Hello Jed -

The PIN for the
SE71 security card
is: 0100#

Olé

John Luyten 925 930 7336

1-3 Matt's PDR Dec. 1 2000
4 Revision of Offset Gregorian Gain Table
4-7 Spillover calculations: Cassegrain, Gregorian, Prime Focus; skirt on 7
8 Ground Screen
10 Spillover plot: Cass, Gregorian, Prime Focus
11 Pattern Comparisons
13 Shroud picture
14 Crude Shroud/Feed Interaction
16 Careful Shroud/Feed Interaction
19 Direct feed tip reflection
20 Direct backlobe reflection
21 Insulation Properties of spherical foam ball
22 Jan 15 ATA board 2001
22 Heat conduction along pyramid
24 skin depth for copper; successive thinning of pyramid walls
25 Thermal conductivity of Copper — conductivity in skin
27 Run of temperature in foam
29 Inductance of straight wire in space
30 Matt's design for antenna 3/6/01
31 Size of Bend Waists
32 Input connection schemes
34 up/down converter meeting
35 Possible Passive Balun for the Unbalanced LNA Amps
36 LNA Temp vs. physical temperature
37 More on cooling RaRoftip
40 More measurements on passive balun: phase 7/23/01
41 sensitivity of balun to nearby grounds.
42 Tests of Krytar 3db coupler
42 potential phase switch for 1st LO
46 tests of double balun S_{11} and S_{22}
49 Effect of phase switching the first LO
52 MOU board Sept. 17, 2001
53 effect of metal bar on top of balun
56 feed tests at JPL
57 High Tc superconductor
58 Top circuit board design
60 interior of feed drawing
61 11/12/01 Sandy's JPL array plans; first active balun design
62 gain stability of LNA; transmission line formulas
64 One dimensional heat flow for K(T). Cu and SS
68 First noise calculation for input sections
70 Low temperature thermal conductivities

72 Equiangular Spiral for input calibration
82 ATA-TAP meeting 2/8/02 || Rogers Dielectrics
83 Input match for spiral antenna
88 Pictures of prototype feed.
90 Feed patterns
92 Noise summary @ 10 GHz
93 feed patterns
96 Noise diodes
97 equalizers for NITRO amps.
104 At what frequency will self-cal work?
105 layout of calibration module
106 design input region of feed pyramid.
111 Input line impedances and Losses
111 Match of Glass Seal Reflections
115 RFI level from the BIMA correlator
117 Thermal Loads in the Dewar
124 Review meeting
126 More on the calibration System
129 Jan. 3 2003 MOU meeting
129 Shielded Room Specs from Rick Fisher and Tom Lacuelabe
130 meeting with Semely + Niklas 1/29/03 good active balun
134 Coupling between seal spiral and small spiral antenna
135 Thermal Loads on Heat Shield
137 Tests of 100-50 Ω matching box
138 Required Gain stability for OTF (first try)
140 Thermal calculations for dewar interior
142 ATA sensitivity relative to Arecibo
143 Properties of crystalline quartz: thermal and electrical
144 Tests of Double baluns with various covers
151 Expected Operation at 327 MHz
152 Thermal stability for OTF mapping

Matt's PDR [December 1, 2000]

Leo, Tom intro.

John Dreher general intro [Software and DSP effort behind]

Dave de Boer intro to the antennas - discusses the optimization of the Gregorian layout, spillover, defocusing,

A lower edge illumination lowers A_e but also lower T_{sys} and less interference sensitivity - might be a good direction
note standing-wave problem at low frequency for out-of-focus case.
need to optimize the shroud.
Matt: wind sock position for stow
show discussion

Matt #1 AzEl pop. for larger sizes — Some ideas here
Linear actuators popular
Andersen quality good.

M shows both ants - downwind is the preferred stow.
any ideas about the shroud cover
spec: 3 1/4 arc minute pointing, 1/5 beam at 12 GHz (!)

Cass: rolled ring on back, rings, spars to edge at 4 pts, ring at edge.

Offset Gregorian: many tension/compression spars go to the 2 ends of an elevation beam - use shear structure?
Eric thinks fab on site is the best - Sergio disagrees
Matt not sure - Andersen rather keep tool at home.

Total costs are similar between budgets of ~$8200
7.0 m Gregorian ~21K

Winds. highest 98 percentile, normal 0-15 mph, degraded > 45 mph (Dave de Boer)
survival: 100 mph
73 is the highest wind of the last 6 years.

Extra edge added to 5m - [Probably not needed - JW]

Counter weight? What wind loads vs. gravity?

Wind Forces, then gravity

back drives - snow loads - safety issues

Azimuth Bearing - casting, cost issues

slew speed spec ~ 3°/second -

Ball Screws Sergio: lubrication problems

Reed Switches w magnets for absolute encoding - reliability
otherwise it's an incremental encoder

optical fiber in the azimuth cable wrap -

interesting elevation encoder mounting

many deflection tests

need a better casting design for stiffness

Andersen 4.2 m had a measured RMS of .023"
JPL did an analysis of deflections ~~with~~ on this dish -

Committee Comments

Need more margin of safety 45-60 mph small incr. cost

good to be able to move the antennas but too costly -

Locate the prototype in a windy place if possible

Manufacture: be able to divide the reflector in 3 parts for shipping - keep the center section intact.

Concern about jack screw drive on the elevation - may be too slow - ball screw may be dangerous - counter weight maybe a good idea for safety -

what slew speed is ok? too avoid counter weights

which configuration is it? — A choice must be made —
Matt needs guidance here.

Recommend absolute encoders, if it's not too expensive

Sergio recalls Heidenhein should have 19 bit for a few hundred dollars — ~~Eric~~ even a synchro may be ok. ~12 bits

corrosion may be an issue on the antennas —

SB: servo needs to be looked into — Allen is optimistic — Chicago made a nice design for Daisy —
get going on the software (for the general project)

Put a boot on the jack, rubber boot,

A design is needed for the Gregorian —
what do we do with the snow?
alignment is tricky —
SB worries out the GBT — he thinks a factor of two in cost —
bigger lever arm

The committee needs to see Matt's design.
Tony thinks the design is worth it.
Concentrate Matt's effort on this.
Counter wgts are probably needed —

[Seems to be some fear of building another GBT]

Tony wonders why not a bigger mount

Make a tighter budget — 20k seems a reasonable total —

AR: Make some iterations on the design. They moved too quickly with their design.

Matt has a good test set-up — elevation drive is needed —

A Revision of the Offset Gregorian Gain Table

Based on Keeping the Feed in Focus

ν	λ	η_{dep}	η_i	η_{spill}	η_{IN}	F_{eff}	A_o	l_d	$e^{-(\frac{\pi l_d}{\lambda})^2}$	Cuel spill 2/3 l_d	~~Cuel spill~~	η_T	
11.25	2.67	1.00	.986	~~.756~~	.832	.65	.211	.0072	.915	.4%		.61	.61
10.00	3.00	'	.985	~~.817~~	'	'	'	.0077	.932	.5%		.62	.62
8.75	3.43	'	.984	~~.756~~	'	'	'	.0082	.947	.5%		.63	.63
7.50	4.00	'	.982	'	'	'	'	.0088	.961	.5%		.64	.64
6.25	4.80	1.00	.991	~~.756~~	.832	.65	.211	.0097	.973	.6%		.65	.65
5.00	6.00	'	.978	~~.817~~	'	'	'	.0108	.983	.7%		.65	.65
3.75	8.00	'	.975	~~.756~~	'	'	'	.0125	.990	.8%		.65	.65
2.50	12.00	'	.969	.817	'	'	'	.0153	.996	.9%		.66	.65
1.25	24.00	'	.957	'	'	'	'	.0217	1.00	1.3%		.65	.65
1.00	30.0	'	.952	~~.756~~	'	'	'	.0242	'	1.5%		.65	.65
.75	40.0	'	.944	~~.817~~	'	'	'	.0280	'	1.7%		.64	.64
.50	60.0	1.00	.932	~~.756~~	.832	.65	.211	.0342	1.00	2.1%		.63	.63
										+1% w/thread		OK,	

$l_d = (.20485)(.211)\sqrt{\lambda(m)} = .0442\sqrt{\lambda(m)}$ ✓

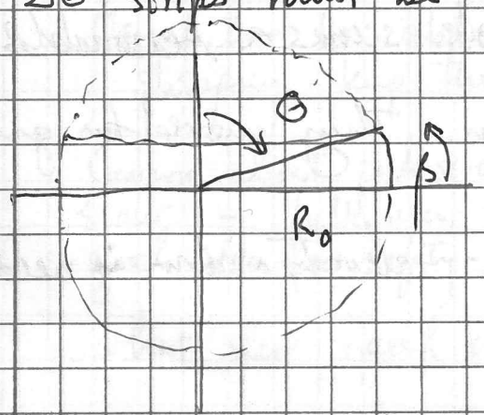
$D \cong 5m \qquad d = 2.4m$

$\eta_i \cong 1 - 2l_d$

(carefully & spillover ✓)

Some accurate spillover calculations

The feed data (I p.133) gives axially symmetrical sidelobes integrated to various polar angles Θ_E. That gives feed power in strips of constant $\Delta\Theta$. When the antenna is tipped the spillover lies between planes: one normal to the antenna axis and the other the horizon. The angular area between these planes is a lune, and the power in the constant $\Delta\Theta$ strips must be corrected to the lune area.



Sphere of radius R_o
Lune of angle β
polar strip of width β

Surface Area of the Lune $S = 2R_0^2\beta$ (a quote)

Surface Area of the polar strip of width β (half of the strip): $R_0^2\int d\Omega = R_0^2\int_0^{\pi} d\phi \int_{\pi/2-\beta}^{\pi/2} \sin\theta \, d\theta = A$

$$= R_0^2\pi\,(-\cos\theta)\Big|_{\pi/2-\beta}^{\pi/2} = \pi R_0^2 \cos(\pi/2-\beta) = \pi R_0^2 \sin\beta$$

The ratio of angular areas: $R = S/A = \frac{2\beta}{\pi \sin\beta}$

$R \to 1$ as $\beta \to \pi/2$ and $R \to \frac{2}{\pi}$ as $\beta \to 0$. &c.

Spillover for the Cassegrain

Feed Pattern from I p. 133/134

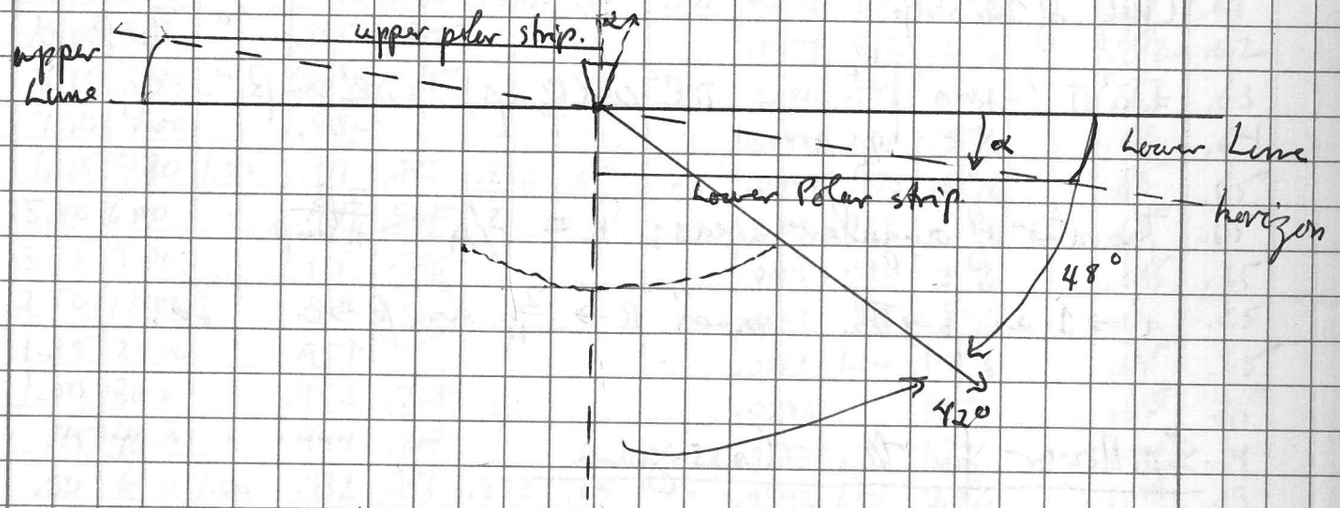
β	θ_e	% in half spherical strip	R	total	×300k		β upper
0	42	0	.637	0	0		0
8	50	.019	.638	.012	3.6		10
18	60	.024	.647	.016	4.8		20
28	70	.031	.663	.021	6.3		30
38	80	.039	.686	.027	8.1	9.1	40
42	84	.049	.697	.034	10.2		
50		.049	.725	.036	10.8		50
60		⋮	.770	.038	11.4		60
70		⋮	.828	.041	12.3		70
80		⋮	.903	.044	13.2		80
90		.049	1.00	.049	14.7		90

Spillover for the Gregorian

β	
0	3
10	3
20	3
30	3
40	2.8
50	2.6
60	2.0
70	1.4
80	0.6
90	0

Spillover for the Prime Focus with the better feed pattern of I p 133/134

Spillover at the zenith is .096



Tipping the antenna axis α lowers the horizon on one side by α and raises by α on the other side. The two sides should be considered separately.

The ratio between the surface half polar strip and the Lune is (again) $R = \frac{2\alpha}{\pi \sin\alpha}$

$\alpha = 10°$ right hand side (lower) lose the lune: $\alpha = 10$, $R = .640$
The energy between $\theta_e = 80$ and $\theta_e = 90$ is gone: $\frac{.916 - .886}{2} R = .0064$

left hand side (upper) gain the Lune $\alpha = 10$, $R = .640$
Gain the energy between 90° and 100° θ_e: $\frac{.932 - .941}{2} R = -.0029$
Net gain: $\Delta_{10} = .0029 - .0064 = -.0035$

$\alpha = 20°$ Right hand side: $\alpha = 20$, $R = .650$ $\theta_e = 90$ to 70: $\frac{.915 - .861}{2} R = .0111$
Left hand side θ_e: 90 to 110 $\frac{.941 - .915}{2} R = .0085$
Net gain $\Delta_{20} = .0111 - .0085 = -.0027$

$\alpha = 30°$ Right side $\alpha = 30$, $R = 0.670$; θ_e: 90 to 60: $\frac{.915 - .867}{2} R = .0161$
Left side; θ_e: 90 to 120: $\frac{.952 - .915}{2} R = .0124$
Net gain $\Delta_{30} = .0124 - .0161 = -.0037$

$\alpha = 40$ Right side $\alpha = 40$ $R = .691$; θ_e: 90 to 50: $\frac{.915 - .857}{2} R = .020$
Left side θ_e: 130 to 90: $\frac{.957 - .915}{2} R = .018$
Net gain $\Delta_{40} = .018 - .020 = -.002$

$\alpha = 48$ (one edge at horizon here) $R = .7177$

right lost $\quad 90-42: \quad \frac{.915-.819}{2} R = .0344$

left gain $\quad 138-90: \quad \frac{.975-.915}{2} R = .0215$

net gain $\quad \Delta_{48} = .0215 - .0344 = -.0129$

$\alpha = 60$ no change lower side $\quad R = .770$

$180-90: \quad \frac{.981-.915}{2}(.770) = .0025$ net gain

$\alpha = 70 \quad R = .828 \quad 160-90: \frac{.985-.915}{2}(.828) = .029$ net gain

$\alpha = 80 \quad R = .903 \quad 120-90: \frac{.995-.915}{2}(.903) = .0361$

$\alpha = 90° \quad R = 1.00 \quad 180-90: \frac{1.000-.915}{2}(1.00) = .0425$

Add these to the .096

Zenith	fractional spillover	x 300
0	.0960	28.8°
10	.0925	27.8
20	.0933	28.0
30	.0923	27.7
40	.0940	28.2
48	.0830	25.0
60	.0985	29.6
70	.1250	37.5
80	.1321	39.6
90	.1385	41.6

Skirt possibilities $\quad f = (.65)\,5\,m = 3.25\,m$

a) continue skirt to 50° (+8°): $8° \times f = 0.45\,m$ radius

$0.058 \times 300 = 17K$ left

b) continue skirt to 60°: $18° \times f = 1.02\,m$ radius

$.048 \times 300 = 14.4K$ left

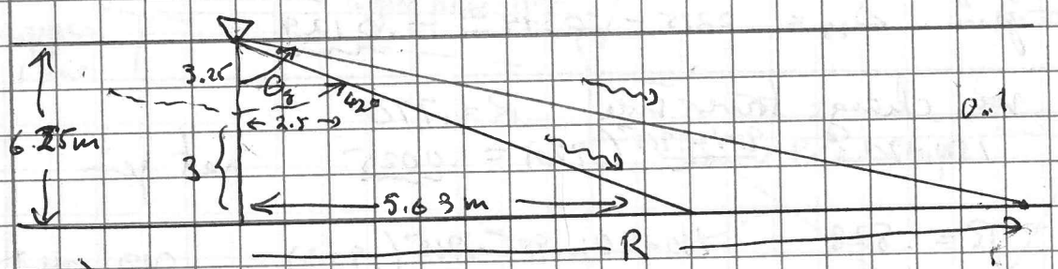
c) continue to 70° $\quad 28° \times f = 1.6\,m$

$.034 \times 300 = 10.2K$ left

d) 80° $\quad .019 \times 300 = 6K$ $\quad 39° \times f = 2.61\,m$

Possible Ground Screen (D = 5m)

With $F = .65$, $f = 3.25$ for $D = 5m$



Solid angle out to θ_g : $\Omega = \int_0^{2\pi} d\phi \int_0^{\theta_g} \cancel{\cos\theta}\, d\theta = 2\pi \sin\theta_g$

Inside fraction $= \frac{\Omega}{2\pi} = \sin\theta_g = \frac{D}{\sqrt{D^2+(6.25)^2}} = \frac{1}{\sqrt{1+\left(\frac{6.25}{D}\right)^2}} \cong 1 - \frac{1}{2}\left(\frac{6.25}{D}\right)^2$

For large D.
For the inside fraction to be 0.9 so that 10% spillover becomes 1%. (This leaves out the shadowed region out to 5.63 m.): $.2 = (6.25/D)^2$ and $D = 14m$

The fraction of solid angle inside 42° is $\sin 42 = .6691$
It's the fraction between this and 1 that should define R, that should be .9.

$\frac{\sin\theta_g - \sin 42^\circ}{1 - \sin 42^\circ} = 0.9$ Get $\theta_g = 75^\circ$
and $R = 6.25 \tan 75^\circ = \underline{23.7\ m}$

This is substantially more. For the tipping, the shield has to cover beneath.
For each antenna, this would be 1761 m^2
For 500 that's 880,000 m^2
But the same screen will serve adjacent antennas.
This argues that the screen be everywhere under the antennas and 24 m beyond each border.
Roughly a disk of 750 m diameter : 440,000 m^2
Thats 880 m^2 for each antenna.
If the cost saving of the prime focus relative to the Cregorian is \$5000 the break-even cost is $\underline{\$5.6/m^2}$

i) Avecibo gnd screen is \$60/m^2
ii) JD's builder's estimates are (for siding) $\geq$ 20 \$/m^2
iii) ACE hardware Al is $\geq$ \$23/m^2
iv) Matt's bare steel $\geq$ \$4.2/m^2 and woven wire is \$1.29/m^2

Solid angle out to θ_g: $\Omega = 2\pi \int_0^{\theta_g} \sin\theta \, d\theta = 2\pi(\cos\theta_g + 1)$

Inside fraction: $\frac{\Omega(z)}{2\pi} = 1 - \cos\theta_g$

Outside fraction: $1 - (1 - \cos\theta_g) = \cos\theta_g = \sin E$

Forgetting the effect of the inner dish shadow, make $\cos\theta_g = 0.1$ for the outside fraction, then $\theta_g = 84.2°$

$\tan\theta_g = R / 6.25\,m = 9.9$

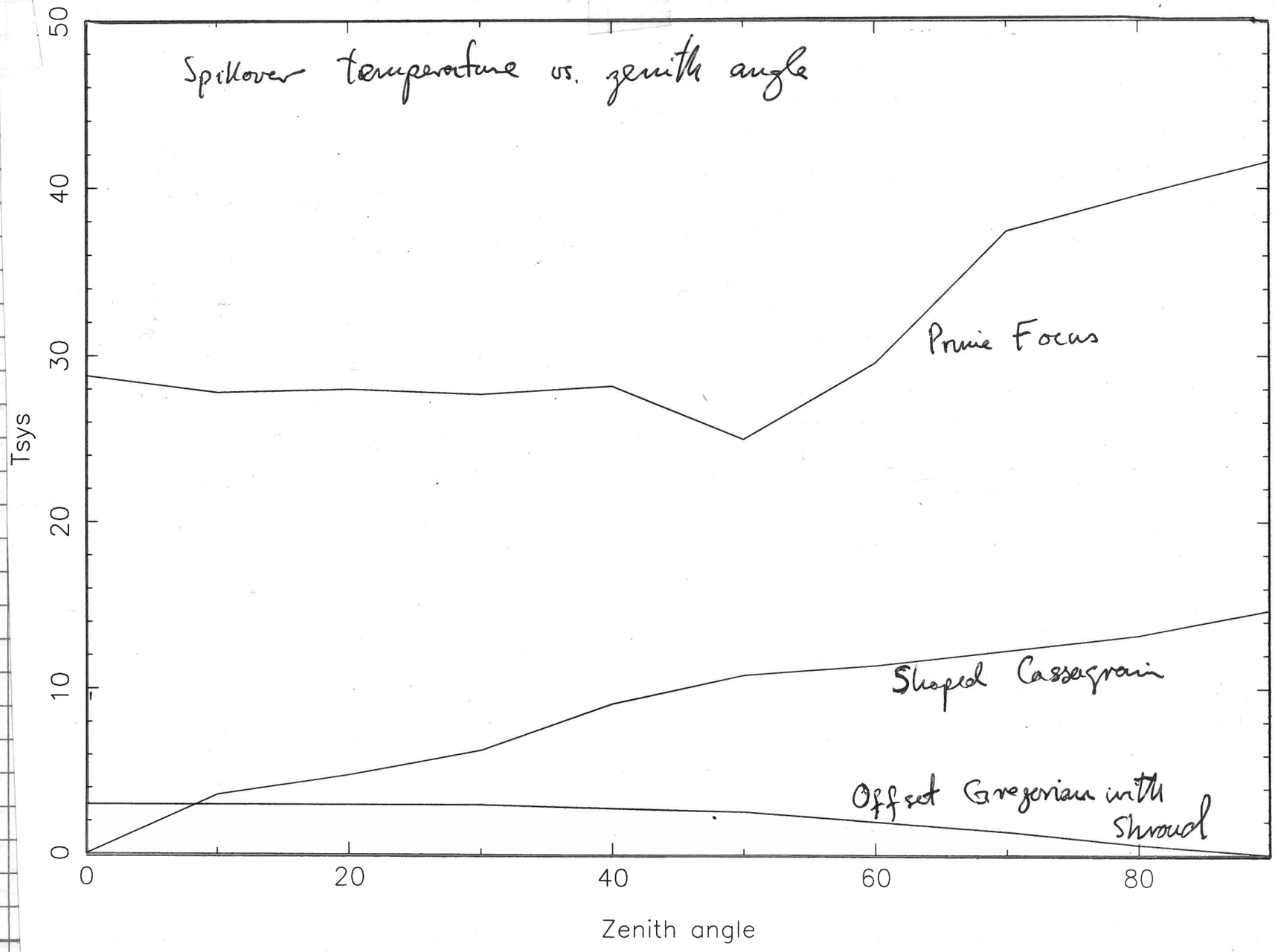
so $R = 62\,m$ The shadow of the dish is negligible.

The area of the screen for one antenna is $12{,}000\ m^2$ the screen must be continuous everywhere and extend to $\sim 62\,m$ beyond the antennas on all sides

A screen of $\sim$ ~~702~~ 820 m diameter is $\sim$ ~~450,000~~ $528{,}000\ m^2$

That's $\sim 1800\ m^2$ for each antenna.

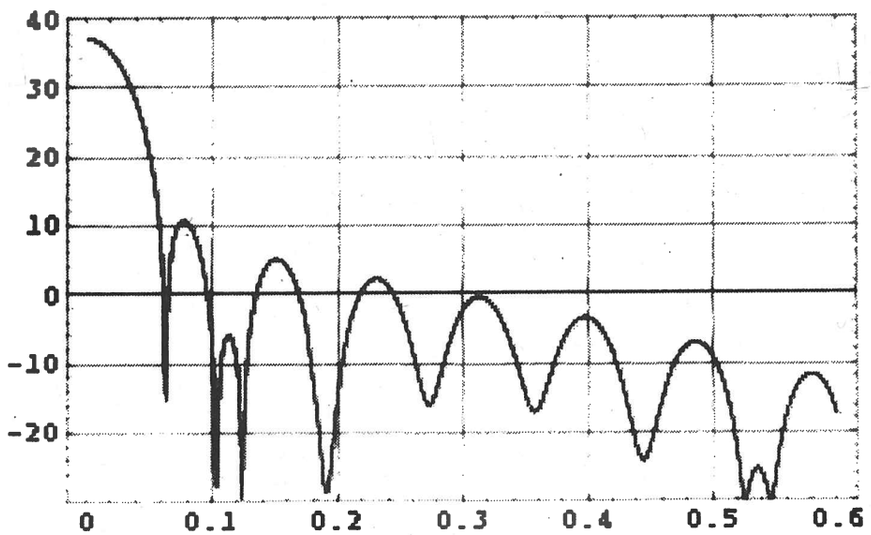
For the prime focus saving of \$5000. That's \$5.0/m² as the break even cost, including labor and maintenance, of course.


Spillover temperature vs. zenith angle
Prime Focus
Shaped Cassegrain
Offset Gregorian with Shroud
Tsys
0
10
20
30
40
50
0
20
40
60
80
Zenith angle

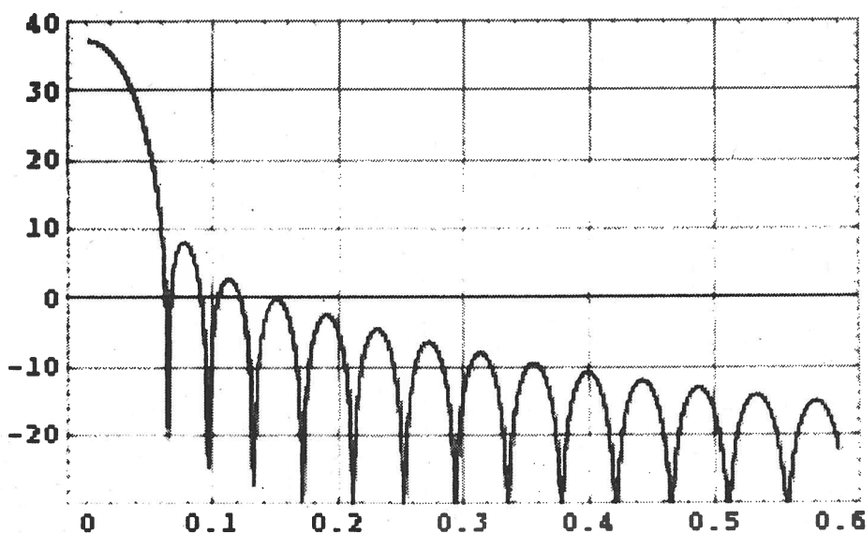
Pattern Comparisons [paraxial diffraction]

I Prime Focus with diffraction from the feed.
- compared with clear aperture (offset Gregorian)

$\lambda = 20$ cm - high frequency behavior



prime focus
$D = 5$ m



clear aperture
(offset Gregorian)

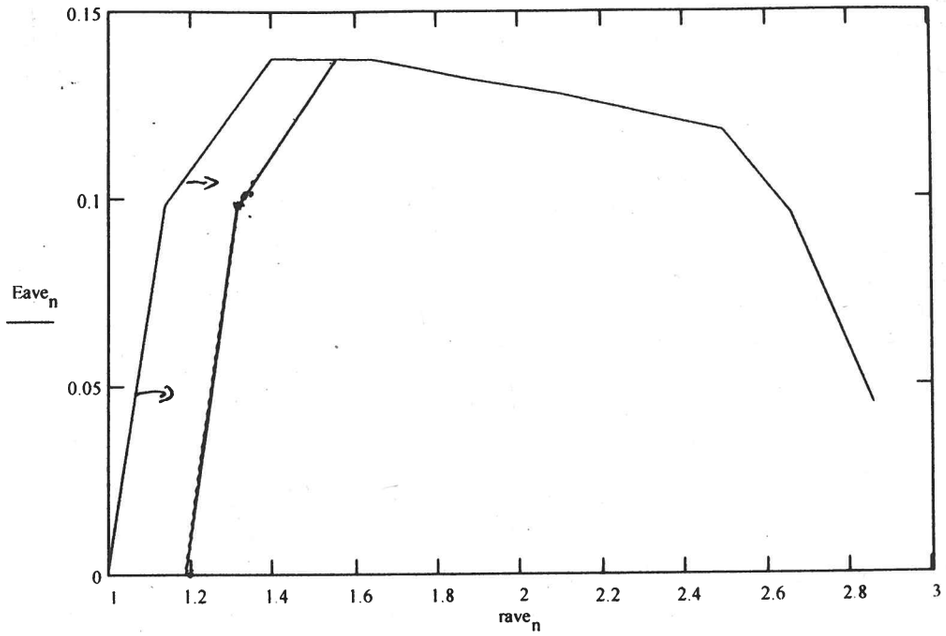
$D_p = 5$ m
$D_s = 2.4$ m

II Shaped Cassegrain and clear aperture Gregorian

$\lambda = 2cm$

high freq. approximation

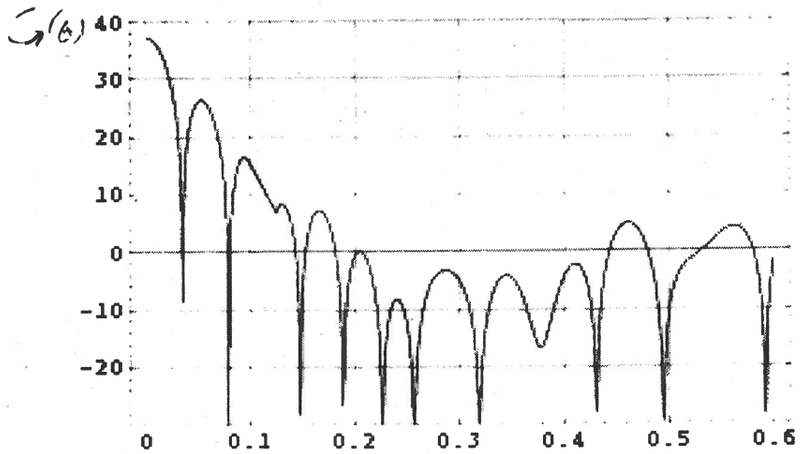
Cassegrain Shaped Feed Pattern adjusted to 2.4m diameter



Cassegrain

$D_p = 6m$

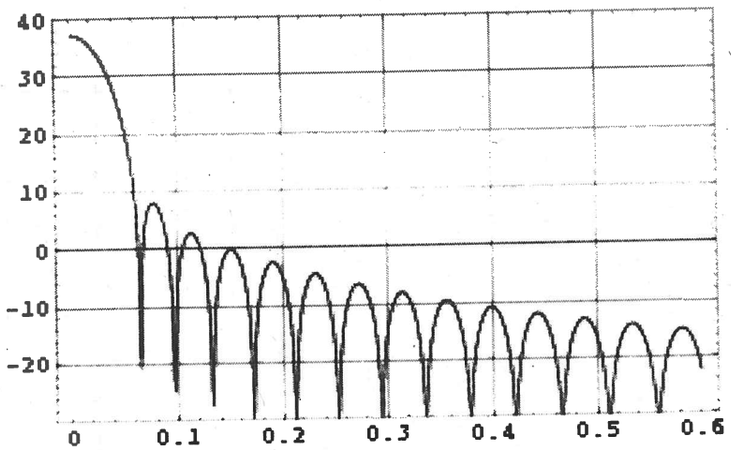
$D_s = 2.4$

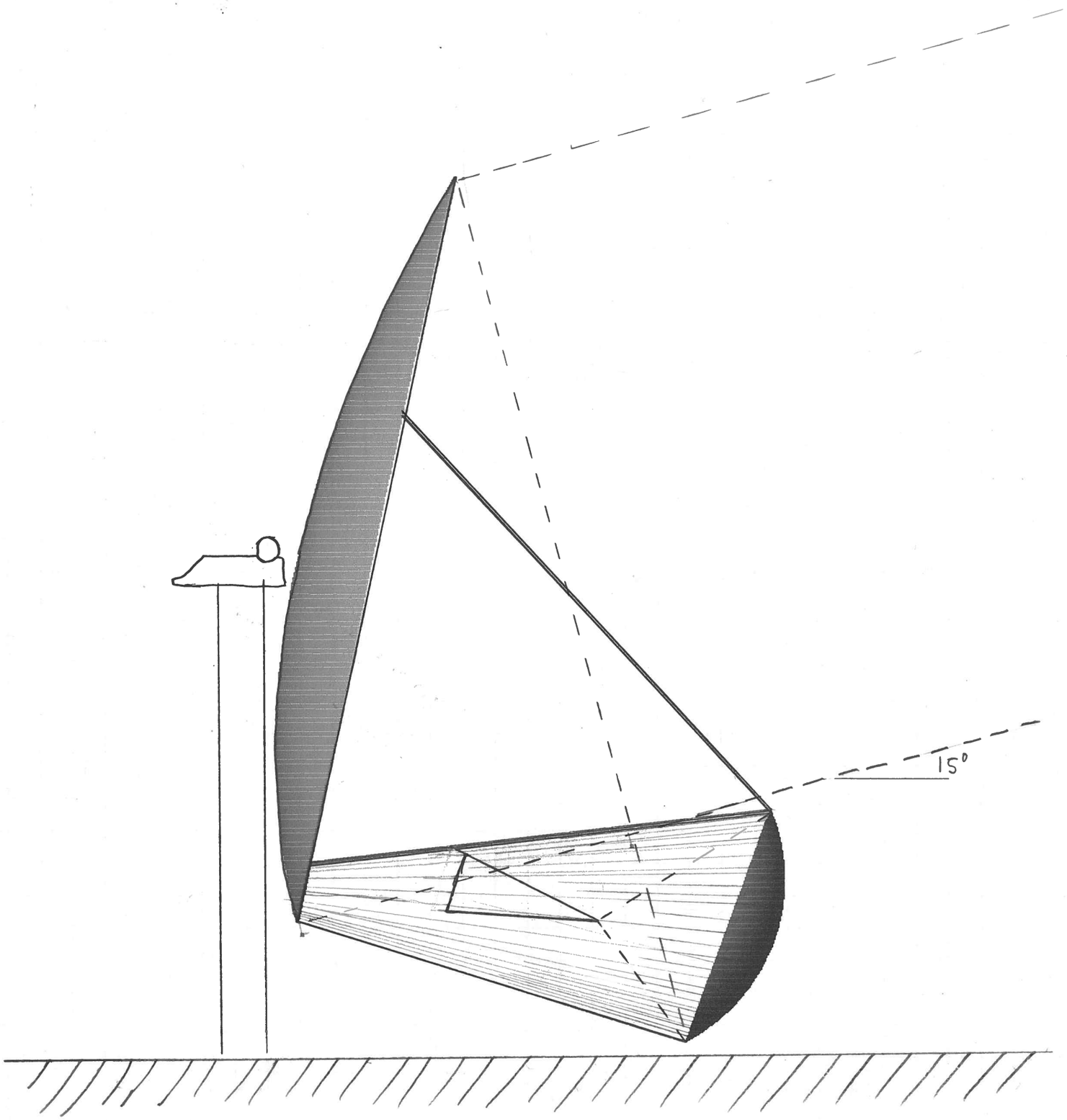


Gregorian

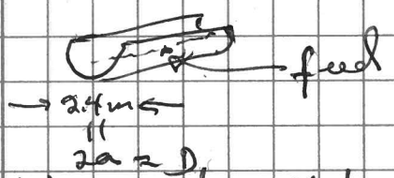
$D_p = 5m$

$D_s = 2.4m$




15°

Shroud/Feed Interactions



Waveguide Modes: Lowest TE_{11} mode cutoff: $k_c a = p_{11} = 3.8$

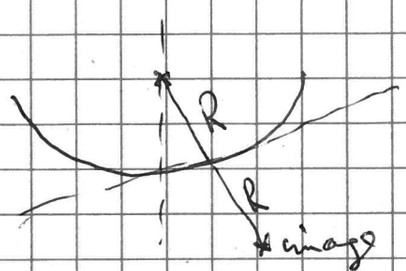
$$\frac{2\pi}{\lambda_c} a = 3.8 \qquad \lambda_c = \frac{\pi(2a)}{3.8} = \frac{\pi}{3.8} D = \frac{\pi}{3.8} 2.4 = \underline{2.0\ m}$$

The lowest frequency, corresponding to 2m λ, is 150 MHz

Back Reflection of the feed signal

The worst case appears to be a side echo from the cylindrical wall back into the feed. Radiation into other directions is reflected so that it doesn't return into the feed.

A rough calculation: Think of each piece of the shroud wall as a section of an infinite plane. Find the echo from each piece and sum it over all the directions toward the cylinder.



The image is at 2R.

Radiated power is $P(\theta,\phi) = P_i \frac{G(\theta,\phi)}{4\pi}$

The image must have the same strength $P_i \frac{g}{4\pi}$.

The receiving cross section is $A = \frac{g\lambda^2}{4\pi}$

The received echo is $P_i \frac{g}{4\pi} \Delta\Omega$, where $\Delta\Omega$ is the solid angle of the receiver $(2R)^2 \Delta\Omega = A$; so $\Delta\Omega = A/4R^2$

$$P_r = P_i \frac{g}{4\pi} \Delta\Omega = P_i \frac{g}{4\pi} \cdot \frac{A}{4R^2} = P_i \frac{g}{4\pi} \cdot \frac{g\lambda^2}{4\pi} \cdot \frac{1}{4R^2}$$

$$= P_i \left(\frac{g\lambda}{4\pi R}\right)^2 \cdot \frac{1}{4} \qquad \text{so} \qquad P_R/P_i = \frac{1}{4}\left(\frac{g\lambda}{4\pi R}\right)^2$$

p. 143 gives the feed gain as $14.1\,|E_f(\theta)|^2 = G(\theta)$

where $E_f(\theta)$ is the pattern of p. 75. At 90° $E_f = .05$,

so $G(90°) = 14.1\,|.05|^2 = .035$ (−14.5 db)

$$P_r/P_i = \frac{1}{4}\left[\frac{(.035)(.3)}{4\pi\,(1.2)}\right]^2 = 1.2\times10^{-7} \qquad \propto \frac{1}{R^2}$$

($\lambda = .5$ m)

Summing this around the half cylinder coherently should give (for the power) an additional factor of $(\pi)^{-1} \cong 10$

So, the coherent echo is 1.2×10^{-6} (−59 db)

The gain should be independent of frequency, so the coupling should be proportional to λ^2

At $\lambda = 0.6$ m (500 MHz), it is −53 db.

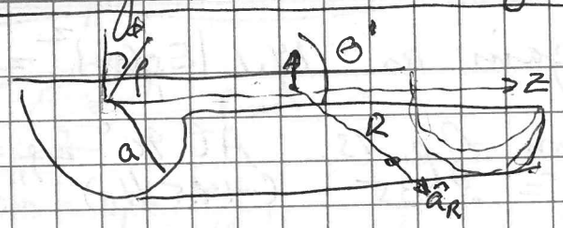
This is small, but it <u>may be too crude</u>. (a little)

Is the shroud in the far zone of the feed? Yes

As seen from the side, the extent of the radiating region $\cong \lambda/2$. Then $\frac{2D^2}{\lambda} = \frac{2\,\frac{\lambda^2}{4}}{\lambda} = \frac{\lambda}{2}$

For $\lambda = 0.6$ m $\lambda/2 = 0.3$ m, smaller than 1.2 m.

More careful calculation of the signal reflected by the shroud



Describe the radiation by the feed $G(\theta',\phi')$ with its polarization corresponding a current element as shown in the primed system.

$P(\theta',\phi') = P_i\, G(\theta',\phi') \rightarrow G(\theta',\phi')$ with $P_i = 1$.

$\vec{S}\cdot\hat{a}_R$ (Poynting vector) $= \dfrac{\eta\,|H|^2}{2}$ at the cylinder

$\vec{S}\cdot\hat{a}_R\, dA = P(\theta',\phi')\, d\Omega \qquad dA = R^2 d\Omega$ so.

$$\vec{S}\cdot\hat{a}_R\, dA = G(\theta',\phi')\frac{dA}{R^2} = \frac{\eta\,|H|^2}{2}\, dA.$$

The field is $H_{\phi'}$ in the source coordinate systems.

Calculate the reflected field from physical optics: $\vec{J} = n\times\vec{H}_i$ ($n = \hat{a}_R$)

The implied cross product is a little complicated, but the final integral for the reradiation will have a stationary phase point, and only the cylindrical plane $z=0$ will be important.

$$H_{\phi'} = \sqrt{\frac{2\,G(\theta',\phi')}{\eta R^2}}\,\sin\theta'\, e^{-jkR}$$

Only J_ϕ will radiate back $\qquad J_\phi = 2H_{\phi'}\, f(\theta',\phi',\phi)$

f is the cross product.

$$J_\phi = 2\sin\theta'\sqrt{\frac{2G}{\eta R^2}}\, e^{-jkR}\, f$$

The back radiated E field (with R large) is

$$dE_{\theta'} = J_\phi\,\frac{\sin\theta'\, j\omega\mu\, e^{-jkR}}{4\pi R}\, dA \quad ; \quad dA = dz\, a\, d\phi$$

$$dE_\theta = \left\{ 2\sin\theta' f \sqrt{\frac{2G}{\eta R^2}}\, e^{-jkR} \right\} \sin\theta' \frac{j\omega\mu}{4\pi R} e^{-jkR}\, dz\, a\, d\phi$$

$$E_\theta = \int_{\pi/2}^{3\pi/2} \int_{-\infty}^{\infty} \frac{2\sin^2\theta' f}{4\pi R^2} \sqrt{\frac{2G}{\eta}}\, j\omega\mu\, e^{-2jkR}\, a\, d\phi\, dz$$

The power received back is $P_{rr} = \frac{|E_\theta|^2}{2\eta}$ per unit area. The receiving area is $A_e(\theta',\phi') = \frac{G(\theta',\phi')\lambda^2}{4\pi}$

$$P_r = \frac{|E_\theta|^2}{2\eta} A_e = \frac{|E_\theta|^2}{2\eta} \frac{G(\theta',\phi')\lambda^2}{4\pi}$$

There is a stationary phase point in the z integration

$$R^2 = a^2 + z^2 \qquad R = a\sqrt{1+\frac{z^2}{a^2}}$$

$$2jka\sqrt{1+\frac{z^2}{a^2}}$$ Do this one first

<u>The Stationary Phase asymptotic expansion</u>

If $I = \int_a^b g(t)\, e^{jx\,h(t)}\, dt$, and $h'(\tau) = 0$

Then $I \cong \sqrt{\frac{2\pi}{x\,|h''(\tau)|}}\; g(\tau)\, e^{jx\,h(\tau)}\, e^{js\pi/4}$

where s is the sign of $h''(\tau)$.

Write $h(z) = \left(1+\frac{z^2}{a^2}\right)^{1/2}$; $h'(z) = \frac{z}{a^2\sqrt{1+z^2/a^2}}$ $\qquad h'(0) = 0$

$$h''(z) = \frac{1}{a^2}\left\{ \frac{\sqrt{1+\frac{z^2}{a^2}} - z\cdot\frac{2z}{a^2\sqrt{1+z^2/a^2}}}{\left(1+\frac{z^2}{a^2}\right)} \right\} \qquad h''(0) = \frac{1}{a^2}$$

$x = 2ka$ $\qquad h(0) = 1$

In the plane $z=0$, $f=1$ and $\theta' = \phi$ and $G \to G(90°)$

Assume that $G(90°)$ is independent of ϕ, that the $\sin\phi$ dependence takes care of it.

$$\int_{3\pi/2}^{\pi/2} \sin^2\phi \, d\phi = \left(\frac{\phi}{2} - \frac{\sin 2\phi}{4}\right)_{\pi/2}^{3\pi/2} = \frac{1}{2}\left(\frac{3\pi}{2} - \frac{\pi}{2}\right) = \frac{\pi}{2}$$

$$E_\theta = \frac{\cancel{2}a\, j\omega\mu}{4\pi R^2} \sqrt{\frac{2G}{\eta}}\, e^{-j2ka} \left(\frac{\pi}{\cancel{2}}\right) \sqrt{\frac{2\pi}{2ka \cdot \frac{2}{a^2}}}\, e^{i\pi/4}$$

$$= \frac{\cancel{a}\, j\omega\mu}{4a^{\cancel{2}}} \sqrt{\frac{2G}{\eta}} \sqrt{\frac{2\pi}{\cancel{2}k/a}}\, e^{-j2ka} e^{i\pi/4} = \frac{j\omega\mu}{4a} \sqrt{\frac{2G \cdot \cancel{2}\pi a}{\cancel{2}\eta k}}\, e^{-j2ka} e^{i\pi/4}$$

$$= \frac{j\omega\mu}{4a} \sqrt{\frac{G\pi a 2}{\eta k}}\, e^{-j2ka} e^{+i\pi/4}$$

The power received back is $P_r = \frac{|E_\theta|^2}{2\eta} A_e = \frac{|E_\theta|^2}{2\eta} \frac{G\lambda^2}{4\pi}$

$$= \frac{G\lambda^2}{4\pi} \cdot \frac{1}{2\eta} \left(\frac{\omega\mu}{4a}\right)^2 \frac{2G\pi a}{\eta k} = \frac{G^2\lambda^2}{8 \cdot 4^2} \frac{\omega^2\mu^2}{\eta^2 k} \cdot \frac{2}{a} = \frac{G^2\lambda^2}{4 \cdot 4^2} \frac{\omega^2\mu^2}{\mu/\epsilon \cdot \omega\sqrt{\mu\epsilon}\, a}$$

$$= \frac{G^2\lambda^2}{4 \cdot 4^2 a} \frac{\omega\mu\epsilon}{\sqrt{\mu\epsilon}} = \frac{G^2\lambda^2}{a\, 4 \cdot 4^2} \omega\sqrt{\mu\epsilon} = \frac{G^2\lambda^2}{a\, \underset{4_2}{4 \cdot 4^2}} \frac{2\pi}{\lambda} = \frac{G^2}{\underset{32}{8}a} \pi\lambda$$

$$\boxed{\frac{P_r}{P_t} = \frac{\pi G^2}{32} \cdot \frac{\lambda}{a}} \sim \frac{G^2}{10} \frac{\lambda}{a}$$

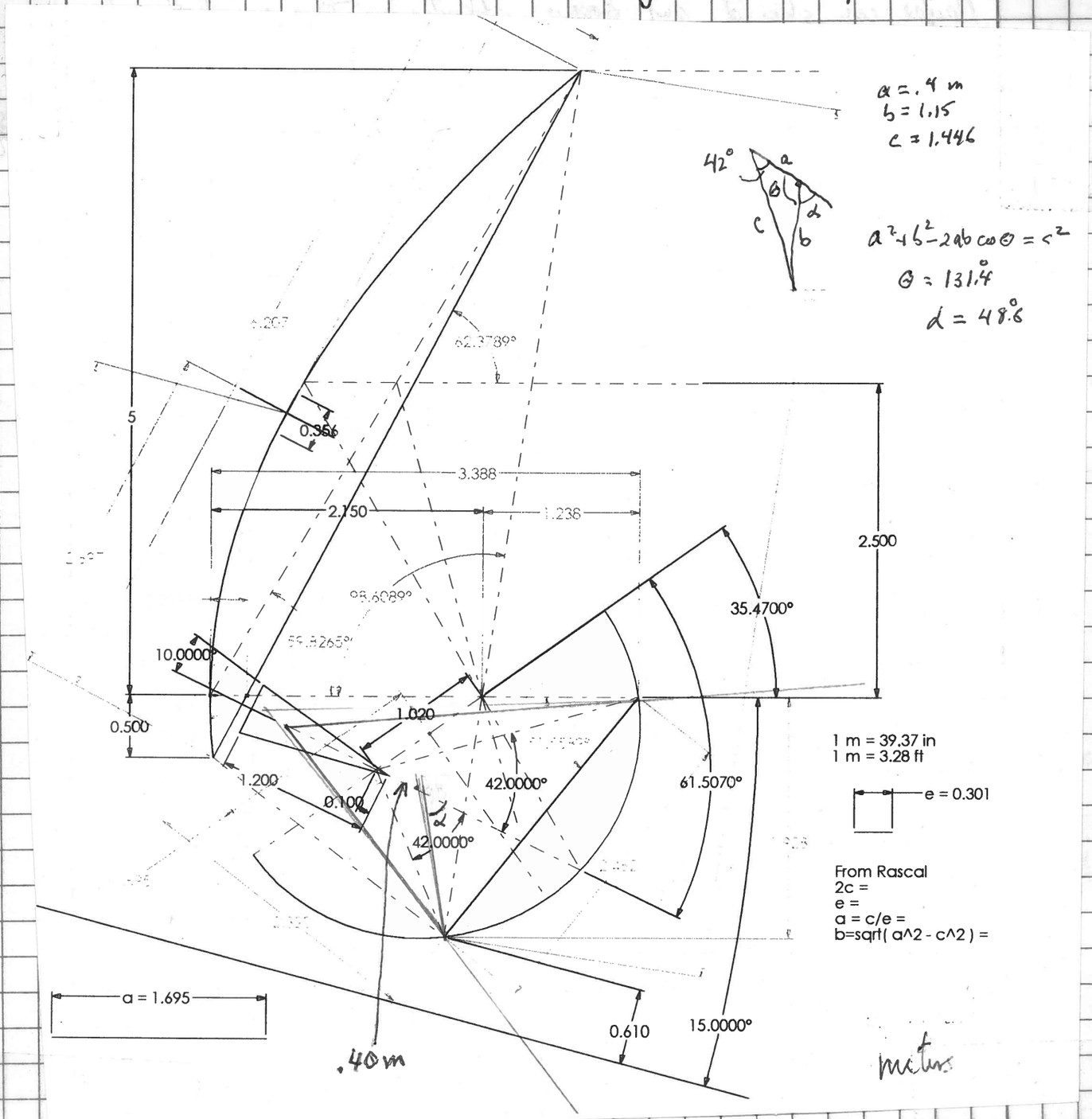
This should be more accurate than the previous estimate.

$G(90°) = .035$ (−14.5 db)

For the longest wavelength $\lambda = .60$ m (500 MHz)
$a = 1.2$ m

$$\frac{P_r}{P_t} = \frac{\pi (.035)^2\, 0.6}{[32][1.2]} = 6 \times 10^{-5}, \quad -42 \text{ db}$$

<u>Another Reflection</u> If the feed is moved forward in order to bring the phase center of a longer wavelength into focus, the tip (active at higher frequencies) may see a direct echo from the edge of the secondary. In the figure below this happens when the tip is moved forward by 0.4m. This amount of motion corresponds to .40m/1.41 = .28m wavelength in focus. In the diagram, $\alpha = 48.6°$. If the cylindrical formula applies $G^2/10\,(\lambda/a)$ for the shortest wavelength (appropriate to the tip): $G(48.6°) = .44$, so (-3.5db $(.44)^2/10\,(\frac{.03}{1.2}) \Rightarrow -33.2$ db! Probably ok. But keep displacement ≤ 0.4m



Reflection of the direct back lobe

The gain in the back direction is ~ −3.5 db (?). The biggest problem could be at the lowest frequency with the radiating region nearest the end, which is closest to the primary reflector.

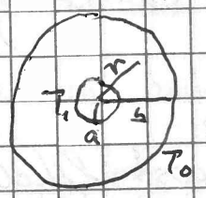
$G^2/10 \quad d/\lambda$

For $d \sim 0.5$ m at $\lambda = .6$ m,
$G^2/10 \sim -17$ db, a little high.

Maybe we should put some absorber there, at the back.

N.B. It's probably lower than this on the new feed, which radiates only 8% in the back direction instead of 16–17%

Insulation properties of a spherical ball of beadboard.



$$\nabla^2 T = 0 \quad \text{so} \quad T = A + B/r$$

$$T_0 = A + B/b$$
$$T_1 = A + B/a$$
$$(T_0 - T_1) = B\left(\frac{1}{b} - \frac{1}{a}\right) = \frac{B(a-b)}{ab} \quad ; \quad B = (T_0 - T_1)\frac{ab}{a-b}$$

$$A = T_0 - B/b = T_0 - (T_0 - T_1)\frac{a}{a-b}$$

Altogether $T(r) = \dfrac{T_1 a - T_0 b}{(a-b)} + \dfrac{(T_0 - T_1)ab}{r(a-b)}$; $= T_0$ at b and T_1 at a.

Heat Flow Through the outer surface $\quad H = 4\pi b^2 K \dfrac{\partial T}{\partial r}\Big|_{r=b}$

$$H = 4\pi b^2 K \left.\frac{(T_0 - T_1)ab}{-r^2(a-b)}\right|_{r=b} = -4\pi K (T_0 - T_1)\frac{ab}{(a-b)}$$

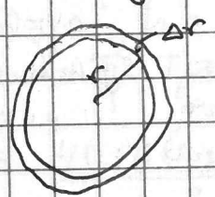
For $b \gg a$ $\quad H = +4\pi K a (T_0 - T_1)$ $\quad$ independent of b.

Good insulators like rock-wool, have $K \sim .034$ watts/m°K
Metal alloys have $K \sim 14$-120. If stainless has $K = 14$,
$14/.034 = 411$ That's how much smaller the crossection of stainless would need to be.

For $b = 15\text{cm} = .15\text{m}$, $2b = .30\text{m}$ and $a = .25\text{cm} = .0025\text{m}$, $b \gg a$,
and $K = .034$, with $T_0 = 300\text{K}$ and $T_1 = 80\text{K}$:

$$\text{Heat flow} = +4\pi(.034)(.0025)(220) = 0.235 \text{ watts}$$

Compare the crossection of a thin cylindrical shell to the cross sectional area of the cylinder.



$$\frac{\pi r^2}{2\pi r \Delta r} = \frac{r}{2\Delta r}$$

If the ratio of K's is 411,

then $r/2\Delta r = 411$ for equal heat flow.

Jan. 15 ATA Board @ SETI Inst.

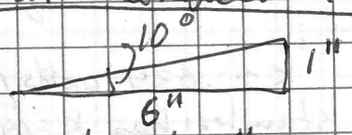
1. Jack summarizes the tech-spec. choice for the offset Gregorian.

2. John summarizes the other ATA specs.
   150 for Ae/Tsys. seems now most likely
   Jill: what cost for various specs? Mike/John Drill — big discussion. Mike: There can be few changes beyond this point.

Issues: start planning the various buildings
6m dish agreed upon.
DSP: John Bunton, ~~Brad~~ Brent Carlson (?)

ATATAP June 11 or 12

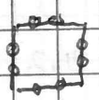
Heat conduction along the pyramid.



A break at 6"

I. A thin layer of copper plating over plastic ~ 1 mil thick

4" = 4000 mils 4000 x1 = 4000 square mils of copper

II 8 pins  30 mil diameter: 706 sq mils, x 8 = 5600 Total
or 20 mil diameter 314 , x8 = 2500 Total

With 1 mil of copper on each: π 30 = 94, x8 = 753 sq mils of copper
or π 20 = 63, x8 = 502

III 30 mil wall of stainless: 4000 x 30 = 120,000 sq mils
20 mil wall 4000 x20 = 80,000 .0005 m²
10 mil wall 4000 x10 = 40,000

IV heat flow through 8 30 mil stainless pins 30 mil ⇒ .00076 m
cross section π/4 (.00076)² = 4.5x10⁻⁷ m², x8 = 3.6x10⁻⁶ m²

Try rock gradient along a 1/4" pin. 1/4" = .006" m

Then $\frac{\Delta T}{\Delta x} = \frac{100}{.0064m} = 15,200\ ^\circ/m$

$H = KA\frac{\Delta T}{\Delta x} = (14)(3.6\times10^{-6})(15,200) = .79$ watts

8x20 mil pins would give .35 watts

1/4" is long for the electrical connection; 1/8" would be better. Would 1/2 mil of copper be ok for the skin effect? Also note the low temperature will have a higher thermal conductivity

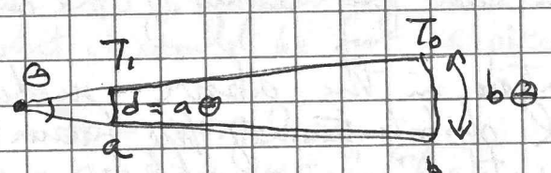
Gradient over the rest of the pyramid for plated stainless.
cross section is $5\times10^{-4}\ m^2$ for the 1" square 20 mils thick.
For a gradient of 200 over a length of 60 cm (.6m)

$\frac{\Delta T}{\Delta x} = \frac{200}{.6} = 333\ ^\circ/m$ $\quad H = (14)(5\times10^{-4})(333) = 2.3$ watts

---

<u>Better Model</u> Each wall of the pyramid, or each antenna arm, is constant thickness. Hence its approximately a cylindrical flow

ok



A fixed thickness t.

$\nabla^2 T = 0 \rightarrow T(r) \propto \ln(r) + \text{const.}$

$$T(r) = \frac{(T_0 - T_1)\ln(r/a)}{\ln(b/a)} + T_1 = \frac{(T_0-T_1)}{\ln(b/a)}(\ln r - \ln a) + T_1 \checkmark$$

$$H = \text{Heat Flow} = (b\theta)\,t\,k\,\frac{\partial T}{\partial r}\Big|_{r=b} = \frac{b\theta\, t k (T_0 - T_1)}{\ln(b/a)}\cdot\frac{1}{b} = \frac{\theta t k (T_0 - T_1)}{\ln(b/a)} \checkmark$$

For 4 walls: $4H = \frac{4\theta t k (T_0 - T_1)}{\ln(b/a)} \checkmark \cong \frac{4\theta t k (T_0 - T_1)}{-\ln a}$ (no) if $b >> a$.

Take $\theta = 10^\circ$, $b = 15cm$ (or 1.2m), $a = d/\theta = \frac{.15''}{10^\circ} = \frac{.38cm}{10^\circ} = 2.2cm$, $\underbrace{t = 5\times10^{-6}m,\ k = 401}_{\text{thin copper}}$
and $T_0 - T_1 = 300 - 80 = 220$

$$4H = \frac{4(.17)(5\times10^{-6})(401)220}{\ln(15/2.2)} = 0.16 \text{ watts just for the copper}$$

For stainless steel $k_s = 15$ so $k_s/k_c = \frac{15}{401} = .0374 = \frac{1}{26.7}$

Skin depth for copper: $\delta = \dfrac{1}{\sqrt{\pi f \mu \sigma}}$

At 300K $\sigma = 0.596 \times 10^{8}$ /ohm-m
$\mu = 4\pi \times 10^{-7}$

at 5GHz

$\delta = \left[\pi (5\times10^{9})(4\pi\times10^{-7})(.596\times10^{8})\right]^{-1/2} = .92\times10^{-6} = .92\,\mu m$

so a layer which is 5×10^{-6} is $\gtrsim 5$ skin depths. $\delta \cong 3\mu m$ @ 500 MHz

---

A 5μm layer is 0.2 ~~mm~~ mils and it is thermally equivalent to (26.7)(.2) = 5.3 ~~mil~~ mils of stainless.

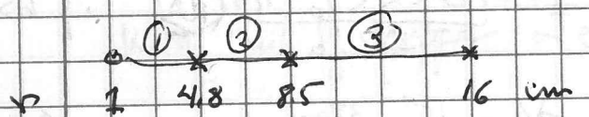
A 20 mil thick wall for the pyramid with 5μ copper plated on is equivalent to 25 mils of stainless with 5 times the thermal conductivity of just copper. The result is a heat flow of

$5 \times 0.16 = 0.8$ watts of heat

This doesn't include the arms of the antenna.

The temperature distribution in the above model is very different from the radial distribution of the foam. See the figure. Whereas the temperature rises quickly away from the center in the foam it rises more slowly along the metal walls of constant thickness. In parallel, the foam will add ~~drawn~~ heat to the pyramid. Thinning the walls toward the center would make the distribution more similar.

<u>An example of successive thinning of the pyramid walls</u>



$t_1 = 5$ mils, $t_2 = 10$, $t_3 = 20$

boundary conditions e.g. @ r = 4.8cm. $T_1 = T_2$

$$K_1 \frac{\partial T_1}{\partial r}\Big|_{4.8} = K_2 \frac{\partial T_2}{\partial r}\Big|_{4.8}$$

$T(1) = 80$ $\quad T(16) = 300$

Region ① $T_1(r) = a_1 + b_1 \ln r$ $\quad 1 \le r \le 4.8$

② $T_2(r) = a_2 + b_2 \ln(r/4.8)$ $\quad 4.8 \le r \le 8.5$

③ $T_3(r) = a_3 + b_3 \ln(r/8.5)$ $\quad 8.5 \le r \le 16$

At $r=1$ $\quad a_1 = 80$

At $r = 4.8$ $\begin{cases} a_1 + b_1 \ln(4.8) = a_2 \\ 5 b_1/4.8 = 10 b_2/4.8 \end{cases}$ $\quad 80 + 1.57 b_1 = a_2$, $\quad b_1 = 2 b_2$

At $r = 8.5$ $\begin{cases} a_2 + b_2 \ln\left(\frac{8.5}{4.8}\right) = a_3 \\ 10 b_2/8.5 = 20 b_3/8.5 \end{cases}$ $\quad a_2 + 0.57 b_2 = a_3$, $\quad b_2 = 2 b_3$

At $r = 16$ $\quad 300 = a_3 + b_3 \ln\left(\frac{16}{8.5}\right)$ $\quad 300 = a_3 + .63 b_3$

Simple substitutions give: $b_1 = 108.9$, $b_2 = 54.5$, $b_3 = 27.2$, $a_2 = 250.97$, $a_3 = 282.5$.

b_3 is the output gradient term. For the uniform piece its $(T_0 - T_i)/\ln b/a$. This is 79.3, greater by 2.91 (3 times). So the heat flow is 1/3 what it would be for a piece that's uniformly 20 mils thick.

$T_1(r) = 80 + 108.9 \ln r$, $= 280.8$ @ $r = 4.8$

$T_2(r) = 250.97 + 54.5 \ln(r/4.8)$, $= 282.1$ @ $r = 8.5$

$T_3(r) = 282.5 + 27.2 \times 2 \ln(r/8.5)$, $= 299.7$ @ $r = 16$

This distribution is much closer to the spherical distribution, so there should be much less cross heat flow between the foam ball and the pyramid.

Thermal conductivity of Copper

T (°K)	K watts/m°K
300	401
80	557
60	829
50	1250
9	24,900

more

T(°K)	K(watts/m°K)
300	401
250	406
200	413
150	429
100	482
90	508
80	557
70	647
60	829
50	1250

How much copper is needed on the stainless?

Reflection coefficient from a metal surface.

voltage: $\Gamma = -1\left\{1 - \dfrac{2}{(1+i)\sqrt{\dfrac{\sigma}{2\omega\epsilon_0}}}\right\}$; $\left|\dfrac{2}{1+i}\right| = \sqrt{2}$

For Copper at 300K $\sigma = .6\times10^{8}$ /ohm-m

At 5GHz $\sqrt{\dfrac{\sigma}{2\omega\epsilon_0}} = \sqrt{\dfrac{.6\times10^{8}}{2(2\pi\cdot5\times10^{9})}\dfrac{36\pi}{10^{-9}}} = 10^{4}$ ✓

$-\Gamma_c = 1 - 1.4\times10^{-4}$

Resistivity of stainless steel: $\rho = 90\times10^{-6}$ ohm-cm at 20°C. [$\rho = 1.72\times10^{-6}$ ohm cm for copper] 1.72×10^{-8} ohm-m

$\sigma_{ST} = 1/\rho = 1.11\times10^{6}$ /ohm-m

Here (at 5Hz) $\sqrt{\dfrac{\sigma_{ST}}{2\omega\epsilon_0}} = \left(\dfrac{1}{7.27}\right)\times10^{4} = 1.4\times10^{3}$

and $\Gamma_{ST} = -\left\{1 - 10^{-3}\right\}$ <- Γ_c

With one skin depth of copper on the stainless the greater loss in the reflection from the stainless is seen through a round trip through two skin depths of copper: $e^{-2} \sim 0.1$ and the effect of the stainless should be negligible.

Conclusion: Even just one skin depth of copper is probably adequate

Alternative estimate $\epsilon_{ST} = 1 - \Gamma_{ST}^2 = 1 - (1-10^{-3})^2 \cong 2\times10^{-3}$ fairly low

e.g. $\Delta T = 300\,\epsilon_{ST} = 0.6°K$.

$\epsilon_{Cu} = 1 - \Gamma_{Cu}^2 = 1 - (1 - 1.4\times10^{-4})^2 = 2.8\times10^{-4}$

Pass this through one skin depth of Copper $e^{-\delta} = e^{-2} = 0.1$

Cref $\epsilon_{ST}(0.1) = 2\times10^{-4}$; and $300K \times 2\times10^{-4} = .06K$.

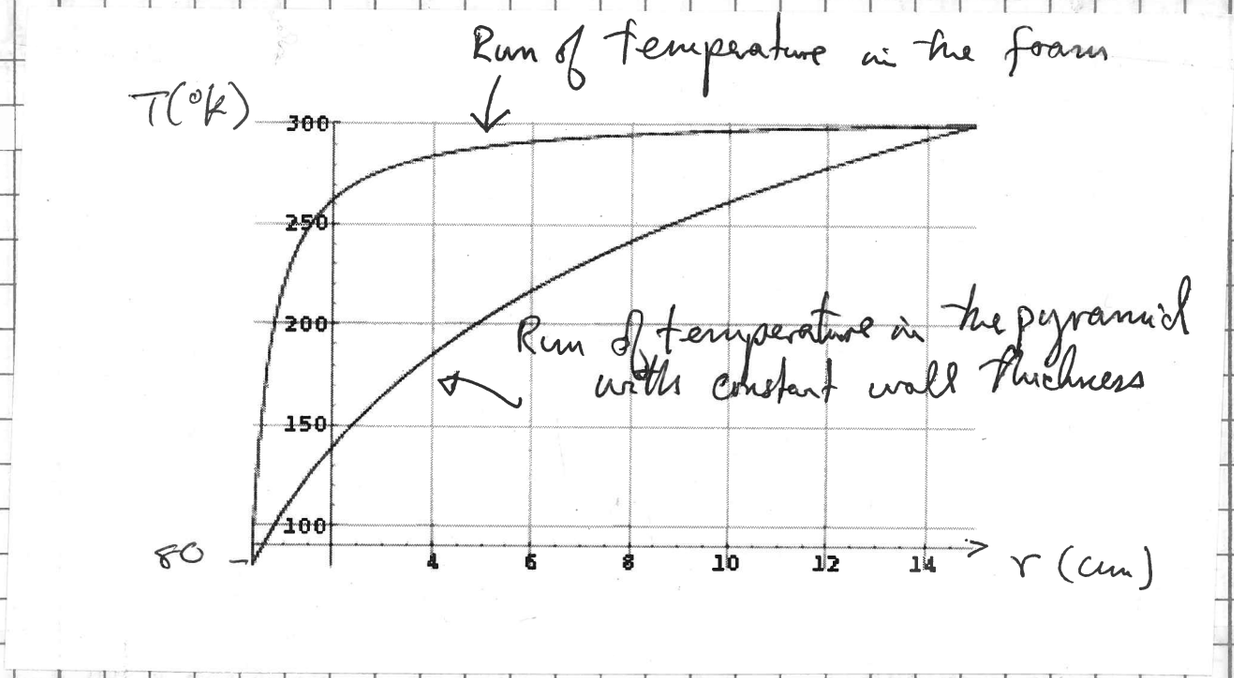
$t = (e^{-\delta})^2$; $e^{-\delta} = .36$, $(.36)^2 = .13$

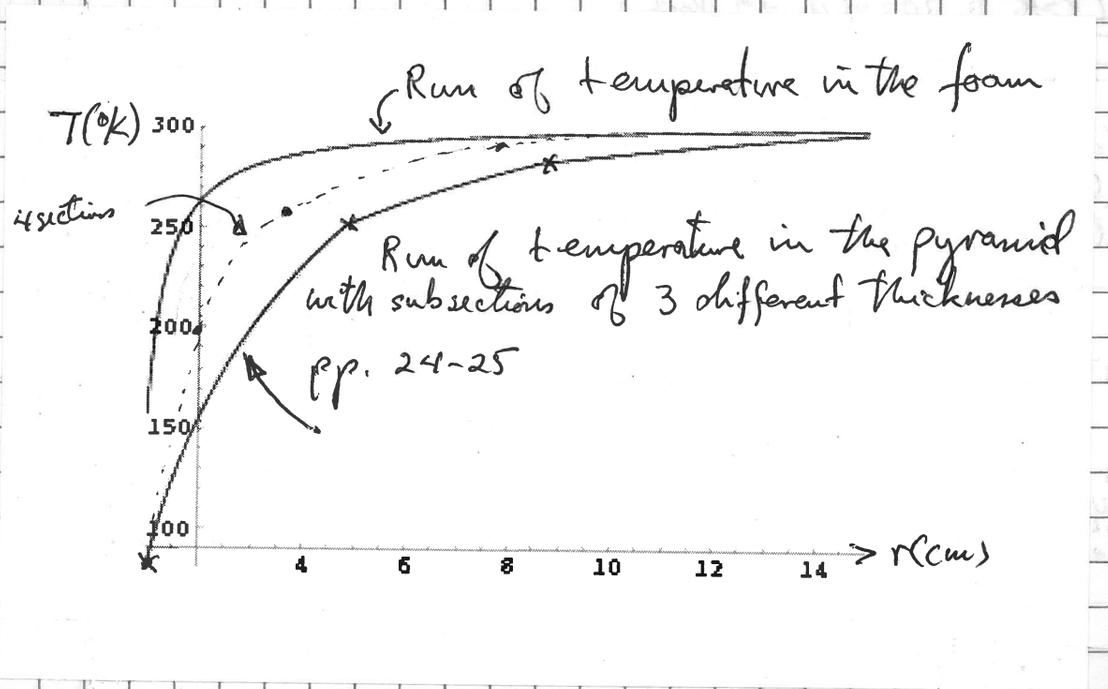
Once more $T_{out} \leftarrow$ [ ] $\epsilon_{ST}(300)$ 300K

$T_{out} = \epsilon_{ST}(300)\,e^{-2\delta} + (1 - e^{-2\delta})\,300\,\epsilon_{Cu}$

$= (2\times10^{-3})(300)(.13) + (1 - .13)\,300\;2.8\times10^{-4} = (.6)(.13) + .073 = 0.151$

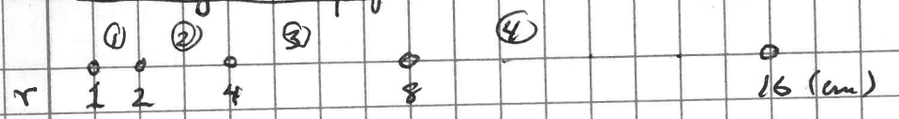
$.078 + .073$





The heat flow in this case is <u>3 times less</u> than that of the above.

Thinning the pyramid walls into 4 sections



$t_1 = 5,\ t_2 = 10,\ t_3 = 20,\ t_4 = 40$

Region 1 $\quad T_1(r) = a_1 + b_1 \ln r$ $\qquad T_1(1) = 80$

2 $\quad T_2(r) = a_2 + b_2 \ln(r/2)$

3 $\quad T_3(r) = a_3 + b_3 \ln(r/4)$

4 $\quad T_4(r) = a_4 + b_4 \ln(r/8)$ $\qquad T_4(16) = 300$

At $r=1$: $\quad a_1 = 80$

At $r=2$: $\begin{cases} a_1 + b_1 \ln 2 = a_2 \\ 5b_1 + 10b_2 \end{cases}$

At $r=4$: $\begin{cases} a_2 + b_2 \ln 2 = a_3 \\ b_2 = 2b_3 \end{cases}$

At $r=8$ $\begin{cases} a_3 + b_3 \ln 2 = a_4 \\ b_3 = 2b_4 \end{cases}$

At $r=16$ $\quad 300 = a_4 + b_4 \ln 2$

With a little arithmetic find.

$T_1(r) = 80 + 170.5 \ln r$
$T_2(r) = 197.6 + 82.25 \ln(r/2)$
$T_3(r) = 256.4 + 42.63 \ln(r/4)$
$T_4(r) = 285.9 + 21.3 \ln(r/8)$

Points are plotted on previous page.

At the bottom end with an 1/2 of copper: $t = 5$ mils $= .00013$ m
$d = .15'' = .0038$ m
$k_{st} = 15$ watt/m°K

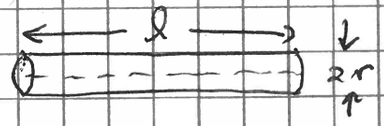
Pyramid
Heat Input (4 walls)

$$4H = 4\,d\,t\,k \left.\frac{\partial T}{\partial r}\right|_{r=1cm} = 4(.00013)(.0038)15(170.5)/(.01)$$

$$= \underline{0.5 \text{ watts}}$$

Careful here. 3/2 of Copper are required for 500 MHz, not 1. Stainless is also required. The real answer should be 1.5 - 2.0 watts.

Inductance of a straight wire in space [Montgomery et al, p.168]



The inductance per unit length is $L_o = \frac{4\pi \times 10^{-7}}{2\pi} \ln\left(\frac{2l}{er}\right)$ henrys/m (w)

With $r = 10$ mils $= .00025$m and $l = .075'' = .0019$m.

$$L_o = 2\times10^{-7} \ln\left[\frac{(2)(.0019)}{(2.72)(.00025)}\right] = 2\times10^{-7} \ln(5.588) = 3.4\times10^{-7} \text{ h/m}$$

For the length, $l = .0019$m; $L_o l = .65\times10^{-9}$ Henrys $= \underline{.65 \text{ nH}}$

If $r = 5$ mils $= .00013$, $\underline{L_o l = .92 \text{ nHenrys}}$

If $r = 20$ mils $= .00050$m $\underline{L_o l = 0.3 \text{ nH}}$

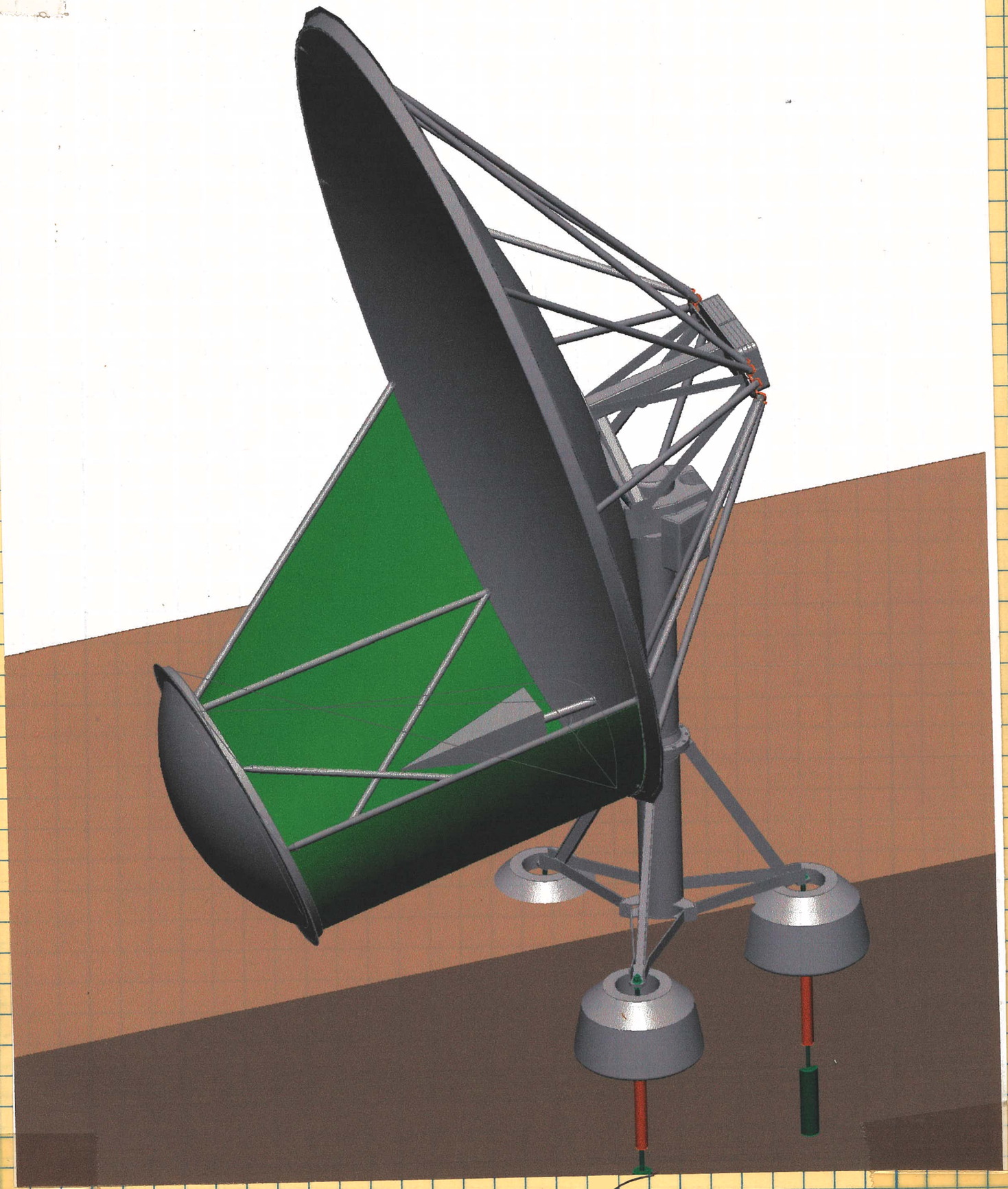
$r = 40$ mils $= .001$m $\underline{L_o l = 0.13 \text{ nH}}$

r (mils)	$L_o l$ (nH)
5	.92
10	.65
20	.30
40	.13

For $r = 40$ mils (1) and $l = 50$ mils (1.2 mm)

$L_o l = .08$ nH

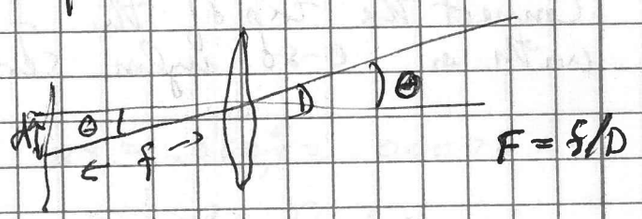
Matt's Design for 6m offset 3/6/04



Size of the beam waist for the primary mirror

It's (approximately) the ~~[illegible]~~ geometric image of the primary diffraction beam in the focal plane.

$$\Theta_{1/2} = \frac{1.2\lambda}{D}$$



$$d = f\,\Theta$$

$$d_{1/2} = f\,\Theta_{1/2} = F\cancel{D}\,\frac{1.2\lambda}{\cancel{D}} = 1.2\,F\lambda$$

$$4d_{1/2} = f\,4\Theta_{1/2} = 4.8\,F\lambda = 1.9\lambda \cong 2\lambda \quad \text{for } F = 0.4$$

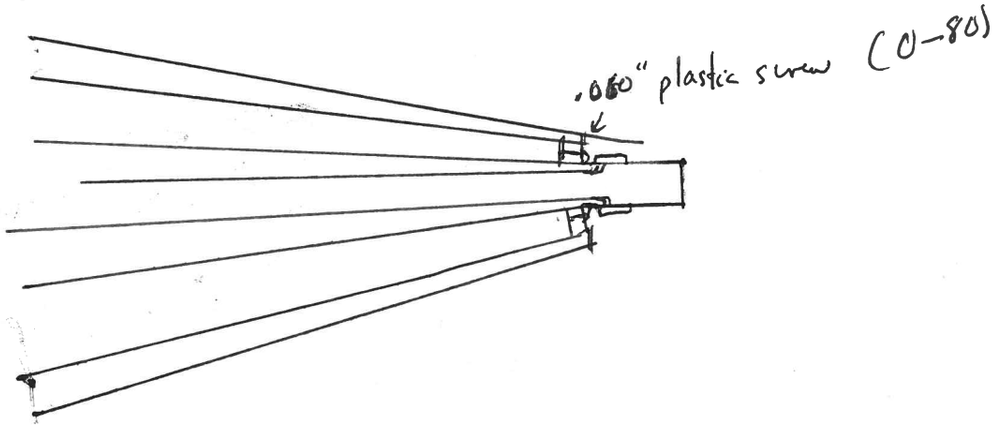
$\underbrace{4d_{1/2} \cong 2\lambda}$ At $4\Theta_{1/2}$ for the primary, $G_{max} \sim 0$ dB.

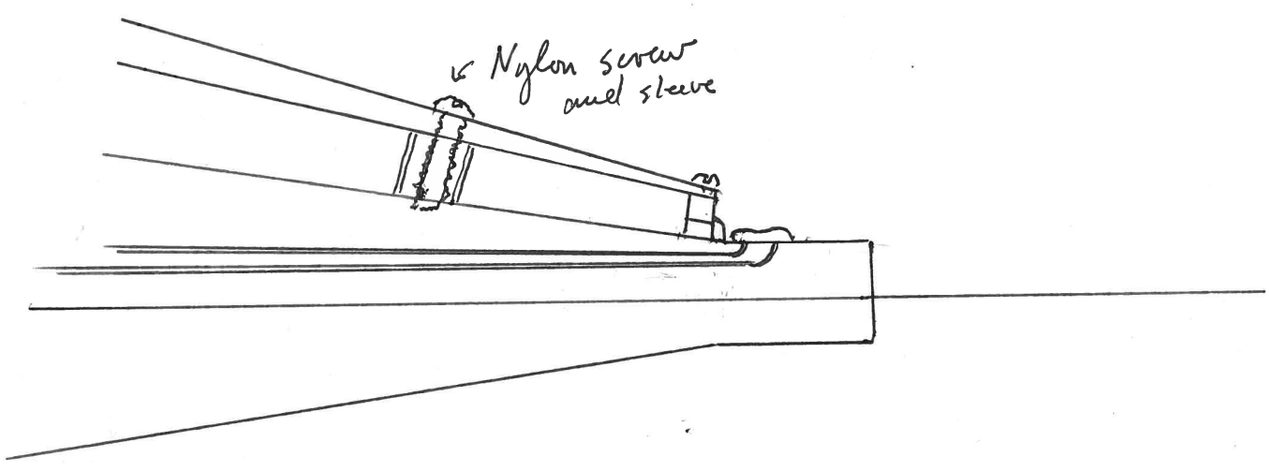
For the secondary, where $F = 0.65$, $4d_{1/2} = 4.8F\lambda = \underline{\underline{3\lambda}}$

So, using 4λ should be quite conservative.

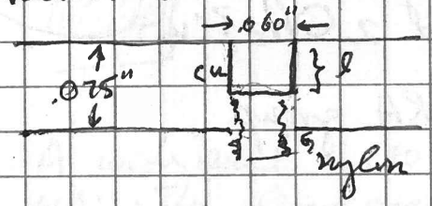
Possible Input Connection scheme for a room temperature amplifier

Connect the tip of the feed arm to the pyramid with a 0-80 nylon screw (.060" diameter)





Rough estimates of the inductance and capacitance of the metal sleeve around the top of the screw.



I. <u>Inductance of the post</u> $\qquad 2r = .060'' ; r = .00076\ m.$

$$L_0 l = 2\times10^{-7} l \ln\left\{\frac{2l}{er}\right\} = 2\times10^{-7} l \ln\{967 l\}$$

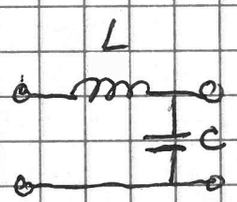
l (mils)	m	$L_0 l$ (h)	X_L (10 GHz)
40	.0010	-4.9×10^{-12}	
45	.0011	1.3×10^{-11}	
50	.00125	4.2×10^{-11}	
55	.00138	7.9×10^{-11}	4.9 Ω
60	.0015	1.1×10^{-10}	
65	.00165	1.5×10^{-10}	

II <u>Excess Capacitance of the post</u> $\qquad \Delta c = \epsilon A\left(\frac{1}{.075-l} - \frac{1}{.075}\right)$

Use $\epsilon = 1.5\ 10^{-9}/36\pi$ $\qquad A = \pi r^2 = \pi(.00076)^2$

$\epsilon A = 2.4\times10^{-17}$

l	m	Δc (fd)	X_c	$Z_0 = \sqrt{L/C}$
50	.00125	1.1×10^{-14}		65 Ω
55	.00138	2.1×10^{-14}	768 Ω	61
60	.00150	3.4×10^{-14}		57
65	.00165	7.0×10^{-14}		46



$$Z_0 = \sqrt{\frac{L}{C}}$$

March 19 up/down converter meeting

What IF band to the ~~GHz~~ digitizer? 1 GHz
BPF 13.8 - 14.8 increase to 2 GHz? (no)

Steve Mass JPD met with the NFRA group.
1) active balun looks ok -
2) 18 db roll off in gain
3) LO leakage very high - with single ended mixer

What polarization isolation is needed? better than 30 db
chips should be identical
Go for separating the polarizations -

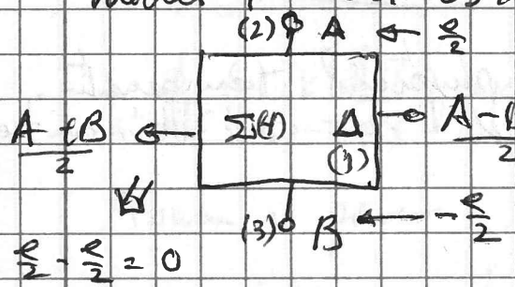
Can one make a Cauer-Tschebyschef as a band pass filter at 13.8 - 14.8 GHz? What image rejection can one get?

RF/IF monitoring - at some point after BP filter and before the ADC.

Passive Balun for the balanced LNA amps

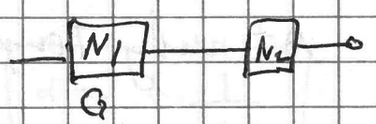
e is the antenna input voltage

A wideband 180° hybrid can be used.
NB: The Omnispectra Catalogue lists a 1-12.4 GHz model {# 2031-6340-00 psi}. It has max losses of 2db.



$$\frac{A-B}{2} \Rightarrow \frac{\frac{e}{2}-\left(-\frac{e}{2}\right)}{2} = \frac{e}{2} \rightarrow \text{RCVR}$$

The following amp at 300°K temperature may be 1db noise figure. Adding the 2db losses from the hybrid gives $N_2 = 3$ db (2)



$$N_t = N_1 + \frac{N_2 - 1}{G} \Rightarrow N_1 + \frac{1}{G}$$

With G = 20db (100) the second stage contribution is .01 or 3K. With G = 30db (1000), the second stage contribution is ~ 0.3K.

What are the effects of imperfections in this unit?
→ Measure the one that was delivered from Krytar

1) Make Full sized antenna elements
2) Could Prada finish the calculations? or Ya Ya?
3) Sandy's current run (~July) 45-50 k at room temp.
   25 db gain (x300)
4) Greg did a pattern simulation which showed that adding a 0.5" box to the nose of the feed had no effect on the pattern.

Receiver temperatures (LNA) vs. physical temperature.
Sandy guesses that his LNA will do 45-50 k at room temperature

His other measured values

T	T_N	$10 + 0.11(T-10)$
10°K	15	10
80K	23	23
300K	45-50	47

The interpolation is pretty good. At nearby temperatures:

100 K	25
120 K	27
150K	30k

More on the possibility of cooling the receiver at the tip.

The conductivity of copper depends on T (p. 25), so this should be included in the temperature distribution calculation.

Derivation of the temperature equation:

$dq = \rho c \, dT$ heat storage

$\nabla \cdot \vec{j} + \dot{q} = 0$ flow conservation

$\vec{j} = -K \nabla T$ heat flow due to the temperature gradient

$$\frac{dq}{dt} = \rho c \frac{dT}{dt} = -\vec{\nabla} \cdot \vec{j} = \nabla \cdot (K \nabla T)$$

with partial derivatives: $\rho c \frac{\partial T}{\partial t} = \nabla \cdot (K \nabla T)$

For the spherical heat flow T is $T(r,t)$ only.
But since k is $k(T)$, it is also $k(r)$.

The steady state distribution satisfies $\nabla \cdot (k \nabla T) = 0$

Let $k(T) = k_0 T^{\gamma}$, which can probably fit the data of p. 25.

$\nabla \cdot (k_0 T^{\gamma} \nabla T) = 0$ with $T = T(r)$ becomes.

$$\frac{1}{r^2}\frac{d}{dr}\left(r^2 k \frac{dT}{dr}\right) = 0 \qquad \frac{d}{dr}\left(r^2 k_0 T^{\gamma} \frac{dT}{dr}\right) = 0$$

Assume a power law solution $T \propto r^{\alpha}$

$$\frac{d}{dr}\left(r^2 T^{\gamma} \frac{dT}{dr}\right) = \frac{d}{dr}\left(r^2 (r^{\alpha})^{\gamma} \frac{d}{dr}(r^{\alpha})\right) = 0$$

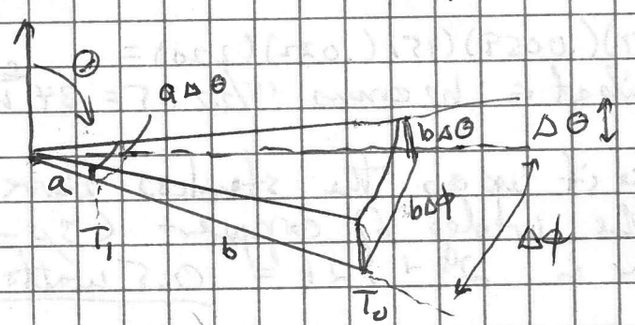
$\frac{d}{dr}\left(r^2 r^{\alpha\gamma}\, \alpha r^{\alpha-1}\right) = 0$ is satisfied if $2 + \alpha\gamma + \alpha - 1 = 0$
$1 + \alpha\gamma + \alpha = 0$

or $\boxed{\alpha = -\frac{1}{\gamma+1}}$ When $\gamma = 0$, $\alpha = -1$ the normal solution

so a general solution is $T(r) = A + \frac{B}{r^{\frac{1}{\gamma+1}}}$ not quite

No, this is just a start on the calculation. It's non-linear

If the pyramid wall is to be thinned out toward the tip so that the run of temperature from the cold tip is close to what it is in the foam ball, approximating the wall as a wedge in spherical coordinates would be a good choice. It should have the same temperature distribution as the foam ball.



$\nabla^2 T = 0$

$$\frac{1}{r^2}\frac{\partial}{\partial r}\left[r^2 \frac{\partial T}{\partial r}\right] = 0$$

(with no temp. dependence of k.)

As for the ball

$$T(r) = \frac{T_1 a - T_0 b}{(a-b)} + \frac{(T_0 - T_1)ab}{r(a-b)}$$

$$T(a) = \frac{T_1 a - \cancel{T_0 b} + \cancel{T_0 b} - T_1 b}{(a-b)} = T_1$$

$$T(b) = \frac{\cancel{T_1 a} - T_0 b + T_0 a - \cancel{T_1 b}}{(a-b)} = T_0$$

$\Delta\phi = 10° \rightarrow 0.17$ rad

$a\Delta\phi = .15$ in $= .38$ cm

so $a = .38/.17 = 2.2$ cm $= \underline{.022\ m}$

For $a\Delta\theta = \underline{5\ \text{mils}} = .013$ cm $= .00013$ m, $\Delta\theta = .013/2.2 = \underline{.0059\ \text{rad}}$

Heat flow through outer surface: $H = (b\Delta\theta)(b\Delta\phi)k \left.\frac{\partial T}{\partial r}\right|_{r=b}$

$$H = b^2 \Delta\theta\Delta\phi k \left.\frac{(T_0 - T_1)ab}{-r^2(a-b)}\right|_{r=b} = \Delta\theta\Delta\phi\, k(T_0 - T_1)\frac{ab}{-(a-b)}$$

$$H \cong \Delta\theta\Delta\phi\, ka(T_0 - T_1) \quad \text{for } b \gg a.$$

I. <u>For pure copper with $\langle k \rangle \cong 450$</u>

a) for $a\Delta\theta = 5$ mils, and $\Delta\theta = .0059$, $T_0 - T_1 = 300 - 80 = 220$

$$4H = 4\Delta\theta\Delta\phi\langle k\rangle a(T_0 - T_1) = 4(.17)(.0059)(450)(.022)(220) = \underline{8.7\ \text{watts}}$$

b) for $a\Delta\theta = 3\mu$, 1 skindepth @ 500 MHz, at the input

$\Delta\theta = 3\times10^{-6}/.022 = .00014$.

$$4H = 4(.00014)(.17)(450)(.022)(220) = \underline{0.21\ \text{watts}}\ \text{total}$$

II. <u>For pure stainless steel,</u>

a) $a\Delta\theta = 5$ mils and $\Delta\theta = .0059$, $K_s = 15$ watt/m°C

$$4H = 4(.17)(.0059)(15)(.022)(220) = \underline{.29\ \text{watts}}\ \text{total}$$

5 mils at a becomes $15/2.2 \times 5 = 34$ mils at b.

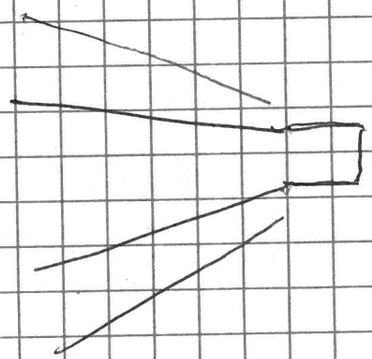
If one could make it up as the stainless above (5–34 mils) plated with the wedge of copper (3μ – 20μ), the total power flow is $.29 + .21 = \underline{0.5\ \text{watts total}}$

The cross section

Doubling the stainless to 10 mils at the input, 68 mils at the top, would double the stainless heat loss to .58 watts. The temperature distribution would match that of the foam and there would be no cross flow. The foam would add another .235 watts flow.

stainless (10 mils at input)	0.58
copper plate (3 μ at input)	0.21
foam ball	0.24
	1.03 watts total

Adding the antenna arms made in the same way would roughly double the load. A little more because the copper must be on both sides, but the foam needs to be considered only once. Altogether about 2 watts total.
Wires must also be included.



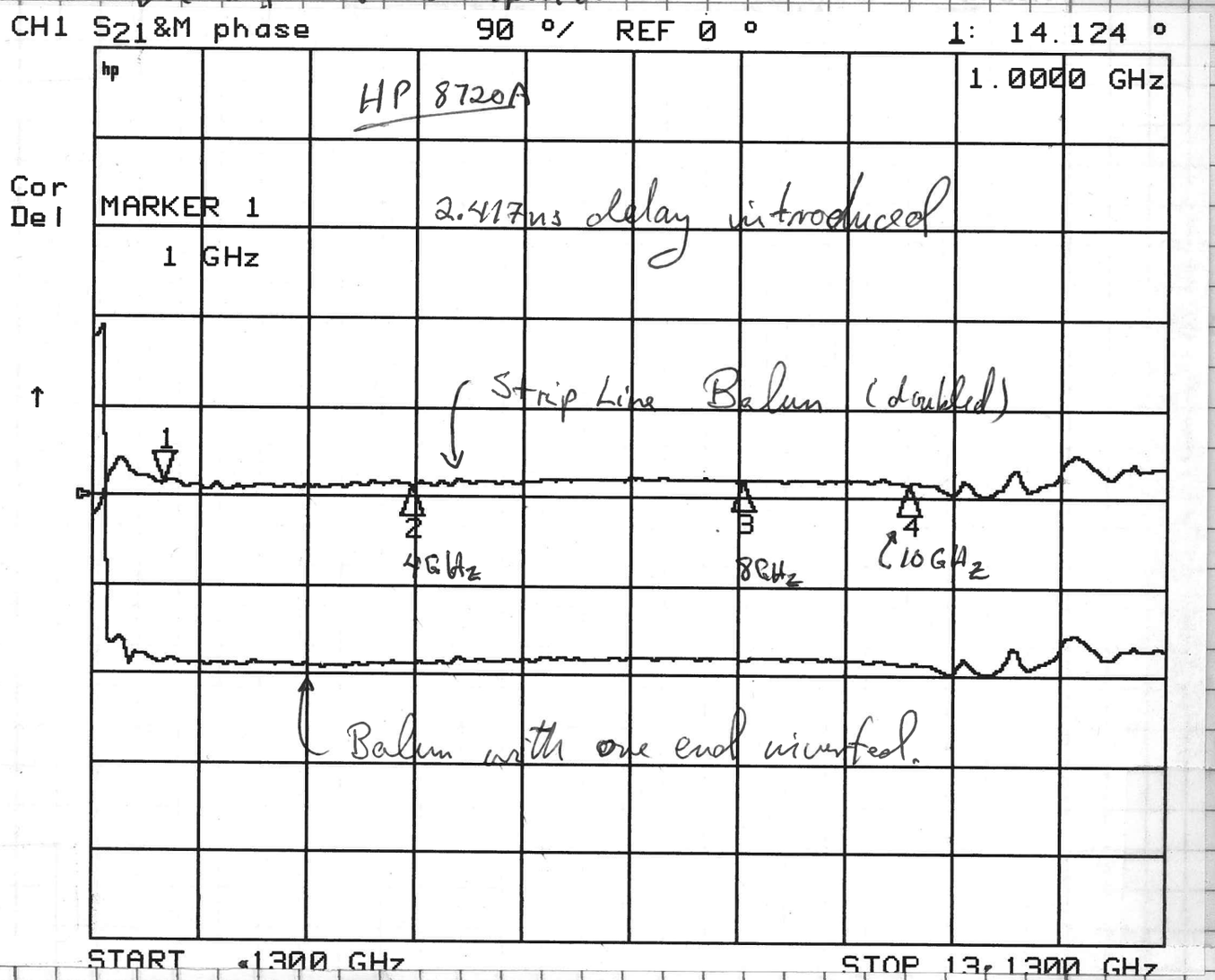
Operating at a higher temperature would lower the heat loss. e.g. at 150K the heat loss would be 68%. At 150K the LNA temperature is ~ 38K. If the new LNA chips get to 15K at temp = 60K, instead of 23 as at present, we might get 20K from that new LNA a 150K temperature. With 9K from the partially cold antenna no transmission line losses, ~~and~~ 6K from the atmosphere and 4K from space, we should get 39K total.

Further measurements on the prototype passive Balun designed by Gary.

7/23/01 NJW

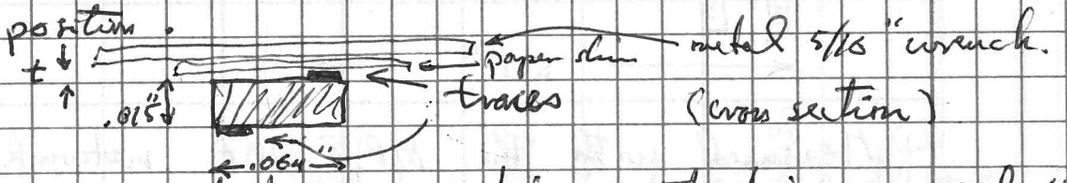
I. Two units were built. The first makes a transition from 50Ω unbalance (SMA connector) to the balanced line at 220Ω and then back through an identical balun to the 50Ω unbalance (SMA connector). The second unit is identical except that on the return to 50Ω the line is inverted and the ground at the unbalance is on top.

After the HP network analyzer was calibrated, the first unit was inserted and S_{21} measured. Phase vs freq. was displayed and 2.417ns of delay was entered to produce a (nearly constant) phase over 1-10 GHz, see below. Then the second unit was inserted in place of the first. They both have the same length. With no further adjustment, the phase changed by 180° as expected.



II. Sensitivity of the balanced line of the balun to nearby grounds.

The same double balun was used and the change in transmission loss at 8 GHz was noted as a function of the separation (distance) from the line to a nearby metal ground. The location on the line was at the 220 or balanced position.

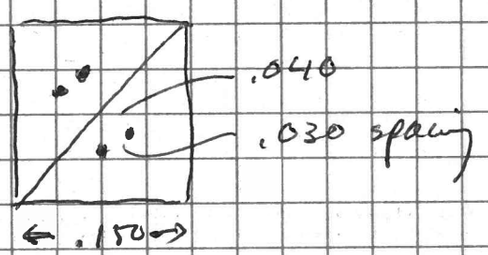


The distance was set by paper shims. The shims were each 4 mils

t	loss @ 8 GHz (total)	
.004" paper only	− 2.15 db	± 0.1 db drift
.004 + bar	− 3.30	
.008 + bar	− 2.76	
.012 + bar	− 2.50	
.016 + bar	− 2.46	
.020 + bar	− 2.41	
.024 + bar	− 2.50	
removing bar	→ − 2.40 db	(same drift here)

At separations ≳ .020 inch (or even .016 inch) the bar had little, if any, effect. The field lines must be largely confined to the dielectric.

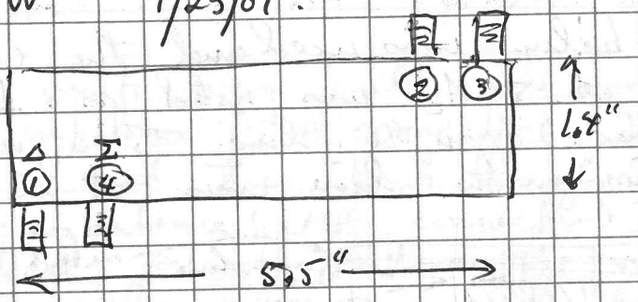
On the feed, the traces are soldered to short .010" wires that are soldered into the end circuit board.



The strip line should be ≳ .040" away from any metal ground. Losses (at 8 GHz) should be ≲ 0.1 db (2.5%)

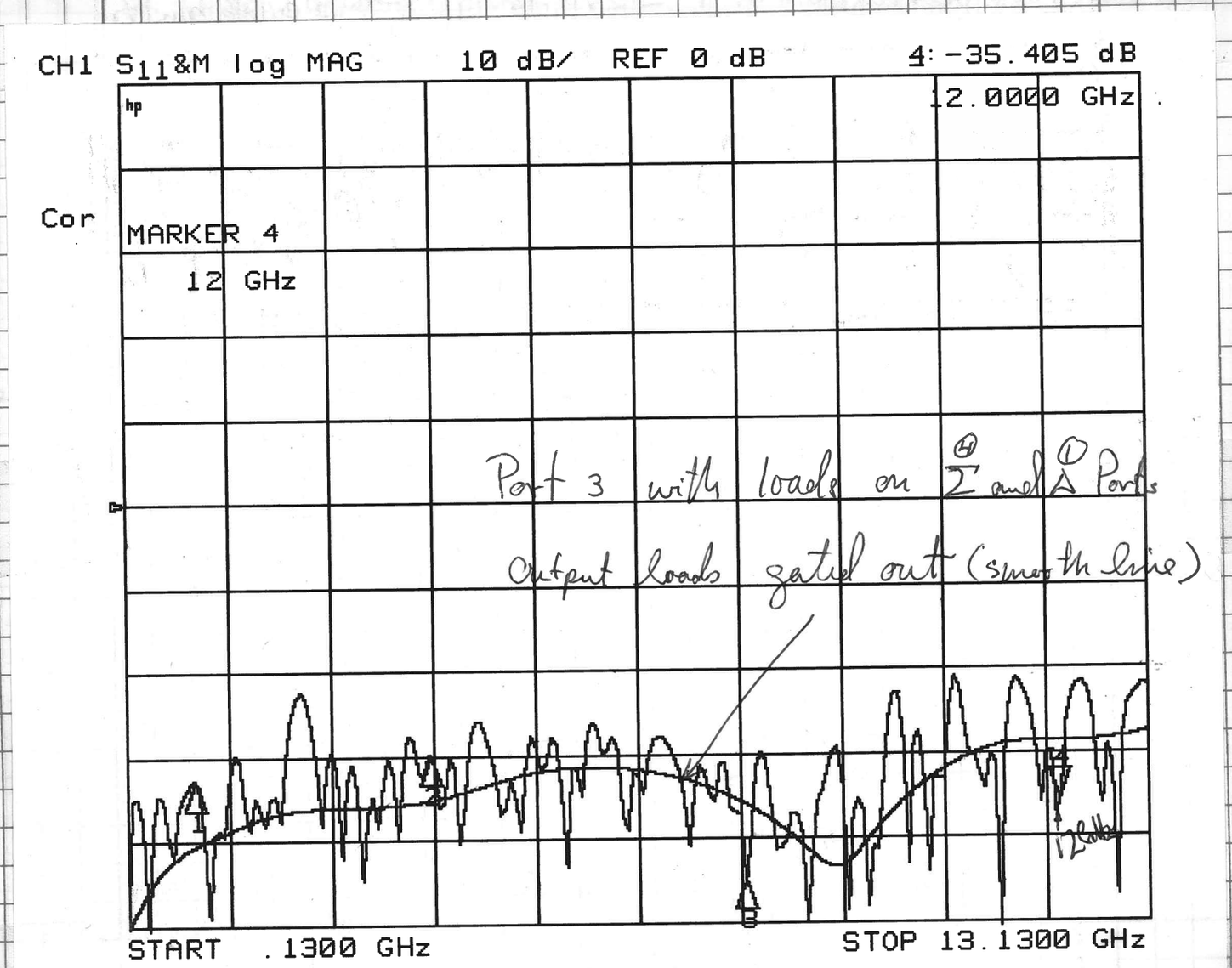
Tests of the KRYTAR model 4010124 180° 3db coupler.

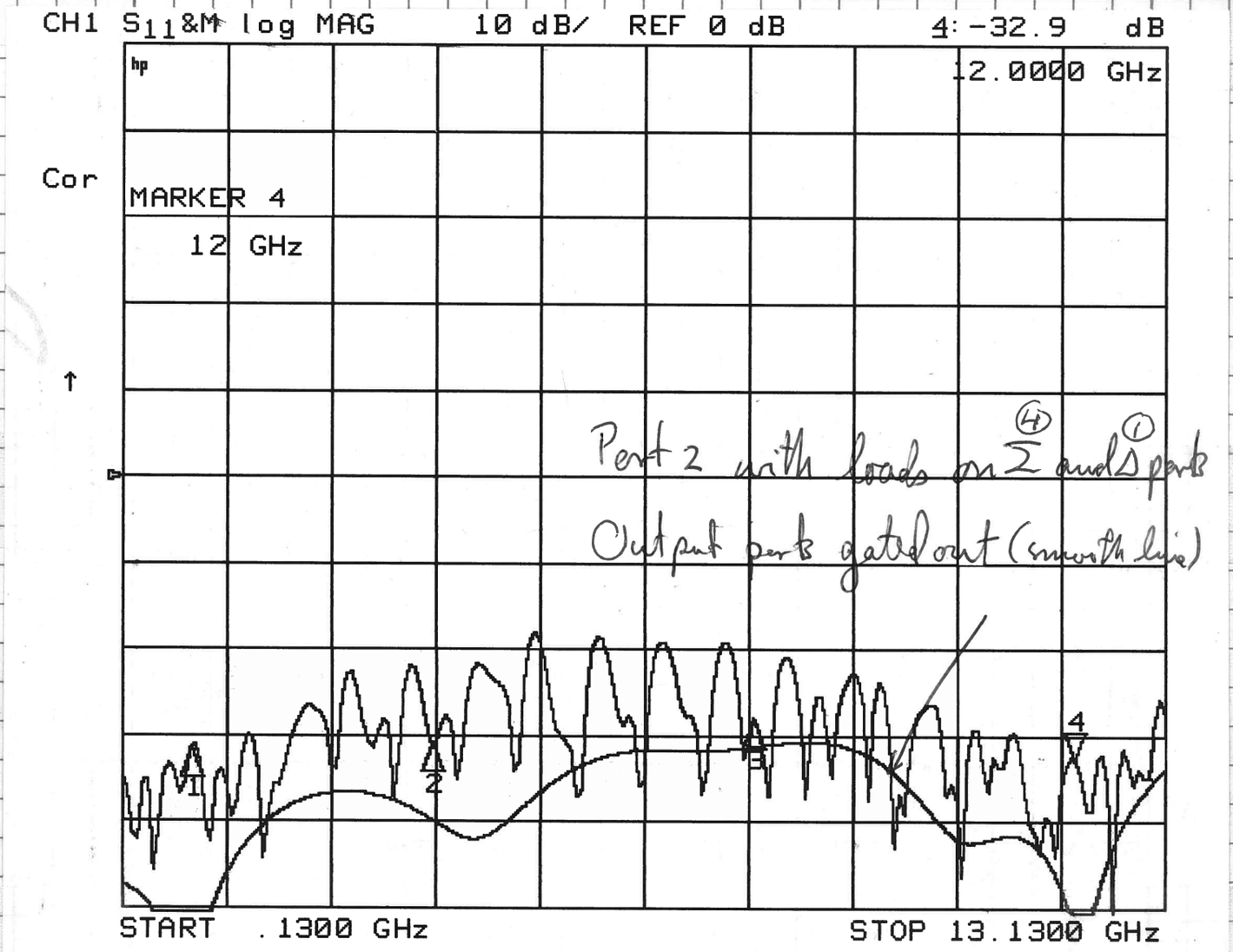
WJW 7/23/01.



Measured with the HP 8720A network analyzer.

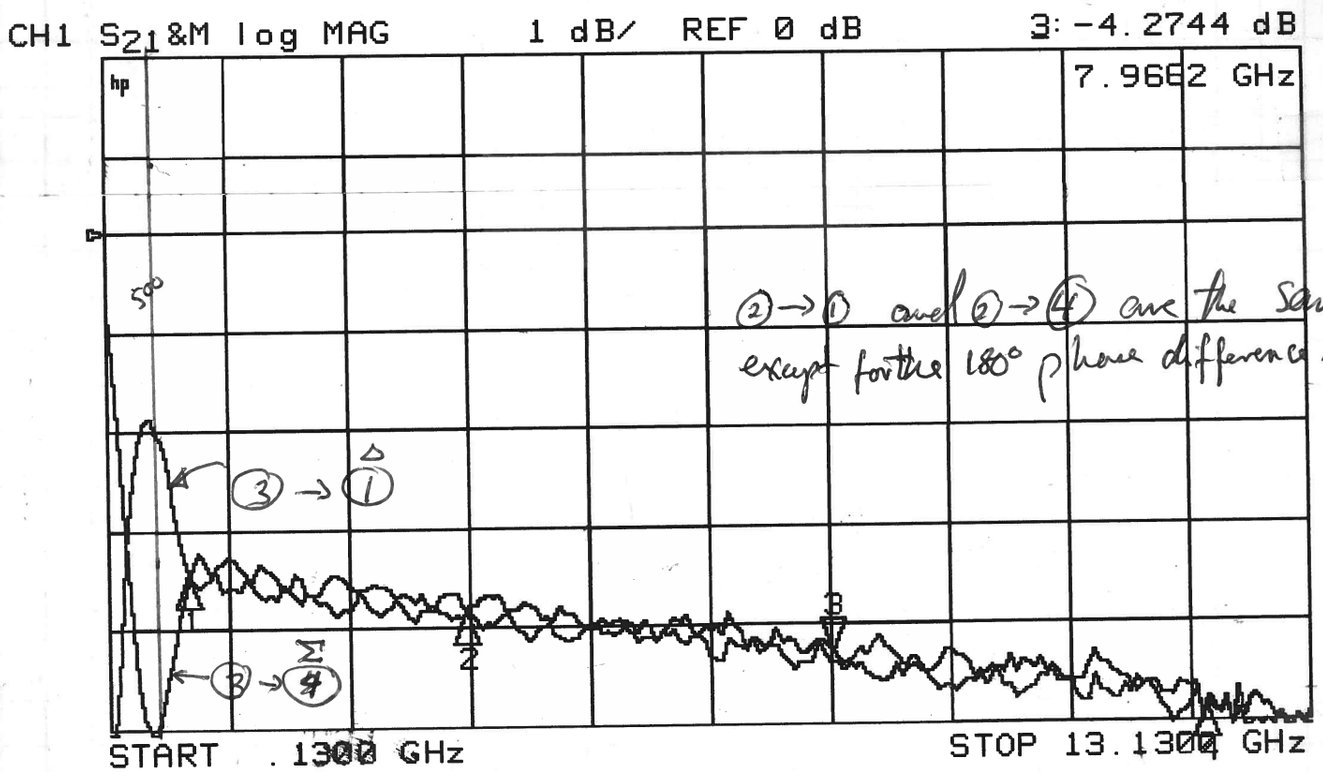
Input match and phase and amplitude balance are recorded.

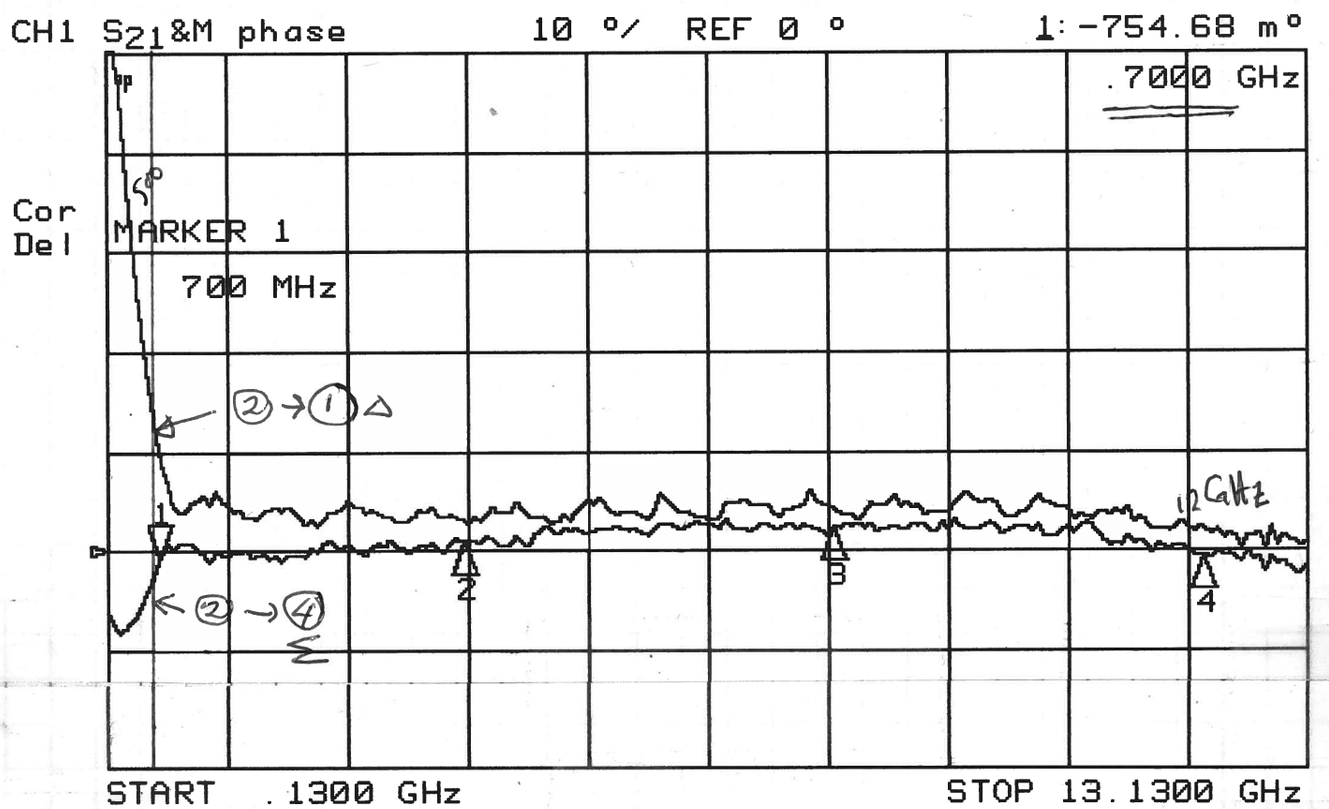




The loads are pretty good here. Even at 10 GHz, they are ≲ -25 db.

The next page shows transmission phase and amplitude balance.

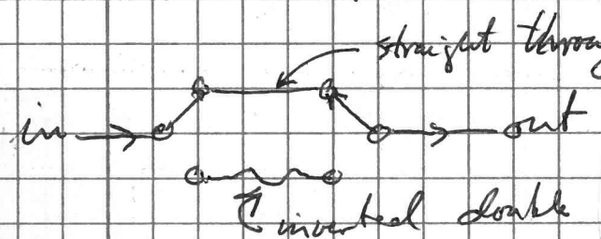

CH1 S21&M log MAG 1 dB/ REF 0 dB 3: -4.2744 dB
7.9662 GHz
50°
② → ① and ② → ④ are the same except for the 180° phase difference.
③ → ①
② → ④
START .1300 GHz
STOP 13.1300 GHz


CH1 S21&M phase 10 °/ REF 0 ° 1: -754.68 m°
.7000 GHz
Cor
Del
MARKER 1
700 MHz
50°
② → ①
② → ④
12 GHz
START .1300 GHz
STOP 13.1300 GHz

7/3/01 WJW

A Potential Phase Switch for the ATA first LO

This is based on the two double baluns tested on p. 40.



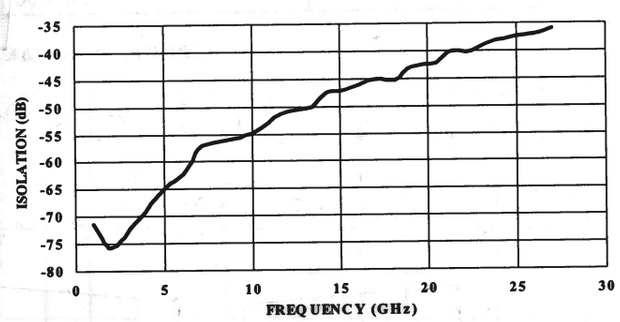
This will be for 1.5 GHz - 2.5 GHz instead of 1-10 GHz.
Only 1 octave BW is required, not 3-4 octaves.

The MA Com SPDT switch MA4SW210 might work.
Price ~ 4$ / unit

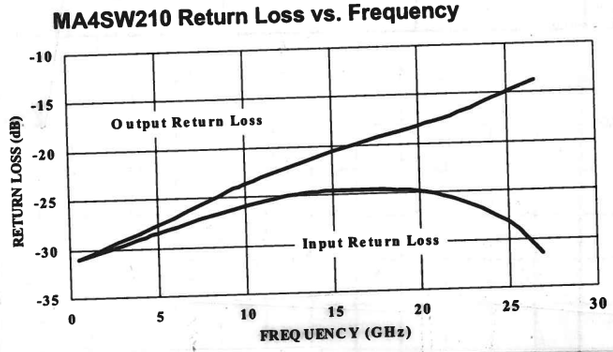
MA4SW210 (SPDT)
Electrical Specifications @ T_A = +25°C, +/- 20 mA Bias Current

Parameter	Frequency	Minimum	Nominal	Maximum	Units
Insertion Loss	6 GHz	-	0.4	0.7	dB
	13 GHz	-	0.5	1.0	dB
	20 GHz	-	0.7	1.2	dB
Isolation	6 GHz	48	63	-	dB
	13 GHz	40	50	-	dB
	20 GHz	34	42	-	dB
Input Return Loss	6 GHz	20	27	-	dB
	13 GHz	18	25	-	dB
	20 GHz	15	25	-	dB
Switching Speed[1]	-	-	20	-	ns
Voltage Rating[2]	-	-	-	50	V

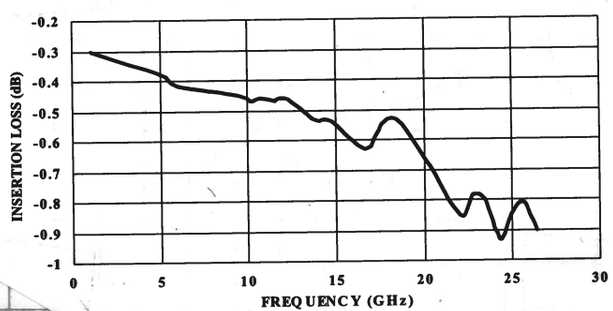
MA4SW210 Isolation vs Frequency

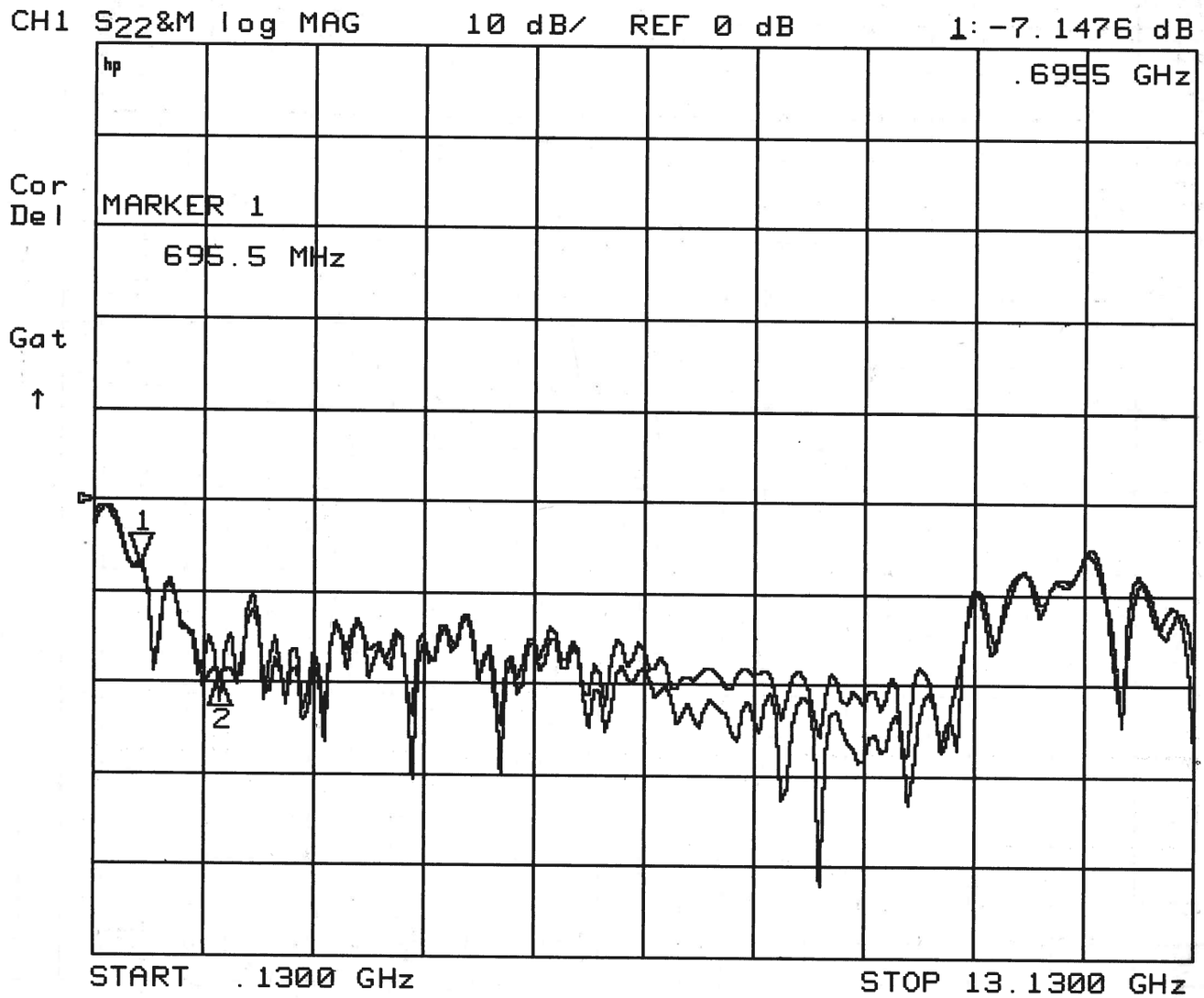


MA4SW210 Return Loss vs. Frequency

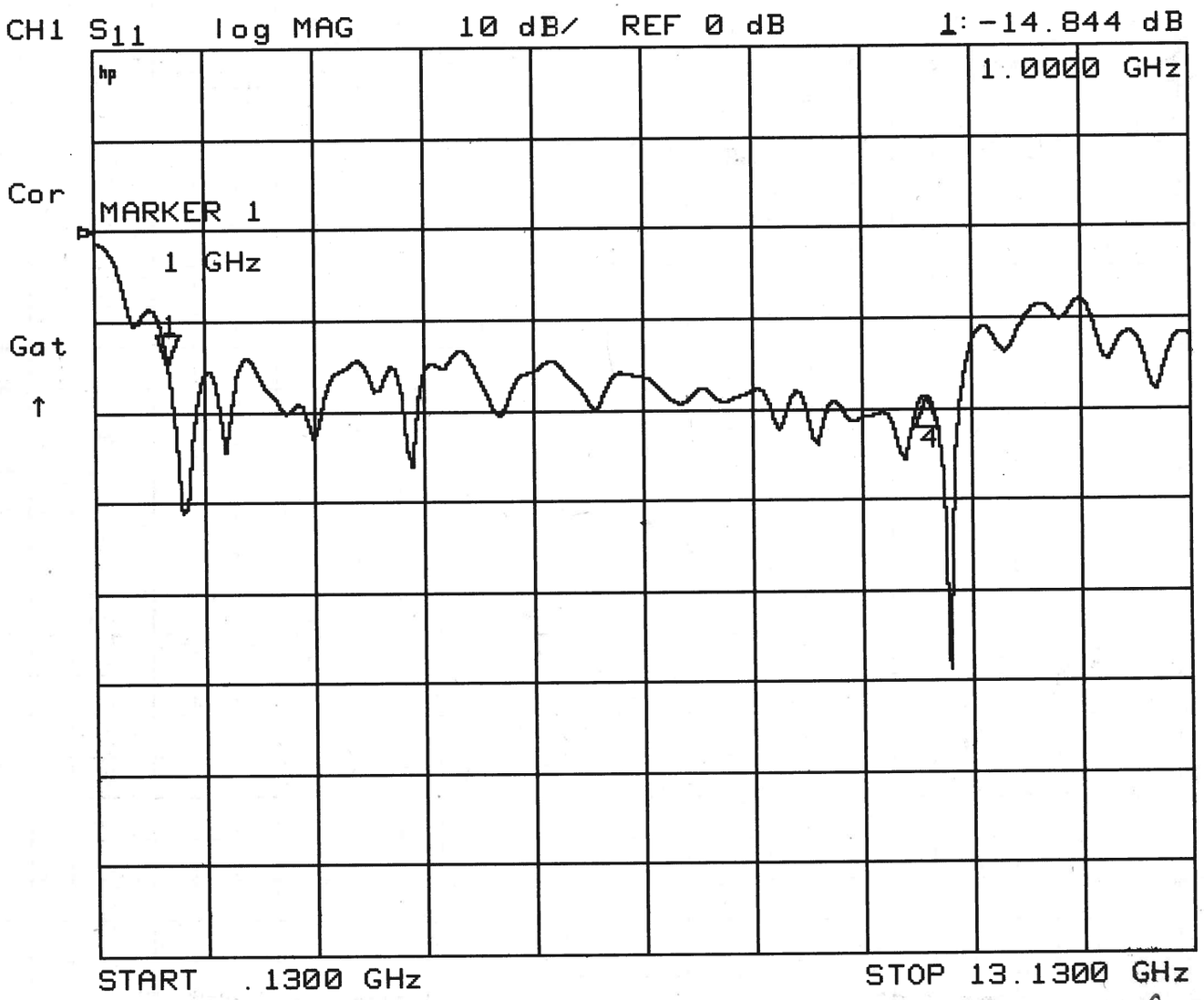


MA4SW210 Insertion Loss vs Frequency

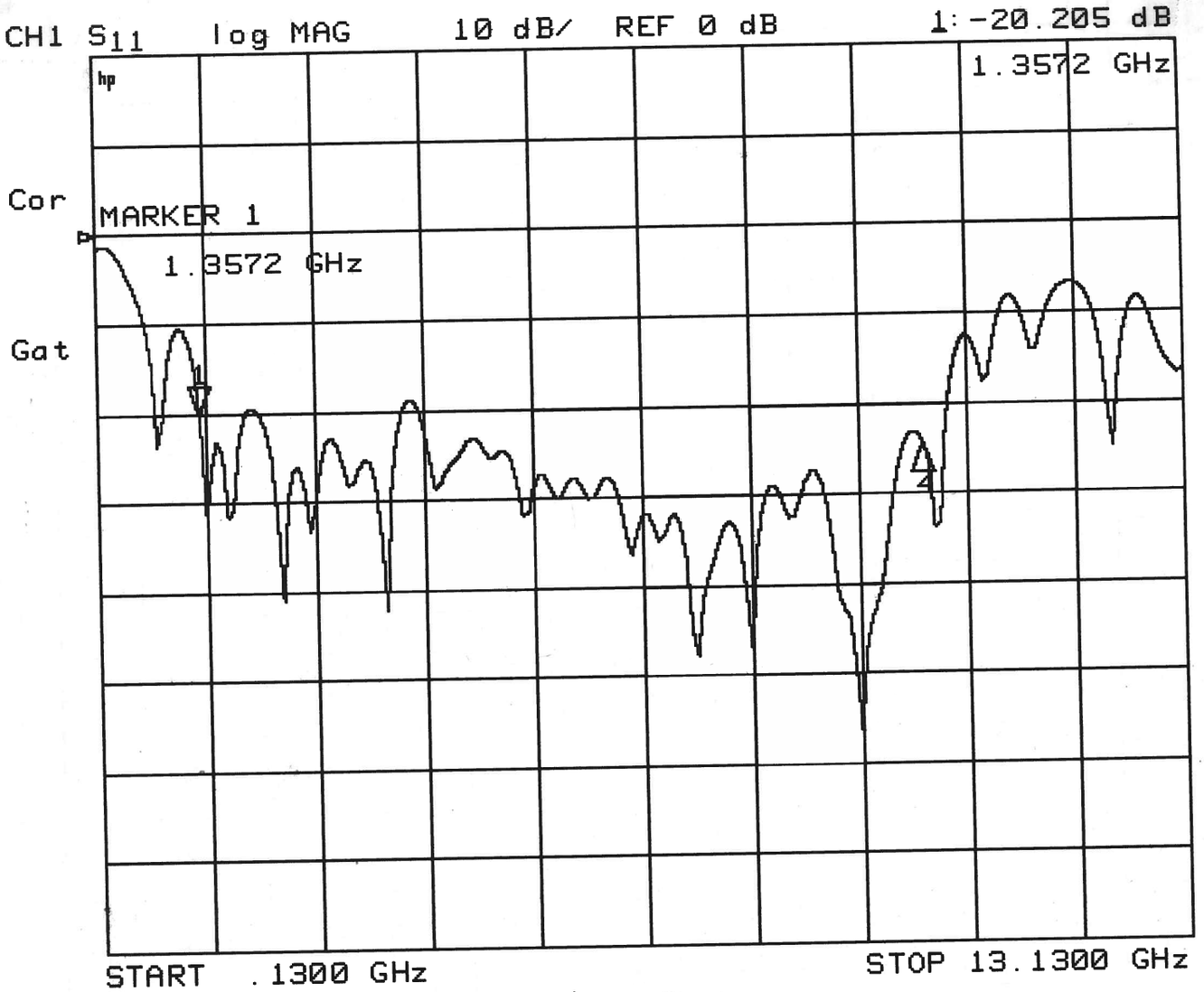




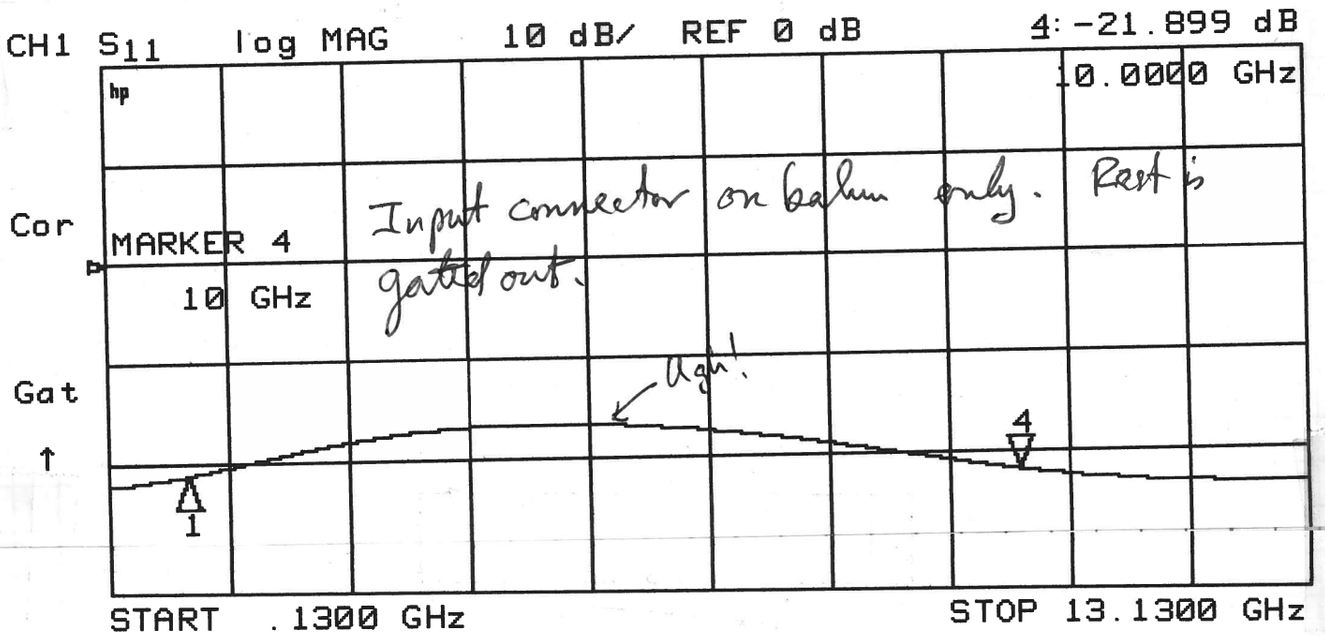
The double balun and the inverted double balun are overlayed here. The agreement is very close even outside the 1-10 GHz design range. The input and output connectors are slightly (?) mis matched (atn -15 dB), see p. 48.



double balun with output connector gated out.
Input connector is <u>not</u> gated out.



double balun with good load on output; input connector gated out. Also output connector gated out.



The effect of phase switching the 1st LO

Assume that the line to the first LO contains the phase switch of p.45 operated by a Walsh function (or some such).

I. For the cross correlator

The signal received at antenna i is:

$$x_i(t) = \int_0^\infty A_i(\nu) \cos[\omega t + \phi_i(\nu)]\, d\nu \quad , \quad \omega = 2\pi\nu$$

$A_i(\nu)$ and $\phi_i(\nu)$ are the random amplitude and phase.

After $x_i(t)$ is shifted to baseband and corrected for delay and fringe rotation, and the phase switch applied, it becomes:

$$x_i'(t) = \int_0^\infty A_i(\nu) \cos[\omega_{iif} t + \phi_i(\nu) - \psi_i]\, d\nu$$

$\psi_i(t)$ is the slowly varying phase switch $\therefore \psi_i(t) = \pi/2\, W_i(t)$
ω_{iif} is the baseband frequency variable.

The correlation is carried out for all antenna pairs (i,j).

$$\int_{-\infty}^{\infty} x_i'(t)\, x_j'(t+\tau)\, dt = C_{ji}(\tau)$$

After some algebra, with the substitutions, (see the BIMA control memo),

$$\int_{-\infty}^{\infty} x_i'(t)\, x_j'(t+\tau)\, dt = \cos[\psi_i - \psi_j] \int_0^\infty I_{ij} \cos[\omega_{iif}\tau + \Delta\phi_{ji}]\, d\nu$$

where $I_{ij} = A_i A_j$ and $\Delta\phi_{ji} = \phi_j(\nu) - \phi_i(\nu)$

$$\int_{-\infty}^{\infty} x_i'(t)\, x_j'(t+\tau)\, dt = \cos[\psi_i - \psi_j]\; C_{ji}(\tau) = C_{ji}(\tau)$$

$$C_{ji}(\tau) = \int_0^\infty I_{ji}(\nu) \cos[\omega_{iif}\tau + \Delta\phi_{ji}(\nu)]\, d\nu$$

The complex power spectrum is found from the Fourier Transform

$$I_{ji}(\nu)\, e^{-i\Delta\phi_{ji}(\nu)} = \int_{-\infty}^{\infty} C_{ji}(\tau)\, e^{-i2\pi\nu\tau}\, d\tau$$

$$\cos[\varphi_i - \varphi_j] = \cos\varphi_i \cos(-\varphi_j) - \sin(\varphi_i)\sin(-\varphi_j)$$

$$\cos\varphi_i = \cos[\tfrac{\pi}{2} W_i(t)] = 0$$
$$\sin\varphi_i = \sin[\tfrac{\pi}{2} W_i(t)] = W_i(t)$$

So $\cos[\varphi_i - \varphi_j] = 0 + W_i(t)\, W_j(t)$

$W_i(t)\, W_j(t) = W_{i+j}(t)$ another member of the set.

The output is just $W_{i+j}(t)\; C_{ji}(\tau) = O_{ji}(\tau)$

Synchronously detect this by doing $\frac{1}{T}\int_0^T W_{i+j}(t)\,\{\quad\}\,dt$

This can be done by having two registers and switching $O_{ji}(\tau)$ into them synchronously with $W_{i+j}(t)$. Then subtract the result after an even number of periods T of integration.

II. For the beamformer — the first LO is switched by $W_i(t)$

$V_j'(t)$ is the input to the first mixer

$V_i(t)\, W_i(t)$ is the output with $V_i(t)$ freq. shifted

Some interference $I_i(t)$ will get in, and some switched interference may get in $V_j(t)\, W_j(t)$

The digital output will be

$$O_i(t) = V_i(t)\, W_i(t) + I_i(t) + V_j(t)\, W_j(t)$$

Before summing for the beam but after phase and delay correction, multiply the output by $W_i(t)$

$$W_i(t)\, O_i(t) = V_i(t) + I_i(t)\, W_i(t) + V_j(t)\, W_{i+j}(t)$$

Sum the voltages for the beam and do a power detection with an integration over the Walsh period.

$$\frac{1}{T}\int_0^T \left(\sum_i O_i(t)\right)^2 dt$$

$$= \frac{1}{T}\int_0^T \left\{\sum_i \left[V_i(t) + I_i(t)V_i(t) + V_j W_{ij}\right]\right\}\left\{\sum_j \left[V_j + I_j W_j + V_j W_k\right]\right\} dt$$

The cross terms where powers vary on the time scales of the Walsh functions integrate out, leaving

$$\sum_i \overline{V_i^2} + \sum_i \overline{I_i^2} + \sum_i \overline{V_k^2} + \text{noise and rapidly fluctuating interference.}$$

① The second and third terms might be eliminated by subtracting an off beam.

② The Walsh function switching could be rapid as long as the intervals are still longer than, say, 10 μsec, so that delays are unimportant.

This is not quite right. All the signal voltages are the same

So $\sum_i^N O_i(t) = NV(t) + \sum\left[I_i(t)W_i(t) + V_j(t)W_{ij}(t)\right]$

$$\frac{1}{T}\int_0^T \left\{NV(t) + \sum I_i W_i(t) + V_j(t)W_{ij}(t)\right\}^2 dt$$

Output power $= \overline{(NV(t))^2} + \sum \overline{I_i^2} + \sum_i \overline{V_j^2}$

MOU Board Sept. 17, 2001 at UCB

JD photonic systems has sent its report - 3 options, no vicsels - ~$1500/channel (each polarization)

Matt will use a gear drive for azimuth

all PDR's are done
a new study of shipping the reflectors whole is underway

John's personal plan puts us 3 mos. behind -

Leo: Matt Dexter will likely come. He could help -

Leo on permitting: a) Bidwell lease - now at OP
OP lawyers and Bidwell seem in agreement -
(b) USFS: must work through planning office - hope for completion of permit ~ 3-4 months
(c) public meeting went well, 2 local papers - high enthusiasm.
(d) Visitor Center - make it larger if poss. (from Intermountain News Editor)

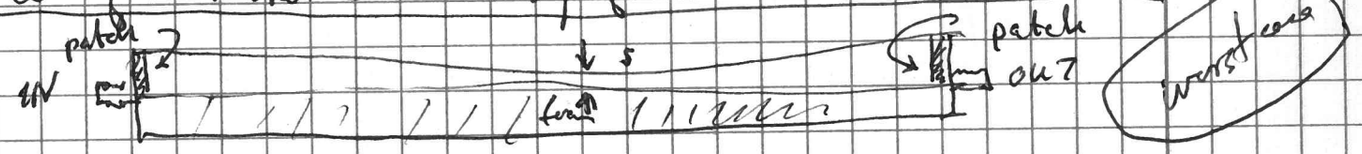
Correlator prop. not funded - but the development is going ahead -

Mike on the NSS good news on the development -
"blue Wave Board" bought 4 units : 2 pol, 2 sites
Motorolla bought the company - will it continue with board?
software system looks very good
October (2002) deployment expected - 25 MHz

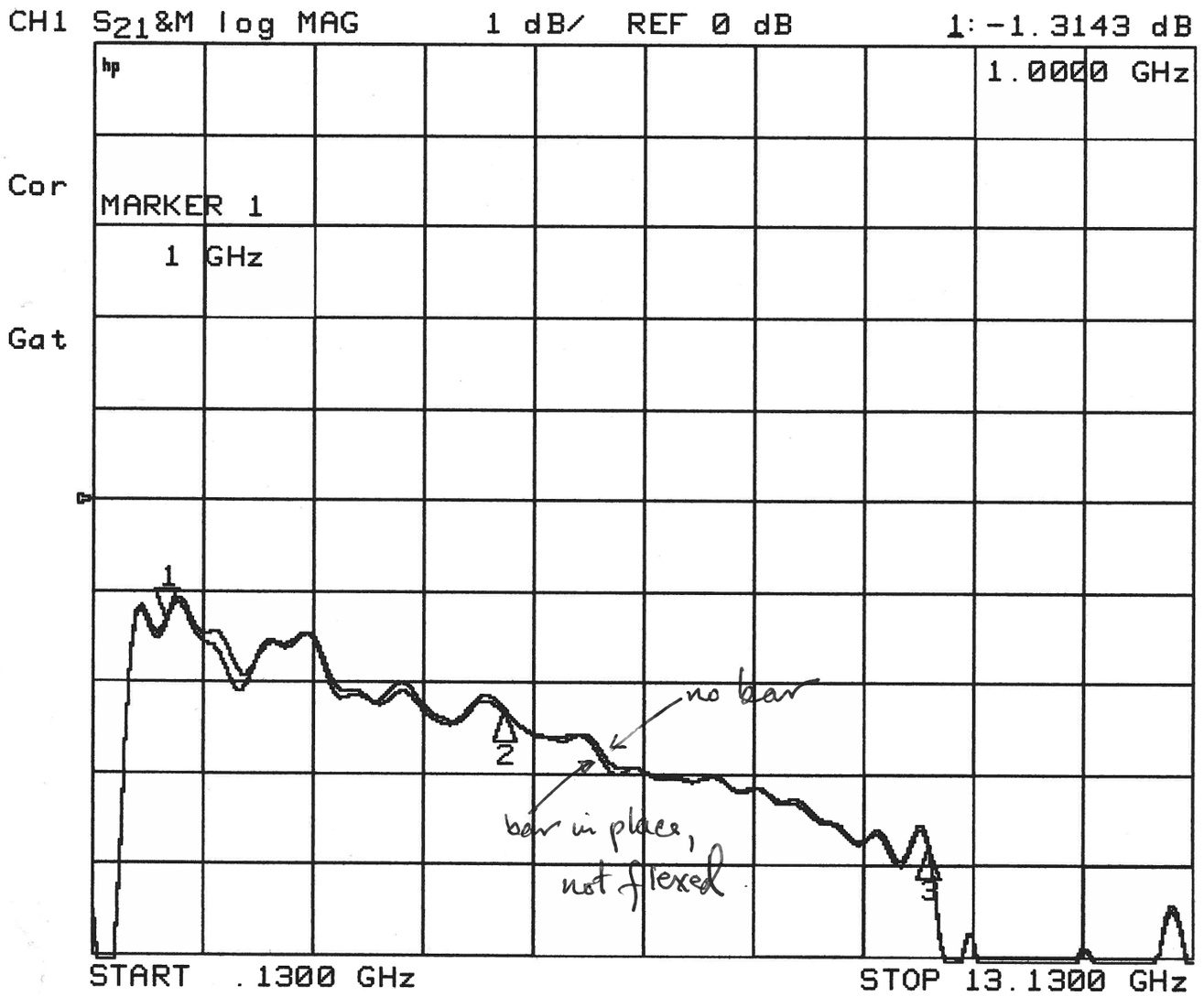
Replication 1.1 M$ for 2 pol., 2 sites, 25 MHz

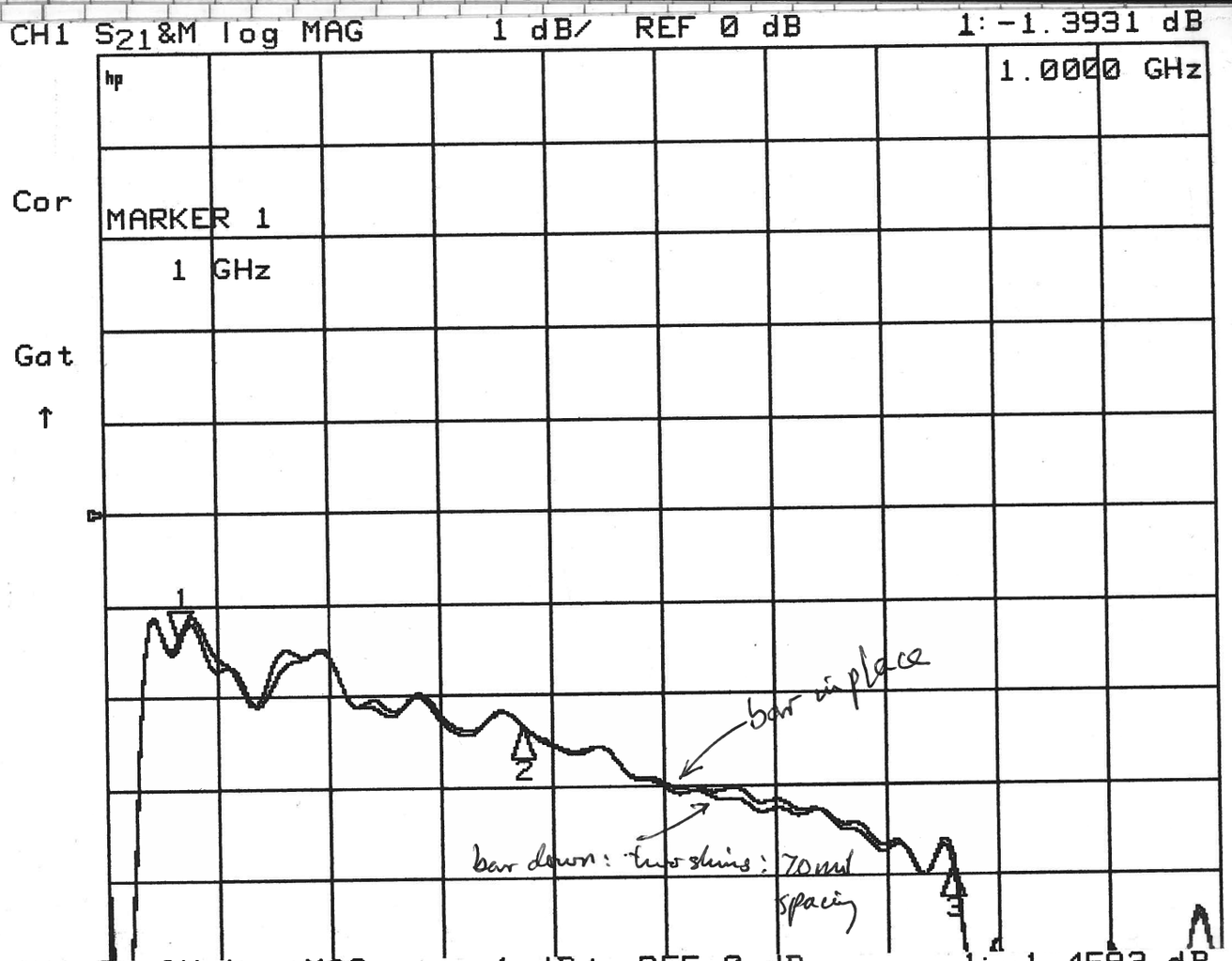
Tom: funding officer

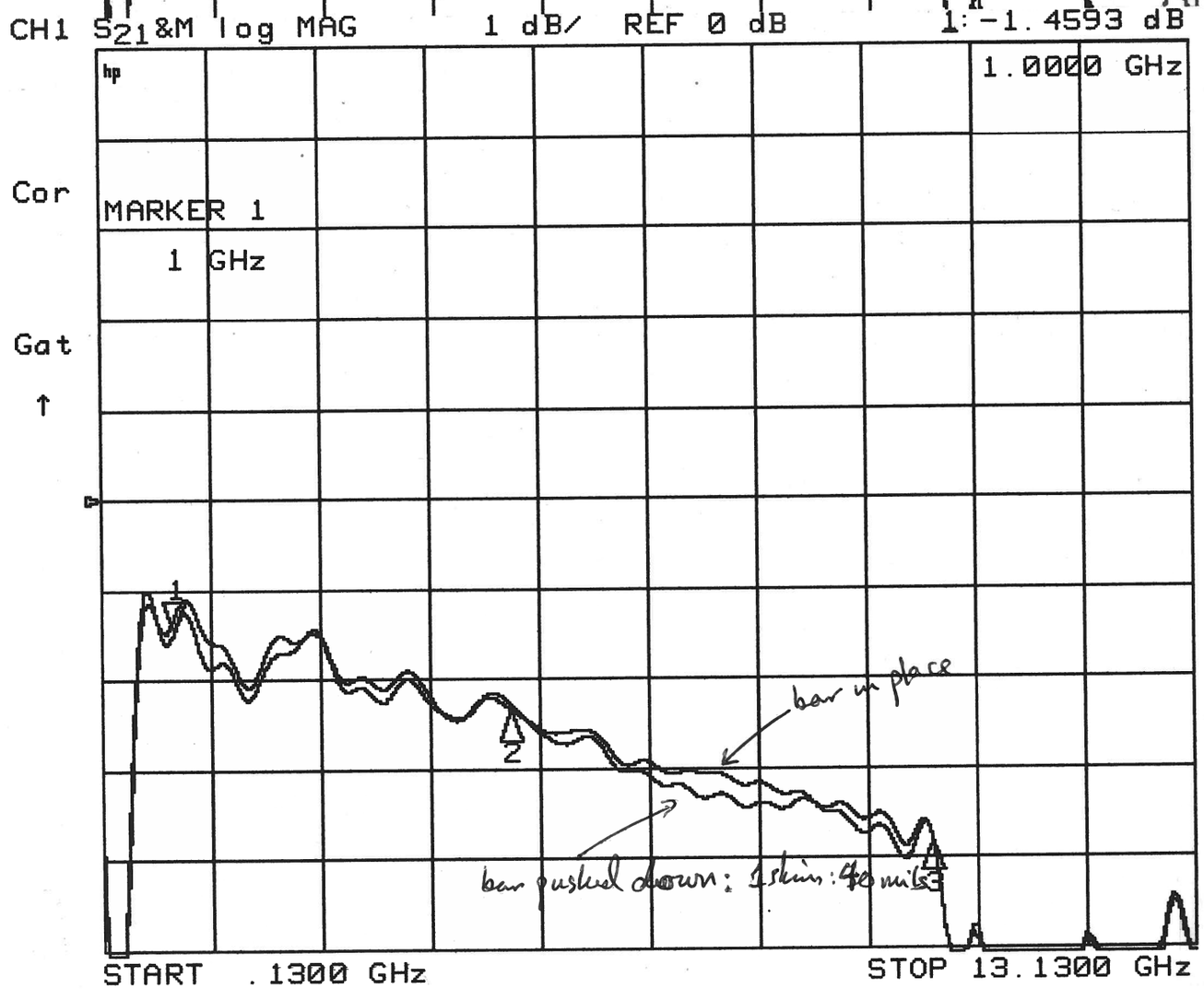
Effect of a metal bar on top of the double test balun.

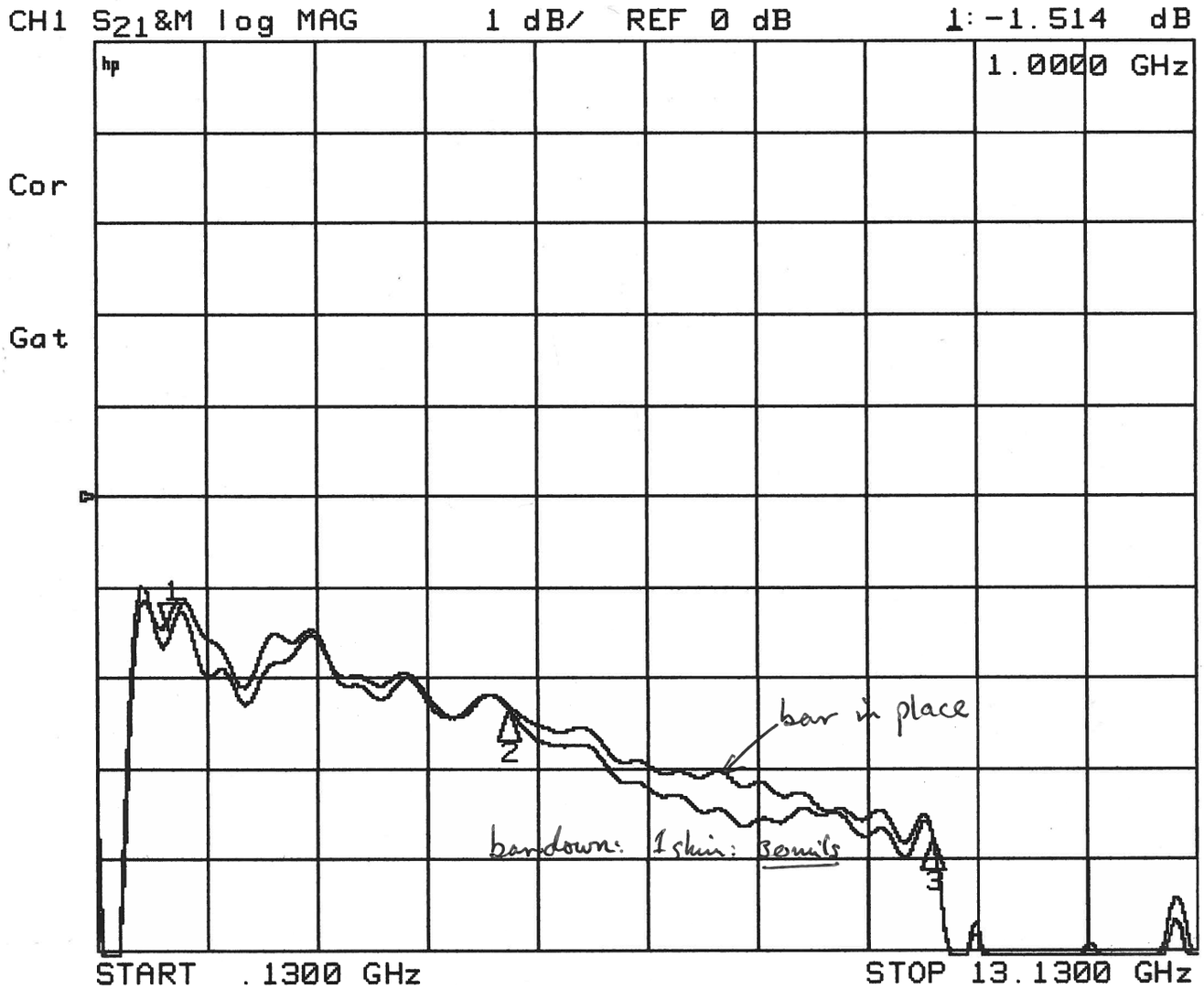


S_{21} is measured with the bar in place or off and with it pushed down to different spacings from the trace.
There are absorbing patches in place on the input and output flanges to reduce the effect of a coupled mode on the top of the trace.

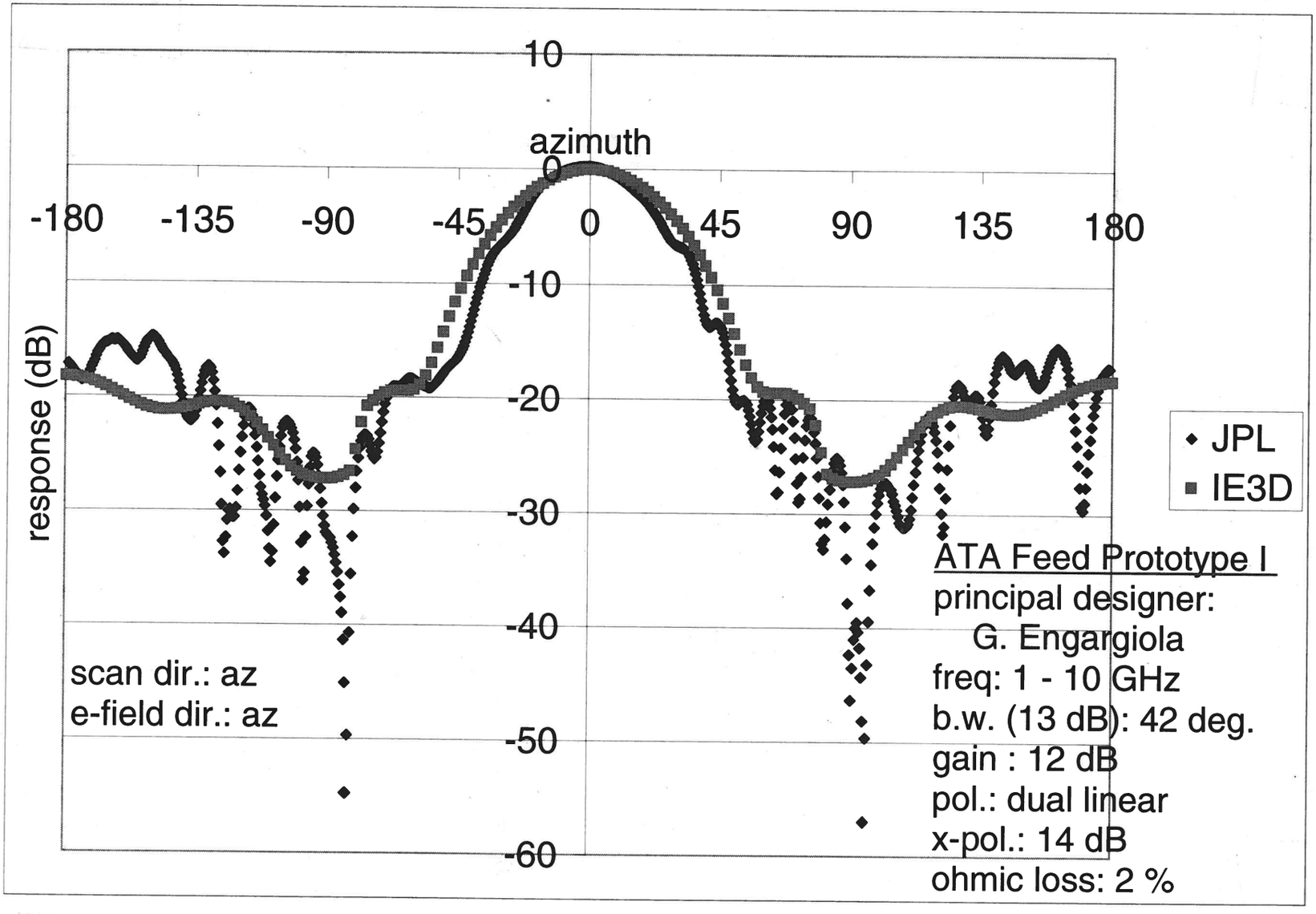



CH1 S21&M log MAG 1 dB/ REF 0 dB 1: -1.3931 dB
hp
1.0000 GHz
Cor
MARKER 1
1 GHz
Gat
1
2
3
bar in place
bar down: two skins: 70 mil spacing


CH1 S21&M log MAG 1 dB/ REF 0 dB 1: -1.4593 dB
hp
1.0000 GHz
Cor
MARKER 1
1 GHz
Gat
1
2
3
bar in place
bar pushed down: 1 skin: 40 mils
START .1300 GHz
STOP 13.1300 GHz


CH1 S21&M log MAG 1 dB/ REF 0 dB 1: -1.514 dB
hp
1.0000 GHz
Cor
Gat
1
2
3
bar in place
bar down: 1 skin: 30mils
START .1300 GHz
STOP 13.1300 GHz

Feed Tests at JPL.



1996 Pub For: Jack Welch

data on single-crystalline Y-123 due to the high T_c of 106 K [101]. The overall trend of $R_s(f)$ of HTS resembles that of the conventional superconductors Nb and Nb_3Sn. Figure 10 presents an experimental extension of Fig. 9 to submillimeter wave frequencies [124].

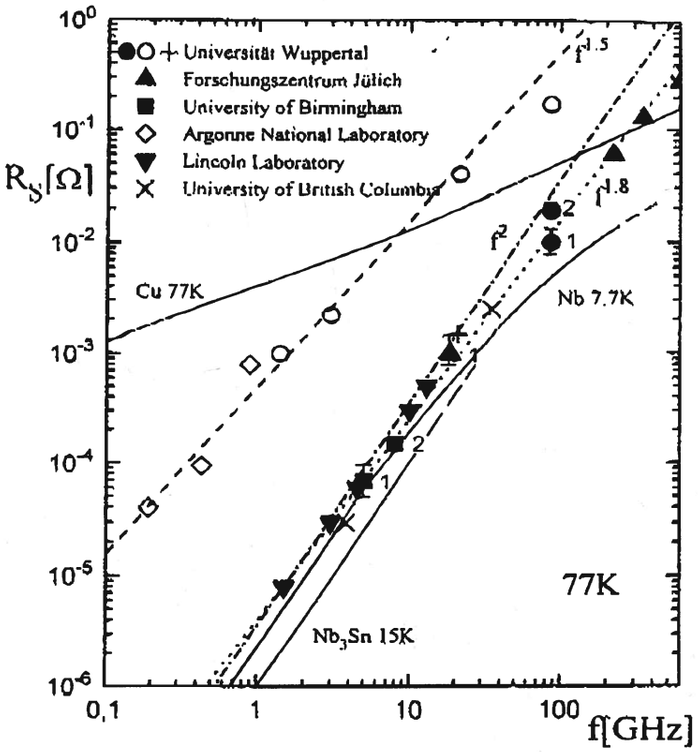


Fig. 9: Frequency dependence of R_s(77 K) of high-quality Y-123 single crystals (crosses) and epitaxial films (filled symbols) compared to polycrystalline samples (open symbols) and Cu. Data on Nb and Nb_3Sn are given at corresponding reduced temperatures [39].

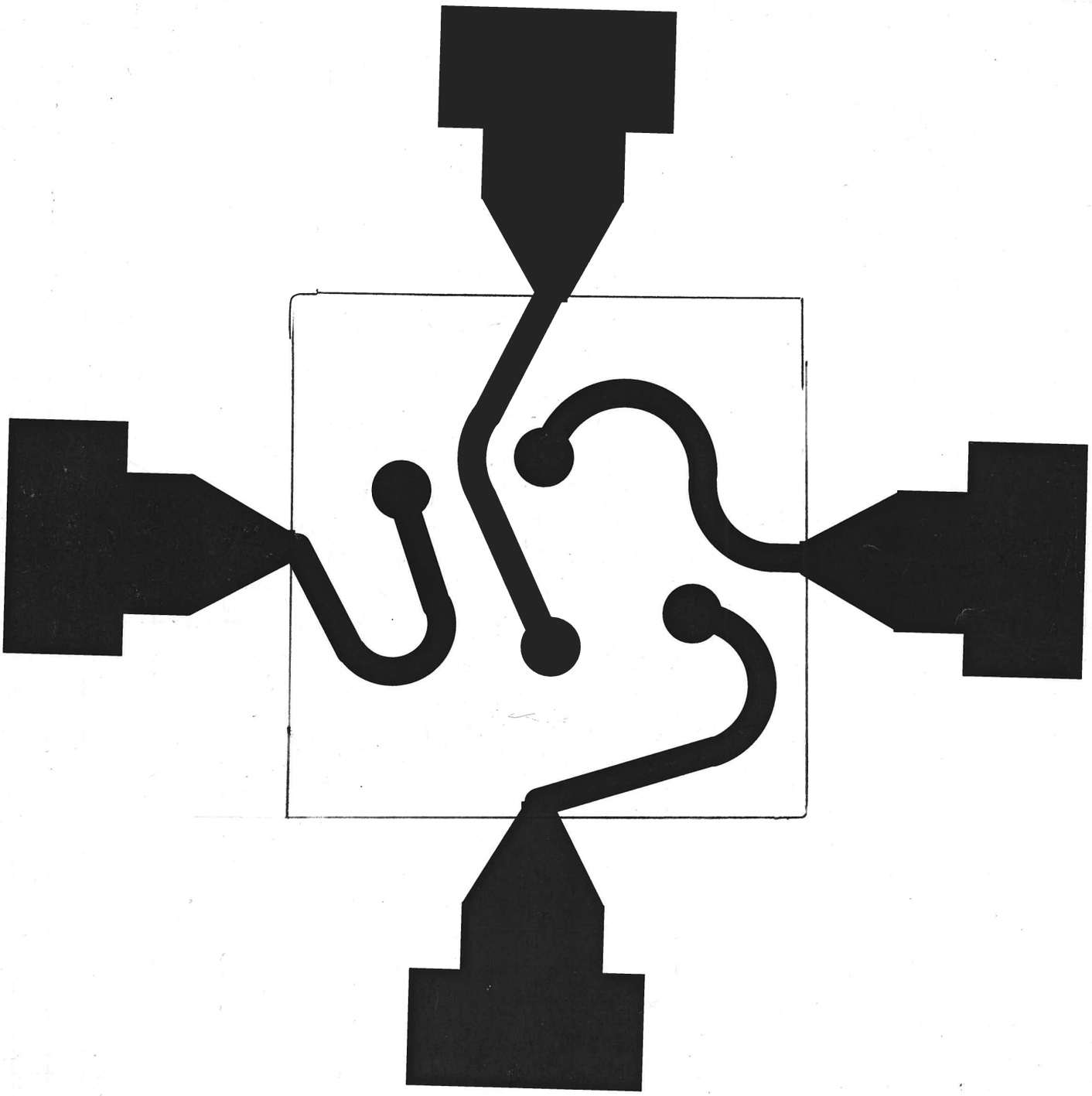
Care has to be taken in scaling the surface resistance of HTS generally quadratic in frequency. Deviations from the power-law $R_s \propto f^b$ with b=2 result from a finite value of the conductivity ratio $y=\sigma_1/\sigma_2$ (cf. Eqs. (3) and (4)) [125]:

$$b = 1 + \frac{y^2}{2(1+y^2)} \cdot \left\{ \frac{1}{1 - 1/\sqrt{1+y^2}} - 2 \right\} \tag{11}$$

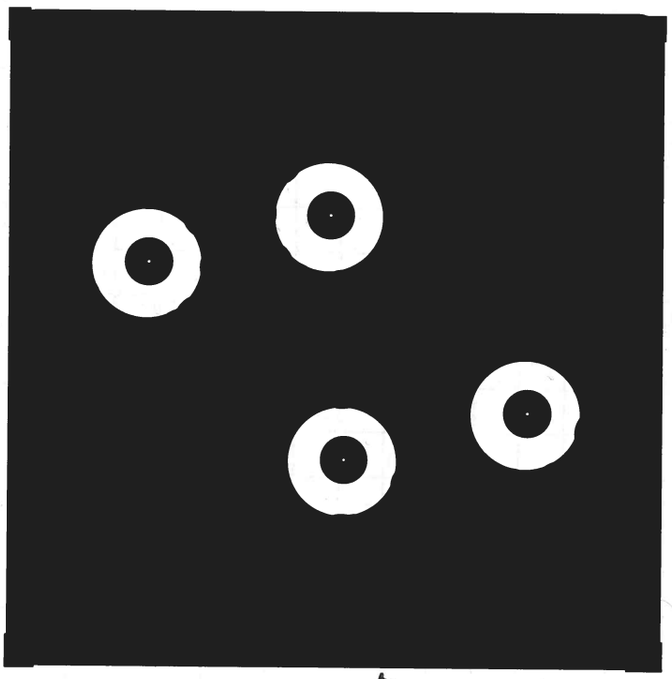
In the limit y>>1, Eq. (11) yields b=1/2 in accordance with the normal skin effect. In the opposite limit, b approaches 2 like $b=2-y^2$. From the definition of y, the dependence of b on the frequency f and the normal-state resistivity ρ can thus be approximated as $b=2-\text{const.}\cdot(f/\rho)^2$. Using b=1.5 at f=87 GHz, T=77 K and ρ=50 μΩcm (resulting in const.=0.165 {μΩcm/GHz}^{2}) [39], the data in Fig. 9 are quantitatively reproduced. In more general terms, Eq. (11) is to be supplemented, according to Sec. 2.1.1, by:

$$y = \frac{x_N(\omega,t)\cdot\omega\tau(t)}{1 - x_N(\omega,t) + \{\omega\tau(t)\}^2} \tag{12}$$

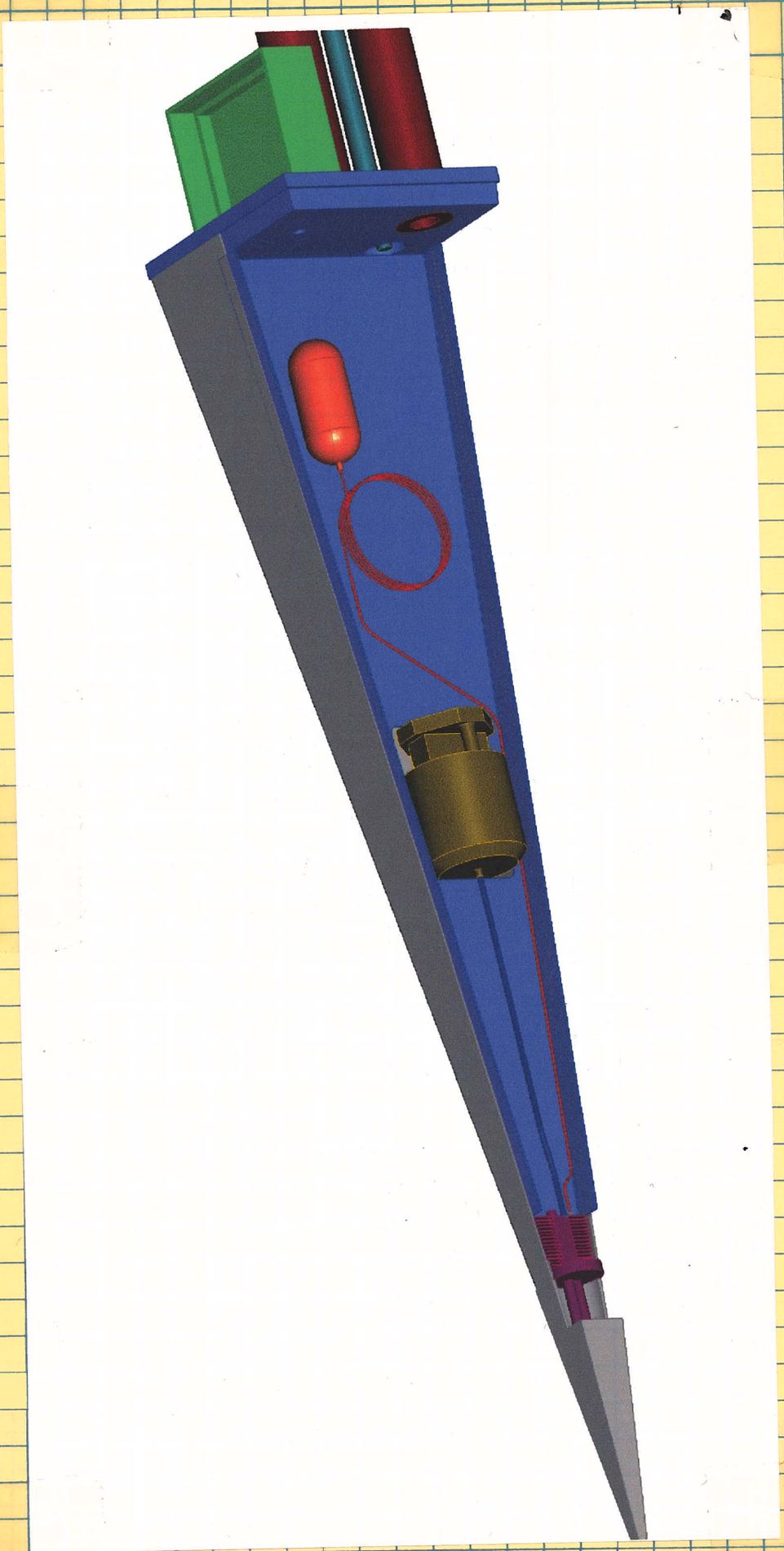
Equations (11) and (12) summarize the phenomenology of the frequency dependence of R_s. The absolute level and the scaling behavior of b are affected by the presence of



see I p. 130



→ 24 mils ←



11/12/2001 meeting

Some Issues

Balun status
Tsys w and w/o cooled LNA — 29 (cooled) K — 28 uncooled
LNA gain stability
LNA production status
Antenna design status.

Sandy's JPL array development -

I. a) 7-43 GHz upgrade of ATA feed
kildal and H. Cortez to study } comparisons
b) 8/32 GHz rcvrs
c) do a JPL/CIT interferometer pair

II. a) 12m 60 GHz antenna
b) 100 x 12m antennas
c) 100 x 8/32 GHz RCVRs

III. Build New Mexico Plant
EVLA array 100x12m -

Tsys estimates — 66 amp. (42K min) — 42K cooled (12 min)

Special Rm ~ 125K

Chips have leakage current problems

Sandy is prepared ~~for the LNA~~ to do the cooled LNA mount -

JL: ① how to mount feed arms on outside of pyramid
② what's inside
③ He has a design for transmission line input
Prefers that -

27db of gain?

Do transmission line calculations — use a grounded septum?

Put a 40db LNA into the dewar
need a bias tee to balance the gains

Dewar – pressure sensors (?)

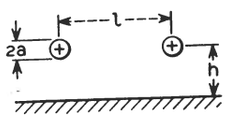
Use the transmission line input lines, not the balun –

But there are 2 noise sources. Is the noise doubled?
→ Sandy will check this.

→ Jack will check out the balun for the Dewar mounting –
Jack and Greg will check out the cooled balun. conical bore.
Use the Krytar 180° hybrid < Feb 1.

Gain Stability $\frac{\delta T}{T} = \frac{\Delta G}{G} = 40\,\text{ppm}\sqrt{\text{Hz}}$

A few transmission lines from Shelkunoff & Friis

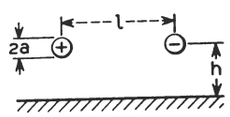


$$K = 30 \log \frac{2h}{a}\left[1 + \left(\frac{2h}{l}\right)^2\right]^{1/2}$$

unbalanced mode.

$$= 69 \log_{10} \frac{2h}{a}\left[1 + \left(\frac{2h}{l}\right)^2\right]^{1/2}.$$

($2a$ is small.)



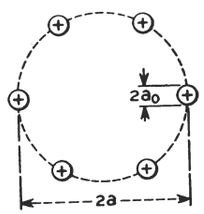
$$K = 120 \log \frac{l}{a}\left[1 + \left(\frac{l}{2h}\right)^2\right]^{-1/2}$$

balanced mode

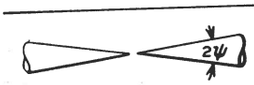
$$= 276 \log_{10} \frac{l}{a}\left[1 + \left(\frac{l}{2h}\right)^2\right]^{-1/2}$$

$$= 120 \log \frac{2h}{a}\left[1 + \left(\frac{2h}{l}\right)^2\right]^{-1/2}$$

$$= 276 \log_{10} \frac{2h}{a}\left[1 + \left(\frac{2h}{l}\right)^2\right]^{-1/2}.$$



The radius of a conductor equivalent to a cage of n wires, $a_{\text{eff}} = a\left(\frac{na_0}{a}\right)^{1/n}$.

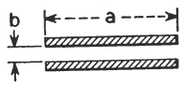


$$K = 120 \log \cot \tfrac{1}{2}\psi = 276 \log_{10} \cot \tfrac{1}{2}\psi.$$

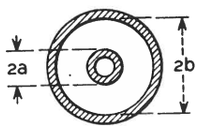
594

APPENDIX I

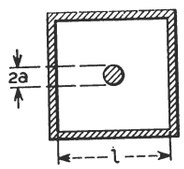
Characteristic Impedances of Transmission Lines



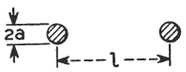
If $b \ll a$, $K = 120\pi \dfrac{b}{a}$.



$K = 60 \log \dfrac{b}{a} = 138 \log_{10} \dfrac{b}{a}$. $= 138 \log_{10}\left(\dfrac{2b}{2a}\right)$



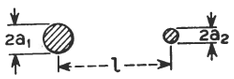
$K = 60 \log \dfrac{1.078l}{2a} = 138 \log_{10} \dfrac{1.078l}{2a}$.



$$K = 120 \cosh^{-1} \frac{l}{2a}$$

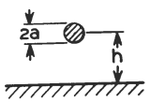
$$= 120 \log \left[\frac{l}{2a} + \sqrt{\left(\frac{l}{2a}\right)^2 - 1} \right].$$

If $2a \ll l$, $K = 120 \log \dfrac{l}{a} = 276 \log_{10} \dfrac{l}{a}$.



If $2a_1 \ll l$, and $2a_2 \ll l$,

$K = 120 \log \dfrac{l}{\sqrt{a_1 a_2}} = 276 \log_{10} \dfrac{l}{\sqrt{a_1 a_2}}$.



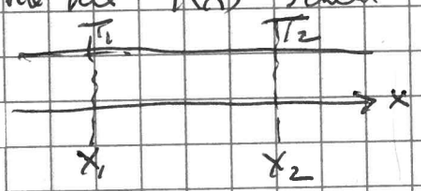
$K = 60 \cosh^{-1} \dfrac{h}{a}$.

If $a \ll h$, $K = 60 \log \dfrac{2h}{a} = 138 \log_{10} \dfrac{2h}{a}$.

Heat flow in one dimension for a temperature power law heat conductivity

① Scale the temperature power law to the temperature at the top of the range: $k = k_0 \left(T/T_u \right)^{\gamma}$

② Write the $T(x)$ scaled to the upper position limit, x_2



$$T(x) = T_0 \left(\frac{x - x_0}{x_2} \right)^{\alpha}$$

Find α, x_0, and T_0

k is the thermal conductivity

The equation for steady heat flow is $\frac{d}{dx}\left(k \frac{dT}{dx}\right) = 0$

$$\frac{d}{dx}\left[k_0 \left(\frac{T}{T_u}\right)^{\gamma} \frac{dT}{dx} \right] = 0 \implies k_0 \left(\frac{T}{T_u}\right)^{\gamma} \frac{dT}{dx} = \text{const.}$$

$$\frac{d}{dx}\left\{ T_0 \left(\frac{x - x_0}{x_2} \right)^{\alpha} \right\} = \alpha T_0 \frac{(x - x_0)^{\alpha - 1}}{x_2^{\alpha}}$$

$$k_0 \left[\frac{T_0}{T_u} \left(\frac{x - x_0}{x_2} \right)^{\alpha} \right]^{\gamma} \alpha T_0 \frac{(x - x_0)^{\alpha - 1}}{x_2^{\alpha}} = \text{const.}$$

$$k_0 \left(\frac{T_0}{T_u} \right)^{\gamma} \frac{(x - x_0)^{\alpha\gamma}}{x_2^{\alpha\gamma}} \alpha T_0 \frac{(x - x_0)^{\alpha - 1}}{x_2^{\alpha}} = \text{const.}$$

$(x - x_0)^{\alpha\gamma + \alpha - 1}$ is the variable factor, requiring $\alpha\gamma + \alpha - 1 = 0$

$$\boxed{\alpha = \frac{1}{1 + \gamma}} \quad \text{or} \quad \frac{1}{\alpha} = 1 + \gamma$$

Boundary Conditions $\quad T_1 = T_0 \left(\frac{x_1 - x_0}{x_2} \right)^{\alpha}$; $T_2 = T_0 \left(\frac{x_2 - x_0}{x_2} \right)^{\alpha}$

a) $$\boxed{T_0 = T_1 \left(\frac{x_2}{x_1 - x_0} \right)^{\alpha}}$$

b) ratio $$\frac{T_2}{T_1} = \frac{T_0 \left(\frac{x_2 - x_0}{x_2} \right)^{\alpha}}{T_0 \left(\frac{x_1 - x_0}{x_2} \right)^{\alpha}} = \frac{(x_2 - x_0)^{\alpha}}{(x_1 - x_0)^{\alpha}}$$

$$\left(\frac{T_2}{T_1} \right)^{\frac{1}{\alpha}} = \left(\frac{x_2 - x_0}{x_1 - x_0} \right) = \left(\frac{T_2}{T_1} \right)^{1 + \gamma}$$

$$x_2 - x_0 = (x_1 - x_0)\left(\frac{T_2}{T_1}\right)^{1+\gamma} \qquad x_2 - x_1\left(\frac{T_2}{T_1}\right)^{1+\gamma} = x_0\left(1 - \left(\frac{T_2}{T_1}\right)^{1+\gamma}\right)$$

$$\boxed{x_0 = \frac{x_1\left(T_2/T_1\right)^{1+\gamma} - x_2}{\left(T_2/T_1\right)^{1+\gamma} - 1}}$$

Heat Flow $\qquad H = A K \dfrac{dT}{dx} \qquad A$ = cross section

$$H = A k_0\left(\frac{T}{T_u}\right)^{\gamma} \frac{d}{dx}\left[T_0\left(\frac{x - x_0}{x_2}\right)^{\alpha}\right] = A k_0\left(\frac{T}{T_u}\right)^{\gamma} \alpha T_0 \frac{(x - x_0)^{\alpha-1}}{x_2^{\alpha}}$$

$$= A k_0\left(\frac{T_0}{T_u}\right)^{\gamma}\left(\frac{x - x_0}{x_2}\right)^{\gamma\alpha} \alpha T_0 \frac{(x - x_0)^{\alpha-1}}{x_2^{\alpha}} = A k_0\left(\frac{T_0}{T_u}\right)^{\gamma} \frac{\alpha T_0}{x_2^{\alpha\gamma}\, x_2^{\alpha}} (x - x_0)^{\alpha-1+\alpha\gamma}$$

But $\alpha - 1 + \gamma\alpha = 0$ $\qquad$ or $\alpha + \alpha\gamma = 1$

$$\boxed{H = A k_0\left(\frac{T_0}{T_u}\right)^{\gamma} \frac{\alpha T_0}{x_2}}$$

dimensions: $m^2 \dfrac{\text{watts}}{m\,°K} \dfrac{°K}{m}$ = watts

Copper

T (°K)	K (watts/m°K)	$401(T/300)^{-.1}$	$401(T/300)^{-.25}$	$401(T/300)^{-.15}$	$482(T/300)^{-.5}$	( )
300	401	401	401	401		
250	406	408		412		
200	413	417	443	426		
150	429	430	476	445		
100	482	447	527	473	482	482
90	508				508	513
80	557				539	551
70	647					
60	829					
50	1250					

Best fits $\quad 300 > T > 100 \qquad 401(T/300)^{-.15}$

$100 > T > 80 \qquad 482(T/100)^{-.6}$

Max error is ~3%.

Stainless Steel 304 and 347

$T(°K)$	(304) K(watts/cm°K	$.152(T/300)^{.46}$	(347) K
75	.080	.080	.076
100	.095	.092	.091
150	.115	.111	.111
200	.130	.126	.126
250	.142	.140	.138
273	.147	.146	
300	.152	.152	.148

For the input transmission line in the dewar. (Copper)
Divide it up into 2 regions: $0 < x \le 2''$ for $100 < T < 300$
$2 \le x < 8''$ for $80 < T < 100$

a) First part $300 - 100$ $\quad x_1 = 0,\ x_2 = 2'' \quad \gamma = -.15,\ x_1 = 0,\ x_2 = 2$
$T_2 = 300,\ T_1 = 100;\ 1+\gamma = .85$

$$T_0 = T_1\left(\frac{x_2}{x_1 - x_0}\right)^{\alpha},\quad \alpha = \frac{1}{1+\gamma} = \underline{1.177} \qquad 3^{.85} = 2.5442$$

$$x_0 = \frac{x_1\left(T_2/T_1\right)^{1+\gamma} - x_2}{\left(T_2/T_1\right)^{1+\gamma} - 1} = \frac{-2}{2.5442 - 1} = \underline{-1.2952}$$

$$T_0 = 100\left(\frac{2}{+1.2952}\right)^{1.177} = \underline{166.76}$$

$$\longrightarrow T_a(x) = 166.76\left(\frac{x + 1.2952}{2}\right)^{1.177} = 100.00 \text{ at } x = 0;\ = 300.14 \text{ at } x = 2 \checkmark$$

b) Second part $\quad 100 - 80\ K,\ x_1 = 2,\ x_2 = 8,\ \gamma = -.60,\ T_1 = 80,\ T_2 = 100$
$\alpha = \frac{1}{1+\gamma} = 2.500,\ 1+\gamma = 0.40,\ (T_2/T_1)^{.4} = 1.0934$

$$\underline{x_0 = -62.240}$$

$$T_0 = 80\left(\frac{8}{2 + 62.240}\right)^{2.50} = \underline{.4378}$$

$$\longrightarrow T_b(x) = .4378\left(\frac{x + 62.240}{8}\right)^{2.5} = 80.00 \text{ at } x = 2,\ = 100.00 \text{ at } x = 8 \checkmark$$

Heat Flow $H = A K_0 \left(\frac{T_0}{T_u}\right)^{\gamma} \frac{T_0 \alpha}{X_2}$

Case a $X_2 = 2\text{cm} = .0508\text{m}$, $T_0 = 166.76\,°K$, $\alpha = 1.177$, $T_u = 300$
$K_0 = 401$ watts/m°K, $\gamma = -.15$

$$H_a = A_a(m^2)\, 401 \left(\frac{166.76}{300}\right)^{-.15} \frac{(166.76)\,1.177}{.0508} = A_a(m^2)\, 1.69\times10^{6} \text{ watts}$$

Case b $X_2 = 8\text{in} = .2032\text{m}$, $T_0 = .4378$, $\alpha = 2.5$, $\gamma = -.60$,
$T_u = 100$, $K_0 = 482$

$$H_b = A_a(m^2)\, 482 \left(\frac{.4378}{100}\right)^{-.60} \frac{(.4378)\,2.50}{.2032} = A_b(m^2)\, 6.75\times10^{4} \text{ watts}$$

$H_b << H_b$ for the same cross section.
For the same heat flow, $H_b = H_a$, $A_b/A_a = 25.03$
(and same material)

John Lugton's plan for the cold run (second part) is 23cm (9.0 inches) not 6". Work out the Temperature distribution for that length here.

$100 - 50K$ $X_1 = 2''$, $X_2 = 11'' = .2794$, $\gamma = -.60$ $T_1 = 80$, $T_2 = 100$
$\alpha = \frac{1}{1+\gamma} = 2.5$, $1+\gamma = 0.40$ $(T_2/T_1)^{.4} = 1.0934$

$$X_0 = \frac{2(1.0934) - 11}{1.0934 - 1} = -94.36 = \frac{X_1 (T_2/T_1)^{1+\gamma} - X_2}{(T_2/T_1)^{1+\gamma} - 1}$$

$$T(x) = T_0 \left(\frac{X - X_0}{X_2}\right)^{\alpha} \quad \text{so} \quad T_1(X_1) = T_0 \left(\frac{X_1 - X_0}{X_2}\right)^{\alpha} \quad \text{and.}$$

$$T_0 = T_1 \left(\frac{X_2}{X_1 - X_0}\right)^{\alpha} = 80 \left(\frac{11}{2 + 94.36}\right)^{2.5} = 0.3522$$

$$T(X) = .3522 \left(\frac{X + 94.36}{11}\right)^{2.5} = 79.993 \text{ @ } X = 2'', \quad = 99.999 \text{ @ } X = 11''$$

Now $H_b = A(m^2)\, 482 \left(\frac{.3522}{100}\right)^{-.60} \frac{(.3522)(2.5)}{.2794} = 4.50\times10^{4} A$, so $\frac{A_b}{A_a} = \boxed{16.7}$

Work out the noise contribution for these two sections.

From I. p 118, 119 $\quad R_s = .0020\sqrt{f\rho}$ cgs.

I. First for the long cold section $\quad T(x) = .3522\left(\frac{x+94.36}{11}\right)^{2.5}$ inches

Convert to cm: $T(x) = .3522\left[\frac{x+239.67}{23}\right]^{2.5}$ °K $\quad x$ in cm $\quad 5 \le x \le 23$ cm

Assume a constant wire diameter $\quad OD = .020\ (\text{in}) = .0508\text{ cm} = D$

For $Z_0 = 220\,\Omega$, $S/D = 3.2$

$$R = \frac{2R_s}{\pi D}\frac{S/D}{\sqrt{(S/D)^2-1}} = \frac{R_s(2)(3.2)}{\pi(.0508)\sqrt{(3.2)^2-1}} = \cancel{1.0527R_s}\ 13.1926\,R_s$$

$$\alpha_p = R/Z_0 = \cancel{1.0527R_s/220} = 13.192\,R_s/220 = .060\,R_s$$

With $\rho(T) = .75\left(\frac{T-20}{132}\right)^{1.35}10^{-6}\ \Omega\text{ cm} = .75(\ )^{1.35}\times10^{-8}$ ohm-in $\{10\,GHz = 10^{10}\,Hz\}$

$$R_s = .0020\sqrt{f\rho} = .0020\left[.75\left(\frac{T-20}{132}\right)^{1.35}10^{-8}\right]^{1/2}\sqrt{f}$$

$$\alpha_p = .060R_s = (.060)(.0020)\left[f_{10}10^{10}(.75)\left(\frac{T-20}{132}\right)^{1.35}\times10^{-8}\right]^{1/2}$$

$$\alpha_p(T) = .0010\left(\frac{T-20}{132}\right)^{.675}\sqrt{f_{10}}\ \text{cm}^{-1} \rightarrow .0006\sqrt{f_{10}}\ \text{at } T = 80K$$
$$.0004 \text{ at } 5\,GHz$$

Emission from the line with the gradient here, $T(x) = .3522[\ ]^{2.5}$

$$T_B = \int_2^{11} T(x)\,\alpha_p[T(x)]\,dx$$

$$= \int_5^{23}\left\{.3522\left[\frac{x+239.67}{23}\right]^{2.5}\right\}\left\{.001\sqrt{f_{10}}\left[\frac{.3522\left(\frac{x+239.67}{23}\right)^{2.5}-20}{132}\right]^{.675}\right\}dx$$

$$\cancel{= 1.0616\sqrt{f_{10}}}$$

$2.414\sqrt{f_{10}}$ °K (for 10 mil line)

For 20 mil copper wire at $T = 273K$, $\alpha_0(273) = .0016\sqrt{f_{10}}\ \text{cm}^{-1}$

II For the region with the larger temperature gradient

$$T_2(x) = 166.76\left(\frac{x+3.289}{5}\right)^{1.177}$$

again $\alpha_p = .0010\sqrt{f_{10}}\left(\frac{T-20}{132}\right)^{.675}$ cm^{-1}

$$T_{B_2} = \int_0^5 T_2(x)\,\alpha_p(T_2(x))\,dx = \int_0^5 T_2(x)\left\{.0010\sqrt{f_{10}}\left(\frac{T_2(x)-20}{132}\right)^{.675}\right\}dx$$

$= 1.291\sqrt{f_{10}}$ With 10 mils OD instead of 20, get 2.6K

New noise summary at 10 GHz

		Notes
Feed (3% @ 10 GHz) @ 300 K	9.0	probably const.
tep circuit board 1% at 200 K	3.0	
transition line	1.3 (2.6)	$\times\sqrt{f_{10}}$
cold line	2.4 (4.8 for 10 mils)	$\times\sqrt{f_{10}}$
spillover ~~tipsoldercon.~~ ~1%	3.0	
	18.7 (22.4)	

NB. 1% geometric spillover, ~1% diffraction on average

2.74 T_{3B}	2.74
atmos at z ~ 50 (~~[illegible]~~) 10 GHz	~~8~~.00
	~~[illegible]~~
	9.74

Dick's atmos program: 1.5K @ 1 GHz for Hat Creek (10 mm H_2O)
2.4K @ 10 GHz " "

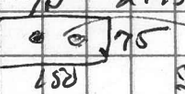
total is 30K.

Plus LNA @ ~ ~~15K~~ 12K ~~32K~~ 42 at 10 GHz
34K ~~32K~~ at 4 GHz

$N_0 B_0$ Heat flow in the input lines: For the input 2" use 10 mil kovar w/ .0001" Cu plating
Then $A_a = 3.05\times10^{-5}$ cm^2 for one pin and 1.22×10^{-4} cm^2 for 4 pins
For $A_b/A_a = 16.7$, solid copper pins of 10 mil diameter are just right.

The heat flow is $H = A(m^2)\times1.69\times10^6$ watts $= 1.2\times10^{-8}\times1.69\times10^6 = .021$ watts

Characteristic impedance for 200 Ω requires $S/D = 2.73$ in air
With D = 10 mils, S = 27.3 mils.



get 40 mils

D = 20 has S = 54 mils
probably ok.

For $\epsilon = 2$, $S/D = 14$ and S = 140 mils.

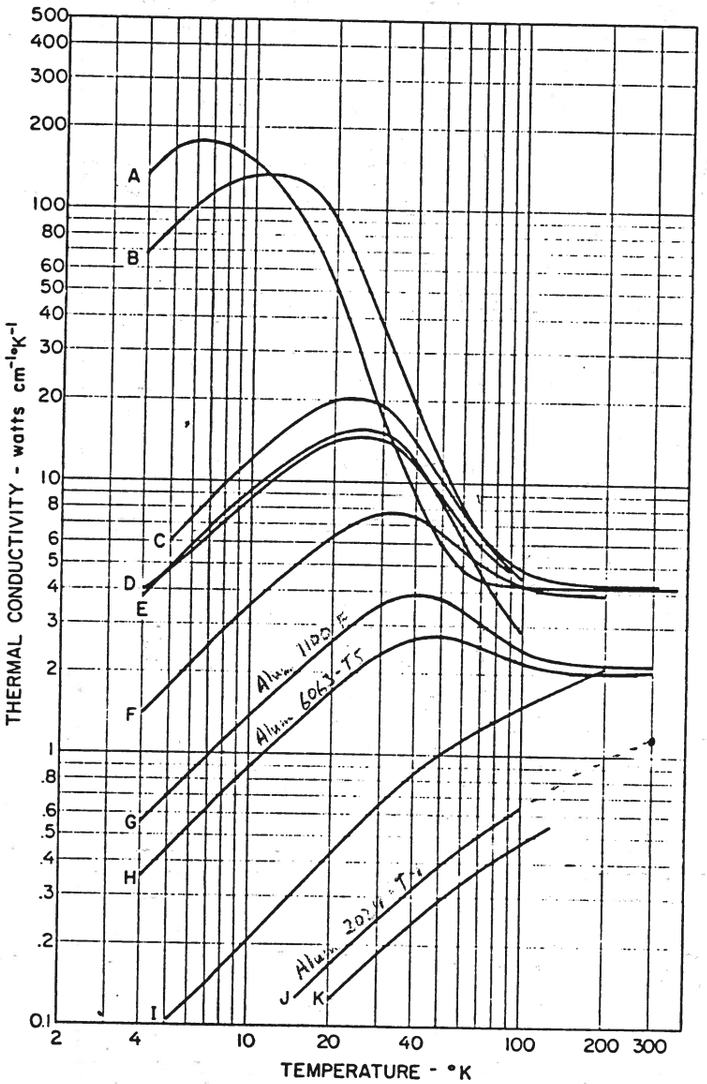


FIGURE 10.7. Low-temperature thermal conductivities of some metals having relatively high conductivities. A, silver 99.999% pure [39]; B, high purity copper [42]; C, coalesced copper [42]; D, copper, electrolytic tough pitch [42]; E, aluminum single crystal [43]; F, free-machining tellurium copper [42]; G, aluminum, 1100 F [43]; H, aluminum, 6063-T5 [43]; I, copper, phosphorus deoxidized [42]; J, aluminum, 2024-T4 [43]; K, free-machining leaded brass [42].

LOW-TE

THERMAL CONDUCTIVITY - milliwatts cm⁻¹ °K⁻¹

FIGURE 10.8.
low conducti
lium copper
Inconel
fluoroethylen

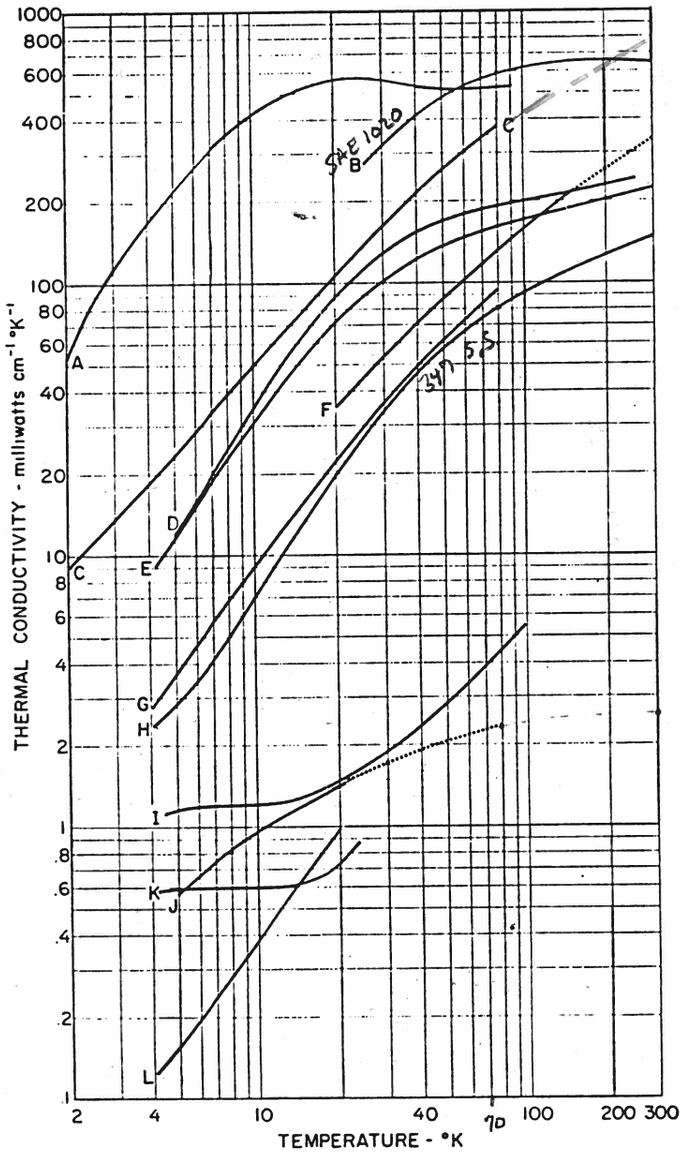


FIGURE 10.8. Low-temperature thermal conductivities of some solids with relatively low conductivities. A, 50-50 lead-tin solder [44]; B, steel, SAE 1020 [39]; C, beryllium copper [44]; D, constantan [39]; E, Monel® [39]; F, silicon bronze [42]; G, Inconel® [39]; H, type 347 stainless steel [39]; I, fused quartz [39]; J, polytetrafluoroethylene (Teflon®) [41]; K, polymethylmethacrylate (perspex) [39]; L, nylon [44].

plastics also p290 (preceding)

aving rela-
pper [42];
aluminum
m, 1100 F
; J, alumi-

A scheme for broadband calibration of the ATA

Plan: Mount a planar equiangular spiral antenna near the edge of the secondary reflector. Back it with absorber.

I. Spiral properties. [Cheo, Rumsey, and Welch; IRE PGAP, 1961, AP-9, November, p.527]

For infinitely many arms, complementary, and equal excitation, $n=1$ is the 2 arm spiral.

$$\boxed{\text{Amplitude pattern}: \quad A(\theta) = \frac{\cos\theta \left(\tan\frac{\theta}{2}\right)^{n} e^{(n/a)\tan^{-1}(a\cos\theta)}}{\sin\theta\sqrt{1+a^2\cos^2\theta}}}$$

spiral defined as $r \propto e^{-a\phi}$, small a is tighter.

$n=1$ corresponds to the 2-arm spiral

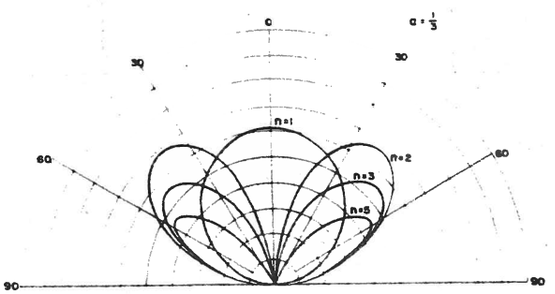


Fig. 5.

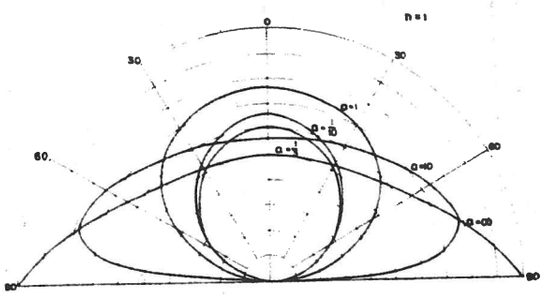
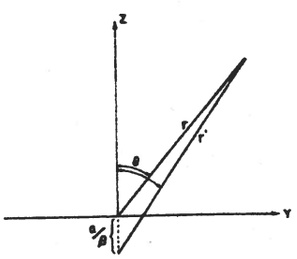


Fig. 6.



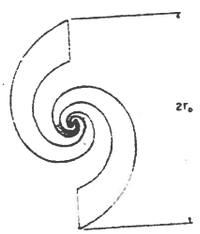


Fig. 1.

Fig. 2.

Fig. 3.

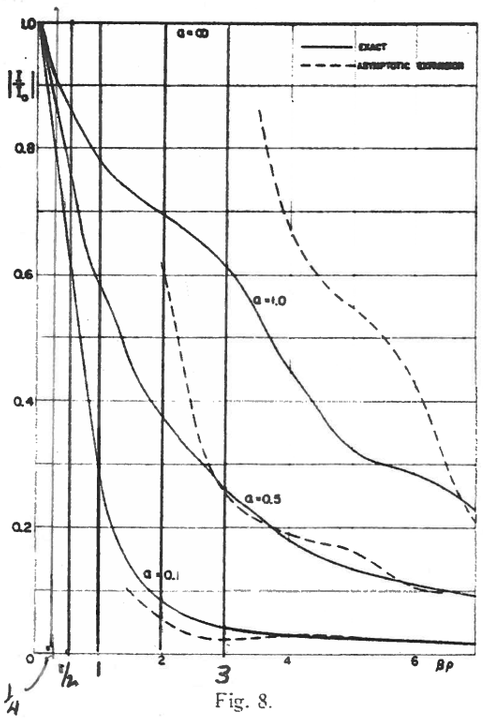


Fig. 8.

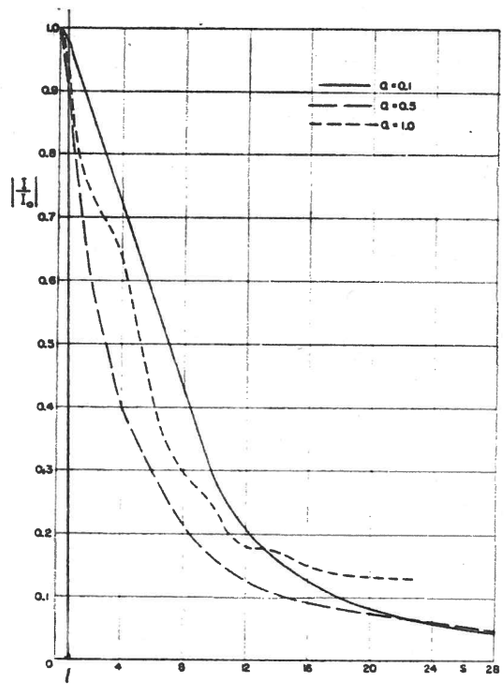


Fig. 10.

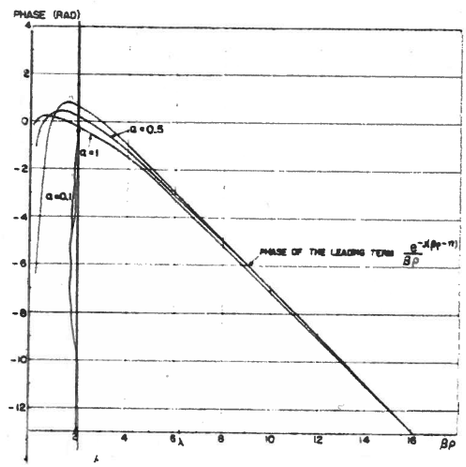


Fig. 9—Phase variation of current distribution as computed by numerical integration.

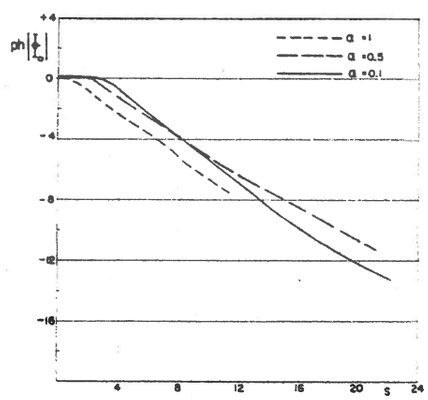


Fig. 11.

The necessary size is fixed by the longest wavelength, $\lambda = 60$ cm. Choose a small, ~ 0.1. Then the phase and amplitude curves, Figs 8 and 9, suggest a radial cutoff of $\beta\rho \cong 3$

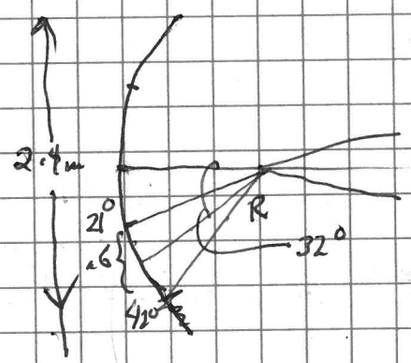
$$\beta\rho = \frac{2\pi\rho}{\lambda\rho}, \quad \text{for} \quad \frac{2\pi\rho}{\lambda} = 3, \quad \rho = \frac{\lambda}{2}, \quad \text{and } 2\rho = \lambda$$

$\boxed{2\rho = \lambda = 60 \text{ cm}}$ for the diameter; note that $\frac{60}{20} = 3$ cm (?)

II. The coupling between the feed and the planar spiral is

$$k = g_1 g_2 \left(\frac{\lambda}{4\pi R}\right)^2$$

$0.6/2.4 = 1/4$



$R = FD = 6.05/2.4 = 1.56m$

g_1: From I p. 72, $|E|^2 = .187$ at $32°$ relative to $|E|^2 = 1$ at $\theta = 0°$. The antenna gain (feed gain) is 12db at $\theta = 0°$, therefore 4.7db at $\theta = 32°$; 4.7db $\Leftrightarrow$ 3.0 for g_1

g_2: From Figure 3, it looks like the gain should be about 6db. Subtract 3db from this for the absorption of the back lobe. A more accurate calculation can be made, but $g_2 = 2$ is ok for the moment. g_1 and g_2 are independent of frequency.

$\boxed{\text{Exact } g_2 = 1.85}$ based on $A(\theta)$ above in Mathematica

For $\lambda = 0.6m$, $k = 3.2 \cdot \left(\frac{.60}{4\pi \cdot 1.56}\right) = .0056$, -22.5 db

At shorter wavelengths $k \propto \lambda^2$ and at 10 GHz (λ 3cm), the coupling has dropped by $(3/60)^2 = .0025$, -26 db to -48.5. At 12 GHz $(2.5/60)^2 \Rightarrow -27.6$ db, or $\boxed{k \cong -50\text{db}}$

This λ^2 (6db/octave) coupling can be compensated by an equalizer to provide a frequency independent calibrate signal.

III. What size calibrate signal is desirable?

With Temp ~ 50K, a $\Delta T = 1°K$ calibrate signal is probably a good choice.

For $BW = 100$ MHz, in $t = 1$ sec. $\Delta T = 50/[10^8]^{1/2} = .005°K$

For HI, $\Delta V = 1$ km/sec $\rightarrow \frac{1}{210}$ MHz ~ 5kHz.

In 1000 seconds $\Delta T = \frac{50}{\sqrt{5000 \cdot 1000}} = .02°K$

With $\Delta T = 1$ K and K ~ −50 db at 2.5 cm, a 10^5 K source is needed. The noisecom diodes put out ~ 15 db above 300 K, ~ 10,000 °K. This is too small, but a broadband amplifier with ~ 10 db gain will bring it up! e.g. this NITRO unit. (see p. 96)

RF NITRO Communications Inc

Preliminary Information

Cascadable Broadband GaAs MMIC Amplifier DC-12 GHz

NBB-300

Version 1.3

Features:

- Reliable Low-Cost HBT Design
- 13 dB Gain, +13.7 dBm P1dB @ 2 GHz
- High P1dB of +14 dBm at 6.0 GHz and +10.3 dBm at 14.0 GHz
- Single Power Supply Operation
- 50 Ω Input/Output Matched for High-Frequency Utilization.

Applications:

- Narrow & broadband commercial & military radio designs.
- Linear & Saturated amplifier applications.
- Gain Stage or Driver Amplifiers utilized in microwave radio and optical designs such as PTP, PMP, LMDS, UNII, VSAT, WLAN, Cellular, and DWDM systems.

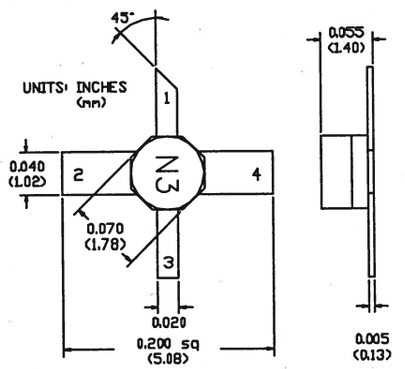
Description:

RF Nitro's NBB-300 Cascadable Broadband GaInP/GaAs MMIC amplifier is a low-cost high-performance solution for your general-purpose RF and microwave amplification needs. This 50-ohm gain block is based upon a reliable HBT (Heterojunction Bipolar Transistor) proprietary MMIC design, providing unsurpassed performance for small-signal applications.

The NBB-300 is packaged in a low-cost, hermetic, surface-mount ceramic package, providing ease of assembly for high-volume tape-and-reel requirements. The NBB-300 incorporates external d.c. decoupling capacitors, which limit the low-frequency response. Designed with an external bias resistor, the NBB-300 provides flexibility and stability to your requirement.

The NBB-300 is available in either packaged or chip (**NBB-300-D**) form, where its gold metalization is ideal for hybrid circuit designs. Packaged parts are available in 1,000 or 3,000 piece-per-reel quantities. Connectorized evaluation board designs are also available for characterization purposes.

Package: 4-Lead Ceramic Micro-X



Electrical Specifications: V_d= +3.9V, I_{cc} = 50 mA, Zo = 50Ω, T_A = +25 °C.

3.9 x (.05) = 0.195 watts

Parameter	Test Conditions	Units	Min.	Typ.	Max.
Small Signal Power Gain – S_{21}	f=0.1 to 1.0 GHz	dB	12.0	13.0	
	f=1.0 to 4.0 GHz	dB	11.0	12.8	
	f=4.0 to 6.0 GHz	dB		12.0	
	f=6.0 to 12.0 GHz	dB	9.0	10.0	
	f=12.0 to 14.0 GHz	dB		9.0	
Gain Flatness - G_F	f=0.1 to 8.0 GHz	dB		±0.6	
Input and Output VSWR	f=0.1 to 4.0 GHz			2.4	
	f=4.0 to 6.0 GHz			2.0	
	f=6.0 to 20.0 GHz			1.6	
Bandwidth - BW	BW3 (3dB)	GHz		12.5	
Output Power @ 1-dB Compression	f=2.0 GHz	dBm		13.7	
	f=6.0 GHz	dBm		14.0	
	f=14.0 GHz	dBm		10.3	
Noise Figure - NF	f=2.0 GHz	dB		4.0	
3rd Order Intercept – IP3	f=2.0 GHz	dBm		+27.1	
Reverse Isolation – S_{12}	f=0.1 to 12.0 GHz	dB		-15	
Device Voltage - Vd		V	3.6	3.9	4.2
Gain Temperature Coefficient G_T/T		dB/ C		-0.0015	

IV. An alternative would be to add another stage of gain and put the spiral antenna on the Galin pinning. This would dilute the interference with the normal operation to less than 1%.

An equalizer is needed to compensate the λ^2 coupling. The bridged T?

The Z_{in} for the self-complementary spiral is
$Z_{in} = 189\,\Omega$

V. The input circuit

[Noise Diode] — [2: NBB-300] — [Equalizer] → G

The coupling should scale by the relative areas of the primary and secondary antennas (for the same angular location of the spiral on the reflector). i.e. at $\theta = 32°$

$$\left(\frac{2.4}{6.1}\right)^2 = 0.1548 \quad , -8.1 \text{ db}$$

$-48.5 - 8.1 = -56.6$

The individual amplifier gains are typically 10 db @ 10 GHz (8-12)

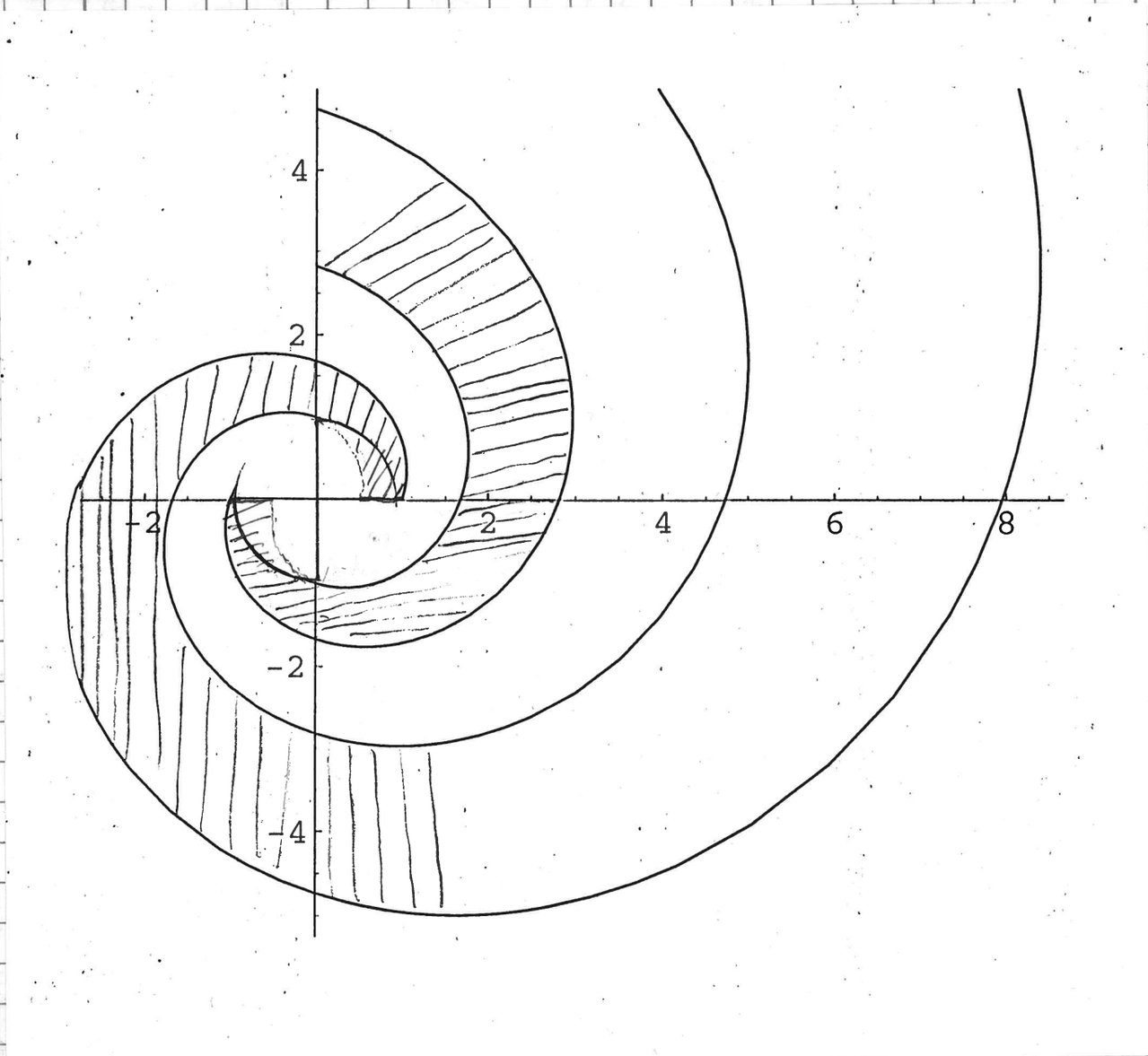
Noise @ 10 GHz: $+15 + 10 + 10 = +35$ db $\rightarrow$ 9.5×10^5 °K $\sim 10^6$ K.

10^6 K $- 56.6$ db $=$ 2.1 °K $= \Delta T$

We'll probably lose 3 db in the equalizer/impedance match.

VI. The polarization It's circular, which means that we get half the signal into each linear polarization.

$r = e^{\frac{1}{3}\phi}$ here, i.e. $a = 1/3$



VII Size of the input region for the shortest wavelength.
Figures 8 and 9 showing the amplitude and phase near the input suggest $\beta \leq \frac{1}{2}$ At $\lambda = 3$ cm $\frac{2\pi a}{\lambda} = \frac{1}{2} \rightarrow a \sim \frac{\lambda}{4\pi} \sim \frac{\lambda}{10} \sim \frac{30\,\text{mm}}{10} = 3\,\text{mm}$

$\rightarrow$ for $\beta \leq \frac{1}{4}$ $\frac{2\pi a}{\lambda} = \frac{1}{4}$, $a = \frac{\lambda}{8\pi}$ $d = 2a = \lambda/4\pi \rightarrow \frac{30\,\text{mm}}{4\pi} \sim \boxed{2\,\text{mm}}$

$\rho = e^{a\phi}$ $(ds)^2 = (d\rho)^2 + (\rho\, d\phi)^2$ $ds = d\rho \sqrt{1 + \left(\rho \frac{d\phi}{d\rho}\right)^2}$

$\frac{d\rho}{d\phi} = a e^{a\phi}$ so $ds = d\rho \sqrt{1 + \left(\frac{1}{a}\right)^2}$

$\rho \frac{d\phi}{d\rho} = \frac{e^{a\phi}}{a e^{a\phi}} = \frac{1}{a}$ $\Delta s = \Delta\rho \sqrt{1 + \left(\frac{1}{a}\right)^2} = \Delta\rho\sqrt{10}$ for $a = 1/3$

A first attempt at matching the spiral to 50 Ω coax

From Matthei, Young, and Jones p. 255 on, use λ/4 transformers multi-section.

Fractional Bandwidth parameter: $w_g = 2\left(\frac{f_2 - f_1}{f_2 + f_1}\right) = 2$ for $f_1 = 0$.

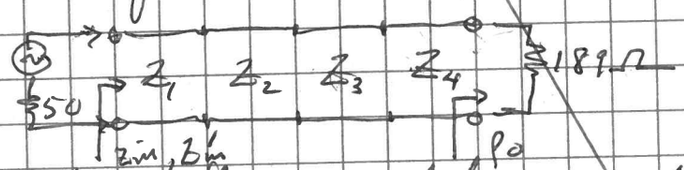
Impedance Ratio $R = 189/50 = 3.77$

$0 < f < 11.5$ GHz

The mean frequency $f_0 = 11.5/2 = 5.75$ GHz ; $\lambda_0 = 5.2$ cm

For four λ/4 sections: $Z_1 = Z_2 = Z_3 = Z_4 = \sqrt{R} = 1.942$, if $w_g = 2$.

$Z_0' = \sqrt{R}\, Z_0$



This is the simplest case. All four sections have the same $Z_0 = 1.942 \times 50 = 97.1\ \Omega$. It's just one wavelength of 97-Ω cable for the matching section.

$$\rho_0 = \frac{189-97}{189+97} = .3217, \quad |\rho|^2 \doteq 10\% \quad \text{as compared to} \quad \frac{189-50}{189+50} = .5816, \quad |\rho|^2 \doteq 34\%$$

Back down the line get $\rho_0 e^{-i\frac{2\pi L}{\lambda}} = \rho_{in}$ [N.B. it should be $e^{-i\frac{4\pi L}{\lambda}}$]

$$\rho_{in} = \frac{Z_{in} - Z_0}{Z_{in} + Z_0} \longrightarrow Z_{in}' = Z_0' \frac{(1+\rho_{in})}{(1-\rho_{in})} = Z_0' \frac{\left(1+\rho_0 e^{-i\frac{4\pi L}{\lambda}}\right)}{\left(1-\rho_0 e^{-i\frac{4\pi L}{\lambda}}\right)}$$

Relative to the 50 Ω line, the input reflection coefficient is

$$\rho_{in} = \frac{Z_{in}' - 50}{Z_{in}' + 50} \qquad x = \frac{4\pi}{\lambda} L = \frac{4\pi}{\lambda}\lambda_0 = \frac{4\pi f}{f_0}$$

$$\rho_{in} = \frac{\frac{Z_0'}{Z_0}\left[\frac{1+\rho_0 e^{-ix}}{1-\rho_0 e^{-ix}}\right] - 1}{\frac{Z_0'}{Z_0}\left[\qquad\qquad\right] + 1}$$

$$e^{-ix} = \cos x - i \sin x$$

With a little algebra

$$\frac{1+\rho_0 e^{-ix}}{1-\rho_0 e^{-ix}} = \frac{(1-\rho_0^2) - i2\rho_0 \sin x}{1-2\rho_0\cos x+\rho_0^2} = f_1(x) + i f_2(x)$$

Then $$\rho_{in} = \frac{\frac{Z_0'}{Z_0}\left[f_1(x)+if_2(x)\right]-1}{\frac{Z_0'}{Z_0}\left[\quad\right]+1} = \frac{R_0^2(f_1^2+f_2^2)-1+i2R_0 f_2}{(R_0 f_1+1)^2+R_0^2 f_2^2}$$

where $R_0 = Z_0'/Z_0$ and finally

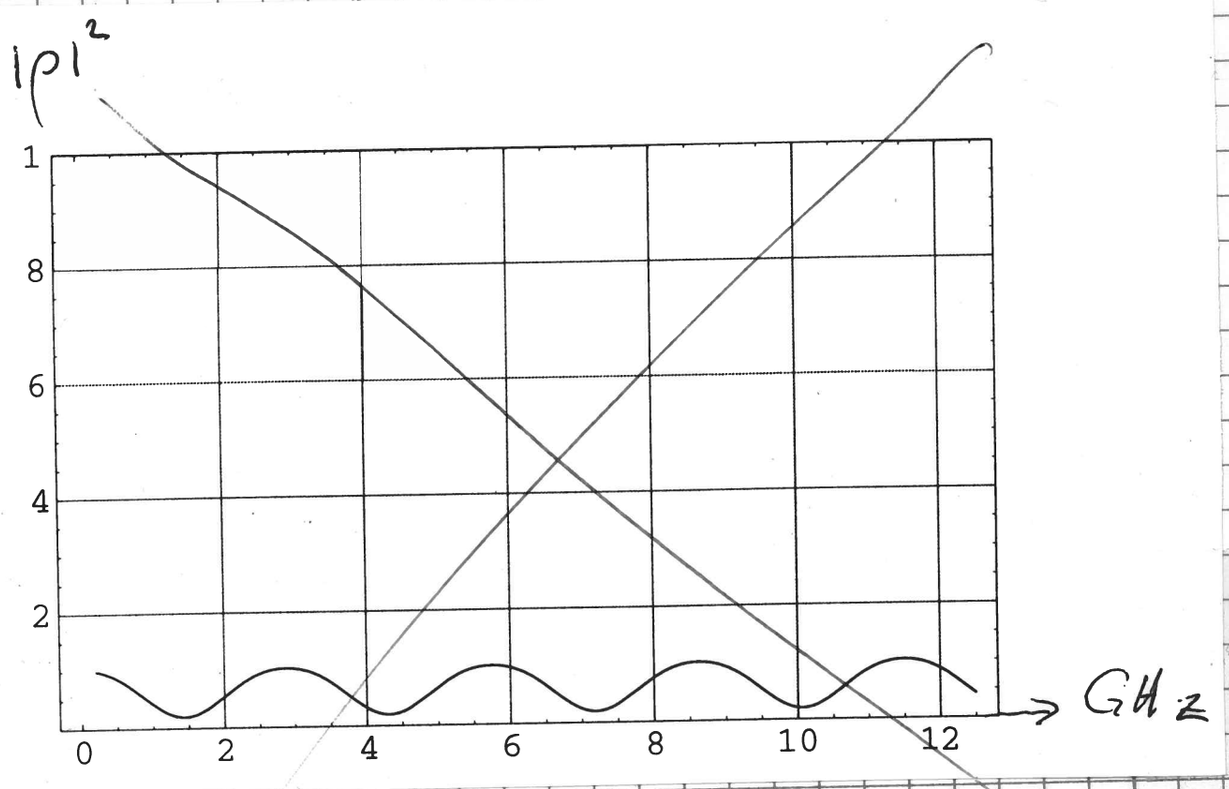
$$|\rho_{in}| = \frac{\sqrt{\left[R_0^2(f_1^2+f_2^2)-1\right]^2+4R_0^2 f_2^2}}{(R_0 f_1+1)^2+R_0^2 f_2^2}$$

$$f_1(x) = \frac{(1-\rho_0^2)}{1-2\rho_0\cos x+\rho_0^2} \qquad f_2(x) = \frac{-2\rho_0\sin x}{1-2\rho_0\cos x+\rho_0^2}$$

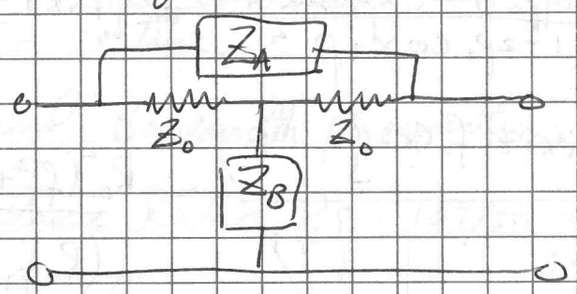
$R_0 = Z_0'/Z_0 = \frac{97}{50} = 1.942$, $R_0^2 = 3.771$

$\rho_0 = .3217$, $\rho_0^2 = .1035$, $1-\rho_0^2 = .8965$, $1+\rho_0^2 = 1.1035$, $2\rho_0 = .2020$

With $x = \frac{4\pi L}{\lambda} = \frac{4\pi f}{f_0} = \frac{4\pi}{5.75} f = 2.1855 f \Rightarrow 2.1855\, t$

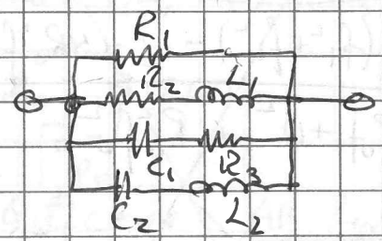


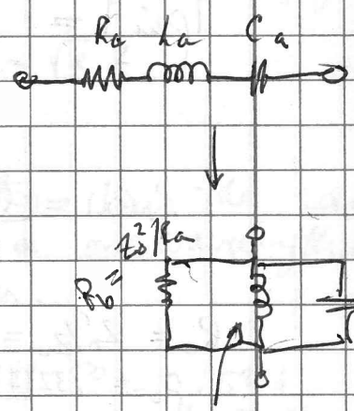
For the λ^2 equalizer the bridged T will probably work, but it will require a microstrip line version.



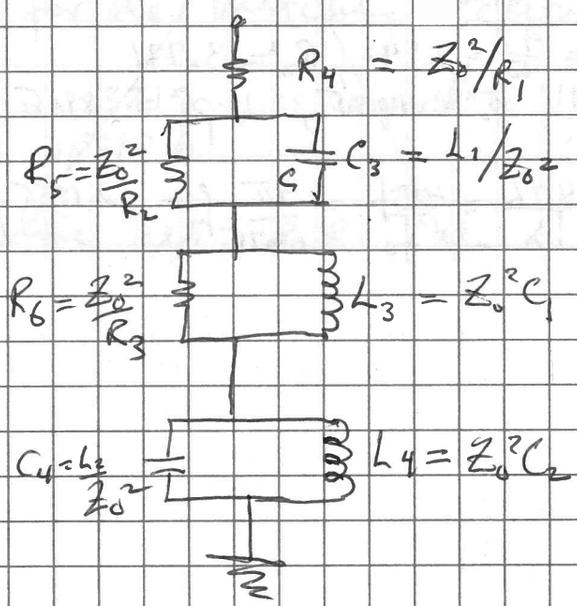
It's matched if $Z_B Z_A = Z_0^2$

If Z_A is this:

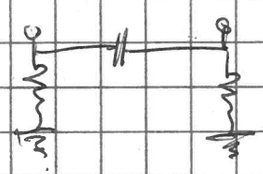




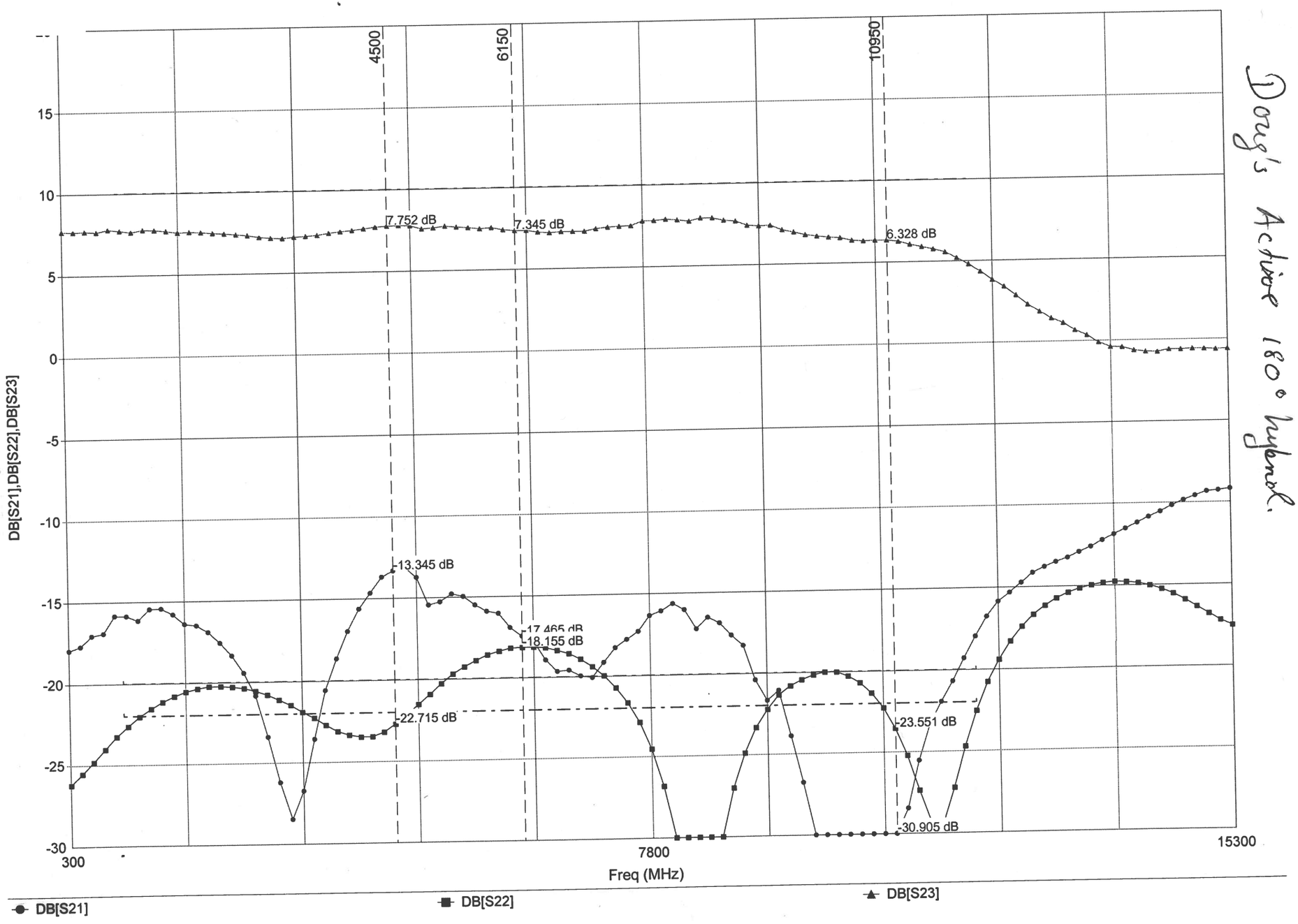
Then for Z_B:



A series capacitor should equalize the 6db/octave, mostly



C:\Program Files\GENESYS\WORK\ATA\2-amp-hybridD.wsp: Graph1 February 1, 2002 17:37:59



Doug's Active 180° hybrid.

ATA - TAP Meeting 2/8/02

JD : (a) Dave Deboer project manager
(b) → one CDR ~ April 2002 (probably late)
(c) delay 3-6 years

DD Overall Review - included a lot on the feed + antenna.
JW needs to add details → i) balun vs. twin line
make sketches
ii) more on calibrate signal
iii) use figure which shows flip and balun
iv) phase switch?
v) ± .75 db over Log Period

advantages

NB Further work on the phase switch (JW)
check Analog Devices temp. regulator - price and regulation

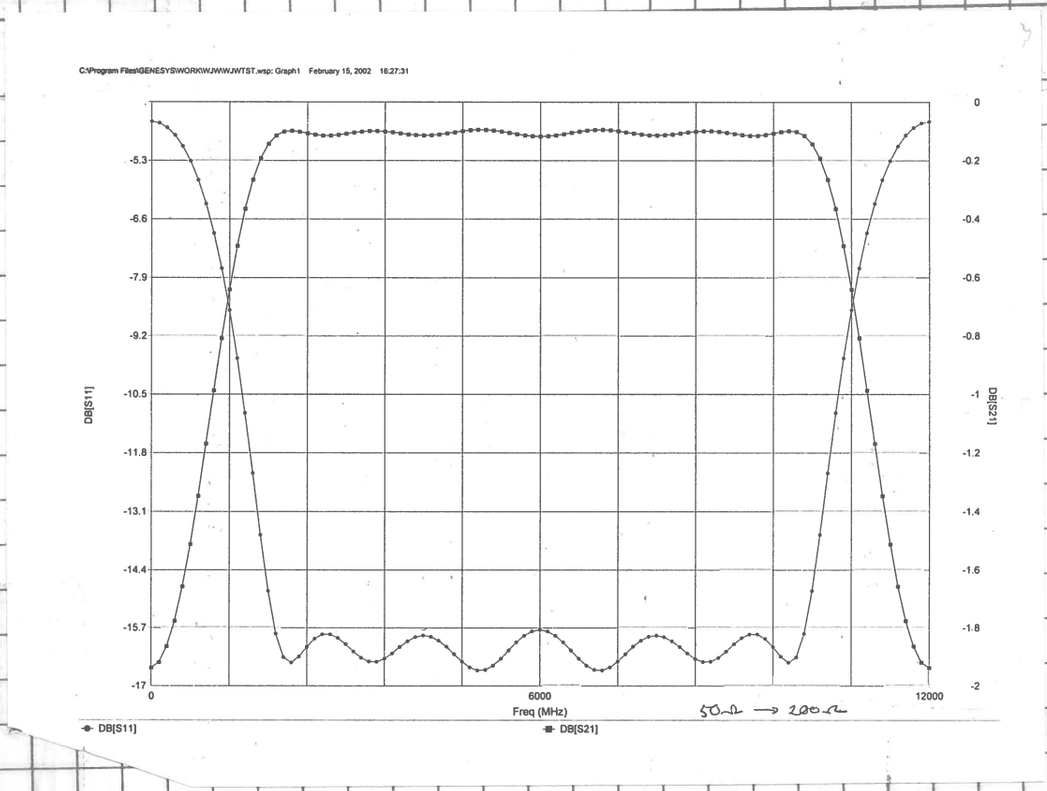
Rogers Dielectric board materials
5880 $K_e = 2.2$ @ 10 GHz Diss = .0009
RT Duroid 5870 K_e (1MHz) = 2.428 Diss. factor (1MHz) = .0014
K_e (10MHz) = 2.452 (10GHz) = .0023
½ oz: Foil electrodep. treated side RMS = 75 μm untreated 15 μm
rolled 50 15

RT Duroid 6002 $K_e = 2.94$ @ 10 GHz and $\tan\delta = .0012$ @ 10 GHz

TMM #	K_e	$\tan\delta$	@ 10 GHz
3	3.27	.0020	
4	4.50	.0020	thermally stable
6	6.00	.0023	more so than alumina
10	9.20	.0020	

RO 3003 - .06 db/inch @ 10 GHz
RO 4003 - .12 db/inch @ 10 GHz
RO 4350 - .20 db/inch "

Calibrator output power: KTB @ 10^6 = 7 $= (1.4\times10^{-23})(10^6)10^{19} = 1.4\times10^{-7}$ watts
to spiral



50Ω → 200Ω

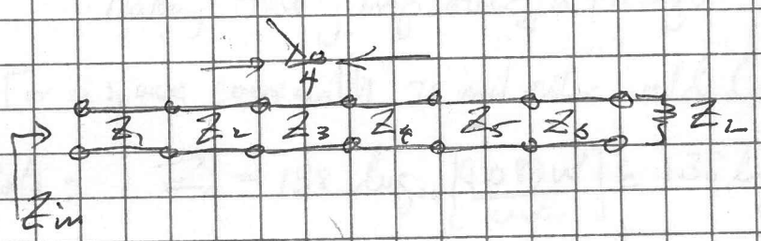
A. 6 λ/4 stubs in series: center $f_0 = 6$ GHz; 50Ω in 200 Ω out.
The characteristic impedances adjusted by hand using Eagleware. (0.6% steps)
Final values: $Z_1 = 76.3\,\Omega$, $Z_2 = 96.4$, $Z_3 = 122.3$, $Z_4 = 150.4$, $Z_5 = 175.3$, $Z_6 = 191.5$

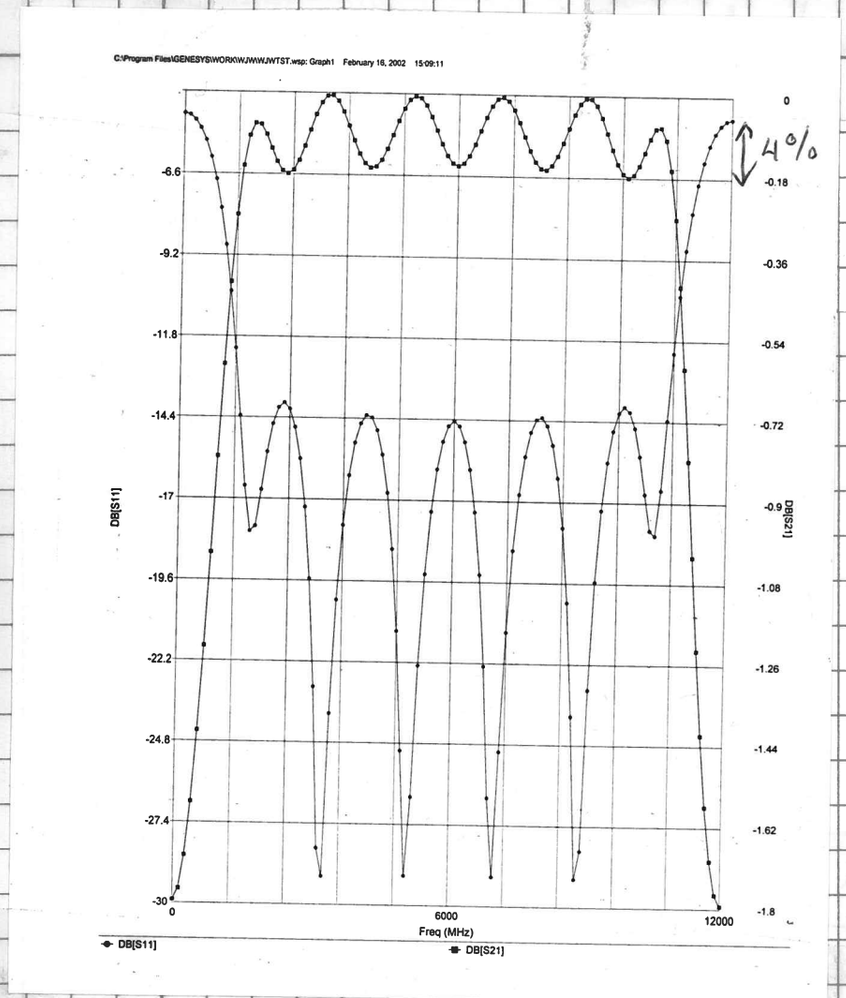
B. 6 λ/4 stubs in series; center $f_0 = 6$ GHz, 50 Ω 189 Ω out
Transmission from 1.5 GHz – 10.5 GHz is ~ –.03 db
–0.5db at 1 GHz and 11 GHz, –1.3db @ 500 MHz and 11.5 GHz.
Final characteristic Z's: $Z_1 = 66.6$, $Z_2 = 82.2$, $Z_3 = 104.2$, $Z_4 = 129.7$, $Z_5 = 154.9$, $Z_6 = 175.6$
Max S_{11} is –19.2 db.

C. Match to 150 Ω with 6 sections (by hand every time) very flat 1.8 – 11.2
@ –.02db, –.45db at 1 GHz and 11 GHz; –1.0 @ 500 MHz

Z_0's: 65.0, 81.1, 101.6, 122.6, 138.6, 147.6.

D. With the load changed from 150 to 189 Ω the peak mis-match is –14.4db and the transmission oscillates p-p 0.18db, 1 < f < 11 GHz.
All with the above Z_0's. –1.1db at 500 MHz and 11.5 GHz



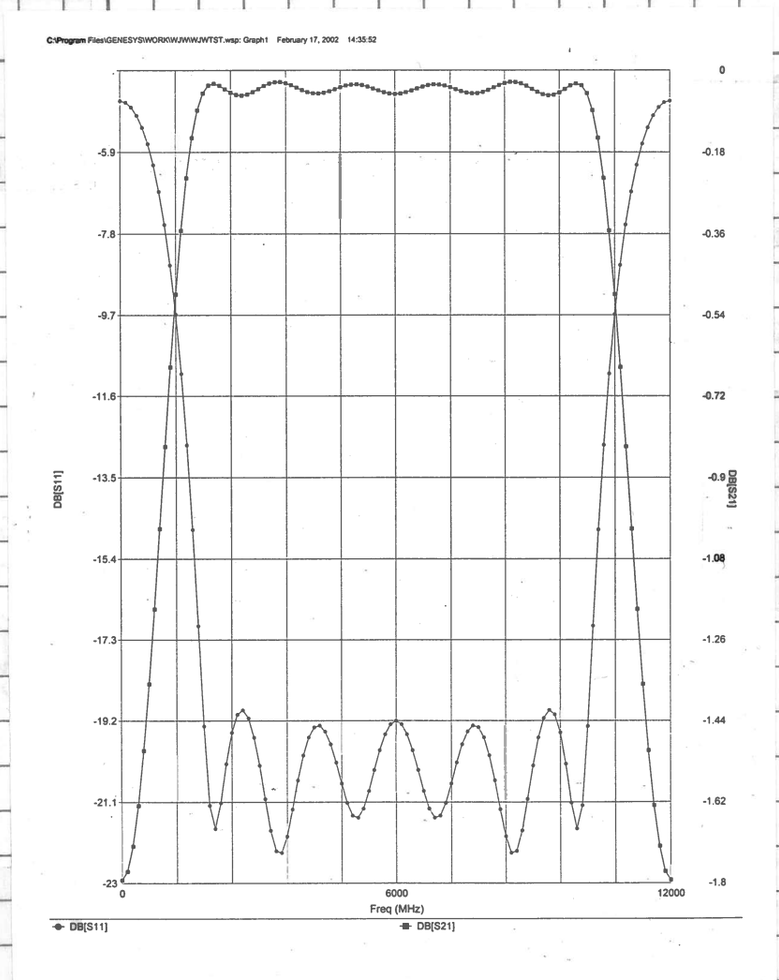


6 section filter set for 150 Ω load, but with 189 ohm load. The λ/4 section impedances are 65.0, 85.1, 101.6, 122.6, 138.6, 147.6

Putting this at the input terminals of the spiral will be difficult to separate from the antenna.

E. 2 λ/4 sections (f_0 = 6 GHz) gives ± 0.3db over 2–10 GHz, (15% p-p) −1.4db at 1 GHz and 11 GHz, and −1.7db @ 5.5 and 11.5 GHz.

The trouble with leaving it unmatched is that the big standing wave will give a very temperature dependent coupling at various frequencies. The maximum attenuation will be 0.66 at the peaks of attenuation, 34% almost 20 times worse than the above case D.



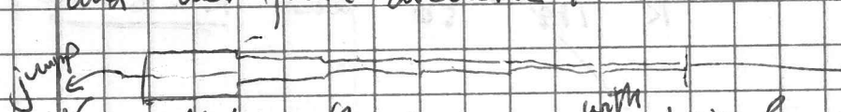
B. Filter to match 189Ω ; 6 λ/4 sections

Stub Impedances: 66.6, 82.2, 104.2, 129.7, 154.9, 175.6

Connection to the antenna. Diameter at the inner edge. If we make $2\pi r = \frac{\lambda}{3}$ for $a = 1/3$, $I = .85 I_0$, probably ok. Then $2r = d = \lambda/3\pi$

For 12 GHz, $\lambda = 25$ mm and $d = 25/3\pi = 2.65$ mm. If that's the inner edge of the inner arms. Then each arm diameter is $\sim d/2 = 1.33$ mm $= 52$ mils. 60 mil width should be ok. Probably even 80 mils would be ok, allowing a 10 mil spilover on each side of the arm for the incoming transmission line.

Use a sequence of square coaxial transmission lines for the filter. Use a fixed inner diameter of 4.4 mils (the inner coax diameter of available cable). Feed it with the 50 ohm coax {UT20-m from Micro Coax}. And open the outer dimension in sequence to get the above impedances. Mill a sequence of rectangular grooves for the filter and use foam dielectric to hold the small wire on center.



Making the jump over with a triangle will minimize the lead inductance

For a square coax with 75 mil outer wall (in air)

Note: $Z_0 = 138 \log_{10}\left[\frac{(1.08)W}{\text{center}}\right] = 138 \log_{10}\left[\frac{(1.08)(75)}{4.4}\right] = \underline{\underline{174.6\ \Omega}}$

Widths of the filter sections, if they were coaxial with square outer surfaces, and with inner diameter of 40.4 mils

$$Z_0 = 138 \log_{10}\left(\frac{(1.08)\ell}{D}\right) \Rightarrow \ell = \frac{d}{1.08}\,10^{(Z_0/138)}$$

Z_0	ℓ (mils)
66.6	12.4
82.2	16.1
104.2	23.2
129.7	35.5
154.9	54.0
175.6	76.3

The filter sequence must be laid along the spiral arm. Assuming that a is 1/3 (for $e^{a\phi}$) and the starting point is at an approximate radius of 2.0 mm. we can estimate the extent in angle ϕ_0 of the filter.

$$\rho = \rho_0 e^{a\phi} \qquad (ds)^2 = (d\rho)^2 + (\rho\, d\phi)^2$$

$$d\rho = \rho_0 a e^{a\phi} d\phi \qquad (ds)^2 = (\rho_0 a e^{a\phi} d\phi)^2 + (\rho_0 e^{a\phi} d\phi)^2$$

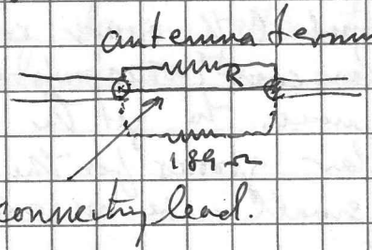
$$ds = \rho_0 e^{a\phi} d\phi \sqrt{a^2+1} \qquad \Delta s = 6 \cdot \frac{\lambda}{4} = 6 \cdot \frac{25\text{ mm}}{4} = 37.5\text{ mm}$$

$$\Delta s = \int_0^{\phi_0} \rho_0 e^{a\phi} d\phi \sqrt{a^2+1} = \rho_0 \frac{e^{a\phi}}{a}\Big|_0^{\phi_0} \sqrt{a^2+1} = \rho_0 \sqrt{1+\frac{1}{a^2}}\left(e^{a\phi_0} - 1\right)$$

For $a = 1/3$, $\rho_0 = 2.0$ mm, and $\Delta s = 37.5$ mm, $\phi_0 = 5.81$ radians, 0.92 turns

---

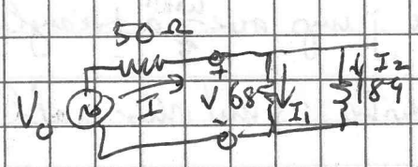
A simpler alternative: match the antenna with a resistor and add an amplifier to make up the loss.



Put a resistor across the terminals to reduce the 189 Ω to 50 Ω

$$\frac{1}{R} + \frac{1}{189} = \frac{1}{50} \quad \text{and} \quad R = 68\ \Omega$$

Power loss



$V = V_0/2$ $\qquad I_1 = V/68 = \frac{V_0/2}{68} = \frac{100 I/2}{68} = \frac{50}{68} I = .735 I$ ✓

$I = \frac{V_0}{100}$

$I_2 = V/189 = \frac{V_0/2}{189} = \frac{50 I}{189} = .265 I$ ✓

The available power is: $P_a = I^2 50$ if the load were a 50Ω match.

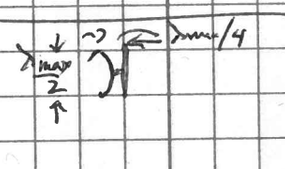
$P_1 = I_1^2 68 = (.735 I)^2 68 = .540 I^2 68 = (.540) 68 P_a/50 = .734 P_a$

$P_2 = I_2^2 189 = (.265 I)^2 189 = .070 I^2 189 = (.070) P_a/50 \cdot 189 = .266 P_a$ ✓

So the antenna gets $.266 P_a \sim (1/4) P_a$ $\qquad$ **−6 db**

This must be made up with an amplifier. Probably 3 of the NITRO amps are required.

The TRW wide-band antenna feed ①



from Sandy: $\Theta_{1/2} = 70°$

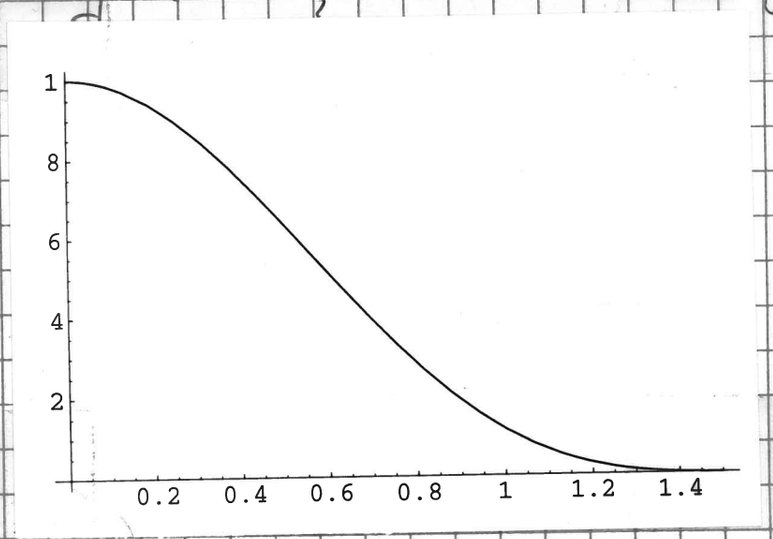
For the LP feed with $\Theta_{1/2} = 42°$

Since $G_0 \propto \Delta\Omega$, $\left(\frac{42}{70}\right)^2 = .36$, −4.4db and the TRW feed gain should be ~ 4.4db lower than the LP feed.

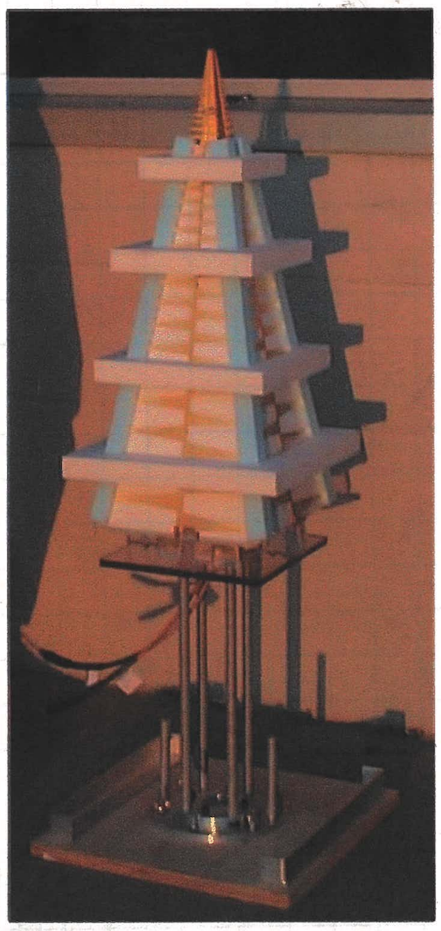
$12.0 - 4.4 = 7.6$ db expected.

Expect optimum f/D ~ 0.5, that's $\Theta_{edge} = 53°$

If $(\cos\Theta)^{3.5} \propto g(\Theta)$ fits the pattern then the spillover between 53° and 90° is about 4%; the back radiation cannot be fitted by $(\cos\Theta)^{3.5}$, so there is no estimate of that.



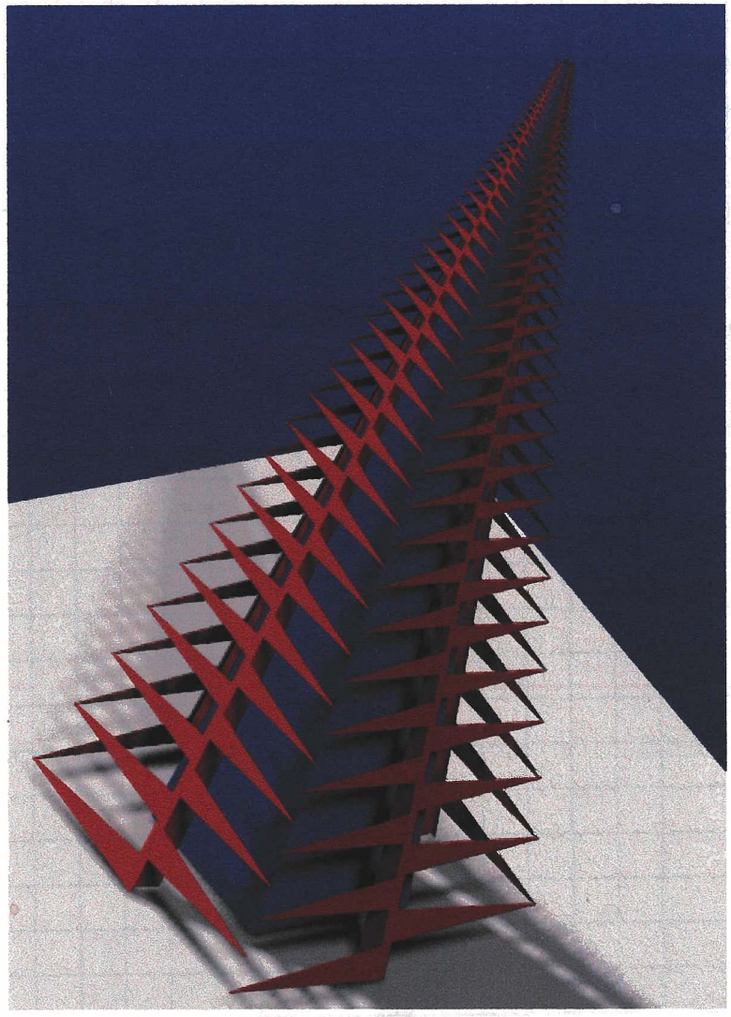
Prototype Feed for Tests



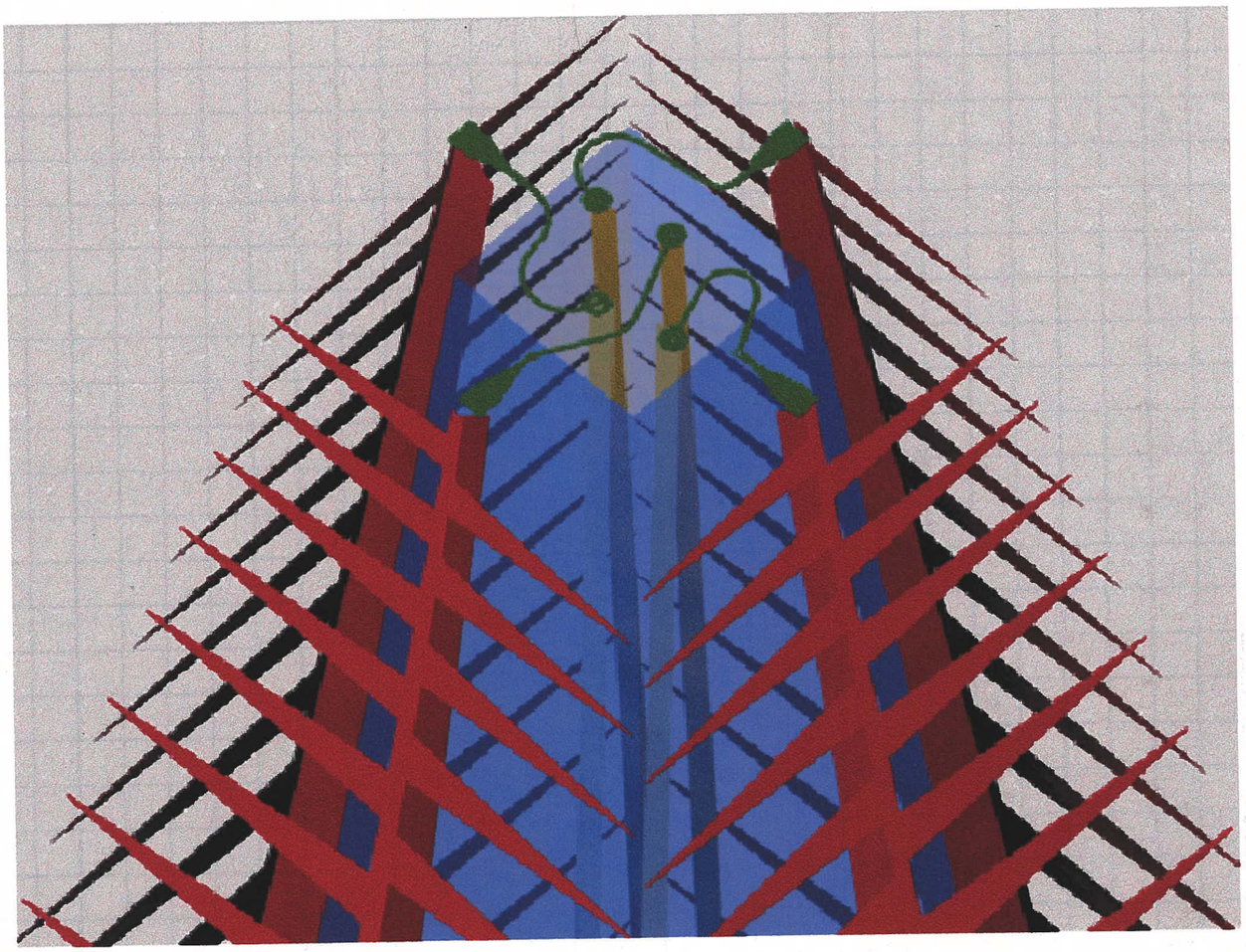
Constructed from thin brass shim and gold plated

Strengthened by fin --- feed pattern improved

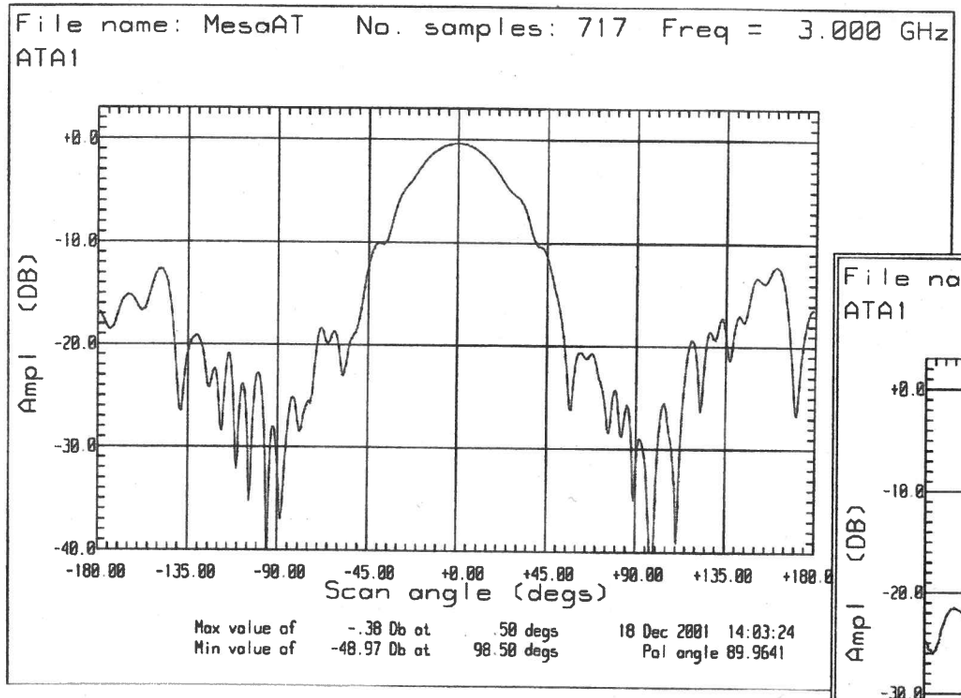
Packaging studies now in progress

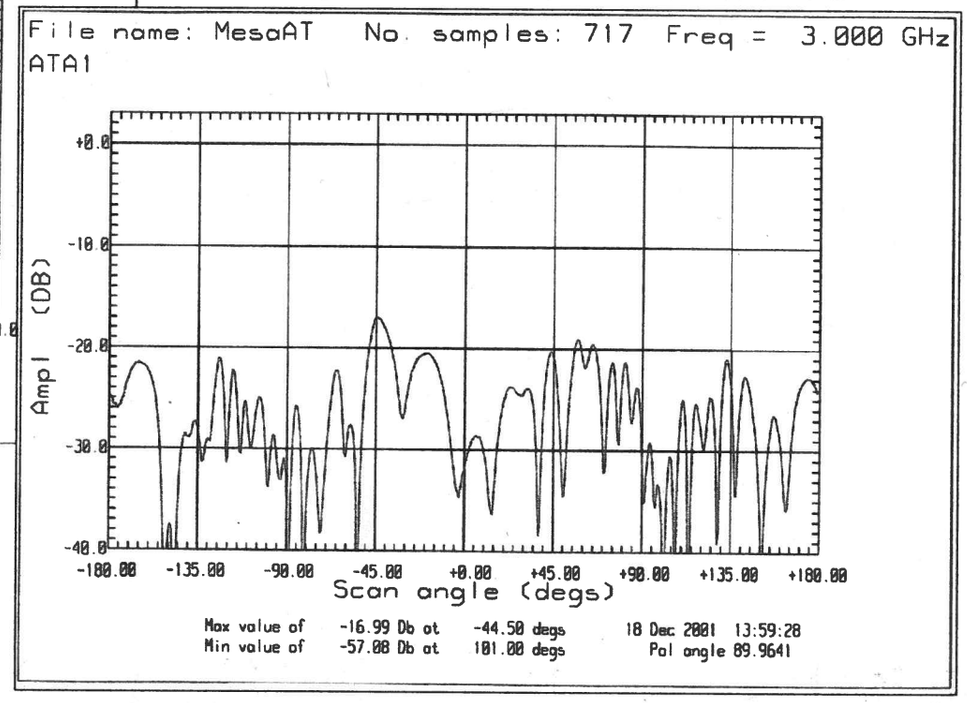


Feed Tip

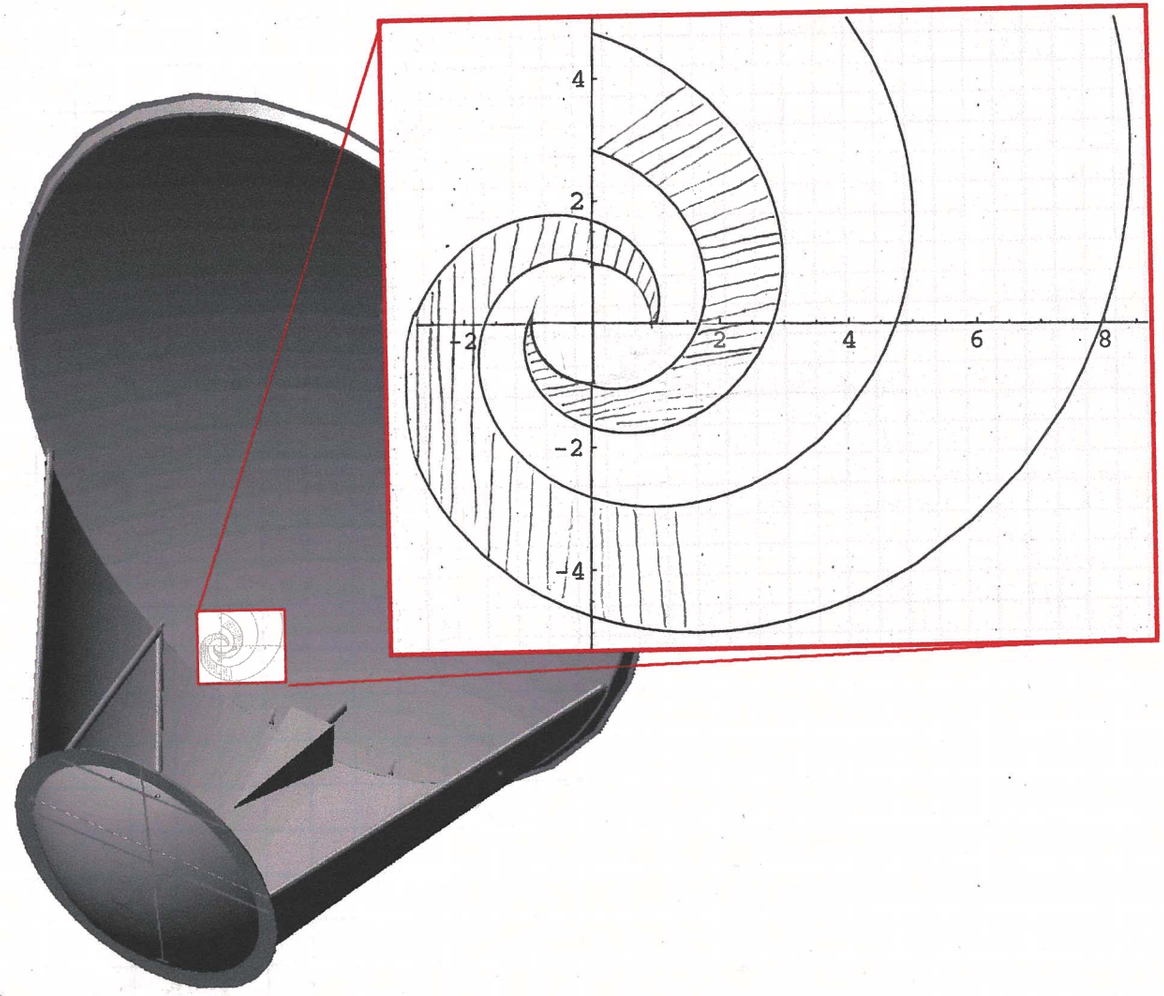


Final Measured Feed Patterns



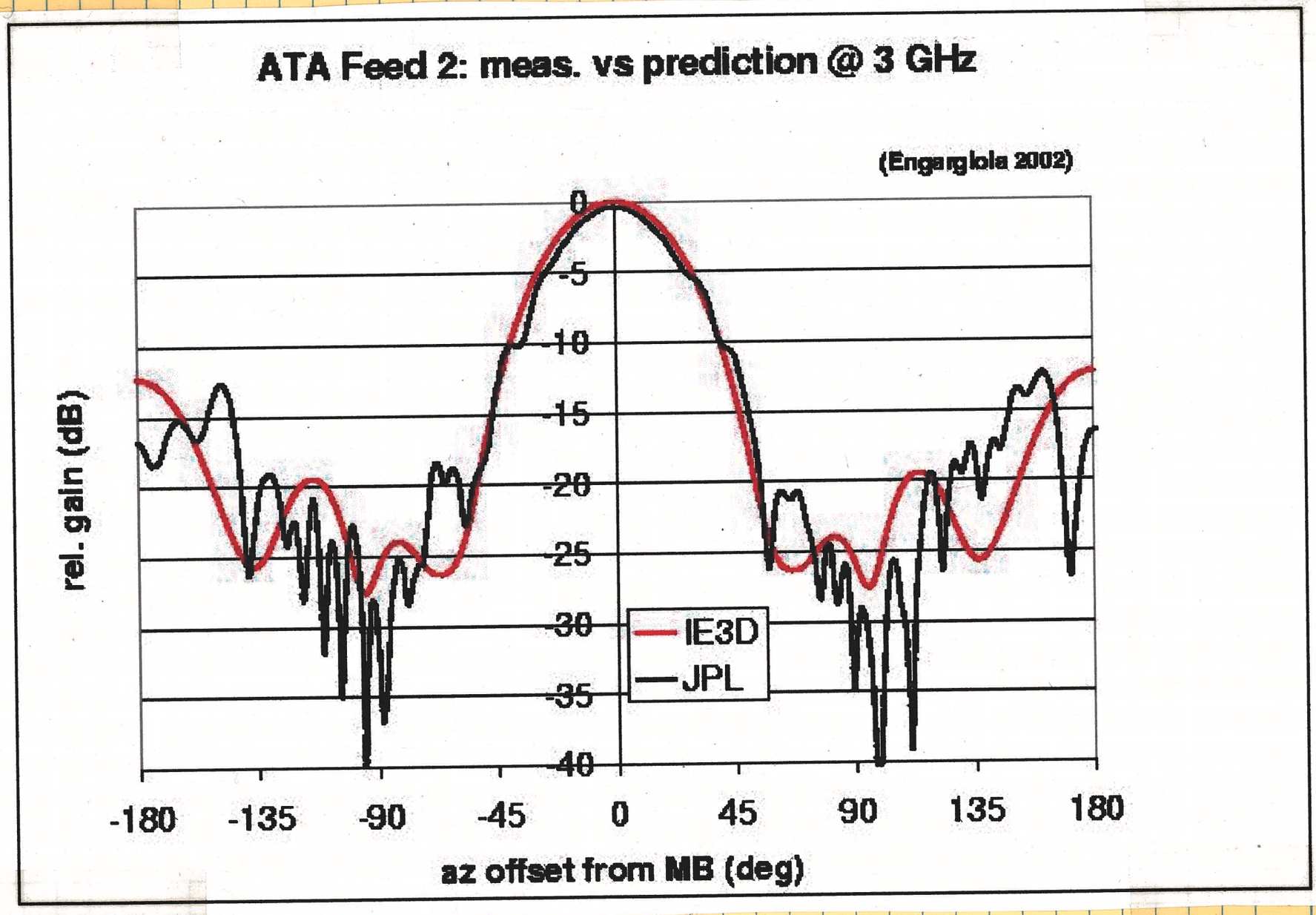


Calibration Source: Equiangular Spiral



Noise Summary for 10 GHz

Item	Temperature	Notes
Feed (3%)	9.0	Const w freq
Tip Circuit Bd	3.0	Varies w freq
Transition Line	2.6 (20mil)	Varies w freq
Cold Line	2.4	Varies w f req
Spillover	3.0	Const w freq
Tbb	2.7	Const
Atmosphere	7.0	4.0 @ 4GHz
LNA	12.0	
Total	42.0	36.0 @ 4GHz

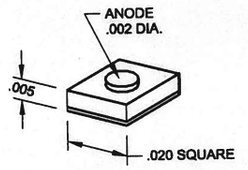


Improved simulation with a finer grid: 15/λ → 25/λ

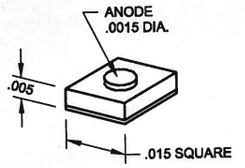
96

NC401 (1-9) $58 each
(500) $50 each

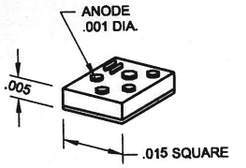
CH1 Chip (inches)



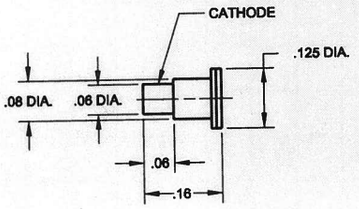
CH2 Chip (inches)



CH3 Chip (inches)

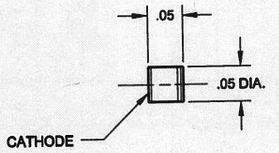


C10 (inches)



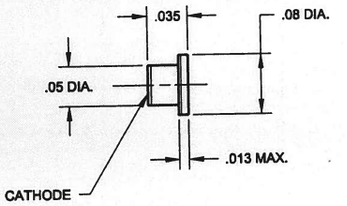
Cp=0.18 pF TYP.
Ls=0.40 nH TYP.

C50 (inches)



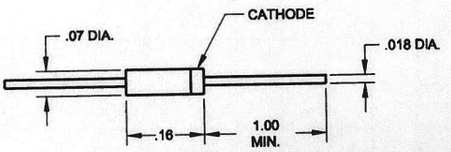
Cp=0.13 pF TYP.
Ls=0.40 nH TYP.

C50H (inches)



Cp=0.23 pF TYP.
Ls=0.20 nH TYP.

DO-35 Package (inches)

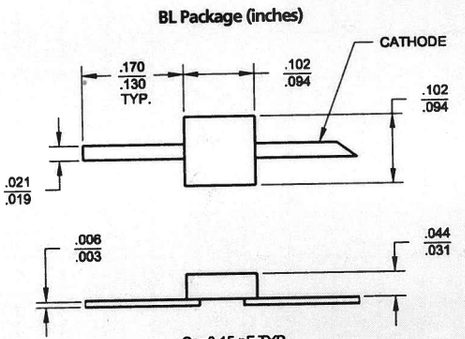


Cp=0.10 pF TYP.
Ls=1.00 nH TYP.

AUDIO & VHF TYPES

MODEL	FREQUENCY RANGE	OPERATING CONDITIONS V_b (V)	I_{op}	R_L (Ω)	MINIMUM OUTPUT (μV/√Hz)	PACKAGE
NC101	0.1 Hz - 100 kHz	7 - 10	30 - 60 μA	2200	3.0	DO-35
NC102	0.1 Hz - 500 kHz	7 - 10	30 - 60 μA	2200	3.0	DO-35
NC103	0.1 Hz - 1 MHz	7 - 10	30 - 60 μA	2200	3.0	DO-35
NC104	0.1 Hz - 3 MHz	7 - 10	30 - 60 μA	2200	3.0	DO-35
NC201	0.1 Hz - 10 MHz	7 - 10	0.2 - 0.5 mA	2200	0.1	DO-35
NC202	0.1 Hz - 25 MHz	7 - 10	0.2 - 0.5 mA	2200	0.1	DO-35
NC203	0.1 Hz - 100 MHz	7 - 10	0.2 - 0.5 mA	50	0.05	DO-35

BL Package (inches)

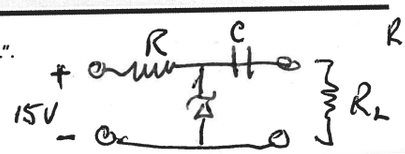


Cp=0.15 pF TYP.
Ls=0.40 nH TYP.

RF & MICROWAVE TYPES

MODEL	FREQUENCY RANGE	OPERATING CONDITIONS V_b (V)	I_{op}(mA)	R_L (Ω)	OUTPUT ENR (dB)	PACKAGE
NC302L	10 Hz - 3 GHz	6 - 8	6	50	30 - 35	DO-35 BL CH1
NC303	10 Hz - 8 GHz	8 - 12	8	50	30 - 35	DO-35 BL CH1
NC305	10 MHz - 11 GHz	8 - 12	10	50	29 - 34	BL CH1
NC401	100 MHz - 18 GHz	8 - 12	10	50	30 - 35	C10 C50H CH2
NC403	100 MHz - 27 GHz	8 - 12	12	50	24 - 28	C50 CH3
NC404	18 GHz - 50 GHz	8 - 12	15	50	20 - 25	C50 CH3
NC405	18 GHz - 75 GHz	8 - 12	20	50	15 - 25	C50 CH3
NC406	18 GHz - 110 GHz	8 - 12	25	50	15 - 25	C50 CH3

1. For chip configuration, add suffix "C".
2. For beam lead configuration, add suffix "BL".
3. For C50H configuration, add suffix "H".



R = adjust for performance

f_c = low freq. cutoff; $C \geq \frac{1}{2\pi R_L f_c}$

Noisecom.com To order call 201-261-8797

R = 500Ω for 5V drop

For the NC401: An ENR of 30db is $10^3 \times 300 = 300{,}000K$

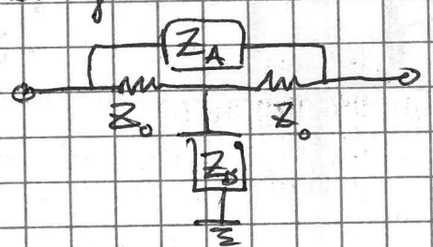
The path loss is −56.6 db. For 1K in each polarization subtract 3db.
Then $-59.6 \times 300{,}000 = 1.096\times10^{-6} \times 300{,}000 = 0.33K$

NB −6db in the matching ~~current~~ resistor and probably 3-4db in the coax, say, −10db total.

$-10db \times 0.3K \Rightarrow .03K$? A factor of 33 (+15db) is needed. It requires 2 amps. for a net gain of +15db, including the equalizing circuit

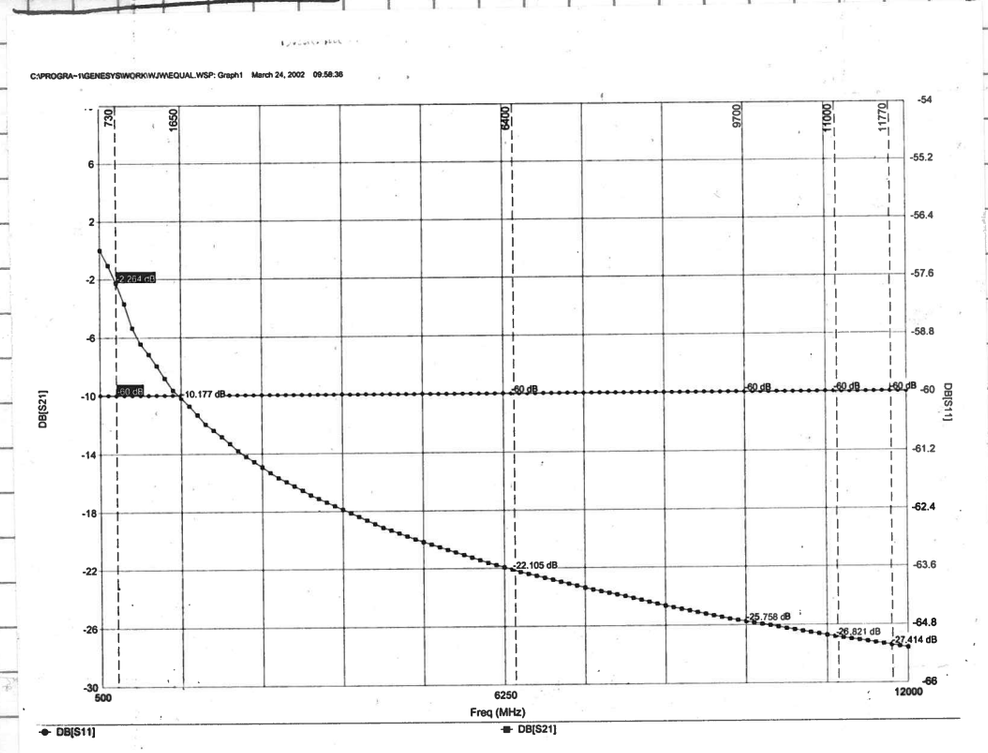
Equalizers for the NITRO amps, Hittite Switch, input cable loss, and antenna Transmission loss!

Use the bridged T circuit here.

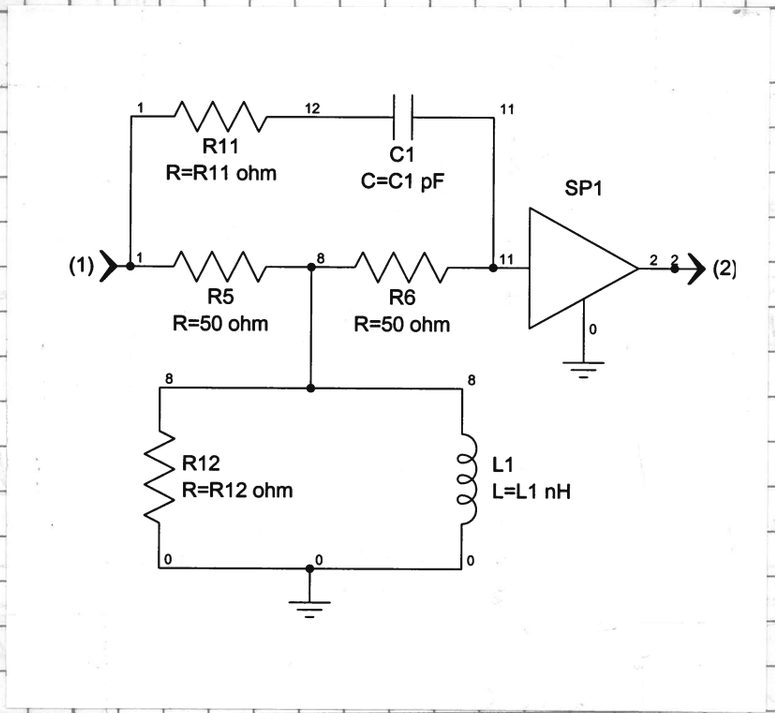


$Z_A Z_B = Z_0^2$ insures that it be matched in and out.

I. Transmission loss S_{11} set to −60db in the file.



SP1



constraints

$$R_{12} = \frac{2500}{R_{11}} = \frac{Z_0^2}{R_{11}}$$

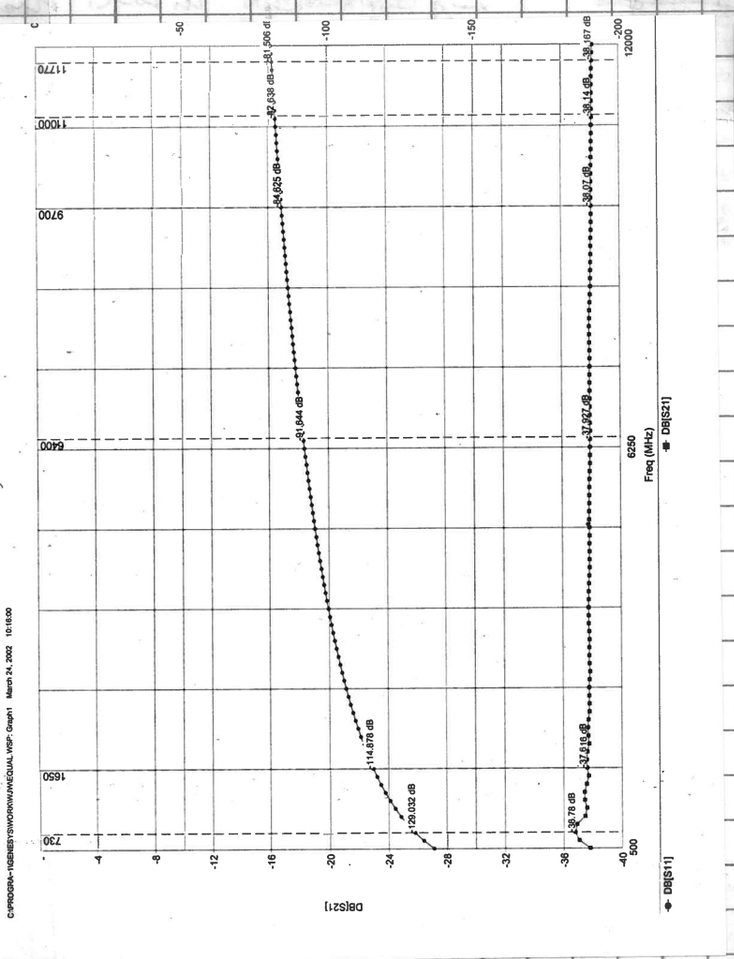
$$L1 = 2.5 \times C1$$

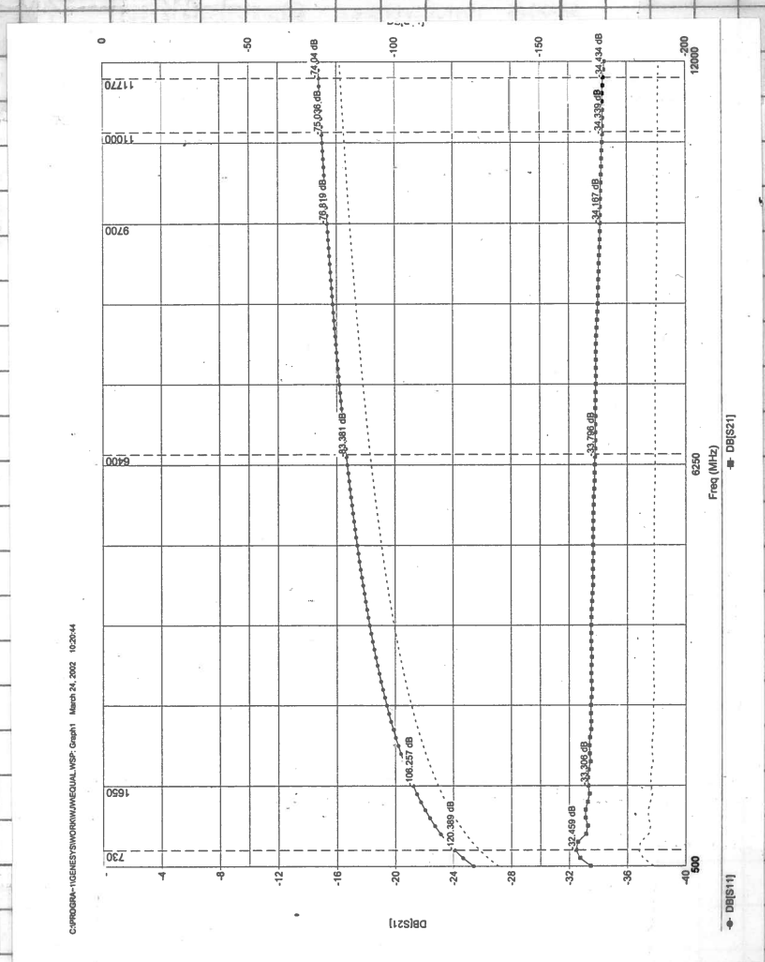
C1 = .08
R_{11} = 1389
R_{12} = 6427

0.5db slope overall

C1 = 0.1Bpf
R_{11} = 1389

1db slope overall





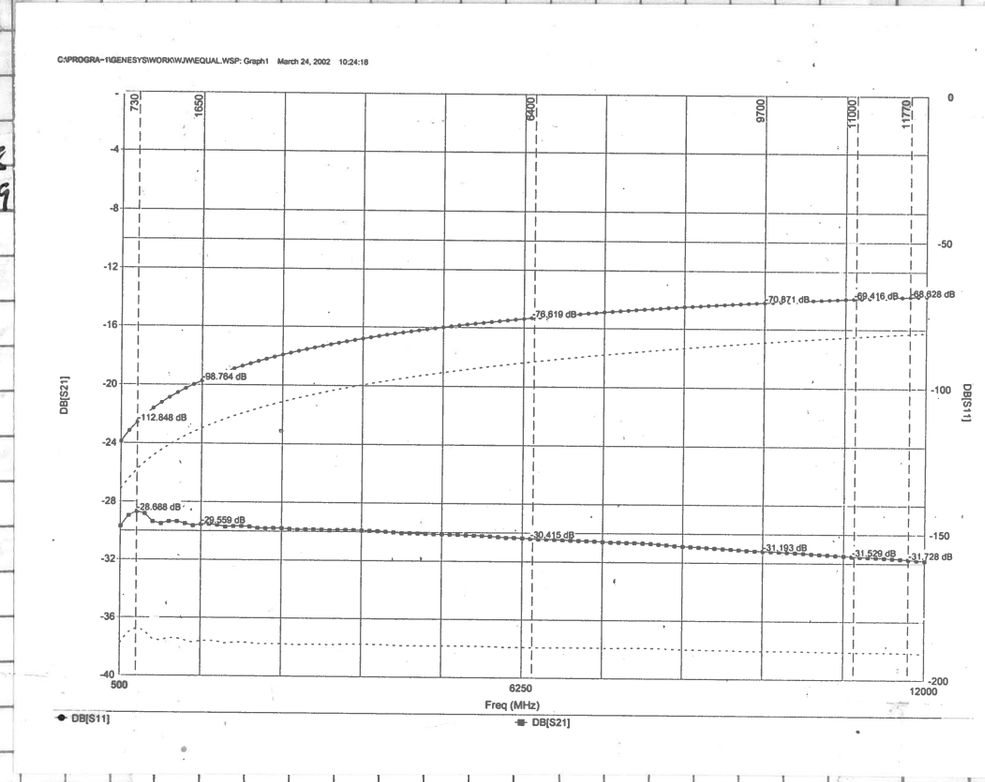
$C1 = .08$
$R_{11} = .559$

R_{11} can be set to zero here with little effect. That also means $R_{12} = \infty$. Replacing $L1$ by a short lowers the transmission by about 6db and raises S_{11}

For $C1 = .08$ $L1 = 2.5 \times C1 = .2$ nh which is small.

The inductor is needed to flatten the slope.

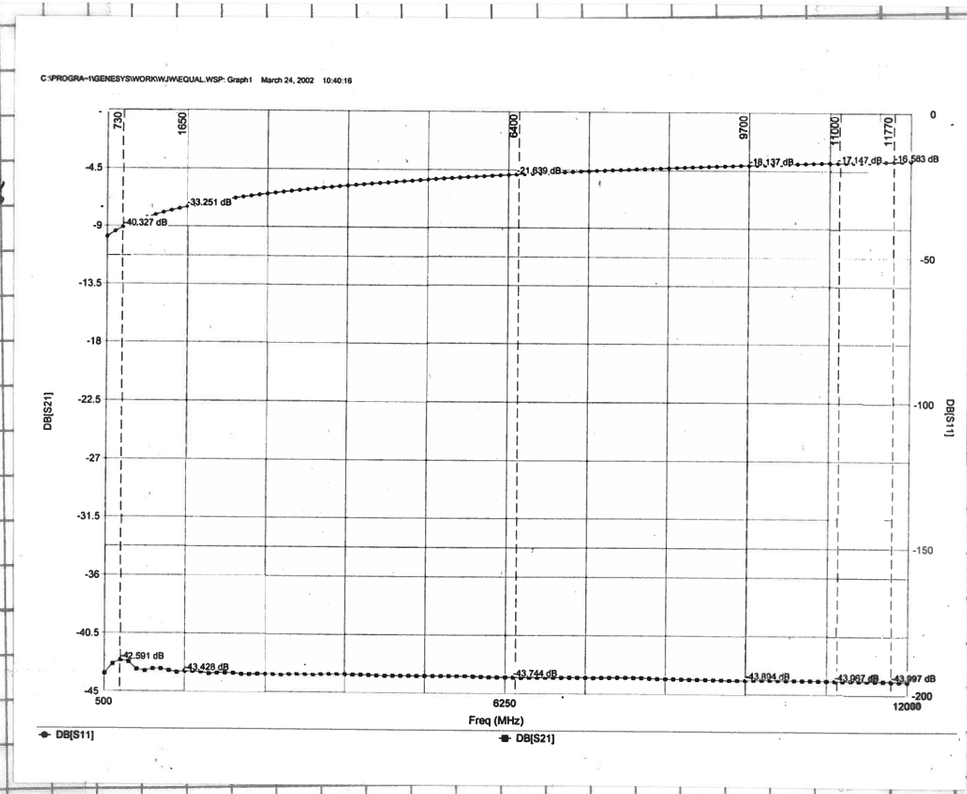
2db slope



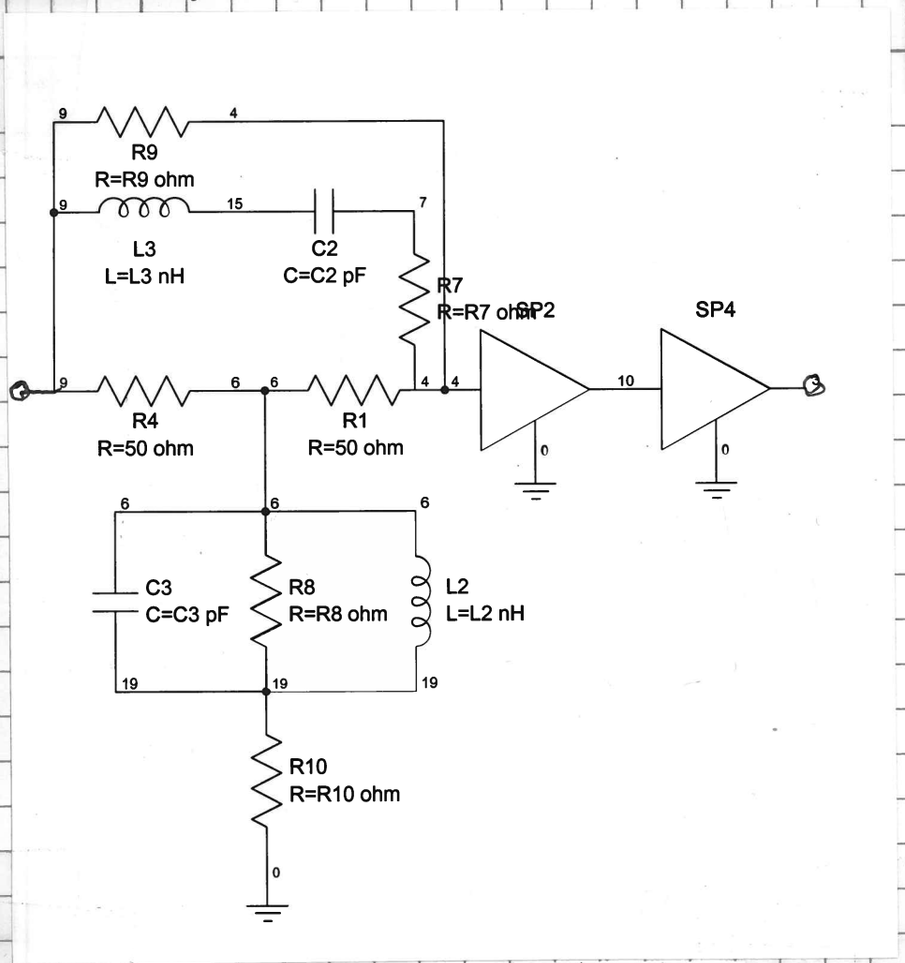
$C1 = .08$

$L1$ set to zero

2db slope



II. Equalizer for NITRO amps.



constants

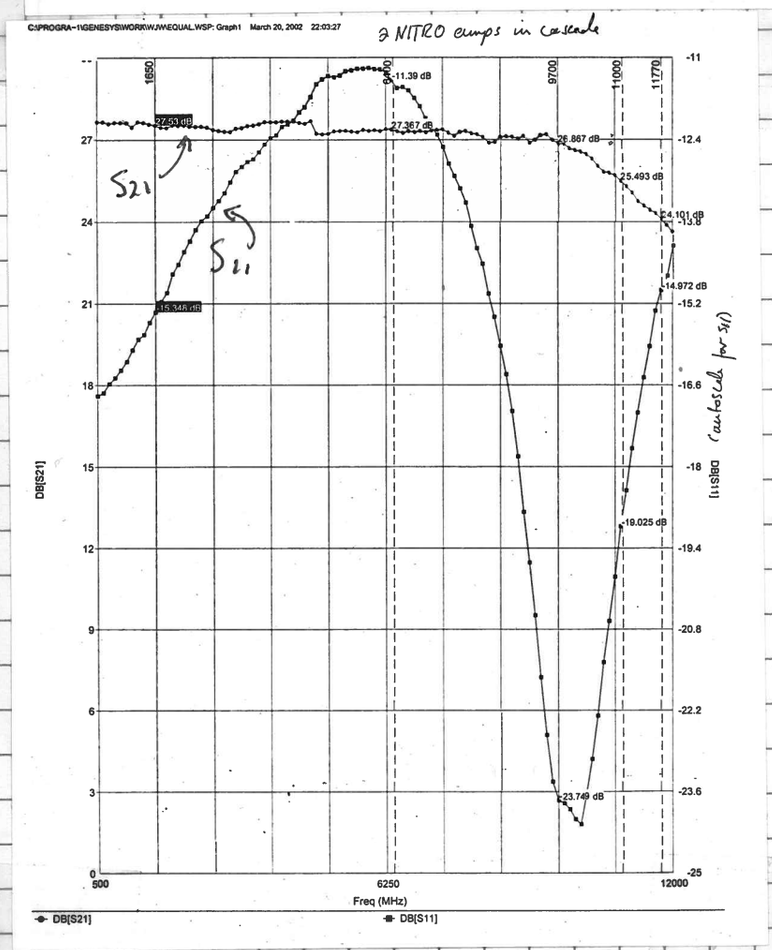
$R_{10} = 2500/R_9$
$R_8 = 2500/R_7$
$L_2 = 2.5 \times C_2$
$C_3 = 2.5/L_3$

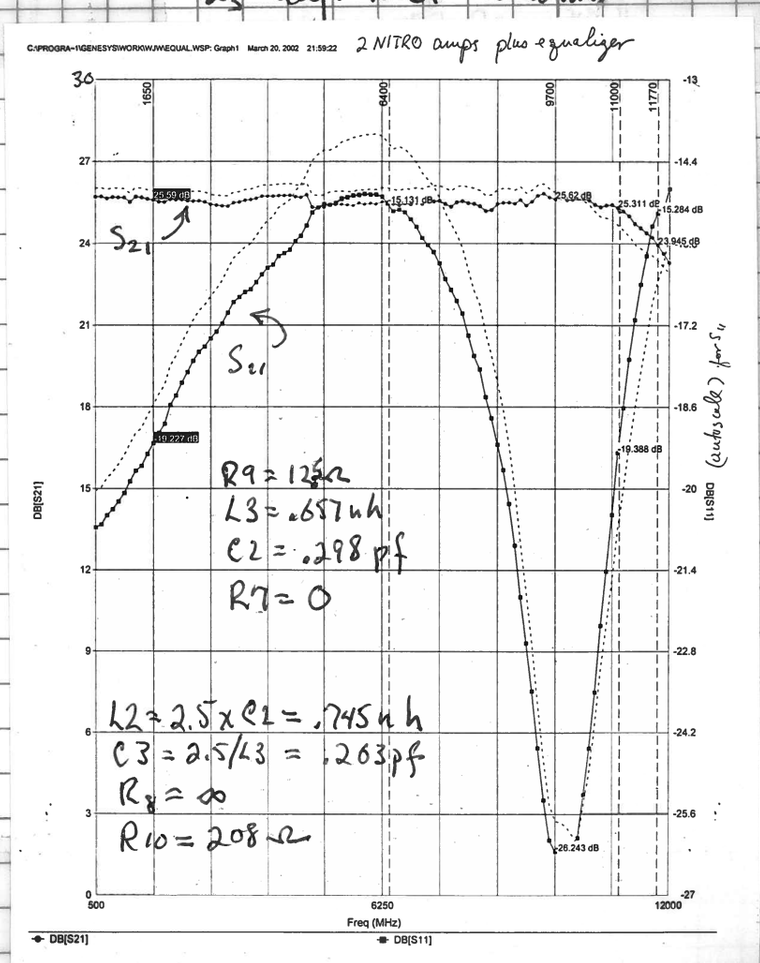
Both tuned circuits have the same resonant frequency. Set this frequency close to the top of the band for maximum transmission there.

Keeping $L_3 C_2 = \frac{1}{\omega_0^2}$

for $f_0 = 12.59\,GHz$ is a further constraint. Then L_3, R_9 and R_7 are the only variables. It turns out $R_7 = 0$ works best. (and $R_8 = \infty$)

R_9 ~~sets the~~ fixes the amount of the overall correction. L_3 adjusts its width.



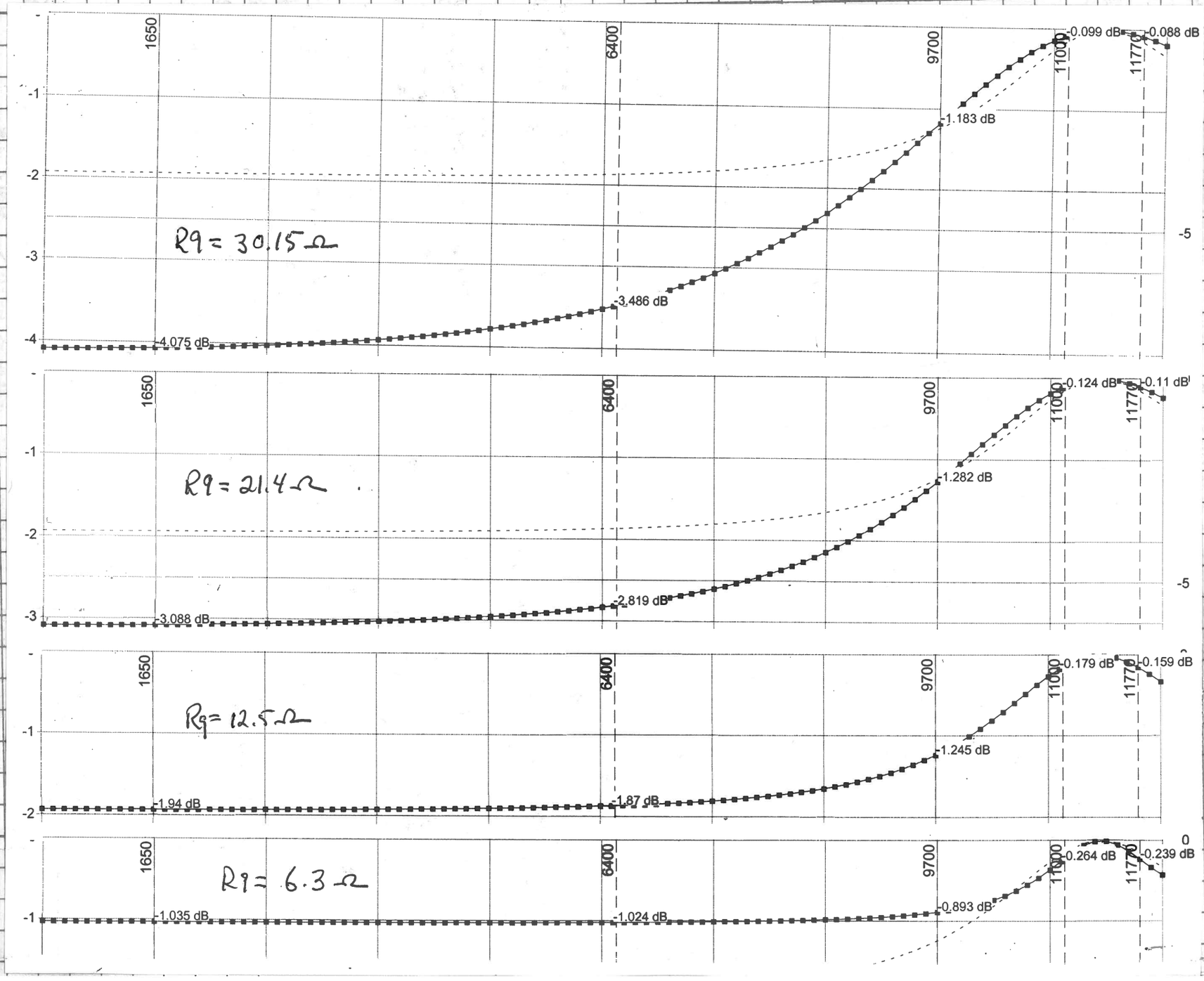


$R_9 = 125\,\Omega$
$L_3 = .657\,nh$
$C_2 = .298\,pf$
$R_7 = 0$

$L_2 = 2.5 \times C_2 = .745\,nh$
$C_3 = 2.5/L_3 = .263\,pf$
$R_8 = \infty$
$R_{10} = 208\,\Omega$

Test the Equalizer with $L_3 = .657\,nh$, $C_2 = .298\,pf$, $R_2 = 0$. Only R_9 (and $R_{10} = 2500/R_9$) is varied.
(fixed) $\frac{1}{2\pi\sqrt{L_3 C_2}} = 11.59\,GHz$

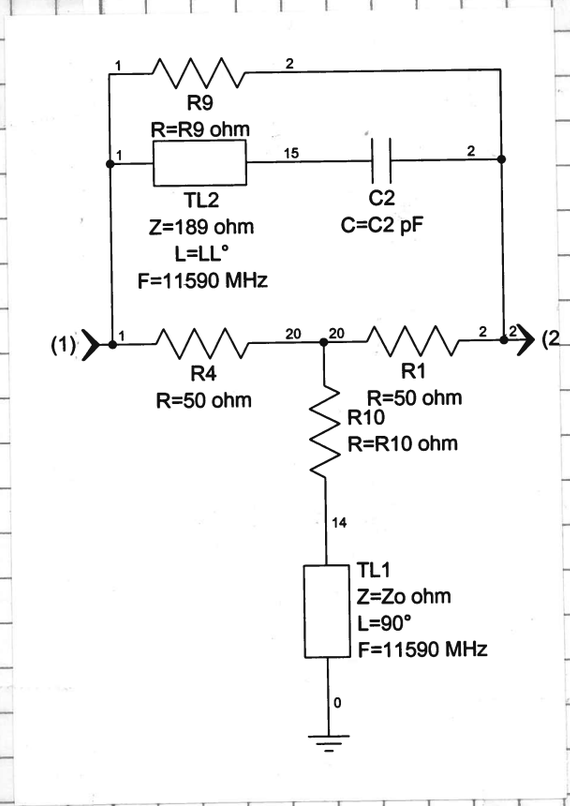
S_{21}



$S_{11} \approx -\infty$

freq

III. Now replace the inductors with lengths of transmission line and resonant circuits with quarter wave stubs.



$$L/\text{unit length} = \frac{Z_0\sqrt{K_e}}{v_0} = \frac{Z_0\sqrt{K_e}\times 10^{-10}}{1181 \text{ mils/sec}} \text{ nh}$$

$$C/\text{unit length} = \frac{\sqrt{K_e}\times 10^{-13}}{1.181\, Z_0} \text{ pf}$$

Obtain mostly inductance for large Z_0, mostly capacitance for small Z_0. The inductance can be in series; the capacitance will be to ground.

Example: Duroid with $K_e = 2.24$, width = 10 mils, $H = 62.5$ mils, $t = 0.2$ mils. Find $K_{eff} = 1.71$, and $Z_0 = 180\ \Omega$.
Then $L/l = .0196$ nh/mil of length.

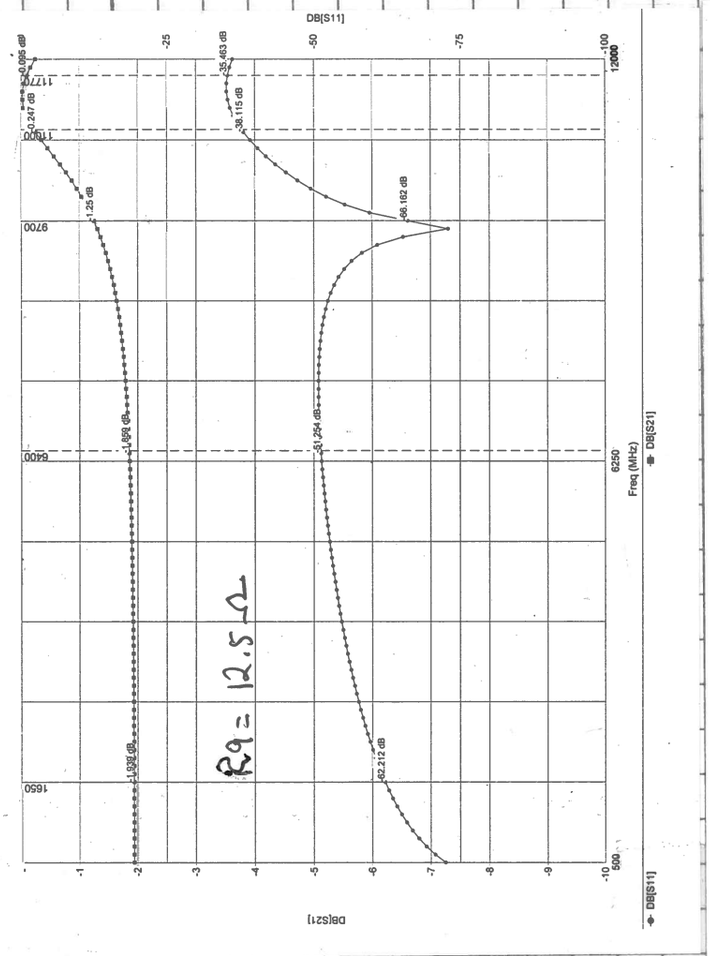
For $L_3 = 0.657$, $l = 33.4$ mils
At 11.6 GHz $\lambda = 2.0$ cm in the duroid; 33.4 mils $= .0424\lambda$ or 15° of phase. $L = LL° = 15°$, $Z = 180\ \Omega$ for TL2.

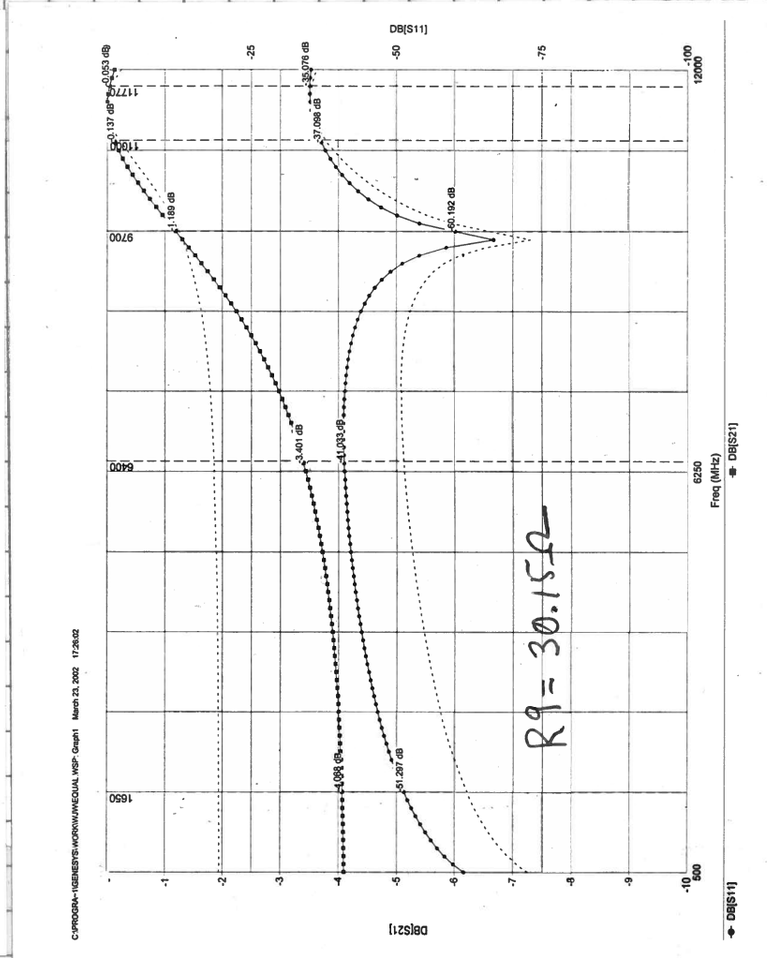
(a) The corresponding cap to ground is .02 pf.

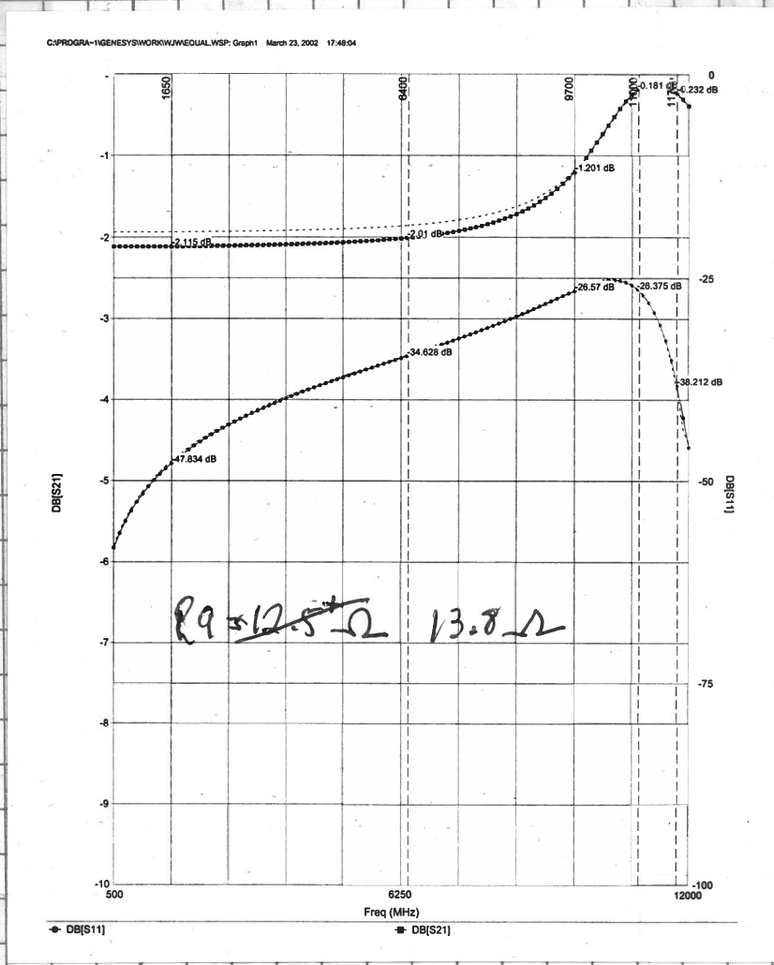
(b) For the tuned circuit in the shunt branch, use a grounded $\lambda/4$ stub. This is TL1. It should be 90° phase at 11590 MHz. Z_0 must be found by trial.

The top 2 figures on the next page show the result of replacing just the shunt branch, L2 and C3, with the grounded quarter wave stub. $Z_0 = 43.1\ \Omega$ to get an exact fit with the original curve. The match S11 is no longer perfect but it is better than −38 db which is good enough. Both the curve for R9 = 12.5 Ω and R9 = 30.15 agree with the lumped element S21 curves within 0.1 db.

In the bottom picture L_3 has been replaced with a length of transmission line of 189 ohms and phase length 15° (as computed). Varying the phase (line length) shows 15° to be the correct length. The match, S11, is now a little worse but ok.







There is now about 0.2db additional loss at 11.5GHz, probably due to the 0.02pf shunt capacitance (to ground) of TR2. Increasing R9 corrects for that, as shown in the figure.

Is the spiral antenna calibration source really needed? Or could one always use self calibration?

Self calibration 6m antenna, $T_{sys} = 50K$, $BW = 100\ MHz$

In 10 seconds $\Delta T = \frac{T}{\sqrt{B\tau}} = \frac{50}{\sqrt{10^9}} = .0016\ K$

For 145 Jy/K $\Delta S = 0.233$ Jy RMS, x3 for good phase $= 0.697$ Jy

Assuming the source is unresolved on all baselines, there will be 350 phase measurements for each antenna. $\sqrt{350} = 18.7$

The source strength must be $0.697/18.7 = .0373 = 37.3$ mJy

For both polarizations, $\div\sqrt{2} = 26$ mJy (best circumstances)

This is the necessary flux for a point source for a pt. source self-calibrator in 10 seconds.

The antenna beam width @ 10 GHz 1219" or $\frac{20.3'}{}$

Beams over the sky 41,000 sq degrees/$(20.3/60)^2$ = 357,000 = $\frac{4\pi}{\Delta\Omega}$

For the 3c catalogue, complete to 10 Jy at 178 MHz,

$$n(F > 26\,mJy, 10\,GHz) = \frac{\Delta\Omega}{4\pi}\, 16{,}200\, F^{-3/2} \left(\frac{\lambda}{1.7m}\right)^{3\alpha/2}$$

α is the typical spectral index $\alpha = .75$

$$n(F > 26, 10\,GHz) = \frac{16{,}200}{357{,}000}(.026)^{-3/2}\left(\frac{.03}{1.7}\right)^{1.13} = (.0454)(239)(.0104)$$

$$= 0.11$$

$\Delta\Omega \propto \lambda^2$ (diffraction)

Note $n \propto \nu^{-3.13}$ so that at 1.42 GHz, $n = 51$

so there's no problem for $\lambda 21$ cm.

The 0.11 source number is optimistic because it assumes that the sources are unresolved on all baselines, doubtful at 10 GHz where the resolution is ~ 8".

Conclusion The calibration source is needed for $\nu \geq 5$ GHz even without considering polarization. The addition of polarization makes the case stronger.

Final layout for the calibration module.

One Nitro amp + equalizer is ~ 12.8 db of gain

The transmission file (SP1) plus equalizer is − 33 db total.
The actual transmission loss at λ18 cm should be −33.0 db
whereas the SP1 transmission file is − 10.2 db at this wavelength.
So the actual overall loss is 23 db greater : − 56 db.

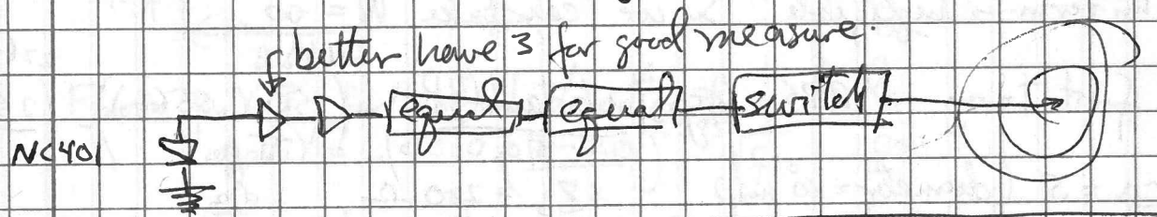
transmission loss (+equalizer)	− 56
antenna match	− 6
circular to linear	− 3
switch + equalizer	− 3
input transmission line + equalizer	− 10
	−78

If the noise diode is +30 db (300,000 K) and we need 1 K,
it's excess is 54.8 db NC401 from NoiseCom

78 − 54.6 = 23.4 db (just barely)

2 NITRO amps should suffice

If the 3.1 db switchable attenuator were used, or,
better yet, 2 in cascade, different calibration signals
would be available, with 2 attens for 60 db, there
should be little leakage noise.



$n(F>26) = 0.11\left(\frac{\nu}{40}\right)^{-3.13}$, $= 1$ when $(\nu/10)^{-3.13} = 9$ $(\nu/10)^{-1} = 2.02$ and $\nu = 5.0$ GHz

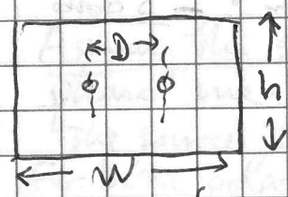
General formula for ATA confusion (continuum) one source/beam

$F(\mu Jy) = 11.6\ \lambda^{2.1}(m)$ μJy // $12.9\ \lambda^{2.1}(m)$ μJy 4/20/08

Design of the Input region of the feed pyramid.

Characteristic impedances and losses taking the finite dimensions of the wall into account.

A formula exists for the impedance of a two wire line (single) in a box. This would apply to the case of having a septum.



wire radius is a, diameter 2a

[S. Frankel, 1942, Proc IRE 30, no. 4] also Marchand, 1947, Wiley appendix

$$Z_0 = 276\left\{\log_{10}\left[\frac{2h\tanh\left(\frac{\pi D}{2h}\right)}{\pi a}\right] - \log_{10}\left[\frac{1+\left(\frac{\sinh\left(\frac{\pi D}{2h}\right)}{\cosh\left(\frac{\pi W}{2h}\right)}\right)^2}{1-\left(\frac{\sinh\left(\frac{\pi D}{2h}\right)}{\sinh\left(\frac{\pi W}{2h}\right)}\right)^2}\right]\right\}$$

For $W = 150$ and $h = 75$, it's the ATA case with a septum.

Look at the second term with $D = 30$. $\frac{\pi D}{2h} = \frac{\pi(30)}{150} = .6283$; $\frac{\pi W}{2h} = \pi$

$\sinh .6283 = .6705$, $\cosh \pi = 11.5967$, $\sinh \pi = 11.5535$

$$-\log_{10}\left[\frac{1+\left(\frac{.6705}{11.5967}\right)^2}{1-\left(\frac{.6705}{11.5535}\right)^2}\right] = \log_{10}\left[\frac{1+.0033}{1-.0034}\right] = \log_{10}\left(\frac{1.0033}{.9966}\right) = .0029 \times 276 = .80$$

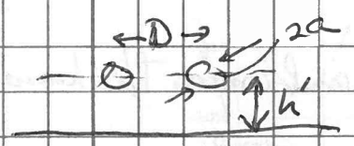
This term is negligible. So we can take $W = \infty$.

The first term $276\log_{10}\left(\frac{2h\tanh\left(\frac{\pi D}{2h}\right)}{\pi a}\right) = 276\log_{10}\left(\frac{150(.5569)}{\pi a}\right) = 276\log_{10}\left(\frac{26.590}{a}\right)$

For $a = 5$ (diameter = 10 mils) $Z_0 = 200\,\Omega$ ok

Note $\tanh .6283 = .5569$ close to the limiting free space value.

$\Rightarrow$ Since $W = \infty$ is ok and there will be no septum, it looks like two wires next to a ground plane should be an adequate model.



$$Z_0 = 276 \log_{10} \frac{D}{a}\left[1+\left(\frac{D}{2h'}\right)^2\right]^{-1/2}$$

I. At the input, $h' = h/2 = 37.5$
With $D = 50$ $\left[1+\left(\frac{50}{75}\right)^2\right]^{-1/2} = 0.83$

For 200 Ω $D/a\,(.83) = 5.3$ $D/a = 6.37$ and $a = 7.8$ mils
for a diameter of 15.7 mils For $D = 20$ here $Z_0 = 170\ \Omega$

This may be a good choice at the input because it will spread field lines out for a better transition to the descrambler board.

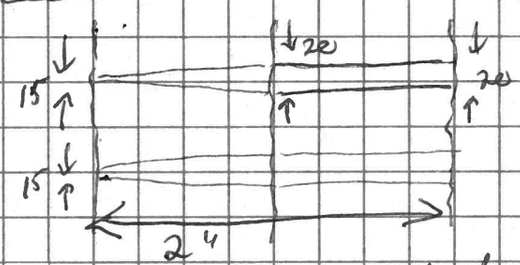
II. At a distance of 1" The lines will be at $\frac{350}{4} = 88$ mils $= h'$

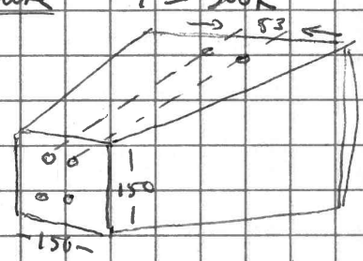
$\left[1+\left(\frac{50}{175}\right)^2\right]^{-1/2} = 0.96 \cong 1.$ It's effectively free space.

$Z_0 = (276) \log_{10} D/a\,(.96)$ $D/a\,(.96) = 5.30$

For $D = 50$, $2a = 18$ For $2a = 20$ mils, $D = 53$ mils

III For a 2" run at room temperature $T = 300K$





From p. 68 for 20 mil diameter twin lead:

$$\alpha_p(T) = .0010\left(\frac{T-20}{132}\right)^{.675}\sqrt{f_{10}} \quad cm^{-1}$$

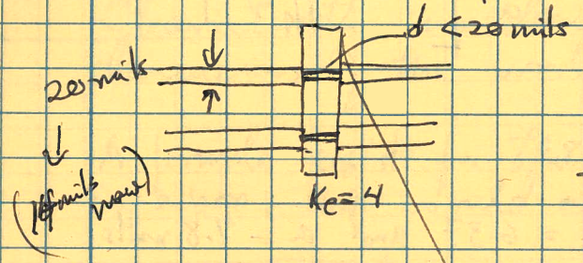
a) In the first inch $\langle D\rangle = 18$ mils. $\alpha_p(300) = .0018\sqrt{f_{10}}\ cm^{-1}$
For 2.5 cm (1") $\Delta T = 300\,(.0018)\sqrt{f_{10}} \times 2.5 = 1.35\sqrt{f_{10}}\ K$

b) In the second inch $\alpha_p(300) = .0017\sqrt{f_{10}}$ $\Delta T = 1.25\sqrt{f_{10}}$

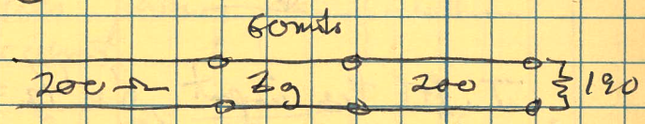
The total is $2.6\sqrt{f_{10}}\ K$

Reflections at the window of the lever

assume $K_e = 4$ for the window glass and window thickness = 60 mils



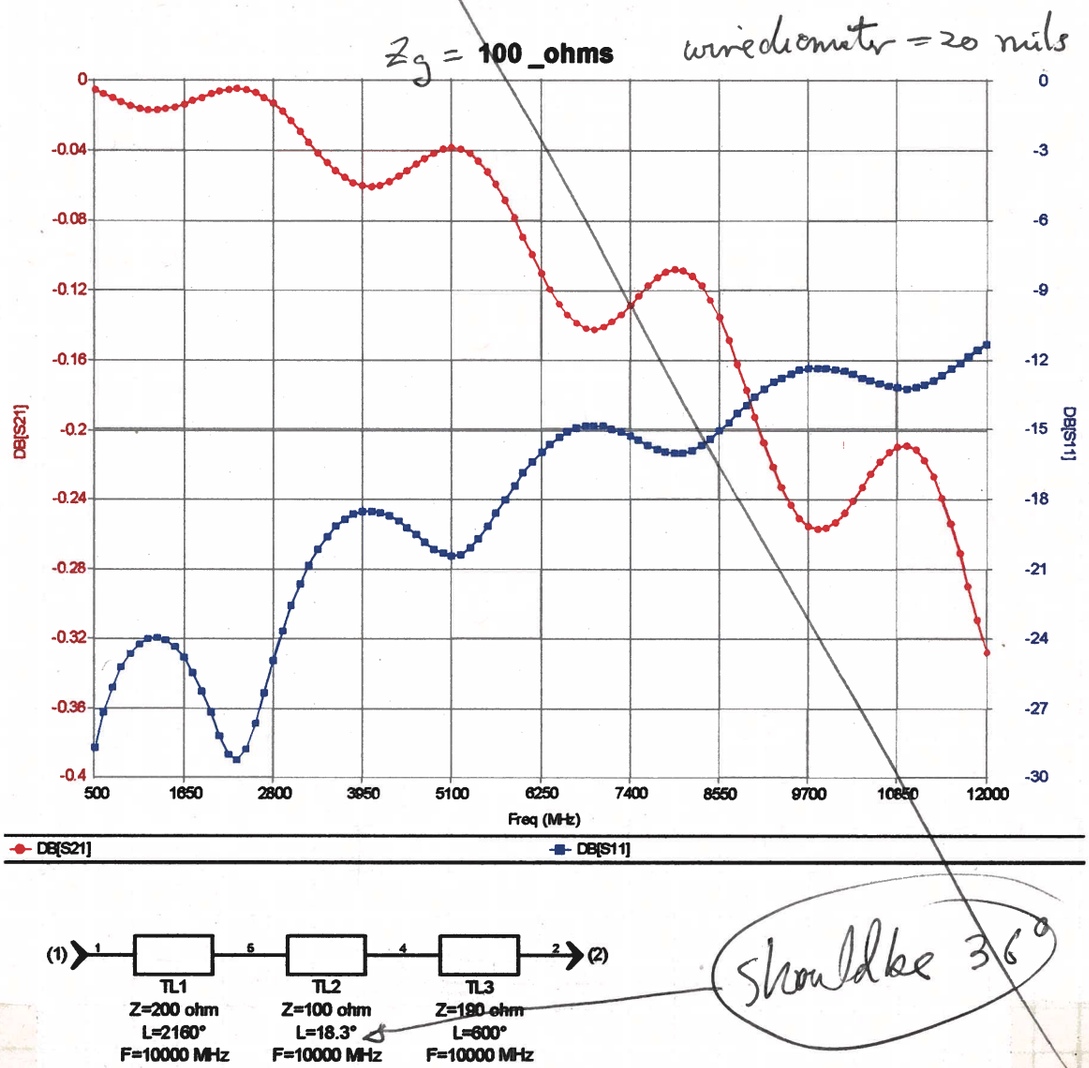
Model



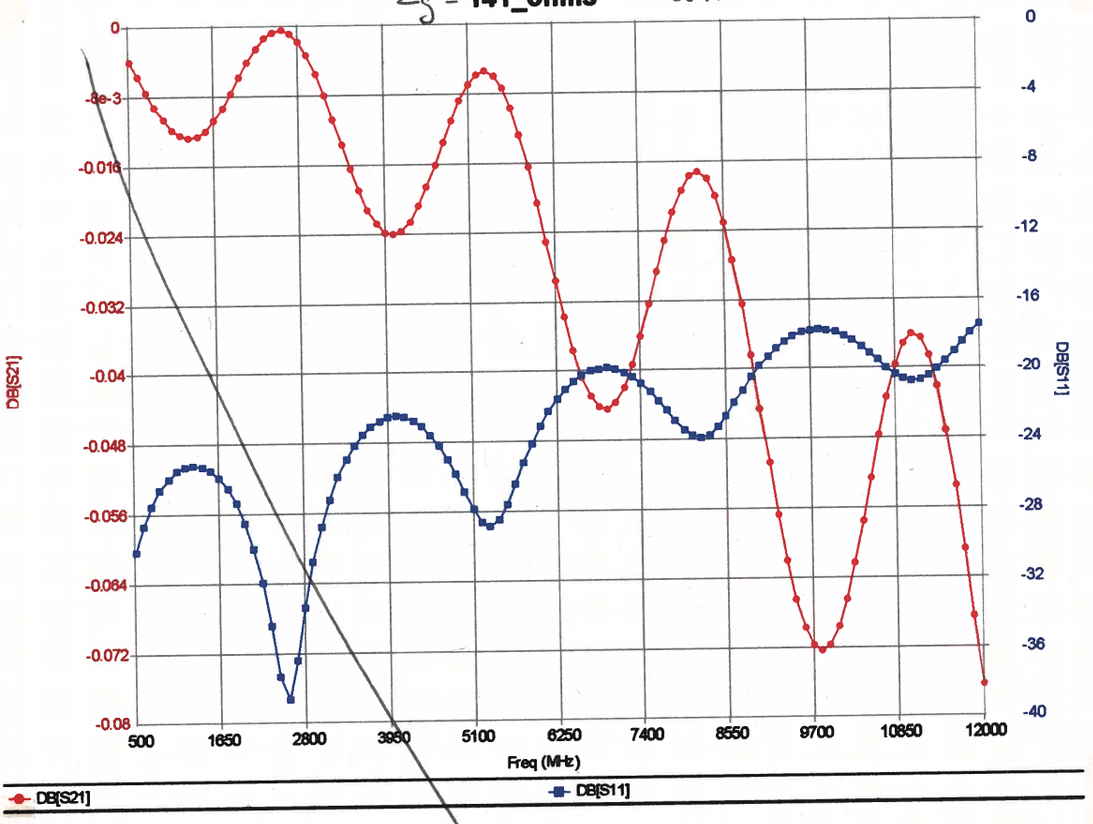
The 120 Ω is a small mismatch at the output as might be expected.

To get to 200 Ω in the glass requires a wire diameter of ~1 mil.

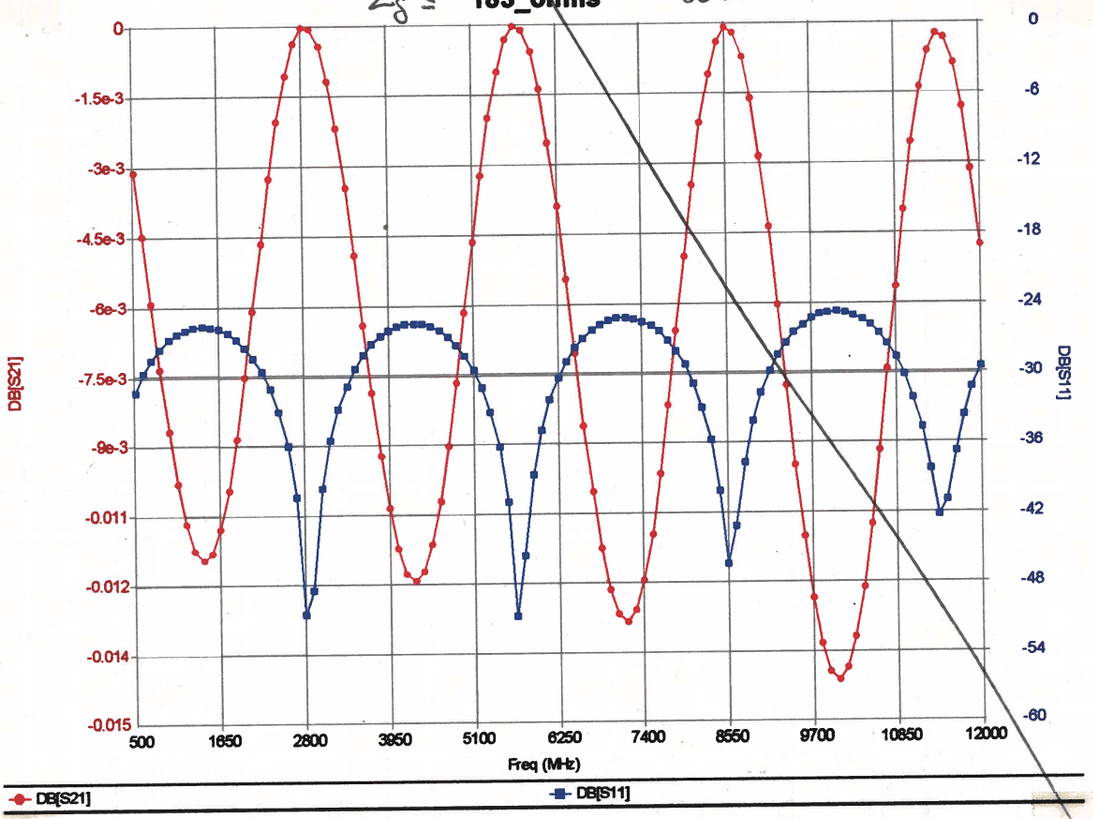
The following show the reflections with @ d = 20 (no change), ⓑ d = 10 mils and d = 5 mils



Z_g = **141_ohms** wire diameter = 10 mils

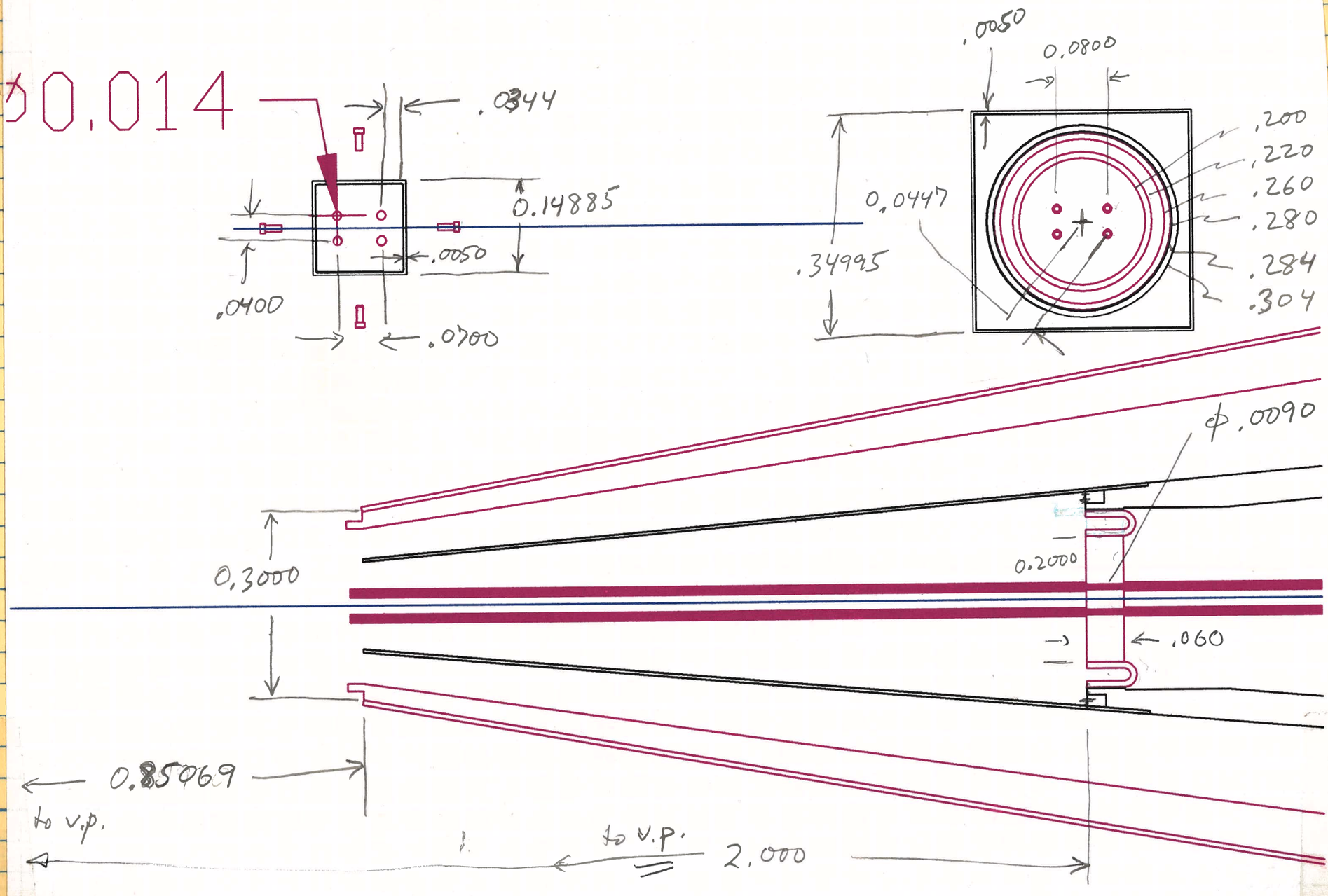


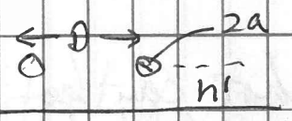
Z_g = **183_ohms** wire diameter = 5 mils



Final Input Layout.

5 in = 1 in





Impedances and Losses

$$Z_0 = 276 \log_{10} \frac{D}{a}\left[1+\left(\frac{D}{2h'}\right)^2\right]^{-1/2}, \text{ for } h'=\infty;\ 276\log_{10}\left(\frac{40}{7}\right) = 209$$

1. input plane: $\underline{D = .040}$ $\underline{2a = .014\text{ in}}$, $h' = .034$ $\quad \sqrt{1+\left(\frac{40}{68}\right)^2}^{-1} = .86$

$$Z_0 = 276 \log_{10}\left(\frac{.040}{.007}\right) .86 = \underline{190\ \Omega}$$; for $a = .0042$

$$276 \log_{10}\left(\frac{.040}{.0042}\right) = 270\Omega$$

$$r = \frac{209-270}{209+270} = -.12$$

2. At the glass seal $\sqrt{1+\left(\frac{40}{135}\right)^2}^{-1} = .96$

$$Z_0 = 276 \log_{10}[(5.714)(.96)] = \cancel{\ } 204$$

3. Losses and noise (p. 68) $\quad R = \frac{2R_s}{\pi 2a}\frac{D/2a}{\sqrt{\left(\frac{D}{2a}\right)^2-1}} = \frac{R_s}{\pi}\frac{(2.86)}{(.014\text{ cm})\sqrt{(2.86)^2-1}}$

$$= 18.9 R_s$$

$$\alpha_p = R/Z_0 = 18.9 R_s/204 = .093 R_s$$

$$= .0015\left(\frac{T-20}{132}\right)^{.625}\sqrt{f_{10}}\ \text{cm}^{-1}$$

At 300K and with length $l = 1.15'' = 2.92$ cm

$$\alpha_p(300)l = .0015\left(\frac{300-20}{132}\right)^{.625}\sqrt{f_{10}}\ 2.92 = .0073\sqrt{f_{10}}$$

Total noise: $\Delta T_n = 300\,(.0073\sqrt{f_{10}}) = \underline{2.2\sqrt{f_{10}}\ K}$

This looks fine $\quad \rho = \frac{1-\frac{190}{204}}{1+\frac{190}{204}} = .0355 \quad |\rho|^2 = -29\text{ db}$

Reflection from the glass seal. $t = .060''$ (1.5 mm), $l = 36°$ @ 10 GHz

dr → For $D = .040$, and $2a = .009$, $Z_g = 276/\sqrt{4}\ \log_{10}\left(\frac{40}{4.5}\right) = 131\ \Omega$

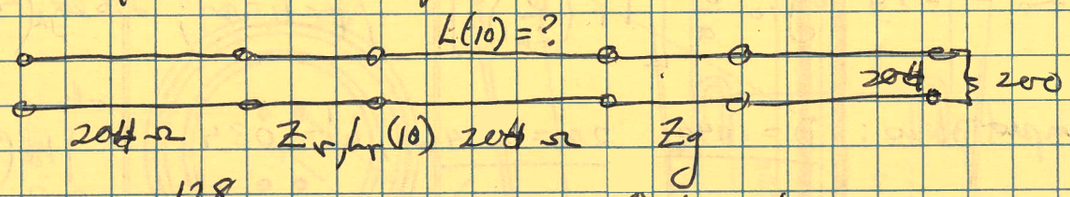
This gives $S_{11} = -11$db and $S_{21} = -.38$db at 12 GHz

For $2a = .007$ $\quad Z_g = 146$, $S_{11} = -13$db and $S_{21} = -.21$db at 12 GHz

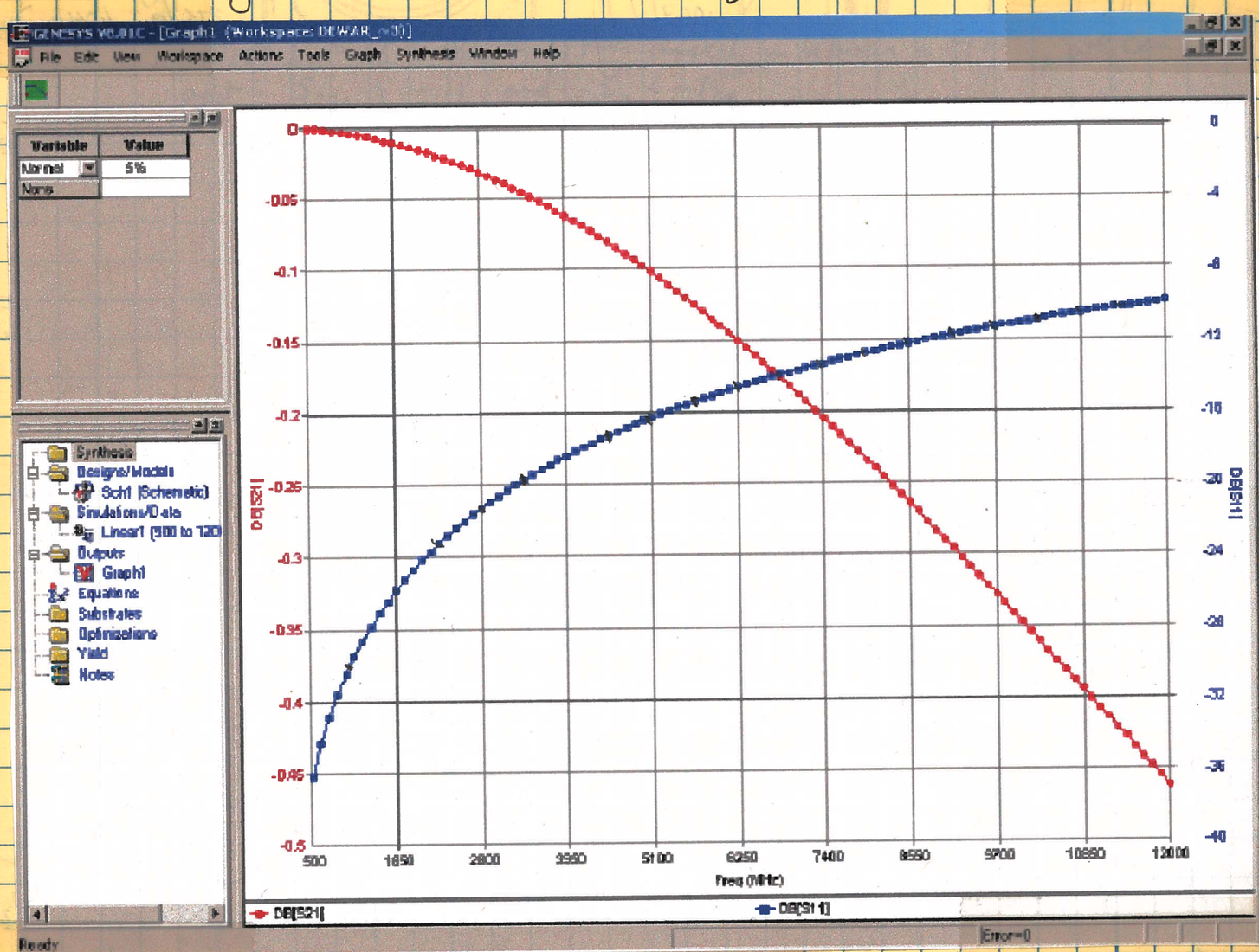
For $2a = .005$ $\quad Z_g = 166$, $S_{11} = -18$db, $S_{21} = -.07$ at 12 GHz

(NB!) watch out for the $\sqrt{1+\left(\frac{D}{2h'}\right)^2}$ factor; it's 0.96

The echo can be reduced by putting a piece of reflecting dielectric upstream of the glass seal.



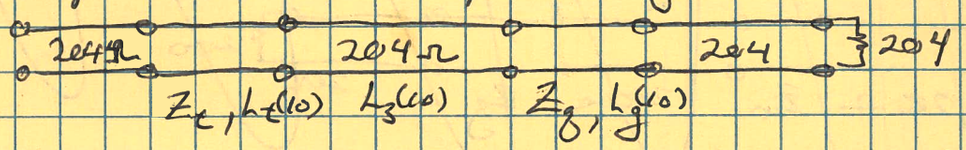
$Z_g = \frac{128}{151}\Omega$ $L(10\,GHz) = 36°$ for $t = .080$ inch



The above is just the echo from the glass window. S_{21} gets up to −.45db at 12 GHz.

5/19/02

Placing a thin Teflon window in front of the glass window at some distance along the transmission line can partially cancel the echo from the glass. $k_\epsilon = 2.0$ for Teflon.



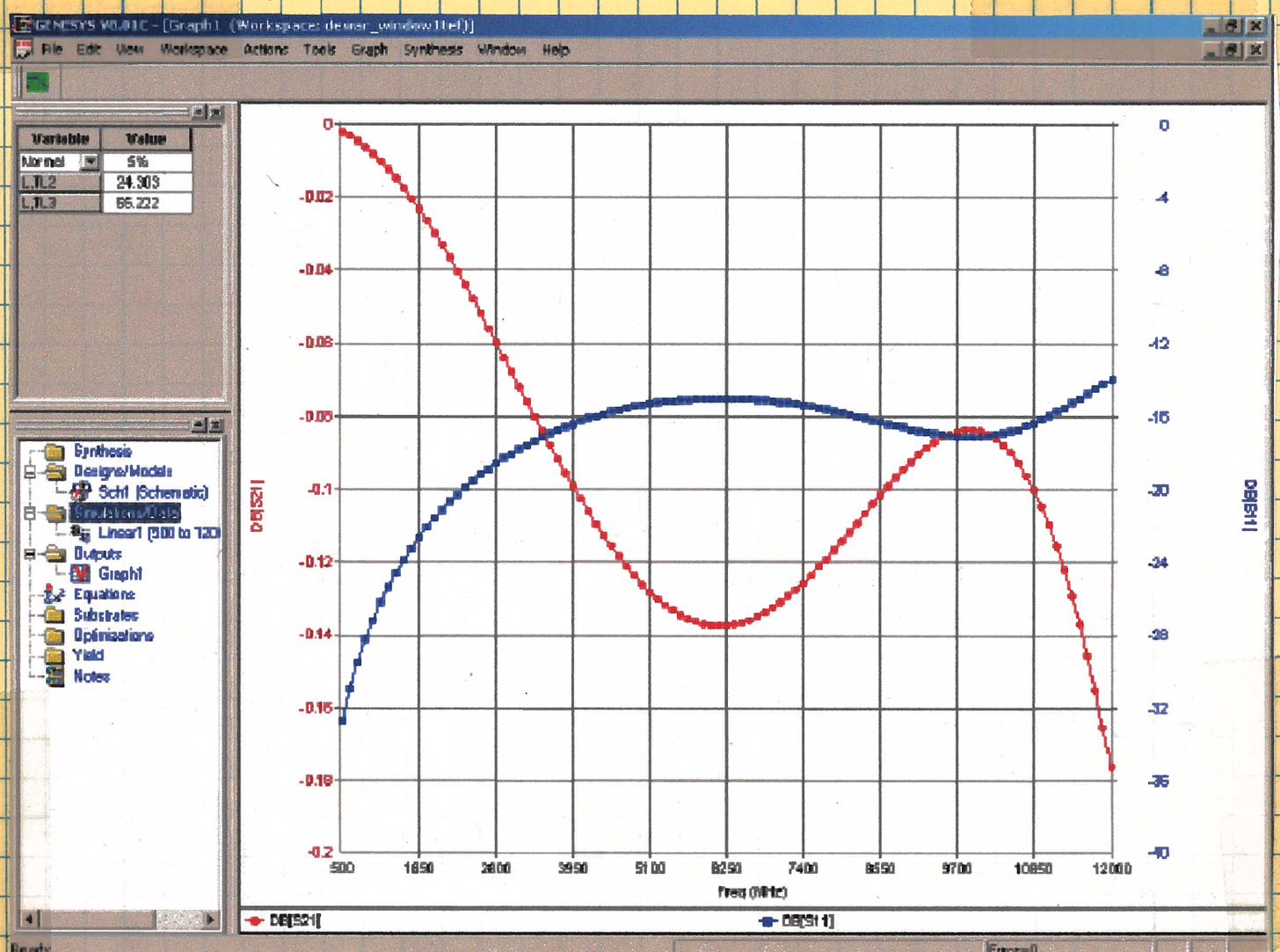
$Z_g = 128\,\Omega$ $\quad L_g(10\,GHz) = 36.6°$ for $t = 60$ mils

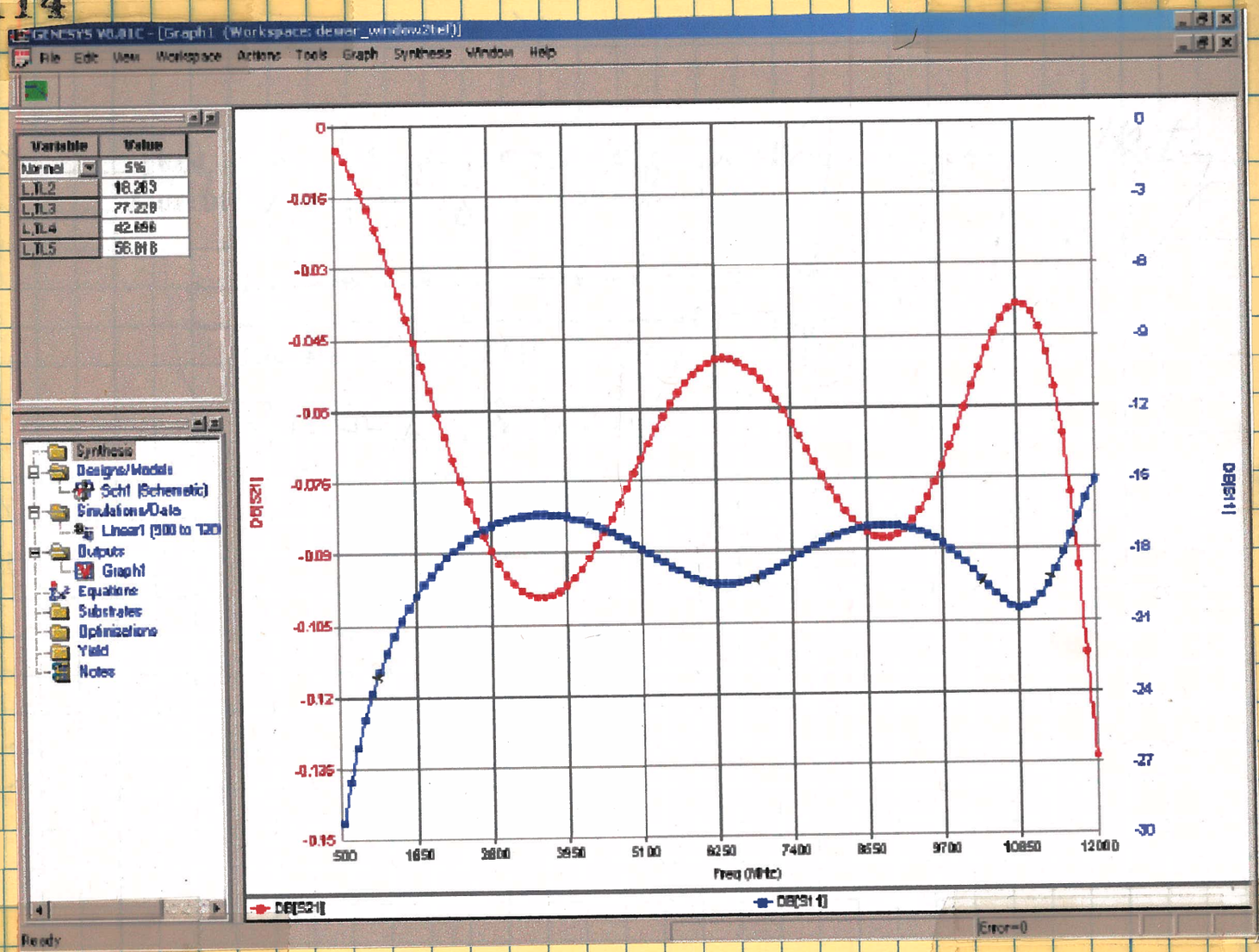
The teflon is on the two wire line with $D = 40$ mils and $2a = 14$ mils. Thus $Z_t = 204/\sqrt{2.0} = \underline{144\,\Omega}$. To get a good match, vary the teflon thickness $L_t(10)$ and the distance to the glass.

The best overall match is for $\underline{L_t(10) = 24°}$ and $\underline{L_3(10) = 55°}$

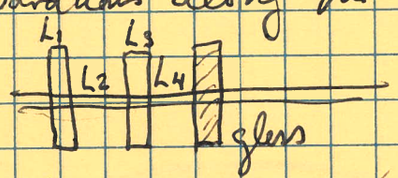
Teflon thickness: $\frac{24°}{360°} \times \frac{3.0\,cm}{\sqrt{2}} = 0.14\,cm = \underline{56\,mils}$

Spacing L_3 $\frac{55}{360} \times 3.0 = .458\,cm = \underline{180\,mils}$ (One teflon window)





Now add a second window; vary the two window thicknesses and their separations along the line.



$L_1 = 18°$ $\left(18/360\right)\frac{3cm}{\sqrt{2}} = .106\,cm = 42\,mils = 1\,mm$

$L_2 = 77°$ $\left(77/360\right)\times 3.0 = .642\,cm = 253\,mils$

$L_3 = 43°$ $\left(43/360\right)\times\frac{3cm}{\sqrt{2}} = .253\,cm = 100\,mils = 2.5\,mm$

$L_4 = 58°$ $\left(58/360\right)\times 3.0 = .483\,cm = 190\,mils$

S_{21} and S_{11} are shown above. Match is ~ -18db

Is sheet teflon available? } L_2 and L_4 can be cut from polyfoam or bead board

RFI levels from the BIMA Correlator

Geoff and Dave have measured RFI from the correlator - at Lab #1. The distance is about 20 m, about the expected nearest ATA antenna.

The strongest signal $\lesssim 1000$ MHz is at 698 MHz (Geoff) measured on the spectrum analyzer connected to the disk-cone antenna through an amplifier and cable. ≈ -41.5 dbm.

The LNA and cable are $+43.0$ db gain.
$-(43.0 + 41.5) = -84.5$ for the signal. (Dave)

The maximum that would be received by an ATA antenna would be if it were pointed to $E = +15°$ toward the building - at a distance of 20 m
At 700 MHz $B_{1/2} \sim 4.2°$ and $15°$ is 3.6 beam widths.
From the pattern of p. 12 II, the peak gain is -4 db,
[The disk-cone gain should be ~ 1.5 (1.8 db)]; down by -5.8 db total.

$$\begin{array}{r} -84.5 \\ -5.8 \\ \hline -90.3 \text{ dbm} \end{array}$$

With -120 db for the shielded room $\quad -210.3$ dbm

Noise power: kTB for $B = 1$ Hz and $T = 40$ K is
$+1.4\times10^{-23}(40) = 5.6\times10^{-22}$ watts
$= 5.6\times10^{-19}$ mw
$= -182.5$ dbm

The margin is $210.3 - 182.5 = \boxed{27.8 \text{ db}}$

In the beam formers there will be further help from the finite fringe rates.

$$\phi \sim \frac{2\pi S}{\lambda}\cos\delta\cos h \qquad \nu_f = \frac{1}{2\pi}\frac{d\phi}{dh} = \frac{S}{\lambda}\sin h\,\frac{dh}{dt}\cos\delta$$

$$\nu_f = \frac{S}{\lambda}\cos\delta\sin h\,\frac{2\pi}{24\cdot3600} = \frac{S}{\lambda}\cos\delta\cos h\;7.2\times10^{-5}\text{ rad/sec.}$$

for $\overline{\cos\delta \cos h} \sim \frac{1}{2}$ $\qquad \nu_f = \frac{s}{\lambda} \, 3.6\times10^{-5}$

$\langle s \rangle \sim 350\,m$ take $\lambda = .60$, the lowest frequency

$$\nu_f \sim \frac{350}{.6} \cdot 3.6\times10^{-5} = .025\,Hz = \underline{\underline{25\,mHz}}$$

<u>Effect of the number of wraps.</u>

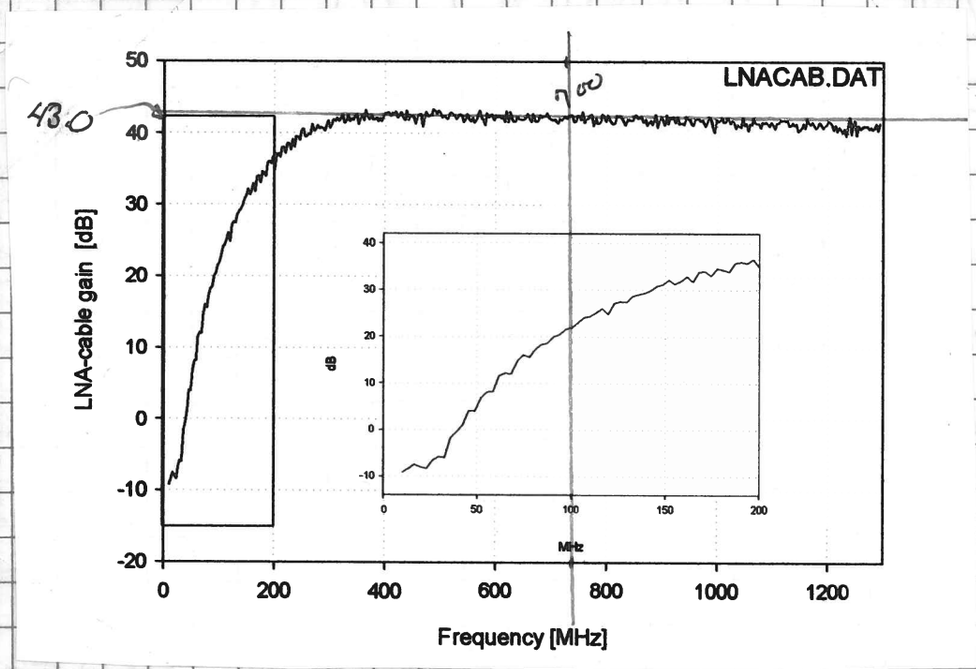
For an integration time of 300 seconds:

$300 \times .025 = 7.5$ cycles.

$$\langle \cos t \rangle = \frac{1}{7.5\,(2\pi)} \int_0^{2\pi(7.5)} \cos t \, dt = \frac{1}{7.5\,2\pi} \sin t \Big|_0^{2\pi 7.5}$$

$$\lesssim \frac{1}{7.5\,2\pi} = .021 \qquad \lesssim \underline{\underline{-16\,db}}$$

This is an additional help in reducing the RFI.

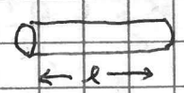


Thermal Loads in the Dewar

Radiative load from the ambient wall. Assume perfect black body radiation at $T = 300K$, a little pessimistic since some angle is taken up by the cold end of the dewar.

$$F = \sigma T^4 = 5.6\times10^{-5}\ \text{ergs cm}^{-2}\text{K}^{-4}\text{sec}^{-1}\ (300)^4 = 4.54\times10^{+5}\ \text{ergs/cm}^2\text{sec}$$
$$= 4.5\ \text{watts/cm}^2 = \underline{45\ \text{mwatts/cm}^2}$$

1. For the 14 mil wire: $.014 \times 2.54 = .0356$ cm.
   A one inch length



$A_{wire} = \pi\,\text{diam} \times \ell$
$= .28$ mwatt, for $\ell = 1''$

This is small compared to the expected heat flow of ~ 7-8 mwatts in each wire. So shielding the wires from radiation seems unimportant.

2. Heat flow in the input glass window

First: can the input leads keep the temperature up at the glass seal? Use solid copper for the 14 mil leads.
Each lead will carry 7 mw into the dewar

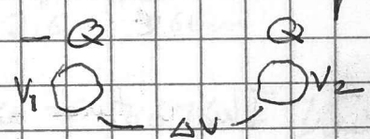
$A_{cu} = 10^{-7}\ m^2$
$k = 396$ w/m-k at 300k
$\Delta x = 1.15'' = .029$ m

$$H = kA\frac{\Delta T}{\Delta x} \quad \text{so} \quad \Delta T = \frac{H\,\Delta x}{kA} = \frac{(.007)(.029)}{(396)(10^{-7})} = \underline{5.1\ K}$$

That may be ok.

Now, what is the effect of the glass seal. One wire near the center with heat flow to the edge of the seal is like the potential analogue for the coaxial line. Find the connection between the capacitance for a length of coaxial line. The offset case is in Genesys.

$$\nabla^2 T = 0 \leftrightarrow \nabla^2\Phi = 0$$



$C = Q/\Delta V$

$$\int dH = \iint k\,dA\,\nabla T\cdot\hat{n} \qquad \text{i.e.}\quad H = \iint k\,\nabla T\cdot\hat{n}\,dA$$

same equations for ϕ and T, and same boundary conditions. The distributions should be the same except for a scale factor.

$$\alpha\Phi = T$$

$$Q = \iint \epsilon \vec{E}\cdot\hat{n}\,ds = -\iint \epsilon \nabla\phi\cdot\hat{n}\,ds \qquad C = \frac{Q}{V_1 - V_2} = \frac{Q}{\Delta V}$$

$$H = \iint K\nabla T\cdot\hat{n}\,ds = \iint K\nabla(\alpha\Phi)\cdot\hat{n}\,ds = \frac{K\alpha}{\epsilon}\iint \epsilon\nabla\phi\cdot\hat{n}\,ds = \frac{K\alpha}{\epsilon}Q$$

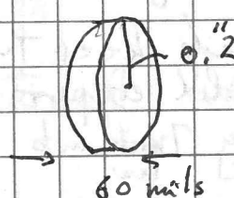
$$H = \frac{\alpha K}{\epsilon}[C\Delta V] = \frac{\cancel{\alpha} K}{\epsilon} C \frac{\Delta T}{\cancel{\alpha}} = \frac{KC}{\epsilon}\Delta T$$

or $\underline{H = G_T \Delta T}$ where $\underline{G_T = \frac{KC}{\epsilon}}$

For the straight wire $G_T = KA/\Delta x = 396\times10^{-7}/.029 = 1.36\times10^{-3}\ w/mm^2$

$G_T = 1.36\ mwatts/^\circ K$

For the glass seal



$R_{outer} = 0.2''$

$2R_{inner} = .01''$

Approximate with a simple capacitance formula with wire in the center

$$Z_0 = 138\log_{10}(R_o/R_i) = 138\log\left(\frac{200}{5}\right) = 221\Omega$$

$$v^2 = \frac{1}{LC},\quad Z_0^2 = \frac{L}{C} = \frac{1}{C}\cdot\frac{1}{v^2C} = \frac{1}{C^2v^2} \quad \text{or} \quad Z_0 = \frac{1}{Cv},\quad \boxed{C = \frac{1}{Z_0 v}}$$

$$C = \frac{1}{(221)(3\times10^8\ m/sec)} = 15\ pf/m$$

60 mils $\Rightarrow$ .0015 m

$$C = (15)(.0015\ m) = .0226\ pf$$

Now with $K = 1.1$ watts/m-k and $\epsilon = 10^{-9}/36\pi$

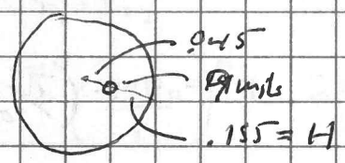
$$G_T = \frac{KC}{\epsilon} = (1.1)(.022\times10^{-12})\frac{36\pi}{10^{-9}} = .0028\ watts/^\circ K = \underline{2.8\ mw/^\circ K}$$

This is twice as good as the solid copper 1.15" wires

With 1 mwatts in the wire into the dewar there will be a 2.5K Temperature drop from the wall to the center

With four wires each with the same load, the total drop should be the same

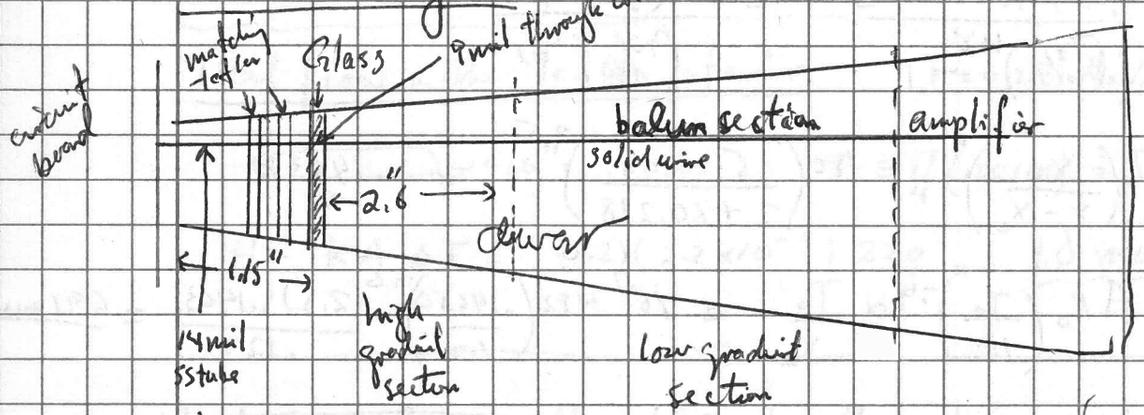
Geneep has a model for Z_0 for an off center conductor. It's about the same.



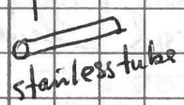
Z_0 goes to 218 Ω in this case instead of 221 Ω

~~ΔT goes to ~~ ΔT goes to 2.48K.

Front End Layout



use bead board between teflon pieces



stainless tube $\begin{cases} 14 \text{ mils od} \\ 10 \text{ mils id} \end{cases}$ $t_{cu} = .15$ mil

High gradient region $\overline{KA(SS)} = 2.4\times10^{-6} \sim$ constant $\frac{W\text{-}m}{K}$ (JL)

Small gradient region $K = 482\left(T/100\right)^{-.6}$ $= 482$ @ 100K, $= 551$ @ 80K for copper

i) no temp. drop over the input 1.15" outside the dewar
ii) 300K − 90K over the 2.6" length of stainless tube .066 m
iii) 90 − 80K over the solid wire assume 3". ?

Region (i): $KA = 2.4\times10^{-6}\ \frac{W\text{-}m}{K}$ 2.6" = .066 m

$$H = KA\frac{\Delta T}{\Delta x} = (2.4\times10^{-6})\ 210/.066 = .0076 = 7.6 \text{ mwatts}$$

NB: For 2.0" instead of 2.6" H ~ 10 mwatts

(ii) over the 3" region with 90°–80° drop.

$T_1 = 80$ $\quad T_2 = 90$

$$k = 482\left(\frac{T}{100}\right)^{-.6} \qquad \gamma = -.60$$

$x_1 = 2$ $\quad x_2 = 5 = .127\,m$

(p. 64)

$$T = T_0\left(\frac{x - x_0}{x_2}\right)^{\alpha} \qquad \alpha = \frac{1}{1+\gamma} = 2.5 \qquad 1+\gamma = 0.4$$

$$x_0 = \frac{x_1\left(T_2/T_1\right)^{1+\gamma} - x_2}{\left(T_2/T_1\right)^{1+\gamma} - 1} = \frac{2\left(\frac{90}{80}\right)^{.4} - 5}{\left(90/80\right)^{.4} - 1} = -60.238''$$

$$T_0 = T_1\left(\frac{x_2}{x_1 - x_0}\right)^{\alpha} = 80\left(\frac{5}{2 + 60.238}\right)^{2.5} = .1463\,k$$

$$H = A k_0 \left(\frac{T_0}{100}\right)^{-.6} \alpha \frac{T_0}{x_2} = 10^{-7}\, 482\left(\frac{.1463}{100}\right)^{-.6} \frac{(2.5)(.1463)}{.127} = \underline{6.97 \text{ milliwatts}}$$

<u>Thermal calculations for the output coaxial lines</u>

<u>Precision Tube</u> makes coaxial cable with pure stainless in the outer jacket and teflon dielectric

JS ~~RM~~ <u>50085</u> has outer wall OD of .085 and dielectric OD of .066. The stainless wall has 10 mil thickness. The stainless is 304 for which $k(T) = 15(300/T)^{.86}$ (p.66)

<u>The temperature distribution</u> over ~~an~~ 4" (10cm) length:

$T_1 = 80$ $\quad T_2 = 300$

$$x_0 = \frac{x_1\left(T_2/T_1\right)^{1+\gamma} - x_2}{\left(T_2/T_1\right)^{1+\gamma} - 1} = \frac{-10}{6.888 - 1} = -1.698$$

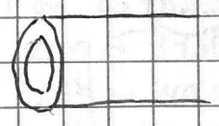
$x_1 = 0$ $\quad x_2 = 10\,cm$

$$T(x) = T_0\left(\frac{x - x_0}{x_2}\right)^{\alpha} \qquad \gamma = .46 \qquad \alpha = \frac{1}{1+\gamma} = .685 \qquad 1+\gamma = 1.46$$

$$T_0 = T_1\left(\frac{x_2}{x_1 - x_0}\right)^{\alpha} = 80\left(\frac{10}{0+1.698}\right)^{0.685} = 269.515$$

$$T(x) = T_0\left(\frac{x - x_0}{x_2}\right)^{\alpha} = 269.515\left(\frac{x+1.698}{10}\right)^{0.685} = 80.000 \text{ at } x=0$$
$$= 360.08 \text{ at } x = 10\text{ cm}.$$

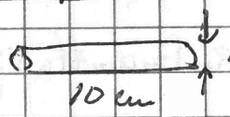
$$H = AK_0\left(\frac{T_0}{T_u}\right)^{\gamma}\frac{\alpha T_0}{x_2(m)} = (1.526\times10^{-6})\,15\left(\frac{269.515}{300}\right)^{.46}\frac{(.685)(269.515)}{.1} = \underline{\underline{40\text{ mw}}}$$



$r_0 = 0.43'' = .00109\text{m}$ $\qquad A = \pi\{(.00109)^2 - (.000838)^2\}$

$r_i = .033'' = .000838\text{ m}$ $\qquad = 1.526\times10^{-6}\text{ m}^2$

<u>Heat flow in the teflon interior</u> $\qquad k = 0.2$ watts/m-K for teflon



$.066'' = .00168 \qquad A = \frac{\pi}{4}(.00168)^2 = 2.2\times10^{-6}\text{ m}^2$

$$H = KA\frac{\Delta T}{\Delta x} = (0.2)(2.2\times10^{-6})\frac{220}{.1} = \underline{1.0\text{ mwatt}}$$

Heat Flow for 6'' cable lengths of 086 cable (SS)

$T_1 = 80 \qquad T_2 = 300 \qquad \alpha = .685,\ 1+\gamma = 1.46$

$x_1 = 0 \qquad x_2 = 15\text{cm}$

(a) <u>Outer conductor</u>

$$X_0 = \frac{x_1\left(T_2/T_1\right)^{1+\gamma} - x_2}{\left(T_2/T_1\right)^{1+\gamma} - 1} = \frac{-15}{6.888-1} = -2.548$$

$$T_0 = T_1\left(\frac{x_2}{x_1 - x_0}\right)^{\alpha} = 80\left(\frac{15}{2.548}\right)^{0.685} = 266.86$$

$$T(x) = T_0\left(\frac{x - x_0}{x_2}\right)^{\alpha} = 266.86\left(\frac{x+2.548}{15}\right)^{0.685} = 79.23 \text{ at } x=0$$
$$= 297.14 \text{ at } x=15$$

$$H = AK_0\left(\frac{T_0}{T_u}\right)^{\gamma}\frac{\alpha T_0}{x_2(m)} = (1.526\times10^{-6})(15)\left(\frac{266.86}{300}\right)^{.46}\frac{(.685)(266.86)}{.15}$$

$$= \underline{26\text{ mwatts}}$$

(b) Teflon $H = KA\frac{\Delta T}{\Delta x} = (.2)(2.2\times10^{-6})\frac{220}{.15} = \underline{0.64\,mW}$

(c) inner conductor: diameter = ~~1.676~~ .510 mm = ~~1.676~~ .510 $\times10^{-3}$ m
Beryllium Copper. Silver plated $A = (\pi/4)d^2 = \cancel{2.2\times10^{-6}\,m^2} = 2.0\times10^{-7}\,m^2$
[TS50085]

K For Beryllium Cu (p.TM) is ~ 400 mwatts/cm-K @ 100K
= 40 watts/m-K ✓

$H = KA\frac{\Delta T}{\Delta x} = (40)(2.0\times10^{-7})\frac{220}{.15} = \underline{11.7\,mwatts}$

This should dominate the silver plate
Relative area of Δr thickness plating: $\frac{2\pi r\Delta r}{\pi r^2} = \frac{2\Delta r}{r}$

☆ Total heat flow is 20 + 0.6 + 11.7 = 33.3 mwatts for the 6" cable.

Calibration with a transmitter at the rim lookout tower.
Distance is ~ 3 km
For field distance: $2D^2/\lambda = 2(6m)^2/.2m = 360\,m = .360\,km$ @ $\lambda = 20cm$
$= 2.4\,km$ @ $\lambda = 3cm$. OK

L = Coupling $= g_1 g_2\left(\frac{\lambda}{4\pi R}\right)^2$; $\left(\frac{0.1}{4\pi\,3000}\right)^2 \Rightarrow -111.53\,db$ @ 3GHz

ATA antenna gain $(g_1) = \frac{4\pi A}{\lambda^2} = \frac{4\pi\cdot\frac{\pi}{4}(6)^2(.65)}{(0.1)^2} \Rightarrow 43.6\,db$

Calibrate the C-band horns. One should have a gain of ~ 10db
Coupling: $-111.53 + 43.6 + 10 = \underline{-57.93\,db}$

A major problem will be multipath, especially at the longer wavelengths, with a source on the rim.

How stable and well calibrated are the satellites?

To get rid of the multipath, use a bright radio source and a low MHz BW correlator. (check with Lynn on the timescale for the correlator. Also for the back ends
How bright the cal source? What Tsys is needed?

Do set up one antenna with a cable output.

Correlation length for 100 MHz is $\Delta X = \frac{c}{\Delta f} = \frac{3 \times 10^{10} \text{ cm/sec}}{10^8 \text{ Hz}} = \boxed{3M}$

Expected antenna gains

38 db at 20 cm (1.50 GHz)
44 db at 10 cm (3.0 GHz)
50 db at 5 cm (6.0 GHz)
56 db at 2.5 cm (12.0 GHz)

for $g = \frac{A_e 4\pi}{\lambda^2}$

For a standard horn on the side with $g \sim 10$ db
$\Delta g \sim 34$ db at 3 GHz
→ $\Delta g \sim 40$ db at 6 GHz

sensitivity @ 6 GHz for 20 m² collecting area.

$KT_A = F A_e$ $K/A_e \sim 120$ Jy/K

The crab nebula is ~ 500 Jy at 6 GHz
expect 5 Jy in the cross multiply. Thats ~ .04K

For $T_{sys} = 100$ K and $\Delta f = 100$ MHz $\Delta T = .01$K in 1 second.
= .001K in 100 seconds
= .0003K in 1000 sec.

Get SN ~ 100 in 1000 seconds (20 mins) Eh

Full Correlation for holography increases the SNR by 100 to 10^4 (40 db)

At 3 GHz get to 46 db

double door on shielded room
oil pressure in the alidadade
Paul worried about a) connections in the dewar
b) possible need to change refrigerator
Lou suggests beginning with 25 room temperature receivers
(He noticed that JL turned off the compressor when he was out of the room)

production receivers need to be marked ~~for board~~ up

[Recall Hat Creek grounding for JD]

extended warranties on components bought early and used late.

for Front End production review as late as Jan. 04, an alternate plan for room temperature feeds is necessary - for the first 25 antennas.

[look further for K-band balanced mixers]

layoff costs? no

Make a lightning path to ground around the bearings.

---

post mortem
amazing progress! good use for radio astronomy as well as SETI
excellent new technology
Dave's overheads: 2" ducts for the air(?) watch out for RFI
2' @ 15 4' @ 45 mph sounds odd -
radome needs attention soon
lightning path needed
JL cryogenics great progress - but how long to get to full production NB! high reliability paramount on this system Consider ambient amps at the beginning April maybe a bit early for that CDR but minimize throw-away

First opt + testing do look for cross out coupling

Goff keep the antennas in motion ASAP

Feed (WJW) get a full system together ASAP need reliability and manufacturability info

Software Very good progress

Douglas Array config. looks fine
Beware of the RFI from everything
Put Double Doors on screen room

JD Good job on budget details -
Good plan to buy up front - but then some contingency is important minimum of contingency is ½ year FTE, even better 10% : some early retrofits may be needed

what component testing? no bed of nails -
but check un-stuffed boards - full just operational input/output tests

More on the calibration system

Antenna coupling at 12 GHz is $\underline{K = -58\,db}$

The input match takes power, $-6\,db$, so $k \cdot ff = -64\,db$.

Choices of cable to the antenna terminals

Precision Tube "Mini-loss" catalogue p. 2.24 copper outer conductor

(1) UC 50031-1 (silver plated) OD: .031" safe bend radius .140"
loss: 42.8 db/100 ft. @ 1 GHz
154 db/100 ft. @ 10 GHz 1.0 db f 8"

(2) UC 50047 (silver plated) OD: .047 bend radius .250"
28.6 db @ 1 GHz
92.4 db @ 10 GHz

(3) UC 50070-1 (silver plated)
19.9 db/100 foot @ 1 GHz
68.2 db/100 foot @ 10 GHz.

The .031" cable loss is 1.1 db at 12 GHz for an 8" length
i.e. 1.0 f 8/10 db

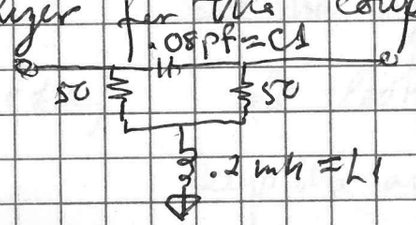
The slope doesn't quite fit the f^{-2} coupling between the antennas, but this cable loss is small so we can just include it with the coupling loss; for $\underline{-65\,db = kcff}$ @ 12 GHz

Starting with a diode noise of 300,000 K (+30 db ENR)
for the NC 401: Get:

$$300{,}000 \times -65\,db = 300{,}000 \times \underline{3.2 \times 10^{-7}} = .096\,K \simeq 0.1\,K$$

To get $\Delta T = 10\,K$ which will sometimes be useful requires a net gain of 20 db in amplifiers, including equalizers, and attenuators

The equalizer for the coupling ends up in the following form



It gives a nearly flat response in series with SP1 in Genesys.

SP1 is set to provide −3db at 300 MHz with a decreasing transmission of −6db/octave to mimik the antenna coupling. The combination of the equalizer and SP1 gives a flat response of −45.0 db. (It drops to −51.0 with a total negative slope over the band of −2db if we set L1 = 0.)

At 300 MHz the coupling is $3.2\times10^{-7}\left(\frac{12}{.3}\right)^2 = .00051$, −33.0db. That's 30 db more loss than SP1 has. (Decreasing the loss was necessary to avoid numerical noise)

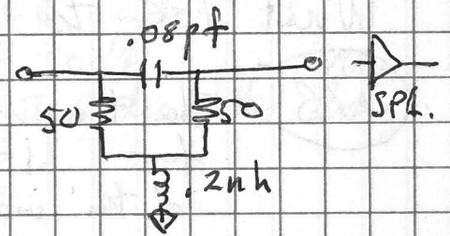
So the total transmission is −75db through the coupling plus the equalizer at any frequency. That's 10 db more than just the 65 db coupling loss at 12 GHz.

−75db × 300,000 = .01K 30db is needed to get to +10K.

Doug's PAM box has 50db of gain. If we remove one amp it goes down by 10−12db. Then the cal signal could be as large as 50−100K. That might be useful for some quick tests.

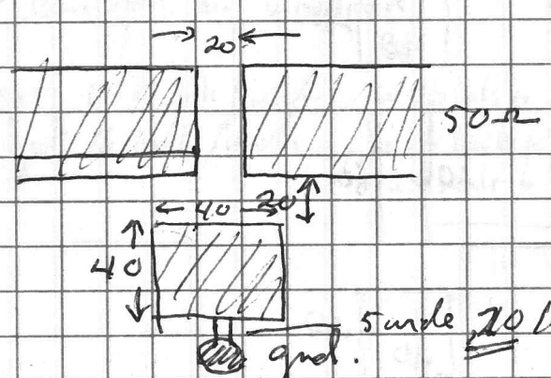
1) Remove the input amp and put in the NC 401.
2) Put in the coupling equalizer in place of one of the coupling capacitors.

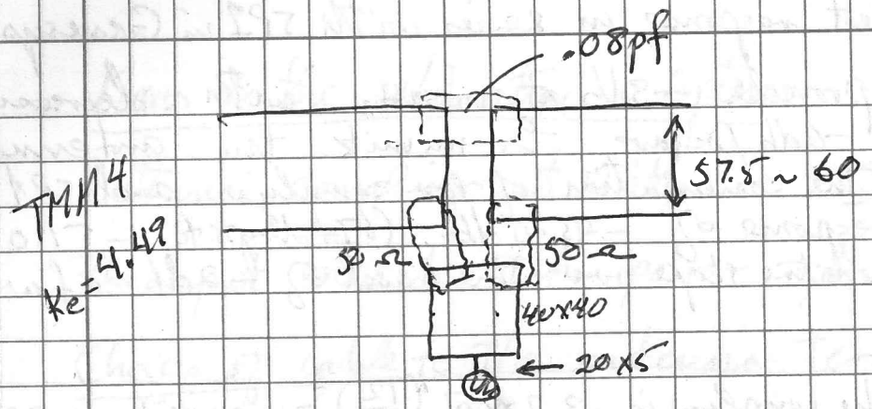
<u>The equalizer circuit</u>



SP1 mimics the antenna coupling loss. This produces a net flat response with total transmission of −75db. The response remains flat but varies by a few db net with changes of ±100% or more to either the capacitor or inductor.

Layout





~~assume h = 20 mil thickness~~

W = 60 and H = 32 gives 49.86 Ω

Use 20x40 components

The pad capacitance @ ~ .1pf is included here. It has only a small effect.

For a first test, it might be good to hook up an SMA connector at the input with the NE401 absent and a short over its position. In this way the overall frequency response (including the equalizer) can be checked. Or maybe just use the spectrum analyzer.

Power levels. Maximum with T = 300,000

$KTB = (1.4\times10^{-23})(300{,}000)\ 10\ GHz = 4.2\times10^{-18}$ watts $= 4.2\times10^{-5}$ mw

With a total maximum gain of 40db get .42 mwatt out, −3 dbm

Pout at 1 db compression is +13.7 dbm

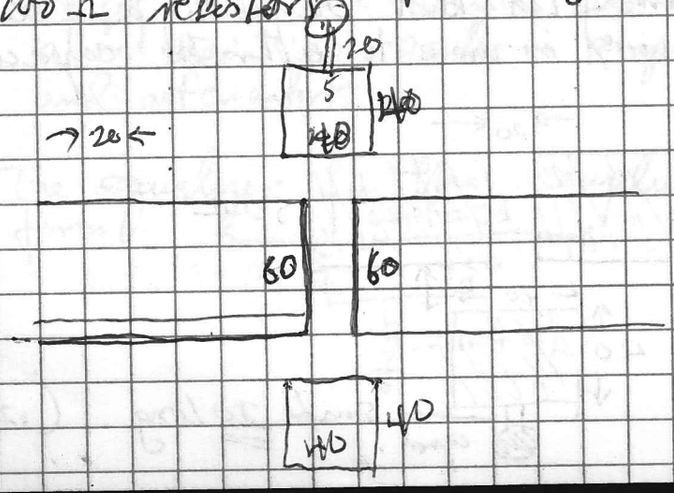
Even if the NE401 puts out 6 db more, there is still a margin of 20 db.

What TEC to use? (Ask Dave what he's got.)

NB The transmission remains flat and $S_{11} \leq -15$ db for $h = 0.4$, so we use 20 mils for the length of the inductor. The transmission is −72 db.

For Testing make up a layout that could take balanced 100 Ω resistors

Gerber File in C:\Program Files\Common File\System\Mapics\1033



Jan 03, 2003 meeting of NOW

Possible further budget changes a) RAL personnel may be increased

b) TAP recommendations

e.g. Telemhá testing – but do we need it? no

budget is approved @ 398 ask for Allen Foundation !?

→ Jack write up intellectual property part + spinoffs
Thursday April 3 10-2 w/ Berkeley

[Call Connie Chang re high speed multimode fiber optics]
our last year's Kvey is too slow

a couple things from Matt
Lyn's stuff memoryless corner turner ?
get note books signed –

From Rick Fisher specifications for a shielded enclosure for the Jansky Laboratory: Table 1 lists minimum plane wave attenuation of 60db for 150 MHz to 2.0 GHz.

A lot of details for building features: lights, ventilation, etc. Use all honeycomb for HVAC vents.

All line filters a minimum of 80db for 150 MHz – 2.0 GHz
Waveguide pipes should be at least 7-8 diameters long: pipes for optical fiber or plastic pipes.

From Tom Lauberker

They are building a room at Penticton 18m – 10m – 3½m
100db isolation for pocket

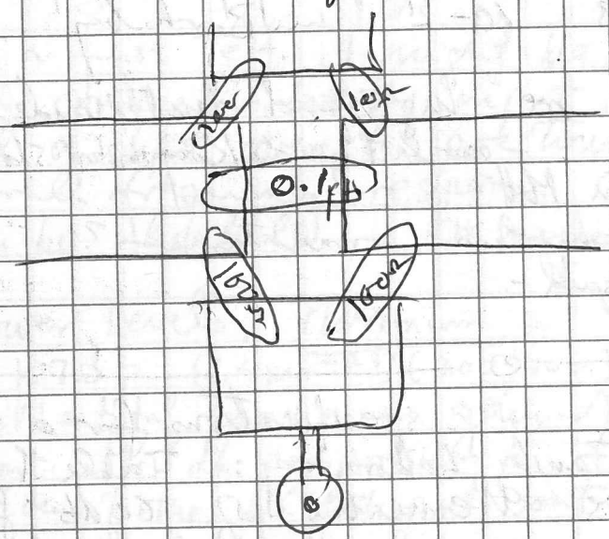
The short strip of line for an inductor TMM4 $K_e = 4.49$
$h = 32$ mils

For W=5 get 19.18 nh/inch, so 20 mils is 0.4 nh
For W=10 get 16.1 nh/inch, so 20 mils is .32 nh
25 mils is .40 nh.

1/29/03 meeting re feed etc. Sandy + Nikko here

Good looking active balun >30db common mode rejection –
Gain of ~27-28 db

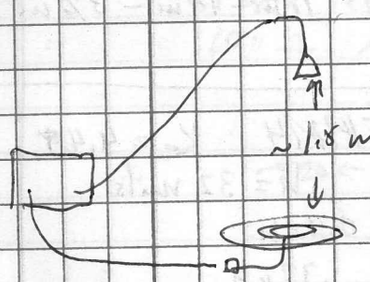
Tests of the equalizer for the the ATA antenna calibration scheme. The unit built is based on the layout of p. 128. The balanced layout was used. A 0.1 pf capacitor was used for the designed value of .08 pf.



The input match is ≲ -15db below 5 GHz. The worst case is -20 db at 9 GHz.

The transmission is close to 6db/octave for 0.5 GHz → 12 GHz.

Coupling between the calibration spiral and the small (2-18 GHz) spiral antenna. The "through" calibration made at ends of cables before connection to antennas.



Expected coupling $\tau = g_1 g_2 \left(\frac{\lambda}{4\pi R}\right)^2$

At 3 GHz

Measured small small spiral gain is 2.0 db (p74)
The calculated cal spiral gain is 2.7 db
$\lambda = 10$ cm $R = 1.5$ m $= 150$ cm.

$$\left(\frac{10}{4\pi \cdot 150}\right)^2 = 2.8 \times 10^{-5} \rightarrow -45.5 \text{db}$$

The input match on the Cal spiral has a loss of −6db (p. 87)

$t = 2.0 + 2.7 - 6 - 45.5 = -46.8$ db.

We also added a 6db pad at the input to the Cal spiral to improve the match (decrease the ripple), so the total expected to be measured is −52.8 db. at 3GHz. The actual measured values are:

ν	t
3GHz	−51.8
6	−56.7
9	−61.5
12	−68.5

ν	t
4.5	−53.8
9.0	−62.2

$\Delta = 8.4$ db

ν	t
3.5	−51.7
7.0	−60.0

$\Delta = 8.4$ db

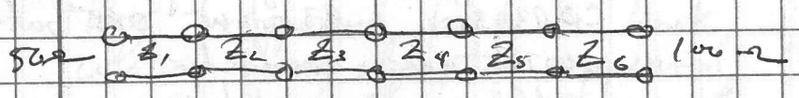
The higher frequency gain is dropping more rapidly than 6db/octave. Some of that may be in the input cable of the Cal spiral.

Copy to Disk on the 8720B Network Analyzer (p. 4-45)

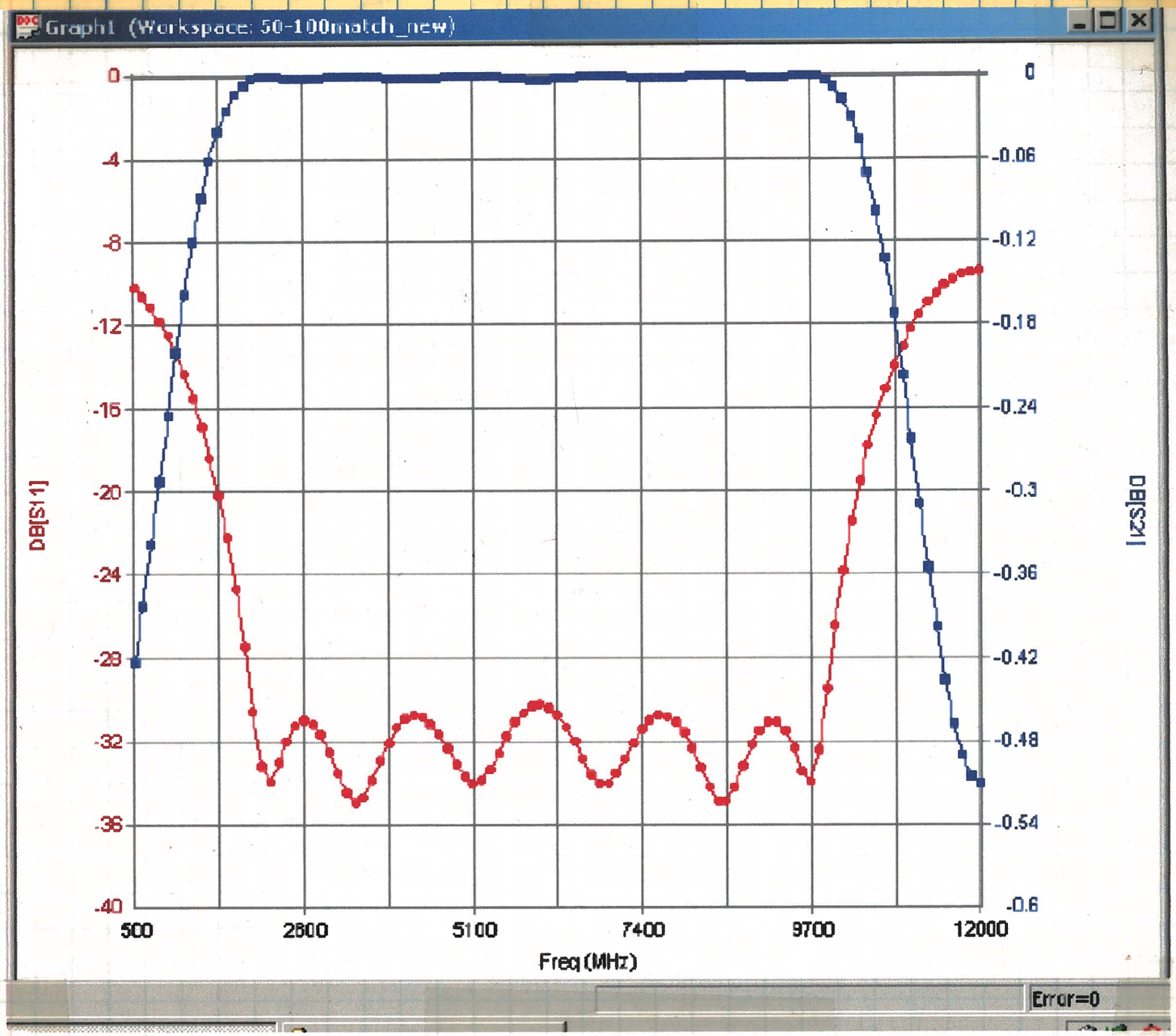
1. Press [Save/Recall] Save File Formats
2. Make sure that GRAPH FMT [JPG] is displayed.
3. Make sure that FILETYPE: GRAPHIC is underlined, Press if nec.
4. Insert 3.5" Floppy Disk
5. Press SAVE FILE to save the display

For connection between output 100 Ω of the balun to 50 Ω, an impedance transformer is needed. A six section λ/4 length can give a wide band match.

The following was optimized by hand using just sections of ideal transmission line in Genesys. Lines are λ/4 at 6GHz.



$Z_1 = 54.92$, $Z_2 = 61.525$, $Z_3 = 70.83$, $Z_4 = 81.17$, $Z_5 = 90.41$, $Z_6 = 96.79\,\Omega$



The plan is to use TMM 4 ($K_e = 4.5$) board which is low loss.

Widths for h = 32 mil board. (Genysis T-LINE)

	Z	W	λ(6 GHz)
Z_1	54.92	50.7	
Z_2	61.525	41.16	
Z_3	70.83	30.96	
Z_4	81.17	22.77	
Z_5	90.41	17.36	
Z_6	96.79	14.40	

Precision Tube (bought by Microcoax)
CA93085 is 93 Ω w .085 outer
.0079" inner conductor
@ 1 GHz loss is 24.1 db/100 ft.
@ 10 GHz Loss is 82.6 db/100 ft.

CIG Associates in Fremont
Charlie Dickerson 790-1193

Microcoax in Pottstown: (610) 495-0110

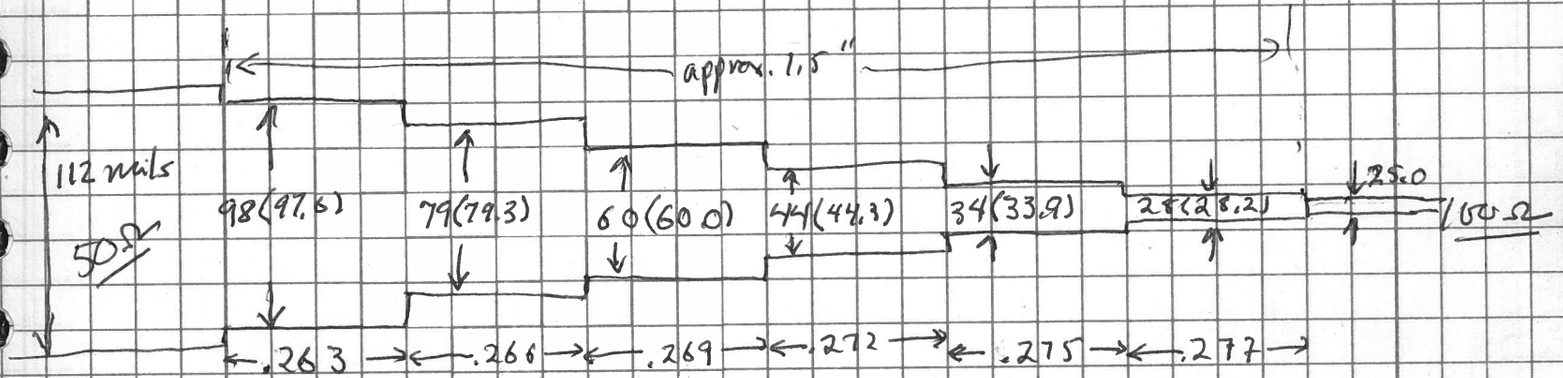
60 mil thick board gives wider traces for the sections

	Z (Ω)	W (mils)	λ/6"	λ/4"
Z_1	54.92	97.64	1.0526	.263 in
Z_2	61.525	79.30	1.0634	.266
Z_3	70.83	60.01	1.0769	.269
Z_4	81.17	44.34	1.0899	.272
Z_5	90.41	33.94	1.0999	.275
Z_6	96.79	28.24	1.106	.277

The wavelengths are found from the TLINE program in Genesys.

G. Files are in:
C:\Program Files\Common Files\System\Mapi\1033

Rounding the widths to the nearest mil affects the details of the −32 db level reflections but not the general level. [See 50-100 match in \1033 for details]



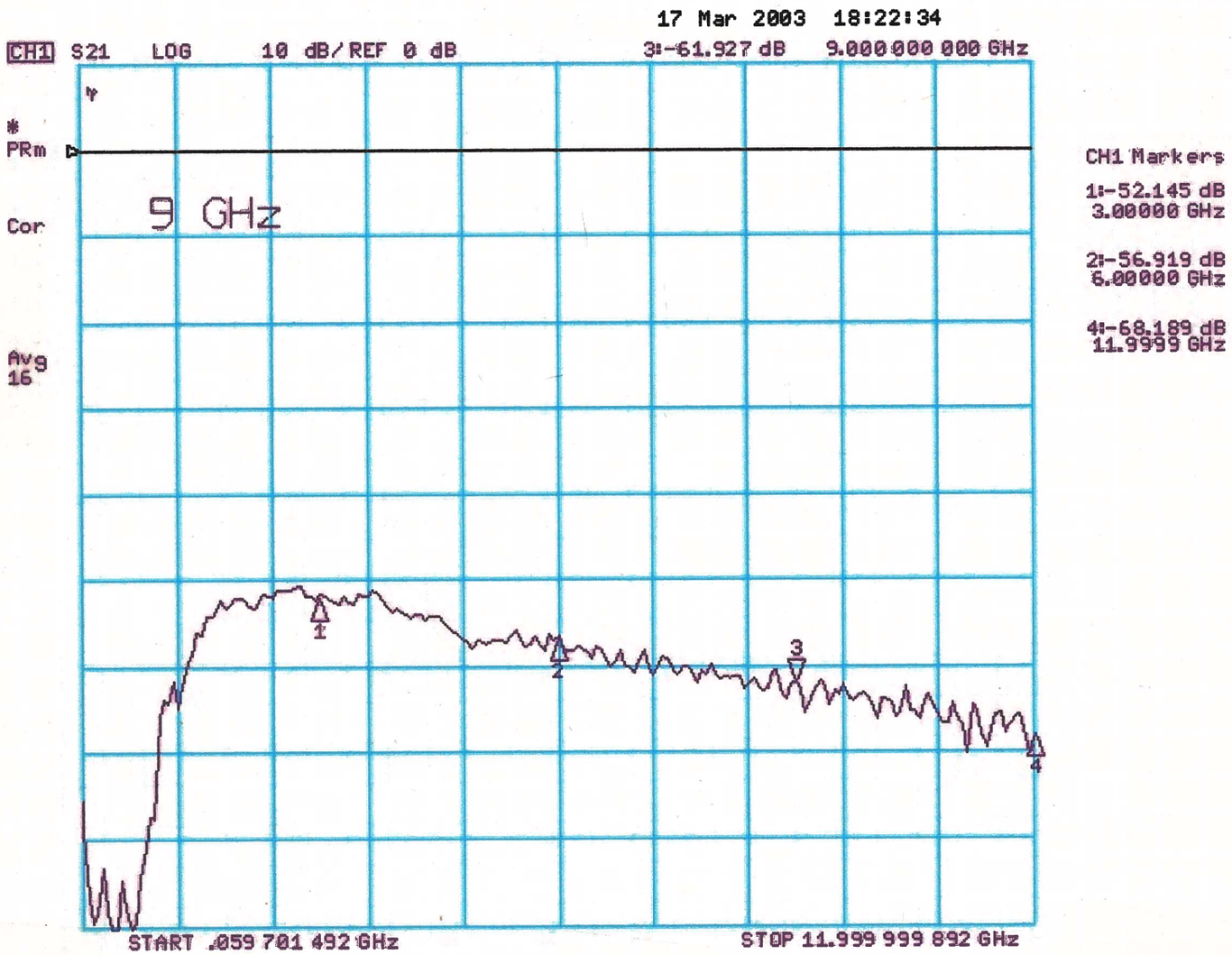
The more exact wavelengths differ from the simple $\sqrt{K_e}$ scaling by about 17% on the average.

100 Ω line is 25.0 mils wide
50 line is 112.2 mils wide

Putting a 50 Ω BNC connector on the 100 Ω output end screws up the transmission badly. Better
(a) we find some 100 Ω connectors, or
(b) we include the 50 Ω connector in the filter

In the first test the 100 Ω cable was wired into the two transformers back to back. It worked ok up to 10 GHz where strong absorption appeared. The box is so wide that many waveguide modes can propagate. It looks like the TE_{03} mode is causing the problem. The box width is 2" = 5 cm. Next attempt will be with narrow interior side walls.

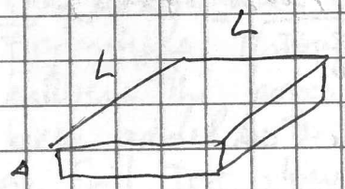
Coupling Between Cal Spiral & Small Spiral Antenna



First tests of the Cal Spiral with a small spiral as the test source (see I for tested the small spiral p.138)

See II p.131 for some more numbers.

Can we use the metalized film resistance and across the coaxial space in the oliver to lower its Q?



$$R_s = \rho\frac{L}{A} = \rho\frac{L}{L\Delta} = \frac{\rho}{\Delta} \text{ or } \frac{1}{\Delta\sigma}$$

Solitron gives the layer thickness: $\Delta = 50\times10^{-6}$ inches. $= .05$ mils $= 1.27\times10^{-6}$ m
The basic dielectric has $K_e = 5.40$ and loss tangent $= .035$ @ 1 MHz.

$$\sigma = \frac{1}{R_s\Delta} = \frac{1}{(150)(1.27\times10^{-6}\,m)} = 5.2\times10^{-3} \text{ mhos/m} \quad (10^{-4} \text{ of Cu})$$

The skin depth in pure copper is $\frac{.066\,m}{\sqrt{f}} = \frac{1}{\sqrt{\pi f \mu\sigma}}$

So δ is 100 times large in this layer material $\frac{6.6\,m}{\sqrt{f}}$

At 1 GHz $\delta = \frac{6.6\,m}{\sqrt{10^9}} = 2.1\times10^{-4}$ m.

whereas the layer is $\Delta = 1.27\times10^{-6}$ m about $\frac{1}{200}$ for 150 Ω/square. This explains why putting some of this material across the Aluminum box around the test balun had no effect.

Using stainless steel for the heat shield may work because it has such low electrical conductivity. This may make the Q in the coaxial mode small enough to suppress the sharp absorptions.

$\rho_{ss} = 90\times10^{-6}$ ohm cm at 20 C, Copper is $1.7\times10^{-6} = \rho_{Cu}$ (20 C)
$K_{ss} = .15 \frac{\text{watts}}{\text{cm K}}$ at 300 °K.

The Wiedemann-Franz relation is that $\frac{K}{\sigma} \propto T$ or $K\rho \propto T$

Since $K = .152(T/300)^{.46}$

$\frac{K}{\sigma} = C_1 T$ but $K = .15(T/300)^{.46}$

$\frac{.15(T/300)^{.46}}{\sigma} = C_1 T$, $\sigma = \frac{.15(T/300)^{.46}}{C_1 T} = \sigma \propto \frac{T^{.46}}{T} = T^{-.54}$

so with $\sigma \propto T^{-.54}$ $\rho \propto T^{.54}$

Thus. $\rho = 90\times10^{-6}(T/300)^{.54}$ ohm-cm.

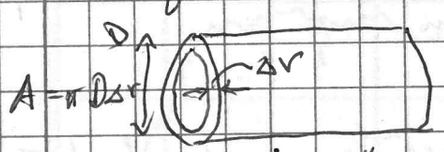
At $T = 80\,K$ $\rho = 44\times10^{-6}$ ohm-cm ; still pretty low

Will the heat conductivity in the SS shield be adequate?

Heat load due to radiation

~~In a 300k black body~~ John Lugten estimates about 50 mwatt of heat load on the shield.

Worst case: suppose that's the total heat flow from one end of the shield to the other end. $K = .08$ watts/cm-k @ 80k.



$\langle D \rangle \sim \frac{1}{2}'' = 1.25$ cm. $\Delta r = .010$ in

$H = \frac{KA}{L}\Delta T = \frac{(.08)\pi(1.25)(.0254)}{15\text{ cm}}\Delta T$

$H = .0005\,\Delta T$

For $H = .05$ $\Delta T = .05/.0005 = 100k$ Temperature drop. !!

Plating a thin layer of copper on the outside should work.

Work out the heat flow to the mean distance $L \sim 8$ cm.

$H = \frac{KA\Delta T}{L} = \frac{(5)(\pi\, 1.3)(.005)}{8}\Delta T = .013\,\Delta T$

For $H = .050$ watts $\Delta T = .05/.013 \cong 4°K$ ok.

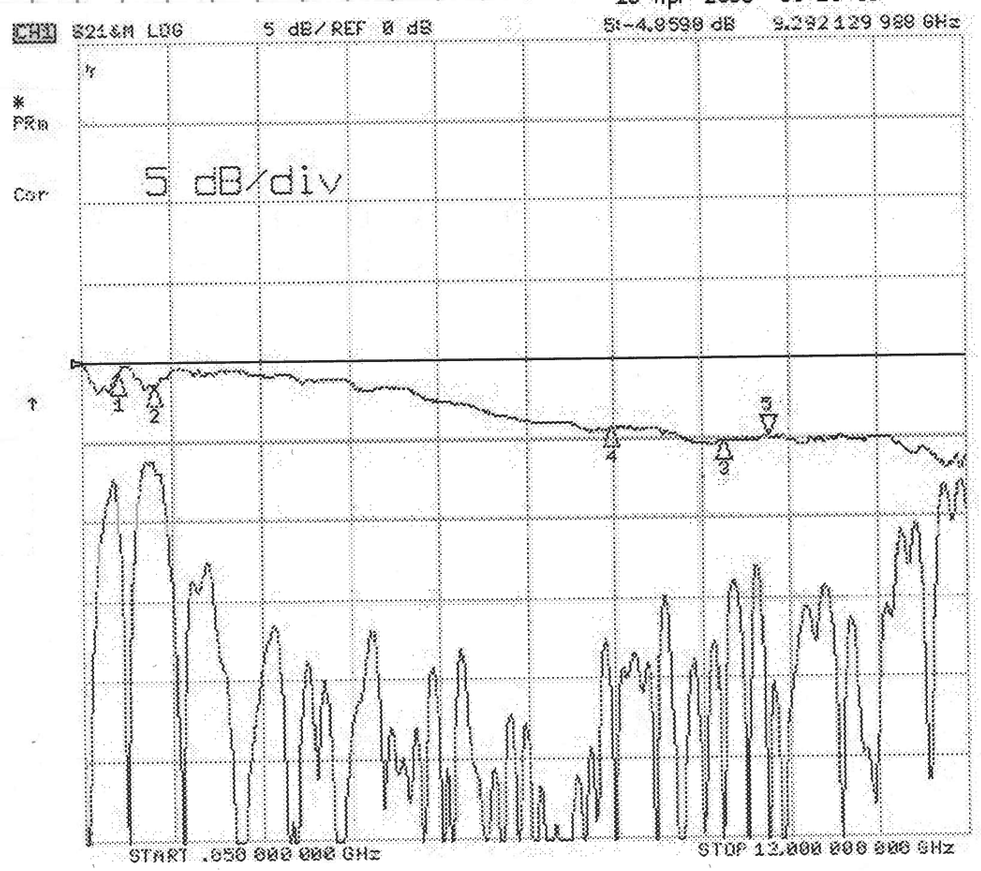
Conclusion Make the heat shield of 304 stainless 20 mils thick should be fine. Plate 2 mils of Copper on the outer side.

Further tests of the 100-50Ω matching boxes.

The boxes are connected back-to-back at their 100-Ω terminals through a short length of 100-Ω coax. (The separation between the boxes is 2.33 inches). Aluminum blocks have been added to the sidewalls to make the interior space narrower. a) First test showed improved transmission but many sharp resonances all over the band. b) Rubber absorber is then stuck to the side walls and to the top cover. The top cover absorber has little effect. The sidewall absorber eliminates the resonances but introduces overall absorption.

The file showing the overall transmission is SCRNOS.JPG. The input match is also shown. The match is overall 15-20 db except at 1 GHz where it is −8 db. However, this is a low frequency where the effects of both boxes should simply add, raising the reflection by 6 db. So the actual 1 GHz reflection should be about −14 db. This is a little worse than the calculated −18 db but it should be ok.

<u>Note</u>: The transmission shows dips at 500 MHz and 1 GHz. These should be due to the high combined reflections there and should be ignored for the individual box transmissions.



Required Gain stability for OTF observing

This is taken from the ALMA memo by Welch et al ( )

There are many on's along the track and one "off" during the turn-around.

ATA slew rate is 2°/second.
t_s is the time on source

$$\frac{\Delta T}{T} = \frac{2}{\sqrt{B t_s}}$$

Work this out for HI first (No)
$\theta_{1/2}$ for HI is 78" (of arc) (this is the array beam)

@ 2°/second slew = 7200"/second = 92 beams/second @ HI
Time on source is .01 sec. = t_s
For $B = 100$ MHz, $2/\sqrt{B t_s} = 2/\sqrt{10^8 \cdot (.01)} = 2/1000 = .002$
The reference is between each measurement and the off (turnaround) which will be much longer but also later depending on the scan time. So $\Delta G/G \sim .002$ for the time between the source and reference, a few seconds at most.

Keeping the scan time ~ turnaround time is efficient. For a slow turnaround time (~ a few seconds) and a slow scan time the gain stability must be better.

At 10 GHz at maximum slew, the time on source is shorter which helps. The limiting time on source could be the correlator dump time. Lynn Samp expect ~ 1 msec. for the minimum correlator dump time.

For spectral line work $\Delta G/G \sim .01$ for a few seconds probably.

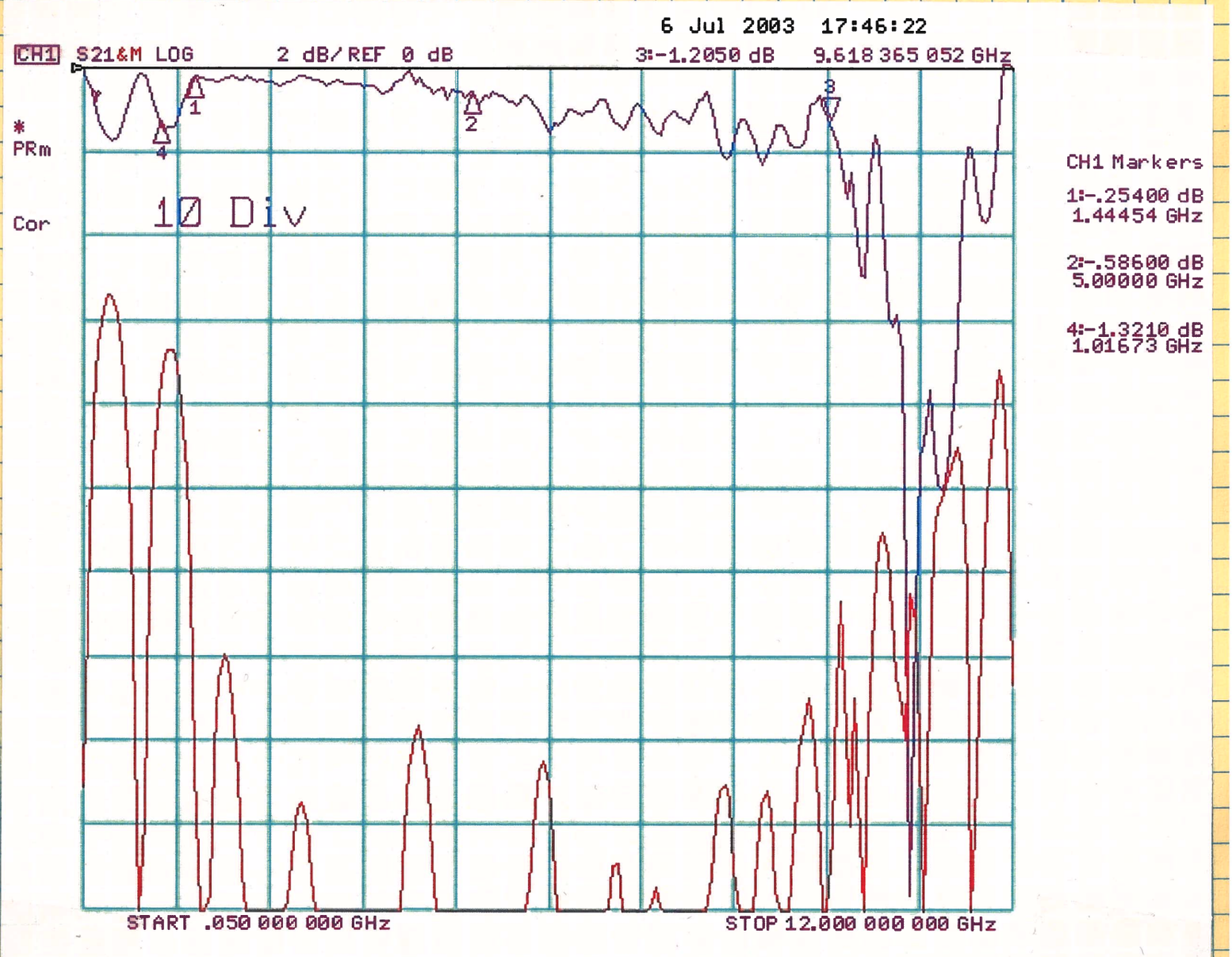
No This uses the synthesized interferometer beam.

see p. 152

Rebuilt the 100-50Ω matching circuit (from p.132)

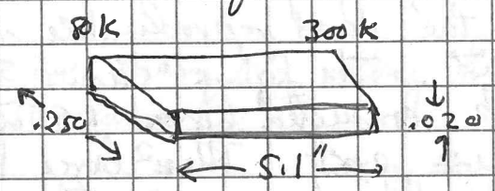
SCRN00.JPG works over 1.0 - 9.6 GHz

Took some time with calibration trying various cables. Two 'work horse' cables gave the most reproducible results. Full 2-port calibration used. The figure shows two matching filters back to back connected by a 1.95 inch length of 100 coax at their 100Ω ports. The 2 large low frequency reflections are probably due to this close proximity making the low frequencies reflection simply add between the two filters. The 9.6 GHz edge may be due to imperfections in the construction. The low frequency pts may well be ok for the seperate matching filters.



Some further Thermal calculations

I. circuit board for bias leads: ~21 traces on a ceramic board.



a) The ceramic board K (TMM-4) $= 0.7\,W/m^2\text{-}K$ at 300K
(It probably drops ~ a factor of 2 between 300K and 80K, but we ignore that for this estimate)

$H = AK\,\Delta T/\Delta x$ End Area $= .280'' \times .02'' \to .0063m \times .00050m = 3.2\times10^{-6}\,m^2$

$H = (0.7)(3.2\times10^{-6})(220)/.13m = .0037$ watts $=$ 3.7mw

b) copper Traces 5 mils wide and .7mil thick (½ oz)

Include the temperature dependence as case ④ of p. 66

$$k = 401\,(T/300)^{-.15} \text{ watts/m.K}$$

$$T(x) = T_0\left(\frac{x-x_0}{x_2}\right)^{\alpha};\quad \alpha = \frac{1}{1+\gamma} = \frac{1}{1-.15} = 1.177$$

$$x_1 = 0 \quad T_1 = 80;\quad x_2 = 5.1'',\ T_2 = 300$$

$$x_0 = \frac{x_1\left(T_2/T_1\right)^{1+\gamma} - x_2}{\left(T_2/T_1\right)^{1+\gamma} - 1} = \frac{-5.1}{(300/80)^{.85} - 1} = -2.457$$

$$T_0 = 80\left(\frac{5.1}{2.457}\right)^{1.177} = 188.97;\quad T(x) = 188.97\left(\frac{x+2.457}{5.1}\right)^{1.177} \checkmark$$

$$H_0 = A\,k_0\left(\frac{T_0}{T_2}\right)^{\gamma}\frac{T_0\,\alpha}{x_2(m)}$$

$A_{trace} = (.005)2.54\,cm = .0127cm = .000127m$
$\times(.0007)2.54cm = .00178cm = .0000178m$
$= 2.26\times10^{-9}\,m^2$

$$H_0 = (2.26\times10^{-9})(401)\left(\frac{188.97}{300}\right)^{.15}\frac{(300)(1.176)}{.13} = 2.23\text{ mwatts/wire}$$

For 21 wires: $21 \times 2.23 = 46.8$ mw, $+3.7 =$ 50.5 mwatts overall

Resistance of the Traces generally $R = \rho \frac{\text{Length}}{\text{Area}}$

For copper $\rho(T) = .75\left(\frac{T-20}{132}\right)^{1.35} 10^{-6}$ ohm-cm fits for $\underline{80 < T < 800}$

The temperature distribution $T(x) = 188.97\left(\frac{x+2.457}{5.1}\right)^{1.177}$, x in inches.

$T(x) = 188.97\left(\frac{x+6.241}{12.95}\right)^{1.177}$, x in cm.

$A_{trace} = 2.26\times10^{-5}\ cm^2$

Resistance/unit length $r(x) = \rho/A = .75\left(\frac{T-20}{132}\right)^{1.35}\frac{\times10^{-6}}{2.26\times10^{-5}} = .033\left(\frac{T-20}{132}\right)^{1.35}$

Substitute the temperature

$$r(x) = .033\left\{\frac{188.97\left(\frac{x+6.241}{12.95}\right)^{1.177}-20}{132}\right\}^{1.35} = .033\left(1.43\left(\frac{x+6.241}{13}\right)^{1.177}-0.1515\right)^{1.35}$$

$r(0) = 0.012\ \Omega$, $r(13) = .0908\ \Omega$

$R = \int_0^{13} r(x)dx = 0.605\ \Omega$ for each trace

For $I = 5$ ma. $I^2R = 0.605(.005)^2 = 1.5\times10^{-5}$ watts $= \underline{1.5\times10^{-2}\ \text{mwatts}}$
for 21 wires, $\underline{\underline{0.32\ \text{mwatts}}}$ total.

Most of this heat is dissipated near the ambient end.

In 1cm length at the warm end $I^2R = (.0908)(.005)^2 = 2.27\times10^{-6}$ watts

Copper specific heat is .09 cal/gm C°, $\rho = 9\ gm/cm^3$; $\rho C = 0.8\ cal/cm^3$

$\frac{\Delta H}{dt} = C_p\frac{\Delta T}{dt}$ 2.27×10^{-6} watts $= 0.8\ cal/cm^3\ \frac{dT}{dt}$; volume $= 2.26\times10^{-5}\ cm^3 \cdot 1cm$

$2.27\times10^{-6}\ \frac{joules}{sec} = 0.8\ \frac{(4\ joules)}{cm^3} \cdot 2.26\times10^{-5}\ cm^3\ \frac{dT}{dt} = 7.2\times10^{-5}\frac{dT}{dt}$

temperature rise with no conduction $\underline{\frac{dT}{dt} = \frac{2.26\times10^{-6}}{7.2\times10^{-5}} = .03\ °/sec.}$

ATA Sensitivity relative to Arecibo @ 1420 MHz

ATA @primary is 2.5° @synth = 78" = 1.3'
no. of beams = $(150/1.3)^2$ = 13,313 beams

The Arecibo beam is 3.0', effective diameter is 226 m.
The ATA collecting area is equivalent to a 113 m dish.
The Arecibo antenna has an instantaneous sensitivity which is $(226/113)^2 = 4$ times greater than the ATA

Arecibo must make $(150/3)^2 = 50^2 = 2500$ pointings to cover the ATA beam. For the time to make these 2500 pointings the ATA is observing every synthesized pixel for 2500 times longer. For the relative time the ATA is more sensitive in each pixel by $\sqrt{2500} = 50$ but Arecibo has 4 times greater instantaneous sensitivity. The overall advantage of the ATA is 50/4 = 12.5 in sensitivity. There is an additional factor of 2 because it does not have to do ons and offs. That leaves a net sensitivity advantage of 25.

The ATA correlator will have a best resolution of 1000 Hz, not 1 Hz. For a single 1 Hz signal, the 1000 Hz channel has less sensitivity by $\sqrt{1000} \cong 30$. This just balances the 25 times advantage.

Conclusion: The ATA with its correlator has the same sensitivity as Arecibo for the SETI survey. However, the frequency coverage is only 1 MHz.

Some properties of crystaline Quartz

from the Web. (Cryptran)

Thermal conductivity : 10.7 (para), 6.2 (perp) w/mK at 300 K
Thermal expansion : 7.1 (para), 13.2 (perp) $\times 10^{-6}$/°C
dielectric constant : 4.34 (para), 4.27 (perp) at 30 MHz
Loss:
Fused Silica has 0.55×10^{-6}/°C
Thermal conductivity is 1.38 watts/m-K
dielectric constant is 3.91 at 1KHz

The effect of the birefringence $K_e = 4.27$ vs 4.34

$$\frac{\lambda}{\lambda'} = \sqrt{\frac{K_e'}{K_e}} = \sqrt{\frac{4.34}{4.27}} = 1.008 \quad ; \quad \frac{\lambda}{\lambda_0} = \frac{1}{\sqrt{K_e}} = \frac{1}{2.07}$$

For the 10" balun → 25 cm

At 6 GHz $\lambda_0 = 5$ cm, λ_{glass} is 2.5 cm.

The length is 25/2.5 = 10 wavelengths.

$.008 \times 10 = .08$, $\times 360° = 28.8°$

This should have some effects on the operation of the crystaline quartz balun, but not much.

The thermal effectiveness of the fused quartz is lower than the crystaline quartz, by about a factor of 100 at 80K.

The average glass cross section is 0.4 in wide x .02 inch thick. = 8000 square mils, as compared with $\pi/4 (14)^2 = 117$ square mils for solid copper. 68 times the area. K_{cu} (100K) = 480
fused quart (@ 100K) is ~ 0.5 watt/mK smaller by a factor of 1000
Krystal quartz (@ 100K) is ~ 50, 100 times better, and 1/10 of copper.

The diameter of solid copper wire needed to carry the heat flow is 14 mils, the same as the needles.

see pp 66-69.

Tests on the double (back-to-back) Cufon balun

Metal Base replaced by a wooden base to avoid the weak TEM mode between the balun line and the mounting plate

Full two port calibration on the HP Network analyzer: High resolution 801 points over .05 – 5 GHz ; $\Delta f = 6.25$ MHz

A u shaped piece was formed that could be placed tightly over the double balun to test for coupling to the balun mode. It's length is 35cm (13.8 inches) approximately 2/3 the length of the double balun line.

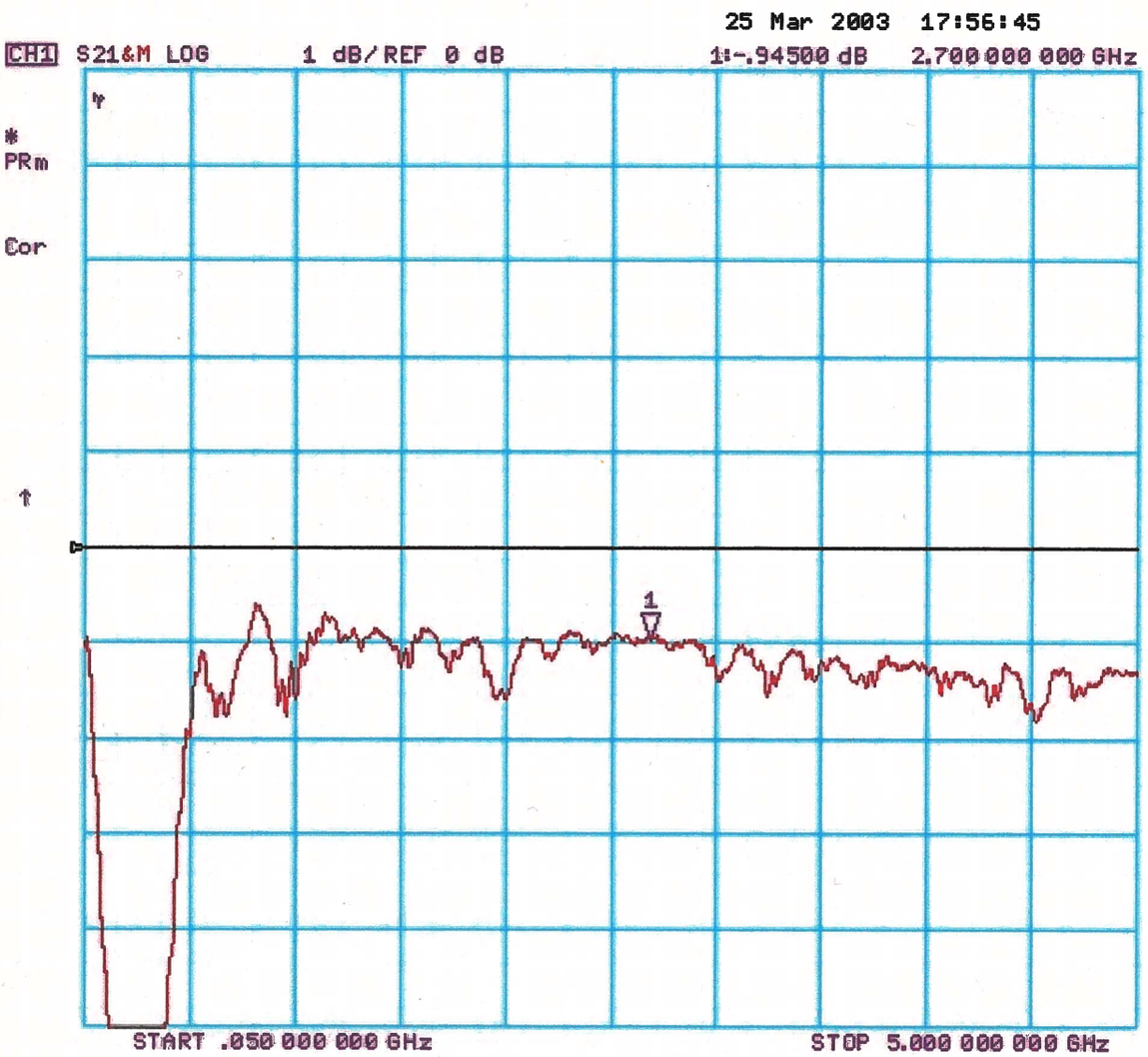
The double balun on Cufon is shown in pencil. The curved coaxial ~~ba~~ center piece is like this

It sits on the wooden base and fits tightly around the cufon. Being u-shaped allows it to be placed on or removed. Its characteristic impedance with the two inner lines as a single central conductor is about 200 Ω, close to that of the central twin line. If there were intrinsic coupling between the central twin line and the coaxial mode placing the u-shaped piece on the balun line should have a major effect on the transmission along the balun line.

No such effect is observed. ⟶

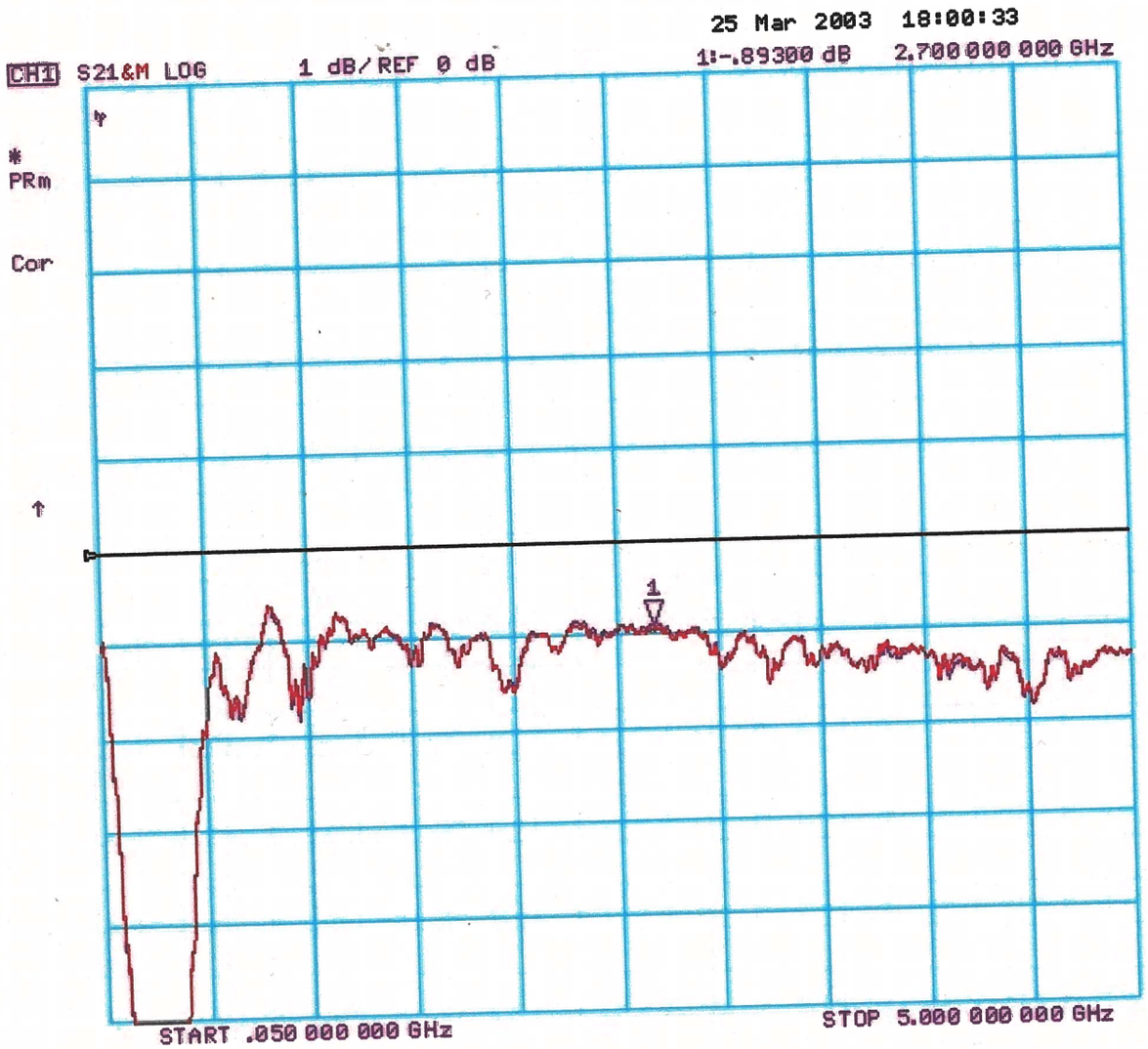
There is just one trace here. The data for this trace is displayed on all the following plots, along with the new trace.

Wooden Base; No Cover



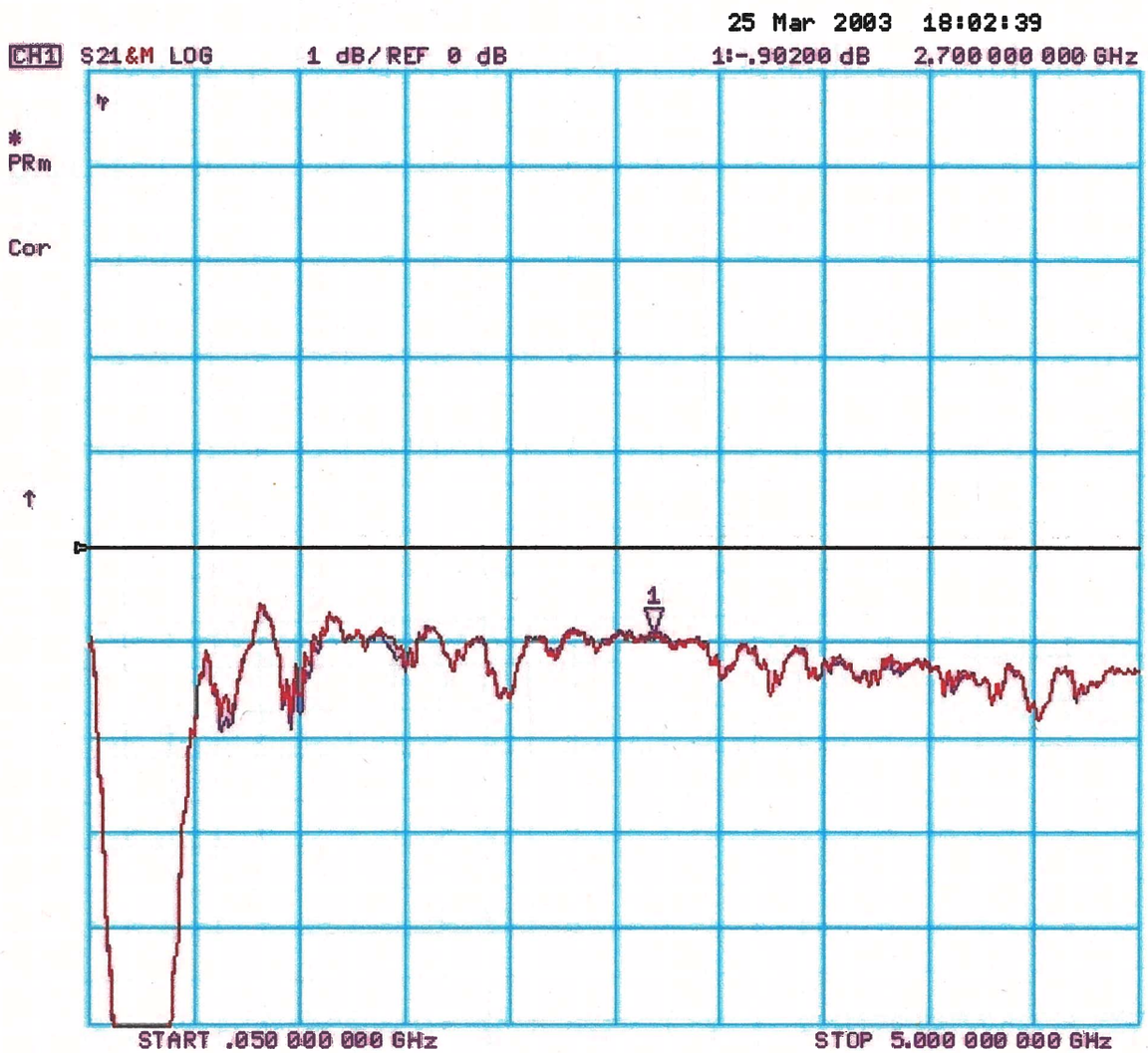
The u-shaped coaxial cover is in place for this measurement. The previous trace is also present. The two traces are almost identical. The transmission at 2.7 GHz actually went up by .05 db (1%)!

Wooden Base; Cover On, 2.5" From End



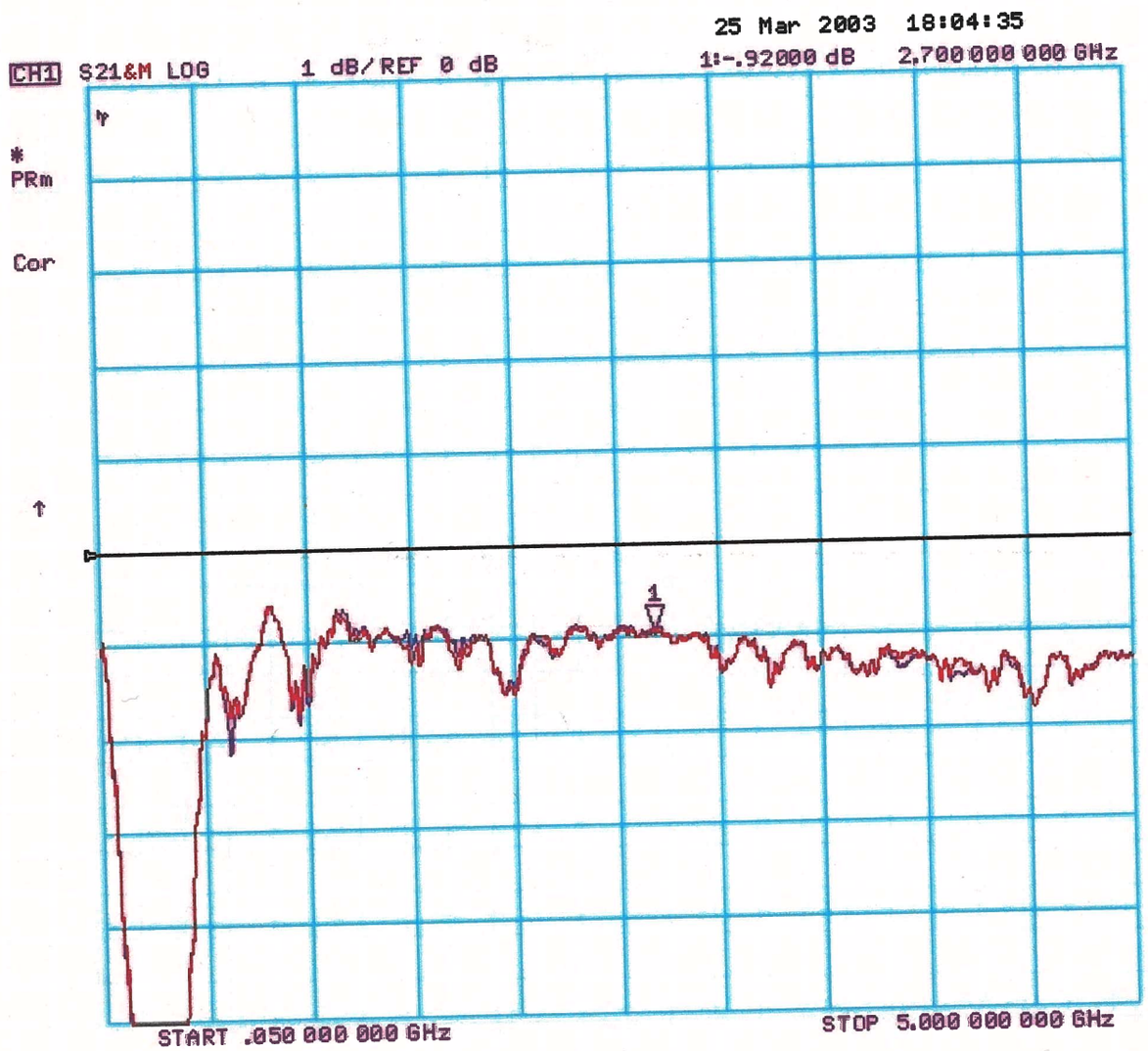
Moving the end of the cover toward the input end connector shows a slightly greater effect.

Wooden Base; Cover 1" From End



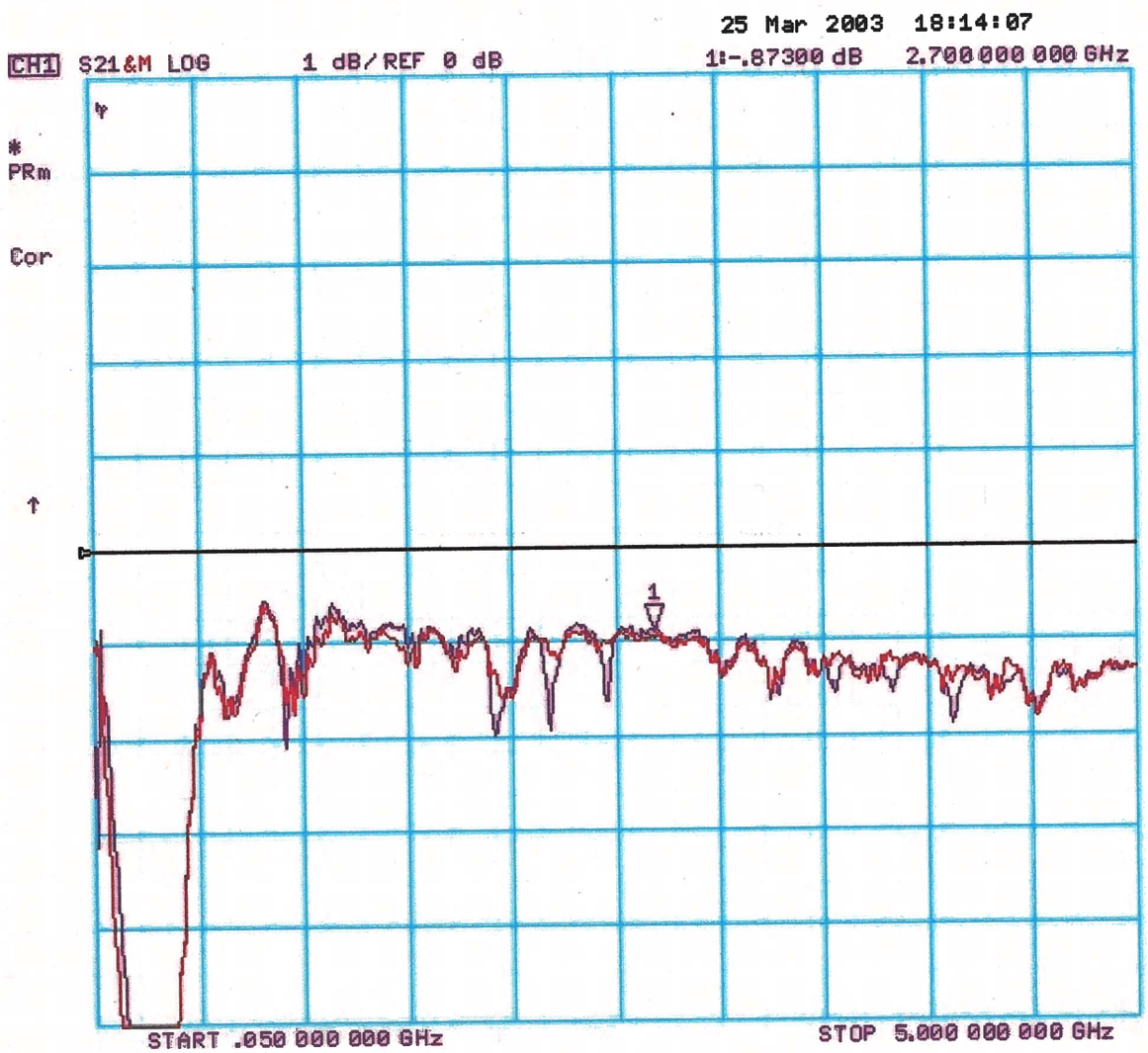
Bringing the u-shaped cover right up to the input connector at the end has a slightly greater effect

Wooden Base; Cover Touching End



If we put the rectangular cover on, with it covering both end connectors, sharp dropouts up to 1 db appear.
If we slide it off of one end, they disappear. Evidently the open ended coaxial line has a poor enough Q not to show the sharp resonances. The open end works better than the black absorber or a stainless steel cover in lowering the Q.

Wooden Base; Aluminum Cover End-To-End



How about Titanium sputtered onto inside of dewar heat shield?

Ti electrical conductivity: 23.8 mohm-cm
Thermal conductivity: 21.9 J/m-sec-deg = 21.9 watts/m-deg.

The CRC book gives values at 25C:
Titanium: 47.8×10^{-6} ohm-cm
Copper: 1.7×10^{-6} ohm-cm
Stainless Steel: 70×10^{-6} ohm-cm.

Expected operation at 327 MHz

The feed will be extended by about 19 inches so that it will work at 327 MHz. The main problem is that the phase center will be about 90 cm (~ 1 λ at 327) back from the secondary focal point. This makes the effective focal ratio F ~ 0.95 instead of 0.65. It is drawn in the layout of p. 19(II).

(1) From the curves of p 79(I) a one wavelength phase error at F = .65 gives a gain factor of $\underline{\underline{0.89}}$

(2) The angle subtended by the secondary is now 57° instead of 84°. From p 77(I) at F = .95 the forward spillover has gone up from 9.5% to 25.4% so the spillover loss factor is $\frac{.75}{.905} = \underline{\underline{0.83}}$ (not total)

(3) At F = .95 the edge illumination is about .465 The edge diffraction is up for this reason and for the longer wavelength.

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

$$= 0.20 \qquad \text{the diffractive loss} = 0.80$$

Altogether $(0.89)(.83)(.80) = \underline{\underline{0.59}}$ (−2.2 db)

This is the loss relative to higher frequency operation.

Repeat the calculation of p 138.

For HI the ATA primary beam is 2.5° (150')

The slew rate is 2°/second approximately ½°/second.
At ½ beam sampling that's one new independent point every 0.5 second of time. $t_s \cong .50$ seconds

For continuum $B = 100\ MHz$ $\quad \frac{\Delta T}{T_{sys}} \cong \frac{2}{\sqrt{B t_s}} = \frac{2}{\sqrt{10^8 \cdot .5}} = .0003$ should be 2

So the required gain stability is $\frac{\Delta G}{G} \lesssim .0003\frac{2}{2} = .03\frac{0}{2}\%$

For a 10°x10° map with a sweep of 5s/line and a turnaround time of ~ 5 secs. That's about 5 seconds between each pixel observation and the "off".

So $\frac{\Delta G}{G} \lesssim .000\frac{3}{2}$ for intervals of 5 seconds.

For line work at HI: $\Delta V = 1\ km \rightarrow 5\ KHz$

For $B = 5\ KHz$ $\quad \frac{\Delta T_n}{T_n} = \frac{2}{[5000 \times .5]^{1/2}} = .04$

so $\frac{\Delta G}{G} \lesssim .04$ is adequate here. $\lesssim 4\%$

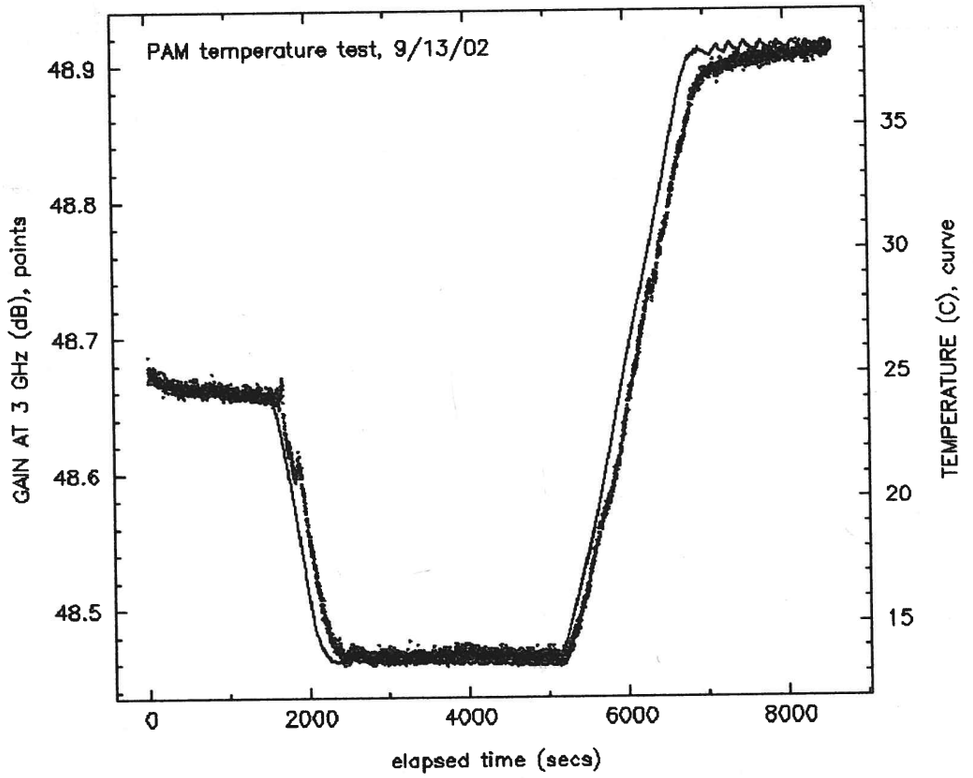
From Plambeck @ 3 GHz $\quad \frac{\Delta G}{G}/°C = 0.4\%/degree$

It looks like .1° stability will do. for the PAM

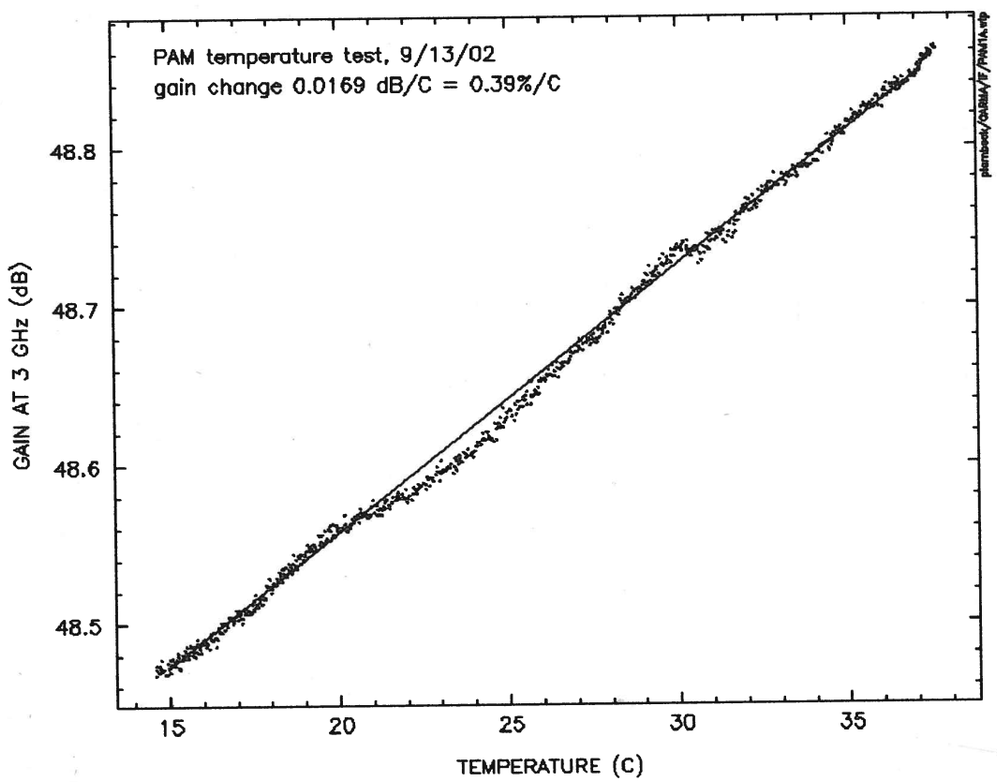
Date: 7 July '83

$$\frac{.004}{.00014} = 28$$

Strictly a stability of $\frac{1}{28}$°C is required



plambeck 16-Sep-2002 12:53



plambeck 16-Sep-2002 12:53