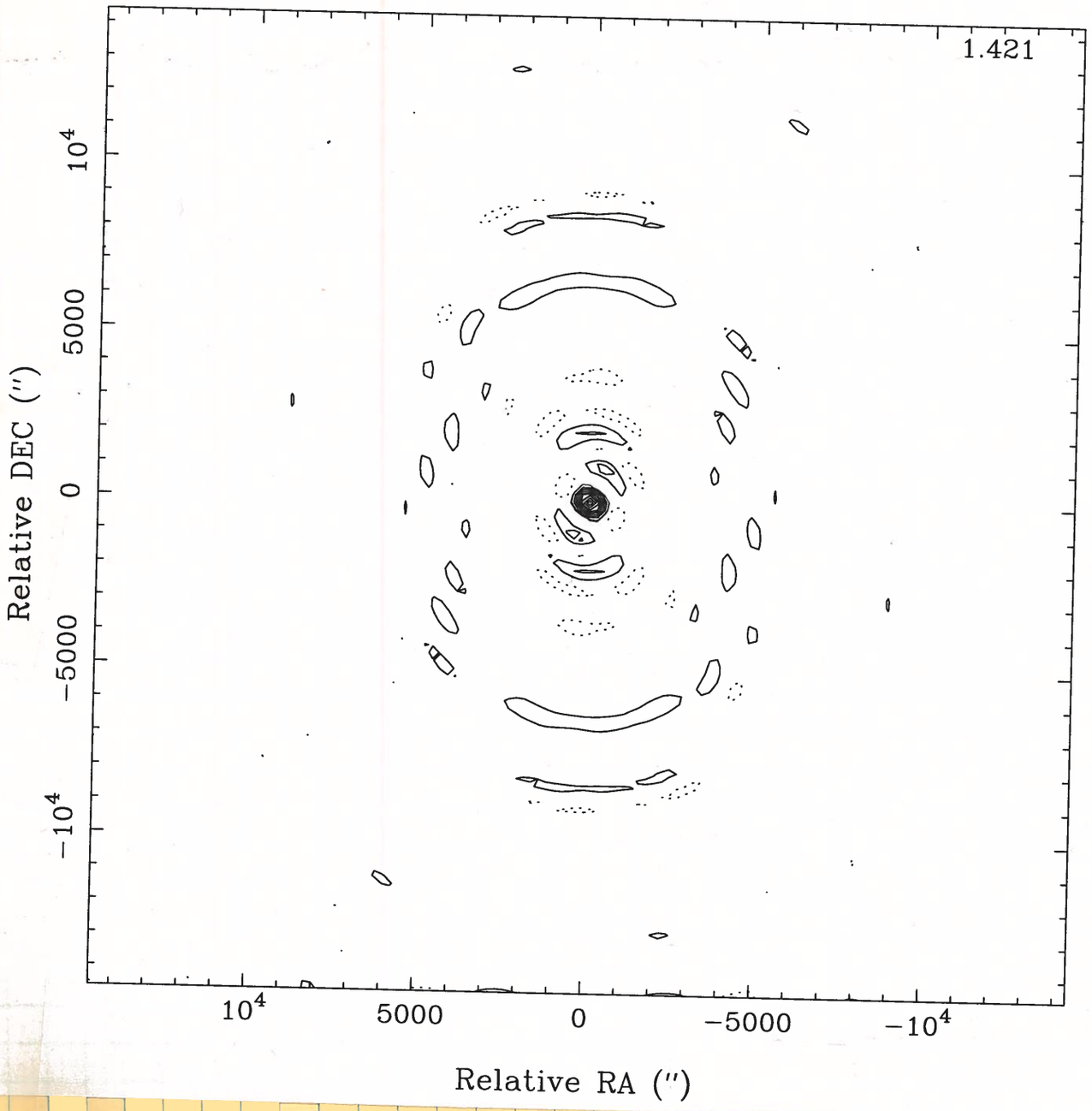
23:23:25.803 40:00:00.000

File: ata−4g.40.bm
Restfreq: 1.420000 (GHz)
Crval3: 1.4210 GHz
Max: 1.00000
Min: −0.221178
Units: JY/BEAM
Beam: 691.0 x 602.2
11.5 x 10

Axes: 128 x 128 x 1
−208.83 x 208.83 x 0.10

Contours: 10

0.100	0.200
0.300	0.400
0.500	0.600
0.700	0.800
0.900	1.000



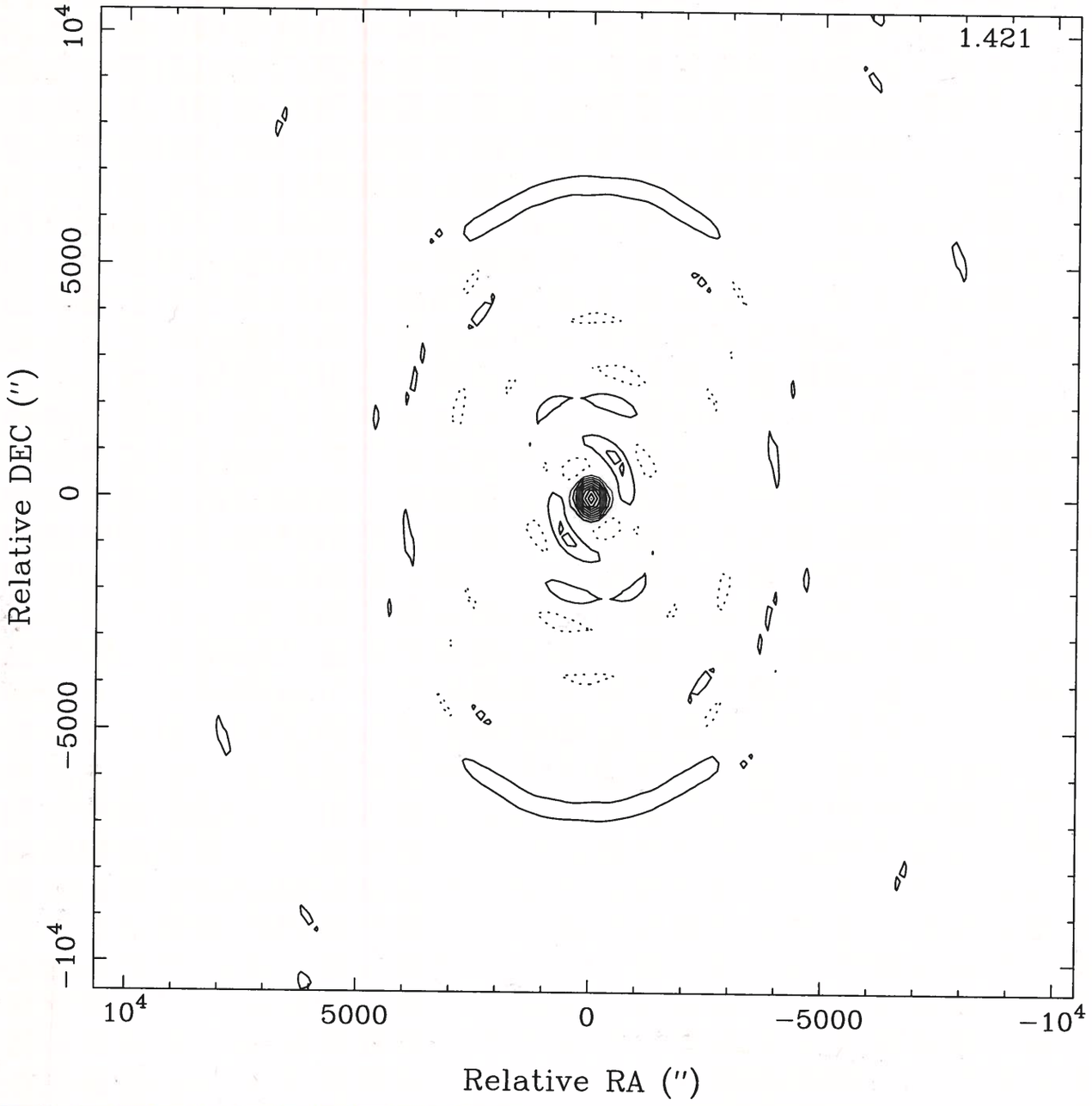
ata−4e.40

23:23:25.803 40:00:00.000

File: ata−4e.40.bm
Restfreq: 1.420000 (GHz)
Crval3: 1.4210 GHz
Max: 1.00000
Min: −0.193667
Units: JY/BEAM
Beam: 655.4 x 526.9
10.9 x 8.8
Axes: 128 x 128 x 1
−228.43 x 228.43 x 0.10

Contours: 10

0.100	0.200
0.300	0.400
0.500	0.600
0.700	0.800
0.900	1.000



ata−4h.40

23:23:25.803 40:00:00.000

File: ata−4h.40.bm
Restfreq: 1.420000 (GHz)
Crval3: 1.4210 GHz
Max: 1.00000
Min: −0.187622
Units: JY/BEAM
Beam: 571.1 x 489.7
9.5 8.2

Axes: 128 x 128 x 1
−166.44 x 166.44 x 0.10

Contours: 10

0.100	0.200
0.300	0.400
0.500	0.600
0.700	0.800
0.900	1.000

Equivalence of FX and XF correlators

How does the LO phase switching work in both cases?

$$C_{ij}(\tau) = \int_{-\infty}^{\infty} x_i(t)\, x_j(t+\tau)\, dt \qquad C_{ij}(\nu) = \int_{-\infty}^{\infty} C_{ij}(\tau) e^{-i2\pi\nu\tau} d\tau$$

For the individual signals

$$x_i(t) = \int_{-\infty}^{\infty} X_i(\nu) e^{i2\pi\nu t} d\nu \overset{\text{or}}{=} \int_{-\infty}^{\infty} x_i(\eta) e^{i2\pi\eta t} d\eta \; ; \; X_j(\nu) = \int_{-\infty}^{\infty} x_i(t) e^{-i2\pi\nu t} dt$$

$$x_j(t+\tau) = \int X_j(\eta) e^{i2\pi\eta(t+\tau)} d\eta$$

$$C_{ij}(\nu) = \int_{-\infty}^{\infty} \left\{ x_i(t) \int_{-\infty}^{\infty} X_j(\eta) e^{i2\pi\eta(t+\tau)} \right\} e^{-i2\pi\nu\tau} d\tau\, d\eta\, dt$$

$$= \int_{-\infty}^{\infty} \left\{ x_i(t) \int_{-\infty}^{\infty} X_j(\eta) e^{i2\pi\eta t} \underbrace{e^{i2\pi(\eta-\nu)\tau} d\tau} \right\} d\eta\, dt$$

$$= \int_{-\infty}^{\infty} \left\{ x_i(t) \int_{-\infty}^{\infty} X_j(\eta) e^{i2\pi\eta t} \delta(\eta-\nu) \right\} d\eta\, dt$$

$$= \int_{-\infty}^{\infty} \left\{ x_i(t) \; X_j(\nu) e^{i2\pi\nu t} \right\} dt = X_j(\nu) \int_{-\infty}^{\infty} x_i(t) e^{i2\pi\nu t} dt$$

$$= X_j(\nu) X_i(-\nu) \overset{\text{or}}{=} X_j(\nu) X_i^*(\nu)$$

In summary,

$$C_{ij}(\nu) = \int_{-\infty}^{\infty} C_{ij}(\tau) e^{-i2\pi\nu\tau} d\tau = X_j(\nu) X_i^*(\nu) = X_j(\nu) X_i(-\nu)$$

Skin depths and surface resistances of a few metals.

	σ (mho/m)	δ_{500MHz} (μm)	δ_{500MHz} (μin)	R_s (ohms)
Silver	6.2×10^7	2.8 μm	110	.005
Al	3.7×10^7	3.7	145	
Brass	1.6×10^7	5.7	224	
Solder	$.7\times10^7$	8.3	325	.017

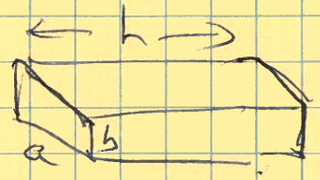
$\delta \propto \frac{1}{\sqrt{f}}$

$R_s \propto \sqrt{f}$

Tests of waveguide losses at 9 GHz in WR90 Guide that is silver plated.

From WRD p. 371:

$$\alpha_c = \frac{R_s}{b\eta\sqrt{1-\left(\frac{f_c}{f}\right)^2}}\left[1+\frac{2b}{a}\left(\frac{f_c}{f}\right)^2\right]$$



$a = 2.30\text{cm}$, $b = 1.05\text{cm}$. $\Rightarrow$ 2.30, 1.05 cm

$\eta = 377\Omega$

$$f_c = \frac{1}{2a\sqrt{\mu\epsilon}} = \frac{c}{2a} = \frac{3\times10^{10}\text{ cm/sec.}}{2(2.30)} = \frac{3\times10^{10}}{2(2.30)} = 6.521\text{ GHz}$$

$$\alpha_c = \frac{R_s}{(1.05)(377)\sqrt{1-\left(\frac{6.521}{9.00}\right)^2}}\left[1+.913\left(\frac{6.521}{9.00}\right)^2\right]$$

$$= \frac{R_s}{395.8(.6892)}\left[1+.4793\right] = R_s\,(.0054)$$

For Silver at room temperature $R_s = 2.52\times10^{-7}\sqrt{f}$

$= 2.52\times10^{-7}\sqrt{9\times10^9}$

$= .0239$.

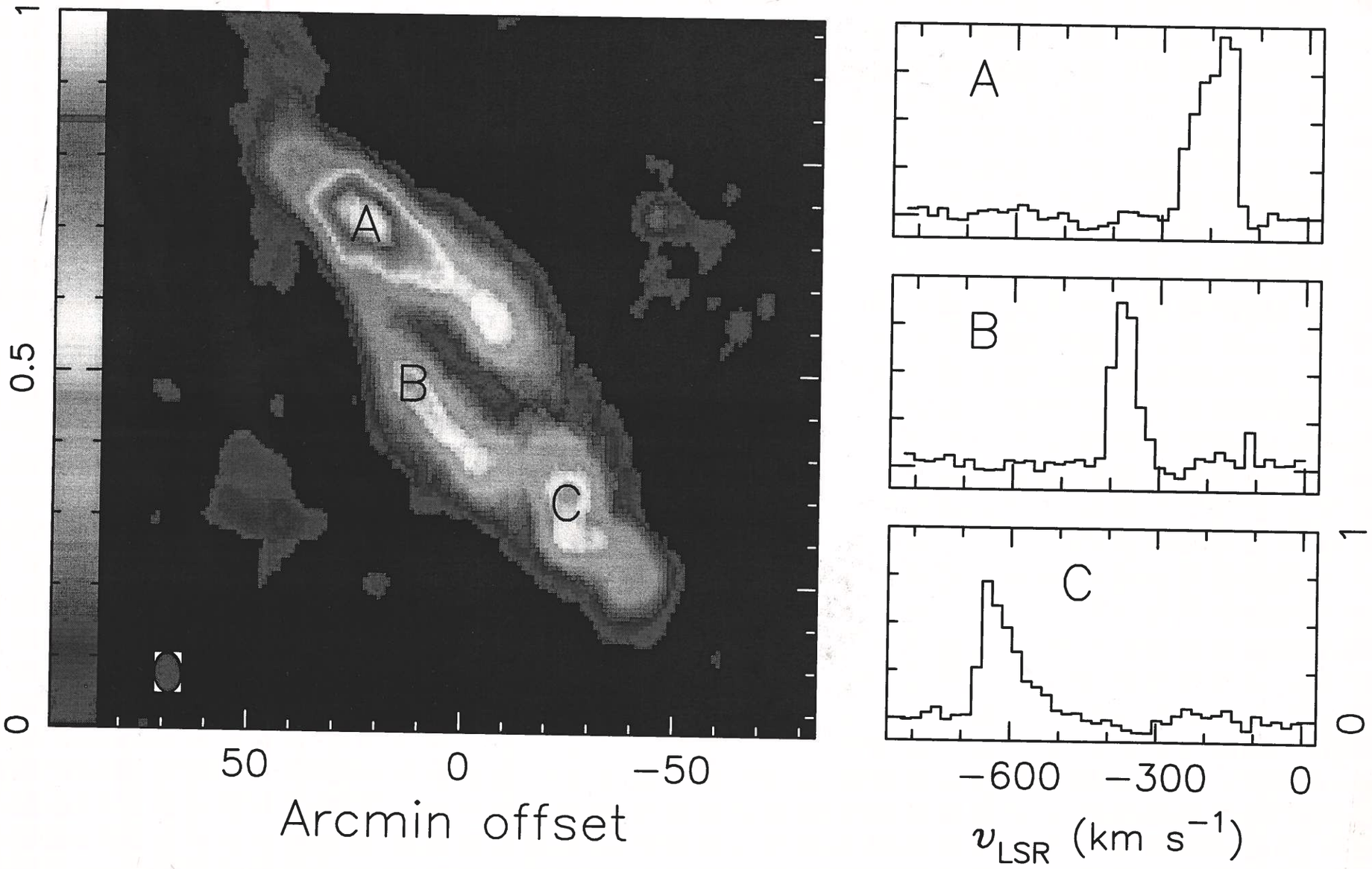
$\alpha_c = R_s(.0054) = (.0239)(.0054)\text{ cm}^{-1}$; $L = 8.00'' = 20.32\text{ cm}$

$\alpha_c L = (.0239)(.0054)(20.32) = .0026$

$e^{-\alpha_c L} \cong 1-\alpha L = 1.000 - .0026 = .9974$, $-.0114\text{db}$

For a normal trip $-.0228\text{ db}$

NB $e^{-\alpha_c L}$ is a voltage ratio, so the waveguide loss is $2\times(.0114) = .0228$ db one way and .0456 normal trip.



M31 in H I AVA-4

A little more on the heat flow from the buried PVC pipes between the node and antennas. from p. 74.

The solution can be written in terms of the Hankel function for the outgoing diffusion wave.

$$\frac{d^2T}{dr^2} + \frac{1}{r}\frac{dT}{dr} + \beta^2 T = 0$$

$T(r)$ is the temperature difference at any point from the value at $r \to \infty$. Thus it is also the mean of the surface temperature, so that $T(r,t) = \mathrm{Re}\{T(r,\omega)e^{i\omega t}\}$

$$T(r) = A\,H_0^{(2)}(\beta r) \qquad \text{where} \qquad H_0^{(2)} = J_0(\beta r) + i\,Y_0(\beta r)$$

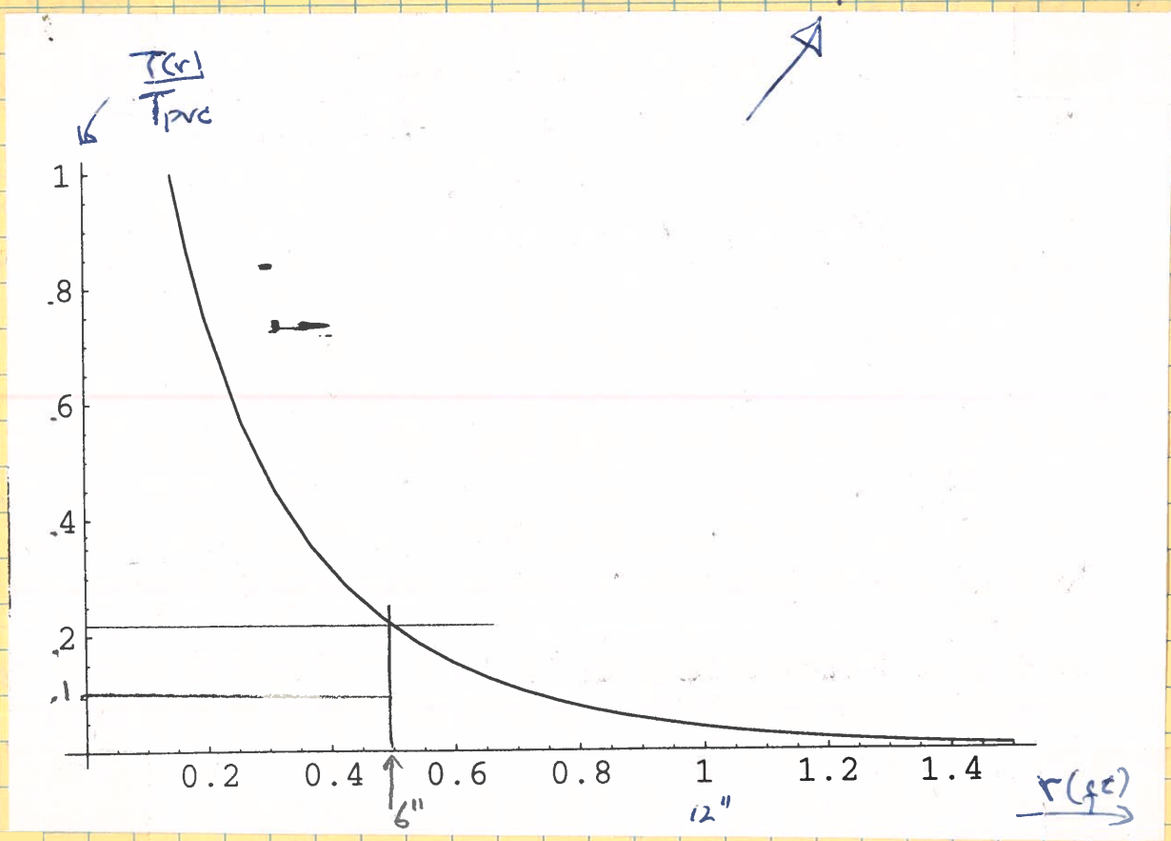
Boundary condition: $T(r) = T_{pvc}\dfrac{H_0^{(2)}(\beta r)}{H_0^{(2)}(\beta r_1)}$

From p. 1 with $p = 1$ day $\beta = (1+i)2.753$ ft^{-1} $= (1+i)33.04$ in^{-1}

$r_1 = 1.7'' = 0.1417$ ft ; $\beta r_1 = 0.390(1+i)$

This is the same for all distances along the pipe: Temp. eps

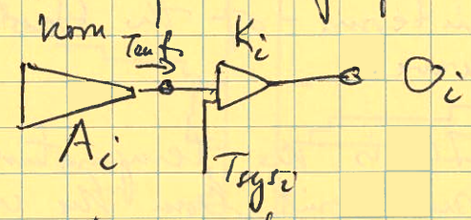
$T_{pvc}/T_{air} = .47$; so $\dfrac{T(r)}{T_{air}} = .47\,\dfrac{T(r)}{T_{pvc}}$ ⟸



Gain and Amplitude Calibration with Interferometry

One antenna plus amplifier and detector

Fig 1



$$G_i = \frac{4\pi}{\lambda^2} A_i$$

I. System temperature, T_{sys_i}, and correlator/AD converter scaling: K_i

$$O_i = O_i = K_i(T_{ant_i} + T_{sys_i})$$

Two input temperatures: T_{1i} and T_{2i}

$$O_{1i} = K_i(T_{1i} + T_{sys_i}) \qquad O_{2i} = K_i(T_{2i} + T_{sys_i})$$

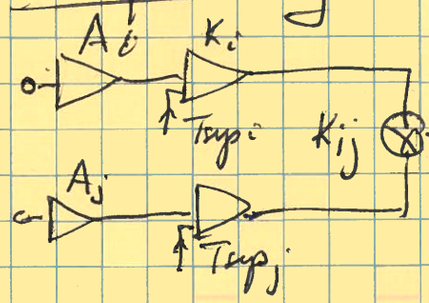
$$Y_i = \frac{O_{2i}}{O_{1i}} = \frac{T_{2i} + T_{sys_i}}{T_{1i} + T_{sys_i}} \quad ; \quad T_{2i} + T_{sys_i} = Y_i(T_{1i} + T_{sys_i})$$

$$\boxed{T_{sys_i} = \frac{Y_i T_{1i} - T_{2i}}{1 - Y_i}} \quad \text{and} \quad \boxed{K_i = \frac{O_{2i}}{T_{2i} + T_{sys_i}}} \quad ; KT_i = SA_i$$

A single antenna observation: $O_i = S K_i \frac{A_i}{K_i}$, zero remnant with a point source of flux S. (and all delays corrected)

II. Interferometry

Fig 2



$$O_{ij} = S\sqrt{A_i}\sqrt{K_i}\sqrt{A_j}\sqrt{K_j}$$

$$= S\sqrt{A_i}\sqrt{A_j}\,K_{ij} \;;\; K_{ij} = \sqrt{K_i K_j}$$

K_{ij} is known.

Let antenna 0 be the standard gain horn with known T_{sys_i}

Ratios

$$\frac{O_{01}O_{02}}{O_{12}} = \frac{\sqrt{A_0}\sqrt{A_1}\,K_{01}S\,\sqrt{A_0}\sqrt{A_2}\,S\,K_{02}}{\sqrt{A_1}\sqrt{A_2}\,K_{12}\,S} = A_0\underbrace{\left[\frac{K_{01}K_{02}}{K_{12}}\right]}_{K_0}S = A_0 K_0 S$$

A_0 is known and $\left[\frac{K_{01}K_{02}}{K_{12}}\right]$ known, so we measure S

It is important that we actually measure O_{10} and O_{12} which depends on $\sqrt{A_0 A_1}$ and $\sqrt{A_0 A_2}$ which have much higher sensitivities than just A_0 for the single dish measurement.

Another ratio is $\frac{O_{12}\, O_{13}}{O_{23}} = A_1 \left[\frac{k_{12}\, k_{13}}{k_{23}}\right] S = A_1 k_1 S$

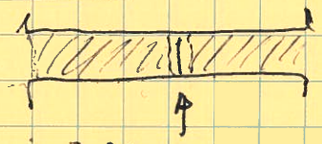
With S known, we get gain A_1 from this measurement.

Do all the correlations, including all the correlations with the standard horn A_0 to improve the S/N on S. From the latter measurement there is redundancy w.r.t. S. For the longer baselines, S will be less due to atmospheric decorrelation and we get a measure of the typical decorrelation. It is really A_i from the various baseline combinations that will show up smaller at the longer baselines. S is constant. This will provide the effective antenna gains for different baseline pairs and should allow accurate (decorrelation corrected) gains for the array. This should allow imaging with very accurate gains which have (mean) decorrelation corrections made for the source observations.

The calibrator with flux S needs to be constant only for the duration of the observation. Otherwise it may change from day to day.

Losses in some plastic materials @ 9 GHz

x-band waveguide



i) polyfoam in the guide in 2 pieces
ii) a piece of the plastic sheet is cut to fit the cross-section
iii) "through" measurements made w and w/o the sheet

I. White rubbery material intended to hold proc hoop

w/o	−.02 db	−.08	average −.093 db ± .01
w	−.10 db		
w/o	+.003	−.103	
w	−.093	−.096	2.16 %

II. painted sunbrella

w/o	+.015	−.128	average −.0915 db ± .02
w	−.113		
w/o	−.013	−.100	
w	−.073	−.060	2.13 %
w/o	+.005	−.078	

Waveguide losses in x-band guide from Jim Gibson.
(from p. 85)

3 pieces of guide, one is "pure" plating which looks matt.

~~All three together @ 8.5 GHz .0927 ± .007~~

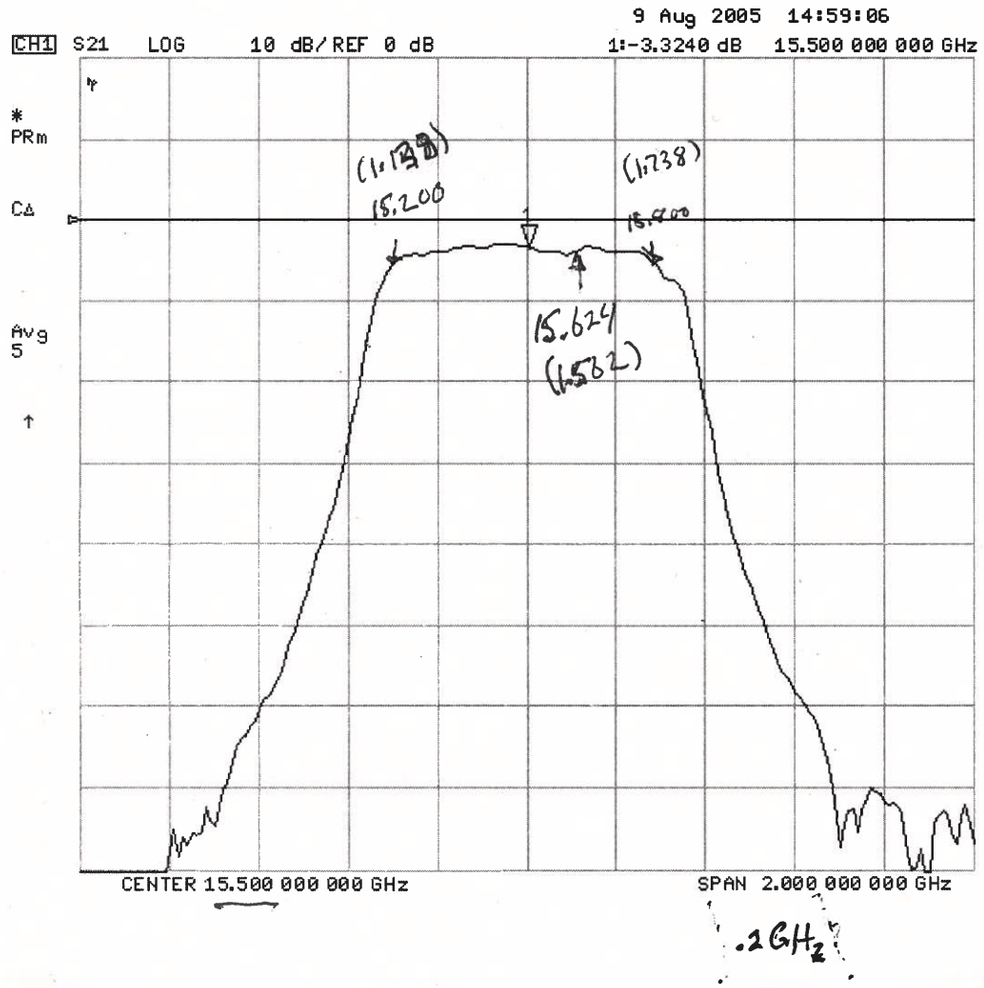
~~2 shiny ones .056 ± .01~~

~~matt finished one .056 ± .003~~

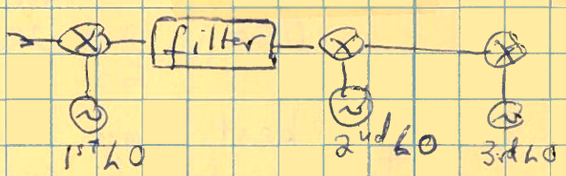
In view of the scatter, they are not very distinguishable.

matt	.04	± .01 db in each case
shiny	.03	
shiny	.04	

9/20/05



ATA oscillator frequencies



LO3 is 1.5624 GHz

LO1 puts a 100 MHz band from anywhere in the range 0.5 GHz – 11.2 GHz in the center of the above filter.

LO2 takes the above band filter and translates it to 1.5 GHz so that LO3 can provide a low pass digitized 100 MHz band.

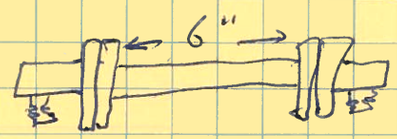
The lowest frequency filter is 0.500 – 0.600 GHz (400 MHz) and its center frequency is 0.550. The first LO translates this band to 15.500 GHz i.e. f_{LO1} = .550 + 15.500 = 16.050 GHz

For higher frequency bands LO1 is tuned to the corresponding higher frequency.

A filter for possible LO#1 excess side band noise.

The simplest filter would be a high pass filter which passes signals only above about 16.050 GHz. A waveguide beyond cutoff at the lower frequencies would work.

Tests of two back-to-back coax to waveguide transitions in X-band with a 6" length of waveguide joining them.



WR 90

According to Harvey the TE_{10} cutoff for WR90 is 6.56 GHz. The NA trace is smooth up to 12 GHz with no substructure and cuts off sharply near 6.56 GHz

f (GHz)	attenuation	fraction of cutoff, 6.56	Scaled to 16.050 GHz	Scaled to 16.174
6.20	−71 db (noise)			
6.30	−60	.9604		
6.40	−50	.9756	15.332	
6.50	−28 db	.9909	15.572	
→ 6.56	−15 db	1.000	15.715	15.837 ✓
6.63	−5.8	1.0107		16.006
⇝ 6.70	−3.8	1.0213	16.050	16.174
6.80	−2.7	1.0366		
7.00	−1.9	1.067		
7.50	−.28 db	1.14		

The best choice seems to be to put 16.050 at 1.0213, so that the formal cutoff guide frequency is 15.715 GHz. The frequency scalings and attenuations are shown above. This gives close to −30 db attenuation at the center of the 15.572 GHz Band Pass filter. The loss at 16.050 is −3.8 db which is reasonable.

A plot of the transmission up to 12 GHz is very smooth with less than −1 db loss maximum above 7.50 GHz

The final choice puts the BP filter center frequency at 15.624 and the 1.0213 point at 16.174 GHz. Now the formal cutoff waveguide frequency is at 15.837 GHz ✓.

$$a = \frac{\lambda_0}{2} = \frac{1}{2}\frac{c}{\nu} = \frac{2.99792 \times 10^{10}\ \text{cm/sec.}}{2[15.837 \times 10^{9}]} = \begin{array}{l} .3726\ \text{in} \\ \cancel{.9465}\ \text{cm} \\ .94649\ \text{cm.} \end{array}$$

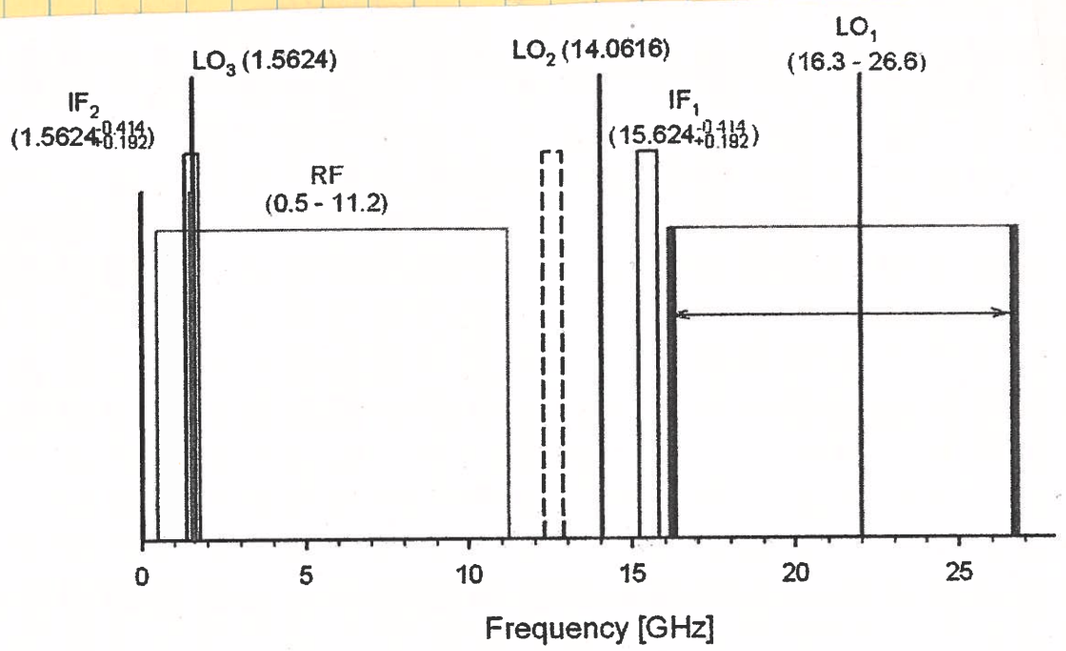
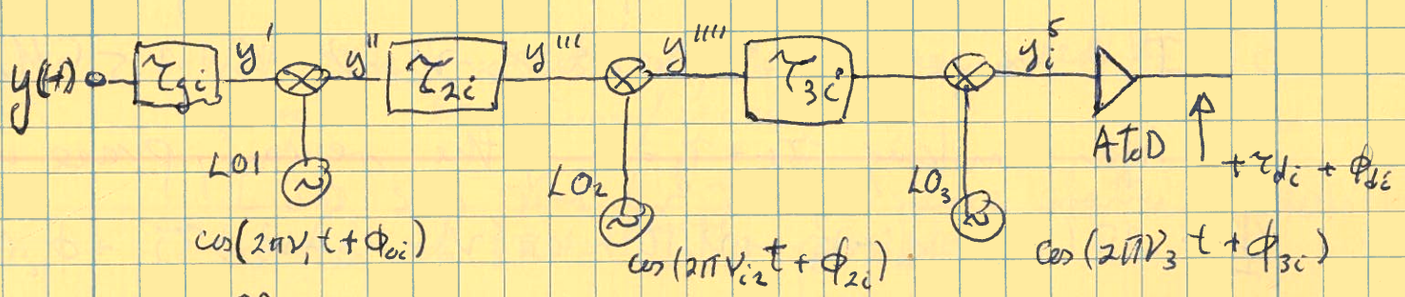


Figure 1: Bands and LO's of the RF converter board.

Phase switching of LO #2 with a Walsh function $W_i(t)$

e.g.:

$$W_i(t) \times \cos(2\pi\nu_2 t) = \cos\left(2\pi\nu_2 t + \frac{\pi}{2} W_i(t)\right)$$



$$y_i'(t) = \int_0^\infty A_i(\nu) \cos\left[2\pi\nu(t + \tau_{gi}) + \phi_i(\nu)\right] d\nu$$

; $A_i(\nu)$ and $\phi_i(\nu)$ are random variables.

At each mixer we keep only the difference frequency. The sum frequency gets filtered out. In addition, the signal passes through the band pass filter, which rejects the other sideband of LO_2. $F(\nu)$ is the BP filter.

Then

$$y_i^5(t) \propto \int_0^\infty A_i(\nu)\, F(\nu) \cos\left[2\pi(\nu_{1i} - \nu - \nu_{2i} - \nu_{3i})t - 2\pi\nu\tau_{gi} + \phi_{0i} - \phi_{1i} - \phi_{2i} + \phi_{3i}\right] d\nu$$

Now the signal is digitized after the third mixer and both digital delay τ_{di}, and digital phase ϕ_{di} are added as often as necessary. The third mixer is the down sampler.

$$X_d(t) \propto \int F(\nu)\, A_c(\nu) \cos\left[2\pi(\nu_{1i} - \nu - \nu_{2i} - \nu_{3i})(t + \tau_{di}) - 2\pi\nu\tau_{gi} + \phi_{di} - \phi_{0i} - \phi_{2i} + \phi_{3i}\right] d\nu$$

ϕ_{2i} is the Walsh function phase switch taken out by part of ϕ_{di}

ϕ_{0i} is the timing of the first LO

Part of ϕ_{2i} is the timing of the second LO.

ϕ_{3i} is the timing of the third LO

$\tau_{di} + \tau_{gi} = 0$ is required for τ_{di}

The terms τ_{2i} and τ_{3i} are common to all receivers and cancel out. (not included)

$\nu_{1i} - \nu_{2i} - \nu_{3i}$ is the fixed overall LO frequency

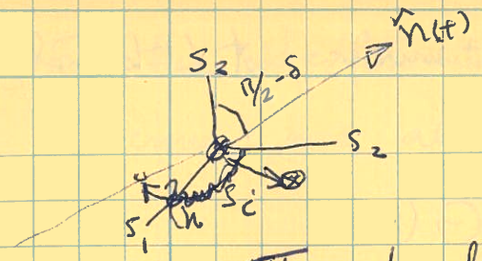
It's range is $0.5 \lesssim \nu_{1i} - \nu_{2i} - \nu_{3i} \lesssim 11.5\ GHz$

After making $\tau_{di} + \tau_g = 0$, the remaining phase is

$$\Phi = 2\pi(\nu_{1i} - \nu_{2i} - \nu_{3i} - \nu)t + 2\pi(\nu_{1i} - \nu_{2i} - \nu_{3i})\tau_g + \phi_{1i}(\nu) + \phi_{0i}$$

Because τ_g is changing in time, the second term is an offset frequency, an approximately steady change of phase in time that must be taken out with ϕ_{di}. It varies slowly.

The first term is common to both telescopes and drops out in the difference.



$$\tau_{g_i} = \frac{\hat{n}(t)\cdot\vec{S}_i}{c} \qquad \hat{S}_i = S_{1i}\hat{a}_i + S_{2i}\hat{a}_2 + S_{3i}\hat{a}_3$$

The unit vector $\hat{n}$ in the (h,δ) coordinate system is the radial unit vector $\hat{n} = \hat{a}_3 \sin\delta + \cos\delta\,[\hat{a}_1 \cos h + \hat{a}_2 \sin h]$

$$\tau_{g_i} = \frac{\hat{n}\cdot\vec{S}}{c} = \frac{1}{c}\left\{ S_{3i}\sin\delta + \cos\delta\left[S_{1i}\cos h - S_{2i}\sin h \right]\right\}$$

[Get $[h(t),\delta]$ from the pointing package. The aberration of starlight must be included.]

$$\tau_{d_i} = -\tau_{g_i}(t)$$

For a BW of 100 MHz, making a step every time the delay steps by 0.3 nsec gives a phase error accross the band of $\sim 12°$, a 2% error at the end.

For the phase: $2\pi(\nu_{1i} - \nu_{2i} - \nu_{3i})$ is maximum for the highest frequency filter (or band)

Just The fringe rate: $\nu_f = \frac{1}{2\pi}\frac{d\Phi}{dt} = (\nu_{1i} - \nu_{2i} - \nu_{3i})\frac{1}{c}\cos\delta\,[S_1 \dot{h}\sin h - S_2 \dot{h}\cos h]$

$$\text{Max } \nu_f \cong (\nu_{1i} - \nu_{2i} - \nu_{3i})\,\cancel{\times}\,\frac{1}{c} S_2 \dot{h}\,[\cos\delta]$$

@ ~10 GHz

$$\cong (11\times10^9)\,\frac{1}{3\times10^8\ \text{m/sec}}\,500\ \text{m}\,\frac{2\pi}{86{,}400} = 1.3\ \text{Hz}$$

With 8 bit resolution, this phase needs to be changed at up to 300 times/second!

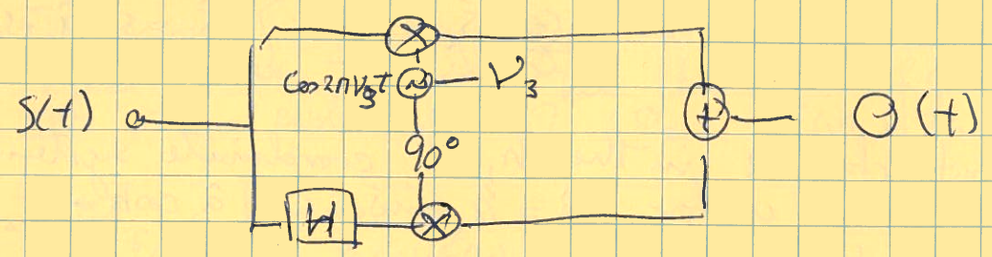
[JD says that for tracking space station updates needed at 1 KHz. It's possible.]

Probably up to 3 times faster for LEO satellites. i.e. ~ 1 KHz.

$\Rightarrow$ For LO = 1.4 GHz and $S_2 \sim 100$ m, $\text{Max } \nu_f = 1.4\times10^9\,\frac{(100)}{3\times10^8}\,\frac{2\pi}{86{,}400} = .034$ Hz

Current max $S_2 \sim 300$ m
$\begin{cases} \text{LO} = 1.4\ \text{GHz}: \nu_f \sim 0.1\ \text{Hz} \\ \text{LO} = 11\ \text{GHz}: \nu_f \sim 0.6\ \text{Hz} \end{cases}$

The Third LO Mixer is an SSB mixer at L-band.



H is the Hilbert Transform.

$$S(t) = \int_0^\infty A(\nu) \cos(2\pi\nu t + \phi)\, d\nu$$

$$\hat{S}(t) = \int_0^\infty A(\nu) \sin(2\pi\nu t + \phi)\, d\nu$$

$$\Theta(t) = \cos 2\pi\nu_3 t \; S(t) + \sin 2\pi\nu_3 t \; \hat{S}(t)$$

$$= \int_0^\infty A(\nu) \left\{ \cos 2\pi\nu_3 t \cos(2\pi\nu t + \phi) + \sin 2\pi\nu_3 t \sin 2\pi(\nu t + \phi) \right\} d\nu$$

$$= \int_0^\infty A(\nu) \cos\left[2\pi(\nu - \nu_3)t + \phi(\nu)\right] d\nu$$

Atmospheric Refractive Effects

$$\Delta E = \left\{ \underbrace{\frac{77.6\,P}{T}}_{\text{optical}} + \underbrace{373 \times 10^5 \frac{e}{T^2}}_{\text{radio}} \right\} \times 10^{-6} \tan z$$

pressures in mbar.

$$e = e_s(T) \frac{U}{100}$$

T	$e_s(T)$
283	12.3 mb
273	6.1
263	2.85
253	1.25
293	23.5

Use mean total radio

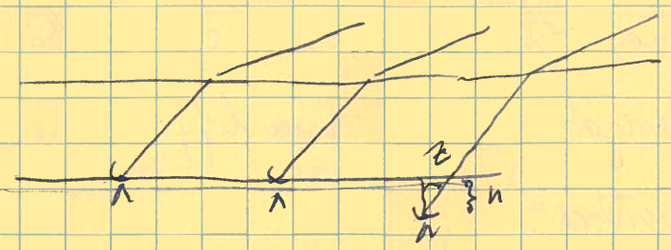
58"

For T = 293 and $e = \frac{1}{2} e_s$ and P = 890 mbar,

$\{287 \times 10^6\} = 57.6''$ and varies only a few arc seconds.

since $\Theta_{1/2}/100 = 12''$ @ 10 GHz, use the mean value.

For the interferometer there is no effect if the heights of the antennas are all ⊥ to the ground. The paths are all the same



There is an effect if one of the antennas is off the parallel.

$$\Delta\Phi_i = \frac{2\pi}{\lambda}\left\{287\times10^{-6}\,h\right\}\sec z.$$

For $h = 1$ meter and $\lambda = 3$ cm $\quad \Delta\Phi_c = \frac{2\pi}{3}\left\{287\times10^{-6}\,100\right\}\sec z$

$= 3.4^\circ \sec z$

At high zenith angles it is worse. e.g. @ $z = 85^\circ$, $\sec z = 3.8$

and $(3.8)3.4 = 13^\circ$

It may be marginal but perhaps should be included. Phase calibration will probably take it out.

{τ = 2 seconds for ATA-8, 8 secs for ATA-32}

Ignoring the fringe rate: $\quad$ max $\nu_f = .034$ at 1.4 GHz, $T = 2$ seconds for 8 correlator

Maximum error: $\nu_f = \nu_{LO}\frac{S_z}{c}h$

integration time: T $\quad \frac{1}{T}\int_0^T \cos 2\pi\nu_f t\,dt = \int_0^{2\pi\nu_f T} \cos x\,dx =$

$$= \frac{\sin 2\pi\nu_f T}{2\pi\nu_f T} \rightarrow \frac{\sin(2\pi)(.034)2}{2\pi(.034)2} = \frac{\sin(.4273)}{.4273} = \frac{.4144}{.4273} = .98$$

not bad and it could be corrected.

For $\nu_{LO} = 5$ GHz, max $\nu_f = .121$; $\frac{\sin 2\pi\nu_f T}{2\pi\nu_f T} = \frac{\sin 1.538}{1.538} = \frac{\sin 88^\circ}{1.538} = 0.65$ correctible

For an 8 sec. int. Time $\begin{cases} 1.4\text{ GHz} & .70 \quad \text{ok for HI} \\ 5\text{ GHz} & \text{wound up} \quad \text{but not this} \end{cases}$

Possibilities for orthogonal function switching (e.g. Walsh functions)

From each antenna two fibers carry the two pol's.

Put the two polarizations into two different up-down converters.

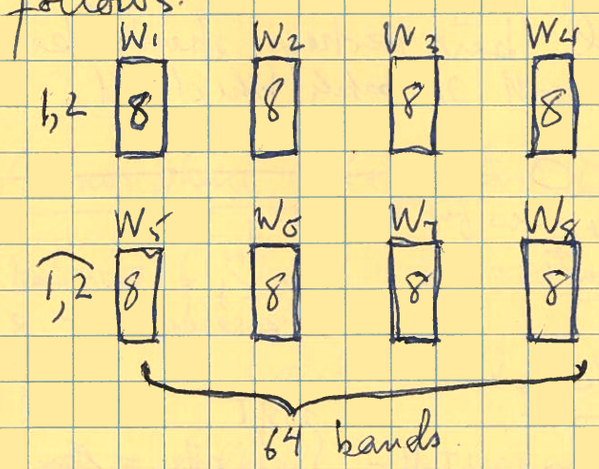
For the A/D converters:

One bin has 12 converter boards, each board has 4 converters

For reasonable isolation we need 4 different switching functions ~~for~~ on each board. Also, we need different functions on adjacent boards. This requires 8 different functions.

For the up/down converters each board gets a single function. Separating the polarizations into different boards will cut down on polarization cross talk.

Distribute the functions among the converter boards as follows.



for the ADC's: $12 \times 4 = 48$ for one bin, i.e., 48 bands.

$\frac{64}{48} = \frac{4}{3} = 1\frac{1}{3}$ bins for the 64 bands.

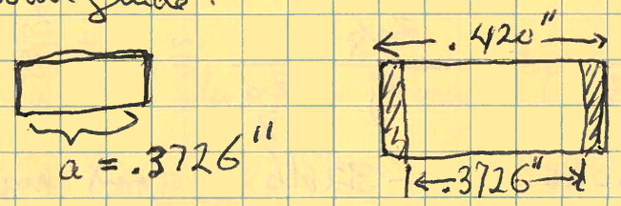
Since there are only 8 functions they can probably be done as square waves with periods that differ by factors of 2

A further simplification is that each polarization on each board gets a different switching function.

Standard waveguide cutoff frequencies

Band	width	height	TE_{10} cutoff	Band
WR 51	.510"	.255"	11.57 GHz	15.0 - 22.00
WR 42	.420"	.170"	14.08	18-26.5 (K band)
WR 34	.340	.170"	17.37	22.033
WR 28	.280	.140"	21.10	26.5-40

To get the filter cutoff at 15.837 GHz use K band coax/waveguide transitions connected to ~~the~~ length of K band guide whose width is reduced. A more complicated scheme would be to build new transitions for the narrower waveguide.



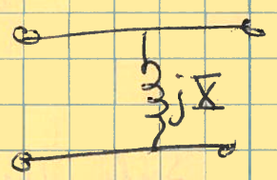
$$\alpha = \frac{a'}{a} = \frac{.3726}{.420} = .887$$

$$\beta = 1-\alpha = 0.113$$

Marcuvitz p. ~~168/169~~ 5.24 (p 296)

For $\beta = 1-\alpha \ll 1$. $(= .113)$

$\beta = .113$



$$\frac{Z_0}{X} \cong \frac{\lambda g}{2a}\left\{\frac{\beta^2(1+\beta)\ln\frac{2}{\beta}}{1-\beta/2}\right\}\left(1-\frac{27}{8}\,\frac{Q+Q'}{1+8\ln\frac{2}{\beta}}\right)$$

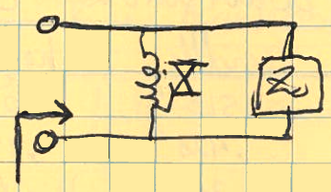
$$Q(\lambda) = 1-\sqrt{1-\left(\frac{2a}{3\lambda}\right)^2} \qquad Q'(\lambda) = 1-\sqrt{1-\left(\frac{2a'}{3\lambda}\right)^2}$$

$$\lambda g = \frac{\lambda}{\sqrt{1-\left(\frac{\lambda}{2a}\right)^2}} \qquad \left\{\frac{(.113)^2(1.113)\ln\left(\frac{.113}{2}\right)^{-1}}{\left(1-\frac{.113}{2}\right)}\right\} = \frac{.0408}{1-.0565} = .0432$$

$$\frac{Z_0}{X} \cong \frac{\lambda}{2a\sqrt{1-\left(\frac{\lambda}{2a}\right)^2}}\{.0432\}\left(1-3.375\,\frac{Q+Q'}{23.988}\right)$$

$$\frac{Z_0}{X} = \frac{.0432}{\sqrt{\left(\frac{2a}{\lambda}\right)^2-1}}\left\{1-.1407(Q+Q')\right\}$$

ν	X(cm)	$\frac{Z_0}{X}$
22	1.36	.034
20	1.50	.042
18	1.66	.052
17	1.765	.062



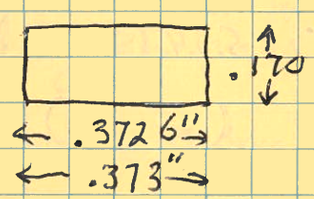
$$Z_L = \frac{jX Z_0}{jX + Z_0}$$

$$\Gamma = \frac{Z_L - Z_0}{Z_L + Z_0} = \frac{\frac{jXZ_0}{jX+Z_0} - Z_0}{\frac{jXZ_0}{jX+Z_0} + Z_0} = \frac{\cancel{jXZ_0} - \cancel{Z_0 jX} - Z_0^2}{jXZ_0 + Z_0 jX + Z_0^2} = \frac{-Z_0^2}{2jXZ_0 + Z_0^2}$$

$$\Gamma = \frac{-Z_0}{2jX + Z_0} = \frac{Z_0/X}{-2j - \frac{Z_0}{X}} \cong \frac{j}{2}\frac{Z_0}{X} \quad \text{for small } \frac{Z_0}{X}$$

For $\frac{Z_0}{X} = .05$, $|\Gamma| \cong .025$ $\quad -32\,db$ good match

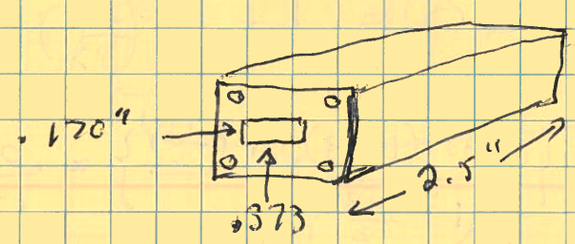
<u>Guide size</u>



$.3726 \cong \underline{.373''}$

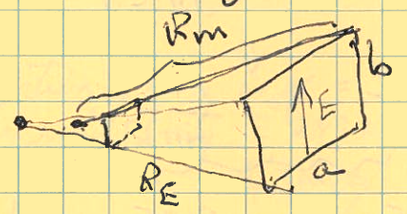
Scale the length to the x-band experiment

$\frac{6.56}{15.837} \times 6'' \cong \underline{2.50 \text{ inches}}$



Frequency Dependence of x-band horn gain.

$a = 4\,cm$ (3.95 cm)
$b = 2.90\,cm$
$R_E = 11\,cm.$
$R_M = 7.5\,cm.$



<u>Schelkunoff and Friis</u> let $\lambda = x$

$$g = \frac{8\pi R_E R_M}{ab}\left\{\left[C(u) - C(v)\right]^2 + \left[S(u) - S(v)\right]^2\right\} \cdot \left[C^2(w) + S^2(w)\right]$$

$C(u)$ and $S(u)$ are Fresnel Integrals.

$$u = \frac{1}{\sqrt{2}}\left(\frac{(\lambda R_m)^{1/2}}{a} + \frac{a}{(\lambda R_m)^{1/2}}\right)$$

$$v = \frac{1}{\sqrt{2}}\left(\frac{(\lambda R_m)^{1/2}}{a} - \frac{a}{(\lambda R_m)^{1/2}}\right)$$

$$w = \frac{b}{(2\lambda R_e)^{1/2}}$$

$$\frac{8\pi R_E R_M}{ab} = \frac{8\pi(11)(7.5)}{(3.95)(2.90)} = 181.00 \checkmark \quad ; \quad \frac{181.00\,\lambda^2}{4\pi} = 14.4\,\lambda^2$$

$$\frac{R_m^{1/2}}{a} = \frac{\sqrt{7.5}}{3.95} = 0.693 \checkmark \quad ; \quad \frac{a}{R_m^{1/2}} = 1.442 \checkmark$$

$$\frac{b}{\sqrt{R_E}} = \frac{2.90}{\sqrt{2.11}} = .618 \checkmark$$

$$u = .490\sqrt{\lambda} + 1.020\,\lambda^{-1/2} \checkmark \quad ; \quad v = .490\sqrt{\lambda} - 1.020\,\lambda^{-1/2} \checkmark$$

$$W = .618\,\lambda^{-1/2} \checkmark$$

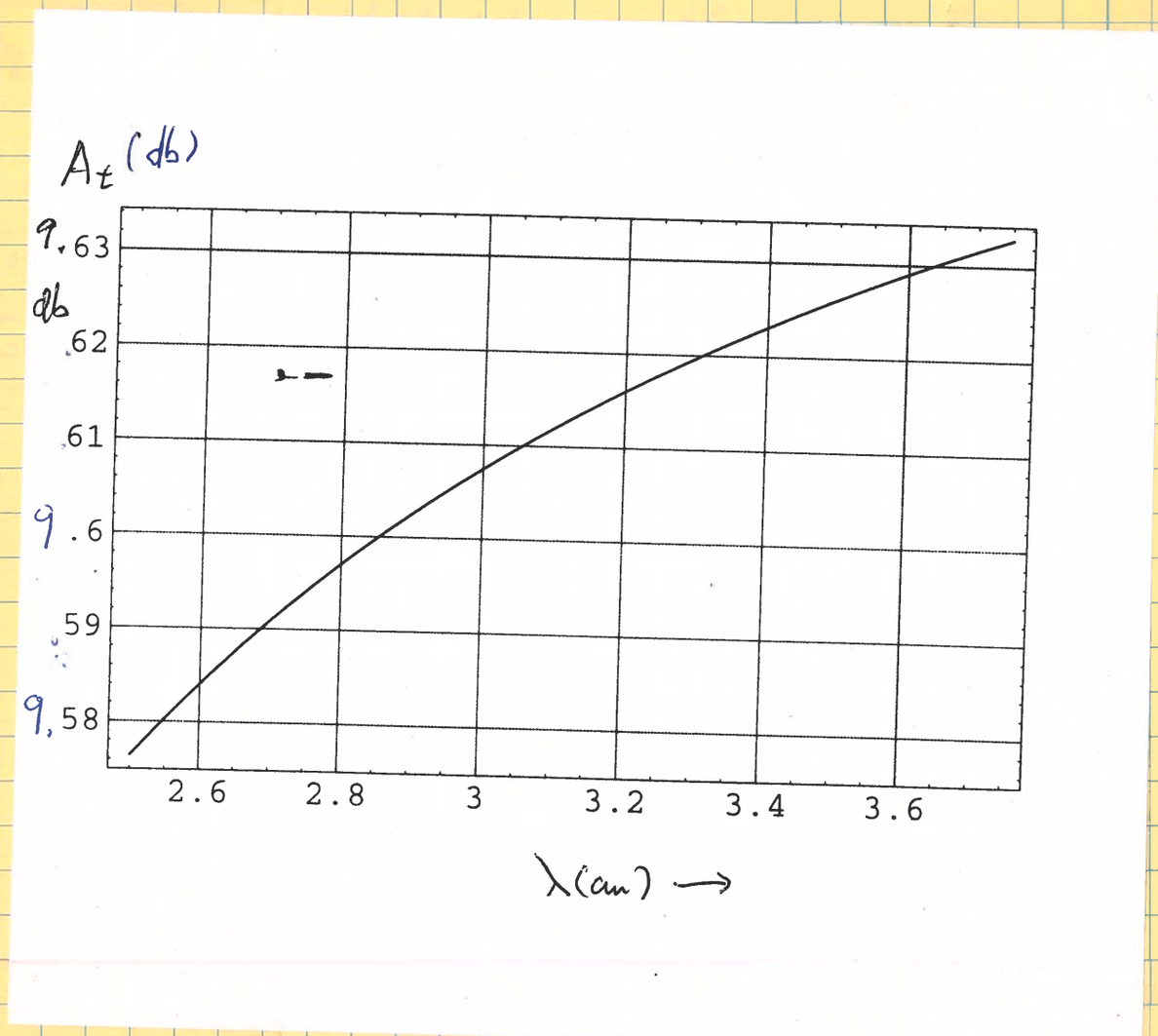
At 8 GHz this gives 8.211 (9.144 db)
notebook I p.128 gives a measured 9.02 + .243 = 9.26 db
transition loss
It agrees within 0.1 db

8 GHz → 3.75 cm
12 GHz → 2.5 cm

$$t = g_o g_t \left(\frac{\lambda}{4\pi R}\right)^2 \qquad \frac{4\pi A_t}{\lambda^2} = g_t$$

$$\text{so } t = g_o \frac{A_t}{4\pi R^2} \longleftarrow A_t = \frac{g_t \lambda^2}{4\pi}$$

For $\lambda = 3.75$, $A[3.75] = \underline{\underline{9.1894\,cm^2}}$ $10 \log [A[3.75]] = 9.6329$ in db.

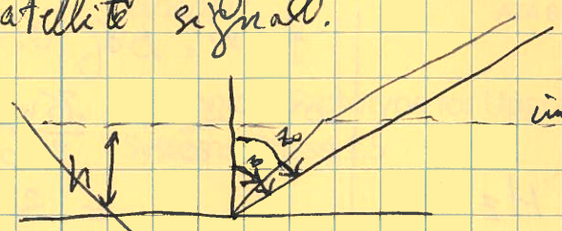


The total variation across the band is .05db (~ 1%)

$A_u(3.75) = 9.187$ $\quad A_u(2.5) = 9.068\ cm^2$

checked ✓

The effect of refraction on the Doppler shift of the GPS satellite signal.



index of refraction: $n = 1 + N$

N is the refractivity $\cong 297 \times 10^{-6}$ (see p. 98)

$\Delta z = z - z_0 = 1' \tan z$

The phase difference between the unrefracted and refracted paths.

$$\Delta\phi \cong \frac{2\pi}{\lambda_0}\left[(1+N)\,h\sec z - h\sec z_0\right] = \frac{2\pi}{\lambda_0}\left\{h(\sec z - \sec z_0) + Nh\sec z\right\}$$

$$v_{eff} = \frac{1}{2\pi}\frac{d\Delta\Phi}{dt}$$

errors here

For the first term $\frac{1}{2\pi}\frac{d\Delta\Phi}{dt}$

$$h(\sec z - \sec z_0) = h\underbrace{(\sec z - \sec z_0)}_{\Delta z}\Delta z = h\frac{d}{dz}\sec z\,\Delta z = h\sec z\tan z\,\Delta z$$

$$= h\sec z\tan^2 z\;1' = h\sec z\tan^2 z\;(.0003\text{ rad})$$

$$\frac{1}{2\pi}\frac{d\Delta\Phi}{dt} = \frac{1}{\lambda_0}h(.0003)\frac{d}{dz}(\sec z\tan^2 z)\frac{dz}{dt}$$

$$= \frac{h(.0003)}{\lambda_0}\left\{\frac{2\cos^2 z\sin z + 3\sin^3 z}{\cos^4 z}\right\}\dot{z}$$

For the second term

$$\frac{1}{2\pi}\frac{d\Delta\Phi}{dt} = \frac{1}{\lambda_0}Nh\frac{d}{dz}\sec z\frac{dz}{dt} = \frac{Nh}{\lambda_0}\sec z\tan z\,\dot{z}$$

The GPS satellite is 20,000 km above the earth's surface. The earth's radius is 6400 km. The GPS orbital velocity is

$$\Rightarrow \quad v_R = \left[\frac{GM}{R}\right]^{1/2} = \left[\frac{7\times10^{-8}\;6\times10^{27}}{2.65\times10^{9}}\right]^{1/2} = 3.98\text{ km/sec.}$$

As seen from the surface of the earth $\dot{z} \cong \frac{3.98}{20{,}000} \sim 2\times10^{-4}$

For the first term at $z = 70°$ with $h = 10\,km = 10^6\,cm$ and $\lambda_0 = 18\,cm$.

$$\nu_{off} = 0.08\,Hz$$

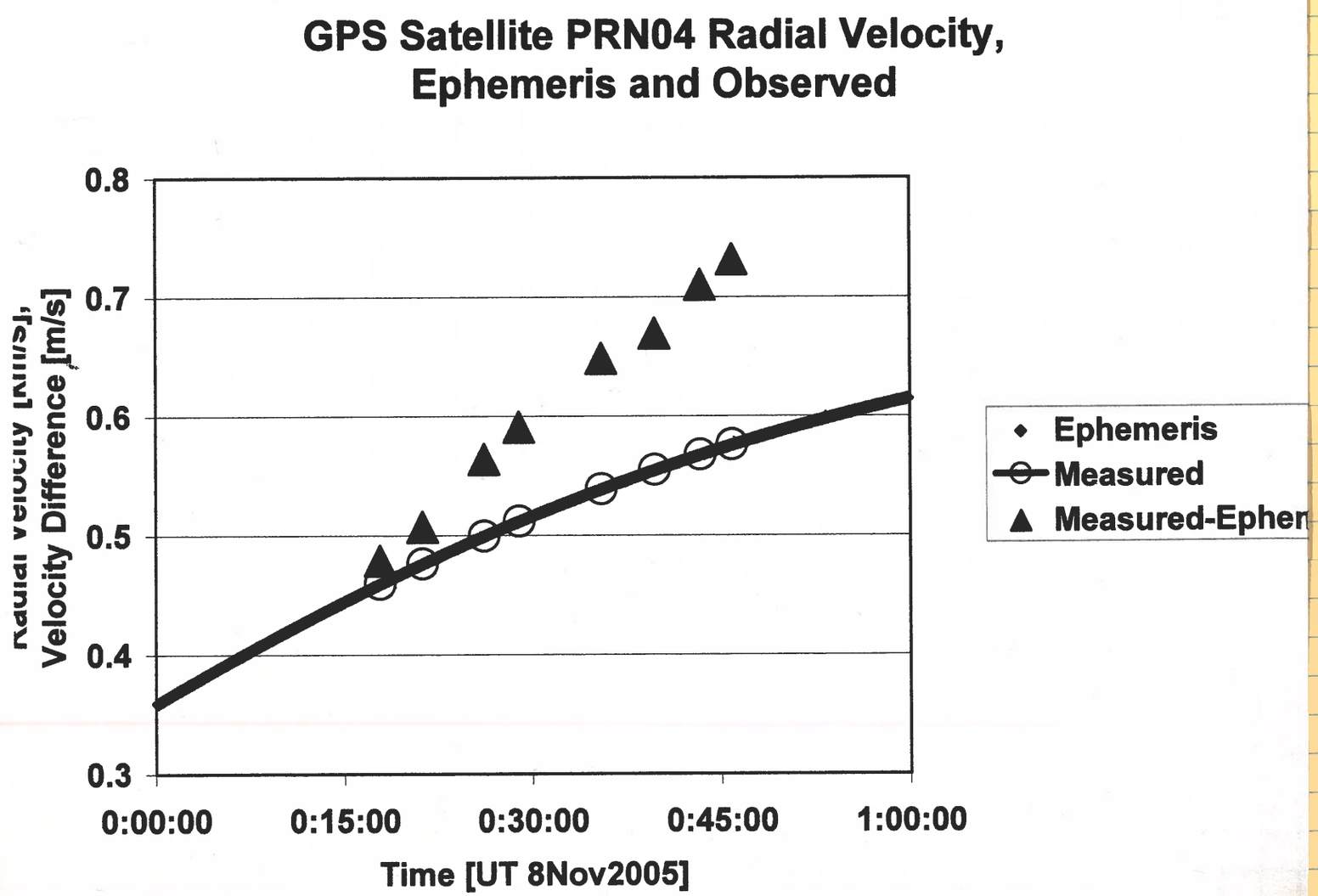
For the second term at $z = 70°$

$$\nu_{off} = .02\,Hz$$

Below: Δv_f 0.15 m/sec.

$$\Delta\nu = \frac{\Delta v}{c}\nu = \frac{.75\,m/sec}{3\times10^8} \times 1.5\times10^9$$

$$= 3.75\,Hz$$



Technology Prototype for Upgrade of ATA Beamformers and Detectors Systems, Phase 1.5

Summary

This proposal outlines a 3 month project, planned for the continued development of new digital back ends for the Allen Telescope Array, informally known as the Antonio digital project.

Project Description

In a recent effort, the SETI Institute evaluated updated, commercially available hardware that could form the basis of a new digital backend system at the Allen Telescope Array (ATA). It was successfully demonstrated that we can take advantage of nearly the full bandwidth of a 40 Gb/s Ethernet link for data transfer between two computers. Subsequent digital processing (correlator, beamformer, beamformer calibration, high speed data stream for transient (FRB) analysis and SETI signal detection) is to be performed on a single server to minimize inefficiencies of data copying. A 64-core Xeon Phi-based computer was tested.

We ported our SETI algorithms (SonATA) to the Xeon Phi with little difficulty, thanks to the fact that Intel supports our previous FFT library (FFTW 3) with a plug-in replacement. Between the Ethernet input and SonATA, we developed beamformer and correlator software on the Xeon Phi computer. This part of the project was a limited success. The correlator software, necessary for conventional astronomy and for beamformer calibration, ran at about 20% of the speed required for our needs.

In this short proposal, we continue the development process with a special focus on the correlator software, but also in testing and integrating one new analog to digital converter (SNAP) into the prototype SETI signal detection system, with 370 MHz RF bandwidth.[1] Increased correlator efficiency is pursued with the aid of a specialist Xeon Phi consultant, Erik Olson, supported at 15 hours/week for the duration of the project. Additionally, we continue to integrate and test all the components of our prototype backend system to see if any component (such as correlator or digitizer) presents further difficulty for a functional system.

This three-month, fixed-term effort will encompass the following efforts:

1) Evaluate analog filter to allow higher bandwidth sampling. Replace the 570-770 MHz bandpass filter at the output of the down converter with 450-900 MHz bandpass (Figure 1). The digitizer that follows this filter runs at 900 MS/s and samples baseband RF between 0-450 MHz, effectively sampling the filter output as -1 alias. The -2 and zeroth aliases also contribute significantly to the sampled data. We define the "useful band" as the part of the frequency spectrum where the -2 and zeroth aliases contribute less than 1%.

[1] The bandwidth 370 MHz is chosen to fit well within the sampler capabilities without stress. This is the design goal the "final" configuration of the RF and SNAP layout for the next generation system: four, 370 MHz IF's, independently tunable over 1-10 GHz.

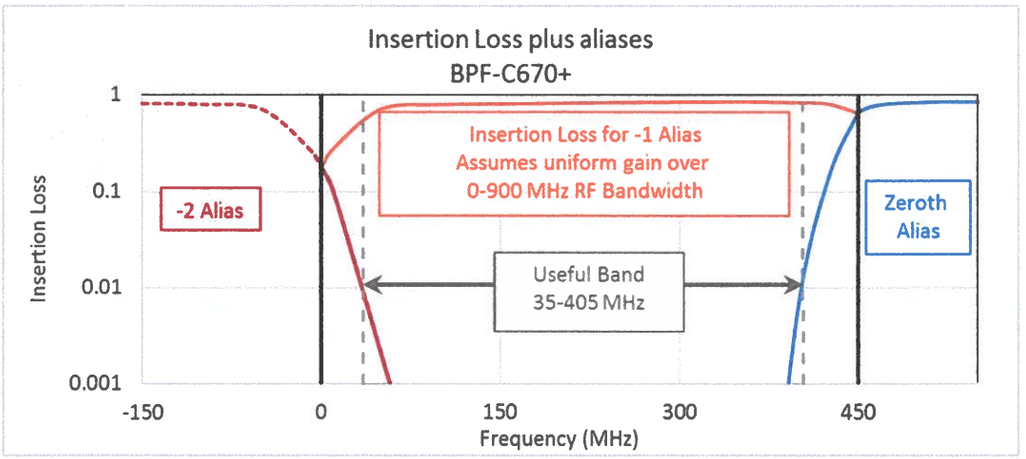


Figure 1: Diagram showing the bandpass filter insertion loss (zeroth alias), plus its -1 and -2 aliases. The digitizer samples baseband between the black lines, 0-450 MHz. We define the useful band as the frequency range where adjacent aliases contribute no more than 1% to sampled power. This figure refers to the filter only, and does not include other sources of gain variation.

2) In close collaboration with developers at Breakthrough Listen at UC Berkeley who have been providing ongoing support, install and test a new analog to digital converter (SNAP) designed by the CASPER group. Breakthrough Listen has agreed to loan us a SNAP board at no charge for an indefinite period to test it. The goals of this work are to program the SNAP to operate at 900 million 8-bit samples per second (MS/s), with a design bandwidth of 370 MHz (not including guard bands).
3) Because this sampler uses four interleaved 225 MS/s samplers to simulate a 900 MHz ADC, it introduces spurs into the sampled power spectrum. We will perform quantitative tests to determine how stable is the uneven gain across the four samplers and to what extent this can be corrected in post processing. If we find that the quad samplers on the SNAP are insufficiently stable, we believe that a different ADC chip can be used in the boards fabricated for the ATA.
4) The SNAP is designed to sample at baseband. We will test the configuration where the analog IF runs with a center frequency of 630 MHz and use digital aliasing to obtain baseband sampling centered at 270 MHz. While sound in theory, we do not know how well IF signals above 500 MHz propagate through the input circuitry of the SNAP. We will measure the signal to noise degradation caused by variable input gain across the 450-900 MS/s bandwidth of the IF. Sample aperture jitter is another source of error that is magnified when aliasing from IF 630 to 270 MHz. We will evaluate the overall performance of the SNAP for use at the ATA. Jack Welch suggested that new SNAP boards should be made with a different substrate more appropriate for IF frequencies 450-900 MHz. The manufacturer (Digicom) has indicated that this change can be made without significant difficulty.
5) One advantage of the SNAP digitizers is the availability of substantial digital processing capability in an integrated FPGA chip. This allows a first stage of channelization (separation into frequency bins) of data prior to output on the network. An initial design of the SNAP firmware is laid out in Figure 1.

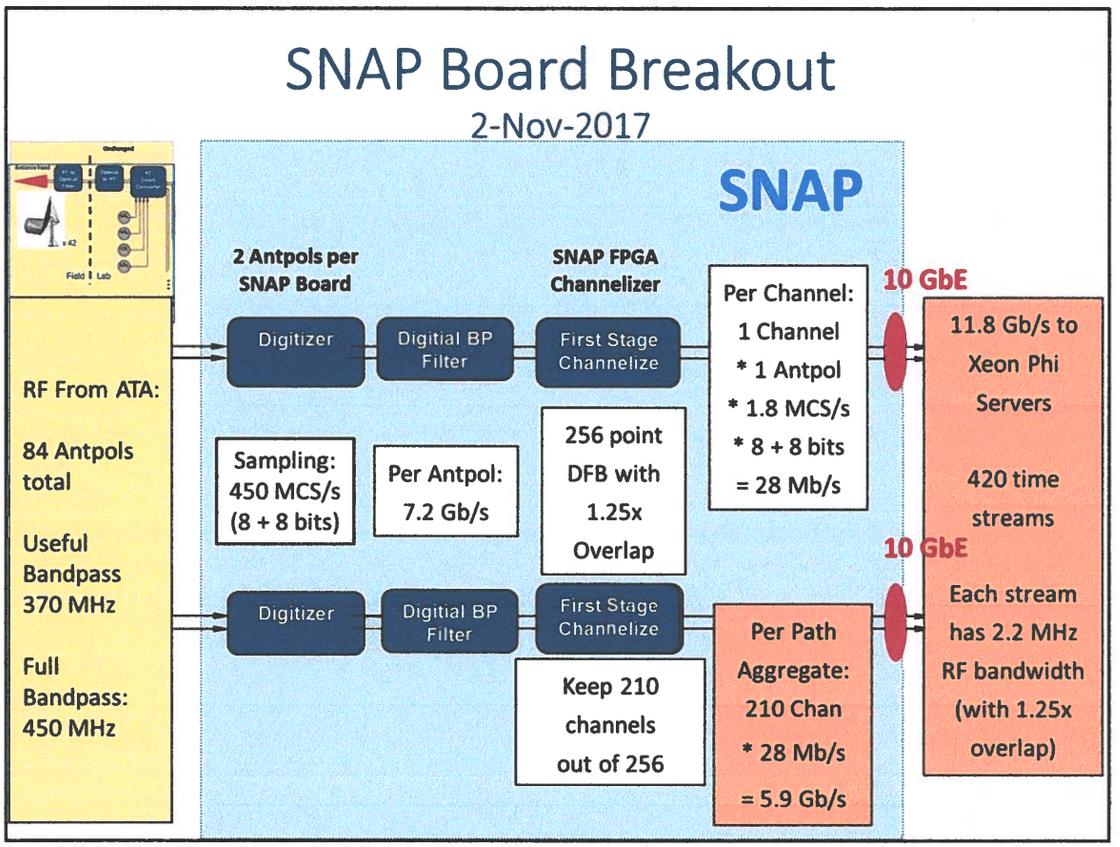


Figure 2: Initial design of SNAP board internals. Each SNAP takes in two antpols at full IF bandwidth and generates two 5.9 Gb/s streams on each of two 10 gigabit Ethernet outputs.

6) A local Xeon Phi consultant, Erik Olson, has been identified with substantial experience programming Xeon Phi Knight's Corner chips. His primary goal is to optimize the prototype correlator code such that it can operate with at least 20 CMS/s bandwidth for 84 input streams (including both polarizations) and full cross correlation. If this goal is achieved with time to spare, Erik will look at first the beamformer code and then the SETI processing code to improve performance. Comparisons will be made with a lightly modified version of the same code running on an Intel I7. Additional Xeon Phi support is provided by Intel, thanks to Jan Hoscan, VP at Intel and a member of the SETI Institute's Board of Directors. We have been given access to support from the top engineers that have been developing Xeon Phi and running highly efficient codes for the DoD. Intel has already generously provided about 20 hours of very helpful engineering time in support of Phase I of this project. Intel has indicated that they will continue to support our Xeon Phi development at a similar level in Phase 1.5.
7) Improve (and refactor) the end-to-end integration of Phase I software including: reception of data at a 40 Gb Ethernet port, casting int8 format to float32, correlators and beamformers followed by SETI analysis.

We note that this work will be carried out in the context of the day-to-day duties of Richards and Harp, including SETI observations, data analysis, ATA maintenance, proposal writing, etc.

Deliverables

At the end of this project we will provide all source codes for review as well as relevant benchmarks for the end-to-end system and all its components. These results will be communicated via email and in bi-weekly teleconferences. At the end of the project, a written report will summarize our results and make suggestions for the next phases of development.

The budget includes support for attendance at "AT-RASC 2018," - the 2nd URSI Atlantic Radio Science Meeting, May 28 - June 1, Gran Canaria, Spain. Gerry Harp will present a presentation/paper on the architecture of the new ATA backend system, with results showing the performance of the digitizer, and the computational processing speed that we attain. This conference is bi-annual, and many other researchers in the SETI community are planning to attend. This is the most important conference of the year for SETI science and URSI encourages papers on new hardware. The budget also covers travel expenses to/from the ATA for staff and consultants. The results of this work will be detailed in a final report prepared for Franklin Antonio for the purpose of evaluating the merits of a Phase II implementation of new digital back ends for the complete telescope array.

Budget Summary

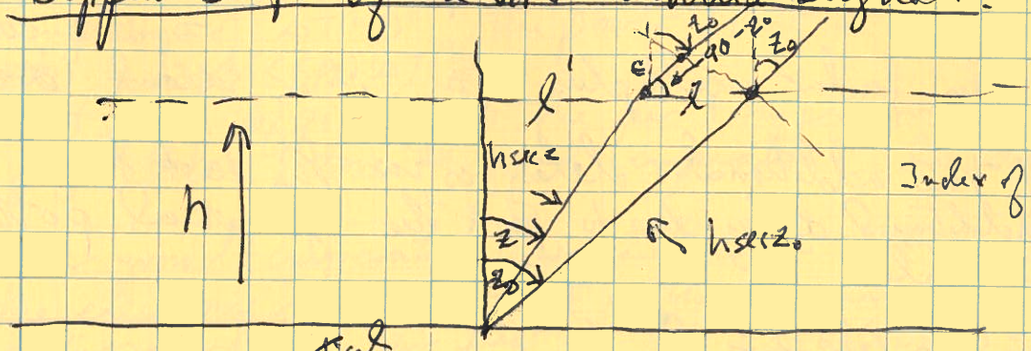
Item	**Cost**
Xeon Phi Consultant – 180 hours	$ 27,000
G. Harp and J. Richards, 50% time for three months	$ 62,000
Additional hardware and materials	$ 2,000
AT- RASC 2018 Conference Attendance and Presentation	$ 7,000
Total	**$ 98,000**

Appendix: Narrative of Erik Olson's Xeon Phi experience

My DARPA Cyber Fast Track (CFT) project was called "Weak Symbol Correlations." The purpose was to detect weak modulated signals by exploiting symbol clock coherency. It was funded for a year and it consisted of two parts. The first part involved novel DSP algorithm research based on a cyclostationary foundation. The algorithms were extremely CPU intensive which led to the second part of building a working prototype that used a 50M sample/second Software Defined Radio as a front-end connected to an Intel Xeon Phi co-processor. OpenMP was used for parallelizing the 60 cores (240 threads) on the Xeon Phi. Optimized code that used the Xeon Phi's 512-bit SIMD vector engine was written. Intel's processor analysis tools were used to understand and resolve the many of the core, cache, memory, and PCI bus bandwidth utilization issues. This required code architectural design changes to work around the bottlenecks. The project ended with a demonstration at the Pentagon during DARPA Demo days 2014.

Erik Olsen's full resume is attached.

Another look at the effect of earth's refraction on the Doppler shift of a GPS satellite signal.



Index of refraction: $n = 1 + N$

The total path optical difference must be calculated between the refracted ray and unrefracted ray.

$\epsilon + nh\sec z$ is the longer optical path
$h\sec z_0$ is the unrefracted optical path

$\tan z = l'/h \quad \tan z_0 = \dfrac{l'+l}{h}$; $l' = h\tan z$, $l'+l = h\tan z_0$

$h\tan z + l = h\tan z_0$, $l = h\tan z_0 - h\tan z = h\left(\dfrac{\tan z_0 - \tan z}{\Delta z}\right)\Delta z$

$l = h\dfrac{d\tan z}{dz}\Delta z$

$\cos(90° - z_0) = \epsilon/l = \sin z_0$, $\epsilon = l\sin z_0 = \sin z_0\, h\dfrac{d\tan z}{dz}\Delta z$

From page 98, refraction $\cong N\tan z = 287\times10^{-6}\tan z = \Delta z = \Delta\epsilon$

Optical path difference: $\epsilon + nh\sec z - h\sec z_0$

$= \sin z_0\, h\dfrac{d\tan z}{dz}\Delta z + nh\sec z - h\sec z_0$

$= h\left\{\sin z_0\dfrac{d\tan z}{dz}\Delta z + (1+N)\sec z - \sec z_0\right\}$

$= h\left\{\sin z_0\dfrac{d\tan z}{dz}\Delta z + \left[\dfrac{\sec z - \sec z_0}{\Delta z}\right]\Delta z + N\sec z\right\}$

$= h\left\{\sin z_0\dfrac{d\tan z}{dz}\Delta z - \dfrac{d\sec z}{dz}\Delta z + N\sec z\right\}$

$= h\left\{\left[\sin z_0\sec^2 z - \sec z\tan z\right]\Delta z + N\sec z\right\}$

$= h\left\{\left[\dfrac{\sin z_0}{\cos^2 z} - \dfrac{\sin z}{\cos^2 z}\right]\Delta z + N\sec z\right\}$

Note $\sin z_0 - \sin z = 2\sin\left(\frac{z_0 - z}{2}\right)\cos\left(\frac{z+z_0}{2}\right) \cong (z_0 - z)\cos z$

$= \Delta z \cos z$

Since this factor multiplies Δz, it is second order

Most of the additional delay is in the last term, $Nh\sec z$
The additional delay due to just the geometrical path is negligible.

$$\Delta\Phi = \frac{2\pi}{\lambda} Nh \sec z$$

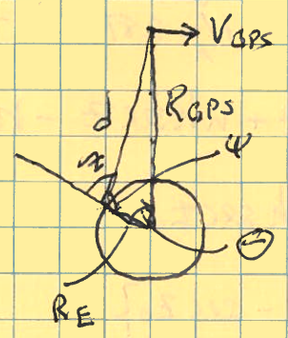
The Doppler shift is

$$\nu_{off} = \frac{1}{2\pi}\frac{d}{dt}\Delta\Phi = \frac{Nh}{\lambda}\frac{d\sec z}{dt} = \frac{Nh}{\lambda}\sec z \tan z\, \dot{z}$$

For $h = 10\,km = 10^6\,cm$, $\lambda = 18\,cm$, $z = 70°$

$$\nu_{off} = \frac{(287\times10^{-6})10^6}{18}\tan 70 \sec 70\, \dot{z} = 123\,\dot{z}\ Hz$$

<u>Possible values for $\dot{z}$</u>



$$\frac{\sin\psi}{R_{GPS}} = \frac{\sin\theta}{d}$$

$z = 180° - \psi$, $\sin\psi = \sin(180 - z) = \sin z$

$$\sin z = \frac{\sin\theta\, R_{GPS}}{d}$$

$$V_{GPS} = \left[\frac{GM}{R_{GPS}}\right]^{1/2} = \left[\frac{7\times10^{-8}\ 6\times10^{27}}{2.6\times10^{9}}\right]^{1/2} = 3.98\,Km/sec.$$

$$\dot{\theta} = \frac{3.98}{26{,}000} \sim 1.5\times10^{-4}\ rad/sec.$$

Near $d \cong R_{GPS}$ $\dot{z} \cong \dot{\theta} = 1.5\times10^{-4}$
Near $d_{min} = R_{GPS} - 6\,earth\,km$
$\dot{z} \cong 1.3\dot{\theta} \cong 2\times10^{-4}$

Thus $\nu_{off} = 123\,\dot{z} \cong$ <u>.02 Hz</u>

Cooling the bunker building with the ground as heat exchanger

Assume the same 3" PVC pipe as for the nodes with 50 CFM @ 5 m/sec
For a temp drop of $\Delta T(°F)$ the power flow is
$P(\text{watts}) = .336\,(CFM)\,\Delta T(°F)$
150 foot length of pipe gets the air temperature down to .024 of input value (above the ground temperature) for dry ground.

Summer ground temperature (worst case) is ~ 70°F (21°C)
If the electronics runs at 90°F in the summer, $\Delta T = 20°F$
the cooling power of the air flow is $P = (.336)(50)(20) = \underline{336 \text{ watts}}$ /pipe

In dry ground temp is down to 10% at a radius of 6" from the pipe centr.
6" is probably the minimum pipe separation. ok

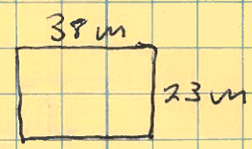
For a strip width of 50m (164 feet) one pipe every 6" is 328 pipes
One out for 75' and one back a foot underneath is one pipe! (no), one way

328 pipes is (328)(336) = 110,000 watts

A Beowulf Cluster with 240 processors in one 19" rack
costs 5200 watts to cool it, as an example

If we doubled the airflow through the pipes, we would double the cooling flow. [but λ goes down requiring longer pipe] ☆
R_e (Reynold's number) $\propto v$ and $h \propto (R_e)^{0.8}$, (p. 66,67)
so the heat transfer to the wall would also double (roughly). So the same length of pipe should work. (no)

Then 336 watts → 672 watts for $\Delta T = 20°F$.
125,000 watts requires 186 pipes, @8" centers that's 124' across
[Check the lifetimes of components @ 90°F.]



124' x 75' → 38m x 23m ; 874 m^2

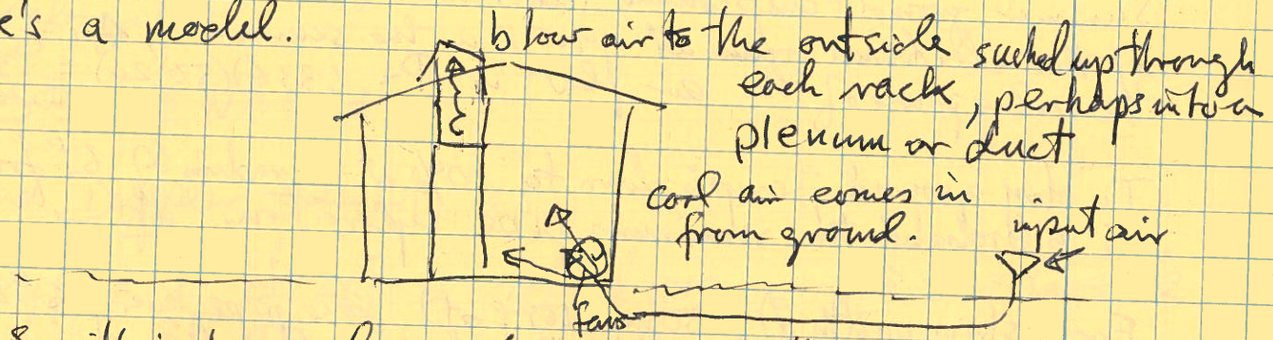
A total depth of ~3 feet should suffice.
Pit: 125' x 75' by 3 feet depth.

<u>Note</u> a) The rapid radial drop in temperature away from the pipe depends on air with a one day period.
b) The rapid drop in temperature along the pipe is much more rapid for one day period air rather than air at a fixed temperature.

☆ This argues for beginning with outside air to go into the pipes and <u>not</u> recycling the air from the Bunker building which would be nearly constant. It needs the cycling up and down of the input air temperature into the pipes for the best smoothing of the pipe air temperature.

J.D. made this point

Here's a model.



So, it's just one layer of pipes with inputs at 150 feet away, with filters for dust.
We could use a depth of 3' or 4' and get down further into the annual period

At 3' (.91 m), $e^{-\frac{.91}{2.1}} = e^{-.433} = .648$

$.65 \times 14.8°F = 9.6°F$, $9.6 + 58.6 = \underline{68°F}$ (20°C)

At 4' (1.21 m) $e^{-\frac{1.21}{2.1}} = e^{-.58} = .56$

$.56 \times 14.8 = 8.3°F$, $+ 58.6 = 66.9° \cong \underline{67°}$

It goes slowly. It looks like <u>3'</u> depth is a good choice. The pit becomes:

⟹ 125' x 150' x 3'

no, see preceding page

<u>Dielectrics</u>

delrin:	$K_e = 3.7$	loss tangent = .005 @ 10 Hz (!)
polystyrene:	$K_e = 2.5-2.6$	loss tan = .0001 @ 100 MHz, .0003 @ 3G (?)
Teflon:	$K_e = 2.0-2.1$	loss tan = .003 @ 3 GHz
FR-4:	$K_e = 4.9$	loss tan = .008 @ 3 GHz
polyethylene:	$K_e = 2.2$	loss tan = .0003 GHz

Rexolite is reinforced polystyrene, mechanically stronger; $K_e = 2.5$

go with this → #1422 has loss tan of .00066 @ 10 GHz
#2300 " .0014 @ 10 GHz, fiber glass reinforced.

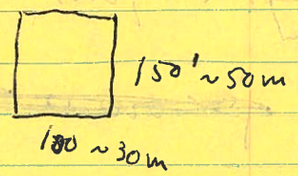
Underground pipes (future?) bldg

Assume 150 CFM air Flow as we are now using for the antennas in 3" PVC buried at 3 m.

With $P = .336\, F(\text{cfm})\, \Delta T(^\circ F) = (.336)(150)(20) = 1000$ watts/pipe

100 KW cooling capacity requires 100 pipes

At one foot centers it's 100 foot width overall. How to dig? 18" centers?



Area ~ (50)(30) = 1500 m^2

Solar const ~ 800 watts/m^2 ~ 1 kw/m^2

100KW/1500 m^2 = 1/15 kw/m^2

Cooling is ≲ 1000 watts per antenna, 500 watts w 50%

with $\Delta T = 20^\circ F \rightarrow 12.5^\circ C$

Observed more like 6°C = ΔT on hottest day.

That's 500 watts (It's also on a 100 foot line)

Wish list (2) Agilent Signal Generators that go up to 30 GHz (2) ~ 30k each

Contract with Niklas Wadefalk and Fraunhofer for 20 more LNAs. 27k

These are needed for receiver replacements

★ Go for 40 call Niklas!

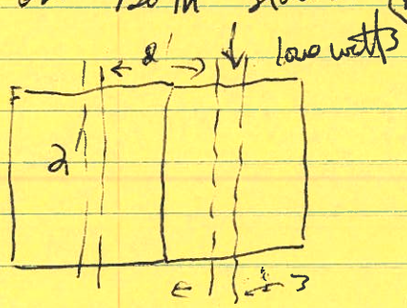
Put the pipes on both sides. (no)

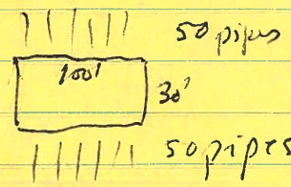
2 foot centers

cost (1) the building

(3) concrete slab

(5) pipes and blowers

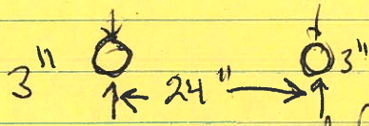




2' X 150' = 300 sq ft. ~ 20 m^2

⟨P⟩ = 1000/20 = 50 watts/m^2

In close, maybe 200 watts/m^2



better model

Could blow 200 cfm for 1300 watts, X100 = 130 Kwatts Total

problem back-filled with sand in the ditches instead of potting soil

solution displace more sand with potting soil

Building: ~~[illegible] x [illegible]~~ (to begin the look-up) [$40/sq ft.]

$\boxed{100 \times 30}$ for the building ($120,000)

40$/sq ft is typical

new bldg: 3000 x 40 = $120,000 ~~for one antenna~~

+ bath room

how much cost for cables from main lab.

High Frequency Circuit Materials - Properties

Product	Composition	Dielectric Constant ε_r[1] Tolerance @ 10GHz	Dissipation[1] Factor TAN δ @ 10 GHz	Thermal[2] Coefficient of ε_r -50° - 150°C ppm/°C (Typical)	Volume Resistivity Mohm cm (Typical)	Surface Resistivity Mohm (Typical)	Youngs Modulus[3] kpsi (MPa) (Typical)			Moisture[4] Absorption D24/23% (Typical)	Thermal[5] Conductivity W/m/°K (Typical)	Coefficient of[6] thermal Expansion 0-100°C (Typical)			Density gm/cm³ (Typical)	Peel Strength lbs/in (N/mm) (Typical)	Flammability Rating	Standard Thickness in. (mm)
							X	Y	Z			X	Y	Z				
RT/duroid® 5870	PTFE Glass Fiber	2.33 ± 0.020	0.0012	-115	$2X10^7$	$2X10^8$	189 (1,340)	185 (1,277)	120 (828)	0.015	0.22	22	28	173	2.2	20.8 (3.7)	UL 94V-0	0.005", (0.127mm) 0.010", (0.254mm) 0.015", (0.381mm) 0.020", (0.508mm) 0.031", (0.787mm) 0.062", (1.575mm) 0.125", (3.175mm)
☆ RT/duroid® 5880	PTFE Glass Fiber	2.20 ± 0.020	0.0009	-125	$2X10^7$	$3X10^7$	156 (1,076)	125 (863)	136 (938)	0.015	0.20	31	48	237	2.2	22.8 (4.0)	UL 94V-0	0.005",(0.127mm) 0.010",(0.254mm) 0.015", (0.381mm) 0.020", (0.508mm) 0.031", (0.787mm) 0.062", (1.575mm) 0.125", (3.175mm)
ULTRALAM® 2000	PTFE Woven Glass	2.40-2.60 ± 0.040	0.0019	-100	$2X10^7$	$4X10^7$	1,700 (11,730)	1,300 (8,970)		0.03	0.24	15	15	200	2.2	18.0 (3.2)	UL 94V-0	0.004", (0.101mm) 0.0101", (0.256mm) 0.0147", (0.373mm) 0.0190", (0.482mm) 0.030", (0.762mm) 0.060", (1.524mm)
RT/duroid® 6002	PTFE Ceramic	2.94 ± 0.040	0.0012	+12	10^6	10^7	120 (828)	120 (828)	360* (2,482)	0.1	0.60	16	16	24	2.1	8.9 (1.6)	UL 94V-0	0.005", (0.127mm) 0.010", (0.254mm) 0.020", (0.508mm) 0.030", (0.762mm) 0.060", (1.524mm) 0.120", (3.048mm)
RT/duroid® 6006	PTFE Ceramic	6.15 ± 0.150	0.0019	-410	$2X10^7$	$7X10^7$	75 (511)	91 (628)	155 (1,070)	0.05	0.49	47	34	117	2.7	14.3 (2.5)	UL 94V-0	0.010", (0.254mm) 0.025", (0.635mm) 0.050", (1.270mm) 0.075", (1.905mm) 0.100", (2.540mm)
RT/duroid® 6010LM	PTFE Ceramic	10.2 ± 0.250	0.0023	-425	$5X10^6$	$5X10^6$	135 (932)	81 (559)	311 (2,146)	0.05	0.78	24	24	24	3.1	12.3 (2.1)	UL 94V-0	0.010", (0.254mm) 0.025", (0.635mm) 0.050", (1.270mm) 0.075", (1.905mm) 0.100", (2.540mm)
TMM®3	Hydrocarbon Ceramic	3.27 ± 0.032	0.0020	+39	$3X10^9$	$>9X10^9$	1,916 (13,210)	1,916 (13,210)	742 (5,116)	[4] 0.04	0.70	16	16	20	1.78	5.7 (1.0)	N/A	0.015", (0.381mm) 0.020",(0.508mm) 0.030",(0.762mm) 0.060",(1.524mm) 0.125",(3.175mm)
TMM®4	Hydrocarbon Ceramic	4.50 ± 0.045	0.0020	+15.3*	$6X10^8$	$1X10^9$	2,000* (13,790)	2,000* (13,790)	752 (5,185)	0.010	0.70	14	14	20	2.07	5.7 (1.0)	N/A	0.015", (0.381mm) 0.020",(0.508mm) 0.030",(0.762mm) 0.060",(1.524mm) 0.125",(3.175mm)
TMM®6	Hydrocarbon Ceramic	6.00 ± 0.080	0.0023	-10	$1X10^8$	$1X10^9$	2,200 (15,168)	2,200 (15,168)	736 (5,075)	0.06	0.72	16	16	20	2.37	5.7 (1.0)	N/A	0.015", (0.381mm) 0.025", (0.635mm) 0.050",(1.270mm) 0.075", (1.905mm) 0.100", (2.540mm) 0.125", (3.175mm)
TMM®10	Hydrocarbon Ceramic	9.20 ± 0.230	0.0023	-38	$2X10^8$	$4X10^7$	2,400 (16,547)	2,400 (16,457)	575 (3,964)	0.09	0.76	16	16	20	2.77	5.0 (0.9)	N/A	0.015", (0.381mm) 0.025", (0.635mm) 0.050",(1.270mm) 0.075", (1.905mm) 0.100", (2.540mm) 0.125", (3.175mm)
TMM®10i	Hydrocarbon Ceramic	9.80 ± 0.245	0.0020	-43	$2X10^8$*	$4X10^7$*	2,400* (16,547)*	2,400* (16,457)*	575* (3,964)*	0.16	0.76	16*	16*	20*	2.77	5.0 (0.9)	N/A	0.015", (0.381mm) 0.025", (0.635mm) 0.050",(1.270mm) 0.075", (1.905mm) 0.100", (2.540mm) 0.125", (3.175mm)
RO4003C®	Hydrocarbon Ceramic	3.38 ± 0.05	0.0027	+40	$1.7X10^{10}$	$4.2X10^9$	3,700 (25,510)	3,900 (26,889)		0.06	0.64	11	14	46	1.8	6.4 (1.1)	N/A	0.008", (0.203mm) 0.020", (0.508mm) 0.032", (0.813mm) 0.060", (1.524mm)
[8]RO4350B®	Hydrocarbon Ceramic	3.48 ± 0.05	0.0037	+50	$1.2X10^{10}$	$5.7X10^9$		1,664 (11,473)		0.06	0.62	14	16	50	1.9	5.3 (0.9)	UL 94V-0	0.004", (0.101mm) 0.0066", (0.168mm) 0.010", (0.254mm) 0.020", (0.508mm) 0.030", (0.762mm) 0.060", (1.524mm)
RO4450B®	Hydrocarbon Ceramic Prepreg	3.54 ± 0.05	0.0040							0.05	0.60	19	17	60	1.86		UL 94V-0	0.004" (0.101mm)
RO3003®	PTFE/Ceramic	[7]3.00 ± 0.04	0.0013	13	10^7	10^7	300 (2,068)	300 (2,068)		<0.1	0.50	17	17	24	2.1	17.6 (3.1)	UL 94V-0	0.005 (0.127mm) 0.010", (0.254mm) 0.020", (0.508mm) 0.030", (0.762mm) 0.060", (1.524mm)
RO3203®	PTFE/Ceramic Reinforced Woven Glass	[7]3.02 ± 0.04	0.0016	13	10^7	10^7				<0.1	0.50	13	13	58	2.1	10 (1.7)	UL 94V-0	0.010", (0.254mm) 0.020", (0.508mm) 0.030", (0.762mm) 0.060", (1.524mm)
RO3006®	PTFE/Ceramic	6.15 ± 0.15	0.0020	-160	10^3	10^3	300 (2,068)	300 (2,068)		<0.1	0.61	17	17	24	2.6	12.2 (2.1)	UL 94V-0	0.010", (0.254mm) 0.025", (0.635mm) 0.050",(1.270mm)
RO3010®	PTFE/Ceramic	10.2 ± 0.30	0.0023	-280	10^3	10^3	300 (2,068)	300 (2,068)		<0.1	0.66	17	17	24	3.0	13.4 (2.4)	UL 94V-0	0.010", (0.254mm) 0.025", (0.635mm) 0.050",(1.270mm)
RO3210®	PTFE/Ceramic Reinforced Woven Glass	10.2 ± 0.50	0.0027		10^4	10^4				<0.1	0.81	13	13	34	3.0	13.4 (2.4)	UL 94V-0	0.025", (0.635mm) 0.050",(1.270mm)

Replacing the delrin standoffs with rexolyte should reduce the losses

Effective Areas of the low and mid range calibration horns

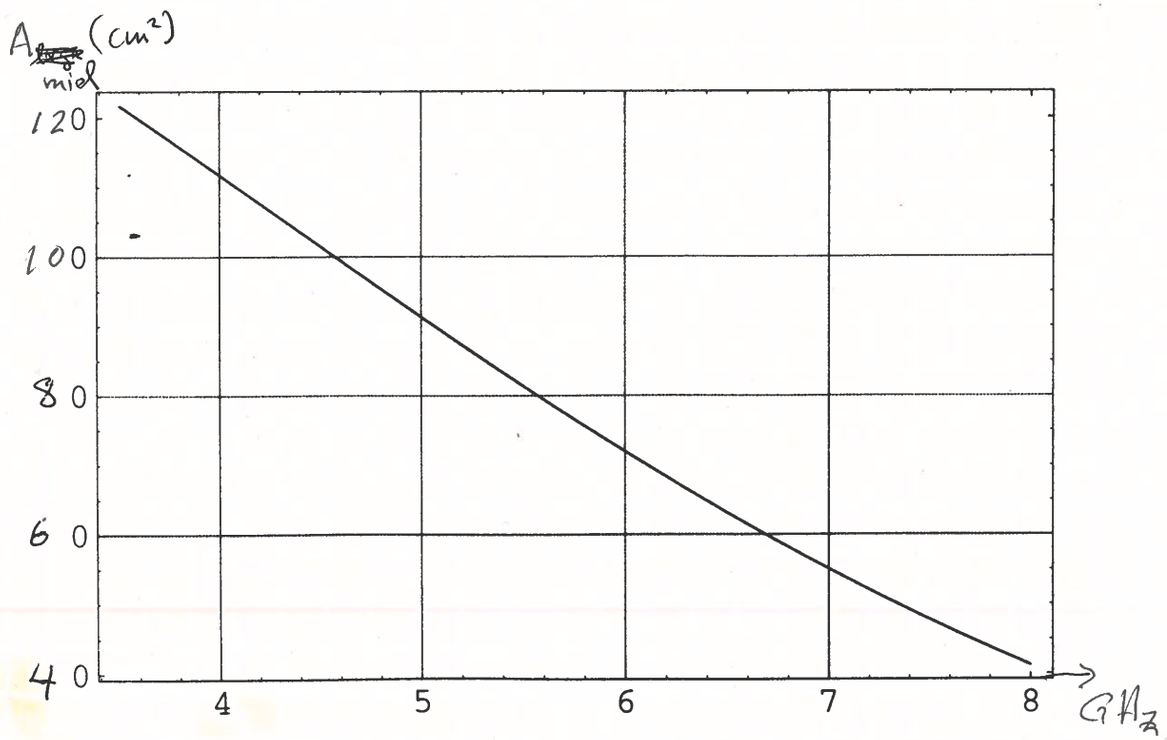
Mid band Horn (Artyom's physical measurements) notation of p. 103

$a = 16.307$ cm
$b = 12.28$ cm
$R_E = 13.67$ cm
$R_M = 11.32$ cm

$u = .1459\lambda^{1/2} + 3.427\lambda^{-1/2}$
$v = .1459\lambda^{1/2} - 3.427\lambda^{-1/2}$
$w = 2.344\lambda^{-1/2}$
waveguide cutoff ~ 3 GHz

Plot the effective area which is $A = \frac{g\lambda^2}{4\pi} \approx A_{mid}$

This horn has a wide opening angle with a lot of phase error. It's area is not constant like the x-band horn. It drops 4.3 db from 4 GHz to 8 GHz



$W_{mid}(f) = 214.3 - 29.75f + 1.0096f^2$

The x-band horn is constant @ 9.19 cm² over 8-12 GHz

Lower Band Horn

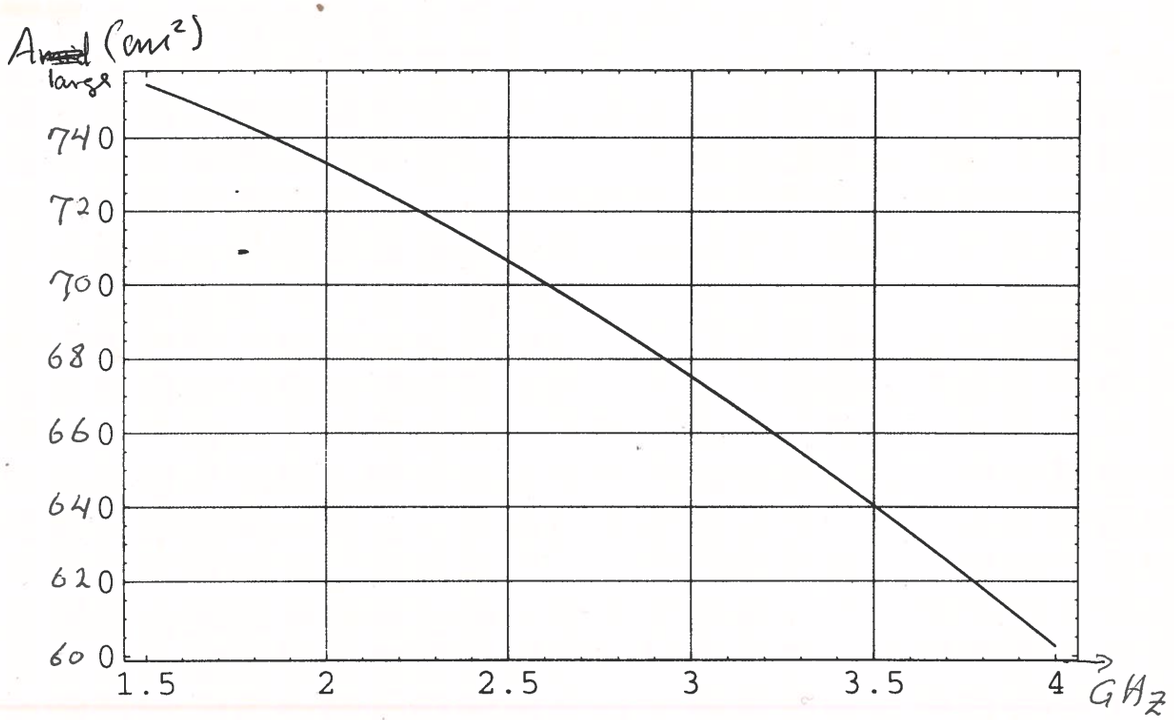
$a = 35.56$ cm
$b = 27.18$ cm
$R_E = 76.84$ cm
$R_H = 65.60$ cm

$u = .1611\lambda^{1/2} + 3.105\lambda^{-1/2}$
$v = .1611\lambda^{1/2} - 3.105\lambda^{-1/2}$
$W = 2.1925\lambda^{-1/2}$

$A = \frac{3}{4}\frac{\lambda^2}{\pi} \equiv A_{eff}$

The ~~gain~~ area drops about 1 db from 2 to 4 GHz

Waveguide cutoff ~ 1.3 GHz



$W_{large}(f) = 796.4 - 15.585f - 8.250f^2$

Some reflection measurements with the passive feed.
This one has polystyrene standoffs, except that the first two sets of standoffs are missing. Also, the feed arms are soldered together. 1/22/06 rental unit Agilent 8523D
Both improved the results. The unsoldered connection was erratic.
Gate set @ (4.61 nsec) (in front of the tip circuit board; antenna only)
Peaks only are identified

	ν	observed	corrected for input cable and balun losses (p.14)
	.585	−17 db	−16.5
not a peak	1.000	−20 db	−18.5
	1.100	−15 db	−13.6
	2.70	−17 db	−15.0
	5.40	−20 db	~~−17.8~~ −16.0
	9.00	−20.7 db	−14.7

(what or where is this?)

Gate set at LNA position (4.056 nsec)

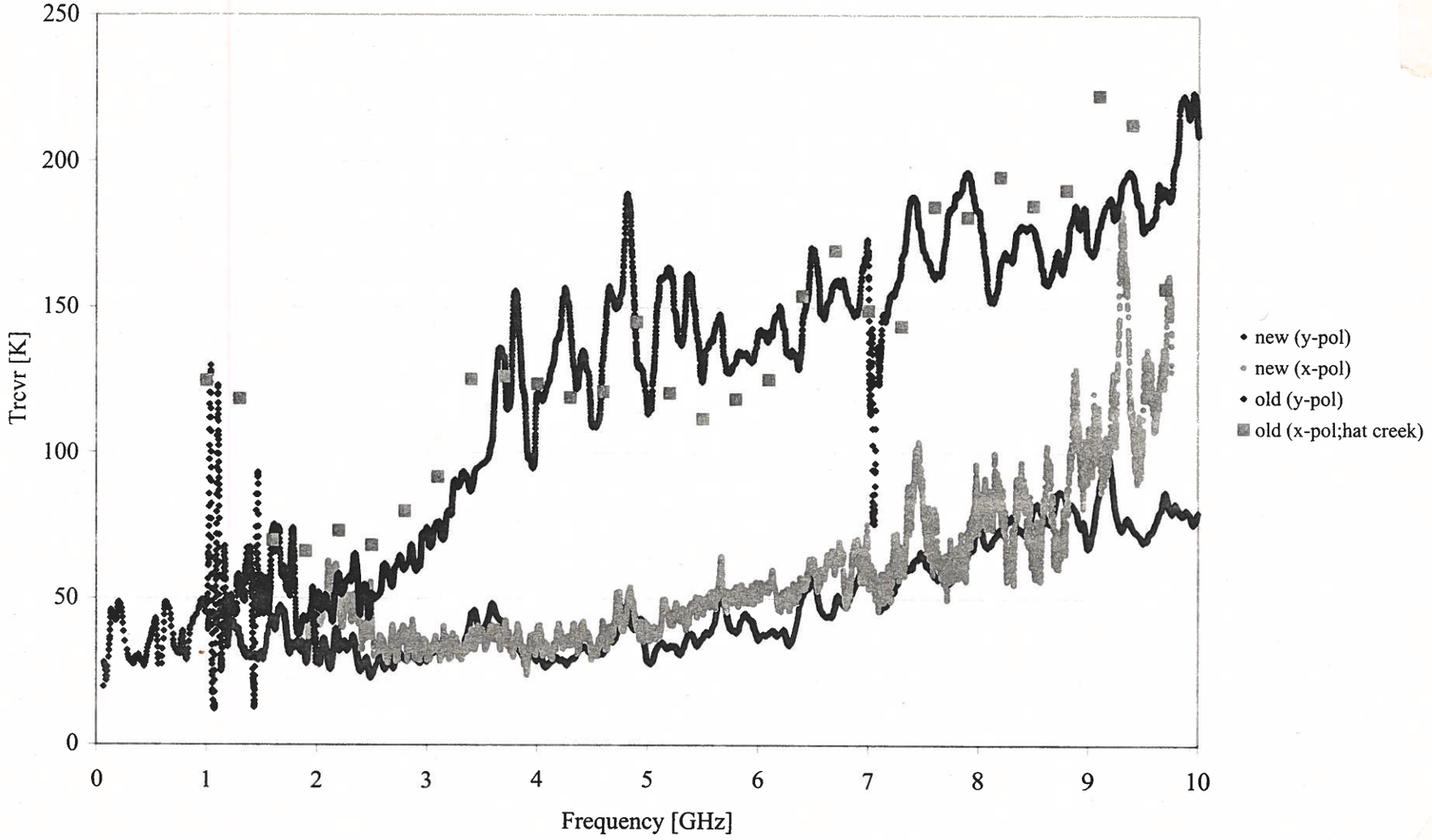
ν	observed	corrected. (p.14 + balun)	← transmission (db) factor
.500	−19	−18	
.638 (peak)	−17	−16	−0.1
.750 − 1.01	< −20	< −18	
1.13	−13	−11.5	−0.3 (.93)
peaks { 2.6	−11	−8.6	−0.7 (.85)
2.7	−11	−8.6	−0.7
5.3 (peak)	−12.4	−8.4	−0.7
5.8	−14.3	−10.3	−0.3 (.93)
8.8	−21	−16	−0.1 (.98)
9.0 (peak)	−15	−10	−0.45 (0.90)
9.5 (peak)	−12	−7	−1.0 (0.8) (.79)
10.4 (peaks)	−14.6	−9	−0.6 (.87)
11.0 (peak)	−13.7	−8	−0.75 (.83)
		−4	−2.2 (.60)

These were visible on the first passive feed. p.19

But they don't appear to be on the feed itself. See above.

NB!

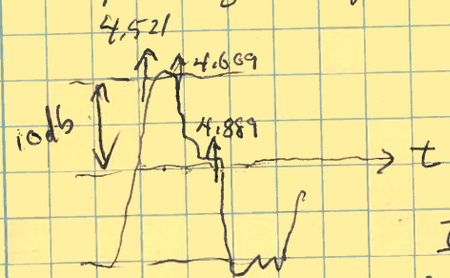
ATA Feed Sensitivity for New and Old Tip Circuit Board Designs
(Truncated Bishop)



Tests of passive tip circuit board (~ 1/12/06 and a little later) with Network Analyzer (some with our NA, some with rental unit)

Sweep .05 – 12 GHz Maury Cal Kit – use tdr option
Razor blade on tip circuit board for short gives spike @ 4.521 nsec.
resolution of spike in time domain { 75 ps for ±10db width
24 ps for ±3db width [.024 nsec.]

201 pt required for it to work.



4.609	4.888
4.521	4.521
.088	.368

(lab notes copied 1/27/06)

It looks like the extent of the tip circuit into the connection to the feed corresponds to the 4.889 ns echo. Δt (one way) = .184 nsec. (resolved, see above)

In free space 1 nsec = 30 cm or 12" 0.1 nsec ≈ 1"
Sweep is to 12 GHz, suggesting resolution of $\frac{1}{12}$ = 1 inch ; $\frac{1}{12}$ nsec = .083 sec
This agrees with the above measured resolution ~~[illegible]~~ (@ 10 db)

The echo remains high from 4.521 to 4.609, .088 nsec. Half of that, .044 nsec, for the one way propagation, is 1/2 inch in free space. The dielectric constant is ~ 2.3 so the distance is ~ 0.3" in the dielectric of the tip circuit board. The extent of the circuit board to edge from the center is close to .25". All the echo is from the tip circuit board. Just in front of that, from 4.60 nsec on, the echos are small. See the previous page (p 114).

Echos back inside the shower.

First piece of plastic at 1.128 in = .0994 ≅ 0.1 nsec.
times 2 = 0.2 nsec. on the screen for the Round Trip
4.521 – 0.2 = 4.321 (first piece of plastic)
4.286 next small piece
4.230 deep null
4.168 nsec sharp piece (glass?)
4.056 deep nulls
3.789 begins broad discontinuity, which extends to 2.41 nsec. then drops low to 1.33 nsec.

Details in the test dewar

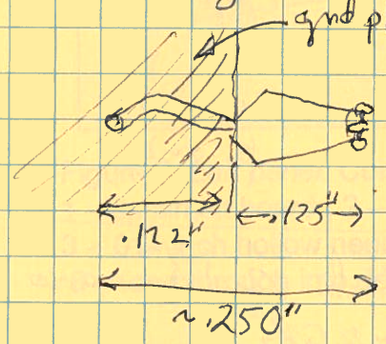
In the passive feed the balun board is RO4003 20 mils thick
5" long $k_e' = 3.38$, $\tan\delta = .0027$ @ 10 GHz ; 50 Ω → 200 Ω
check this

Tip Circuit Board (first try) RO5880 $k_e = 2.4$ $\tan\delta$ (wrd) = .000
20 mils thick It should have $Z_0 = 110\ \Omega$

Eagleware gives 11.5 mils thick for $Z_0 = 110\ \Omega$
Total loss/inch = .088 { .066 conductor, .022 dielectric

It looks narrower, ~9 mils with $Z_0 = 120\ \Omega$ and
total loss/inch = .094 (10 GHz here)

The length on the board is .122" · .122 × .094 = .0115 (1.2%)



@ 10 GHz, $\lambda = 3\text{cm} = 1.18"$
$\lambda/2 = .59"$
$\lambda/4 = 0.29"$

It should radiate well at ~10 GHz!

The second Teflon tip board is Cuflon.

Review 2-wire line losses

III p. 42

$$\alpha_p = \frac{2R_s}{\pi d Z_0}\frac{(s/d)}{\sqrt{(s/d)^2-1}}$$

$$Z_0 = 120\cos^{-1}(s/d) = 276\log_{10}\left(\frac{2s}{d}\right)$$

for silver: $R_s\ 2.52\times10^{-7}\sqrt{f}$

a) For s = 40 mils, d = 14 mils (.036 cm) $\alpha_p = .0024\sqrt{f/10}\ \text{cm}^{-1}$

b) For s = 40 mils (.021 cm), and d = 8 mils ; s/d = 5 and $Z_0 = 274\ \Omega$ (no wall)

$$\alpha_p = \frac{2(2.52\times10^{-7})\sqrt{f}}{\pi(.021)\,274}\frac{5.0}{\sqrt{(5.0)^2-1}} = .0029\sqrt{f/10}$$

Z_0 drops and losses go up with a nearby wall

c) For d = 3.2 (.008 cm), and s = 16 mils $Z_0 = 276\log_{10}\left(\frac{2\times9.1}{3.2}\right) = 208\ \Omega$ (no wall)

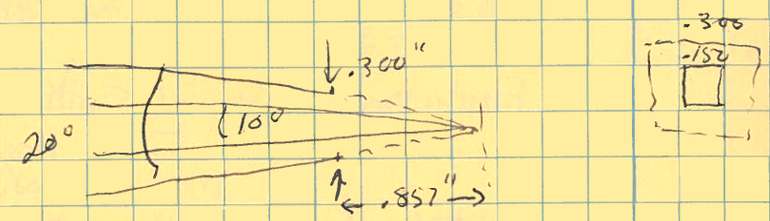
$$\alpha_p = \frac{2(2.52\times10^{-7})\sqrt{f}}{\pi(.008)\,208}\frac{5.0}{\sqrt{(5.0)^2-1}} = .0096\sqrt{f/10}\ \text{cm}^{-1}$$

[check Z_0]

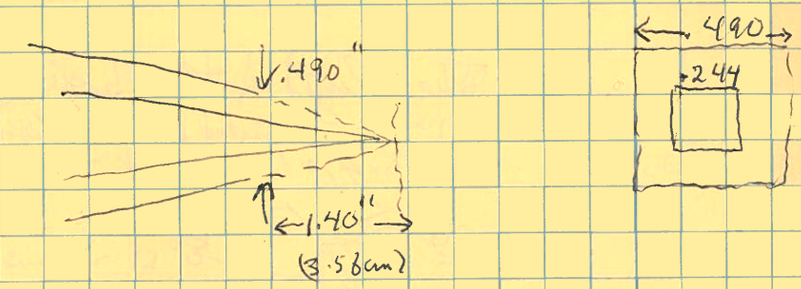
Extending the feed to 25 GHz (27 GHz)

feed scaling
feed angle = 20° (.35 rad)
pyramid angle = 10° (.175 rad)
$\alpha = r\theta$

first tip circuit board

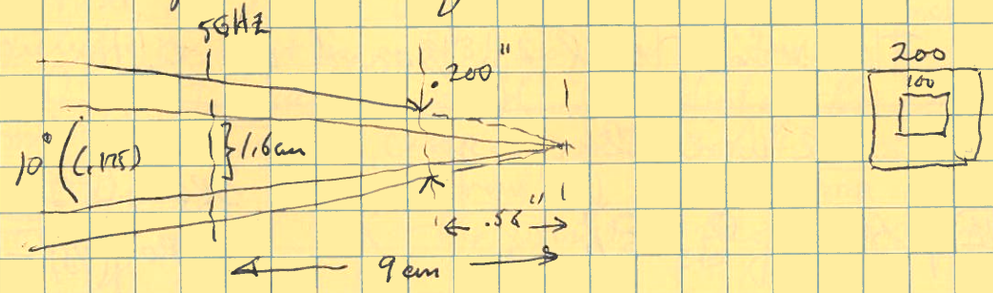


Current tip circuit board



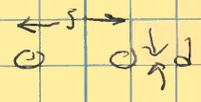
The input line is 8 mils here

Tip Board for 27 GHz; change by factor of 12/30 = 0.400 to get the ½ resonance up to 30 GHz from ~12 GHz



Length change 1.40 − .56 = .84", 2.13 cm

New wire diameter 0.4 × 8 mils = 3.2 mils



For s/d = 5.0, Z_0 = 208 Ω, s = 16 mils
$\alpha = .0096\sqrt{f/10}\ \text{cm}^{-1}$

Additional loss with additional length .0096 × 2.13 $\sqrt{f/10}$ = × 0.020 $\sqrt{f/10}$ 2% %
~ 6K at 10 GHz

Overall there will be ~ 2.6" (present from LNA) + .84" or 6.6 + 2.13 cm = 8.7 cm
If it were all in 3.2 mil (208 Ω) wire, the loss would be 0.009 (8.7) $\sqrt{f/10}$
= .082 $\sqrt{f/10}$, 8.2% at 10 GHz (24K), 8.2 × 1.6 ~ 13% at 25 GHz
24K ≈ 40K in the line

Plan to extend the dewar out beyond the tip circuit board into a glass sphere that surrounds the tip and includes the high frequency part of the feed for, say, $\nu \gtrsim 5$ GHz or maybe lower.

The figures below shows the current feed arrangement beginning with the dewar window in green. The lower figure shows the dewar expanded into the sphere

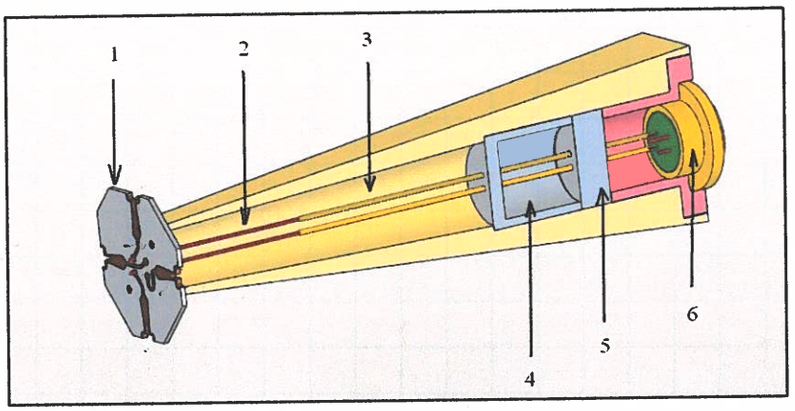
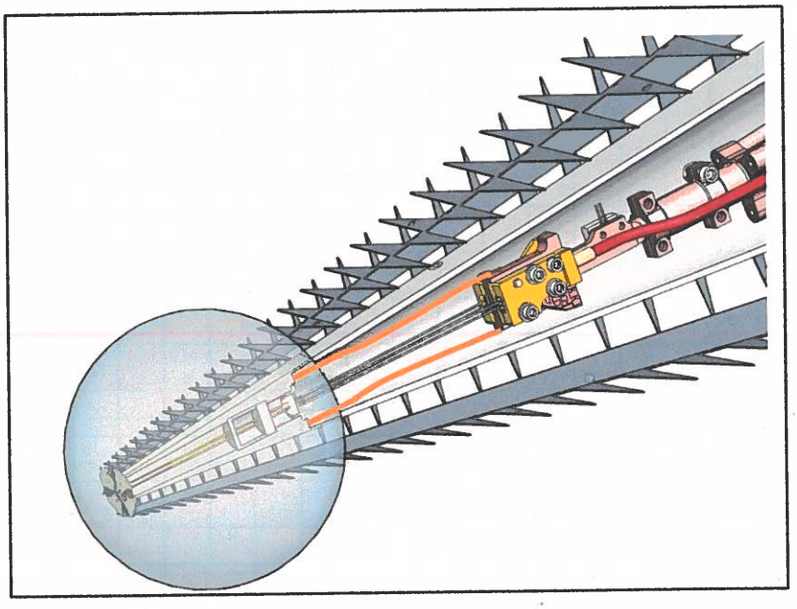


Figure 7: *Left panel:* Current feed through design for ATA
1 – tip circuit board , 2 – .0085 inch wire,
3 - .014 inch hollow needle, 4&5 – Teflon matching lens,
6 – feed through into dewar.



Note orange heat straps

Right panel: Cartoon of glass vacuum bulb at end of feed to extend the dewar and cool the physical feed structure.

A. The radiative heat loads

I. The current structure in the dewar (see p. 77)

a) the central tube

$R = .126'' \rightarrow .32\,cm$ ✓

$L = 5.47'' + .916'' = 6.39\,in \rightarrow 16.2\,cm$ ✓

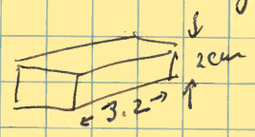
$Area = 2\pi R L = 2\pi(.32)(16.2) = 32.6\,cm^2$ ✓

b) small tubes $R = .063'' \Rightarrow .16\,cm$ ✓

$L = 6'' = 15.2\,cm$ ✓

$2\pi R L \times 3 = 2\pi(.16)(15.2)\,3 = 46\,cm^2$ ✓

c) LNA box approx. 1.25" long .75" on a side



Area $(3.2 \times 2) \times 4 = 26\,cm^2$ ✓

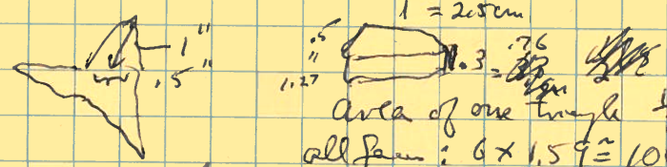
d) input cylinder

$R = .25'' \Rightarrow .64\,cm$ ✓

$L = .9'' \Rightarrow 2.3\,cm$ ✓

$2\pi R L = 2\pi(.64)(2.3) = 9.3\,cm^2$ ✓

e) input frame



area of one triangle $\frac{1}{2}(1.27)2.5 = 1.59$ ✓

all faces: $6 \times 1.59 \cong 10\,cm^2$

edges $(.76)(2.54) \times 6 = 11\,cm^2$

Total area: $33 + 46 + 26 + 9 + 11 = 125\,cm^2$ ✓

estimate of radiative heat load on this surface.

An upper limit, probably a good approximation, to the radiation field within the dewar is that of a black body at ~300K. This gives a flux of heat on every square centimeter of the inner structure that is σT^4 with $T = 300K$

$\sigma = 5.7 \times 10^{-12}$ watts/cm²/K⁴ ✓ at $T = 300$

$5.7 \times 10^{-12}(300)^4 = 5.7 \times 10^{-12}(8.1 \times 10^9)$ watts/cm² $= .046$ watts/cm²

$F = 46$ mwatts/cm² ✓ At 250K get 35 mwatts/cm²

The structure in the dewar is copper or silver which has $\epsilon_m = .04$

The fraction of the heat incident that is absorbed must be the same.

$46(.041) = 1.8$ mwatts/cm² ✓

This heat is taken away by the copper to the cold head.

The total is $(1.8)(125) = 230$ mwatts ✓

There are also the heat inputs from the connectors and supply wires.

II. The added surface

a) from the transmission line it is tiny

b) from the cooled pyramid tip back to 9 cm for 5 GHz [9 cm from the vanishing point.] The width of the pyramid at this point is $9(.125) = 1.6$ cm ✓

The area of the triangle all the way to the vanishing point would be $\frac{1}{2}(1.6)\times 9 = 7.2\ cm^2$ ✓

For all four sides $\times 4 = 29\ cm^2$ ✓

→ Double that for the added area of the feed arms (both sides) ~~(inside only)~~

$2 \times 29 \cong \underline{\underline{60\ cm^2}}$ (probably more) ok

That is about half of the interior ✓

That brings the total load to $1.5 \times 230 = \underline{345\ \text{milliwatts}}$

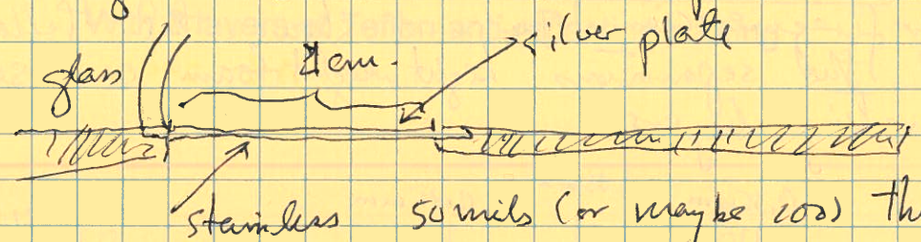
This is small compared with the 6 watt capability of the STI cooler.

Probably it can be cooled to 50K.

B. Heat flow along the surface of the pyramid from the outer part to the region inside the glass.

The surface will be silver plated which will carry the heat.

Let the gradient from 280K to 80 (or 50)K occur over a 1 cm span just inside the glass wall of the glass barrel where it attaches to the outside metal. This region will be consist of stainless steel silver plated.



50 mils (or maybe 100) thick

It requires 3μ of silver to provide one skin depth at 500 MHz.

☆ $3\mu = .00012'' = 120\ \mu\text{inches}$, .12 mils [2μ for 1 GHz]

The thermal conductivity of silver is ~ 2600 times larger than stainless. It would take $(.12)(2600) = 315$ mils of SS to have the same thermal resistance as the silver layer (3μm). 50 mils will add ~~a~~ little to the thermal ~~resistance~~ conductance.

The heat flow is: $K \frac{\Delta T}{\Delta x} \text{Area} = H$ $\qquad K = 400 \ w^t/m^{\circ K}$ (Silver)

width of the pyramid at 5GHz point (9 cm) is 1.6 cm

$A_{m^2} = wt = 4 \times (.016) 10^{-6} (3) = (.064) \times 10^{-6} (3) \ m^2$

$H = (400) \frac{(280 - 80)}{.01 m} (.064 \times 10^{-6})(3) = \underline{\underline{1.5 \text{ watts}}}$ for $t = 3\mu$

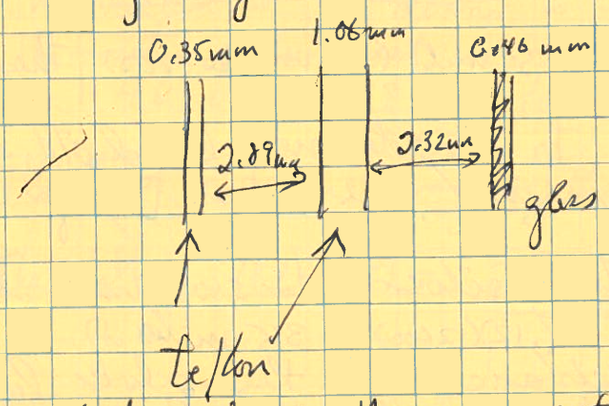
This is comfortably less than the 6 watt capability of the STI cooler. The inner part of the pyramid will be directly connected to the heat exchanger cold head. The thermal connections between cold tip and pyramid are shown in orange above.

C. How will the glass ball affect the electromagnetic properties of the feed?

The glass thickness of a typical light bulb (one example) is .018", 18 mils, which is 0.46 mm. Even at 27GHz (λ = 1.11 cm = 11.1 mm) the glass is only 1/24 of a wavelength thick. Including $k_e = 4$, it's 1/12 of a wavelength.

Look at the reflection of a plane piece of glass for a plane wave normally incident. The upper curve shows the reflection coefficient in blue and the reflection loss in red. $S_{11} = -10$ db at 25GHz and $S_{21} = -.48$ db at 25GHz.

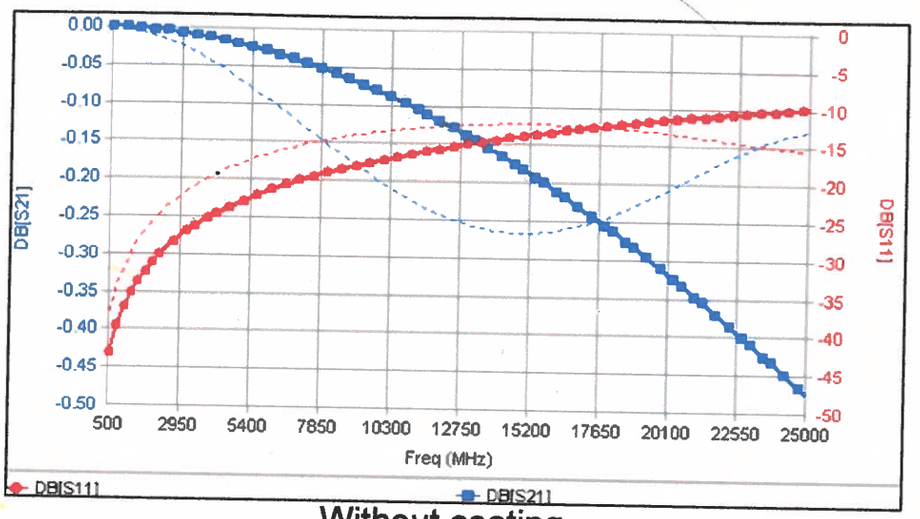
The lower curves show what a two Teflon layer of antireflection coating can do. It's slightly worse at the lower frequencies. The next shows the Teflon layer thickness and the separations. Lightweight foam can serve as the separating layers.



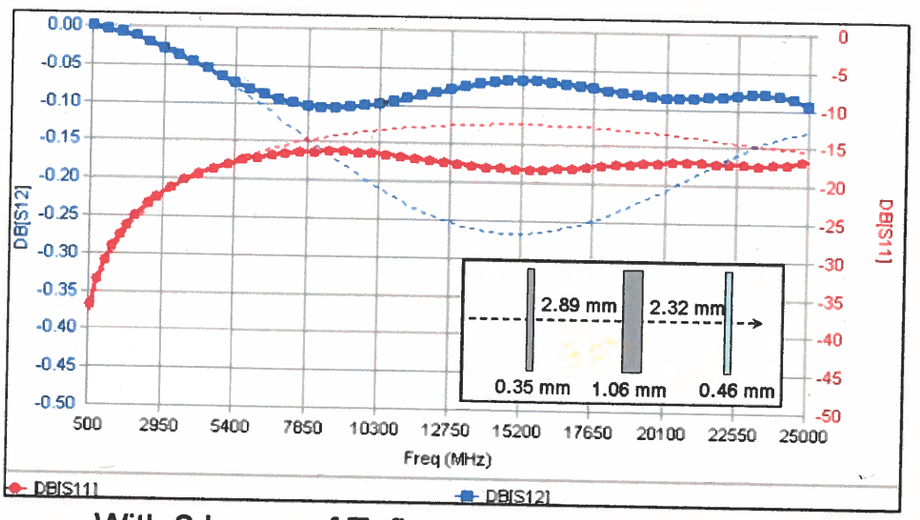
put these layers on the glass.

Simulations done with pieces of transmission line in Eagleware.

Reflection of plane wave from flat glass of thickness 0.46 mm



Without coating



With 2 layers of Teflon anti-reflection coating

Area of the glass ball for a radius of 9cm is 1000 cm². Effective area of the cold interior part is ~ 60 cm². With it's ~~area~~ emissivity of 40%, it won't significantly cool the glass

T_{sys} estimate with the cooled feed. 3/01/06

p. 47 has $T_{sys} = \frac{7.8}{\sqrt{f_{GHz}}} + 17.5\sqrt{f/10} + 16$

without the balun contribution. The 16 K is 10 for rcvr and 8 K from the atmosphere. Cooling the tsp gives the factor 75/300, assuming the same losses, to multiply the 17.5 K.

$T_{sys} = \frac{7.8}{\sqrt{f_{GHz}}} + 4.0\sqrt{f/10} + 16 = \cancel{\frac{7.8}{\sqrt{f_{GHz}}} + \sqrt{f} + 16}$

C_0

f	T_{sys}
5G	23
10G	23

$T_{sys} = \frac{7.8}{\sqrt{f_G}} + 1.4\sqrt{f_G} + 16$

It could be a little lower in the second term with allowance for lower silver losses at the low temperature.

C_0

From the air temperature data of pp 59-60 and the next page we can see three components: the daily variation, the annual variation, and a weather component → is has random phase and periods in the range 3-20 days.

A fit to the data:

For the annual: $T_{an} \cong 286 + 10 \cos(\Omega_{an} t + \phi_{an})$ K

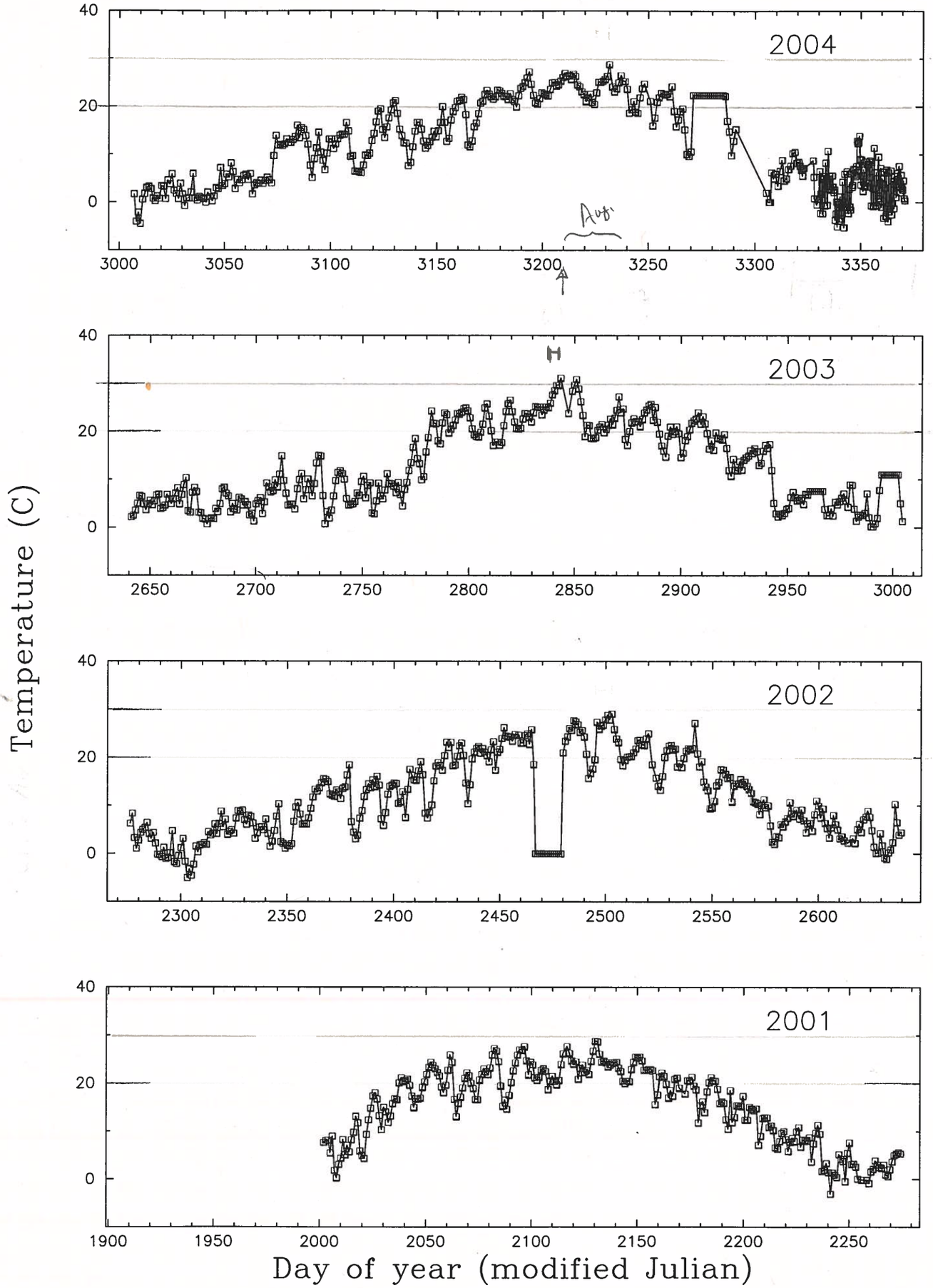
For the daily: $T_{da} \cong 10 \cos(\Omega_{an} t + \phi_{da})$ K

For the weather: $T_{wa} \cong 5 \cos(\Omega_{wea} t + \phi_{wea})$ K

Penetration depths $\delta_T = \sqrt{\frac{2K}{\rho c_v}}$

	δ_T	$3\delta_T$
$\Omega_D = 1$ day	11 cm	33 cm
$\Omega_W = 15$	43 cm	128 cm
Ω_{AN}	210 cm	630 cm

Daily average air temperature (2001–2004)



At 2 foot depth:

daily = 0

annual: $e^{-\frac{61}{210}} = .74$; $7.4 + 286 = 293.4\,K$

weather: $.25 \times 5 = 1.3$; $293.4 + 1.3 = 294.7\,K$, $\underline{\underline{21.7°C}}$ 71°F

At 3 foot depth

daily = 0

annual: $.647 \times 10 + 286 = 292.5\,K$

weather: $.12 \times 5 = .6$ Total = $\underline{293.1K}$, $\underline{20C}$ or $\underline{68.2F}$

What's the equation for the diffusion (annual) with the time varying diffusivity?

From the next page (128) $\alpha = \frac{k}{\rho c} = \alpha(t)$

$$\alpha(t) = \frac{k(t)}{\rho c} \simeq \frac{k_0 + k_1 \cos \Omega_a t}{\rho c} = \alpha_0 + \alpha_1 \cos \Omega_a t$$ for a first order relation.

$$\frac{\partial^2 T}{\partial x^2} = \frac{1}{\alpha(t)} \frac{\partial T}{\partial t}$$ We can separate the variables: $T(x,t) = X(x)\tau(t)$

$$X''\tau = \frac{1}{\alpha} X \tau' , \quad \frac{X''}{X} = \frac{1}{\alpha}\frac{\tau'}{\tau} = \gamma^2 \text{, a constant} ; \quad X(x) = e^{\pm \gamma x} \to e^{-\gamma x}$$

$$\frac{1}{\alpha}\frac{\tau'}{\tau} = \gamma^2 \qquad \frac{d\tau}{\tau} = \alpha \gamma^2 dt \quad ; \quad \int \frac{d\tau}{\tau} = \int \alpha(t)\gamma^2 dt + \ln a$$

$$\ln\left(\frac{\tau}{a}\right) = \gamma^2 \int \alpha(t)\,dt = \gamma^2 \int (\alpha_0 + \alpha_1 \cos \Omega_a t)\,dt = \gamma^2 \alpha_0 t + \frac{\gamma^2 \alpha_1}{\Omega_a} \sin \Omega_a t$$

$$\tau = a\, e^{\gamma^2 \alpha_0 t}\, e^{\frac{\gamma^2 \alpha_1}{\Omega_a} \sin \Omega_a t} = \tau(t)$$

$$T(x,t) = X(x)\tau(t) = a\, e^{-\gamma x}\, e^{\left[\gamma^2 \alpha_0 t + \frac{\gamma^2 \alpha_1}{\Omega_a} \sin \Omega_a t\right]}$$

This must equal $T_0 \cos \Omega_a t$ or $T_0 \left(\frac{e^{i\Omega_a t} + e^{-i\Omega_a t}}{2}\right)$ at $x = 0$.

$a = T_0$ will work. But then we have

$$\frac{1}{2} e^{i\Omega_a t} = e^{[\gamma^2 \alpha_0 t + \frac{\gamma^2 \alpha_1}{\Omega_a} \sin \Omega_a t]} \quad \text{or} = e^{[\ ]}$$

If we put $i\Omega_a = \gamma^2 \alpha_0$, then we must have $\gamma^2/\Omega_a = -i/\alpha_0$

$$e^{\frac{\gamma^2 \alpha_1}{\Omega_a} \sin \Omega_a t} = 1 \qquad e^{-i \frac{\alpha_1}{\alpha_0} \sin \Omega_a t} = 1 \quad ?$$

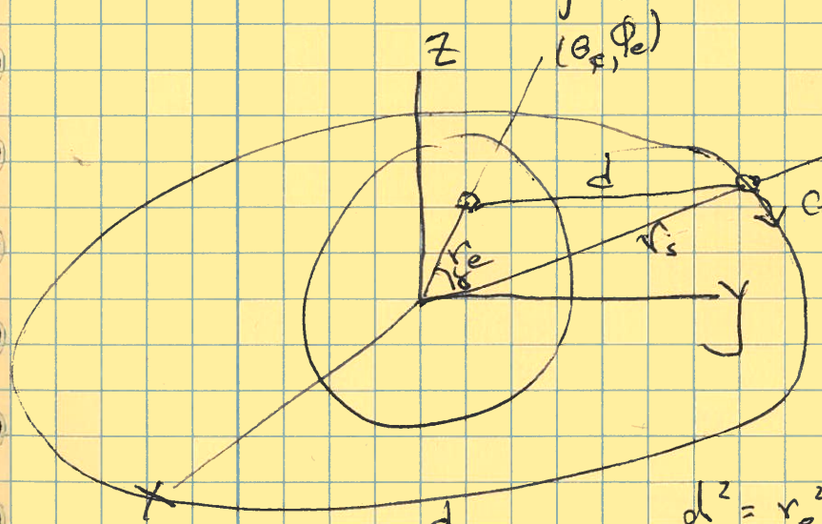
<u>No obvious simple solution.</u>

$e^{\frac{\gamma^2 \alpha_1}{\Omega_a} \sin \Omega_a t}$ has period Ω_a and so can be written as a Fourier Bessel series. It doesn't seem to help. i.e.

$$\text{at } x = 0 \quad e^{\gamma^2 (\alpha_0 t + \frac{\alpha_1}{\Omega_a} \sin \Omega_a t)} \propto e^{\gamma^2 \alpha_0 t} \sum_{n=-\infty}^{\infty} J_n\left(\frac{\gamma^2 \alpha_1}{\Omega_a}\right) e^{i n \Omega_a t} \overset{?}{=} e^{i\Omega_a t}$$

This seems to require that $J_n\left(\frac{\gamma^2 \alpha_1}{\Omega_a}\right) = 0$ for $n > 1$.

The effect of station position error on the observed radial velocity of a satellite, e.g. GPS

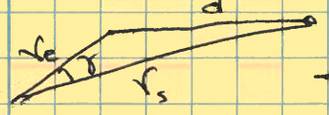


$d = ?$

γ is the angle between the two directions

$$\cos\gamma = \cos\theta_e \cos\theta_s + \sin\theta_e \sin\theta_s \cos(\phi_e - \phi_s)$$

$$d^2 = r_e^2 + r_s^2 - 2 r_e r_s \cos\gamma$$



The antenna has latitude l and longitude L

$$l = \pi/2 - \theta_e \ ; \quad L = \phi_e \ ; \quad \theta_s = \pi/2 - e_s \ ; \quad \phi_s = h_s$$

$$d = \cos(\pi/2 - e_s)\cos(\pi/2 - l) + \sin(\pi/2 - e_s)\sin(\pi/2 - l)\cos(L - h_s)$$

$$d = \sin e_s \sin l + \cos e_s \cos l \cos(L - h_s)$$

For an estimate of the order of the error, put the satellite into the equatorial plane. Then $e_s = 0$, $\sin e_s = 0$, $\cos e_s = 1$

continue on p. 181

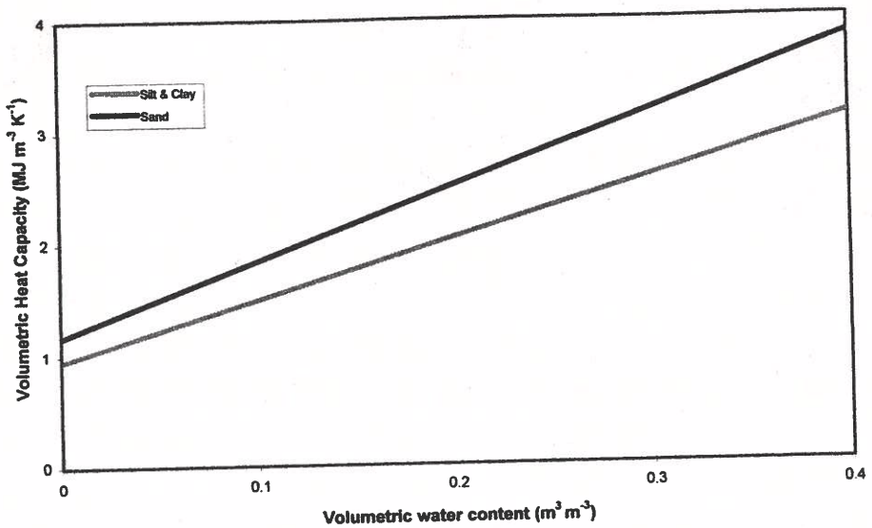


Figure 4. The volumetric heat capacity as a function of water content in three New Mexico field soils.

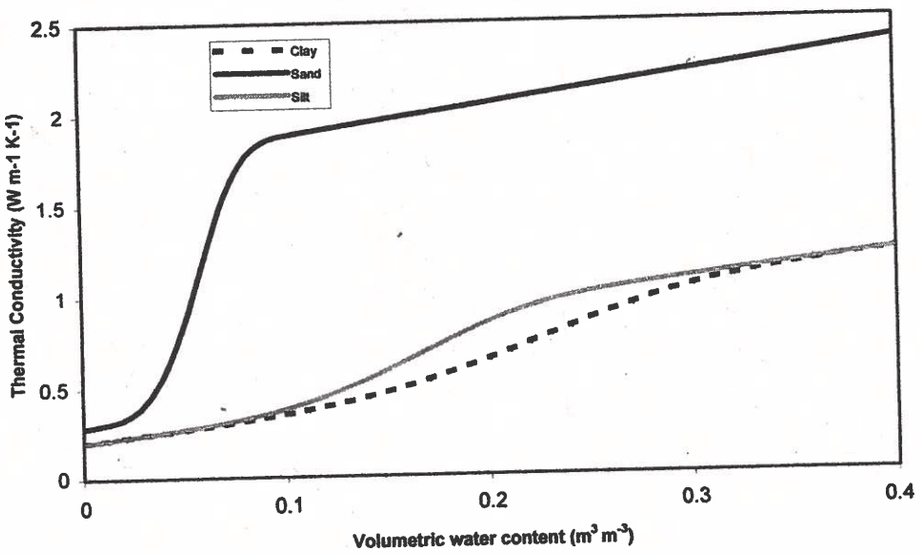
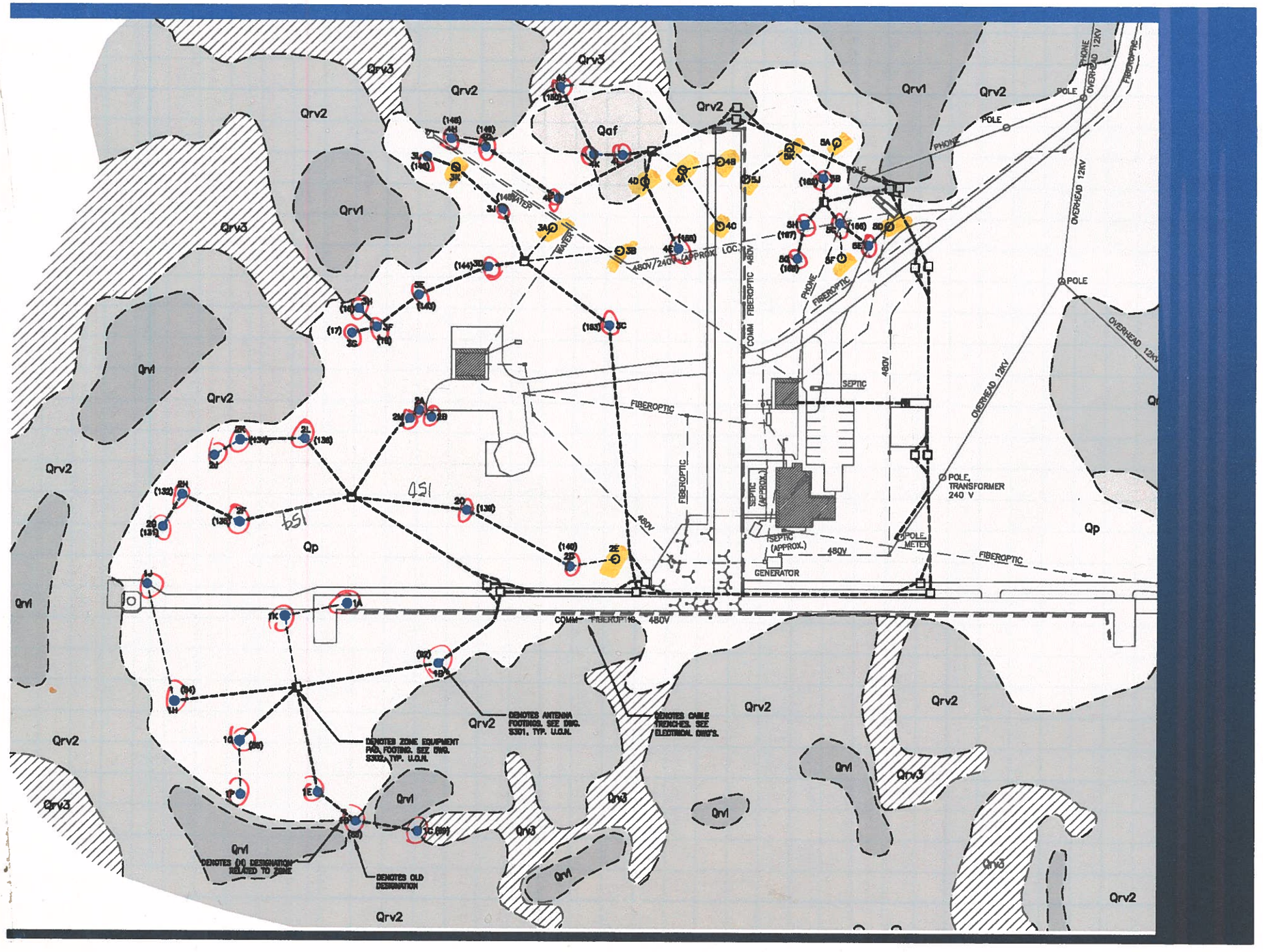


Figure 5. The thermal conductivity as a function of water content in three New Mexico field soils.

6 3/8" ⇒ 1000'


Qrv3
Qrv2
Qrv1
Qaf
Qp
WATER
480V/240V (APPROX. LOC.)
COMM FIBEROPTIC 480V
PHONE
FIBEROPTIC
SEPTIC
480V
OVERHEAD 12KV
POLE
POLE, TRANSFORMER 240 V
POLE, METER
SEPTIC (APPROX.)
GENERATOR
COMM FIBEROPTIC 480V
DENOTES ANTENNA FOOTINGS. SEE DWG. S301, TYP. U.O.N.
DENOTES CABLE TRENCHES. SEE ELECTRICAL DWG'S.
DENOTES ZONE EQUIPMENT PAD, FOOTING. SEE DWG. S302, TYP. U.O.N.
DENOTES (N) DESIGNATION RELATED TO ZONE
DENOTES OLD DESIGNATION
451

3/15/06

① Get lifetime vs. temperature for the FRAs

② Expect inlet temperature of 20°C

③ A student project for the reflection from: ⓐ a glass cylinder
ⓑ " " , w 2 anti-ref. layers.
ⓒ a glass sphere
ⓓ " " , w 2 anti-ref layers

④ Find out about glass bending to pyramid.

⑤ Get a schedule for Jim

✓ ⑥ Write the contract letter for Jim – make budget.

3/15/06 ATA properties

I. Independent pixels / primary beam (snapshot) one pointing
resolution @ HI (1420) is 70" or 1.17'
primary beam @ HI is 2.4° (142') 11,600 beams ~ 12,000
pixels from number of baselines $\frac{N(N-1)}{2} = 66,250$

II. Astrometry accuracy @ 10 GHz: $\theta_{1/2} = 8.4''$ With reasonable S/N, we should get ≤1 sec.; best case would be better.

III. Confusion limit 1.6 $\lambda^{2.1}$(m) mJy: (check this number)
ⓐ 1.4 GHz (.2m) → 0.4 mJy
ⓑ 5 GHz (.06m) → .032 mJy = 32 μJy
ⓒ 10 GHz (.03m) → 7 μJy
ⓓ .7 GHz (.4m) → 1.7 mJy

time to confusion with $T_{sys} = 50K$?

IV. More sensitivity calculations

i) One antenna Jy/K = 98/0.6 = 163 one polarization
115 dual polarization
For 350: 115/350 = 0.33 Jy for 1K.

ii) Time to confusion @ 1.4 GHz; 0.4 mJy → 1.3×10^{-4} K ΔT/pixel

$$\Delta T = \frac{T_{sys}}{\sqrt{BW \cdot t}} \Rightarrow t = \left(\frac{T_{sys}}{\Delta T}\right)^2 \times \frac{1}{BW} = \left(\frac{50}{1.3\times10^{-4}}\right)^2 \cdot \frac{1}{10^5}$$

$$= 1.5\times10^6 \text{ sec}$$

ⓐ $\lambda 6\,cm\ (5\,GHz)$; $.032\ mJy \rightarrow 3.2\times10^{-5}\,Jy \rightarrow 10^{-5}\,K$

$$\tau = \left(\frac{T_{sys}}{\Delta T}\right)^2 \frac{1}{BW} = \left(\frac{50}{10^{-5}}\right)^2 \frac{1}{10^5} = \underline{2.5\times10^{8}\ sec.}$$

From 127:

Then. $\cos\gamma = \cos l \cos(L-h_s)$

$$d^2 = r_e^2 + r_s^2 - 2r_e r_s \cos l \cos(L-h_s)$$

Assume $r_s \cong$ const, and r_e is constant

$$\cancel{2}d\dot{d} = \cancel{2} r_e r_s \cos l \sin(L-h_s)\dot{h}_s$$

$$\dot{d} = \frac{r_e r_s \cos l \sin(L-h_s)\dot{h}_s}{\left[r_e^2 + r_s^2 - 2r_e r_s \cos l \cos(L-h_s)\right]^{1/2}}$$

The maximum looks like its at $\sin(L-h_s) = 1$ i.e. $L = h_s$

Then $\dot{d} = \dfrac{r_e r_s \cos\gamma\, \dot{h}_s}{\sqrt{r_e^2 + r_s^2}}$

$r_s = R_{gps} = 26{,}000\,km$; $V_{gps} = 3.98\,km/sec.$; $r_e = 6400\,km$

$\dot{h}_s = \dfrac{3.98\ km/sec}{26{,}000\,km} = 1.5\times10^{-4}$ rad/sec

$$\dot{d} = \frac{(6400)(26{,}000)\cos l\,(1.5\times10^{-4})}{\sqrt{(6400)^2 + (26{,}000)^2}} \approx 1.0\cos l \quad km/sec.$$

Error $\delta\dot{d} = -1.0\sin l\ \delta l$

At $l \cong 45^\circ$ $\quad \delta\dot{d} = -0.7\,\delta l$

Suppose the error in l corresponds to 1 m

Then $\delta l = \dfrac{1\,m}{6400\,km} = \dfrac{1\,m}{6400\cdot 1000} = 1.6\times10^{-8}$

$\delta\dot{d} = 0.7(1.6\times10^{-8}) = \Delta V = 10^{-7}\,km/sec.$; $\dfrac{\Delta\nu}{\nu} = \dfrac{\Delta V}{c}$

$$\Delta\nu = \frac{\nu\,\Delta V}{c} = \frac{(1.5\times10^{9})(10^{-7})}{3\times10^{5}\ km/sec.} = \underline{0.5\times10^{-3}\,Hz}$$

System temperature tests with "final" feed. (active balun version)

I. First tests with plane 2'x2' absorber in front of feed compared with no absorber feed looking into optical system.

What's measured is $O_{amb}/O_{ant.} = T_{amb}/T_{ant.}$

a) for T_{amb}: The plane absorber fills just 2π solid angle, less at the lower frequencies. From I (p133), the integrated pattern of the feed past 90° is 8.5%. The top half goes to the sky either directly or after reflection from the primary and secondary. With the doors open about 1/2 goes to the ground and half to the sky. The net effect is 1 3/4 x 8.5% goes to the sky

The real ambient temperature is 50F or 10C → 283K. Making just the above back lobe correction, we get

$$283\left(1-\tfrac{3}{4}(.085)\right) + 6\left(\tfrac{3}{4}(.085)\right) = 265.3\,K$$

b) For T_{sky} (p.47) include the brightness of the atmosphere and all the antenna contributions.

i) Geometrical optics spillover $\sim 3K\left(1-\frac{z}{90°}\right)$
ii) Diffractive spillover $8K/\sqrt{F}$
iii) CMB $\sim 3K$
iv) atmospheric emission, typically 3K sec z (J. Ball '86)

$$T_{sky} = 3K\left(1-\frac{z}{90°}\right) + 3K + 8K/\sqrt{F_{GHz}} + 3K\sec z$$

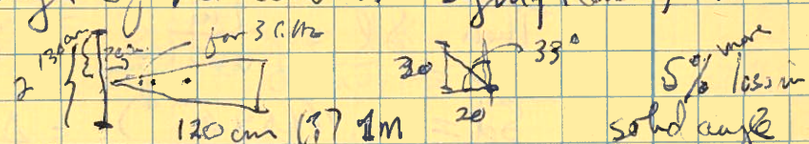
(?)

At z = 45° (sec z = 1.41) and f = 3 GHz

$$3\left(1-\tfrac{1}{2}\right) + 3K + 4.6 + 3(1.4) = 13.3K = 13.3\,K$$

At 2 GHz Get ~ 14K

At 3 GHz the solid angle of the load is significantly less than 2π



This lowers "T_{amb}" by another 15% (?) scale by .75 again

$$.75 \times 13.5\% = 10.1\%$$

From the Y factor: $Y = \frac{O_{amb}}{O_{sky}} = \frac{T_{amb} + T_{sys}}{T_{sky} + T_{sys}}$ for the feed

so $T_{sys} = \frac{\text{"}T_{amb}\text{"} - Y T_{sky}}{Y-1}$

For $T_{sky} = 15$ and "T_{amb}" = 265

$$T_{sy} = \frac{265 - Y15}{Y-1}$$

For $Y = 7db$ (5) $\quad T_{sys} = 47.5$

For $Y = 5db$ (3.2) $\quad T_{sys} = 99K$

At 3GHz, "T_{amb}" → 265 − 15 = 252 with $T_{sky} = 13.3$

$$T_{sys} = \frac{250 - Y13.3}{Y-1} = \frac{250 - 5(13.3)}{5-1} = \underline{46K} \text{ ok.}$$

At 2GHz load solid angle

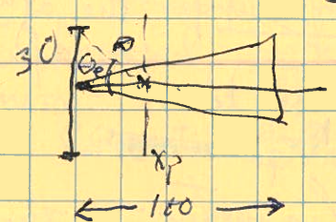
close to edge of sky

Integrated spillover past the load is 18% $\times \frac{3}{4}$ for open doors is 13.5%. So "T_{amb}" = $283(1-.135) + 6'(.135) = \underline{245K}$

Also, $T_{sky} = 14K$

$$T_{sys} = \frac{245 - Y14}{Y-1}; \text{ for } Y = 5(7db), \; T_{sys} = \underline{43.7°K}$$

For several wavelengths find the spillover solid angle



$\theta' = 90° + \tan^{-1} x_p/30$

ν	x_p	$\tan^{-1} \frac{x_p}{30}$	θ_e	spillover $\int_{\theta_e}^{\theta=180} \lvert\psi(\theta')\rvert^2 d\Omega$	259 ×.75	283/+(1−.159)
1	50	59°	31°	.32	.24	215
2	25	40°	50°	.143	.107	258
3	17	30°	60°	.133	.099	258
4	13	23°	67°	.118	.085	259
5	10	11°	79°	.104	.078	261

Assume the load temp is 283K.

The last column is the expected load temperature. The asymptote value will be 265 for higher freqs.

Getting an antenna temperature Scale for an ATA antenna

I. One way is to use the 2'x2' absorber described above.

a) For the antenna temp. on blank sky, use

$$T_{A\,sky} = 3K\left(1-\frac{z}{90^\circ}\right) + 2.8K + 8K/\sqrt{f_{GHz}} + 3K\sec z.$$

The first and last terms add to a nearly z independent and frequency independent 6K. So use

$$\boxed{T_{A\,sky} \cong 8.8 + 8.0/\sqrt{f_{GHz}}}$$ independent of zenith angle.

b) For the temperature when the load is in place, use the table from the previous page.

ν	T_{load}
1	215
2	253
3	255
4	259
5	261
10	265

This is for ambient temperature of 50°F (10C, 283K). It should be scaled.

II The other way is to observe the moon

$$T_{AM} = \eta_B T_M \left(1 - e^{-.69\left(\frac{\theta_M}{\theta_{1/2}}\right)^2}\right)$$

$T_M = 230K$ $\boxed{\theta_{1/2} = 2.33^\circ\,(\lambda/.2m)}$ for the (gaussian fit)

$$\eta_B = 0.88\left(1-.052/\sqrt{f}\right)\left(.97-.023\sqrt{f}\right)e^{-(.0226f)^2}$$

~~gaussian~~ f in GHz

It requires the moon's current diameter. The diameter varies by a maximum of 10% in radius, or 20% in area.

$\eta_B(1GHz) = .79$; $\eta_B(3GHz) = .79$; $\eta_B(5GHz) = .78$;

$\eta_B(7GHz) = .76$ $\eta_B(9GHz) = .75$

all about the same $\langle\rangle = .78$

A few examples from the Moon

$T_M = 230\,K$ (Troitski)

Use just the mean lunar diameter for these numbers. For any particular day the moon's actual diameter must be used. $\langle \Theta_m \rangle \cong 0.5^\circ$

For the Gaussian fit beam. $\Theta_{1/2} = 2.33^\circ \left(\frac{\lambda m}{.2 m}\right)$

$\nu = 1\,GHz$ $\quad \Theta_{1/2} = 2.33\left(\frac{0.3}{0.2}\right) = 3.^\circ 459$

$T_{AM} = .79\,(230)\,(1 - e^{-.69\,(\Theta_M/\Theta_{1/2})^2})$

$= .79\,(230)\,(1 - e^{-.69\,(.5/3.459)^2}) = 182\,(1 - .986) = \underline{2.55^\circ K}$

$\nu = 1.4\,GHz$ $\quad \Theta_{1/2} = 2.^\circ 33$

$T_{AM} = 182\,(1 - e^{-.69\,(.5/2.33)^2}) = 182\,(1 - .9687) = \underline{5.69\,K}$

$\nu = 3\,GHz$ $\quad \Theta_{1/2} = 2.33\,(1/.2) = 1.^\circ 165$ $\qquad$ 1.6db

$T_{AM} = .79\,(230)\,(1 - e^{-.69\,(.5/1.165)^2}) = 182\,(.1194) = \underline{21.7^\circ K}$

$\nu = 5\,GHz$ $\quad \Theta_{1/2} = 2.33\,(.06/.2) = .699^\circ$ $\qquad$ 2.2db 2.5

$T_{AM} = .78\,(230)\,(1 - e^{-.69\,(.5/.7)^2}) = 180\,(1 - .703) = \underline{53.4^\circ K}$

$\lambda = 5\,cm.$ $\quad \nu = 6\,GHz$ $\quad = .77\,(230)\,(1 - e^{-.69\,(.5/.583)^2}) = 70.4$ $\qquad$ 3.0db

$\Theta_{1/2} = 2.33\,(.05/.2) = .583^\circ$

$\nu = 8\,GHz$ $\qquad$ 8.7

$2\,GHz$ $\quad \Theta_{1/2} = 2.33\,(.15/.2) = 1.748^\circ$

$T_{AM} = .79\,(230)\,(1 - e^{-.69\,(.5/1.748)^2}) = 9.97^\circ$

For $F = 1.0\,db\ (1.26)$ $\quad T_M/T_{sy} = .26$, $T_{sy} = 38.3\,K$

$\nu = 10\,GHz$

~~To account for the loss of signal from the sky background when one observes the moon and the moon blocks the background, subtract the term $8.8\,e^{-.69\,(\Theta_M/\Theta_{1/2})^2}$ from the formula on the previous page~~

$$T_{AM} = \eta_B \left\{ T_M \left(1 - e^{-.69\,(\Theta_M/\Theta_{1/2})^2}\right) - 8.8\,e^{-.69\,(\Theta_M/\Theta_{1/2})^2} \right\}$$

$$= \eta_B \left\{ T_M - (T_M + 8.8)\,e^{-.69\,(\Theta_M/\Theta_{1/2})^2} \right\}$$

no

Power levels for the LO downconverters JRF and WJW

For LO#1 and LO#2

At the LO splitter outputs there should be ~~[illegible]~~
- 0dbm @ LO #1 and
- 6 dbm @ LO #2

} These are also the inputs to the Boards.

Each 8 way splitter has ~ 11db of gain.
So the output @ LO1 synthesizer is −15dbm
LO2 " " −2.0 dbm

For these tests the type of antenna 5b4 was used.

Channel A had no extra attenuation before the mixer #1
D did have a 10 db pad before its mixer #1.

Various attenuator settings were used in the PAM
We used both a broad band power meter and the Spectrum analyzer following the DCB. ΔV=3MHz for SA. SAmple mode

PAM attenuators

power e.g. (15,15) 30 total		PM	SA Peak
(15,15) 30 total	chan D	−18.5	−36
	chan A	−11.4	−28
(0,0) 0 total	chan D	+5.8	−15
	A	+5.0	−16
(5,5) 10 total	chan D	6.6	−20
	chan A	5.2	−21
(10,10) 20 total	chan D	−10.1	−27
	chan A	−2.4	−20
repeat (15,15) 30 total	chan D	−19	−36
	chan A	−11.4	−29

At (15,15) mixer 1b not overdriven in either A or D

Changing the input power from (10,10) 20db, to (15,15) 30 db channel A went -20db to -29 db and D went -27 to -36. No evidence of anything over driven in either A or D

Changing the input power from (5,5) 10db, to (10,10) 20db, channel A hardly changed: -19 to -20 db while D changed -21 to -27. There is considerable non-linearity in channel A (with no pad) and less in channel D.

To use channel A requires the 20db of pad in place for no limiting.

Conclusion: Having the 10db pads in place as for channel D allows more head room against overdriving mixer #2, where the non linearities occur

With the 10db pads in place, a level of (10,10) 20db, gives enough head room w.r.t. the mixer. If it were removed, as above, mixer #1 would be over driven.

Another advantage of the pad at the output of the power splitter before the mixer #1 is that it makes a better input match to the mixer. Without the 10db pad, there is a bare low pass filter at the output of the signal power splitter output.

In the pad free state, Koch and I noticed that changing the LO1 frequency for one band made a small change in the band pass shape for another band.

Spectrum Analyzer

looking at c drive

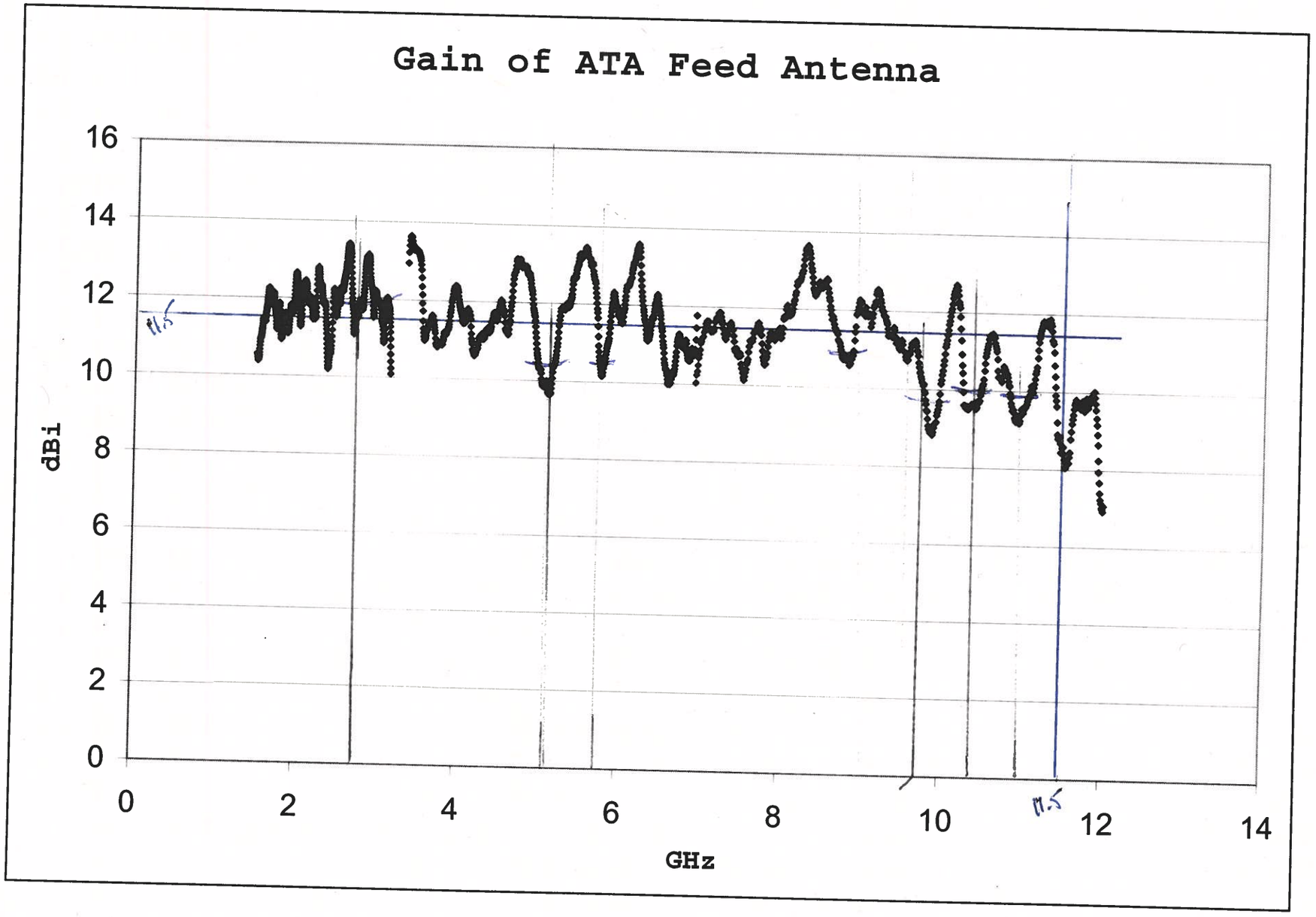
Steps: File (soft keys) Load states

list a) HI (in linear mode, may need reference adjustment
b) GPS off works in sidelobes
c) Broadband

---

atacheck source e.g moon plots circumstances

work from data directory
cd .. /0730
ls works.
ataautstatus brings up a double GUI
atasetazel (aut) (az) (el) /


Gain of ATA Feed Antenna
16
14
12
10
8
6
4
2
0
dBi
11.5
0
2
4
6
8
10
11.5
12
14
GHz

The gain curve of the previous page was made using the standard horns, whose effective areas are given on pp 112–113. The gain formula is given on p. 103.

Hat Creek Computing control

To login from Berkeley

ssh -X obs@maincntl.seti.org pwd - berkeley e => 2

(for all obs)

prompt is obs@maincntl: -X enables looking at Gui's and pictures

pwd: bzr ----

to go to next machine: ssh obs@labusr1

type tsch to get Rick's stuff.

I. skyfreq change: atasetskyfreq d 1420.4 (freq. in MHz)

(have control over a, c, and d; not b)

II. pointing antennas: atasetazel 2f, 2k, 2l "az" "el"

(for example 388° 41° for pole)

to point at a particular source:

i) atacheck moon (e.g.)

it gives the az and el and whether/not it's up

ii) for ephemeris atacephem "source" e.g. sun, moon, or Jupiter

iii) type ls find the files that are there

cd data/jack {blue are directories; green are executables; black are files}

iv) make an ephemeris file here: atacephem sun does it

v) to inspect a file, type "less" filename; e.g. sun.ephem

get "time", "az", and "el"

vi) to use the ephemeris: atatrackephem 3c sources.ephem

(↑ auto)

3c didn't move in azimuth type atasevrostatus

III. Receiver status (back end status)

browser http://hcro.berkeley.edu/~rick/ {travels w rick & Carl Hat Creek Obs

click, then ata backends.

hydrogen will be the c LO.

atafx4spectrum -n --queue help: with no arguments

IV further commands

i) atacascii status

ii) atacmds

typing this will list many commands.
typing a command alone will
produce a list of all the options

Air Flow into the rim box measured by Jeff:

For 5G the flow is 3500 ft/min $= v = \dot{\ell}$; $\ell \cong 1'' = \frac{1}{12}$ ft

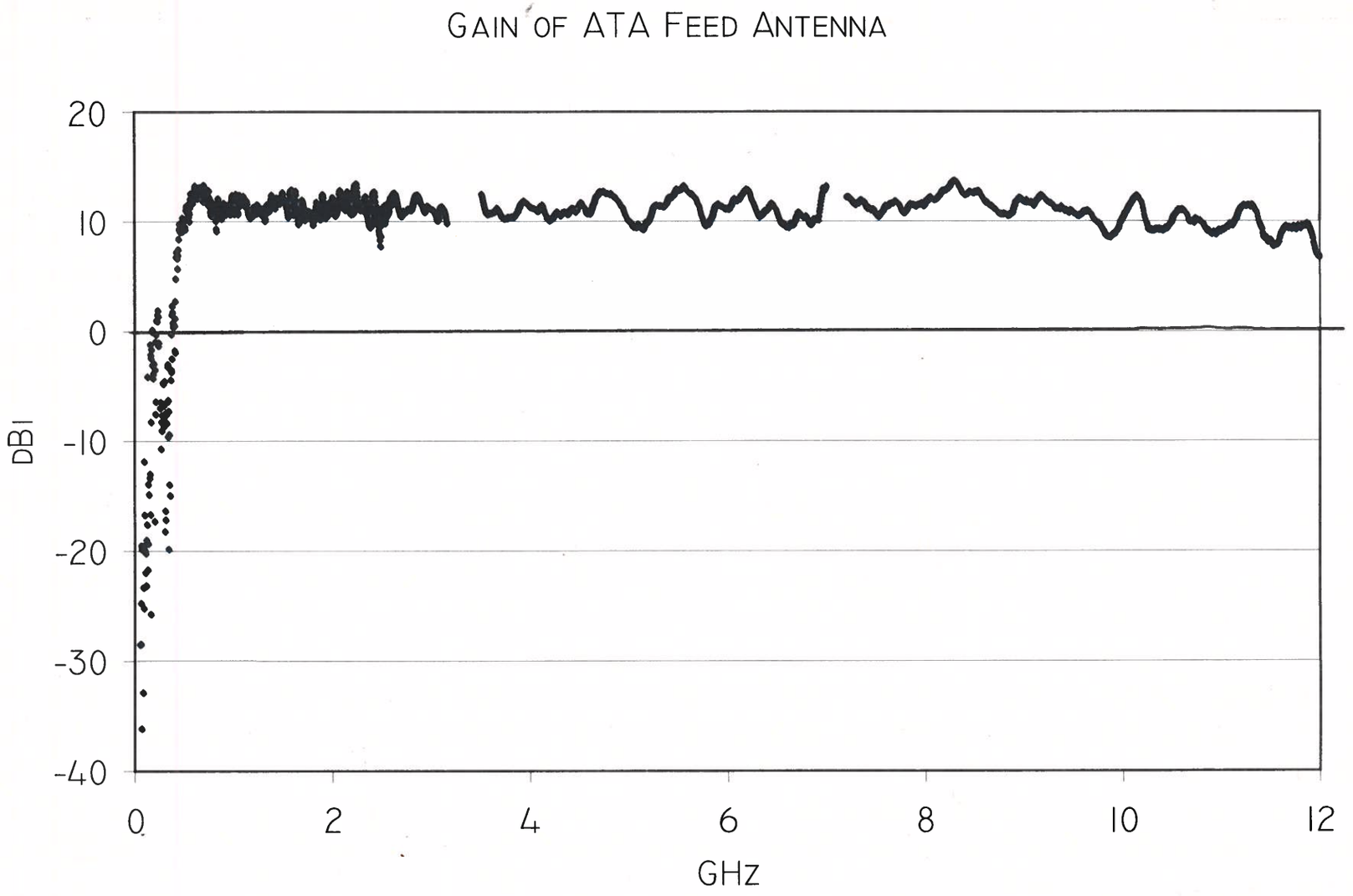
$V = \frac{\pi}{4} d^2 \ell$; $\dot{V} = \frac{\pi}{4} d^2 \dot{\ell} = \frac{\pi}{4}\left(\frac{1}{12}\right)^2 3500 = \underline{\underline{19 \text{ CFM}}}$

Power Flow: $P = (.336)\,\text{CFM}\,\Delta T(^\circ F)$ $\boxed{P.86}$

$= (.336)(19)\,20^\circ F = 128$ watts

Allow for some blockage by the wires, e.g. 2/3, and CFM → 12 CFM

Then $P = (.336)(12)(20) = \underline{80 \text{ watts}}$. This should be more than adequate



$G_0 = 11.5 \pm 1 db$

Effects of the Feed Gain Variation on the Gain and Beamwidth of the overall optical system

There are no large input mismatches for the feed alone. (see p. 116) Some of the drops in the transmission can be attributed to the tip circuit bound mismatches, but it is not much. Hence the variation gain must be a real effect of the feed. Since $\int g \, d\Omega = 4\pi$, $G_0 \Delta\Omega_{1/2} \cong 4\pi$.

When G_0 drops, $\Delta\Omega_{1/2}$ must go up. The effect of this will be a better edge illumination of the primary but more spillover. Also, a better edge illumination will narrow the main lobe from the mirror.

An estimate of these effects. $G_1 \Delta\Omega_1 = G_2 \Delta\Omega_2$

The nominal solid angle to the edge is at 42° on the feed pattern. If the Gain increases so that the same illumination falls at 38°, this feed angle change of ~10% decrease corresponds to a solid angle decrease of 20%. Thus the forward gain of the feed must have increased by 20%, about 1db.

To first order it is the same as if we had kept the edge of the dish at 42° and lowered the edge illumination intensity to what it had been at 46°. Thus we change the edge illumination from .211 to .139. From table of p 76 (I) the illumination efficiency goes from 0.831 to 0.739, a decrease in illumination efficiency of 9%.

At the same time the spillover fraction past the edge decreases from .182 to .157, a decrease of 2.5%, or an increase in gain by 2.5%, a net loss in gain of about $\boxed{6.5\%}$. This is about 1/3 of the feed gain increase.

For a forward feed gain decrease of 20%, the feed pattern fattens by 20% in solid angle or 10% in radial angle. This is equivalent to increasing the edge illumination (at 42° edge) to what it had been at 38°. The spillover goes from .182% to .224% and the illumination efficiency goes from .831 to .893. The net change is an increase in overall gain of $\boxed{\text{about } 2\%}$

The two effects cancel each other out to some extent and overall the system gain remains nearly constant.

The other effect is a slight change in the width of the final pattern with the ~~power~~ gain of the feed.

We can use three distributions of the aperture field from p. 142.

For the nominal distribution, −13.5 db at the edge

$$E(r) = .211 + .786\left[1-(r/3)^2\right]^{1.4} \qquad F = .65$$

$$E(\theta) = \int_0^3 \left\{ .211 + .786\left[1-(r/3)^2\right]^{1.4} \right\} J_0(kr\sin\theta)\, r\,dr$$

$$= \left\{ 1.899\,\frac{J_1[94.25\sin\theta]}{94.25\sin\theta} + 48.23\,\frac{J_{2.9}[94.25\sin\theta]}{[94.25\sin\theta]^{2.9}} \right\}$$

This for $\lambda = 0.2$ m. Plotting the square of this, the power pattern, we find FWHM = $\underline{\underline{2.327^\circ}}$ ←

For the 20% lower feed gain case, the distribution is

$$E(r) = .299 + .701\left[1-(r/3)^2\right]^{1.9} \qquad F = .75$$

In this case, the FWHM = $\underline{\underline{2.327^\circ}}$, $\underline{\underline{3.8\%}}$ narrower

For the 20% higher feed gain case, the distribution is

$$E(r) = .139 + .860\left[1-(r/3)^2\right]^{1.9}$$

In this case, the FWHM = $\underline{\underline{2.458^\circ}}$, $\underline{\underline{5.6\%}}$ wider.

The nominal beam width is 2.327° at $\lambda = 0.2$ m

So, in general $\boxed{\Theta_{1/2} = 2.327^\circ \dfrac{\lambda(\text{cm})}{2\Theta}}$

except for small effects of illumination above.
i.e. ± 4.5%

Table in /vals/welch/HT/ATA_tests/ATAfeed/feed/feed_table.pdf

Temperature Stability measurements on the PAX Temperature

observe the PAM temperature on 18 antennas (no slow sampling)

ⓐ Over a 5 second interval not one temperature changed by 0.1°C, the read out resolution

ⓑ Over a one minute interval just two temperatures changed by 0.1°

A rough estimate would be an rms temperature change of ~.03°C for a 5 second interval

Gain stability for OTF total power observing using auto correlation. p 152(II)

For the HI frequency, 1.4 GHz, with $\Theta_{1/2} \cong 2.5°$ the slew rate can be 2.5°/second. At ½ beam spacing, that's one new independent point every 0.5 seconds of time so t_s = 0.5 seconds. We want the thermal fluctuations, not the gain fluctuations, to dominate.

For continuum BW = 100 MHz

The fractional noise in each 0.5 second is

$$\frac{\Delta T_n}{T_s} = \frac{1}{[10^8 \cdot 0.5]^{1/2}} = .00014 \sim 10^{-4} \quad \text{(thermal)}$$

For a 10° x 10° field (e.g. the M31 field) the sweep will be ~ 5 seconds in duration with a turn around time of 5 seconds. The average time between each pixel and the off time is about 5 seconds. So $\frac{\Delta G}{G} \lesssim .00014$ (modest gain)

→ over a time interval of ~ 5 seconds. Plambeck's measurements of the PAM gave .02db/1° → .0005/0.1°C ✓

So an RMS of .03°C would give us an RMS of

$$\frac{\Delta G}{G} \sim .0005(.3) = \underline{.00015} \text{ for 5 seconds.}$$

and this is comparable to the thermal fluctuations. Also we can make many single dish maps with many dishes.

For observations of an HI line, 1000 channels over 10 MHz, $\Delta\nu = 10\,MHz/1000 = .01\,MHz$
$\times 200 = 2.1\,Km/sec$ resolution

.01 MHz is smaller by the factor 10000 from 100 MHz. Hence the thermal noise is higher by 100 times than for the continuum and $\frac{\Delta G}{G} \sim .015$, very easy.

Again the result of many antennas can be averaged.

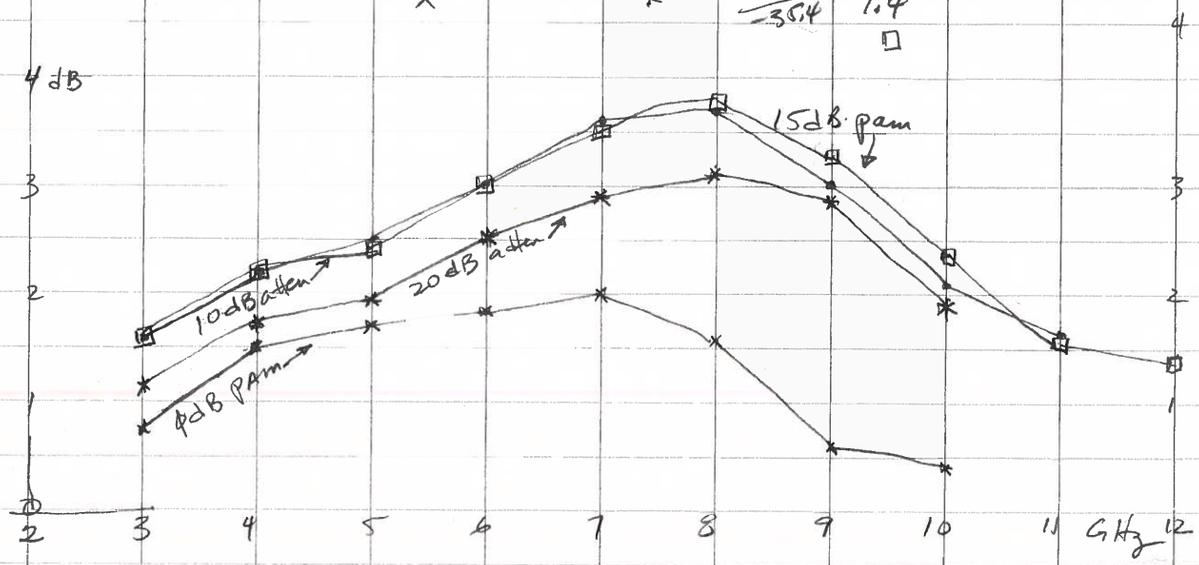
Moon ON/OFF ⇒ measured with Power Meter at output of ATA RFCB output (LO A) changing PAM ATTEN AT Feed 0, 10, 15, 20 dB

midnight – 8-9th Nov '06

Moon – Thru ATA DC – 2K-X pol

Moon elev ~ 48°
off direction = 5°

ATTEN =	PAM 10		φ		20		15	
GHz	dB ON/OFF							
1								
2	XM in band						-19.3 / -19.2	-0.1 XM
3	-12.5 / -14.1	1.6	-5.6 / -6.4	0.8	-21.5 / -22.7	1.2	-16.9 / -18.5	1.6
4	-12.8 / -15.0	2.2	-4.9 / -6.4	1.5	-21.8 / -23.5	1.7	-17.2 / -19.4	2.2
5	-13.4 / -15.9	2.5	-5.9 / -7.6	1.7	-22.5 / -24.4	1.9	-17.8 / -20.2	2.4
6	-11.5 / -14.5	3.0	-4.6 / -6.4	1.8	-20.4 / -22.9	2.5	-16.1 / -19.1	3.0
7	-10.7 / -14.3	3.6	-4.6 / -6.6	2.0	-19.6 / -22.5	2.9	-15.1 / -18.6	3.5
8	-8.1 / -11.8	3.7	-3.6 / -5.2	1.6	-17.2 / -20.4	3.2	-12.3 / -16.1	3.8
	-8.2 / -11.2	3.0	-5.0 / -5.6	0.6	-17.0 / -19.8	2.8	-12.3 / -15.6	3.3
10	-9.7 / -11.8	2.1	-7.2 / -6.8	0.4	-18.4 / -20.3	1.9	-13.7 / -16.0	2.3
11	-22.7 / -24.4	1.7	-17.9 / -18.6				-26.1 / -27.7	1.6
12		•		×		*	-34.0 / -35.4	1.4 □



To get the T_{sys} from observing the moon we use the calculated η_B from p. 48

$$\eta_B = 0.88(1-.052/\sqrt{f})(.97-.023\sqrt{f})\, e^{-(.0226f)^2}$$

along with the moon observing formula:

$$T_{AM} = \eta_B T_M \left(1 - e^{-.69\left(\frac{\Theta_M}{\Theta_{1/2}}\right)^2}\right) \qquad \langle\Theta_M\rangle \cong 0.5^\circ$$

$\overline{T_{AM}} = \overline{T_M} = 230K$ (Troitskii) $\qquad \Theta_{1/2} = 2.33^\circ\left(\frac{\lambda}{.2m}\right)$ ok for long λ.

We can also compare the derived T_{sys} with expected quantity from p. 48 $\quad {}^cT_{sys} = \frac{7.8}{\sqrt{f}} + 6.6\sqrt{f} + 16$

$$Y = \frac{T_{sys} + T_{AM}}{T_{sys}} = 1 + \frac{T_{AM}}{T_{sys}}$$

ν	η_B	Y	T_{AM}	η_{ap}	T_{sys}	${}^cT_{sys}$	η_B'	T_{AM}'	T_{sys}'	${}^cT_{sys}'$	η_{ap}'
1	.79		2.5	.60			0.75			43	.57
2	.79		10.0	.60			0.75				.57
3	.79	1.6db	21.7	.60	48	32	0.74	20.3	45.1	47.9	.56
4	.785	2.2	36.4	.60	55	33	0.73	33.8	51.2	48.4	.56
5	.78	2.5	53.4	.59	68.5	34.3	0.72	49.3	63.2	52.5	.55
6	.78	3.0	71.4	.59	71.4	35.4	0.71	65.0	65.0	54.6	.54
7	.76	3.5	87.3	.58	70.5	36.4	0.70	80.4	64.8	56.6	.53
8	.75	3.8	94.2	.57	67.0	37.4	0.69	86.7	62.0	58.4	.53
9	.75	3.3	117.6	.57	103	38.4	0.68	106.7	93.6	60.2	.52
10	.74	2.3	128.2	.56	186		0.67	116.0	168		.51
11		1.6									
12		1.4									

see p. 48

The higher than predicted T_{sys} suggests that there may be more room temperature ohmic losses in the feed than expected. Increasing such losses will increase the predicted T_{sys} and lower the η_B. (No, the
As a trial, we double the frequency independent feed loss from 3% to 8% and increase the ohmic $\sqrt{f}$ losses by the factor 1.7.

$$\eta_B' = 0.88(1-.052/\sqrt{f})(.94-.04\sqrt{f})\, e^{-(.0226f)^2} \qquad \eta_{ap} = .761\,\eta_B$$

The new theoretical: ${}^cT_{sys}' = \frac{7.8}{\sqrt{f}} + 11.2\sqrt{f} + 24$

This all ~~assumes~~ is for a fixed focus, probably set for 6 GHz
☆ These numbers are close to the feed measurements of p. 115. The issue of the OTX noise floor may still be a problem here. (1/10)

To get the T_{sys} from observing the moon we use the calculated η_B from p. 48

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along with the moon observing formula.

$$T_{AM} = \eta_B T_M \left(1 - e^{-.69\left(\frac{\Theta_M}{\Theta_{1/2}}\right)^2}\right) \qquad \langle\Theta_M\rangle \cong 0.5^\circ$$

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We can also compare the derived T_{sys} with expected quantity from p. 48 $\quad {}^cT_{sys} = \frac{7.8}{\sqrt{f}} + 6.6\sqrt{f} + 16$

$$Y = \frac{T_{sys} + T_{AM}}{T_{sys}} = 1 + \frac{T_{AM}}{T_{sys}}$$

ν	η_B	Y	T_{AM}	η_{ap}	T_{sys}	${}^cT_{sys}$	η_B'	T_{AM}'	T_{sys}'	${}^cT_{sys}'$	η_{ap}'
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4	.785	2.2	36.4	.60	55	33	0.73	33.8	51.2	48.4	.56
5	.78	2.5	53.4	.59	68.5	34.3	0.72	49.3	63.2	52.5	.55
6	.78	3.0	71.4	.59	71.4	35.4	0.71	65.0	65.0	54.6	.54
7	.76	3.5	87.3	.58	70.5	36.4	0.70	80.4	64.8	56.6	.53
8	.75	3.8	94.2	.57	67.0	37.4	0.69	86.7	62.0	58.4	.53
9	.75	3.3	117.6	.57	103	38.4	0.68	106.7	93.6	60.2	.52
10	.74	2.3	128.2	.56	186		0.67	116.0	168		.51
11		1.6									
12		1.4									

see p. 48

The higher than predicted T_{sys} suggests that there may be more room temperature ohmic losses in the feed than expected. Increasing such losses will increase the predicted T_{sys} and lower the η_B. (No, th
As a trial, we double the frequency independent feed loss from 3% to 8% and increase the ohmic $\sqrt{f}$ losses by the factor 1.7.

$$\eta_B' = 0.88(1-.052/\sqrt{f})(.94-.04\sqrt{f})\, e^{-(.0226f)^2} \qquad \eta_{ap} = .761\,\eta_B$$

The new theoretical: ${}^cT_{sys}' = \frac{7.8}{\sqrt{f}} + 11.2\sqrt{f} + 24$

This all ~~assumes~~ is for a fixed focus, probably set for 6 GHz
☆ These numbers are close to the feed measurements of p. 115. The issue of the OTX noise floor may still be a problem here. (1/0)

Central Discussion — what to monitor Dave's 3 page spread sheet

1. Matt brought a blower to use when the alidade cover is off.

Gains and noise figure for RFCB (ATA down converter)

I. Power levels "Down Converter" diagram shows gain levels and power levels. One omission is the −13 db power loss through the 15.51 GHz, 600 MHz filter. Where the output shows −1 db, it should be −14 db approximately. The measured value shown on the previous page is close to that.

II. Noise figure estimate, starting from the back. Assume $\langle N \rangle = 4$ for all the 412 stages.

Use the basic relation: $N \; [G_1 \mid G_2]$ (N_1, N_2) $N = N_1 + \frac{N_2 - 1}{G_1}$

"Down 4 box" assume the Serenza is used with $N = 4$.

Referred to the input of the second mixer ($L = 8$ db, 6.3)

1. $N = 1 + 6.3(4-1) = 20$

2. At input of first amp: $N = 4$ and $G = 100$
$$N = 4 + \frac{20-1}{100} = 4.2$$

---

3. At input of first mixer with mixer loss = 8 db and filter loss = 4 → 12 i
$$N = 1 + 15.8(4.2 - 1) = 51.7$$

4. At input to LP filter w/ 2 db loss in filter and 10 db pad → 12
$$N = 1 + 15.8(51.7 - 1) = 802$$

5. At input to final stage of power divider with 12 db gain
$$N = 4 + \frac{802-1}{15.8} = 54.7$$

6. At input of previous stage with 2 stages of −11 db of loss (~17 db)
$$N = 4 + \frac{54.7-1}{1} = 57.7$$

7. At input of second stage $N = 4 + \frac{57.7-1}{15.8} = 7.8$

8. At input of first stage $N \cong 4 + \frac{7.8-1}{15} = \boxed{4.5}$

Adding one more stage at the input will only slightly improve the noise figure. Adding no more gain here will leave a good output margin: ~~~ 8db with P1db = +7 for the Serenga.

Now move back to the PAM. With 30db loss in the optical system, the effective noise figure at the input to the OTX is, $N = 1 + 1000(4.5-1) = \underline{\underline{3501}}$

Leaving out any attenuation in the PAM will leave 6 stages with at least 50 db of gain. N at the input to the PAM would then be $N \cong 4 + \frac{3501-1}{100,000} = \underline{\underline{4.03}}$

12/7/06

~~This is not correct~~. The OTX/RCVR system has 30db loss but it's noise figure is closer to 45 db, according to both JD's spreadsheet and the Phot Eng. presentation. Hence, we should write $N = 15db$ and $\Delta = 30$ db.

So $N = 32 + 1000(4.5-1)$ [N has now been measured with $2.7 < N < 7.0$ over 1000 - 10,000 MHz. See Below]

$N = 32 + 3501 = 3532$. The extra 15db doesn't make much difference.

At the highest frequency $N-1 = 9$. Then, at the input to the OTX, $N = 32 + 9000 = 9032$.

The PAM has 6 412's, each with $g \approx 10$ and $N \sim 4$

a) Into the final one $N = 4 + \frac{9000-1}{10} = 904$

b) Into the previous stage $N = 4 + \frac{904-1}{10} = 94$

c) With 20 db of pad in, we are back to 9000 at the input to the pad

d) Then we are at 94, two more stages up and 13 just after the first attenuator

Then, with no attenuation we should be at $4 + \frac{13-1}{10} = 5.2$ at the input to the PAM (check N for the 412's at 10 GHz)