Antennas

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[[Mobile Antenna FS Comparisons](#)] [[Phasing Systems](#)] [[Sleeve Baluns](#)]
[[Baluns on Log Periodic Antennas](#)] [[Snap on core material data](#)] [[Short dipoles and problems](#)]
[[End-fed vertical j-pole and horizontal zepp](#)] [[Ground resistance measurements](#)]

Multiple Articles at:

Receiving: Contains information in many articles. Receiving antennas including Beverages and Beverage Antenna Construction, Loops, K9AY, Pennant, Flag, EWE, Slinky Beverage Antennas, vertical arrays, magnetic loop antenna, and terminated loop arrays

Transmitting: Contains information on various transmitting antennas

Single articles:

Balun and Core Selection: Contains information on core selection for transformers and baluns

Baluns, Sleeve balun: How they work

Baluns, transmitting: What they do, how they do it, and how to test them

Balun, Toroid: Winding style, debunking split-winding will improve performance

Crossfire Phasing: Contains information on cross fire phasing and why it is superior for broad bandwidth arrays

Combiner and Splitters: Contains information on Magic-T splitters and combiners, how they work and what they do

Detuning Towers: A quick explanation of how it works and the incorrect idea that you adjust for minimum current! Pass this along, it is a major error to tune for minimum current!

EH antenna: the E-H and CFA antennas. How they work.

Inductors, Loading: A brief tutorial on loading inductors

Mobile and Loaded Antennas: Small loaded antenna systems

(Related page: Inductor spice model)

Omega and Gamma Matching: Contains technical information on Omega and Gamma matches and matching, impedance limits, component selection, component failures

Radiation and Fields: What the terms we use actually describe

Radiation Resistance: A revised (as of Feb 14, 03) tutorial on radiation resistance and how it is used and misused

Traps: Measurements and operation of traps and trap antennas. Including trap antenna loss.

7676 **since May 2004**

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[[How to Check Signals](#)] [[Keyclicks](#)] [[MFJ-259B Calibration](#)] [[RF Noise Powerstroke Diesel](#)]

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[E-ham](#)

[FT1000D Click Mod](#) [FT1000D NB mod](#)

[W4AN Trust Fund. Bill Fisher contributed a great deal to amateur radio. I hope we can give something back to help his family.](#)

Welcome to W8JI's Home Page

Search:

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All contents of this site are the property of W8JI, but I welcome all links from other sites. Please, if you copy something from here (and even if you rewrite it) add a reference link here! This is entirely to protect technical content, because I refine and update things routinely as errors are called to my attention. Like you, I also learn new things every day. As my knowledge improves, I revise technical articles. I'll note revision dates on articles with changes (if the changes affect technical content).

The most important thing any of us can do to make the web an asset is to help each other with review to insure technical accuracy IMPROVES with time! Let's work to make Internet a reliable source of information instead of a collection of folklore!!

Other Recommended Technical Info Sites:

http://www.eznec.com/index.shtml

http://fermi.la.asu.edu/w9cf/

http://www.seed-solutions.com/gregordy/Amateur%20Radio/W8WWV%20Experimentation.htm

http://www.ifwtech.co.uk/g3sek/

QRN USA Lightning Storm Map

For an instant check on USA mainland QRN click on this link or this link to a weather map of thunderstorms.

Some of this site's (W8JI.com) contents are listed below:

New Page Cebik Dipole and other short dipoles. Feedline and tuner analysis!!

Amplifiers

SB-221 Tank Voltages and the parasitic oscillation myth. The real explanations behind amplifier parasitic oscillations and alleged damage by HF amplifier parasitic oscillations. Look at voltage throughout the tank at frequencies between 1 and 300MHz, and decide for yourself. Related article at Arcs, what makes an amplifier arc?

Preamplifiers for receiving. Measurements of receiving preamps. Compare your favorite pre-amp (text corrections 4/27/03)

Arcs, what makes an HF amplifier arc? This explains causes of amplifier arcing in any amplifier, including the TL-922 amp, Ameritron amps, Heathkit SB220 amplifier, and any other HF

power amplifier for amateur or commercial service. Simple failures and design or operating shortfalls are often explained as parasitic oscillations

Bandswitch failures the real causes and cures of bandswitch arcing and failures in HF power amplifiers that are often attributed to poor parasitic suppression causing oscillation

How to load or tune your HF amplifier and why an amplifier arcs when tuned incorrectly

Inductors deals with antenna loading coils but also applies to amplifiers

Related page: Inductor spice model

Relays how to speed and sequence amplifier T/R relays and how antenna transfer relays affect amplifier failures

Tank Circuits two models showing tank voltages, currents, loss, and phase shift

Vacuum Tubes how they work in RF power amplifiers and why vacuum tubes arc in amplifiers Includes details on 3-500Z and 8877 3CX1500 A7 tubes. Learn reasons tubes fail in amps

Vacuum tubes, how they convert d-c to RF how RF amplifier tubes convert dc to RF power

VHF Parasitics and general stability. Is your amplifier unstable? If you own a TL922, SB220 series, 30L1, or other amps with grid "negative-feedback" you will want to read this! (rev 5PM Feb23/03) Spice models. The resistor think might be non-inductive might not be what you think. Includes internal photos of various resistors.

ANTENNAS and antenna related topics

New Page Cebik Dipole and other short dipoles. Feedline and tuner analysis!!

EH antenna: the E-H and CFA antennas. How they work. See how my early theoretical predictions and descriptions compare to actual field measurements of this most recent antenna theory

hoax! For measurements click here.

If you think *baluns at the inputs of tuners* are a good idea, go to this W7EL link and read the section on current baluns. You might be surprised!

Another suggested tuner link is *W9CF.*

Balun and Core Selection: (mostly receiving) Contains information on core selection for transformers and baluns (it isn't a balum)

Balun, sleeve baluns: How a sleeve balun works and what is important

Balun, transmitting: Testing transmitting baluns and untrue folklore about choke, voltage, and current baluns. Dispelling myth that grabbing coax adequately tests a balun

Baluns: winding methods with toroid cores and debunking myth that a split winding improves performance

Baluns: Common-mode noise. How baluns can help reduce receiving system noise

BALUN TEST: New!!! Test data on transmitting baluns.

Beverages: Construction, Endfire Beverages, Broadside Beverages, Beverage Arrays for low band DX and other low noise receiving antenna arrays for 160 meters and other low frequency bands. Also see balun and core selection for Beverage transformers.

Combiner and Splitters: Contains information on Magic-T splitters and combiners, how they work and what they do

Detuning Towers: A tutorial with an important correction of the common error that we adjust for minimum current!! (revised slightly Feb 17, 2003)

Gain. Stacking (Broadside) and end-to-end (Collinear) gain. How it works, dispelling the 3dB myth.

Loading Inductors: A brief tutorial on loading coils (inductors) and inductor Q

Magnetic Loop Antennas: receiving

Mobile and Loaded Antennas: Loading coils, ground losses, and currents in the system. Related page: Inductor spice model

Power Stroke Diesel RF Noise: PowerStroke Diesel noise and how I corrected the problem in my 2003 F-250

Noise and common mode noise. How it gets into the receiving system Power line and other noise sources

Omega and Gamma Matching: Contains technical information on Omega and Gamma matches, impedance limits, component selection, component failures

Phasing, Crossfire: Contains information on cross-fire phasing and why cross fire phasing is superior for broad bandwidth low-band receiving arrays

Phasing Systems: A quick discussion of phasing methods

Polarization and diversity: Think you can have the best of the two worlds of vertical and horizontal? Think again!

Radiation and Fields: Electric field, magnetic field, Fresnel zone What the terms we use actually describe

Radiation Resistance: A revised (as of Feb 14, 03) tutorial on radiation resistance and how it is used and misused

Receiving and receiving antenna pages: Relative ranking of antennas for Topband DX plus various articles on low-band DX receiving antennas...including low band receiving antennas such as Beverage antenna and Beverage antenna construction. Elongated Loops and other loop receiving antennas such as K9AY, EWE, and flag antennas, small Topband or HF low noise receiving Verticals , my receiving antennas (with a demo), end-fire or cross-fire Beverages, common-mode noise in low noise receiving systems, and more.

Skin depth: The best explanation I have found

Transmitting: Contains information on my various transmitting antennas, including my eight

direction four-square

[Traps:](Traps) Measurements and models of traps, including Coaxial and Tribander Antenna traps

Receivers

Diversity reception "diversity" receiving using stereo along with some sound file examples

How I test receivers

Receiver Tests of some transceivers and receivers. Newly added 756PRO FT1000 MK V Kenwood TS870 (modified)

Receiver IM improvement **mods for FT1000** FT1000D

[DX Sound files:](DX Sound files) What some signals sound like here

[1000MP MK V](1000MP MK V)....mods for FT1000MKV receiver IM and transmitter clicks

Key Clicks

Includes sound files of signals with excessive bandwidth, a technical description of what causes clicks, and mods for the FT1000(D) and FT1000MP.

What Causes Clicks....The technical cause and what we can do about clicks.

Keyclick-MP.... An analysis and patch for the FT1000MP 's infamous key clicks

Keyclicks 1000D.... An easy and cheap patch for the FT1000 and FT1000D

Keyclicks 1000MP MK V....mods for FT1000MKV receiver IM and transmitter clicks

Also see Radios.

Noise

Technical article about noise and receiving/ receiving antennas

[Power Stroke Diesel RF Noise](#): PowerStroke Diesel noise and how I corrected the problem in my 2003 F-250

[MFJ-1025/1026](#) Technical Information

[Common-Mode Noise](#) issues with feedlines, and how to avoid problems

MFJ-259B Alignment instructions and some technical information [MFJ259B](#).

[MFJ-1025](#) and 1026 phasing unit technical information.

Local ARES activities will be included later. We have a local repeater that covers a good bit of I-75 between Macon and Atlanta. Please give a call on 147.225(+600) if you are passing by.

Special thanks to my friend Bill Fisher, W4AN, who is now a silent key. It was Bill's encouragement and generosity that made this site available.

73,

 Tom W8JI

This page has been viewed 355761 *times since March 2003. Overall site hits are over 1.5 million for the past year!*

160 meter History

[Home]

Welcome to a new page that will have archives of 160 meter history. First on this page, a download of W1BB's papers. These papers were contributed by Rolf PY1RO and converted to .pdf files by Ron PY2FUS. Information on this page is intended for private viewing. Publication without permission is prohibited.

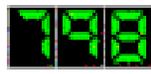History of 160 (about 15mb)

Newsletters

Feb 1962 Bulletin

April 1962 Bulletin

October 1962 Bulletin

December 1962 Bulletin

Dec 62 to Feb 63 Letter

1963? Letter (off screen at edges and incomplete)

since May 2004

©w8ji Mar 2004

HISTORY
of
160 METER DX
by
W1BB

This HISTORY is presented to you, by W1BB, in the hope that
it will be found useful, and interesting, in connection with
your 160 meter activities.

Please advise of any corrections, errors, or omissions noted.
Your help in this respect will be very much appreciated.

Best of luck, and HAPPY HAMMING to you always!!

B.C. N. U. on 160 !!!

73!
        de
            W1BB

HISTORY
of
160 METER DX TESTS
(Dx vs W/Ve)

(The following data for the period from 1931 to 1946 is
taken largely from an article by Austin Forsythe, G6FO,
Editor, in SHORT WAVE MAGAZINE, for April 1946.)

1931 DECEMBER   G to European Continent contacts. Up to 10 watts power.
OK3SK made it with 8 watts power.
(3)

1932 JAN-MAR   G to European Continent, Almost as good as 7mc.

      DECEMBER   Gs organized tests -- W1DBM with 450 watts, directional
360' antenna on tip of Cape Cod -- Heard s5 to s7 up to
o800 GMT. Peak period 0530-0730.

1933 FEBRUARY   G6FO using 9 watts and 66' antenna with counterpoise, worked
W1DBM 0600-0715 for FIRST Transatlantic Crossing. GW5WU
also heard in USA

1934 FEBRUARY   Several more Gs worked Across.

1935 JAN-MAR   G6FO-G2PL-G2II-G6YQ, and GW5WU worked across. VE1EA worked
G2II and GW5WU first time he ever tried 160 Dx.

1936 JAN-MAR   No scheduled 160 meter Dx tests -- However, FIRST USA/AFRICA
160 meter QSO made by W1BB/FA8BG on 2/23/36. VE1EA worked
G2II - G2DQ and G6UJ

1937 JAN-MAR   No scheduled 160 meter Dx tests -- However G2PL/W1BB made
FIRST 5-B.T.O.C. within 24 hours, working on 160-80-40-20-10.
VE1EA worked across on 160 to G2PL W2EQS heard in Wales with
30 watts.

1938 FEBRUARY   G6FO - Short Wave Magazine, arranged series of tests. Heard
W1BB but could not work across to W/VE - Conditions poor.

1939 JAN - FEB   Band cracked wide open Jan. 8-15-22 and Feb 4 & 16 between
0500-0830. Fifty stations worked across including G-GI-GW-HB-
F-FA-SM-OZ-W1-W4-W8. W phones were heard and also identified,
but not worked. Interest mounted high. W1BB outstanding W
station. Gs across were: G2DQ-2PL-2RC-2CF-2MI-2AO-3GH-3JU-5MP-5RI-
5JO-5XH-5QY-6MK-6WY-6GM-6QH-6SQ-6XL-F8ZP-GW2IP-GW8HI-FA8BG.

1940   WAR --- Band Closed

1941   WAR --- Band Closed

1942   WAR --- Band Closed

1943   WAR --- Band Closed

1944   WAR --- Band Closed

1945   WAR --- Band Closed

1946   Band opened up in "G" land 1800-2000 kc. Local working.
No Dx. Loran QRM. Band closed in USA

(The following data from the year 1947 on from the notes
    of WlBB, other interested Dx Hams , and SWLs.)

1947   Local G working -- no Dx. Band closed in USA.

1948   Local G working -- No Dx. Band Closed in USA.

1949 JAN-MAR   Gs heard across again. --- April 9th 1949 W band opened
                up again by FCC 1800-1825 Kc and 1875-1900 Kc. -- Also other
                bands for Western USA - See FCC notice.

      DECEMBER VElEA heard G6BQ and G5JU 12/17/49 11.19PM EST -0415 GMT.
                G3PU and W4NNN made First Post War crossing 12/18/49 with
                QSO. GD3UB and VElEA made second post war crossing 12/19/49

1950 JAN-MAR   -Unorganized tests by leading G and W/VE stations, indicating
                 that band had potentialities for D.x.
                -G2PL and G3PU outstanding Dx stations. Other Dx stations
                 working across were: G3FAB-G6BQ-G6GM-G8NF-GD3UB-KV4AA-G2YS.
                 Heard across were GM8UM and EK1AC
                -Of special interest - WlBB repeatedly heard very weak signal -
                 almost but not quite identified- called "QRZ QRZ QRZ G6?"
                 G6G?N G6G?N G6G?N - a number of times -- this was later
                 identified as G6ZN with only 3 watts power!!!!!
                -VElEA - WlBB - W4NNN outstanding W/VE stations. Other W/VE
                 stations working across were 1OE-3IU-3LII and W2WFZ.
                -Of special importance was first G/W two way phone contact
                 on 160 meters between G2PL and WlBB
                -WlBB-W4NNN-VElEA signals heard by FA8BG and EK1AO in Africa.
                -FIRST 3 element 160 meter beam tested by W2WFZ with G3PU and
                 with excellent results.
                -Eighteen Gs worked EK1AO Jan. 22 -- KV4AA heard by G4LX --
                 ZB1AR worked several Gs -- HZ1KE worked GM2HIK-GD3UB-G6AB-G6BQ
                -W4NNN (Now W6NWX) worked HC1PK during 1950 ARRL DX contest
                 for FIRST USA/South America 160 CW QSO. (Per QST July 1953)
                -HZ1KE QSO'd by G. First G-HZ??? First Asiatic??
                -TA3FAS QSO'd by G2pl - FIRST G-TA on CW and PHONE.

1951 JAN-MAR   -Organized tests -- by G6FO Short Wave Magazine - and WlBB
                 announcements by QST.
                -This was a very successful season, summerized by the following
                 "Outstanding events" from WlBB's report as follows:

                    #1-First crossing this seasonby VElEA working G6BQ on Dec.
                        31, 1950. G6BQ 349 and VElEA 449
                    #2-First N.W./Africa Postwar QSO VElEA/EK1AO Jan 7 - both
                        RST 449
                    #3-FIRST!!!! Contact between Asia and N.A. on record.
                        between VElEA and HZ1KE Jan. 14 1951 - 229 both ways.
                    #4-VElEA has worked 4 continents - 8 countries Post War -
                        total of 10 countries all time.
                    #5-WlBB also worked 4 Continents - 8 Countries post war
                        total of 11 Countries all time.
                    #6-WlBB worked South America HC1JW Jan.21 '51 BB was 469
                        and HC1JW was 567
                    #7-W2UKS QSO'd N.A./S.A. same date 349 - W8BKH also."/HC1JW
                    #8-WlBB and W9CVQ both heard in NEW ZELAND by ZL1AH Ex
                        G6AH and ZL1MP Ex GW6AA-Ex G2II of early 160 meter DX
                        fame. These reports confirmed. Also an unconfirmed repor
                        of W2QHH being heard by ZL1HM. Howy using 20 watts and
                        12' high antenna.

#9-A number of Phone/phone and CW/PhoneQSOs were made.
 G2PL/W1BB p/p F8PJ/W8FXS p/p W1LYV/GW3ZV p/p W1LYV/G2PL
 cw/p
#10-OUTSTANDING and Most Consistent DX station EK1AO - 300 Watt
#11-Also EXCEPTIONALLY OUTSTANDING for 10 Watts!!!! GW3ZV
 using miles of wire for an antenna!! Signals up to 589 at
 times!! This performance deserves a lot of credit.
#12- More US districts worked than ever before!!! W-1-2-
 3-4-8-9- VE-1-3

-DX working across were: G2PL-YS-NM-ACV-FLK-DA   G3PU-SU-DIY
 ERN G5HB-RI-UB-KM-JU-GU-VB   G6GM-QB-BQ-LB   G8JR-NF   GD3UB
 EK1AO HZ1KE GW3ZV-FSP   HC1JW   F8PJ KV4AA.
 Heard but not worked were the following stations: G2DPV,NP and
 G3GGN
-W/VEs working across were: W1BB,EFN,PLO,ORP,ZE,LYV,BEU,SAN,
 SS,KDX,DVS,RQR,BEU,NRE,AXA,MQB --- W2PTV,ESO,UKS,EQS,BFA,BLR,
 PEO --- W3LII,FNF,PMG,LOE,EIS --- W4KFC --- W6WKV,FKS --
 W9FIM,CVQ,PNE, -- VE1EA -- VE3AAZ.   Heard also were the
 following W1AW - W2WC - W2KZX W3OKU,AAA,JAA -- W8FIR,FLH,
 FLV and WØTQD.
-FIRST!!!!! W9CVQ worked ZL1AH December 31, 1951 RST 229 both
 ways - after a series of tries and schedules, making FIRST
 W/ZL 160 meter QSO on record!!!!
-This was an outstanding year -- But??? The SUNSPOT cycle in
 on the down grade - lets see what happens next year!!

1952 JAN-FEB   _Band was very much poorer than last year -- Due to Sunspot
 cycle being on downgrade????
-Outstanding events were as follows:

#1-VE1EA made first crossing working G3PU Nov. 25. 1951
#2-W1BB,W9NH,W2QHH and KV4AA heard in New Zeland by
 ZL1AH and ZL1HM during tests. SPECIAL NOTE!! W2QHH
 using 6L6, 17 watts, 270'wire 12' highat one end and
 3' at the other heard by ZL1HM and CONFIRMED!!!
#3-W1LYV was heard in RUSSIA - received SWL card!!
#4-VE1EA worked HZ1KE for second N.A./AFRICA 160 QSO.
#5-FIRST!!   W9/G QSO on record when W9CVQ worked
 G5JU and GW3FSP December 23rd. 1951 -- but included
 in 1952 "Season" report here
#6-W2QHH made FIRST VP4 QSO w/VP4LZ on February 23. (Again
 using only 17 watts.)
#7-W2ESO, Gene Black put weekly reports of test results on
 Voice Of America broadcasts with special note of
 FIRST W/ZL QSO

-Dx stations working across were: G2PL,AJ. - G3PU,ERN,COJ,DIY.
 G5JU - G6GM,BQ- GW3ZV,FSP. - EK1CW - KV4AA - ZL1AH - VP4LZ.
 ALL Dx heard was worked expect GM3OM who was heard and not Wkd.
-W/VEs working across were - W1BB,LYV,SS. - W2TRK,EQS,K2USA -
 W2RGO - W9CVQ,NH - VE1EA and VE2WW
-It will be noted that while there were several outstanding
 events, contacts to Europe wase way down, and it was seriously
 considered giving up the tests for 1953. It was W1BBs suggestion
 that they be continued to see what would happen.

1953 JAN-FEB   -In spite of poor conditions, there was a lot of activity and
 interest and good contacts were made. This was due in some
 measure to better antennas, beter receivers and MORE ACTIVITY.
-The OUTSTANDING EVENTS for 1953, were as follows:---

#1-W1LYV made first crossing for season nov.9th 1952
    working G6BQ, GW3FSP, G3HYG and G3GGN.
#2-FIRST!!! W5/G QSO between W5ENE and G5JU January 4th.
    G6GM and G3PU were also heard by W5ENE.
#3-FIRST!!! NORWAY/USA QSO between OH3NY and W1BB JAN.
    11th 0600GMT. W1BB using KYTOON (Balloon) antenna 260'
    vertical BB/449 NY/339
#4-NEW COUNTRY, EI9J on for the first time - Thanks to
    W2QHH - with outstanding signal and worked many W/VEs -
    Had special permit from his government for operation
    during the 160 meter DX tests.
#5-W1BB,W3EIS heard in South Africa by ZS3K
#6-W1BB,EFN-W2EQS-W3EIS-W9PNE-W0NWX all heard by ZL1AH
    varying RSTs 229/259 - and various dates during tests.
#7-W1BB experimenting with 260' vertical KYTOON - (Helium
    Balloon) supported antenna - slightly better than
    horIZONTAL ANTENNA _ BUT NOT OUTSTANDING _ WEATHER
    too much of a factor to make it practical. Hope to run
    further tests for confirmation.
#8-FIRST !!! ASIA/USA QSO on record - W1BB worked ZC4XP
    on March 8th.
#9-VE1EA worked ZC4XP for his third ASIAN QSO - having
    previously worked HZ1KE twice.
#10-Several QRP contacts of note were made:
            (a) G6ZN with 3 watts worked K2ANR
            (b) W2QHH and W1LMU with 17 watts each worked across
#11-A number of phone stations were heard and worked across
#12-First Post War W/EI contact - W2WWP/EI9J Jan 11th.
#13-First Post War VE/EI contact VE1YW/EI9J
#14- W9NH completed FIRST 160 meter "WAS" -(Worked all
    USA States) - on record. !!!

    -Dx working across were: G2HDT,FGD,SU,JF-G3PU,GGN,BKF,FGT,GZK,
     ATU,GLW,-G5RI,JU,TN-G6GM,CJ,LB-G8KP,JR. Other countries working
     across were EI9J - GI2ARS-KP4DV-KP4KD-VP4LZ-GW3FSP-OH3NY-KV4AA
    --Heard across were G2PL-G3GKQ-G3US-G5VB-G3HVX-G3FEW.
    -W/VEs working across were:- W1BB,LYV,LMU,AHX-W2EQS,WWP,QHH,WC.
     W3EIS - W5ENE - W9PNE - VE1EA -VE1HJ - VE1YW.
    -W/VEs heard only were - W1QJM,HSC,AYG,SS,TCR,OJM,FYS,QCA,RQR,
     DWO,AW,OE - W2KNZ,JPW,HCW,ESO,AMC,NC-W3HL,AVL,TBG - W4VFL,LRN
     W8NJC,HFG,GDQ - W9NH,MFV,FIM,CZT,BQQ - VE2AIE.
    -January 11th was the outstanding dsy of the tests - Dx was
     gOOD _ QRN was low -- later tests were QRN'd out with terriffic
     QRN, coinciding with the great hurricane and floods in Britain
     and Holland.

APR-SEP -SPECIAL NOTICE!!!
    -In April W2WWP received word from ZL1WW via ZL2ACV on 40 Cw
     that 160 meter faint signals were being heard, and asking
     for schedules.
    -On May 3rd W2WWP 1899 Kc QSO'd !!! ZL1WW 1904Kc RSTs WWP/349
     WW/239, for SECOND ZL/USA 160 meter QSO on Record!!(See
     W9CVQ Dec. 1951 for FIRST ZL QSO) -- This was FIRST W2/ZL
     QSO! - Times were 0900-1030 GMT -- Sunrise this end -- early
     evening ZL time.
    Thereafter W2WWP worked ZL1WW a number of times.
    -W2WWP invited W1BB and W9NH to join in on further tries.
    -W9NH heard first ZL on July 1st 119rst
    -W9NH next worked ZL1WW on July 29th 1022 GMT - NH/449, WW369
    -On the same date W2WWP also worked ZL3RB.
    -W1BB worked ZL1AH Sunday August 16th at a different time,
     0605 for a short "Contact" type QSO. (Sked suggested by ZL1AH)

-W1BB worked ZL1WW, ZL3RB and ZL1AH all on the following
 Wednesday August 19hh 0945-1030. ZL1WW was a solid QSO.
 RB and AH short "Contact" report QSOs.
-W2WWP the same morning, as signals were unusually good, almost
 had a phone QSO with ZL1WW when signals peaked - they heard
 each others modulation but no phone QSO -- CW QSO again OK..
-W1BB exchanged reports with ZL1NX on August 23rd - also Wkd WW
-Tests are continuing (At the time of this typing⊘ during the
 summer months and fall of 1953 to see if enough can be learned
 to establish regular tests in which it would be worth while
 for all Hams to participate - lots of lost motion this summer.
-Many Many thanks to ZL1WW, ZL2ACW and W2WWP for starting this
 thing off.   ZL2ACV
-FIRST !!! 160 meter "WAC" on record made when W1BB QSO'd
 ZL1AH, ZL1WW, ZL3RB, ZL1NX.

-W2WWP the same morning, as signals were unusually good, almost
had a phone QSO with ZL1WW when signals peaked - they heard each
others modulation, but no phone QSO resulted. CW qso again OK.
-W1BB exchanged reports with ZL1NX on August 23rd - also worked
ZL1WW
-Many Many thanks to ZL1WW,ZL2ACVand W2WWP for starting this ZL wk
-Thus the FIRST!!! 160 meter "WAC"-(Worked all Continents on 160)
on record was made when W1BB,QSO'd ZL1AH,ZL1WW,ZL3RBand ZL1NX.

1953 OCT-DEC   -This "SEASON", the organized tests were held per G6FO, Short Wave
1954 JAN-MAR   Magazine, and W1BBwith announcements in QST/CW and "Break-in".
-There was more activity than ever on 160 - more C ountries, and
all 6 Continents heard from.
-Much DX was worked in between the "Organized" tests, and it looks
now as though, through these "organized" tests, that the band is
being opened up more and more to general operation, and year/xxxx
round DX efforts.
-Summary of this seasons OUTSTANDING EVENTS from W1BB's log follow

#1W3RGQ arranged special tests starting October, and the FIRST
   CROSSING this season was on Oct 4th when W3RGQ worked G3PU.
   W1LYV Contacted G3PU later the same evening/ Signals 239/449
#2.FIRST!!! ZL/G QSO!! - 12,000 Miles, October 10th 1953
   o640GMT. ZL1AH worked G6GM. AH339,GM339 Half Hour QSO. On the
   following night, OCT 11th another QSO was made, this time 559
   both ways. On Oct 21st G6GM QSO'd ZL3RB - RB439/GM449 -ZL1AH
   QSO'd G6GM again also.
#3.FIRST W/HB QSO on record -when W1BB worked HB9CM Dec 20 '53
   W1BB 449 HB9CM459 0637 GMT.
#4. FIRST xxxxxxxxx Canal Zone/W QSO on record!!! When W3RGQ
   worked KZ5DE, January 16th 195%
#5.LOWEST POWERED G QSO this season. G2HKU, 5 watts worked W1BB
   Jan 3rd. 0640GMT - Isle of Sheppy. HKG/329 BB 579
#6.FIRST Hawaiian/New Zelamd QSO. KH6MC/ZL1WW March 24th 1000 GM
#7.Some FIRST KP4KD QSOs. - EI9J Jan 31st. G6BQ,Jan31st. VP4LZ
   Feb 21st -VP6EB Jan 31. - VP7NM Jan 31 - VP9BDA, Feb 28th -
   KZ5DE Feb 7th VE 1EA Feb 14th.

-In general, the first part of the TESTS had better conditions tha
than the last.
-An unusual phenomonon was observed in that the path seemed to
shift from good E -W conditions early, to a more NE -SW path, as
evidenced by Canal Zone QSOs with W and G stations

#8. FIRST G/C anal Zone QSO when G3PU worked KZ5DE on Jan 16th%

-We had TWO FAKES this season VR2BJ and LU4DM who fooled several.
-DX WORKING W/VE, were:-G3PU,ERN - G5JU,KM,RI - G6GM,QB,BQ,GJ -
G2HKU - G8JR - GC3EML - GW3ZV - EI9J-HB9CM-HB9T-KZ5DE-KV4AA-
KV4BB-KP4KD-VP4LZ-VP7NM-VP9BDA.
-DX HEARD were G3HVX and G3BKF.
-W/VE WORKING DX were:- W1BB,DBE,LYV,TMA,VDB,EFN,AQE,UJP,ZL --
W2GGL,EQS,AMC,WC,K2ANR,BWR,QHH,WH,PP,UKS - W3RGQ,EIS-W4KFC-W9PNE
W0NWX-VE1EA-VE2AIE0VE2WW.
-W/VEs HEARD only by DX were:-- W1AHX,LMU,QCA-W2JIL,MCU,TRK,AMC
HEY,KV -, W3FWF,TBG,HL,PA.-W4BRB,POB,KMS-W5ANO-W8GDQ,HQ,IKN,EJL -
W9EWC,CZT,NH.
-SWLs!!!!!!!! - Special Mention should be made of the splendid wok
done by the SWLs. To list a few of the most consistent ones, --
(next page)

-Norman Smith, BRS20601 -- ROBERT IBALL BSWL 4648 --- George
Allen BSWL 4588 -- C.L. Bradbrook G3692 --- William Iball,
BRS-20410 -- Mike Birch ISWL/DL2-4832 --- Jack Burgess, ISWL
2561 -- E.M. Hainsworth --- John Whitington ISWL-G3719.BRS
19771 -- Gerald Heslop RSGB A1167/ISWLG4555 --- Ian S Davies
RSGB/A1182 --- John L Hall, BSWL/2333. There were others.
-These very factual and accurate reports assisted greatly and
added mush zest to the 160 meter TESTS. Many many thanks to these
cooperative, and fine fellows - the "SWLs" !!!

#9. FIRST!!! - GC/W QSO on record, when GC3EML made contact with
W1BB December 17th, 1953. GC 3EML rst 229 - W1BB rst 459.

1954 SEP-DEC   -This "SEASON" there was a variety of DX, with the North/South
1955 JAN-APR   path predominating.
-LORAN became more of a problem, with a new station moving in
permanently on 1900Kcs, and an intermittant station on 1850.
This causes considerable QRM, and gives a high artificial noise
level that many weak stations cannot get through.
-Summary, of the OUTSTANDING EVENTS from W1BB's log, follow:-

#1. Season's FIRST USA/Oceania QSO when W8ANO worked ZL3RB Sept.
5th
#2. On Sept 5th G3PU heard W3RGQ 239-349 - and then on Sept.
19th W3RGQ almost worked G3PU for a first season's QSO -
when they heard each other briefly through static Crashes,
but did not call it a QSO.
#3. First Transatlantic QSO of the 1954/1955 season when W2EQS
and G6GM worked each other October 10th 0532GMT. Second
QSO of the season - same date - when W1BB worked G6GM 0607gt
#4. G6GM and G6CJ worked ZL3RB again in Mid-October
#5. FIRST VO/W QS O!!! When W3RGQ hooked W4KVM/VO6.
$6. FIRST LU/W QSO!!! When W1BB worked LU3EL November 28th
at 0707 GMT. EL/559 BB/569. This was followed by LU QSOs
with W8GDQ,W9PNE,WOCVQ,WØNWX,W4ZQ and W3FBX.
#7. FIRST YV/W QSO!!! - Was between YV5DE and W9FIM November 20
1954/ -- Then followed QSOs with W8GDQ,W9PNE,W1BB in order,
the next day Nov. 21st. We are greatful to "Mick" for QSOs.
#8. First HB9/KP4 QSO when "Ev" QSO'd HB9CM on 2/20/55 - this
makes 17 Countries for "Ev"
#9. FIRST (So far as we know) USA/S.A. Phone QSO, when WØNWX
worked YV5FH March 13th.
#10. DXPE DITION!!!! TI9MHB, provided a thrill on 160 and another
First, when opening up from COCOS ISLANDS 2/20/51 0500-0800
GMT, the following were worked. A new Country, with WØNWX
for a FIRST!!! - He also worked in order -- W4KFC-3EIS-1BB-
8DUY-9NH-8ANO-9PNE-3KLA-2QHH-1ZL-TI2BX-9FIM-9CVQ-6KIP/6-
YV5DE and 3RGQ.
#11. First YV/g - or YV/Europe QSO February 6th 1955 when YV5DE
and G5JU QSO'd.
#12. FIRST!!! USA/Columbia QSO, Mar h 20th when W1BB worked HK4DP
at 0540 GMT, followed by W1ZL and W8ANO. Incidentally, this
was HK4DPs First EVER QSO on 160 meters.
#13. FIRST!!! USA/HONDOURAS QSO March 27th. W1BB/HR1LW at 1015
GMT through terrific QRN. Followed by W3RGQ at 0447. This
was also "Loui's (Hr1LW) First CW Contact - and also First
QSO EVER on 160.
#14. FIRST!!! USA/FAR EAST QSO, when W6KIP/6 worked VS6CQ April
3rd 1955 5.05-5.25 AM PST, at DAWN. VS6CQ using only 10 watts
has been trying for USA QSOs 3 years - the only 160 meter
signals heard in that time were VS6CZ, a ZL and then W6KIP/6

#15. FIRST "WAS"/PHONE made by W8GDQ - F.B.!!! This is Second
     160 WAS. (See No. 14 1953)
#16. SECOND CW "WAS" - and THIRD all time WAS, Made by W2QHH

-W/VEs working DX were:-- W1BB,AHX,VDB,LYV,ZL,ORP,SFE/1,ERX,UXS,
  YAA,ME,BIL,WY. -W2EQS,GGL,WHH,WWP,K2EDL -W3EIS,RGQ,KLA,DGM,VAN,FBX
  W4VNE,KFC,ZQ,LXA,INV,LCV - W6KIP/6,VBY -W8GDQ,ANO,KIA,NSF/9,PEV,
  DUY,PCS,WCL - W9PNE,FIM,CVQ - WØNWX -VE2WK,WW .
-Other W/VEs HEARD only by DX were;-W1BMW,WM/2,CF,LQ and VZE -W2MX,P
  PP,HX,SKE,K2BWR,BVN,WH,WZ,CF,DLO -W3LRN,KMS,RWH/4,HIH - W5CZ - W8
  W8PCS,BMJ,SHP/1 - W9CZT,NH - VE2NI - VE3AAZ
-DX working W/VE were:- G2AJ,PL,HX,NX,NN -G3PU,WW,GGN,JOJ,JVI,BKF,
  ERN,JEA -G5JU,RI - G6GM,QB,LB -KP4KD,KP4CC - KV4AA-HB9CM-OK1HI,LM,
  AJB -OK3AL,VV,DG -OK2AG -ZL3RB,WW -ZL1BY -EI9J -GI3IOS - TI2BX -
  TI9MHB- LU3EL - YV5DE,FH-VO3X - W4KVM/VO6 - VP7NM,NG.-HK4DP -HR1LW
  XE 2OK-KH6IJ
-DX HEARD only by W/VE ---G3HRW, G6QB,LB -OK3VV -YU1GM -TI2WR -

1955 JUN-AUG -SUMMER 160 meter DX!!!!! -Some of the boys believed that they could
  dispel the myth that DX on 160 was not possible during the summer
  months. (ZL schedules had been successfully worked in the summer)
  -G3JOJ,G3GGN and G3ERN started the ball rolling to test this out,
  by writing several Ws for skeds during June July August.
  *The Ws responding were W1BB,K2BWR,3RGQ,9PNE,9NH.
  -On July 3rd W1BB's signals were heard by G3JVI at 579 at Sunrise
  time G land Sunday Morning. G3JVIs signals were weakly heard.
  -The following Tuesday morning K2BWR worked G3GGN 1830KCS for the
  FIRST!!! summer crossing on record.
  -On July 12th K2BWR worked G3GGN again.
  -On July 13th K2BWR and W3RGQ worked G3GGN
  -On Sunday morning W1BB worked G3JVI at o325GMT/Heard G3GGN weakly.
  -WWWWWWWWW W1BB's signals were heard regularly EVERY Sunday
  morning for EIGHT weeks at Strengths from 349 to 589
  -G signals were heard and worked only on 3 mornings. Faint signals
  on two other occasiona.
  -The high static level kept the weak 10 watt Gs from coming thru.
  -Sunrise time - G land , as in the winter, was found the ideal time
  for signals to peak and make QSOs

1955 SEP-DEC -This "SEASON" instead of "Tests" every other week end or so, it was
1956 JAN-APR decided in cooperation with ShortWaveMagazine and W1BB to have the
  "Tests" EVERY sunday morning - then, if the in-between S unday
  happened to be an excellent one for DX it wouldnt be missed. This
  through December to Mar h 1956.
  -During SEP-November, VS6CQ,VS6CW,VS6CZ ran special tests with
  W8GDQ and W1BB to try and see if QSO could be made. While these
  schedules were kept regularly every week, no contact or signals
  were heard. VS6CQ "sparking" these tests has now gone back to G.

1955-1956 SEASON — SUMMARY of this seasons OUTSTANDING events, follows, and is noteworthy for the large number of FIRSTS!!!

#1 -"FIRST!!" - DL/W QSO when "Shely", W3RGQ worked DL1FF, (Who secured special permission to operate on 160 from Dec. thru FEB-December 8th at 0510 GMT!! -- Followed by W3FBV, W8GDQ,W1BB same morning -- DL1FF was on whole season, with other DLs.

#2 -"FIRST!!" - YN/W QSO, December 15th when W2QHH worked YN1AA, (loo Watts) for FIRST Phone/CW QSO, on 1805 Kcs. YN was #59.

#3 -ZL1AH worked the long distance to G6CJ three times, on Oct. 13,14, &15 - ~~███████████████████████████~~ Sigs weak.

#4 -"FIRST!!" - HE/W Qso on February 12th when ~~████████████~~ HB1CM/HE, Liechtenstein, (HB9CM on DxPEDITION) worked W1BB at 0505 GMT, after a previous unsuccessful attempt. HB1CM/HE's signals came thru repeatedly and unexpectedly at 0505-0600, rst 229-449. This was an unusual performance on HB9CM/HE's part, and he is to be congratulated.(See separate story, which has been written on this for the magazines, with photos.)

#5 -"FIRST!!" - W/VP2 160 meter QSO ever with TORTOLA, British Virgin Islands, LEEWARD Group, when Howey, W2QHH worked FP/8AK/VP2 (W2BBK) on 1801 at 4.40 AM EST March 5th 1956 to receive 449 just as he was packing up to conclude DxPEDITION.

#6 -"FIRST!!" - Barbados to USA ever QSO, April 6th 1956, when VP6RG, "Ron", a high school Boy, snagged W1BB on April 6th for a report of 429x, BB being 569x. This represented a lot of hard work, perseverance and good planning on VP6RG's part, Of special interest is the fact that the pole supporting one end of the long wire antenna was our in the sea, quite a few feet from shore so that quite a part of the antenna was OVER salt sea water -- NICE !!!

#7 -"FIRST!!" - Recorded USA/CUBA Qso on 160. W1BB and CO2BL worked it May 31, 1956 at 0700 GMT through heavy QRN. BB,559 -BL,449. "Doc" very cleverly and quickly "Converted" an 40 meter antenna for use on 160 - and it worked out -- well, well enough!!!

#8 -W1BB's signals were heard, for FIRST SWL report - by Ernest Sprock, in SURINAM, PJ2CK, South America, on July 1 1956. This is interesting news from a Country rarely heard from.

#9 -An unusual "FIRST" was made, when after a series of schedules, the team of OM/XYL, ZS2GE/ZS2KZ got across to W1BB on July 11th at 0505 GMT, also September 8th 0400-0440 GMT for the very "FIRST" ZS/W Qso on 160. (A special SUMMER BULLETIN to ~~████~~ magazines, covered this event in detail)

#10 -Another important, and RARE, "FIRST" !! was made by CP5EQ, Bolivia, S.A. putting that country on the 160 Dx map, by working W1BB, July 22nd at 0800 GMT to receive RST 119-349 on QSB, with bb at 449

#11 -"FIRST!!" - British Guiana/W Qso by "DAD", VP3AD 119/339 -W1BB the same, on Sept. 12th. The previous week when contact was ~~██~~ about to be established "Dad"s xyl announced the expected Jr. ~~█~~ Opr. would arrive, and the schedule was QRT just in time, but fast!! - A fine little fellow did arrive!! Congratulations!!!

- These summer FIRSTS by W1BB, are the result of testing out his
theory that when it is Summer up North, it is WINTER down South,
and conditions would be best for the Southern Countries & it worked
out that way - but meant a lot of "EAR DRUM BUSTING" by static, for
W1BB to keep these schedules.

- In general it was thought that the EAST/WEST conditions were not
as good this year as last - believed due to sun-spot cycle -
However, as will be noted - a lot of DX WAS worked, and some have
the feeling that 160 meters is the least affected band of all.

- PARTICIPATION was good --Totaling 16/1s,24/2s,16/3s,17/4s,6/5s,
3/6s,2/7s,23/8s,15/9s,8/Øs - one each VE1,2,3. --- Dx as follows:
Gs-26/1s,3/2s,5/5s,4/6s,1/8 --- 3/KZ,4/KP,2/TI,2/GM,2/HB,3/ZL,
3/VS,1/EI,6/VP,1/EL,3/GI,1/OA,9/DL,3/XE,2/YN,5/OK,1/GD,2/KV,1/SP,
1/HR,2/GC,1/HE,1/KH,1/CO/2/ZS,1/CP and one VP/3.

W5CGT

- WORKING Dx were W/VEs,,, W1BB,AHX,EPE,BMW-W2EQS,QHH,K2BWR/W3RGQ,
EIS,FBV,TBG,DGM,MSK-W9PNE-W8IL---HEARD only by Dx - K2JIO. -----

- WORKING W/VEs were - G3PU,GGN,KKP,IGW,JEQ,ERN,JVI,FPQ-G5RI,JU,CV,NS
G6GM-G8JR-KZ5PB,DE,FA-KP1SD,KP,CC ,DH-VP7NG,4LZ,9BM,6RG,3AD-
DL1FF,DA-XE1A,OK-YN1AA-GD3UB-KV4AA-HR3HH-KH6IJ-ZS2GE,KZ-CP5EQ-
CO2BL --- HEARD only by W/VE were G3HWI,ILO-TI2BX-EI9J

PY1-15652/Rolf:

Here is all the latest on 160

Thanks for al your fine assistance this year

If you can help me line up a PY HAM that will QSO me

Or make private schedules - would much appreciate

Happy summer and 73 de Stew/W1BB

ROLF - I am enclosing some PUMP
literature that may be of interest -
apparently you are in same line of
business that I am - eh??? CLENZ s/a
BOMBA!!!

Also   am sending you our 1962
calendar via slow mail

Stew

Please be sure not to forget to line up
some HAM who will make schedules with me
on 160 - I might even send small 160 Tx if
necessary if can find reliable fellow - most so
far havent been reliable - did not keep schedules etc and only wand
wanted me to send them somethings.

FEBRUARY 28 1962   W1BB -- 160 METER DX -- BULLETIN   1961/1962 - No. 4

de W1BB/General Remarks --- Well!!! The last regular TRANSATLANTIC DX TEST has co
and gone, and the CQ-CW-WW Contest is over - so no doubt DX activity on 160 will
begin to slack off - as spring approaches!! --- 1962/1961 Season hasnt been as go
as some expected, or as bad as it could have been. This comparison, for example,
total number of different DX stations worked by W1BB each year, might be of inter
however. 54/55-19, 55/56-26, 56/57-20, 58/59-4, 59/60-5, 60/61-11, 61/62-27!!! I
indicates a definite cycle, corresponding with Sun-Spot activity, and improving!!
--- While 160 DX isnt completely done for this season - more DX will be worked -
apparently we have had about the best of it for this season - or have we????????
--- Dont forget the new "Format" wherein underlined calls indicate QSOs, without
underline, indicates "Heard" only. ---- The reports from participating stations,
SWLers are just grand!! Keep °em up and TKS!!! ----- Reference IQSY - it has not
yet been determined if 160 meters data which might be collected, would add enough
knowledge to Ionospheric conditions to make the venture worth-while, as the 50Mc
Data was for IGY -- More on this later. --- There will probably be one more Bulle
this season. ----- Occasionally an error, sometimes due to W1BB fault, and at oth
times not - creeps into this bulletin --- Please excuse the "Human-Element" O.M.s
However on the whole, information herein is believed to be 99 and 99/100 % Pure!!

W1BB/Summary LOG of SUNDAY MORNING 160 METER DX Activities:::
          1. February 4th 1962, FIFTH Scheduled TRANSATLANTIC Dx TEST --GENER
At ooooz DHJ54/2 1800k was coming thru 549, a good omen for Dx, At 0500 449. Quit
a lot of Snow Static - Vx Cloudy - Band opened up to East, Fair to good - South a
W S, Poor. ------- EAST -Fairly good Dx - Gs Coming thru at 0530 peaking 0730, t
loudest signals lingered until 0820. Gs,3PU,6BQ,3ERN,CHN,5JU got across. Exciteme
of the evening was new Country heard from when SVØWZ reported thru G3PU he was Hr
W1BB 0505-0570 - unfortunately no QSO resulted. Reports as follows --G/SWL/Allen:
VP8GQ,1BB,PPN,TX,2FYT,GGL,KHT,8EVL,K8EVG-VE1ZZ-ZC4BP-8HBR/8 - Thanks Geo!!! Then
G/SWL/Smith: VP5FH,1BB/1,HR3HH,2FYT,8NSF,1PPN,2EQS,VE2AYY-UB5WF--VFB Norm & Tnx!!

G3PU/ERN:: VE3BQL/SU/1803, SVØWZ, 1BB/1,PPN-2FYT,IU,KHT-K2PNF-K8EVG,HBR/8-VE1ZZ,
VE2AYY,-HR3HH ------ G2PL/Peter:: 1BB/1-HR3HH ----- "DUD"/G6CJ, of "Shadow-Edge"
160 Dx fame, is back on 160:: VE1ZZ-1BB/1,PPN-2FYT,KHT. Says friend ZL1AH getting
"Hot" on 160 for ‚SO w/HC1AGI & Others ------ G/SWL/IBALL::1BB/1 -----G/SWL/Davidson
1BB/1-VE1ZZ ----- VE2AYY/MIKE::G5JU-VP5BP.Could have wkd VP5BP except he was too
close to 1800k & BC Harmonic - suggests ALL Dxpeditions and DX stay away from every
multipliers of 10kc, 1800,1810,1820 etc - in future!! ------WEST::Conditions poor
A few 6s hrd weakly - no real DX. K7HDB/Dick active, but no Dx.------SOUTH::HR3HH
was star of the evening QSOing quite a few Vs. Real excitement, when HR3HH reported
thru W1BB he was hearing G3PU - BB Notified PU - tried get them together - but no
QSO resulted. This is first G ever hrd by Hal/HR3HH---HC1AGI came on at 0445 CQing
many times - many Vs called him-few ‚SOs-apparently skip conditions poor and static
bad --- A W was hrd clg VP2VL --- XE2OK also on --- BRAZILIAN/SWL/ROLF/ISWL/PY924
::WØGBV,559/239-K8BBI,349 - VFB & Tnx!!! New Country hrd from on 160!!! VP8GQ::VP8
W6KIP,6CDP,5SOT all Hrd VP2ZA ------- NOW HEAR THIS de ZL1AH/John!!!:-"Stew,I am
gradually readjusting my ideas again as the sun-spot cycle takes a dive. I was told
this week that you had a test this weekend so took a couple of looks at the band.
At 0809 hrd W1BB/1 on 1807kc calling CQ-DX until 0810 RST 339 - would have been
better except for summer static" VFB&Tnx John-sure glad you are back on 160!! We'll
all be looking for you!! -------- HC1AGI/"Don"/W3EIS :: W8NSF,GIY,FW,‚WI-K8BBI --
seems like the 8's had it!!! ------- OTHER PARTICIPANTS - W6KIP,8GDQ,HUT-9JDK-ØCC-
VXO/9-VE1DR DX--G2IM-G3CHN,PEK,ORD,LIi,OQT,NXJ,MZW,NXV,GGS,BJD,GXP,NMF,NEE-5JU-6
ZC4BP-OK1GT-GW3CBY,PZU-GW8PG-GI3NZZ. --- HC1AGI says - "Conditions Lousy" !!!

2. February 11th 1962, Non/Scheduled SUNDAY MORNING 160 Meter DX

GENERAL --- Snow Static QRN/S7 at 0430 as snow storm passed thru Eastern USA, cut
down Dx on what otherwise might have been a real good AM - all in all this was a
rather poor Dx morning. ------- EAST - No Dx hrd at W1BB - Marker DHJ54/2 not hrd
G/SWL/Allen::1BB/349-2FYT/239-CN8PZ-ZC4BP-SVØWZ-EL4A-UB5WF -Tnx Geo, --- GSWL/Smith
1BB/1-2FYT-CN8PZ-Tnx Norm!! ---GSWL/Davison:1BB/1-CN8PZ QSOing G6BQ @ 0516/1905
SOUTH --- HR3HH QSO'd W1BB/1 altho going was ruff w/S7 QRN!! --- HC1AGI/Don, reports
"QRN so heacy all reception hopeless"!!! --- VP8GQ apparently QSO'd W3FBV & W2FYV
No confirmation, --- W9PNE/Brice:: VP8GQ/1801/339-HR3HH ---HC1AGI did hear HR3HH ---
VP3AD/"DAD" Daly was on and QSO'd VP2VL589 Tortolla Island, (Box #45 P.O.)-"Dad"
and VL were both real excited for their FIRST EVER VP3-VP2 160 meter QSO!!! Cong
to both - "Dad" thinks 160 good for another month at least. He also hrd, W1BB/1-
2FYT-K8HBR all 569 --- PY/SWL/Rolf-NITEROI BRAZIL!!! Hrd VP2VL/449 & VP3AD/239 -
sure nice to be getting SWL reports from Brazil - not to get a 160 meter Tx going
there!!! Tnx Rolf! ---WEST - W6KIP/349 QSO'd ZL3RB/139 @ 0800 - CERY FB!!!!!

3. FEBRUARY 18th 1962. - SIXTH ANNUAL TRANSATLANTIC DX TEST

GENERAL - Quite disappointingly poor - of all Mornings - on the last, 6th Annual
Transatlantic DX TEST - and also the ARRL Dx Contest to have the QRN "Gremlins"
work spoiling things!!!!! SKIP, all directions, pretty good - however, S6 to 8
QRN covered most DX stations with a heavy blanket - nevertheless, a few lucky Sts
did succeed in breaking thru for QRM QSOs. ------- NORTH - V1RAN/FP8BX DxPedition
expected on - did not show. ------- EAST - The European Dx conducted a battle royal
with QRN, pushing thru occasionally to QSO on "Peak" conditions. G3PU:VE1ZZ --
G6BQ:VE1ZZ - G3FPQ:1PPN -G6GM:1BB/1 - xxxxxxxxxxxxxxxxxxxxxxxxxxxxxxxx - W3EIV:G6BQ -
W1BB/1:G6GM,BQ,G3FPQ - VE1ZZ:G3OIT and others --- GSWL/Allen: W1BB/1,PPN-K1DSV-
2EQS,KHT-3EIV-K8OR-VE1ZZ-VO1FB-HC1AGI-VE3BQL/SU-SVØWZ - Fine list Geo!!!/Tnx!!
GSWL/Smith;1BB/1-2KHT-VP5BP & other unidentifiable signals Tnx/Norm!! ---
K4KSY/Drayton hrd only DHJ54/2 weakly at 0330 - QRN S7/8 --- GM3IAA/Jim MacIntosh
Hrd W1BB/1 - says "It is strange why our Northern stations do not get across so
on 160!!" Their strictly low QRP has something to do with it no doubt!!! Jim used
to operate a TELEFUNKEN 5 KW Spark Tx at Assiut Egypt - some difference!!! GSWL/
DAVISON:1BB/1/569-1PPN/459-2FYT/449-VE1ZZ/559 - Tnx Henry!! ------- SOUTH - VE3CO/
Noel Eaton, Vacationing Canadian Div. ARRL. Director broke thru as VP5BP, Grand
Cayman Island Working W1BB/1, 3RFA, 4YHD, 8HCW, and probably others for NEW COUN-
QSOs. (On Saturday Night 17th QSO'd W1PPN). --- DON/HC1AGI/W3EIS, apparently was
having QRN problems too. Many W/Ve frantically and frustratingly calling, 0730-0
w/no Ans. However he did QSO W1BB/1,2iu,MSK-4KFC,YHD-6KIP-K6ZH-9PNE-KH6IJ - VFB
WEST - Practically NO Western Dx heard at all except W6KIP and only for brief time

ADDITIONAL PARTICIPANTS - W/VE: 2BWR,SSC,GGL,QHH-3GQF,CJ-4OMW-K4KSY-8GDQ,YSD-K8H
78,1QO,RRH ---Dx: VP8GQ-GI6TK-GM3NHO-GM3IAA.

FEBRUARY 25th 1962 - CQ-160-WW-CW-CONTEST --- GENERAL - Hoped for A
Contest/Dx conditions did not materialize - principally due to high static levels.
Much Dx was in there but could not be copied or worked thru QRN - although some, w
All in All, a good TEST was had however - everyone seemed to have a grand time, an
a lot of new stations were heard from!! CQ will carry the results. W/VE - W1TX/Roy
FP8BX,WØVEH/VP9-HC1AGI-EI9J-HR3HH & others for score of 10,430 in 11 hrs -Not bad-
had bad 60 cycle line noise!! ---W1BB/Stew: VO1FB-WØVEH/VP9-FP8BX-HR3HH-HC1AGI-G3C
GD3UB.--- W2UKS/Bill - KH6IJ-EI9IW-G6BQ-WØVEH/VP9-HR3HH-GD3UB-XE2OK-FP8BX - Tnx Bi
(Not CSO(d) ---VE2AYY/Mike:-"Had a ball"!! 152 QSOs including FP8BX-EI9J-HR3HH-
WØVEH/VP9- now has 10 Countries on 160 - says Cdx poor to West. Hrd XE2OK,GD3UB,G3
Score 11,760 ----- Willie/W8GDQ - had some fun, but snow static terrifgic - missed
few good ones. ------- EAST - A LOT of Europeans were on for the test, but between
QRMing each other, and covering up DX, they had a lot of trouble with the QRN also
Gswl/Allen:W1BB,FRR,MR,PPN,TX-2EQS,FYT,IU-K2BWR,DGT-3ECR,EIV-4KFC,YHD-8GDQ,HGW-VE1
VE2AYY,3KEE-VO1FB-WØVEH/VP9-HR3HH-EL4A QSOing Gs & OKs but no Ws - Nice going Geor
G3PU/ERN, says "Withrout doubt, best opening of season BUT ruined over here by lot
Gs▪▪▪ in Contest calling "CQ-DX" on top of W/VEs - they were even on W1BB's Fy ev
when you yourself were calling CQ or working another station - I cant understand
some of these chaps!! On top of that, W stns did not seem to be tuning our Fys - I
packed up after 50 Mins fruitless calling!!! Ern heard: W1BB-2FYT,HUG,IU-K2DGT-4KF
4YHD-8GDQ,HGW,NSF-K8HBR/8-WØVEH/VP9-VE1ZZ,2AYY-VO1FB-HR3HH-FP8BX!!! --- G3CHN/Roge
W1BB,ME,PPN,FRR,2FYT-3GQF-4KFC,YHD-EL4A-EP2BK-OK1ZL,NR-UB5WF-VE1JX-HB9T-ZC4PB-VO1D
VØVEH/VP9 589 at times!!! and many others hrd - however "T"was a bit of a carve-up
at times" ne says but very enjoyable, bringing in-no less than 4 new ones!!! K8HBR
HC1AGI-HR3HH-WØVEH/VP9-KH6IJ-XE2OK. Score 12,792 for 152 QSOs - 3 Dx and 39 Mult.
Hrd no Eu in FEB - Al leaves for his job - sailing on the Great Lakes, shortly - T
loaded ship antenna that he appropriates for 160 DX goes with him - best luck Al!!
Gswl SMITH:W1BB/1,AWP-2FYT,UWD,IU,GGL-K2DGT-4UK-8HGW-K8HBR/8-VE1ZZ,2AYY-9PNE-WØVEH
Norm also sent in FB list for 24th also - tnx & Congrats!! ---- CYPRUS HEARD FROM
AGAIN ZC4PB/GERRY BATEMAN - With low power and poor antenna - single wire 15-20' H
made 8 Eu. QSOs Hrd Between 0222-0430z the following: WØVEH/VP9,W1BB,PPN,ME-4YHD-3
3GQF-8HGW-K8HBR-No VEs hrd. Stan called Ws using 1827Kc but ND - Unfortunately - T
a million Stan!!! ----W1PPN/Cliffe: G3CHN-GD3UB-G3IGW-EI9J. ------ WEST - W6YC/Gen
using 235 ft LW on small 30x120 lot QSO'd 25W States, VE,KH6 & XE1 - heard W1 & VE
and also HC1AGI/579 but could not QSO him!! ---- NORTH - Nil!! ----- SOUTH - HC1A
::1BB,TX-K2PNF,DGT-3GQF,EIV-4AE-8GIY,GDQ-K8HBR/°-9YT-8GBV-VE3KE-XE1OK-HR3HH-YN1AA.
"Don" says his problem hearing DX - his European listeners mail is terriffic - but
he doesnt hear 'em - "Two whole nights without sleep - wonderful conditions - but
only 59 QSOs!! Don corrects BULL No.3 - HC1AGI is 11,700 Ft high - not 1170!! ---
OOOOooooOPs!!! Sorry!! (Don also had nice string of 23 stations QSO'd on 24th) ---
NEW COUNTRY HEARD FROM - PJ2AE/Stan - Who says he cant operate on 160 - regs - but
thought we would be interested in signals hrd, as follows:: W1BB/1,PPN,ME,MMN-2FYT
NVD,IU-K2DGT-3WV,GQF-4YHD-8HGW-K8RRH,MBR/8-9YT-WØVEH/VP9-ØBDS. We certainly are
interested and Tnx millions Stan - and get after those PJ Authorities to allow 160
ADDITIONAL PARTICIPANTS None not already mentioned!!!

MISCELLANEOUS TOP-BAND ITEMS ------- W6KIP/Alex QSO'd HR3HH & HC1AGI - Is disappoi
there is so little DX to West. Has "Rectification" Problems with his antenna tower
causing QRN - says "Noise here remains tremendous problem and I dont know WHAT I▪▪
I'd do without noise balancing effects of receiving antennas - although it cuts do
on signals - it still gives much better signal to noise ratio, and I seem able to
all stuff fellows in QUIET spots hear"!! -------GSNF/FLASHBACK!! Jan 7th Harold Hr
VE2AYY/57-VE3KE/57-VE1ZZ/57-W1BB/57-HC1AGE/33-W2UWD/45-K3KMO/45-W8IC/33-3NTX/33-1C
▪▪▪▪- Tnx Hal! -------▪▪ GM3IAA/JIM, A real O.T. -1914-1915 R.E. Wireless Ass'
O.T. - Heard Jan 7th 1BB/1-569,1PPN/349,1TX/349. He is Ex VS1AA-2AA,2AF. In 1934
improved WINDON antenna which is written up in RSGB handbook -- He is looking for
Contacts!! Nice to have you with us Jim - Good Luck!!! (Has just wkd 1BB & 1PPN!!!
--- BRAZILIAN/SWL-ISWL/PY9247-Rolf Rasp --Checks in with Vy nice ltr & fm now
will log 160 stations and QSL - has been HB9SWL 5 yrs - also DL,PK1,PK3 - has conv
into 10 tube BC/Rx Home Made. Sure nice to hear fm PY - Tks!!! ------VE3DU/Dave
is on 1898 Kc every Sunday 0500 - says quite a fw VE past 1900

W∅AIH/VE3 - Loran does not bother there (Lucky boys!) However TV puts birdies 1825
1827 messing up Eu Dx-Thinks Cdx not nearly as good this year as last (Dont underst
that dave!!!) Jan22:G3CHN,HR3HH,HC1AGI - Tnx Dave!! -------- W6KIP/Alex sez Europea
who want to work W5,6,7, MUST use 1800-1805Kc - hopes for 160 in KL7 someday - Try
for ELAD & VO∅LL on 160! -------- K7HDB,JAN27:2FYT,UWD rapid QSB -Jan 28 1BB/1 ---
Does anyone have a "Yen" for WHALE MEAT??? write Dick!! -------- W2TR/Geo -"We dont
Hv much time for HAM wk, but when we do, it is the OLD TOP BAND, although things a
rough at times, we enjoy 160 meters best of all!!! --------K4KSY/Drayton - "Notice
that West Coast stations come in hr quite early or late - sigs break thru literally
everything is quiet on top band - then suddenly they appear-early as 2030est-sever
nights exceptionally good ∩5 low as 1980kc - Alas!! cant seem to wk °em!! Does thi
fit with study E/W Cdx?? ----- WE ARE SORRY that one of our most reliable & well K
160 DXers BRICE/W9PNE is plagued with ulcer trouble requiring strict diet - but wo
of all enforced sleep schedules that cut out much 160 Dx activities - better obey
"DR's orders" tho - get it over with - and then BACK TO 160!! Best luck Brice Fm a
of us. -------- W∅CDP/BING says skeds w/ G3FU,6HB-EL4A-DL1FF on 1999 did not wk out
Hrs VP8G∎ well on 1827 - HC1AGI best on 1995 -------- OH3NY/Matti - Continues effor
to get FINLAND Govt allow 160 agn - Keep up the good wk Matti!!! -------- HC1AGI/Do
rpts FB string QSOs Jan 14 & 28th --14th:1BB,ME,2KQT,FYT,IU,WA2KPH,8GDQ,JIN,K8I,Q,
HBR/8,K6ZH,KH6IJ,∅NWX,CDP,IFH --28th:1BB/1,2IU,FYT,UWD,GGL,K2LWR,3RFA,G∩F,6KIP,8FG
FW,IQSmLOF,HRV,NSF,HGW,K8HBR/8,∅EUQ,CDP,IFH,VP8GQ,VE1ZZ,ZL3RB!!! W∅AIH/VE3 --------
EL4A/Ken - reports via Ham Radio "Been hvg trble w/Kw rig but expect hv fixed soon
-------K7ICW/Al - has new Hy-Tower coil loaded Vert. 160 ant!!! Says DX poor this
but QSO'd KH6IJ-line noise teriffic - going to hit CQ-WW-160-CT hard this yr agn!!
--------V2UKS/QSO'd VP5BH/1804Kc Feb 19 - Hrd FP8BX/1827k,20th-sez QRN HIGH?! -----
SORRY TO HEAR?W8ANO/Waldo - has sold 160 meter Tx&Rx-gone SSB!!! He'll be back - s
day!!! -------Willie/W8GDQ Moans that cdx havent been good as expected this yr & w
power limit/line noise etc has had rough time - sorry!!!!

BOY oh BOY/V2UKS/Bill!!!!!!!! Feb 21st truned in an incredible list Dx Hrd - UB5WF,
∩0∎∎∎VP3AD-HC1AGI,HR3HH,G3CHN & Others ---W1BB sed to himself - there's something
here worth knowing - SOOOoooooo wrote Bill, asking if by chance he had real QUIET
location - and here's the answer of 160's UTOPIA!!! "I am Located on an island 2½
miles out in the Atlantic Ocean Fm mainland -plenty water all sides-plenty deep -
noise level S2-S3 normal - in Loran S5 - we have NO industry in or nearby/not allo
I'm using Heathkit "Q" Multiplier HD-11 w/SX28 & L.W. 200' long over large parking
lot & 40° high. QTH center of city w/4&5 story hotels arnd - resort city/fairly qu
in winter all directions except N - Ant runs E/W-end into shack - ant coupler-Viki
II Tx/260w--sure can hear °em/darned if can wk °em!! Wl keep trying (And he DID!!)
"Ed-Note" -W1BB also had superb reception when band was open one Sunday AM, and he
was at Summer QTH 300 miles into Maine away from noises - hearing and working UB5,
ZC, many others - Dx sounded like 20 meters - all of which goes to prove that the
most important ingredient for 160 Dx is a quiet QTH in the clear.

PROPAGATION - We 160 Meter fellers are especially interested in this - and so for
everyone's benefit we copy herewith all the latest dope - Courtesy "FLORIDA-SKIP".
"PROPAGATION CLEARED UP" -!!??- Propagation charts show an easing up of the rate a
which Dx is easing off, This is proof of the slowing up of the slowdown. ---- It
should be noted that the slowing up of the slowdown is not as good as an upturn of
the downturn which is a good deal better than a speed-up of the slow-down. It does
suggest that the propagation is about right for an adjustment of the readjustment!
---Turning to long skip, we find a definite decrease on the rate of the increase,
which clearly shows a letting up of the letdown -----Of course, if the slowdown sh
should speedup, the decrease in the rate of the increase of short-skip would turn
into an increase in the rate of decrease of long skip. In other words, the decel∕a
would be accel∕erated. ---The charts suggest a leveling off, followed by a gentle
pi∎∎up, rather than a faster pick-up, a slow down of the pickup, and finally a
leveling off again of the pickup. --- At any rate THE PROPAGATION IS RIGHT FOR A
PICK-UP this winter - that is, if you have a 10 Kw rig and a 25 Element beam!!!°"
forecaster Johnny Short Skip/W4SKIP. --PS - How's DX fellows??????? There you have
it fellows - and this information should be VERY helpful in working 160 meter DX?!

       73 de Stew/W1BB and HAPPY HAMMING to you - Especially on 160!!!

PYswl/Rolf

Here is all the latest on 160 for you with compliments W1BB

Thanks for yourletters March 8 and 19th

The reports were very interesteing and studied carefully and entered in SUMMARY log here.

Do not know VP2VL QTH - suggest you write VP2VI W2YH T.W. WINTERNITZ, Yardley Road, Menham N.J. - He will know or DX Magazine, Don Chesser, Burlington Ky - they will surely know. Sorry I cant help more.

I sincerely hope that you can get someone on 160 for schedules with me next fall - or send me likely suggestions to write to.

That SX71 is a real fine receiver - VFB - Know you will like it,

VERY VERY VERY FB on you and PY1NFC -- and with 40 watts if you can build it to cover 160 we can QSO - and make HISTORY - since there has never been a PY/W QSO on 160 meters - and if and when you are ready if you will make private schedules with me , you can win the W1BB CERTIFICATE for FIRST QSOs - I will give you all the needed information on Antennas - and help you in any way possible - just let me know what you need - as long as you will make the skeds with W1BB only until a FIRST QSO. And I assure you that I have a sustem that will work if you are faithful and persistent!!!

HAPPY HAMMING and 73 de Stew/W1BB

W1BB
STEWART S. PERRY
36 Pleasant Street
Winthrop, Mass.

APRIL 18 1962   W1BB -- 160 METER DX -- BULLETIN   1961/1962 - No. 5

de W1BB/General Remarks --- This last 1961/1962 season BULLETIN has been delayed a bit for unavoidable reasons, but the enclosed FLASH supplement was sent to all the Ham Magazines on time. This will be the last Bulletin until the 1962/1963 season opens this fall - unless by chance there may be enough activity to make a summer Bulletin necessary, ------- Yes - it HAS been a GOOD season, a VERY good season on 160!! with indications of possibly better next!! We say "Season" because during the winter months there is more activity and Dx worked in this latitude - the band is quieter, nights are longer, and weather males it more conducive to staying indoors at the rig. HOWEVER, we must not forget that while it is summer here, it is "Winter elsewhere and there could be Dx minded Hams willing to work 160. In fact, W1BB wkd ZS in August, and CP in July. Most of us Northern Hemisphere people are so happy to enjoy our outdoor activities in the warm summer months, that our 160 Dx, generally speaking will be QRX until about Oct or Nov. Another advantage however, these periods of activity and rest give more ZEST to the Test when it resumes!!! SOOOooo generally speaking QSK till fall and b.c.n.u. then. ------- We cannot end this season, (or any other), without saying many many thanks and appreciation to the SWLs whose prompt and complete reports have added so much to our 160 meter tests. Outstanding, as usual, were reports from G/SWLs, Robert Iball-Norman Smith-George Allen,--Excellent reports from SWLs - Gs/John Hall, Henry Davison, Jim Sleight -VK4 Charles Thorpe- PY/Rolf Rasp -- not to mention a number of others who sent in spot individual reports. Next year, it is hoped to make available to any W/VE Hams who are interested and write for them, copies of these reports. "Cheers" to these boys - and fellows,keep up the good work!!! -------W1BB wishes to send thanks and 73 to all the 160 meter gang with appreciation for their QSOs, reports, cooperation and all matters which have led to greater enjoyment of DX hunting and testing on

W1BB/Summary LOG of SUNDAY MORNING 160 Meter DX activities!!!!

MARCH 4th 1962, Non/Scheduled SUNDAY MORNING 160 Meter Dx --- Generally speaking this was a dandy morning for Dx to Europe - Stations "in" for 2½ hours!!

W/Ve-REPORTS:- W1BB QSO'd 11 stations!! GM3IAA-G6GM,BQ-G3OIT,IGW,RJA,GWX,PU,FPQ,E
OQT - some up to rst589!! ---W1PPN Topped W1BB with 13!!! GM3IAA,WØVEH/VP9,A3-G3R
OQT,KOX,PU,ERN,FPQ,IGW,OIT,MBN-G6BQ- also AM/SSB QSD with G3FPQ - Cliffe on AM
Congrats Cliffe!!! --- VE1JX:: -G5JU,3CHN,IGW,FPQ --- W8GDQ/W1llie::G3PU,W6KIP,"H
Eastern guys working stuff but couldnt pull 'em through"!!! ---W2UKS/Bill- Hrd Onl
W6KIP,WØVEH/VP9,G3PU,G6BQ ---WØAIH/VE3 QSO'd G6BQ ------(EAST)- VE3DU/Dave - G3CH
PU,FRQ,6BQ says "Good opening but plagued with S6 intermittant line noise and coul
NOT make QSO through it"-Sorry Dave!! -----G/swl/Smith::1BB/1,PPN-2GGL,FYT,IU,UWD,
KHT-8NSF-K8HBR/8OVE2AYY-VE1JX-"No exciting stuff,a W did call a VP" -Tnx Norm!! -
Gswl/Allen::says "A very good time was had by all-best condx since I came back on
trail!. Spent much time on 1.9-2.0Mc hrd nothing-though pptomistic-Willie/8GDQ doe
have the punch of old-wonder if his VERTICAL deteiorated?? -(NO!! Power limit now
50w instead of 200 his area) - anyway real nice to hear him around again" -Hrd, 1B
1/589,AW/349,TX/449,PPN/559-2IU/339,FYT/559,UKS/449,GGL/349,TR/459,KHT/349,UWD/459
K2BWR/229,K2HLI/239-3NOQ/239,GQF/239-K5TMF/559-8GDQ/469,NSF/339,FW/358-K8HBR/8/339
VE1JX/559,2AYY/449-WØVEH/VP9/239 - Tnx Geo!! -----G3OJI/Jim had an exceptionally
list 1BB/1,PPN-VE1JX-K8HBR/8,8NSF-VE2AYY-W2IU,FYT,6FW,K2HLI,2UWD TNX!! -----Gswl/1
1BB/1,PPN,2FYT,VE1JX -----G3PU/Ern - "Without doubt best morning of season"!! 1BB
PPN,2EQS,FYT,GGL,IU,UWD-8GIY-K8HBR/8,8NSF,WQV,VE1JX,VE2AYY,VE3ABG -----Gswl/David
W1BB/1/569-VE1JX/569-W1TX/449-2FYT/239-VE2AYY/469-1PPN/459-8NSF/339 - and proud o
his FIRST W8!!! -----G3OQT/Dick - Hrd 1PPN QSOing/569-WØVEH/VP9/439 CW then chang
to phone hrd Cliffe 3/5 and VP9 2/3 - also 1BB,AW,2FYT,IU,8GDQ,1PPN, VE1JX,VE2AYY
WØVEH/VP9. -----(WEST):N11 -----(NORTH);N11 ----- (SOUTH)- ZL3RB reports,"condx
unusual, extreme slow QSB,sigs rising to medium peak slowly, dying right down thru
period of 5 minutes. ZL3OX hrd W6HRG/459 -We logged W6KIP,5SOT,6HRG,8CDP. It gave
ZL gang real thrill hearing Stateside clg "CQ-ZL" and we really appreciate the Ws
looking for us. Participating were ZL3OX,ZL3BC,ZL3RE,ZL3CO and ZL3RB"- Tnx ZLs!!!
Another participant this morning was XE2OK.

MARCH 11th 1962 - Non/Scheduled SUNDAY MORNING 160 meter Dx --- Early
ski condx seemed fairly good and the band did open up fairly well. -----(EAST)-1B
hrd EP2BK briefly 119-239 0310z,later0335z G6BQ came thru 579 QSOing W1BB. At 0500
few other Gs hrd,G3PU coming thru real well. WØ/AIH/VP9,QSO'd a number of Ws ---
G3OJI/Jim-"Cdx rather poor-Please ask W2FYT NOT to send QSX 1801kc as many Gs star
calling there and make mess of Dx band Hr - HC1AGI/VFB!! Hrd1BB/569,HC1AGI/559.3G
439,2FYT/569,1PPN/339 --- G3FPQ/Hrd HR3HH ---Gswl/Smith:1BB/1,2FYT,IU,3GQF --- de
G3PU/Ern:1BB,PPN,2EQS,FYT,IU,3GQF --- G/swl/Anthony Quest;1BB/449-K2DGT/339-3GQF/3
W9PNE/Brice:: says his son had emergency appendectomy this night so vy QRL,however
did hr G3PU/229. Brice feeling better himself - using NAVIGATOR/50w. CQ/CW Contest
105x38 ---Gswl/Allen:1BB/1/349-2UKS/229-4KFC/229-K2DGT/349/3GQF/459-2EQS/239 -----
GM3IAA/Jim:1BB/1,PPN,3GQF all 349 - Jim is planning new antenna for next year !!!
WEST:;;;; No Western Dx hrd hr or being wkd by others ------(SOUTH)::: Real exci
ment when W1BB alerted Gs to fact that HC1AGI was on 1826 - thereupon G6BQ QSO'd
for FIRST!!!. G/HC 160 Meter QSO ever. HC1AGI gave G6BQ 449. Congratulations!!!!!
In addition HC1AGI QSO'd a vfb string K8HBR/8,3GQF,2EQS-This one the reward of man
hours of trying and a new antenna by Chas -2GGL,K8HUT,2UWD,1PPN. Don hrd & Cld G3
several times but he did not answer. Interesting that G6BQ who was alerted could
Hc and G3PU also alerted at same time could not - proves that 160 meter skip can
be either "in" or "Out" because of short distances separation between receiving QT
HC1AGI hrd HR3HH/579 and W1BB/579. ----- HR3HH had Fb sig QSOing W1PPN,2GGL and m
other Ws -----W9PNE;HC1AGI/449,HR3HH/569 -----(Other PARTICIPANTS)- Ws,1TVN,2DGT
TR,UKS,4KFC,MUTK/1,8HUTk,FW,GDQ,HBR/8,TYSk,FW,GDQ,ATPwa,VLO,ØTUT --G3ERN,LIQ,PGN,GM3
OK1NR, SWL/Allen,Smith.

MARCH 18th 1962 - Non/Scheduled SUNDAY MORNING 160 Meter Dx ---(GENERA
W1BB had to go to Maine on a business trip/spent one night at old Homestead 300Ml
N. Boston in quiet QTH - took SX100 and Ranger/50w along. Too busy to spend much
time on air but did QSO EI9J and Hrd G3PU,ERN-6BQ-OK1ZL - what impressed most, wa
the quietness of reception. At W1BB's home QTH Boston, the noise level reads arou
250 to 300 usually, while at this quiet place in Maine it was only 140 to 160 - o
about HALF and DX signals sounded just GRAND!!! Imagine what a difference this
makes in weak signal reception - to get away from the Electronic Clouds that
surround big cities and communities, just like the "Smog" in California - or Smok
and haze -same idea. Ph for these Island and remoht QTHs for 160 meter Dx!!!!

(EAST) - EI9J:W1BB/1øMaine -G3ERN QSO'd UO5AA - Other QSOs were made. ---G3PU/Ern
"Heard only Wø I have hrd this season WøNWX,1802kc 0610z rst229/459. Also hrd weak
Wøon 1999 believe WøCDP. Records!!";;;1PPN,TX, 2EQS,2IU,UKS,UWD-3ECR,GQF,øNWX - VFB
W2UKS/Bill::G3PU,XE2R&C,HC1AGT & QSO'd W5SOT!! ------ Gswl/Davison::1PPN/459, thru
se re QRN ------ Gswl/Allen::K2DGT/339-øVEH/VP9/349-K2BDM/339-1PPN/449-3GQF/449-
2IU/329-2UKS/339 ------Gswl/Smith::2DGTk,1PPN,TX-3MSK,ECR,GQF-2GGL-IU ----- (SOUTH)
VP3AD/Dad hrd 1PPN/449 0432 and W3GQF/449 0450 other signals in there but QSB and
static made copy difficult so closed up early "Guess 160 has had it for this seaso
Tnx Dad and b.c.n.u. Next for sure!! -----(WEST) -----(NORTH) ----- Other PARTICIPA
Gs3OIT-GI6TK,VO1FB -W/VFB: K1TKK,6HRG,8UDR,BHUk,øNWX,TUT,LQTk ---SWL/Davison,Allen
Smith. Activity and Participation dropping off as Spring/Summer approaches!!!

MARCH 24/25th 1962 Non/Scheduled SUNDAY MORNING 160 Meter DX Activitie

(GENERAL) - Dx participants dropping off - QRN very high due to snow/rain static - b
skip generally quite good -W1BB QRT at 0555 -----(EAST)- W1BB got on early again f
schedule with EP2BK - hrd each other but no QSO. Also made contact March 22 and 24
0220z approx. Static very heavy. W3GQF and W2EQS tried for Bob/EP2BK also. W1BB hr
him 119/229 to 000. W3GQF hrd him 349 W2EQS could not copy. Skip not as good as ni
before when W1BB not solid QSO resulted. QRN Vy high also - at 0320 G6BQ came thru
with CQ QSOing W1BB,his 449/BB559-his signals at 0400 still 449 QSBing in and out-
QRN high. W2KQT also took part DXing QSO'd G6BQ/Jack getting 349/449 and giving 46
W2IU QSO'd G6BQ getting 33/49 Giving 469 ----- Gswl/Norm Smith:: 1BB/1,AW-2KQT,IU,
K2PNF-3FBV,GQF says "One or two weak unreadable Ws including a W5 at 0617"!! Norm
going to slack off for the season - thanks for some wonderful reporting and Assist
Norm OM. -----W2UKS rpts no DX hrd "Appears that springtime is here" -----(SOUTH)
(WEST) -(NORTH) (Other PARTICIPANTS) -2PSQ,4VWU,8CJN,9ZQOk,øCDP-G3ERN-SWL/Allen & Smit
----

COASTAL COMPARISON NET -(West) ----- K7HDB/Dick writes "There is a tremendous diff
renœ between signal levels at the Northern Vs Southern stations and I would like
ge some sort of NET of Coastal stations going to compare reports of signal level
during these DX tests in an effort to determine the Southern ends of this zone of
absorption or whatever it is that is creating the relatively poor Cdx at the more
Northerly stations. Mainly interested in stations close to shore where Loran is li
limiting factor in reception rather than power line noise" Those interested, write
Dick - sounds like VFB idea!!!
INTERESTING !!! W1EAE/Bigelow - of EMN CW traffic NET says that on 80CW with lower
sunspot Numbers, LONG skip, and short Ground wave are setting in to such an extent
on 80 that it is virtually impossible to conduct their normal NET. Will have 8 to
stations on and NET Control can only hear 2 or 3 - they are eyeing 160 for possibl
solution to their traffic NET Meetings 6.30-7.30 PM EST. Best of Luck Big and boys
ZL STATIONS -W6KIP/Alex de ZL3RB/Mick, says more ZL activity. Look for ZLs 0720/08
z around 1975-2000Kc. ZL3BC/Joe,ZL3CO/Garth,ZL3OX/Jack & ZL3RE/Art - all active he
Also ZL1AH. This one FOR THE RECORD!!!! "ZL" --FIRST !!! ZL/W QSO ever was between
ZL1AH/John Eightman Dec 31st 1961 and W9CVQ - the Second between W1BB and ZL1AH AU
16th 1952. ZL1AG has two other FIRSTS!! ZL/G and ZL/VS6. Following closely on thes
ZL3RB/Mick became active and due to John's absence, and his own splendid efforts
quiet QTH has been the mainstay ever since. We understand ZL1AH/John will reactive
PROPAGATION de Rod Newkirk W9BRD- "The North Atlantic propagation path to Europe i
highly susceptible to Magno-Ionospheric irregularities which do not significantly
affect South American & Pacific Paths according to Termann" Guess this accounts fo
those shut-outs while HCs & KH6s BOOM in!!! Thanks a lot Rod - Most interesting!!!
SILENT KEY VP7NM/Chas - 160 Meter Hams especially, are saddened by the death of Ch
"C.N." Albury. Many of us have his card confirming 160 QSO and are proud of it. He
put the BAHAMAS on the 160 DX map. His presence and activity will be greatly misse
IN "ZL" LAND EXCITEMENT WAXED HIGH Per ZL3TB/"Mick" and W6KIP/Alex who report that
during the CQ/WW/160/cw/Contest, Mick/"There were so many 160 meters signals on bo
ends of the band QRMing each other from 0700-1200GMT - I never thought I'd live to
hear the 160 meter band dish out signals like it. Although I was the only one to
contact W6KIP, - ZL3CO,BC,OX all heard an earful. Stations were too numerous to lo
stronger signals were W6KIP,WA6CDR,HC1AGT,W6HXT,W6HRG,K6SHG,W8HG,WøGBV - all audib
to 1200 when I fell asleep!! - 5 hours of Dx signals without a break or fade -Wow!
ZL3OX Hrd 6KIP,K7HB,W6HRG. On Sunday Band was poorer but I hrd WøNWX,2PIQ,W1BB/25
Mick HAPPY It sure would have gladdened your heart to hear that crowd - FUN!! St

160 DXPEDITION ????? --- K4KSY/Rev. Drayton writes "Wonder if any of the gnag woul
be interested in a DX-pedition to one of our Barruer Islands down off the Coast, W
lived there a couple of years and found conditions absolutely ideal!!! The place i
EDISTO Island - 1½ miles off the lower South Carolina Coast. ---Heard Low Fy Dx
th re that is considered practically impossible elsewhere - European BC stations
consistently - "Scads" of stuff on 160 and 80 -- Next winter possibly some of the
Yankee group going South might like to stop off for a weekend - have access to an
authentic ANTE-BELIUM Mansion just 20 feet from the salt water and poles already
set for antenna. Only power line nearby is the small one to the house. Nearest
neighbor is 2½ miles away. Perfect for anyone DXing in a 50 watt way!!!" ---Anyone
drooling besides W1BB??? Thanks a million Drayton and We'll see!!! (Incidentally
watch for some 160 meter dope in "73" Magazine in near future!!!!)
MISCELLANEOUS TOP BAND NEWS ITEMS -------K7ICW/Al, had a "Ball" in Dx Contest - wk
74 stns/29 sections including HC1AGI & KH6IJ for Total 4500 Points - Hrd XE2OK,
W0AIH/VE3 and says "These 10th wave verticals are for the birds - altho better'n N
NUTIN!!! ------W2UKS/Bill, in that ideal 160 meter QTH on an island, has 2 feet o
seawater all over the place during that March Northeaster ------Steven GIBBS/GM5M
of Ex ZC4AK fame!!! reports from Benghazi, Libya, hearing on Feb 11th, 4 Continent
VE3B/L/SU,ZC4PB,SV0WZ and W1BB/1 plus a lot of other Europeans. Conditions not too
good however/ Steve is 160 meter Dxwise famous for giving two DXers their 160 mete
WAC. W1BB/ZC4XP, March 3 1953 and EXACTLY three years later (As he points out) -
W8GDQ/ZC4AK March 3 1962 -,quite a coincidence - thise dates eh????? Thanks Steve
you prove that it is impossible to keep a good DX man down and we hope to hear you
one of these days as 5A sumpin or other!!! ------Kn3RKD/Ron/-3/10/62 heard W1BB's
1809kc CW 579 on a 3 band citizens USL transceiver converting into a BC radio on
dual conversion!!! Can you figger that one out??? ------- HC1AGI/DON/(W3EIS) lists
CQ/CW/CT QSOs on 2/24: W2FYT,K2BWR-3RFA,ECR,4KFC,YHD,SQE-5DWB-6KIP,HRG-K6HXT,WA6CI
7UBI,K7ICW-8HGW-0NWX,CDP,EWH,GJI,SDN-9WJB,YTV-KH6IJ ---2/25: 1BB/1,TX-K2PNF,DGT-
3G T,EIV-XXXXXXX,4AH-K6PJY-8GDQ,GIY-K8HBR/8-9YT-0GBV-YN1AA-VE3KE-HR3HH-XE1OK, and
Co amts Don "Called 10 U.S. guys for each one worked - never heard so many TIN es
in my life!!! ------PY/SWL/Rolf Rasp Feb 25th hrd 6KIP-K6HXT-4KFC,HYD-K2IUT-6HGW-
KSRRH-9TGB-0GBV-KH6IJ-HC1AGI- also unidentified VE2 - excellent Rolf-Tnx!! ------
W0AIH/VE3 Rev Paul - Has new antenna ideas for 1962/1963 - hope summer Wx very
conducive to installing same!! QSO'd KH6IJ in CW/CQ/WW/160/DX Contest - also G6BQ
March 4th - CONGRATULATIONS!!!. ------Sven Elfving/Polar Bears Radio Club/SWEDEN
Hrd following Dx this season: VE3BQL-HR3HH-OX3CB-EP2BK-W0AIH/VP9-VP8GQ-VP5BP-ZC4P
OK1ACU-UB5WF-HB9QA-VE1ZZ-GI6TK-EI9J-W1BB-5N2GUP-HC1AGI -- VFB Sven - and your SWL
Bulletin also very nice -Tnx!! ------ G3PU/ERN CONGRAYULATIONS!!!! His 160 meter
Dx Score at the end of the 1961/1962 season is WAC, 37 Countries!! Greatest
disappointment this year -,almost but not quite QSOing HR3HH. On Feb 18th Ern QSO
VO1FB and Hrd VP5BP ------- THAT 160 Meter ISLAND Dx PARADISE we mentioned last Mo
W2UKS/Bill's Town OCEAN CITY N.J. sure took a pounding in N.E. Storm in March - se
"We've really had it with terrible storm and excessive high water mark - hundred
of cars a total loss -250 homes damaged beyond repair - dozens of homes washed out
to sea - dozens destroyed by fire -loss of life 5 including C-D worker, Chief of
Police - I have been helping with Emergency work 75 phone and CW. My own car had
$300 water damage - what a mess!!!" ------- Bing/W8CDP reports 1961 CQ/160/CT,
160x43 - 1962 163x45 including HC1AGI,KH6IJ,XE1OK but gripes!!! "Ws too busy on l
end NOT covering high end, comments, I am thinking of a strictly 160 meter station
and location with special low noise receiver" -- Bing pleads for boys to cover TO
END of 160 says "Istill believe I can get signals across to Europe from We if tho
fellows will just listen - this view also shared by W5SOT,DWB,6KIP Etc. I'll be o
1998 each weekend"!!!(See G3PU Report for March 18th) -------W9PNE/BRICE, heard
EP2BK/449 during CQ/CW Contest QSOing EI9J - thought he was a pirate so did
c l him - alas!!! Contest 105x38 VFB Brice. ------ K7HDB/DICK - during CQ/CW/CT
blew H.V. Transformer, of all times!!! Quite interested in ITBY reporting - says
scientists find "Light radiation belt similar to Aurora Borealis" Over Oregon --
believes that this may account for blanketing 160 there - Could Be!!! This ioniz
belt goes apparently around the Globe - wtite K7HDB for further info if intereste
PY/SWL-ROLF RASP -Still listening - no Dx lately ------G3QLT/Dick - is to be
Congratulated on several crossings this season. QSO'd W1BB,UO5AA,1PPN-2UWD-UB5W
VO1FB!!!
HAPPY SUMMER-HAPPY HAMMING and 73 de Stew

!!! --- 160 METER FLASH de W1BB --- !!!

THE EP2BK story!!!!

EP2BK works VE1ZZ -- W2IU -- W1BB -- WØVEH/VP9 !!!!!!!!!! FLASH!!! On 160!!

Everyone was thrilled with the appearance of EP2BK/Bob at ABADAN, IRAN - the FIRST time that Country has ever been heard on 160 - QSOing quite a few European

THEN!! during the CQ-160-WW-CW Contest, FEB24, a brief, not too good but definite contact was made with VE1ZZ/John, for the FIRST ~~xxxxxxxxxxxxxxxxxxx~~ ~~xxxxxxxxxxxxxxxx~~ EP/VE -- Congratulations to both Bob and John!!!! During the same morning EP2BK was also heard by W4KFC/Vic, (And others) peaking 588 and thought to be a "Phoney" because of his strength - however, EP2BK also heard W4K at 589 and SWL'd

Following this, on March 8th EP2BK arranged regular transmissions and schedu with W1BB (First at 0300, then 0220Z, to attempt a W contact and QSO - finally making definite, but "Touch and Go" contacts with W1BB on March 22,24 and 26th - but no "Solid" 100% QSOs.

FINALLY!!! on March 27th at 0150Z, the **FIRST ever** Solid 100% QSO, EP/W was made between W2IU/Sam (Ex W9NH) and EP2BK, and followed by W1BB as No.2 at 0220Z their regular schedule time. Bob's signals varied greatly from "nil" to 449 & 53 peaks. W's maximums were W2IU/339-W1BB/449.

IN ADDITION -- Bob/EP2BK also contacted WØVEH/VP9 at 0245 for a not too soli nevertheless definite QSO!!! WØVEH/VP9 peaked at 569 in IRAN!!!

CONGRATULATIONS !!!! Go to Bob/EP2BK for his splendid operating, patience, a perseverance!!!! And also to "Sam"/W2IU for this FIRST!!!

Dick

CONDOLANCES and sincere regrets go to W3GQF/who indirectly made the W2IU QSO possible, when he made this earlier than usual schedule with EP2BK, invited W2IU in on it, and then for some reason or another failed to make contact himself. Also to Charlie/W2EQS, who came in on the opening rather late, and also failed to make it to IRAN. This was all the more disappointing because of the fact that this was EP2BK's last morning on the air in IRAN before closing down and leaving the Country. Also to Shely/W3RGQ who was also heard to make a call this last AM.

BOB accomplished this VFB feat of 160 meter DX using an 813 at 100 watts - into a 15 meter Lazy H antenna 75 Feet high - with an AR88D receiver. He has als worked many Europeans, and we are anticipating a complete report from him soon giving the complete story on his 160 meter activities!!!

THE HC1AGI 160 Meter DXPLOIT!!!!!

Another **FIRST!!!** when HC1AGI and G6BQ QSO'd at 0610z March 11th for the FIRS ever HC/G QSO on 160 Meters. G6BQ was 449 in HC-land. CONGRATULATIONS to both Don/HC1AGE (W3EIS) and Jack/G6BQ VFB indeed!!!

160 sure has been "places" this season!!!!!

EI4A --Ken and Traut Bale expect to be on for a while longer this season. W1BB has sent them crystals for 1827½ and also 1803.0 Kcs Watch for them!!!!

73 de Stew/W1BB

OCTOBER 15 1962   W1BB - 160 METER DX - Bulletin   1962/1963 No. 1

de W1BB/General Remarks ----- Hello Fellows!!!! - W1BB checking in with you on wha
promises to be our BEST 160 meter Dx season in a long time!!! -"Chatter" collected
since spring ended the 1961/1962 season will be found herein. The band was more op
this summer than last, with VP8-ZL3-VK3-EI9-G6-W6-XE-FP-CW making real news!! FIRS
TRANSATLANTIC QSO (W3RGO/G6BQ) was a month earlier than usual this year. Much new
with some suprises is expected!! Many stations W/VE-Dx have used the summer to go
advantage for new antenna and ground system construction. As the Sunspot Number go
lower and lower, 160 Meter conditions get better and better, so we are apparently
for an extremely interesting and exciting season!!! SOOOOOOOooooooooo join in on t
FUN - get those antennas up high and strong for the winter storms - your SWR 1/1 -
check your receivers and transmitters for best performance - mark your calendars
store up a bit of extra rest for those "wee small hours" when 160 is at it's best.
HAPPY HAMMING - best of luck and best of Dx HUNTING on 160!!! 73 de Stew/W1BB.

ANNUAL TRANSATLANTIC AND WORLD-WIDE TOP BAND 160 METER DX TESTS ----- Will be hel
as usual, and this year on Sunday Mornings, December 2 & 16th --January 6th & 20t
February 3rd and 17th, from 0500-0730 GMT. During these tests, special efforts wi
be made to contact European, Asian, African and other 160 meter DXers throughout
Transatlantic World, for general DXing, and particularly to also make observation
and to contribute to the study of propogation conditions on these frequencies. Th
ANNUAL 1.8 Mc Activity since 1932 has been planned as TESTS and not Contests - as
an unusual operating activity, without competition. W/VE stations will call "CQ D
TEST" the first 5 minutes of each hour, and each alternate 5 minute thereafter,
LISTENING the second 5 minute period, and each alternate 5 minute period after th
unless working or hearing DX. Dx will of course, use the opposite periods for
transmitting and receiving. When DX is coming thru, the idea is to work it regard
of time periods.

Most W/VEs East of the Mississippie River will operate 1800-1825kc segment, while those WEST of the River will use 1976-2000kc, clustering near 2000kc. Most DX will usually be found bentween 1795-1850kc, while some will be found near 2000kc (ZL). DX stations should cover 1975-2000kc carefully for unusual and choice W/VE Contac W/VE stations should send their reports for these TESTS to W1BB, for tabulation, Recording in the booklet " THE 160 METER DX HISTORY" in the making, and especiall for transmittal through BULLETINS to ARRL-CQ-SWM-Western Radio-DX-RADIO ZS and ot G,GM,GI,GW etc should report to L.H. Thomas, DX Editor Short Wave Magazine, 186 Winchelsea Rd., Hastings Sussex, England. Other DX may send their reports to eith of the above. Please DO report accomplishments, Happenings, Rumors and Data!!! W1 will award Certificates to all DX stations working W1BB, if requested, and to DX SWLs who send 7 verified reports of W1BB's signals - also IF requested. Certifica for past work on this band are available - contact W1BB for further details. ---- Working DX on 160 is indeed challenging, and extremely interesting. Conditions su as static, BC Harmonics, QRN, QRM, LORAN, NOISE, QSB, all require extreme patienc with perseverance, a top-notch station and A-1 operating ability. There are many FIRSTS yet to be made and many unusual conditions to cope with. One gets anew the thrill of early Amateur Pioneer days, when HAMS were relegated to "TWO HUNDRED METERS DOWN", and Paul Godley made the FURST Transatlantic Crossing via Amateur Radio, thereby inaugurating the first of what has now become Yearly TESTS, enjoye by many 160 meter enthusiasts. In addition you will make many friends, among the finest and most sincerely devoted HAMS you would ever care to meet!!! OK?? b.c.n.

160 mete
CONTESTS ----- W1BB and many other W/VE 160 meter HAMS will give special/attentio to DX stations during the ARRL and CQ DX Contests and also the special CQ 160 Met CW Contest, (Which will come earlier this year - watch CQ). W1BB invites any special schedule arrangements from any of the DX gang desiring to make a real try for an extra band multiplier. AIR/MAIL your intentions to W1BB in this respect, i advance if possible, otherwise W1BB will generally cover all 160 DX Frequenc

W1BB 160 METER DX BULLETIN -----Will be issued during the 160 Meter DX Season, as material is obtained, to the above/mentioned HAM magazines, via air-mail, so that they can publish whatever may be of general interest & the most important informa on TOP BAND activities. There is no mailing list to individuals, although W1BB wi oblige and send copies to any W/VE stations supplying s.a.s.e. - and to any DX station who reports regularly to W1BB on 160 Meter DX Activities - also SWLs. W1B also will send FLASH news items to Magazines on any special events and happenings

FIRST 1962/1963 TRANSATLANTIC QSOs ----- Congratulations to W3GWF/Dick/359 who on Sep. 16th QSOd Jack/G6QB/449 at 0524z one of the earliest openings on record!!! Second crossing was made by W1BB/1/489/0405 from farm in Maine Sept. 29th Third was K3MBF same date. Of interest FIRST SIGNALS were heard "Across the Pond" by G/SWL/Norman Smith Sep 1 0423 W2EQS and W3GWF/459 - Sep. 2 W3FBV/weak CQ answered by G3PU and G3OLI - Sep7 0353 W2IU/459 - Sep8 W3GQF-Sep9 G3PU Calls W3FB Looks like an excellent season shaping up with openings so early!!!!

SUMMER DX - VP8GQ!!! ----- 160 Meter DXers were thrilled with the lowering SunSpot numbers and the pick-up in 160 Dx to hear VP8GQ in the South Orkney Islands comin thru frequently this summer QSOing a number of W/VEs also EI9J "Paddy" to whom go special congratulations. Credit for VP8GQ's appearance and following info goes to Dick/W3GQF who when working him on 80 June 30 asked him QSY to 160 and try - the was "right" and results immediate!!! W3GQF/589 VP8GQ/579 0300z/1827k for a solid 20 minute QSO - contacts followed with W3RFA & W2FYT. W3BKE almost made it but not quite as signals faded. July 1 W3GQF again and W2KQT -July 2 0300z W2IU,KQT,TKG, V GP,D July 3 EI9J/0202,W1BB,W2KQT,VE3QU and W3GQF for his third QSO with VP8GQ W3GQF valiantly but unsuccessfully tried to hook VP8GQ and W6KIP/Alex but he was out of town. Finally July 9th VE1ZZ/VP8GQ QSO. Signal reports averaged S78 for W/ and S48 VP -- these results still further confirm W1BB's often expressed views tha the band is best open to the South summers, after having had QSOs with CP5EQ and other South Americans during prevoius summers. These latest W/VE/EI QSOs confirm!

Further confirmation comes from PY/SWL/Rolf Rasp in R.J. Brazil S.A. who reported
JUN 16 0620 hearing W6KIP/Alex/579 QSO W6ZH/Herb/559-589 1999k. "Their signals we
so clear and loud that I did not miss a word. Never thought 160 signals could be
loud"!!! To which could be added that W3GQF/Dick QSO ZL3RB/Mick MAY6 0955 - VFB!!

VK-EL-Dx!!!!! ----- The VKs with their new liberties on 160 are pushing for 160 Q
with W/Ve & Dx. On Sun AM SEP2 W2FYT was suprised to hear a weak stn clg on his f,
A sort of contact was established before QSB out. The station turned out to be a
call uncertain - believed to be VK3AMZ on 1803k at 0614 - since VK3AGS has been i
identified and it is believed he was the station. News spread and the 160 boys we
"HOT" after this exotic DX!! On Sep9 VE3DU W1EFN-Maybe others- hrd ZL3RB 1015z on
CW & A3 but no QSO. Then the band to VK Cracked open!!! On SEP20 K3MBF W1EFN W2FY
broke the ice for FIRSTS QSOing VK3HG. On FRI AM SEP21 W1BB/1 on farm in Maine an
W3GQF QSOd VKeAKR whose signals peaked 579 staying in from 1000-1100z Steadily!!!
W3SO/459 QSOd VKeAKR/359 SEP28 as did Chas/W2EQS for his 34th Country!!More to co

ENVIABLE RECORD/W3GQF!!! ----- Dick kept the 160 band warm QSOing this summer & F
VP8GQ-W0VEH/VP9-FP8CA-ZL3RB-XE1OK-G6BW-VK3AKR-G3OOT-GW8PG-G3ERN-G3PU and has had
least one W6QSO every month!!! Says maybe cdx will open up DX year/round. Suggest
helpful "Markers" - DHF/1831k- and when WWV is off::: OMA/2500k/Prague/Czech/1000
MMX 1000cps/A2-or FFH/Paris/300w-HBN/Swiss/300w -JJY/Tokyo/2000w-LOL/Buenos Air
2000w-MSF/Rugby/500w-ZLFS/NZ/30w - Thanks Dick - nice work!!!!

THANKS TO CHARLIE/W2EQS ----- who is banging away at his key on 20 and typewriter
"Airmail", stirring up 160 DXers in foreign lands for this years hunting!! Report
PY2BZD-HK1W-HK4DP,1HV-GC2CNS-OA5G-VP3AD-YN1AA-YU1GM,AD-ZB1BJ-OE3SE-ZE8JJ-HI8XAG
all may be on. Some have worked 160 before but others are new ones. Ken/EL4A is
hoping to be on 160/but not for sure/expecting new Jr/op etc. - Charlie/EQS sport
a new 160 doublet which netted him the VK and intends to do his share of Dx Hunti

EARLIER THIS YEAR/VP8GQ/Peter Hobbs (G3LET) ----- opened up one of the rarest 160
spots on the map appearing JAN28 QSOing in order HC1AGI/569/0800-W0IFH/469/0627-
W1BB/1/469/0640-W0NWX/579/0653 - FEB17 W0NCS/569/0625-W3FBV/569/0640. Peter is to
thanked a million times for these FIRST SOUTH ORKNEYS 160 QSOs. G3PAG, His very
meticulous QSL Mgr is to be thanked too for FB QSLing - Peter's address is FALKLA
ISLANDS DEPENDENCIES SURVEYS Base "H" Signed Island, South Orkneys - but QSL via

SAINT PIERRE/FP8BX/W1RAN/"Ned"-----Who made 160 so interesting for many of us dur
contests says "Grand time on TOP-BAND . Golly could I hear Europeans!! They were
S789 and standing out over Ws as early as 0130Z!! VFB NED - Nw we All like try FP

EI9J and 160 in EI Land -----"Paddy" says until recently licenses for 160 were no
generally granted, but now han be had for the asking. At request of W4KFC, last
season Paddy whipped up quick final/ got on with well known FB results. EI9J wks
CW mostly but found 160 refreshing after the commercial and HAM QRM on 80. During
CQ Contest "Paddy" was disturbed by many Gs calling him when he wanted Ws & other
rarer DX & Multipliers!! "EP2BK answered 2 of my CQs/was so loud/ took him f
phony & didnt ans.-ALAS!! Later wkd him however!! Am sure that his signals were t
most remarkable of the test! Now that I have the gear and permit an looking forw
to being up on 160 the 1962/1963 season!!!" We'll all be happy to hear you Paddy!

COINCIDENCE ZCs QSOs w/Ws !!! ----- Stephen Gibbs/GM3MBS/ZC4AK now 5A3CJ/Libya
refers to p4 W1BB DX BULL No.5 1961/1962 pointing out that ZC4XP was Sid Parks,,,
making this comment "Stew-I bet if you examine your log you will find that ZC4AK/
W DQ QSO MAR3'62 and ZC4XK/W1BB QSO MAR3 '53, exactly 8 years apart to the day,
would be at about the same time GMT" ----- And SURE ENOUGH!!! ZC4XK wkd BB 0400z
ZC4AK wkd Willie/W8GDQ 0417z - Pretty close eh????? You boys who want to QSO ZC n
know the right magic date and TIME!!! Thanks for interesting observation Steve!!

W1BB.----- Celebrating his 50th year of HAM Radio - made his 80th 160 Country QSO
working VP8GQ - The next 40 to DXCC/160 will be the hardest - although we MAY mak
it before "Silent-Key" time!!! Great game this 160 - enjoyed every minute of it

Gswl/Allen/APR8 Hrd Wa2KWH,5SCT,ØCDP,2IU,1AW. George was real excited SWLing ØCDP his furthest WEST yet!! Still trying for W6&7 ----- W2IU/Sam has FB 125° antenna To QSO'd EP2BK w/it & should be ready for more FB DX to come!! ----- WØCDP/King/1998k APR8 hrd G3ERN/129-QSOd ZL3RB/Mick 0820 339/both VFB/Congratulations!!!! ----- K1PPN/Paul hd rig trble-missed most Dxtests-got real thrill however hearing FIRST/ JAN7 ----- W2GHK/Stuart(Hammarlund Co) says thru his influence their new HX50 SSB has 160 coils available!! Expects be active 1962/1963 w/inverted"L" - Best of Luck ----- WØPSF/Larry, Hoisington Kansas hopes activate 1962/1963 after stretch/army. M hve real FB ant on 400' tower!!! bonu Larry!! ----- GARY/K1KSH, fm USAF/Turner/AF says 50/50 chance he may become rare 160 DX within 6 Mos-were hoping Gary!! ----- W8HGW/Les,(Parkview Electric Co), who started 160 1961/62-looking fwd to more act this season/listen fr him!! -----Fm G3IGW/"MIKE" WHITAKER,a fine rpt FEB/10 QSOd ZC4PB fr his First/Ever QSO on 160!! FEB/11 QSO VE3BQL/SU -- On CQ/WW/DX/TEST-14 countries-16Mult-9136 Score!!! JAN7&MAR3 each 7 Ws QSOs-FEB/18VE1ZZ&VO1FB - Wishes Ws would give more data on aerial system& Pwr on their QSLs also wonders why there are no W/SWLers giving 160 reports to DX boys -- (W1BB wonders too!!!) There just doesnt seem to be any W/VE 160 meter SWLers!! Also comments "Over the years we hav noticed a N/S skip with E/W "Cut" - even on 160 - this happens on 80 a lot and is more noticeable as greater level of activity shows up - not necessarily due to ATOM bombs - but just normal variation on ionized layers"!! Thanks Mike - fine ide -----G3LIQ/Denis, gives a list of outstanding stations he rec'd viz, Ws9JIN,NSF,VI K8HBR-3GQF-K2DGT,IU-1AW,1BB-W6KIP,APR22!!!!! - HP6AKZ,HC1AGI,HR3HH,EI4A,CN8PZ,EP2B also QSOd/W1BB - VFB Denis and more good hunting 62/63!! ----- G3OJI/Jim,Planning new 160 ZEPP at new QTH/quiet unobstructed. May try vert kite ant!!! Best/Luck Jim -----W8TSD/Dave, was vy active this summer on the 160 Meter "Mills Trophy Race Net on Lake Erie. Also participating were K8YOO Chm-K8SCW,W8QUR,K8YON,W8IME,RZ4,OFG,GJ & K8LFI - VFB boys but how about some 160 DX????? ----- W7UBI/Keith , Expects 2b active 62/63 w/DX100/SX71.Inverted "L" Marconi- doing lots of W/Coast/SSB - bonu!! ----- W8TSD/Dave, again - says at close of RACES everything went fine and dandy - all there are 500 HAMS in Toledo area and 100 on 160 & More coming - Toledo Emgey NET AREC/1812Kc have around 70 sign-in each month's DRILL Sunday 1PM - Excellent w ----- W6HRG/FRANK, reports QSOs between 1.8/2.0 Mc Stns hve occurred each month during the summer!! Frank QSO'd KØPAU/57-89 at times JUL31 A3 - WA6AJY & WA6CDR QS him on CW. W3GQF & WØRFT hv been wkd by a number of W/Coast stns during summer! Fr believes more results would occurr if stns tried summertime DX more!! ---------- SORRY TO REPORT VK4/SWL/Charlie Thorpe, 81 Dawson - Rockhampton-Queensland-Austral had back operation - wl B in cast fr 6 Mos from HIPS to under arms - must stand mo making it hard for SWLing - however will nevertheless cover 160 as well as possibl 62/63 season -Great Spirit Charlie!!! Tnx ------- 8/28/62-K1KSH/Gary says "ALHOA" Hawaii and hopes to get on 160/MARS stn soon - still expecting to "Pop-up" on 160 fm some rare & EXOTIC DX spot one of these days!!! ----- PY/SWL/ROLF RASP,Hrd W6ZF 559/589 & W6KIP/579 JUN/16 and copied entire QSO ----- W1AUQ/"ED"-is showing inter in 160 & may be on!! FB/Ed -----WILLIE/W8GDQ, checks in 62/63 - hrd weak ZL call b Tried VK3AKR who QSO'd W8HGW.Willie has 28 Countries- sure nice to have you Willie one of our old reliables on 160 checking in again for the 160,season!! --------G/SW Norman Smith re orte hearing W1PPN,2EؘS,FYT,IU,KHT,K2PNF,3GQF,8HGW SEP30 ----- G/SWL/ANTHONY M QUEST/LEEDS/ISWL/G5967 also hrd same stations - except 8HGW-Tnx/To ----- WILLIE/W8GDQ, QSO'd VP8GQ SEP23 ------ FIRST VK ever QSO expected to be credited to K3MBF - Nice work DAWSON!!! SEP20/QSO/VK3HG!!!! ------- W1EFN/Walt, sporting special new 160 ANT280° Horiz & 579° wire in vert section 90° off ground and HT9 190w was SECOND VK QSO on record w/VK3HG/SEP30 ------ Comments also that W6s come in frequently around 6.30ᴀᴍ LOcal time - sunrise about - with suprising strengths at times. QSO'd WØVEH/VP9 JUL21 FP8CB/JUL27 - rag chews more than DX but enjoys DX when possible - worked lots of Mobiles on 160 6his summer mostly W8s ----- GW8PG/Gus, reports operation from G8PG/AmGreasby SEP23 hrd W3GQF, W1BB/1- SEP30 as GW8PG "Boy what a party" - Wkd W3GQF and Hrd W2FYT/339-2EؘS/569-2GGL/449- K2PNF/559-2IU/569-2KHT/559 -- Gs QSOing Ws were G3PU,ERN,OQT & 6BؘW ---OCT7 Band in bad shape - NO DX!!! Gs3LIQ,LLE,PGN.NEO,LYW,OQT,RBP Participating.
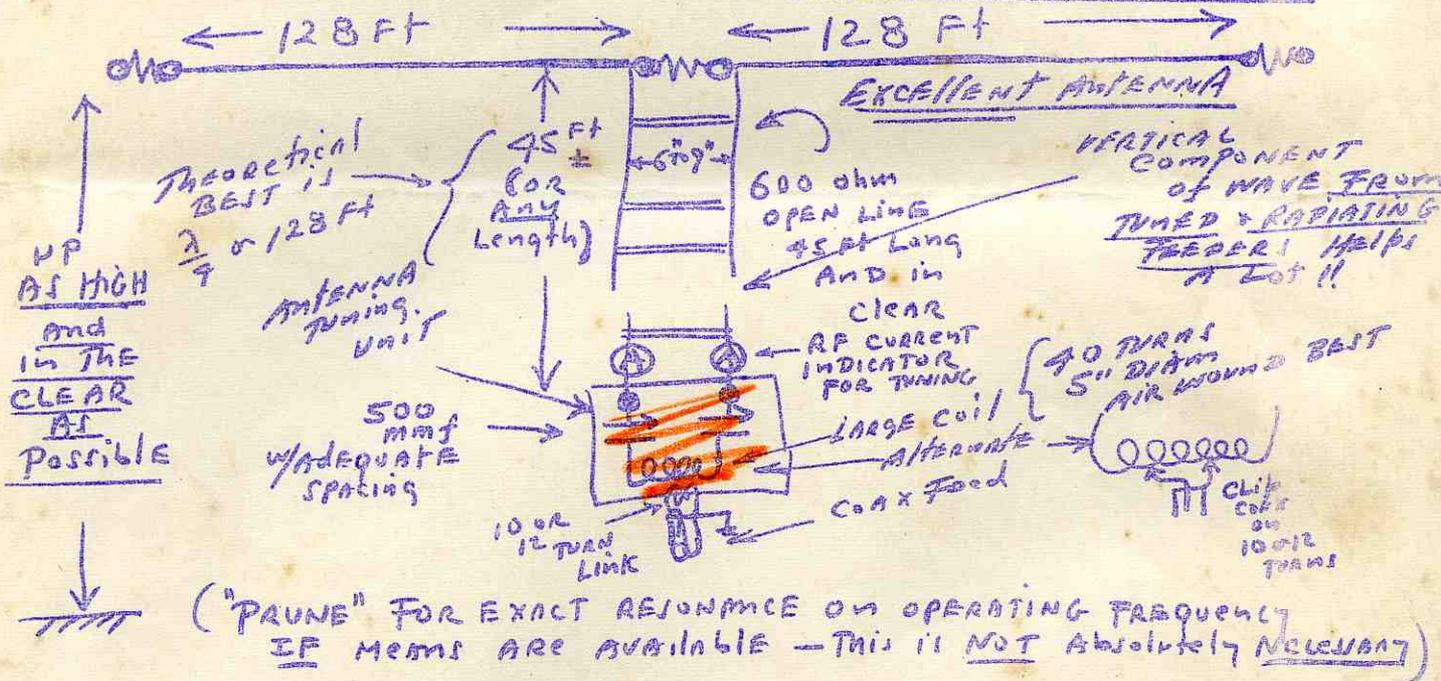
FIRST VK/W QSO CREDIT !!!!! Does go to K3MBF - this has been confirmed by W1EFN The station was VK3HG SEP20/62 o515EST - Congratulations DAWSON!!!!!!
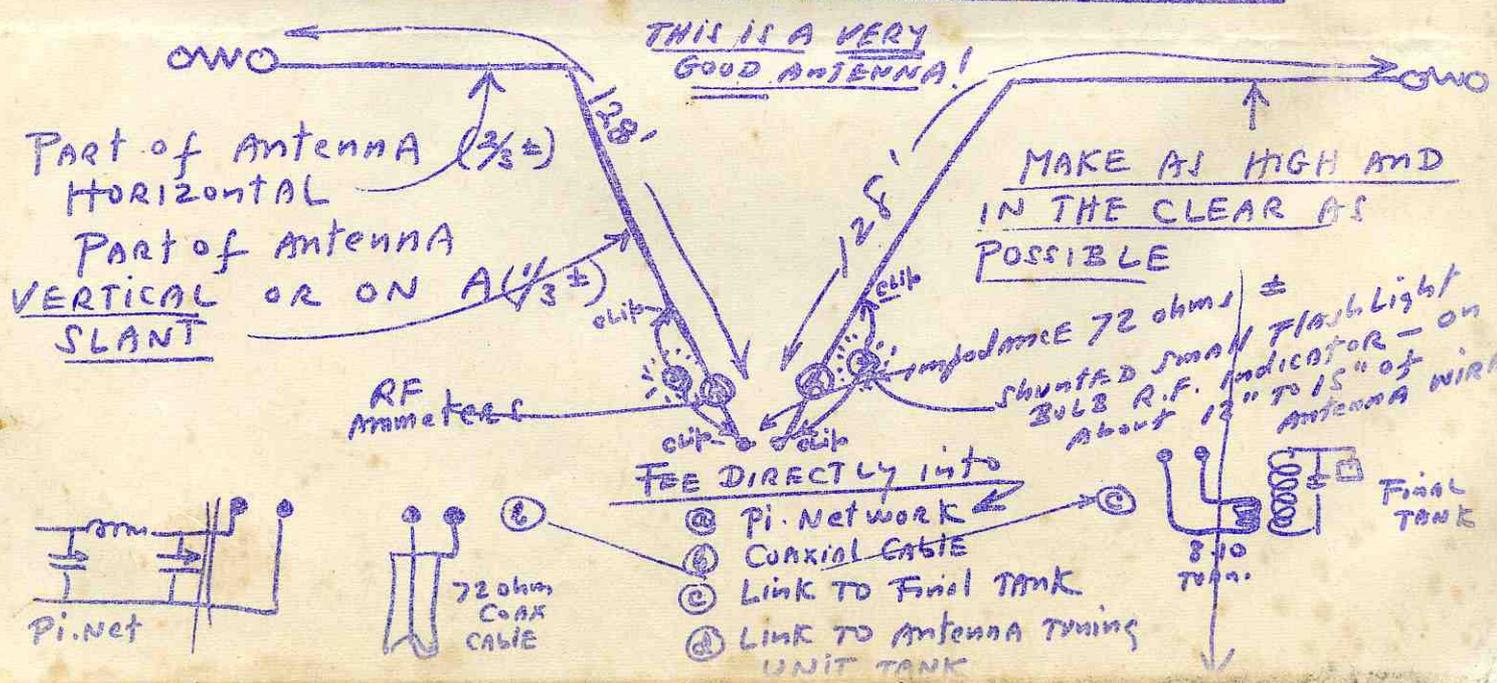HAPPY HAMMING fellows - and best of DX on 160 -- 73 de/Stew

W1BB

ANTENNAS FOR ONE SIXTY

LISTED IN ORDER OF DESIREABILITY
FOR BEST DX WORK

NO. 2 — HALF WAVE DOUBLET — w/ OPEN WIRE FEED

THIS BEST ANTENNA FOUND BY ACTUAL TESTS

← 128 Ft → ← 128 Ft →

EXCELLENT ANTENNA

UP AS HIGH And in THE CLEAR AS Possible

Theoretical BEST is $\frac{\lambda}{4}$ or 128 Ft

45 FT ± (or Any Length)

Antenna Tuning Unit

6'9"

600 ohm OPEN LINE 45 Ft Long And in CLEAR

VERTICAL COMPONENT of WAVE FROM TUNED & RADIATING FEEDERS Helps A Lot !!

RF CURRENT INDICATOR for TUNING

LARGE COIL

ALTERNATE COAX FEED

40 TURNS 5" DIAM AIR WOUND BEST

CLIP COAX ON 100Ω TURNS

500 mmf w/adequate spacing

10 OR 12 TURN LINK

("PRUNE" FOR EXACT RESONANCE ON OPERATING FREQUENCY IF MEANS ARE AVAILABLE — This is NOT Absolutely Necessary)

NO. 3 — HALF WAVE DOUBLET — BENT and Brought DIRECTLY INTO

SHACK FOR FEED — 2nd BEST ANTENNA

THIS IS A VERY GOOD ANTENNA!

Part of Antenna ($\frac{3}{3}$ ±) HORIZONTAL

Part of Antenna VERTICAL or ON A ($\frac{1}{3}$ ±) SLANT

128'

128'

clip

clip

MAKE AS HIGH AND IN THE CLEAR AS POSSIBLE

RF Ammeters

impedance 72 ohms ±

clip

shunted small Flashlight BULB R.F. Indicator — on about 12" TO 15" of antenna wire

Final TANK

8-10 TURN

Pi·Net

72 ohm COAX CABLE

FEE DIRECTLY into
ⓐ Pi·Network
ⓑ Coaxial Cable
ⓒ Link TO Final TANK
ⓓ Link TO Antenna Tuning UNIT TANK

NO. 4   ¼ WAVE — SINGLE WIRE ANTENNA
WITH "GOOD!" GROUND   3RD BEST

(THIS IS EQUIVALENT TO Half of ANTENNA #2)

!GOOD!
Simple and
EFFECTIVE !!
Antenna

¼λ 125'

½ ± vertical TO ½

½ TO ⅔ ± Horizontal

HIGH
AND
AS IN THE
CLEAR AS
Possible

FEED

ⓐ Pi Network Direct
ⓑ Coaxial Cable Direct
ⓒ Couple to Final Tank w/Link
ⓓ Couple Directly to "
ⓔ Link to Final Tank or Ant Tuning Unit TMK

R.F. Ammeter or Bulb R.F. indicator   Flashlight Bulb

8"   8"
clip   clip

GOOD GROUND Connection !!!
IMPORTANT —
SHORT TOO!

IMPORTANT TO HAVE BEST GROUND Possible

16 TO 10 TURN Link

500 mmf Condenser may be Needed

TAP ON 8 TO 12 TURNS From Bottom Adjust to Load

ABOUT TURNS TO Load
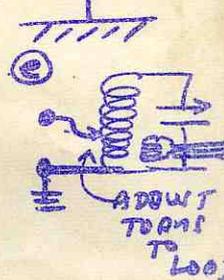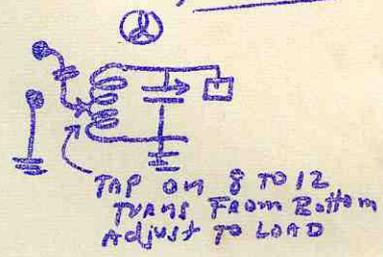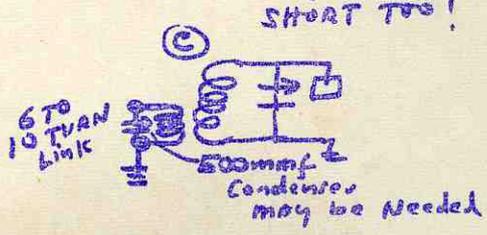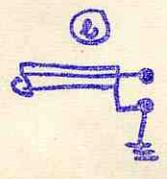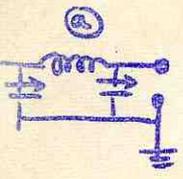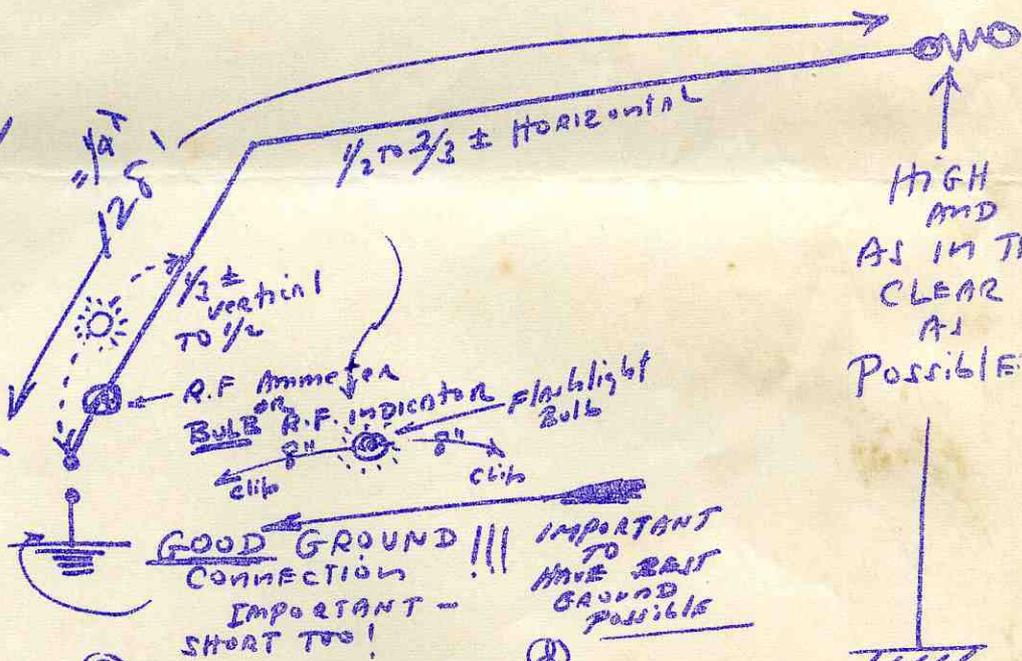
NO. 1   ½ WAVE ZEPP ANTENNA — 4TH BEST

SEE NO. 1
EXACTLY SAME REMARKS APPLY
AND SAME Component VALUES

← 256 Ft →

45' or any Length
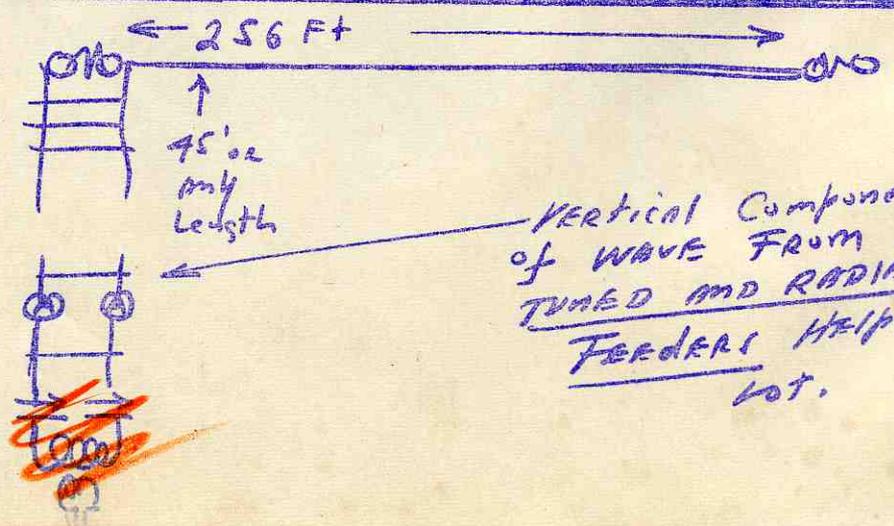
Vertical Component of WAVE FROM TUNED AND RADIATING FEEDERS Helps A Lot.

NO. 5 — HALF WAVE DOUBLET — COAXIAL FED — 5"

← 128' →   ← 128' →

If Local Man/Made QRN is HIGH — This is a good Antenna To Help Cut Receiving Local QRN

75 ohm Coaxial Line — any length

As HIGH and in THE CLEAR As Possible
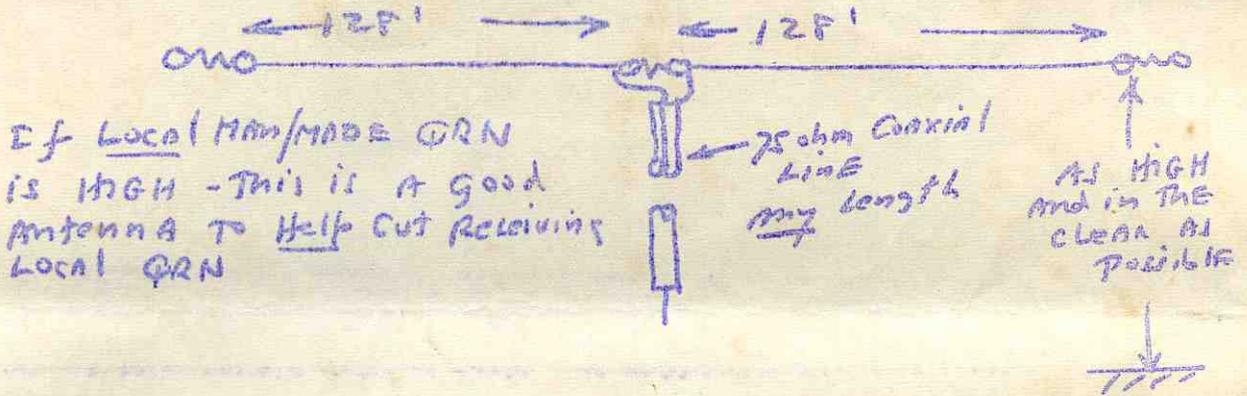
NO. 6 — 160λ Folded Dipole w/ 300 ohm Ribbon

OR MUCH BETTER !! w/ OPEN 300 ohm Line

← 256' →

128'   128'

any length FEED →

300 ohm Folded Dipole

Ribbon or "open" Line

As HIGH and in THE CLEAR As Possible

TX
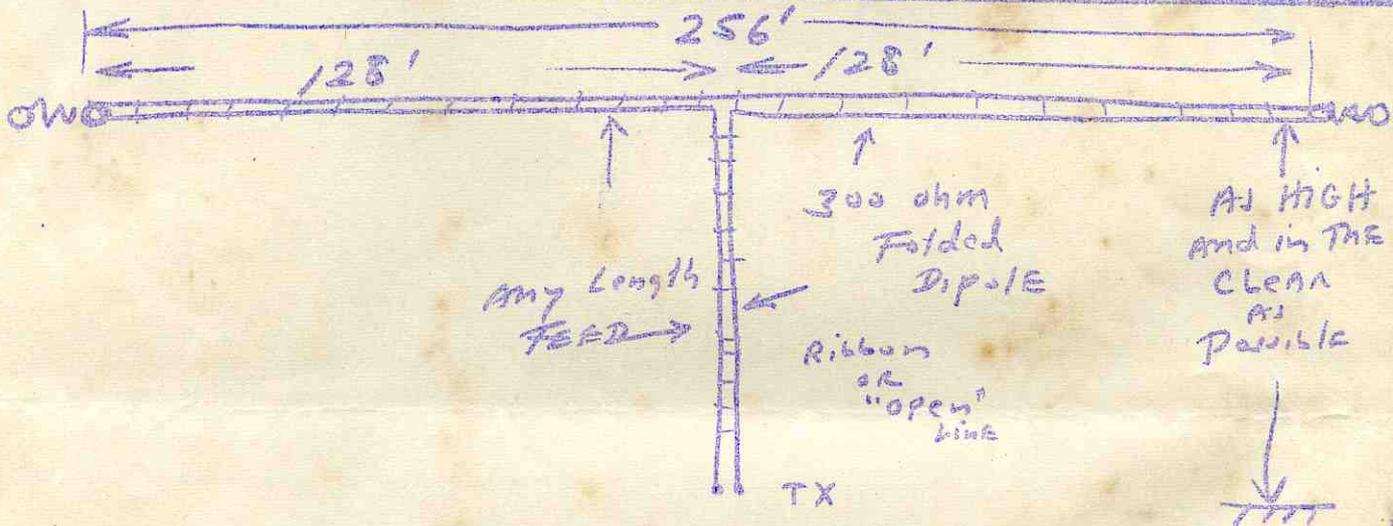
NO. 7 — VERTICAL 160λ Antenna
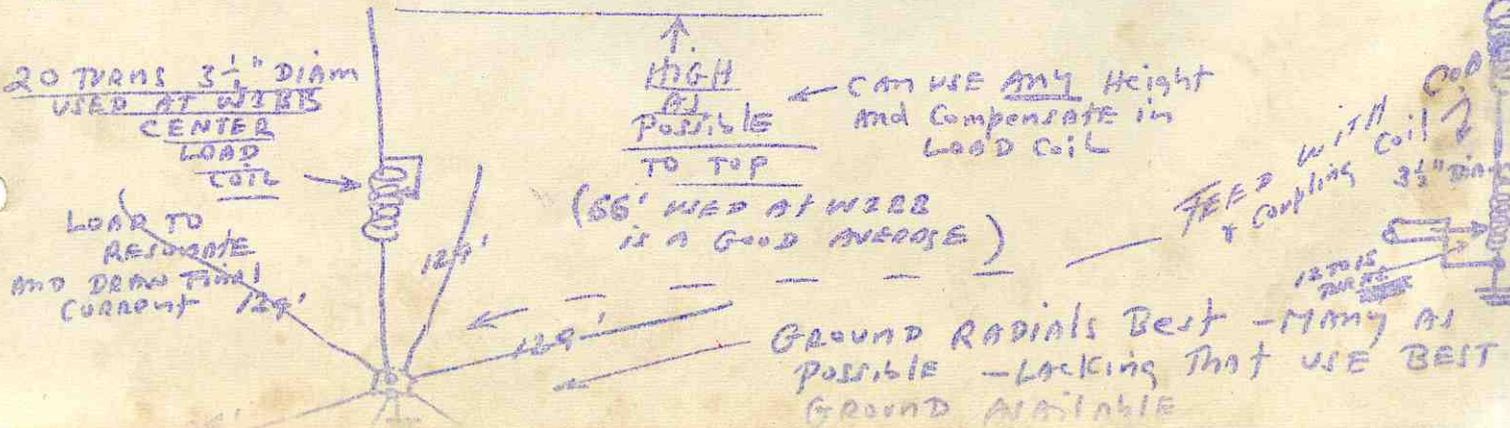
20 Turns 3½" DIAM USED AT W1BB CENTER LOAD COIL

HIGH As POSSIBLE TO TOP

← CAN USE ANY Height and Compensate in LOAD Coil

FEED WITH Coupling coil 3½" DIAM

LOAD TO RESONATE AND DRAW FINAL Current

128'

128'

(66' USED AT W1BB is a Good AVERAGE)

GROUND RADIALS Best — MANY As POSSIBLE — LACKING That USE BEST GROUND Available

W1BB
STEWART S. PERRY
36 Pleasant Street
Winthrop, Mass.

PYswl/Rolf

Here is all the latest on 160 O.M.!!!

Nothing on 160 this morning - see log - sorry

Starting New Year will make Wed and Fri at 0600

And as soon as OK with you

Make it Tue and Thursday at 0600 instead Wed/Fri - OK????

CONGRATULATIONS yourfine work there on 160

K9GZK also reported to me your hearing him - VFB!!!

Am waiting to hear from you're crystals - tolerances

SOOOOOoooooo again Rolf - HAPPY NEW YEAR

And here's to a contact REAL SOON NOW!!!!

73 de Stew/W1BB

73! de W1BB/Stew

DECEMBER 31 1962   **W1BB - 160 METER DX - Bulletin**   **1962/1963 No. 2**

de W1BB/General Remarks ----- Fellows, if things have been a bit late, or some of
your reports and letters not acknowledged or answered as promptly as usual - Pleas
Excuse - Alice/W1DQF, and myself were called home from 12 days vacation in KP4/KV4
lands, when very recently, Dec7th, we had the second death in our immediate family
with the passing of our Son's wife, 30, leaving two little children, which we are
now helping to care for - all of this upset us and our normal activities in many
ways - however, we expect to be back "on the beam" again real soon - thanks for yo
patience and understanding!!!   DX this fall was good up to October 1st, with
the earliest Transatlantic crossing on record. Then, rather suddenly from Oct1 to
Nov25 Dx was Nil - since there has been Dx but spotty - no real good openings
only a couple "fair" ones - Why conditions which started out so promising, have be
so much poorer than expected, no one knows - some feel the Pacific Nuclear and USS
tests might have been responsible - others that it is just the "Way the ball bound
, normal unpredictableness of 1.8 - which incidentally makes it interesting and al
intriguing. The First scheduled TRANSATLANTIC DX TEST Dec2 washed out - the 2nd De
16th only fair although participants were not lacking on either side!!! - At any
the frequency of openings lately has improved a wee bit, and lets hope for the bes
     Dont forget the CQ/WW Contest on 160 Jan26/27 and best of luck!!! This should
a grand event!!!   SOOOOOOOooooo Best wishes to all for a HAPPY NEW YEAR, with
l   of HAPPY HAMMING - especially if it is on 160!!!!!

W1BB/Summary LOG of 160 Meter DX Activities!!!!
     Oct21 --- Poor morning to Europe-fair to South - medium w/ve activity - Wx-
VFB/warm/clear/calm -felt like gang just "warming-up" for the real DX ahead!! Alt
VP8GQ put on a fine performance!!!   W/Ve p/W1BB,ME,2EQS,UWD,IU,3QQF,CBM,FBV,4OM
8GDQ,K8IQQ,MRB,9FNE,ØCDP,RHI,GIH,VE2UQ,3QU ---(p/ etc stands for Participants!!!)

EAST::: all-nob a signal or marker hrd a per cus/word because of poor cndx seems
degenerated into "Rag-Chewing". p/G3IGW,OUK/A,CIT,MYI,REP,RAA/A,OZF,OQT,AGG-6BQ,2.
WNF,PG/A. Tnx Gus!   SOUTH:::: VP8GQ 1800.5 our "Southern (Dx) Star" 0500-0630z
weak QSB zero/459, av 339 QSO'd Ws,1ME,TX-2EQS,UWD-0CDP-9PNE(?),1BB-K8MBR in order
was hrd by Py/Swl/Rolf Rasp!! Says "Hrd VP8GQ QSOing one W after another, BUT the
whole time through I did not hear one TRACE of the Ws - only VP8GQ- dont you think
this propogation was rather strange??" Yes Sure Do Rolf!!! For Roy/1TX his QSO
with VP8GQ was his 10th Country!! (Oct.7 VK3AKR his 9th!) (p/ means "Participants

OCT28::::: Extremely poor Dx Cdx. WWV sending "W3" Wx,clear, cold, windy,FB.
W/VE:::: Ws,1BB,ME,TX-2UWD,KHT-3GQF-8HCR-9PNE,CUY-VE3DU participated, calling and
listening for Dx but nary a peep and by 0600 most had QRT!!

NOV4:::: Again Cdx Vy poor-no Dx at all except W3GQF reported hrg DL1FF 0300
559max - HI8XAG & 5B4FB expected but no show - There seems to be some barrier, that
signals of any strength do not penetrate, since Oct1 - Nuclear Tests??   W/Ve:::
p/Ws,1BB,BHQ,ME,WY-K1LTM-2EQS,IU,KHT,UWD-Wa2IUQ-3GQF-4CXW-Wa6CDR-8CIM-K8RYU-9UKV,P
8NWX,VXO-VE3FF   EAST:::: No QSOs but G8PG/a hrd W1BB 229-339 o530z. DJ1BZ, GM3PBA

NOV11:::: Condx still generally Vy poor especially to East. W/Ve Participants
dropping a bit also.   W/Ve:::: p/Ws,1BB,UZ,TX-K11MO-2FYT,KHW,JGQ-3GQF,FBV-K7DEK/
8IQS,GDQ-K8WGW,BBI,IQQ,HSW-9PNE-K9JDK-0IFH,GBV-VE3DU,QU.   EAST:::: Cdx vy poor w
no European markers or Dx coming thru!! However GW8PG reported hrg W2FYT briefly a
0438z clg CQ DX -- Gswl/QUEST hrd a N. African but no other Dx. p/G3IGW,HR3HH,GM3P
EI9J,DL1FF,VP8GQ   SOUTH:::: VP8GQ came thru weakly 0440z QSOing W2FYT. Our old
friend HR3HH/"Hal" put on best show of evening 0600-0700z QSOing Ws1BB-8HGW-K8IQQ-
0IFH,VXO and others.   WEST:::: K7DEK/7 provided some welcome Western DX and new
states for a few QSOing VE3QU,DU-W8GDQ,HGW-K8IQQ and others - Was hrd by W1BB S2/3
Wa6CDR active QSOing W0GBV/others!!

NOV18:::: Cdx still generally poor-QRN rather high-No Dx whole evening exce
VP7NY "Local-Dx"   W/VE:::: p/Ws 1BB-2DHQ-3FBV-K3MBF-4IQU-8TSD-k8RYU-9PNE-VE2UQ,3M
2FYT,UWD,GGL,EQS-1BHQ.   EAST:::: GW8PG/Gus says morning a "Dead-Loss"-Blizzard cu
poor lines-became good cook over fireplace but no Dx!!! SOUTH:::: VP7NY/Lowell
put on a fine show wkg Ws 1BHQ-2FYT,UWD,GGL,EQS-3PAP,GQF,PMV-8QWI-K8RYU,MRB-9PNE-
VE3QU,AGX. Cdx so poor his sigs varied 000-569 quick QSB 2 hours much of time
unreadable in QRN/QSB WEST:::: A fine clean out FIRST!!!!! when the OM/XYL team
WB6AFH/WB6AFI (Ex W9VER/K5VLG) QSO'd VR30 Christmas Island on 1991k Nov18. VR30/44
6s/559.-----(Under call K5VLH QSO'd ZI3RB&KH6IJ Fm N. Mexico early 1962 and as W8F0
QSO'd G-VE-VP7-KP4-KV4-YN4-YV5 in past years) CONGRATULATIONS!!!!!! Jean and Dean
Understand VR30 will be on regularly when he gets new Tx now being built by W6VBY.
Note!! W6HRG also worked VR30-Details not rec'd hr.

Nov25::::: AT LAST the band showed some signs of opening up and quite a numb
of Dx QSOs resulted - our best morning in some time!! Willie/W8GDQ says dont forge
to cover 1876k fr ZLs and also claims "MOONBOUNCE" QSO w/W1BU on 160!!! Hi!!!
W/Ve:::: Sam/W1BU says QSO w/W8GDQ was 1st 160 QSO since coming to Mass fm Ohio
'40s - using pp814s/rig fm 20yrs garage storage-has tilted ant 200° high NE end-n
to have Sam on again!!! Maybe sometime Sam and Willie will explain this "Moonboun
stuff on 160!!!!! p/Ws1BB,ME-2FYT,EQS,UKS,IU,GGL-3GQF-6KIP-Wa6CDR-8FW,HGW,GDQ-0VX0
VE7AN.   EAST:::: G3IGW hrd W2FYT,GGL,IU and briefly contacted W1BB. UO5AA on 180
G8PG/Gus Qsb/Qrm heavy sigs weak but wkd W1BB 1st time & Hrd W2IU -G3IIQ hrd W2FYT
G/Swl/QUEST QUEST h/1BB449,2GGL549,FYT339 - nw hrd 19 Countries 160-logged 15 Dls
this yr ---Gswl/Smith h/1BB,ME,2IU,FYT,GGL after some blank weeks!! ---- G3OQT/D
"The Dx Blanket which covered everything OCT & NOV seems to hv lifted nw-hrd W2FYT
449,2FYT/449 & 1BB/549-Qrm Hr Fantastic!!! Making himself an 80° vertical Beer Can
ant wl top load it"also wonders why there are no W Swls- says Gs wld like rpts!!!
Gswl/Iball h/1BB,ME "Have never known such activity/QRM over here!! p/G5OQT,PU,IG
PXZ,LIQ/8MYI,REP,JFY,OZF,NEO-6BQ-DL1FF,DJ1JZ,UB5CI,UB5WF,UO5AA,OH2NB,OK1KAU,KSO,A
Z,YD,KVE,2KJU,5B4PB,OY7ML   SOUTH:::: W8GDQ/Willie wkd YV5AGD as did W1BB,W3GQF
and others. VP2VL/W0NWX Bob was on QSOing 1BB,3GQF,8GDQ and many others - believe
Bob intends retire VP2land someday and work DX to his hearts content alongside
next door neighbor KV4AA/Dick who sure sets a good example - Dick an ex/160 Dxer
is washed our on 160 by "Regs" while on VP2 Island few miles away VP2 can wk 160!
W2UKS/Bill rpts wkg VP2GX/1810k 2334z-VFB!!   WEST:::: W6KIP/Alex had 17 eastern
"Without trying" hrd HC1DC-VP2VL/S3 -Couldnt find YV5AGD" - Sri Alex!! Next time.

Dec2:::: W1BB was in KP4 land/unable observe conditions but understand were generally poor!! W2UKS/Bill rptd band vy NOISY & "Dx sure is scarce to date"!!! has HQ145x new/March '62 for sale $175.00!!! VFB Cdx.   EAST:::: GW8PG/(G6PG)Gus 1MF229-has driven 300 miles this winter to try wk 160 fm GW8PG --- Gswl/QUEST says "N. Dx Hrd" p/G3IGW,OII,OQT,MYI,GRL,OIB,LIQ/a,PQA-6BQ-GW8PG-GM3ITN,PBA

Dec9:::: W1BB returning fm KP4-KV4 lands unable ck conditions - however fm rpts apparently not too bad!!! NOTE- the night before DEC8 W3GQF worked an O.T. an famous early 160 meter DXer - and noted DXer all bands G2PL Peter - and hrd G6BQ, 3QT,IGW-G6GM and DL1FF!!!   W/Ve:::: W9PNE hrd DL1FF239 15 mins-G3PU339 Briefly - EAST:::: Gswl/Allen W1ME,TX-2EQS,UWD,FYT,GGL,KHT,IU-VFB!! W1TX wkd G6BQ-G3PU --- G8PG/a/Gus hrd 2FYT449-2IU229,EQS559,GGL339,UWD339 sigs weak/QSB Predawn Peak0612-postdawn peak0747 ---"Harry"/G6GM 78 Yrs young putting out VFB Sig as usual!!! --- Gswl/QUEST; Ws1ME,TX-2EQS,FYT,GGL,KHT,UWD says PAØPN,SS-OH2YV,NB on!! ---- Gswl/ Smith [illegible] Ws1ME-2IU,FYT,GGL says Ws 3 deep between 1806-8 & Fish- fones real bad!! ---Gswl/Iball-1TX359-2FYT569-GGL559-EQS449-IU459-3GQF339 using HC1155 Rx.   SOUTH:::: VP8GQ trying QSO Gs DLs/N.D. --- VP3AD/Dad hrd W2FYT, 569/0410-UWD449/0425-UKS459.0436-8IRE559/0432 - Called all of them on 1801/N.D. Dx/Participants G2DCG-3QT,PQA,PU,LIQ/a,PHO,OIT,CHN,REP,IMV,PIQ,ERN,GRL,NHE,PG KFT,IGW,NJR-5JU-6GM,BQ,QB-8PG/a,JR-DL1FF,FL-PAØPN,SS-OH2NB-GW3PPF,VP3AD.

Dec16:::: This Second Annual TRANSATLANTIC Dx TEST was much better - but still spotty and with lots of snow static & Loran - Nevertheless quite a few QSOs ensued between 0500-0800 with a fwir number of participants and several outstandi QSOs. As a "Warm-up" W1TX on the 14th QSO'd G2PL,3ERN,6BQ,6TK and 15th G6BQ,3ERN, PQA   W/VE:::: Roy/1TX wkd G3PU -has new VFB antenna on only 50' Lot - results are what count!! W9PNE/Brice hrd DL1FF 30 min peak 239 also G3PU 0700z 339 rapid QSB. NOTE!!! Dec 15 Brice wkd VP8GQ solid got 570 gave 459 for his 35th Country!!!! (Coincidence - VP8GQ was also W2EQS/Chas's Nr 35 Country!!) --- CONGRATULATIONS'' to WØVXO/Herb for the 2nd Transatlantic QSO EVER from WØ when he QSO'd G3PU559 re 3 VFB!! WØNWX has only other WØ-European 160 QSO on record. He also QSO'd G3PU 11 yrs ago!!- Wish could print all of WØVXO's letter/Space limited-however he hrd DL1FF569/0730z w/Horiz Doublet & NOT AT ALL w/Vert. Could have kicked myself for not putting up doublet sooner he sez. NOTE WØVXO & WØNWX both also hrd Nov11 by PY/Swl Rolf Rasp 0545z - FB!! p/1BU-2FBA,IU,FYT,GGL,TMG-K2BWR-3GQF-4VYB,KTA-Wa4FJ 8IRE,HRV,OOR,QWI,QIY-K8UNP-9PNE,CUY,HIY-K9GZK-ØVXO,IFH-VE1ZZ,2IL,UQ-3DU,AGX-1BB EAST:::: G3PU's signal was terrific and outstanding, staying in from 0500-0800, peaking 579-averaging 559 dipping 229 with slow QSB and wkg quite few Ws includin WØVXO & W9PNE!!!!(CONGRATULATIONS to Ern also on his first QSO with VP8GQ this AM DL1FF was next-QSOing quite a few but not up to usual strength. G6BQ,3QT,IGW,8JR also made QSOs. (The previous evening Dec15 G2PL569 wkd W3GQF and band was quite open half hr with fair signals) ---- Gswl/QUEST hrd 1BB,ME w/30' piece wire at g level!! and nite before 1TX,ME-2FYT,KQT VFB!! ----Gswl/Allen hrd 1BB,ME,TX,VP8GQ using series tuned "LOOPSTICK" 2/short 12" screened lead!!! Some goings-eh??? say it helps cut QRM/N -sumpin to try!!! ---- Gswl/Iball hrd 1BB only Cdx bad Winds above storm force   SOUTH:::: VP8GQ QSO'd G3PU/0507z CONGRATULATIONS Peter - W listened to both sides of QSO!! Night before - 15th G6BQ also QSO'd VP8GQ for po FIRST - (On 15th QSO'd W9PNE and Brice's son W9CDP at 0400z) VP8GQ that is. ----- VP3AD/"Dad"hrd 2FYT569-K2BWR559-W9HLY459-W2FYT589-W1BB559-W1TX559-W8HRV459-W2GGL4 VE3AGX449 between 0300-0510z -----KP4AXU on 3.5Mc works 1.8 Mc Stns Cross-band -- PYswl/Rolf Rasp sends impressive list hrd 0500-0600z, VP8GQ589-W1BB589-2GGL589- 8IRE449-2FYT459-1MF459-2IU579-VE3AGX449-1TX459-K9GZK349-84WI259 Condx outstanding QRM/high-QSB/strong. Also 14th Dec hrd WØNWX569/0700,W6KIP589/0713 and 15th VP8GQ 338/579 QSO G6BQ-0515z   WEST:::: No stations further West than W9/WØhrd and none reporting to W1BB   NORTH:::: VO1BD heard clg DL1FF/0435z-believe no QSO. Dx-G3ORH,OQT,ERN,NVO,GGS,ROZ/a,RKA,RIQ,OFA,PWK,LIQ/a,OIN,KDV,MCY,PSB,PQA-6BQ,TK 8JR-ZL2BE-OH3NY-OK1MGmAEZmKSH,#CEA-GM3AEYgAWF-SVØWZ-GW3CBY NOTE !!!! During this week the band openedcup Dec 21st and G3PQA put in a terrific signal QSOing W3GQI W1BB,ME-2IU-UWD-8HRV. G6BQ QSO'd W1BB,TX - and W1BB QSO'd VP8GQ ((9PNE also QSO'd PQA,

Dec23:::: WWV sending "W-3", which is extremely poor Cdx - and so they were- not a trace of DX and all participants QRT early. Strange when mornin before G3PQA put on such good show. Understand terrific storm in G/land

An extremely interesting letter from George Jacobs/W3ASK comments on our FLASH
bulletin Nov.11 ---"I hve mentioned the possibility of A & H bomb explosions causi
absorption in the ionosphere, with some friends of mine in this field. They are of
the opinion that the enegry contained in the many bombs exploded recently by both
the .S. and the USSR is still negligible to the amount of energy that would be
necessary to induce a long period absorption effect. With Bomb-testing apparently
halted, how have conditions on 160 been?" - (George, since testing stopped conditi
seem to be coming back to normal for this time of year, but extremely slowly and v
very spotty - whether there is any correlation or not is hard to tell - but some
still feel that there is, others that there might be!!) Conditions hard to explain

FINLAND ON 160!!! ----- Here is another Country available - word from several sour
OH3NY,GW8PG,etc - HO HAMS allowed 1820-1845k & 1915-1955k - 10 Watts. Our old frie
Matti Paiavo/OH3NY, also OH2NB are on - Matti says "Cdx much poorer than 11 yrs ag
sigs must go over Magnetic N Pole - will be active even though new QTH hr much poo
for ant - 73 to all!!"

PROPAGATION FORECASTING ----- Fm W0TUT/Joe a most interesting comment --"Your
coincidence of QSOs p3 Bull#1 62/63 reminds me paragraph June 62 RSGB as follows -
"IEE procedures 3/62 explains new method ionospheric forecasting being used by th
D.S.I.R. based on identification of "Epochs" during which corresponding Cdx have
prevailed past yrs, rather than plotting completely new forecasting maps as previo
It has been found usually possible identify some period within the previous 10 yrs
when almost identical radio condx occurred. The result of this new system, is a
great increase in accuracy of predictions made several months in advance" - Tnx Jo
Sorry abt ur S9 line noise washing you out on 160 Dx - Joe wants info on D.F. to
find noise - can anyone help?? If so send Joe the references - Joe Check QST/ARRL.

ULTIMATE NAVIGATIONAL AID!!!!!" ----- Fm Boston Globe 11/8/62 -"A new Navigational
aid developed by John A Pierce, Harvard Scientist,(who also developed LORAN)-Named
OMEGA - 8 times better than LORAN - Range 8000 Miles (Loran only effective 1000 Mi
Three OMEGAs would replace 100 Lorans and give greater coverage -$10,000,000. cost
per installation however. Expected operational not less 3 yrs nor more than 5 yrs-
been in development 12 yrs. Funded by three Services through office of Navigationa
Research - apparently operates on higher Fy than LORAN (Dont know fy yet) -S00000o
maybe here's relief in sight from our LORAN QRM on 160 - We HOPE!! Anyone more dop
                                                                    on this????
HEARD ON 160 (Without mentioning any names) BY W1BB ----- "A fellow
can stretch his ears -when digging for Dx on 160 9 or any other band - but especia
160 -to the point where he THINKS he hears sumpin!! Did YOU hear ever so weak a si
on 1827?? I thought I did!!! - Gosh, I doubt if my ears will ever go back into sha
" - Now isnt it the truth???!!!! W1BB at least knows JUST what you mean, O.M.!!!

MARKER DHJ54/2 ----- Per G3PQA "Has packed in and is replaced by DHJ59/2 which is
much weaker station - so if you dont hear DHJ - dont despair of Dx"- Tnx John !!

CYPRUS - Fm G2BVN/Steve comes word that 5B4PB is Famagusta, Cyprus & Now active -
usually fm 0300-0450z 1801k alternate odd numbered days. Static high/but will TRY!

160 in FLORIDA ----- de W4IYT/Andy Florida Skip Ed. - "You said there is nothing
keeping us fm at least listening in Fla to 160!! Oh YES there is Stew - LORAN!!!!!
I'd like invite you Fla just ONE Night - ANY night and "Listen" to 160!! You proba
couldnt stand the 40 to 60 DB over S9 racket fm our HiPwr LORAN Stns - Man it's JU
AWFUL!!!! - lets hope however some day they'll turn off those QRM makers an albow
us back on"!!! Tnx/Andy-real sympathies and hopes for a better future!!!

"FLORIDA SKIP" Mag/W4JO "Work For Joy Only", Frank. Advt Mgr sends extremely FB
comments --"160 been dead band Fla many yrs-missed band much- W4EB/W4JO organized
"Miami 160 phone gang" in 1930-Frank used 201a Mod W/247-still has it -(In Attic)-
i 1935 Mag "HARMONICS" born devoted 160/Phone/Hamming-Net met 8AM Sundays - so yo
see 160 is vy dear to hearts of O.T.s arnd Miami - we used Long Wires/Low Pwr - Dx
was anything out of the state' tho I once QSO'd Europe with tt 201A - wl be list
Transatlantics - If poss thru Loran -was THERE at the very FIRST & 2nd TRANSATLANT
w/cans only " Frank O.O.T./Tnx Vy interesting Comments - Hope 160 back to Fla Soon

AMERICAN SHORT WAVE LISTENERS CLUB -----Jim Howard, Ed, says "Thinking of offering awards to ASWLC members who report overseas 160 meter Dx stations - might make it Club Contest" - this would help Dx stns a lot!! Tnx much Jim - Hope you do it!!!

SOUTH AFRICAN RADIO LEAGUE/Michael C Bosch/ZS2FM,East London ----- Writes " Our branch has embarked on a project to revive interest in "Top-Band" in view of low M expected during Min Sunspot Cycle. Following stns are on - ZS2CE,CD,FF,FM,GW,JH,KA KW,LC,CF,OL - some VFB/Most Xtal 1950k - ZS2FM has rock for 1987 - pwr 10w,1987kfr OK!!!! Julius Reitmayer/Club Station OK1KCI ----- says, trying hard for W contacts Requests contacts & Contestants especially for their "TP""TEST" 2nd & 4Th Mondays 160 - fm 1900-2100z - Exchange rpts RST plus QSO Nr ie 579013 - Call CQTP. One QSO rpts Multiplier is No. QSOs during 1st half hour 1900-1930z. Send TEST Logs within 3 days to Central Radio Club PO Box 69 Praha 1 Czechslovakia - OK??? (Time NG W/Ve

MISCELLANEOUS TOP-BAND NEWS ITEMS -----G/swl/Norman Smith was season's earliest,la latest, most persistent SWLer!-His rpts kept up all summer! Oct6 band sounded STIN no activityNorm kept at it finally digging out W8HGW & VP8GQ fm mud 30th VFB!!! -- GW8PG/Gus,Oct14 h/G6BQ,3IGW,OQT,OIT,PQS,NQF,GSS,active/NoDx-Gus vy happy/Psed w/ne friends/Activities 160-says "More Antenna dope & SWL rpts fm W/Ve,Pse"!!! DL1FF/Ar Per Gswl/Quest)who can be counted on to give 160 real working over, has VEE beam,5 high towards WEST being hrd terrifically loud in Europe-broke thru to W/VE 449 No DL9KP,1W,60S also active. -----EI9J/PADDY (11/6) Hopes be on soon - says "Regardi your saying that Trans-Equatorial Cdx best summer - I could agree - Over last 12 y Jun&Jul hv bn two best months in yr - QRN rarely high hr-we get no extremes - I am particularly thinking of 80/same goes for 160 Tnx Paddy/benu ----- OY7ML/MARTIN, g spcl Lic operate 160 on 1825k during TRANSATLANTIC DX TESTS. First night on wkd 20 stns inc. many Gs, EI9J,DL1FF,GW3JI - Hrd a N African,could not raise-using 10w - Congratulations Martin!!!! ----- USA/SWL/Adrian Fallert/Hamilton Ohio taking part Dx logging per GW8PG/Gus who was hrd by him Sep30,0600z,449. This is VFB to have U SWL!! -----Gswl/QUEST rpts DJ2KS,5KR,1FF,KB,7IL,1W,60S active 160-also NOV7th OY7ML "Exploded" on 160 wkg string Gs,DL1FF fr FIRST ever DL/OY 160 QSO!!! OH2NB w DL1FF Nov13 & G3MYI on CW&SSB for a 1st!!! 5B4FB emerges fm QRN now and then-TnxTo VP3AD/DAD wl be on regular TRANSATLANTIC TESTS only whenever poss. 1801k - can mak no special skeds at other times/time limited - had rig trouble/now fixed - Good nr G3OQT/Dick says "I want to cross pond SSB - Anyone interested please contact me fo series skeds weekday mornings-(So not QRM boys hr weekends/make myself unpopular)!! OK SSB gang write R.F.Mclachlan 27 Lake Rise,Romford Essex Eng - Dick nw has new ant bent 1/2 wave -has crossed pond 4 times since Sep30!!!'VFB!! -----PYswl/RolfRa doing FB listening 160 Hrd W1BB several times on "CQ DX TEST" calls & other ?????? weak signals - band bad PY land however NOV. ----- G2HKU/Ted has new QTH "Now actu above sea level" wl b on soon. -----G3CHN/Roger"Patchy season so far-G3FPQ wkd VP8 OH2YV,NB & PA0XN,SS active" Tnx Roger -----EI4A/Ken"It doesnt look like I'll ever on fm Liberia agn/consfiscated all gear Sep27-wl be lvg spring-hope go some rare Dx Country next to make up for lost time" Congrats to Ken&Traut on VFB yl Jr Op!! ExGD5MIK/Rex 160 Dxer nw VE2BKK is at Perry Iowa "Long live 160-wish I could be mo active" -----GM2BUD/Dave wl b on starting Jan 9 - His SON graduated "Associate"-Ro College Science & Technology & Wkg DeHaviland Aircraft - VFB Dave/Congrats-get him interested 160!! -----ZS2FM&ZS6XQ&"RADIO ZS"Mag plugging 160 in S.A. - bound to I more ZS 160 activity soon!! -----G3PQA/John"Wl be glad make special skeds with any anytime" -----G2DCG/Ernest plugging 160 hard-wkd UA3NB,DJ2KS,OY7ML,GD5UB,OK1AA1, DJ3KR,1KB,2KS -----PA0XN anxious for Ws-Nuf Sed!! -----JIM/GM5IAA(ExVS1AA) wl hve new mast soon & active 160 agn -----G3OIT/Keith wkd OH2NB,DL1FF& Others -----W1WY/ FRANK regrets time doesnt permit much time on 160-has 75A4-intends some activity-offers give good CQ Publicity to "HOT" 160 News - Tnx Frank -----W9PNE/Brice "Neve hrd cdx so poor-dec3-WWV/"W-3/4" for nearly 2 Mos nw-only few times up to N5-You I no idea hw difficult it is to wk 160 Dxfm out hr - Noise level low however - testi various ants/interesting & Confusing results -Wkd 80 exclusively during CQ/WW DX t after 6th Straight Certificate-Hope!! Congrats Brice!! -----K4KSY/Rev DraytonCooper Moved to Southport N.C.-on soon as can solve ant problem/little space-wl use new SP-600-JX-14 Rx - benu Dray on 160!! -----W8ANO/Waldo ex avid 160 Dxer listens 160 on old BC/Rx but still no Tx-too bad!! -----W6KIP/Alex moans "Out here I dont know 160 could be worse!!" All I can hear is big juicy power leaks arnd clock" Sri/Alex
73 de Stew/W1BB and b.c.n.u. on 160 fellows HAPPY NEW YEAR!!

!!! FLASH !!!

MIDDLE NOVEMBER
!!!-- 160 METER DX report de W1BB --!!!

Dx in all directions since October 1st have been extremely poor, although some spotty DX has broken through to W/VE, namely, VP8GQ (Weak) - HR3HH good signal - and DL1FF weak and briefly for 20 minutes (No QSOs) using his Vee beam and calling CQ DX --- European "MARKERS" have been conspicuously absent except for the briefest moments.

This raises a question as to what has happened to 160 meter Dx. Up until October, it was our best year - with the earliest Transatlantic crossing on record and lots of Southern Summer DX including VP8, ZL, VK etc and apparently continuing to improve while the SUNSPOT Numbers increased, as expected.

However - rather suddenly from the first of October on, it has been almost as though a DX barrier had been thrown up, which even the strongest DX signals can not penetrate.

Many are wondering if the Pacific NUCLEAR tests could be responsible for upsetting the Ionosphere and casting a "blanket" over 160??????

All 160 Meter DXers are active however, and holding on hoping for the "Break" to come soon!!!!!

Most active W/VE PARTICIPANTS are≡ Ws1BB,TX,ME,UZ,WY-K1IMO-2EQS,FYT,IU,UWF KHT,-3GQF,FBV,-4CXV-5KG,NU,SOT-6KIP-Wa6CDR-K7DEK/7,K7HDB-8GDQ,IQS,CIJ-K8IQQ-9PF UKV-K9JDK,-0NWX,IFH,GBV,VXO-VE1ZZ,3QU,DU,FF   DX PARTICIPANTS are≡ G6BQ,G3ERI
                                                              -8HGW
IIQ,PU,CHN,OIT,PQA,IGW-██████ PU- G5RI-G6CM-G8PG - GM3IAA - GM3IOS - GD3UB - HR3HH - VP8GQ - EI9J - UB5WF and others.

Stewart /S Perry/W1BB

(During Week of December 2 W1BB
will visit KV4AA "Dick" in the
Virgin Islands --To discuss possibility
of 160 Meter DX operations from "YASME and future DXpeditions...

1962/1963

DECEMBER 23 ~~7~~ ~~#~~ 16 and ~~17~~ ~~8~~ ~~#~~
JANUARY 6 ~~7~~ ~~8~~ = 20 and ~~21~~ ~~22~~
FEBRUARY 3 ~~4~~ ~~5~~ = 17 and ~~18~~ ~~4~~

During these periods, Extra Special efforts will be made to contact European, Asian, African and other 160 Meter DX Amateurs throughout the world, for general DXing, and also yo make observations, as a contribution to the study of PROPOGATION conditions on these frequencies.

An active group of English and DX Amateurs, in Cooperation with USA Amateurs are behind this effort in carrying on these Tests, which have been a yearly operating activity since 1932. All 160 meter stations throughout the world are invited and also urged to participate in as many of these TESTS as possible!!!

Most W/VEs will operate 1800-1825 Kc Segment EAST of the Mississippie River, while those to the WEST will use 1975-2000 Kc, Clustering near 2000 Kcs Usually. DX will be found mostly between 1795 and 1850 Kcs, the best segment being 1800-1825. Some DX will be found near 2000Kcs. DX stations should cover near 2000Kcs carefully for unusual and choice USA Contacts!!!

W/VE stations will call CQ DX TEST the first 5 minutes of each hour, and each alternate 5 minute period thereafter, listening the second 5 minute period, and each alternate 5 minute period of the hour, for DX. The DX stations of course will do just the opposite. Thus, each station alternately calls for 5 Minutes, and then listens for 5 minutes throughout the TEST — unless working or hearing DX. Care should be taken particularly by all participants to carefully set their clocks to WWV, etc. to be sure of keeping in step with these 5 minute periods, and not cause QRM by transmitting during the listening periods. These calling and listening periods of course apply only when no DX is being heard or worked. When DX is coming through the idea naturally, is to QSO regardless of the periods — BUT, here again, adherance to the 5 minute periods, even for a QSO, would help to minimize QRM.

W/VE stations should send their reports for the TESTS, of activity and unusual accomplishments to W1BB, Stewart S. Perry, 36 Pleasant Street, Winthrop, Mass., USA for recording in the booklet " THE 160 METER DX HISTORY", and for transmittal through BULLETINS to A.R.R.L.—CQ—SHORT WAVE—WESTERN RADIO and other magazines, etc. for publication. G,GM,GI,GW, Etc. should send their reports to G6QB, L.H. Thomas, DX Editor, Short Wave Magazine, Turkey Road, Bexhill or Sea, Sussex, England. --- OTHER DX, may send their reports to either of the above as desired. Do REPORT!!!

Remember!!! --- These are "Tests", and not CONtests -- for pleasure in an unusual operating activity, without "Competition", for a change, and to develope some propogation information as a worthwhile contribution to the art of Radio.

A.R.R.L. and CQ - DX CONTESTS

W1BB --- And many other W/VE, 160 Meter HAMS will carefully cover these contests on the same operating schedule as above for the Transatlantics. W1BB invites any special schedule arrangements from any of the DX gang desiring to make a real try for an extra band multiplier!! AIR/MAIL your intentions to W1BB in this respect, in advance if possible, otherwise W1BB will cover the same frequencies as mentioned in paragraph 3 on page 2 of this Special No. 1 Bulletin. CQ are proposing to run another 160 Meter CW Contest early in 1961, and it is hoped that DX stations will participate in this operating activity also. See CQ for further details.

W1BB 160 METER DX BULLETIN

W1BB --- Will issue this BULLETIN during the 160 Meter DX Season, as material is obtained, to the above/mentioned HAM magazines etc, via air-mail, so that they can publish from it, the important information on "TOP-BAND" activities. There is no mailing list to individials, although W1BB will oblige and send copies to those supplying S.A.S.E.s
Best of luck ALL - and again, 73 de W1BB/Stew.
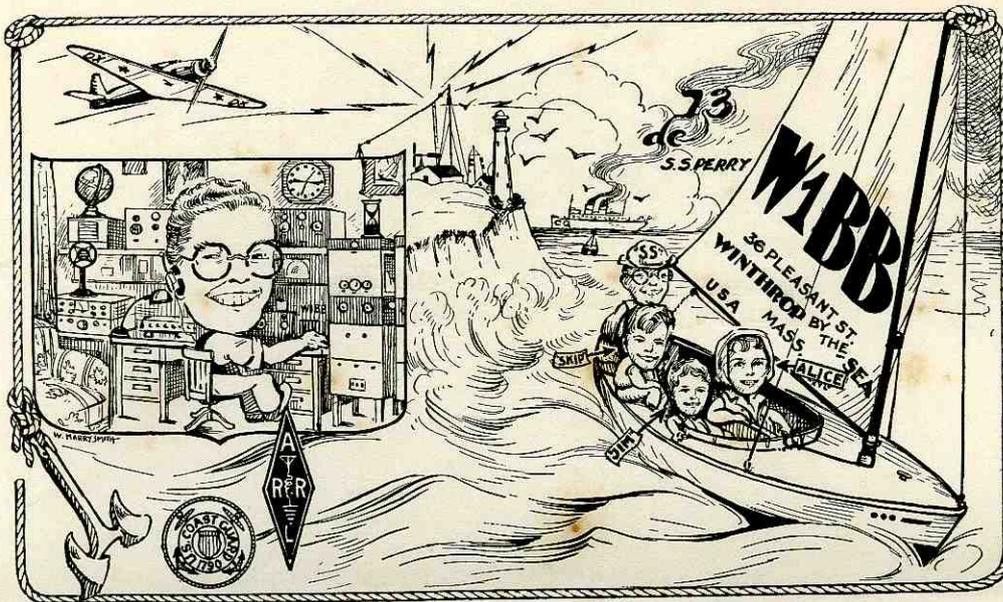
Dear O.M.:-

At the start of another Season, W1BB sends GREETINGS with 73!!, and the hope t
you will take an important part in the 160 meter DX activities!!!

Working DX on 160 is challenging, and extremely interesting. Conditions such a
static, BC harmonics, QRM, Loran, QSB, noise, all require extreme patience, with
perseverance, a top-notch station, and A-1 operating ability. There are SPECIAL,
TRANSATLANTIC and WORLDWIDE 160 Meter DX "Tests", and the ARRL and CQ, DX Contest
to participate in. (See below in this bulletin). There are many "FIRSTS" yet to be
made. Many unusual conditions to observe and cope with. One gets anew, the thrill
of the early Amateur Pioneer days when HAMS were first relegated to "200 meters d
down", and Paul Godley made the FIRST Atlantic crossing via Amateur Radio. In
addition, you will many friends among the finest and most sincerely devoted HAMS
you would ever care to meet -- these "Top-Band" boys !!!

Your participation here will also help to promote Amateur occupancy of the ban
and assist in keeping this 160 meter operating priviledge. In addition, you will
help to develope PROPOGATION information of real value to the art.

Through these BULLETINS issued during the season by W1BB, reports on your
activities and accomplishments, and all information of 160 meter interest, will be
sent to QST-CQ-SHORT WAVE-DX-WESTERN RADIO-VCM-DXing HORIZONS-FLORIDA SKIP-Radio
Propogation Editor/CQ, magazines, etc., for publication. Be sure to keep W1BB well
informed of all your 160 Meter exploits and results, etc.

In case you personally are not particularly interested in 160 Meter operation,
"Test" schedules, DX Contacts, or unable to work this band - PLEASE pass this
BULLETIN along to someone who is. Also, W1BB would appreciate if you would send th
calls of all HAMS in your Country who might be interested in 160 Meter DX, and ab
to work this band. Please post this BULLETIN in your Club, and give 160 meters as
much publicity as possible!! Particularly as to DX!!! THANK-YOU, very much indeed

W1BB 160 METER DX ACTIVITY

W1BB --- Will be on regularly EVERY Sunday Morning, (Baring unexpected sicknes
business demands or Emergencies), from 0500-0730 GMT, calling and listening for
alternate 5 Minute periods, November through March, looking for 160 Meter DX
Contacts!!! Please be sure to give W1BB a call!!

W1BB --- Will transmit, Mostly on 1802½ or 1807½ Kc. However, 1817½ and 1822½
will also be used at times depending on conditions, QRM, Interferance, etc.

W1BB --- Will NOT work V/VEs during DX schedule hours - unless something of an urgent nature comes up, or band conditions are useless for DX - will Concentrate/... on ...

W1BB --- Will generally follow the regular 160 meter "Test" procedure. When no DX is being heard or worked,, W1BB will call, CQ DX TEST, the first 5 minute period of every hour, and each alternate 5 minute period throughout the hour. W1BB will listen carefully for DX stations, the second 5 minute period of each hour, and each alternate 5 minute period throughout the hour.

W1BB --- Suggests that DX stations call W1BB on 1797½ - 1802½ or 1827½ Kcs, as these are the most QRM/free spots, usually. These SPOT frequencies will be covered with EXTREME care by W1BB at ALL times. The band 1800-1830 will be covered carefully too, and frequencies up to 2000Kcs checked regularly.

W1BB --- Welcomes, and invites special schedules with any DX station that desires to make contact - at any time of year, and on any special, (Authorized), frequency designated by the DX station. Send AIR/MAIL letter to W1BB, with full details on the proposed schedules, and also any schedules on HF bands that may be necessary to make full arrangements or exchange of reports etc.

W1BB --- Appreciates REPORTS on his signals, from any station, at any time of year, with full particulars - whether you make QSO schedules or not. This includes SWLs, some of whom have submitted MOST accurate and valuable data. These reports assist in propogation studies being made. Please AIR/MAIL them weekly. Postage gladly reimbursed if desired. Special reporting FORMS supplied on request. All reports acknowledged. Please DO report!!! The more the better!!!

W1BB --- Will operate mostly from the HOME 'TH, "Winthrop by the Sea", BOSTON, USA, with 200 watts, a 260' doublet antenna and 75A4 receiver. Some operation will be from W1BB/1, either at the WINTHROP YACHT CLUB, with quiet location and special antenna entirely over salt water!!, or from "Villa Mon Repos", at Harrington, Maine, 300 Miles Northeast from Boston, a quiet country location, with 260' long "Vee" antenna - 260' on a leg.

W1BB --- Will mail a handsome appropriately lettered "AWARD" Certificate, to all DX stations working W1BB on this band, if they request it. In addition, for DX stations where their W1BB/QSO is a "FIRST" recorded "Top-Band" QSO between that Country, and USA, an extremely SPECIAL and attractive lettering job, in color will be done on the Certificate by a Commercial Artist, to Adequately and Especially commemorate the event. This Certificate is in effect a "KING-SIZE" QSL, measuring 10" x 14", beautifully colored, on heavy durable white paper, suitable for wall mounting or framing. DX SWLs who send in 7 or more Verified reports on W1BB's signals, or have in the past, may also have a Certificate upon request. DX stations who have worked W1BB in past years, and any previous "FIRSTS", may have one of the Certificates, back dated to cover the event, if they request it. Sixteen of these special "FIRST", Colored certificates have been awarded thus far, and in each case the recipient has written appreciatively advising that it was framed and/or posted conspicuously in the shack!!

W1BB --- Hopes that you will join in on the fun, and the challenging work of QSOing DX on 160 - and enjoy yourself and help others to make that much desired contact!! Write W1BB if you need further particulars. Keep W1BB fully informed of all 160 meter activities possible as suggested above, for the BULLETINS and reports on your doings to HAM magazines for publication, as their Editors decide.

W1BB --- Says, 73!!! HAPPY HAMMING and b.c.n.u. on 160!!!!!! -- de Stew/W1BB

ANNUAL TRANSATLANTIC and WORLDWIDE "TOP-BAND" 160 Meter DX TEST !!!!

These special "TESTS" will be held this Season on the following SUNDAY Mornings from 0500 to 0730 GMT, in accordance with the listings given below:--

Balun and Transformer Core Selection

[[Home](#)] [[Up](#)]

Related articles at

[Balun Test](#) contains model of "perfect" dipole currents.

 [Sleeve Balun](#) shows how a sleeve adds impedance, useful for VHF and higher baluns

Receiving [Common Mode Noise](#) shows how lack of a balun can contribute to system noise (it applies to transmitting antennas as well)

[Longwires, Verticals, and Baluns](#) shows how unbalanced antennas can have similar problems

[Transmitting baluns](#) on testing transmitting baluns

Occasionally errors are made regarding core selection. This especially includes baluns, where on occasion some very strange ideas surface. One rather odd but somewhat popular idea is that adding a mixture of core types will allow both high power operation and high choking impedance in baluns by slowly reducing current through a balun. Other misleading claims are that extreme values of core ui, such as values in the 10,000 or higher range, are necessary on 1.8 MHz and higher. Other ideas tend support use of excessively low permeability cores for the same application.

When I recommend a core, the material selection is always based on actual measurements with proper test equipment on a bench as well as in the actual end-application.

Core Material

I mainly use 73 material for receiving applications in *LOW POWER* applications between .1 and 30 MHz. 73 and similar core materials generally minimize the turns count required without inducing excessive loss. One of the best indicators of correct core selection is looking at the turns required. You'll notice most of the transformers I use have only one or two turns for every 100 ohms of impedance. My 75 to 450 ohm Beverage transformers, for example, only require two-turn primary and 5-turn secondary windings. *A low "turns count" is a good indicator the correct core size and*

core material is being used.

For high power applications at HF it is often necessary to use lower permeability cores. There are two reasons for this:

- Lower permeability cores generally are available with higher curie temperatures. They operate at high temperatures without losing their magnetic properties.
- Lower permeability cores have higher Q (lower loss tangent) at a given frequency. This means a larger part of the impedance is associated with *lossless* reactance rather *dissipative* resistance. They turn a smaller percentage of power into heat, and that is very important at high power levels.

Permeability changes with frequency. As frequency is increased from zero eventually core impedance peaks. Above the frequency where impedance peaks the impedance of the core (and the *effective* permeability) actually decreases.

A downward slope in permeability with increasing frequency is useful for controlling impedance in broadband transformers, but we should be careful to avoid excessive slope. Excessive initial permeability can easily move the operating area too far out on the downward slope of impedance.

A transformer or inductor operating on the downward slope of a high ui core requires extra turns to maintain critical impedance and often requires more turns than a lower ui core. The upper frequency limit will decrease, and this may reduce useful bandwidth in the desired frequency range.

Using excessive initial permeability means winding becomes more tedious (it takes more turns). The wire has to be smaller and more fragile to fit a given core window. Temperature stability is often reduced while losses increase over an optimum core material selection. In addition, stray capacitance increases needlessly, reducing bandwidth and increasing unwanted stray coupling.

Do **NOT** pick cores solely by considering initial mu, since that value is taken at dc. You should consider *characteristics measured at the operating frequency!*

Always remember this general guideline. Les wire length (as long as winding impedance is sufficient) results in better transformer bandwidth. The best designs place maximum conductor length INSIDE the magnetic core window, and minimum conductor length OUTSIDE the core window.

Heating

At higher power levels, it is necessary to move to lower loss tangent and higher curie

temperature materials like 65, 61, or (in extreme cases) 43 materials. Even a fraction of a dB loss produces significant heating in small cores when power level is in the kilowatt range. *The loss DIFFERENCE in non-resonant applications between lower and high ui ferrite cores isn't significantly different, but **heating** can be much less!*

We often assume heat means a core is very lossy or is "saturating", but **t**his often isn't true. We must consider the power level, duty cycle, and ability of the core to dissipate heat and look at the full picture.

Very small cores, such as small thin .5 inch diameter cores used on bead-type choke baluns, can only dissipate a fraction of a watt in open air. It sometimes helps to put temperature in perspective by visualizing how hot a 60-watt light bulb runs in normal operation. When we consider the core's size, it usually has significantly less surface area than the bulb. The core also has poor thermal conductivity, and is often stuffed in a container preventing any type of air circulation.

Consider the construction of a typical bead balun, enclosed in PVC and heat shrink tubing. As little as 20 watts dissipated out of 1500 watts can produce damaging heat in tiny beads enclosed in a PVC tube. 20 watts out of 1500 is less than 0.1dB loss, yet it overheats the core!

The problem is almost always a heating problem, and not a core-loss problem. It is almost never core saturation, unless the core is subjected to very low average power and very high peak power levels. It is best that we worry about heat and the number of turns we use, not actual power loss, when selecting a core.

Core Style

Soft-iron cores (soft magnetically) increase inductance because they increase *flux density* near a conductor for a given current. With only a small amount of flux "concentration", there can not be a large increase in inductance or impedance. We need a significant increase in flux to have a significant increase in impedance.

The area outside the core window does NOT have a closed magnetic path surrounding the conductor. ***The presence of the core has a minimal effect on impedance of any conductor area outside the core window.*** Most of the flux from external wires is in air, rather the core. With only a portion of the flux surrounding the outer conductors cutting the outer layers of the core, the ***useful*** impedance contribution of wire outside the core window to system impedance is minimal. Conductor length outside the core window mostly adds unwanted stray reactance and leakage flux. If we MINIMIZE the wire length exposed outside the core, and we generally have a more effective inductor, choke, or transformer.

This effect can be easily conformed in a simple experiment using an antenna analyzer. Connect a short wire across the output of an analyzer, and measure the impedance. Lay a core against the wire,

and observe the very small impedance increase. Now pass the same wire through the core center, and observe the large impedance change. This illustrates why the winding's wire length on the outside of the core is wasted, mostly contributing to undesired effects.

Core Dimensions

The area inside the winding-window of a soft-iron (soft magnetically, not physically) core is cut by all of the flux lines, and this area has a very large effect on impedance. The core concentrates the magnetic flux surrounding a current-carrying conductor into a very small area, and the thickness of the core moving away from the area of the conductor very rapidly has less effect.

- The additional impedance caused by placing a core over a conductor or conductors is almost entirely proportional to the core's internal length (window depth) paralleling and surrounding the conductor or conductors.
- The core diameter or radial thickness only has a small effect on impedance.

Doubling the core area **parallel** with a conductor roughly doubles winding impedance. The same is NOT true for an increase in core wall thickness, core thickness barely affects impedance.

I prefer binocular cores for most low-power applications and side-by-side stacks of cores (making a large "binocular core") for high power broad-band applications. This type of core arrangement almost always minimizes the amount of conductor hanging "outside the window". With very little conductor hanging "outside" the core window, there is less "needless" wire adding undesired stray capacitance and series resistance. For a given core material and impedance, conductor length can often be reduced to about one-third of a similar impedance choke (or transformer) using a conventional single-hole core or single stack!

Low Power Measurements

Phase Inversion and Choke Baluns

Some of my receiving system designs use phase-inversion transformers. Phase-inversion transformers are identical to (and interchangeable with) choke baluns or line-isolation transformers. For HF receiving applications, 73-material binocular cores are wound with six passes of #26 twisted-pair enameled wire. I use Fair Rite Products 2873000202 cores (about 1/2 inch square and 1/3 inch thick 73 material).
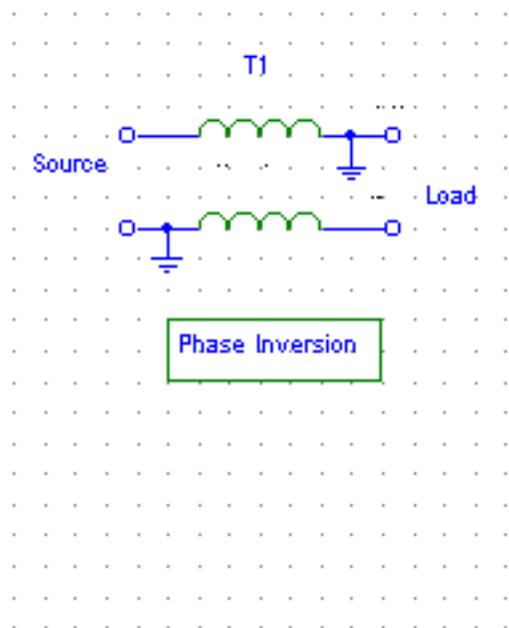
Here are measurements of a sample inversion transformer at 2.5MHz using accurate (and fairly new) commercial equipment:

Load Value (ohms)	Loss (-dB)	Phase Error (degrees deviation from 180)
100	~ 0	<1
33	~ 0	-1.2
10	0.2	-2

This shows inverting transformer construction is good, since even a 10-ohm impedance load works well!

Beverage Matching Transformer
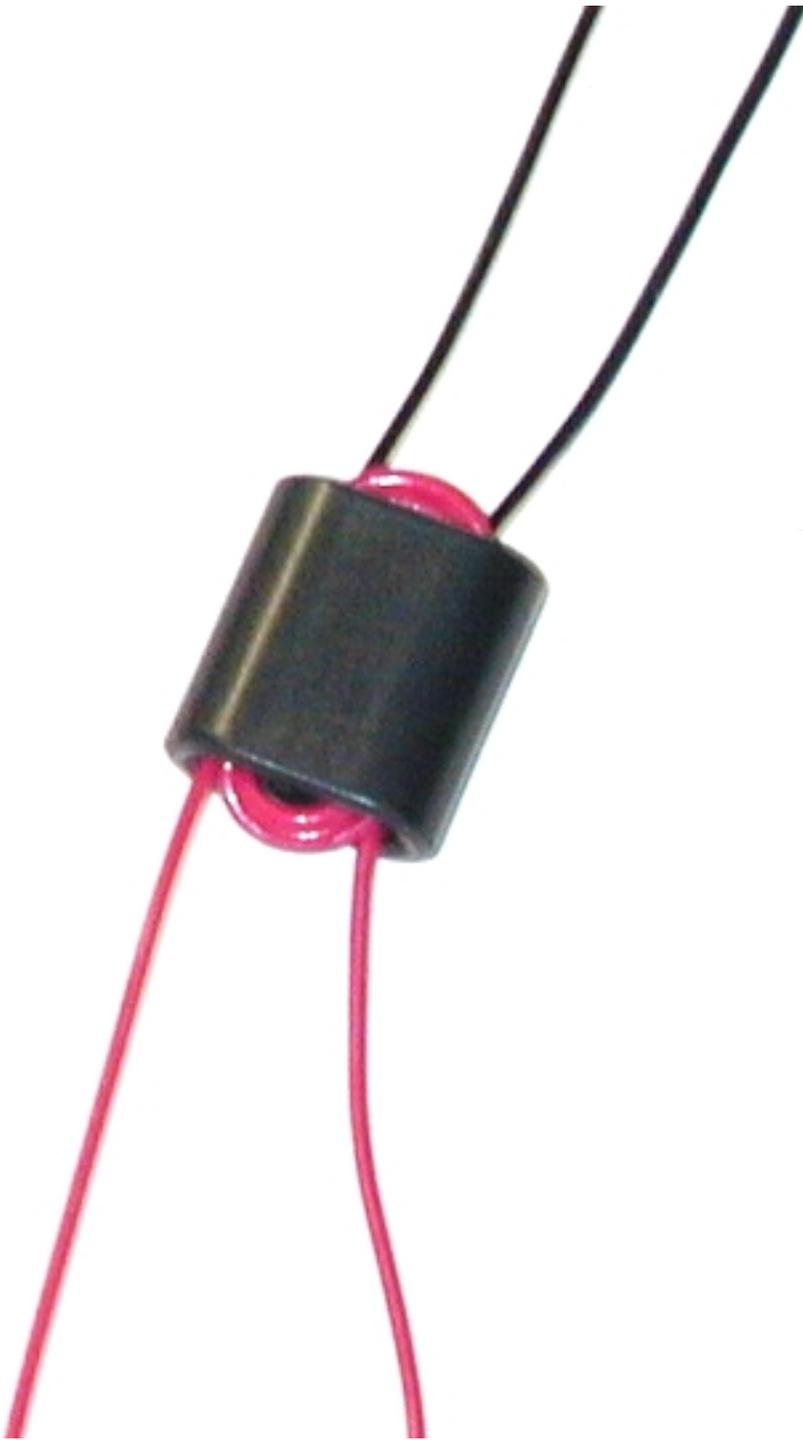
Sometimes I use designs for a long time, and forget how I decided they were OK. I recently received an e-mail questioning the number of turns in my Beverage transformer design, so it seemed like a good time to re-confirm the design.

I retested a 2:5 turn ratio transformer using a single FairRite Products 2873000202 core (about 1/2

inch square and 1/3 inch thick 73 material) two different ways on a generator/ network analyzer/vector impedance test set.

Total loss of two back-to-back transformers was .84dB at 1 MHz increasing, not decreasing, linearly to .98dB at 30MHz. The actual transformer loss would be .42dB at 1MHz increasing to .49dB at 30MHz.

Doubling turns increased the attenuation slope. While 1MHz loss decreased to .69dB per pair, 30MHz loss increased to 1.21dB. This was for a PAIR of transformers connected in series to make a 1:1 transformer. This of course removes mismatch losses, so it is twice the real transformer loss. Actual loss would be .35dB @1MHz increasing to .61dB at 30MHz.

Measuring a second way, I terminated the transformer in 470 ohms. Loss measured .65dB at 1MHz when mismatch loss was included. Since receivers have wide ranges of input impedance, any mismatch error might help OR hurt actual system loss. Factoring out mismatch loss the second measurement indicated about .53dB 1MHz loss with the original 2-to-5 turn transformer and .43dB with twice the turns.

Every measurement has tolerances, and the two different methods do provide different losses because measurement errors affect results differently. Still, it is safe to say doubling turns has a negligible effect on 1MHz loss (which is around .45dB).

Balun Test

[[Home](#)]

Related pages:

[Sleeve baluns](#)   [Balun and Transformer Core Selection](#)

Impedance and SWR Test

This page shows measurements of various baluns and how various baluns compare.

Choke Impedance

This data shows the common mode impedance of the balun. In general, the highest impedance at the operating frequency or over the operating frequency range is desirable. This impedance isolates the antenna from undesired signals on the feedline shield, and prevents antenna terminal voltage from exciting the feedline with unwanted currents. Common mode impedance is directly related to the care in design and construction.

Pay particular attention to the impedance peak in air-core baluns. For narrow-band applications they make excellent baluns.

Unfortunately common-mode impedance is all over the place, as this Smith Chart plot shows:
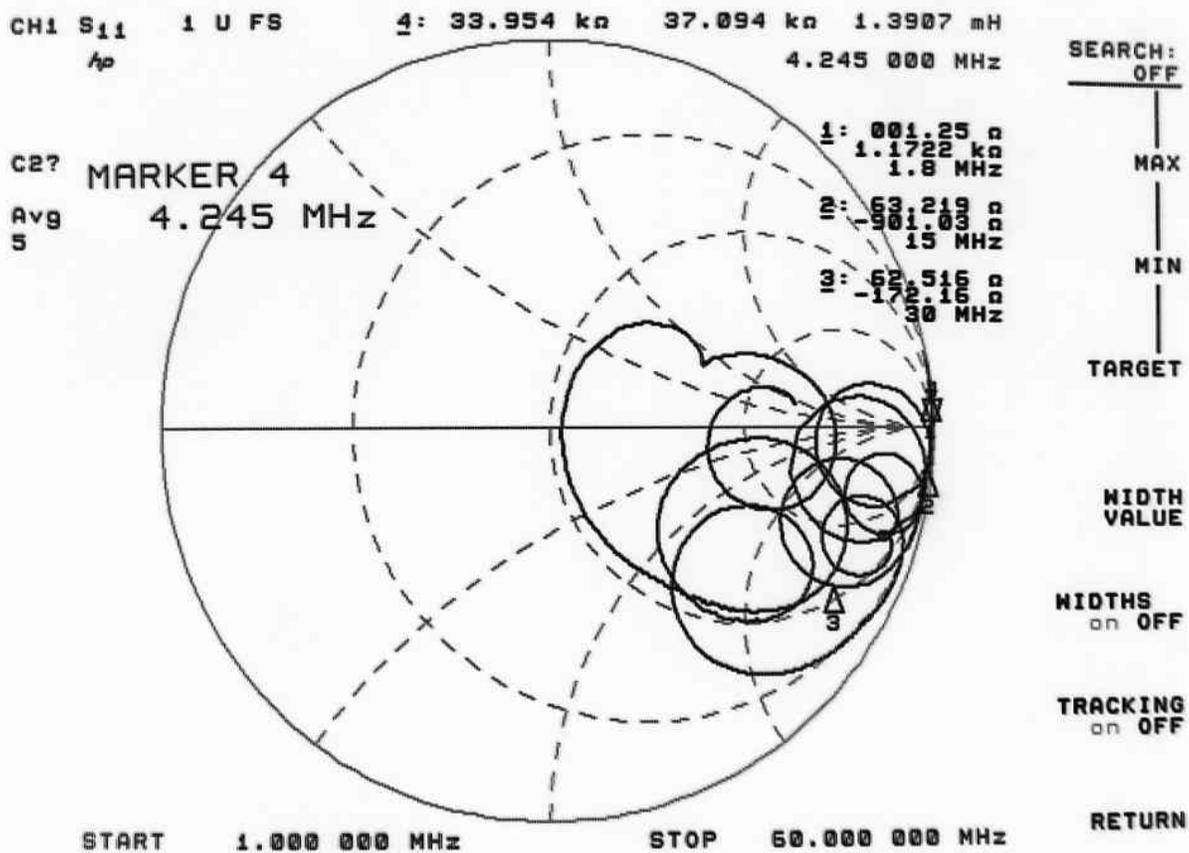
The air-core balun is good only for a three or four-to-one frequency range, unless you pick a winding style and size that places unwanted series resonances outside desired bands.

In contrast a good core-type balun looks like this:

```
CH1  S11    1 U FS      4:  4.488   kΩ   -341.5   Ω    70.168 pF
     hp                                   6.641 875 MHz        SEARCH:
                                                                  OFF

C2?   MARKER 4                         1:  554.37  Ω               MAX
                                           1.4196 kΩ
Avg      6.641875 MHz                        1.8 MHz
  5                                    2: 835.06  Ω               MIN
                                          1.8361 kΩ
                                            15 MHz
                                       3: -153.28  Ω            TARGET
                                           855.16  Ω
                                            30 MHz

                                                               WIDTH
                                                               VALUE

                                                              WIDTHS
                                                              on OFF

                                                             TRACKING
                                                             on OFF

                                                             RETURN
  START    1.000 000 MHz       STOP   60.000 000 MHz
```
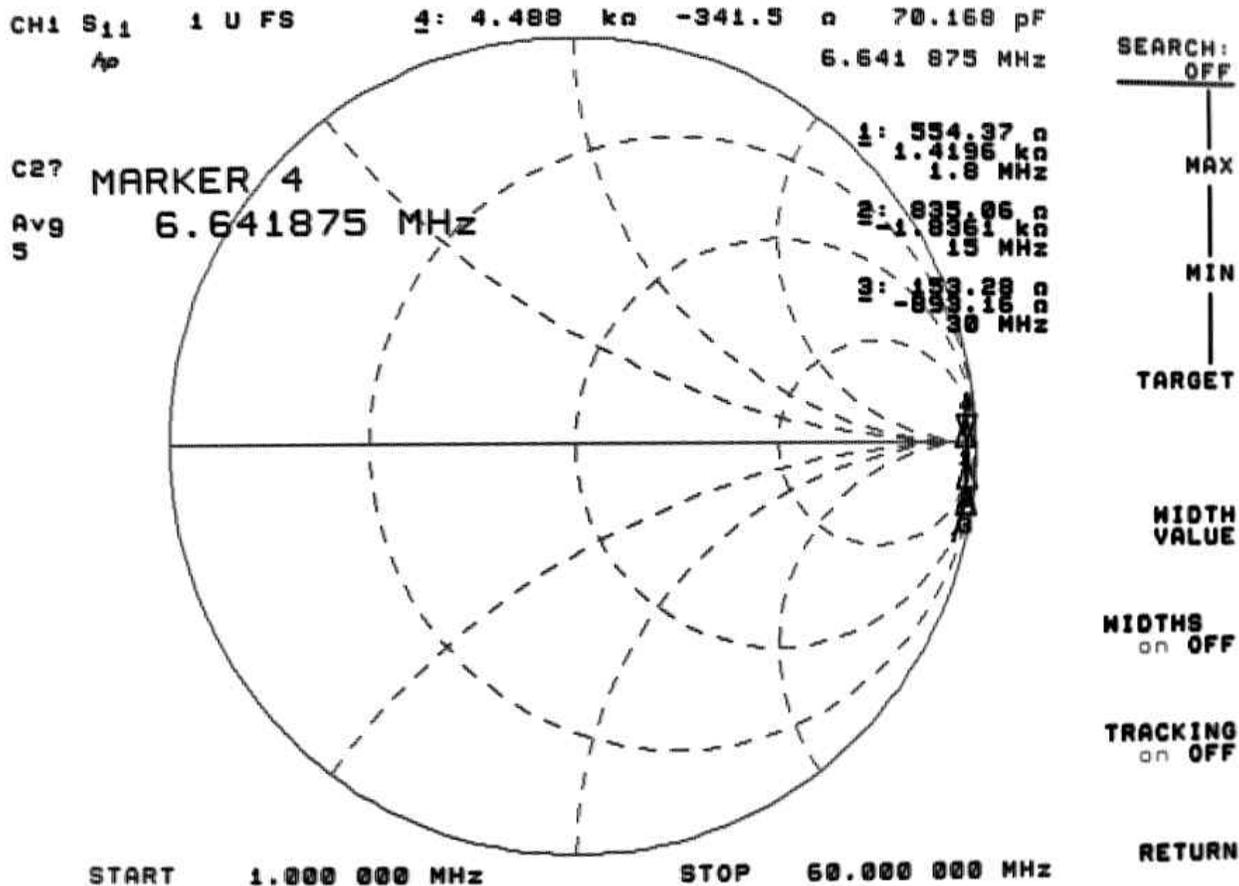
SWR

The lowest SWR is desirable, although any mismatch can often be compensated by adjustment of antenna dimensions. This SWR mainly comes from incorrect wire impedance inside the balun. It may be caused by excessive length of internal leads, or incorrect cable or winding impedance inside the balun. It generally is a construction related problem.

Measured SWR and Choking Impedance

The following data is measured using a currently certified network analyzer with low capacitance test fixture:

| Centaur | DXE | W2DU(1) | W2DU(2) | Force 12 | Scramble | Solenoid | W2AU volt |

Choke Impedance								Note 2
R+X@1.8	84 129j	554 1.4k j	378 617j	230 325j	169 286j	1.67 245j	1 +1.18k j	.488 5j
R+X@15	3.76 2.7k j	835 -1.84k j	727 -611j	761 -10j	883 -105j	11.97 -850j	62 -895j	1.36 42j
R+X@30	143 -729j	153 -893j	284 -440j	610 -296j	538 -381j	73 162j	68 -168j	8.2 68j
Max Z@F	17 MHz	6.65 MHz	7.16 MHz	15.3 MHz	13.24 MHz	6.42MHz	4.25 MHz	60 MHz
R+X at max Z	5.87k -943j	4.5K -340j	1.3K -13j	770 -20j	914 2.25j	42.7k 0j	34K 37K j	75 286j
Min Z@F						27.68MHz	11.7Mhz	
R+X at min Z						10 -2j	198 -252j	
SWR								
F SWR=1.25	6.8 MHz	65 MHz	21.15 MHz		20.2 MHz	note 1	note 1	20.9 MHz
1.8MHz	1.07	1.02	1.03		1.03			1.43
15MHz	1.58	1.04	1.17		1.18			1.2
30MHz	2.16	1.08	1.37		1.39			1.35

note 1: SWR not measured because construction and cable type affects SWR

note 2: This is a W2AU voltage balun. It is only shown as a example of poor shield isolation offered by voltage baluns if the antenna is not perfectly matched to the balun with the feedline exiting the balun at right angles. This type of balun is unsuitable for non-symmetrical systems such as off-center-fed antennas, verticals, or antennas with the feedline paralleling the antenna (even at a fairly large distance).

The W2DU baluns were manufactured by Unidilla. (1) is a Maxi balun and 2 is a 10-40 meter model.

The DX Engineering balun is the dipole balun type DXE-BAL050-H05-P

The scramble-wound choke was about 20 feet of RG8X in a six-inch diameter "bundle".

The solenoid balun was about 60 feet of RG-8X on a 4" PVC thin wall drain pipe coated with rubberized roofing tar.

Power Dissipation and Feedline Common-mode Current Estimates

Note: This section revised 1/2/2003 to correct model error. Please report any other errors to me!

Balun power dissipation is estimated using Eznec to simulate a perfectly balanced dipole.

Here is a copy of the model used:

160m balun EZ

Please be aware I made no special effort to create a "bad antenna" other than I intuitively understand what the worse case condition of feedline length would normally be and I selected that length. I dropped the wire representing the feedline vertically from the center of a perfectly balanced dipole, and made that wire 1/4 wl long.

Here is a view of the model with no balun:

EZNEC
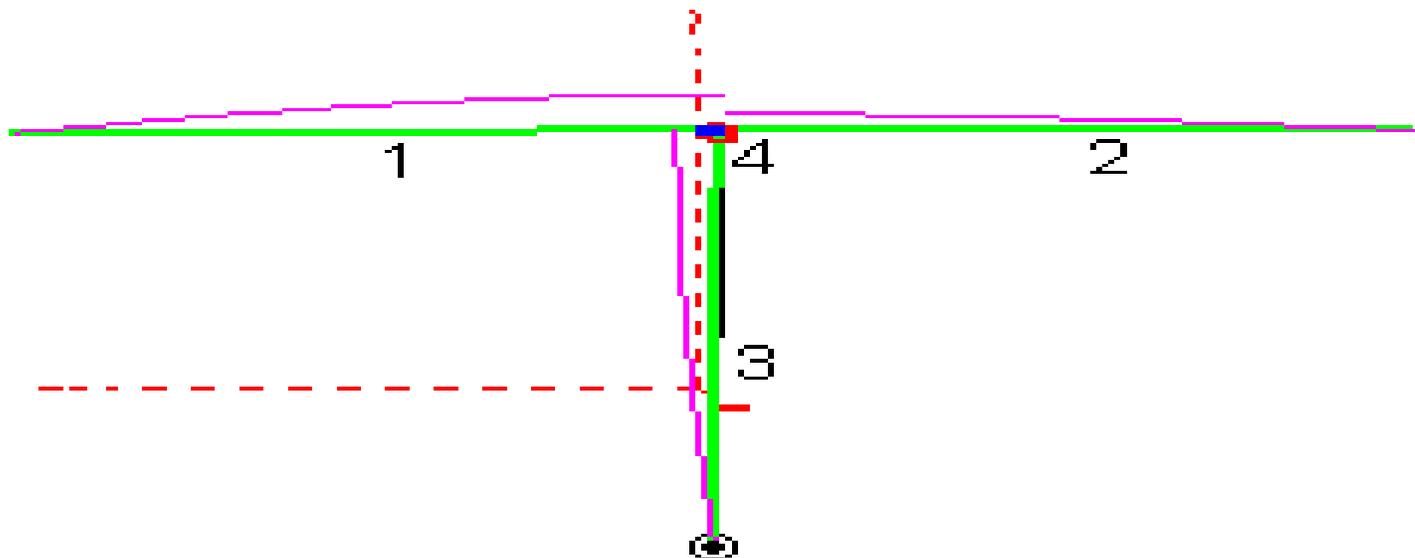
SWR is 1.46:1 power is 1500 watts

Currents at 1500 watts are approximately:

5.65 amperes into wire 1

2.63 amperes into wire 2

3.73 amperes into wire 3 (coaxial cable shield)

Using this model (a 135 foot high 160-meter dipole) we can add each of the balun impedances in the coaxial cable shield and estimate feedline current and power dissipated in the balun:

Centaur	DXE	W2DU(1)	W2DU(2)	Force 12	Scramble	Solenoid	W2AU
0.90 amp	0.12 amp	.25 amp	.47 amps	.57 amps	.87 amps	.16 amp	.4 amps

69 watts	8.5 watts	25 watts	51 watts	55 watts	1.3 watts (26w TL loss)	.03 watt (74w TL loss)	34 watts

From this we can see the following:

- Adding more beads is very inefficient. **W2DU(1)** has about twice the beads as **W2DU(2)**, yet it has 53% of the current and 49% of the power dissipation! This does increase power rating by a factor of four, but it is still too low to prevent balun heating.
- The **solenoid** has (by far) the lowest choking or balancing power loss, but it is 60-feet of RG-8X wound on a 4" diameter form in a single layer. It adds transmission line loss of 74 watts, but since the area is so large it will not overheat.
- The **scramble wound** balun has inadequate impedance since wire length is only 20 feet. It has low loss, but it really isn't acting like a balun (at 6.5 MHz it would be super, having 42K ohms of impedance).
- Of the baluns above, only the DXE, scramble wound, and solenoid would not be overheated in normal operation for continuous Morse CW transmissions.

Perspective of Heat

Think about the heating this way. Imagine you had a 60-watt light bulb. Nearly all of the applied power is turned to heat, and the surface area of the bulb and conduction through the base radiates that heat. Would you hold a 60 watt light bulb?

Now picture a balun core with a surface area a fraction of the size of the light bulb. This core area is enclosed in a case that often has poor thermal conductivity.

- The large air-core baluns mainly produce heat from transmission line losses.
- W2DU and other style baluns mainly have CORE losses. Transmission line losses are negligible since the transmission lines are very short.

Currents

Every ampere of current not going onto the cable shield goes to the dipole's shield-fed leg! The total is not the exact sum, because of phase differences. Here is a view of currents in the dipole with the DXE balun:

EZNEC
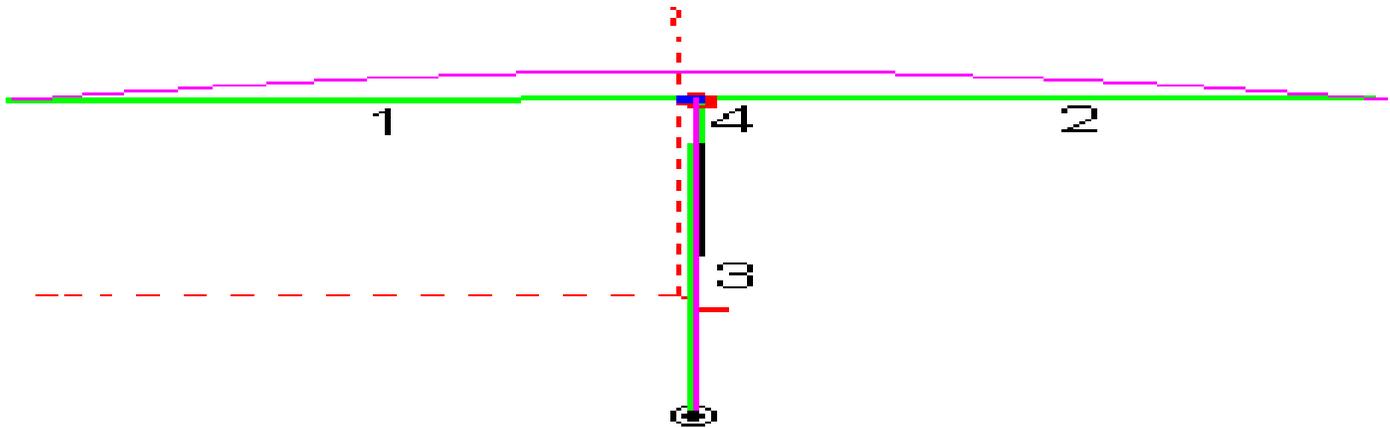
Currents are:

Wire 1   4.52 amperes

Wire 2 4.52 amperes

 Wire 3   .12 ampere

Remember power radiated increases by the square of current. The feedline shield current is now .12A compared to 3.73A without a balun! The feedline radiates about .1% of the power it radiated without a balun. While pattern distortion on transmit may not hurt, the feedline radiation probably aggravates RFI and allows noise to couple into the antenna when receiving.

Remember the model is worse case in the NO BALUN condition. This does not mean every system or most systems will be this bad. This example was only intended to show how bad balance can be and how much power baluns (even with a matched load) can dissipate!

I'll add more data as time permits.

Sleeve Baluns

Sleeve baluns are normally used at VHF and higher. As a general rule, they are not practical at HF.

The sleeve is connected like this:
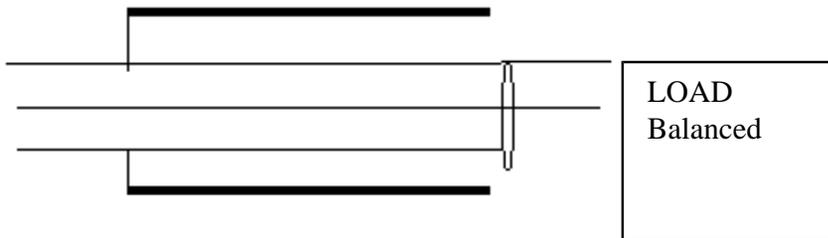
The left hand side of the drawing, once past the sleeve connection point, is UNBALANCED. The right hand side is balanced. The *choking impedance* of the balun is the impedance between the outer sleeve and the outside of the shield of the coaxial cable inside.

The choking impedance represents the same type of impedance we would have from any common-mode choke, such as a coil of cable or a string of ferrite beads would produce. A sleeve balun only works where it is ¼ wl electrical length, or odd multiple thereof.

Obviously we want the highest possible choking impedance, because the balanced load terminal voltage on the shield side to unbalanced (shorted end side) voltage and impedance sets the current though this balun (as it does with any choke balun).

We can use TLA (free with the purchase of ARRL Antenna Books) to calculate balun's choking impedance.

Below is a calculation of sleeve balun choking impedance with 100-ohm Zo 1dB loss sleeve. I used a .5vF inside the sleeve.

```
MG TLA                                                                  _ [8] X
 ┌──────────────────────────────────────────────────────────────────────────┐
 │ Auto      ▼ │ ▢ │▣│▦│ ❖ │ ▦│▤│ A │                                        │
 └──────────────────────────────────────────────────────────────────────────┘

        TLA (Transmission Line), Copyright 1993-1997, ARRL -- by N6BV
                     Ver. 1.0, Mar 05, 1997

   Other Transmission Line
   Length of line:    35.13 ft.
   Frequency:       3.500 MHz
   Transmission line characteristic impedance:  100.0 - j  2.57 Ohms
   Matched-line loss, dB per 100 ft.:    1.000 dB
   Velocity factor of transmission line: 0.500
   Maximum voltage rating of transmission line: 10000.0 V
   Matched-line attenuation =     0.351 dB
   Resistive part of impedance at load:   .00001
   Reactive part of impedance (- cap., + induct., ohms) :  0
   SWR at load: 11184810.00
   SWR at line input:        24.74
   Additional line loss due to SWR:  57.966 dB
   Total line loss:  58.317 dB  (100.0%)

   At line input, Zin = 2474.04 - j 63.68 Ω = 2474.86 Ω at -1.47°
   At    1500 W, max. rms voltage on line:   1927.1 V
   Distance from load for peak voltage =    35.1 ft.
   Maximum rms voltage rating of cable: 10000 V estimated

   Impedance (Z), Frequency (F), Main Menu (M), Antenna Tuner ([T]), Exit (X): _
```

Balun choking impedance 50-ohm Zo sleeve and .351dB internal sleeve loss (sleeve 35 feet long and matched loss 1dB per hundred feet).

```
MK TLA                                                          _ |&| X
Auto     ▼ □    🔍 🔲 🖳 🖨 A
        TLA (Transmission Line), Copyright 1993-1997, ARRL -- by N6BV
                        Ver. 1.0, Mar 05, 1997

     Other Transmission Line
     Length of line:    35.13 ft.
     Frequency:      3.500 MHz
     Transmission line characteristic impedance:   50.0 - j  1.29 Ohms
     Matched-line loss, dB per 100 ft.:   1.000 dB
     Velocity factor of transmission line: 0.500
     Maximum voltage rating of transmission line: 10000.0 V
     Matched-line attenuation =     0.351 dB
     Resistive part of impedance at load:   .00001
     Reactive part of impedance (- cap., + induct., ohms) :  0
     SWR at load:  4793489.50
     SWR at line input:       24.74
     Additional line loss due to SWR:   57.966 dB
     Total line loss:  58.317 dB  (100.0%)

     At line input, Zin = 1237.02 - j 31.84 Ω = 1237.43 Ω at -1.47°
     At    1500 W, max. rms voltage on line:    1362.6 V
     Distance from load for peak voltage =     35.1 ft.
     Maximum rms voltage rating of cable: 10000 V estimated

     Impedance (Z), Frequency (F), Main Menu (M), Antenna Tuner ([T]), Exit (X):
```

Impedance is now 1237 ohms compared to 2474 ohms for 100-ohm sleeve Zo.

As you double sleeve Zo, all other things equal, choking impedance doubles. What happens if we only change loss in the 100-ohm line? We can do this either by doubling the loss per hundred feet or doubling Vf at the same loss, making the sleeve twice as long (and twice the loss).

```
TLA                                                              _ 回 X

Auto  ▪ □    🖭 🖾 📇 🖨 A

          TLA (Transmission Line), Copyright 1993-1997, ARRL -- by N6BV
                          Ver. 1.0, Mar 05, 1997

Other Transmission Line
Length of line:   70.25 ft.
Frequency:     3.500 MHz
Transmission line characteristic impedance:  100.0 - j  5.15 Ohms
Matched-line loss, dB per 100 ft.:   1.000 dB
Velocity factor of transmission line: 1.000
Maximum voltage rating of transmission line: 10000.0 V
Matched-line attenuation =     0.703 dB
Resistive part of impedance at load:  .00001
Reactive part of impedance (- cap., + induct., ohms) :  0
SWR at load: 11184810.00
SWR at line input:       12.39
Additional line loss due to SWR:  60.631 dB
Total line loss:  61.333 dB  (100.0%)

At line input, Zin = 1239.04 - j 63.79 Ω = 1240.69 Ω at -2.95°
At    1500 W, max. rms voltage on line:   1365.1 V
Distance from load for peak voltage =    70.3 ft.
Maximum rms voltage rating of cable: 10000 V estimated

Impedance (Z), Frequency (F), Main Menu (M), Antenna Tuner ([T]), Exit (X):
```

Now we have 1240 ohms with a 100-ohm Zo sleeve. The reason we have less Zo is the attenuation or loss is twice as high. If you look at Matched Line Attenuation above, it is now .703dB versus .351dB in the 0.5 Vf case.

From the above, we observe the following characteristics in a sleeve balun:

1.) The highest possible choke sleeve impedance (largest ratio of balun sleeve diameter to outside of transmission line) is desired. We won't have a good balun if the choking Zo (ratio of sleeve inner diameter to coaxial shield outer diameter) is small.

2.) The balun requires the lowest possible loss over the length of the sleeve. It forms a transmission line from the inside of the sleeve to the outside of the coax. The coax jacket is a dielectric, so we need to keep a lot of air inside of the choking sleeve or the coax jacket will increase loss and reduce impedance, both being very undesirable.

3.) The velocity factor of the sleeve, based on the dielectric between the sleeve and the shield of the coaxial cable we are

trying to balance or choke, is very important to length of the sleeve.

The following construction guidelines apply:

The cable should have a good low-loss jacket or a very large air or low loss dielectric gap between the shield and the sleeve. Since energy is normally confined to the inside of a coaxial cable manufacturers are not concerned about jacket losses. They use outer materials with long life, not low RF loss. It is advisable to use a filler material with a high volume of air to maximize sleeve impedance and minimize sleeve losses.

It is also advisable to use the largest practical diameter sleeve with the smallest diameter coaxial cable inside to maximize choking impedance.

The sleeve length has to account for velocity factor of the sleeve, since the sleeve forms a coaxial transmission line with the outer conductor of the coaxial cable it is intended to choke or decouple.

Common-Mode Noise

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 [Sleeve Balun](#) shows how a sleeve adds impedance, useful for VHF and higher baluns

Receiving [Common Mode Noise](#) shows how lack of a balun can contribute to system noise (it applies to transmitting antennas as well)

[Longwires, Verticals, and Baluns](#) shows how unbalanced antennas can have similar problems

[Balun and Core](#) selection for transformers and baluns

[Transmitting baluns](#) on testing transmitting baluns

Common Mode Noise Isolation

Common-mode currents can be detrimental to antenna system noise or directional performance. A quick look at systems using common-mode currents demonstrates how effective they are at causing radiation. If common mode currents can radiate effectively; they can also transfer unwanted signals and noise *into* our antennas when receiving.

Many antennas actually function because of common mode currents. Two popular examples are the CFA and [EH antennas](#). Both become significantly poorer radiators if common-mode currents on feedlines are eliminated. Another recent example appearing in Antennex's compact antenna articles was a thick stub "vertical" with no counterpoise. You find an example on the [baluns and vertical](#)s page of how poor some large grounds can be.

A receiving example of an antenna that works because of common-mode excitation is the "Snake"

antenna. This system accidentally (or intentionally) induces common-mode on a cable shield in order to receive signals. The entire shield picks up signal, the Snake is simply a reverse-fed random wire lying on the ground.

There are many examples where designers intentionally use common-mode currents. Examples are found in textbooks, such as the "Antenna Engineering Handbook" by Jasik on and around page 22-6.

The antennas below, copied from Jasik's textbook, outline the derivation of a skirt collinear antenna from a simple feedline with the open end terminated by a "stinger".

22-6   VHF AND UHF COMMUNICATION ANTENNAS

Two or even three skirts may be added to the mast as shown in Fig. 22-5c, but reduced excitation of the lower sections diminishes the effectiveness of each additic skirt.[6] Thus, while the multiple-skirt coaxial antenna resembles a collinear arra; in-phase half-wave elements, its gain is not as great. Typical gain values are ta lated in Table 22-2, but since no test data are available for this type of antenna, values given are merely engineering estimates.

(a)   (b)   (c)
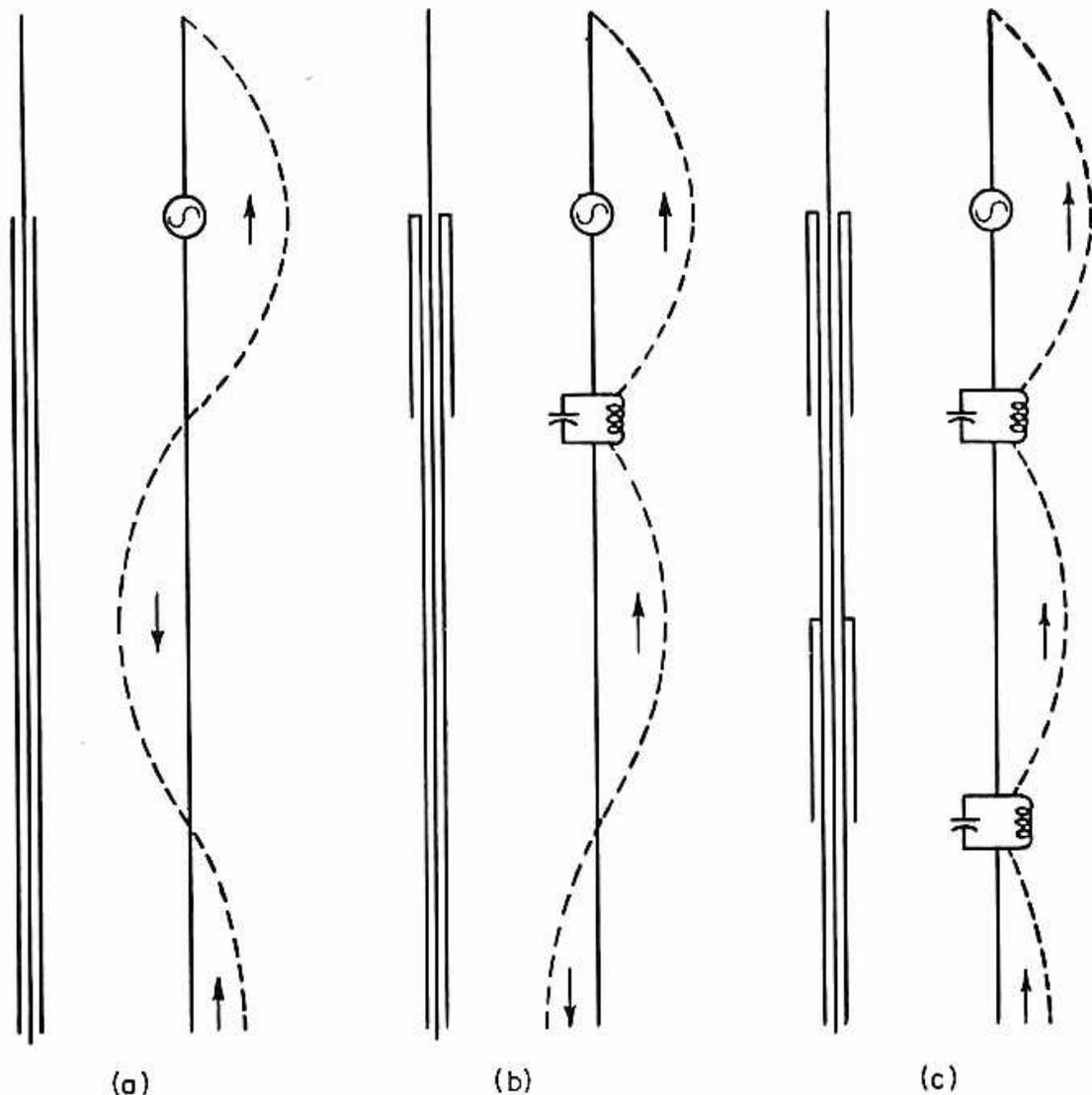
Fɪɢ. 22-5. Evolution of multiple-skirt coaxial antenna.

(a) (b) (c)

FIG. 22-5. Evolution of multiple-skirt coaxial antenna.

The relatively large diameter-to-length ratio of a practical skirt produces an effect which requires that the exterior length of the skirt be reduced by a factor of 0.8 to 0.9 and the interior length by from 0.95 to 0.98.[6] This is accomplished cutting the skirt to the proper exterior length and inserting a dielectric slug in skirt to increase its electrical length.

The center conductor termination in these drawings could easily be a ground rod (in the case of a Snake) or an antenna like a Beverage or loop. The termination does not have to be an "open circuit" 1/4 wl stinger that intentionally radiates!

Looking at (a), we find by hanging any low impedance on the end of a coaxial cable the shield is excited by common-mode current.

The electrical equivalent is just as if the transmitter or receiver (generator symbol in the drawings) is located at the end of the shield. This causes the outside of the shield to act like a longwire antenna.

Unless the coaxial shield connects to a zero resistance ground, current with flow on the shield.
Looking at (c), we find even multiple sleeves appearing as parallel tuned high-impedance circuits do not fully decouple a shield! It takes grounding and series impedance to do a good job.

Common Mode Currents and Receiving Antennas

Analyzing our antennas, we often forget grounds are not perfect. We make assumptions that four radials, or worse yet two radials, form a perfect groundplane. Even a groundplane antenna many wavelengths from earth with four radials has considerable common-mode currents on the feedline. Consider the following model of a "perfect" Ten Meter groundplane using four perfectly horizontal 1/4 wl radials spaced every 90-degrees with a 1/4 wl feedline hanging vertically and attached to the radials. The main element current was set at 100.
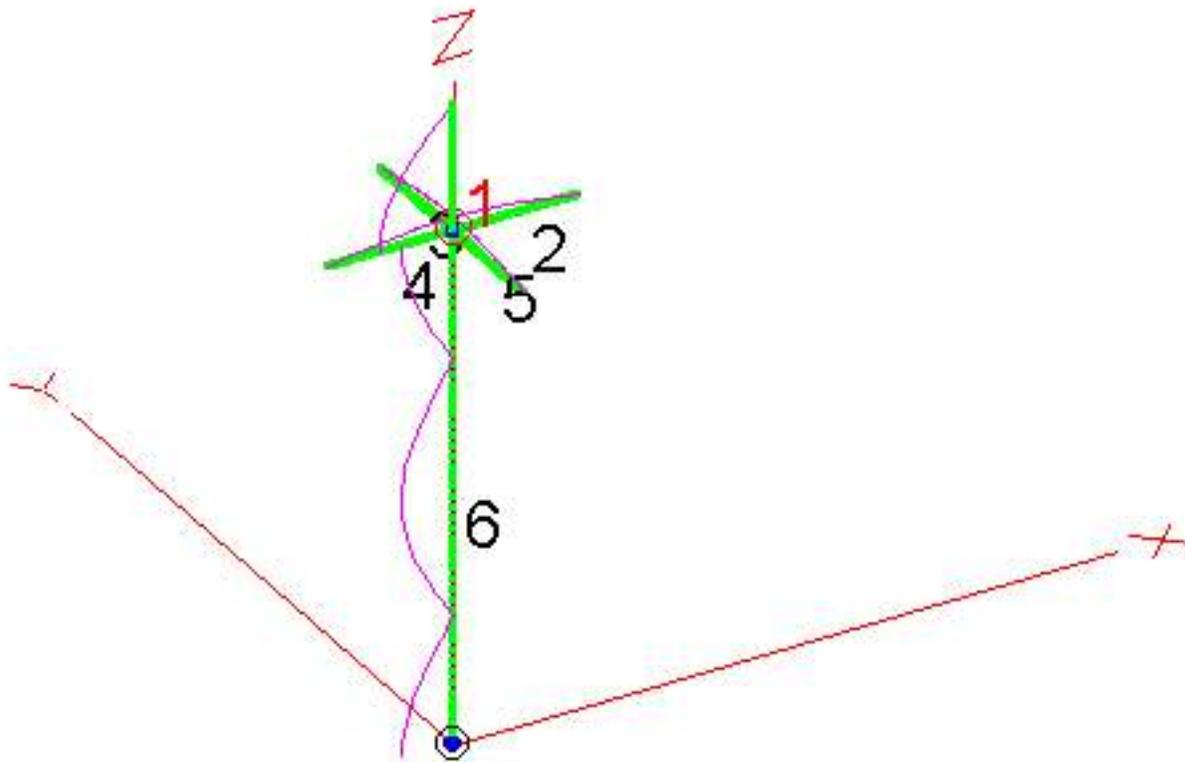
EZNEC ver. 3.0
Groundplane 12/14/02 6:20:40 PM
--------------- CURRENT DATA ---------------
Frequency = 29.95 MHz.

Wire No. 1 Main Element	100
Radial wires	19.71
Feedline shield at GP	69.96
1/4 wl from GP	2.12
1/2 wl from GP	71.3
3/4 wl from GP	.70557
At ground end of feedline	71

A glance at radial current shows the bulk of ampere-feet (ampere-feet, or current over spatial distance, determines E-M radiation levels) is on the feedline shield, not the antenna! Radiation from the feedline would be severe, yet most amateur antenna designers claim with only four radials, or worse yet two radials, no balun is needed! The claim that four radials makes a "perfect ground" is false.

Why do we depend on a simple ground rod with 50 or more ohms RF resistance to clamp a coaxial cable shield to ground?

Receiving Systems

Admittedly the above antenna is a worse-case example of feedline length and grounding, but even the best cases could cause problems. A best-case system might be "nearly perfect" when transmitting, but it could be a disaster receiving when even minor amounts of conducted noise are present on the station ground. Noise paths exist through station wiring. Only the shunting impedance of ground connections and series impedance of the feedline shield prevent excessive unwanted noise ingress at the antenna feedpoint.

Very small levels of conducted unwanted noise often go unnoticed in large high-level transmitting antennas. Noise ingress would not be an issue if local noise levels on power lines are very low, especially if the antenna has substantial common-mode feedline noise rejection.

If a feedline is very long and lies directly on or is buried in the earth, ground losses can attenuate conducted noise or unwanted common-mode signals. Unfortunately, we almost never know if the feedline shield is contributing noise, because it is nearly impossible to measure the common-mode noise contribution of the feedline!

Measuring Common-Mode Noise

We sometime hear suggestion that we test a system for noise ingress by disconnecting and replacing the antenna with a dummy load. This idea actually has no theoretical foundation at all. The dummy load would change system common mode greatly.

The only real test would come from a dummy load with the same connections and impedances (both differential and common mode) as the actual antenna. In other words the test load has to be the actual antenna to keep feedline common mode ingress the same! Obviously this is a useless test!

The best approach is to use preventative measures in initial system design and installation. Quite often the cost of being safe is less than a few percent of the initial system expense.

Analyzing Systems

This circuit is simplification of typical common-mode paths in Beverage, EWE, and other similar antenna systems:
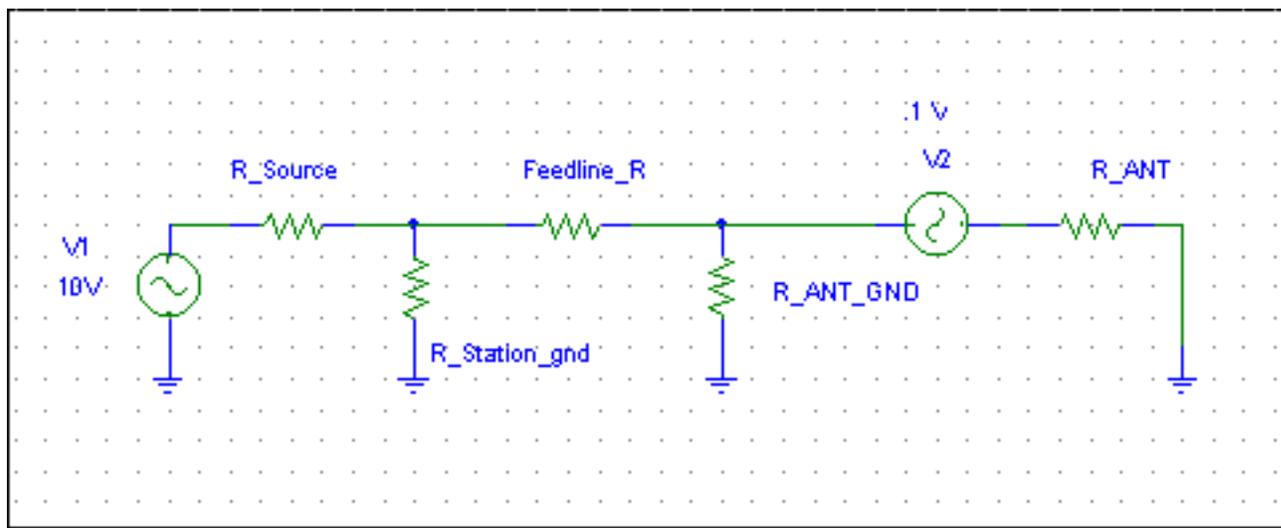
R_Source and V1 represent the source creating voltage across R_Station_Gnd, the station's ground impedance.

Feedline_R is the equivalent series-impedance of the feedline shield.

Current through the feedline shield path develops a voltage across R_Ant_gnd, which represents the earth connection ground impedance at the antenna.

V2 is a voltage source representing desired signals, while R_ant is an impedance representing the sum of the coaxial differential input impedance presented to the antenna (from the desired signal path into the coax) and the actual antenna impedance.

Using the circuit below, we can find the attenuation. Assume:

R_source is 90-ohms

R_station_ gnd is 10 ohms

R3 (the coax shield) is 500 ohms

R5 is the combined series resistance of antenna impedance and impedance presented by the feedline matching system, is 1000 ohms

In a typical system where a single six-foot or deeper rod (the earth's skin depth prevents deeper ground rods from decreasing resistance substantially) is driven into typical soil, R_ANT_GND will typically be between 40 and 120 ohms, assume 100 ohms.

We have the following results:
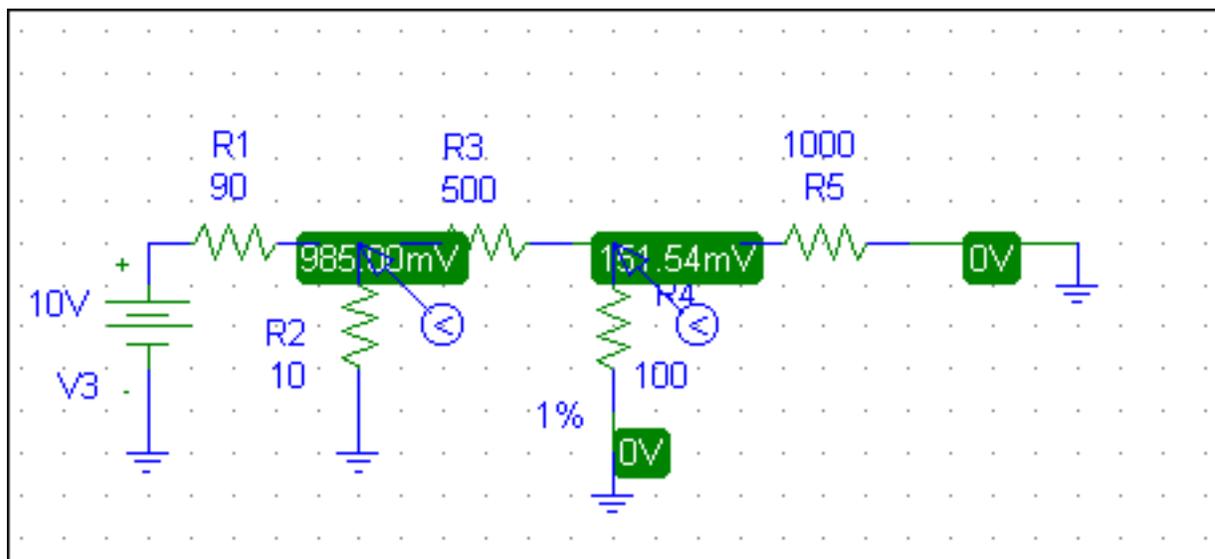
Using the model above, only ~1 volt of common-mode voltage across the station ground results in .152 volts driving the feedline *exactly as a signal from the antenna would*. Path attenuation from station ground to the feedline's differential input at the antenna is 20log 151.5/985 or **16.26dB.**

Changing the ground resistance to 10-ohms results in:
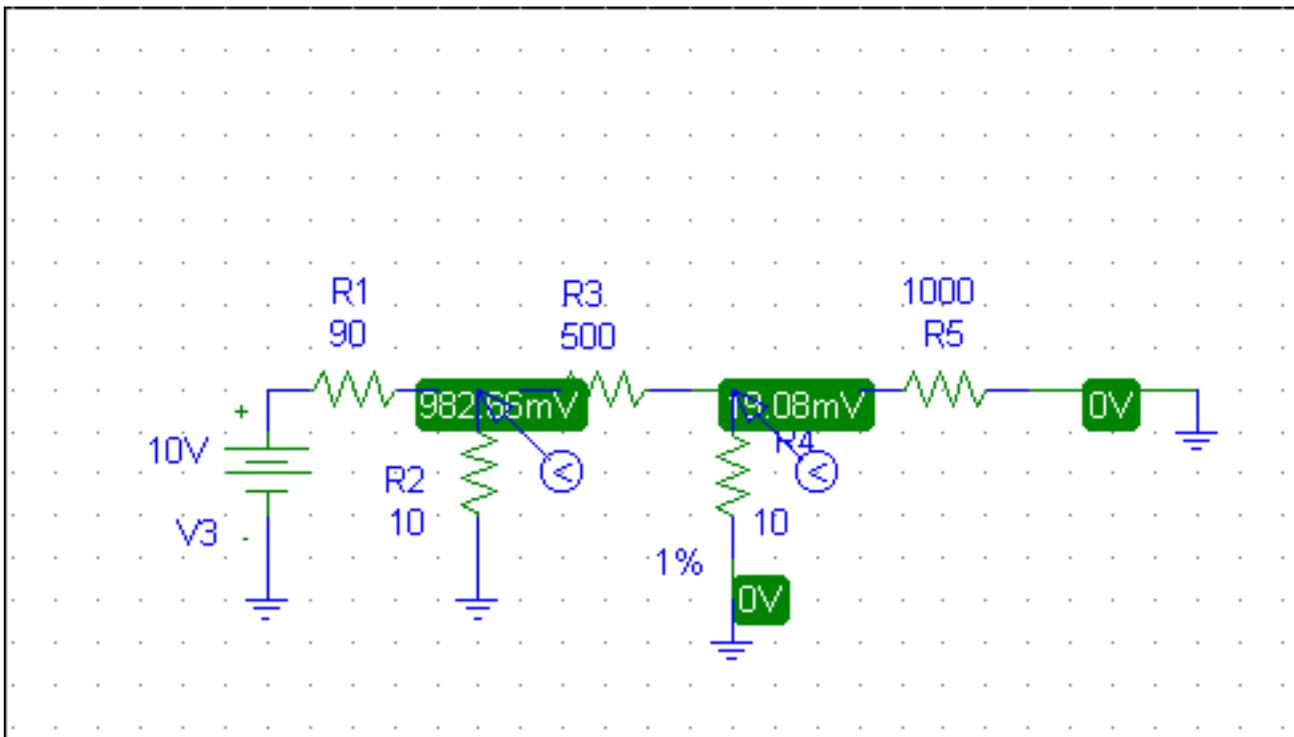
19.1/982.7 or *~34dB attenuation* of common-mode noise. Increasing R3 by adding beads has a similar effect. If R3 is effectively made ten-times larger, attenuation is in the 30dB range.

Obviously it takes a combination of reducing ground resistance and/or adding series impedance on the cable shield to significantly isolate any low-noise receiving antenna from conducted ground noise over

the feedline's shield.

We sometimes observe much less noise on transmitting verticals after installing a large effective ground system. Decreasing ground impedance at the antenna reduces common-mode excitation of the antenna feedpoint and reduces noise ingress, although adding a [feedline choke](#) would sometimes help.

Solutions

A typical isolation scheme would be to use an isolated primary and secondary in the matching transformer, and ground the feedline shield some distance away from the antenna's signal ground. This will introduce several thousand ohms of reactance in the common-mode signal path, as well as provide another path to earth for common-mode noise.

Another method, in cases where the feedline can not be isolated through a floating primary in a matching transformer, is the use of multiple independent ground rods with a series of choke baluns between each. This forms a multi-section pi attenuator, making even modest choke impedances effective. As an additional benefit, lightning paths are disrupted by this method.

Summary

Noise contribution can vary with time. A receiving antenna's ground connection resistance varies with soil moisture, and sources of noise come and go. As noise levels and grounding changes noise contribution as a ratio to antenna noise will change. The fact we can not readily measure noise contribution by substituting dummy loads further complicates the issue. Real systems are vastly more complex than the simple analysis above.

Since we can't easily measure noise contribution, we shouldn't take chances. It makes no sense to gamble that unwanted signals (from wrong directions) or noise are so low that they will never contribute to noise in a special antenna installed to reduce noise and interference.

While isolating feedline common mode effects from the antenna and antenna's ground may not reduce noise, isolation can generally be achieved at virtually zero time and material cost. With the low cost of prevention in mind, it is shortsighted at best and foolish at worse to not isolate a feedline shield from any low-noise antenna's signal ground path.

Follow these rules:

1. **With small magnetic loop antennas, make sure the antenna is [properly balanced](#)**
2. **Lay feedlines on or bury them in the ground so earth losses reduce shield current**
3. **Be sure shield connections are good**

4. **If noise levels are high or antenna sensitivity is very low, isolate the feedline from the antenna as it approaches the antenna by using [choke baluns](#)**
5. **Ground the feedline a few dozen feet away from the antenna**
6. **Avoid autotransformers. Instead use isolated primary and secondary winding transformers**
7. **Use an independent ground on the antenna. Do NOT connect it to the coaxial cable shield**

Noise

[Home] [MFJ-1025/1026] [Common-Mode Noise] [RADIATED and CONDUCTED NOISE]

Also see: Receiving Antenna Design

To see world QRN (delayed): Lightning world map

To see my local QRN (nearly real time): Lightning USA Map

My local noise level quiet night from NE is -127dBm, 350Hz BW, from pair of ~800ft broadside Beverages ~375ft spaced.

The noise that limits our ability to hear a weak signal on the lower bands is almost always an accumulation of many signal sources. Below 18 MHz, the noise we hear on our receivers (even at the quietest sites) comes from terrestrial sources. Receiver noise is generally a mixture of local groundwave and ionosphere propagated noise sources, although some of us suffer with dominant noise sources located very close to our antennas.

Our locations fall into three basic "radio" categories that may or may not be related to our actual communities:

Note: noise levels quoted in this text are the average of three independent studies by Bell Labs, FCC Land Mobile Advisory committee, and the Institute for Telecommunication Sciences. Rural data are actual measurements of summer noontime and winter midnight noise at my location, several miles from high voltage transmission lines and far from any industrial or suburban populations.

Urban

In urban-type noise situations, noise arrives from multiple random sources through direct and groundwave propagation from local sources. One or more sources can actually be the induction-field zone of our antennas (in most cases the induction field dominates at distances less than $1/2\lambda$). Urban locations are the least desirable locations because typical noise floors average 16dB higher

than suburban locations. There is often no evidence of winter night noise increase on 160 meters, since ionosphere-propagated noises are swamped out by the combined noise power of multiple local noise sources. Much of the noise sources are utility distribution lines, because of the large amount of hardware required to serve multiple users. Other noise sources are switching power supplies, arcing signs, and other unintentional man-made noise transmitters.

Suburban

Suburban locations average about 16 dB quieter than urban locations, and are about 20 dB noisier than rural locations. Noise generally is directional, arriving mostly from areas of densest population or the most noise-offensive power lines. Utility high-voltage transmission lines are often problematic at distances greater than a mile, and occasionally distribution lines can be problems. The recent influx of computers and switching power supplies has added a new dimension to suburban noise.

There is often a small increase in nighttime winter noise at exceptionally quiet suburban locations. This increase occurs when propagated terrestrial noise equals or exceeds local noise sources.

Rural

Rural locations, especially those miles from any population center, offer the quietest environment for low-band receiving. Daytime 160 meter noise levels are typically around 35-50 dB quieter than urban, more than 20 dB quieter than suburban locations. Nighttime brings a dramatic increase in low-band noise, as noise propagates in via the ionosphere from multiple distant sources.

Primary local noise sources are electric fences, switching power supplies, and utility lines. I can measure a 3 to 5dB daytime noise increase in the direction of two population centers, Barnesville (population 7500, distance 6 miles) and Forsyth (population 10,000, distance 6 miles) Georgia.

Typical daytime noise levels, measured on a 200 foot omni-directional vertical, are around -130 dBm with a 350 Hz bandwidth (noise power is directly proportional to receiver bandwidth) . Noise power increases about 5 to15 dB at night, when the band "opens". As in the case of suburban systems, directional antennas reduce noise power.

Nighttime is the "big equalizer", reducing the advantage of location as distant noises increase with improved propagation.

Polarization

Noise is generated by randomly polarized sources. Noise polarization is *filtered* by the method of propagation.

Noise arriving *from the ionosphere* is randomly polarized. It arrives at whatever polarization the ionosphere happens to favor at the moment. It has the same ratio of electric to magnetic fields as a "good" signal.

Noise arriving *from sources within a few wavelengths* of the antenna is also randomly polarized. It has NO dominant field, it can either be electric or magnetic field dominant.

Noise arriving *from ground wave sources* some distance from the antenna is vertically polarized. This is because the earth "filters out" any horizontal components. Horizontal electric field components are "short circuited" by the conductive earth as they propagate, and are eliminated.

To hear a demo of noise and directivity, go to the DX Sound page.

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MFJ-1025/1026

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Noise Phasing and MFJ-1025/1026 Data

The original concept of the 1025 is based on a bridge phasing system. Many phasing boxes, such as the QRN Squasher, do not rotate phase. They are based on the concept that a 180-degree inversion allows rotation of phase when mixed through a potentiometer, but that does not occur. The only thing mixing a 180-degree shifted phase signal with another base phase signal does is vary the amplitude, and produce an abrupt 180-degree flip at the zero-crossing point. Some people have copied the flawed QRN-Squasher concept, and offer similar products.

I recently reviewed another phasing system design (on paper) which is implied to be the "ultimate". It uses a simple R/C phase advance circuit, not a bridge. The maximum phase rotation of such a system is limited to less than 90 degrees. With 180-degree or zero-degree inversion in a later stage, that allows only a 150-160 degree slice of phase choice. Even transposing inputs would only bring that to 320 degrees. Not only that, dynamic range is limited by a very poor choice of operating parameters for the FET driving the phasing system.

To have a perfect product you need 360-degree rotation with no level change. Even with perfect design, phasing systems are tedious to adjust. Other systems rotate phase over a wide range, but have severe level change as phase is rotated. The effect of a level change is you have to "walk" back and forth with phase an level every time you adjust phase! When level changes with phase adjustment, phasing units are even more unpleasant to adjust than a proper system is.

With a *well designed* bridge system, amplitude shift is minimal. The MFJ1025 series, despite being manufactured as cheaply as possible, outperforms any other unit I have tested (by a large margin) over its intended frequency range of 1.8-20 MHz.

The phasing system at W8JI, used to pinpoint signal directions of transmitters or jammers more than to actually cancel noise, is a more sophisticated version of the 1025 with higher dynamic range. It uses tandem bridges, and is calibrated in degrees shift. Phasing is spread over a wide, linear, control range. Phase rotates a full 360-degrees with less than 1dB amplitude shift. It is possible to resolve as little as 1- degree directional difference in arriving signals.

A Word About Dynamic Range and Noise Floor

160-meter DX'ers have the most demanding noise floor and dynamic range requirements of anyone

using a phasing unit. 160 ops not only must hear exceptionally weak DX stations using small inefficient antennas and low power, they must contend with multiple 1500-watt signals (often with excellent antennas) parked just hundreds of Hz away! CW operators often listen to CW signals with 250Hz BW systems, pushing noise floors up to 50 times lower than 10kHz filters would produce at any location. Amplifiers and other components must also handle the large powerful window of AM BC stations just below 160, as well as the SW BC stations above. At my house the accumulation of signals from stations *outside* Amateur bands is enough to light a small 12v light bulb when using a 15dB amplifier!

Obviously what works well at MW for sorting 10kHz spaced wide bandwidth signals apart is entirely different. Here we have the strong window of stations, but we are looking in that strong window for other stations that often are strong (compared to 100-watt transmitters with poor antennas). Noise floor is also much higher, and so dynamic range and noise floor are of almost no concern. The critical performance parameter is mostly adjustment range, and not dynamics.

SW BC listeners need something a bit different yet.

Everyone should be cautious assuming a device optimized for one application somehow makes it fit the other.

The only drawback of the higher noise floor of the MFJ-1025/1026 is you *must place it after any amplifiers used with low-level low-noise antennas.* The MFJ-1025/1026 noise floor is typically around 14dB.

The dynamic range and noise figure of my personal unit is excellent. Unfortunately such a system is impractical for amateurs and SWL's because of cost. Each unit has about $300 in raw material cost, plus the assembly time. The FET's alone are 28-volt devices capable of several watts RF power while providing a 1.5dB noise figure. But for typical operation the 1025 is almost as good. If you want to spend several hundred dollars, I can sell you a system that works almost perfect in almost any application!

Even with MFJ's well-known QC issues and affinity for building things cheap (not good) I still recommend the MFJ-1025 over all other mainstream units. It has the potential to be better than anything else with only minor changes, and is certainly betetr than all other mainstream units like the ANC-4 or QRN Squasher just as it is shipped. If you want a custom modified unit, and can not make the changes yourself, I would be happy to help.

GAIN MEASUREMENTS

The following table are gain measurements for the MFJ-1025/1026. Wiltron Network Analyzer (50 ohm detectors), transmission loss mode.

Stock Unit and Filters 50-ohm input and output

Aux Port Gain

Frequency   Gain (dB)

0.5MHz	-35
1.2 MHz	-16
1.5 MHz	-9.3
1.7MHz	-6.7
1.8MHz	-5.4
1.9MHz	-5.2
2.0MHz	-4.8
2.5MHz and higher	-4.4

From the above losses, deduct -2dB (or add 2dB!) if the input protection light bulb is bypassed or replaced with a jumper wire. I.e. 1.8 MHz gain becomes -3.4dB with the lamp removed.

Main Port Gain

Frequency   Gain (dB)

0.5MHz	-34
1.2 MHz	-14.1
1.5 MHz	-8.0
1.7MHz	-5.8
1.8MHz	-5.0
1.9MHz	-4.2
2.0MHz	-3.7
2.5MHz and higher	-2.3

Bypass is essentially lossless.

Gain Compression and IM Dynamic Range

With gain controls wide open and pre-amp disabled, blocking begins at about -3dBm. Third order intercept is about -21dBm. This is generally well beyond what most receivers are capable of obtaining. (Click here to see the **receivers** page on this site.) The drawback of the 1025/1026 is the lack of any form of filtering, other than the high pass filter. This means the MFJ-1015/1026 sees a very wide window of signals.

With that in mind, it is questionable if removing diodes actually is worth the risk of RF or ESD damage. It would be much better to add a simple bandpass filter system for the band you are operating. However, there is some improvement offered by removing diodes. IM and blocking also can be improved by increasing quiescent current in Q5 and 6, as well as Q4. The bias modification is generally too complex for most people to do, and adding a bandpass filter in from of the unit would be much more effective anyway.

Removing Diodes and Lamps

IM dynamic range will improve about 6dB with this modification. The lamp will improve gain and noise figure of the AUX port (reduce loss) by about 2dB.

If you *never* transmit through the unit:

1.) Remove C5

2.) Remove D1,D5 at the output, remove D12 and 13 (or D10 and 11 in 1025) and D6 and 9 at the inputs.

3.) Replace Lamp1 with a jumper wire.

 NOTE: If you transmit through the unit do NOT remove C5 and D1 and D5.

Sensitivity to Load Impedance

Like any typical amplifier system, this unit is sensitive to load impedance. Since it has an emitter follower at the output, it likes to see a low load impedance. Gain will increase about 5dB if you add a 1:4 step up transformer at the emitter of Q4.

The following are modifications or changes to the MFJ-1025/1026:

Gain Improvement Mod.

Increases gain and dynamic range by approximately 5dB (assumes your receiver has a 25-ohm or higher input Z):

1.) Cut the trace from C6 to D1/D5 near the small dip relay.

2.) Add 1:4 step-up transformer, low impedance side towards C6 and high impedance side to load.

Transformers are available, including US shipping, for $6.

Filter Modifications

The MFJ-1025/1026 contains BCB filters that roll off at about 1.9MHz. To extend operation lower in frequency:

Remove L3, L4, L5, L6.

Short C8 and 16

Increase the value of C12 and 13 by the same factor as the decrease in frequency. For half the frequency, double C12 and 13.

Caution: DO NOT remove filters if you intend to use this unit for weak-signal reception on Amateur bands above 1.8 MHz, unless you are sure you do not have strong BC signals.

C12 and C13 mods will reduce upper frequency limit proportionally to the change.

Extending Phase Range

Phase range of the MFJ-1025/1026 is limited by the ratio of R16 to reactance of C12 or C13. Normally phase will rotate through at least 130-degrees. SW3B inverts phase 180-degrees, and this moves the phase shift to the other side of the circle. Typically the MFJ-1025/1026 has about 280-degrees of phase rotation.

The remaining 80-degrees can be covered a number of ways. You can swap inputs (only when using it entirely passive), or modify the phasing network. I also have successfully added a relay that swaps the outputs of the low-pass input filters. A second method is to add an additional capacitor in series with the ground lead of the bottom of R16 (the wiper remains grounded) back to the junction of C14 and R17.

I'd recommend swapping the inputs, it is easier and does not degrade amplitude response or bandwidth.

Receiver_Tests

Here a few measurements of receivers, using 500Hz filters.

MDS is a measure of sensitivity. -135dBm is 10dB more sensitive than -125dBm. This number doesn't matter very much in what you actually hear. <u>**The real test is if you hear a very noticeable noise increase when you connect an antenna to the receiver. If you hear an obvious noise increase when you connect an antenna instead of a dummy load, your receiver is sensitive enough! You should check sensitivity at the quietest time with the narrowest selectivity you use on every antenna you use.**</u>

Contrary to folklore and hyperbole, there isn't a receiver sold today that can dig into noise more than others on CW based on sensitivity or the use of a DSP or multiple DSP systems. The exceptions are:

- Increased selectivity will reduce noise
- Poor AGC design or detector problems can cause mixing of signals and noise

If you read a review that claims a receiver made weak signals appear from nowhere, you better keep a wary eye on the rest of the review. It is possible for a receiver to be abnormally bad, but it is not possible for a receiver to work better than other properly working receivers based on sensitivity.

BDR is blocking dynamic range. This is the point where a strong signal either 2 or 10kHz just starts to make your receiver lose sensitivity. The bigger the number the better, ESPECIALLY at 2kHz spacing. <u>**The number you want here is probably around 80dB or more if you live in a reasonably quiet location and work weak signals on crowded bands. If you run two transmitters on the same band or have a neighbor who operates near your frequency, you almost certainly need more dynamic range. I'm in a very quiet rural location and have very directive antennas, and 80dB blocking DR suits my requirements just fine most of the time.**</u>

IMDR is intermodulation dynamic range. This is the single most important number when comparing receivers. This is where two or more strong close-frequency signals mix and generate a new phantom signal or multiple tones in a adjacent frequency SSB signal mix with themselves and make what sounds

like splatter. The measurement is made just at the point where the phantom signal level is high enough to interfere with the weakest signal your receiver can detect. IMDR is a measure of how badly your own receiver causes problems you might blame on other people. Bigger numbers mean better receivers. It is most important the **2kHz** number be good. The 10kHz test number doesn't mean nearly as much, because almost any radio is good enough at 10kHz or wider. **Some number above 80dB is enough to stay out of trouble 99% of the time. If you are in a noisy location, you obviously need less performance. 85dB keeps my receivers at the point where poor quality external signals cause nearly all off-frequency problems. With 85dB IM3DR only a few of the strongest stations cause my receiver to make its own internal problems.**

Be aware the very good Drake receivers in the test below are essentially gutted and rebuilt receivers. They DO NOT use Drake mixers or amplifiers! **Stock R4C's** suffer from severe problems. See the notes!

CLICK HERE TO SEE SORTED BY CLOSE-SPACED IM3 PERFORMANCE, THE MOST IMPORTANT TEST!

Model serial	MDS	2kHz BDR	10kHz BDR	2kHz IMDR	10kHz IMDR
FT1000MP 8N370165	-135	73.5	84	62* 75.5** note 1	70.5* 80** note1
FT1000D	-134	68	85	72	81
FT1000D/W8JI mods	-136	73	87	89	88
FT1000MK V stock	-131	74	86	60 note1A	77 note 1A

FT1000MK V W8JI mods	-131	80	89	85	90
IC-751A	-140	83.5	98	79	91
IC-756 PRO	-133	N/A	78	67	77
IC-7800	-127	80	98	78	88@5kHz
R4C S/N 17816 stock	-132	57	109	48	82 note2
R4C S/N 22291 stock	-139	80	116	68	86 note3
R4C 17816 med mod	-140	96	115	92	103 note4
R4C heavy mod	-143	127	131	118	119 note5
TS-870 Inrad 400Hz IF filter mod (AIP)	-139 (-125)	74 (82)	94 (98)	86 (81)	90 (91)
TenTec Orion	-129	98	100	92	92

Note 1: The FT1000MP has the same inherent problem as the FT1000 and FT1000D. A noise-blanker FET is left on *even when the noise blanker is turned off with front panel NB and NB-W switches*. This FET connects directly to the IF, and overloads when strong signals are present. Unlike the FT1000 and FT1000D, the front panel noise blanker level control affects the FT1000MP IMD performance. * is with the NB gain control fully advanced, ** is with the NB gain control fully off. ***In ALL cases the NB off-on switch is OFF!!***

Note 1A: Yaesu copied the same basic NB from the FT1000D, and so this radio also has a NB problem even when the NB is switched off. This test was with NB gain menu set for minimum gain. The radio is worse when NB gain is increased in the NB gain menu.

Note 2: This is a MOSFET 2nd mixer R4C, aligned and in good shape but with no changes. This receiver is almost useless for weak signal CW work of any kind because of poor IM performance and filter blow-by.

Note 3: This is a tube 2nd mixer R4C, aligned and in good shape but with no changes. This receiver is almost useless for contesting or crowded bands because of poor IM performance and bad filter blow-by.

Note 4: This is the same receiver in note 2, but with the following changes:

1. Increased screen voltage on RF amplifier to 130 volts.
2. 6J6 first mixer with oscillator injection increased to 5 volts.
3. Diode double-balanced 2nd mixer with +10dBm injection.
4. 10dB of gain removed from first mixer and first IF and redistributed behind the narrow filter.
5. Sherwood Engineering 600Hz roofing filter.
6. New AGC system. After narrow filter only, no AGC on front end.

Note 5: Gutted and rebuilt with solid-state double balanced high-level mixers. No vacuum tubes in signal path. Sherwood 600Hz roofing filter.

This page has been viewed 2728 **times since revised on June 28, 2004**

Receivers

[[Home](#)] [[Receiver_Tests](#)] [[FT-1000 Noise Blanker Mod](#)] [[Orion and IC-7800](#)]

Related pages: [Checking signal BW](#)   [FT1000 NB](#) [FT1000MK V NB](#) [External receiver preamps](#)

My receiver Tests

[Sherwood's Tests](#)

We hear many things about R4C's (and other receivers), we love to talk about radios. Unfortunately most information comes from hearsay, useless subjective opinions, or from subjective reviews.

The only commonly published sources of measurements are ARRL or RSGB reviews. While the ARRL and RSGB do a good job of reviewing equipment, they publish somewhat useless wide-spaced data for receivers. Even with excellent wide-spaced performance, close spaced performance can be horrible!

Why Test at Wide Spacing?

Most manufacturers and many magazines test at wide spacing. When we have on-the-air interference problems, it is almost always with a station a few kHz or less away. Why would most of us care about a test or data at 20kHz or wider, when the bothersome signals are a few kHz up or down from us? Wide-spaced tests inflate performance, and gives us meaningless numbers for real-world performance.

Wide-spaced testing only evaluates RF amplifier and first mixer performance. Common design problems are easily and often missed when wide spacing is used. The weakest link is almost always downstream of the first mixer. There are several specific examples I can give:

1.) Receivers using DSP-based filtering systems for primary narrow selectivity.

2.) Receivers with poor 2nd mixer design (the R4C)

3.) Design errors in noise blankers, such as the **Yaesu's noise amplifier** design error

It's no wonder receivers have shown very little performance improvement over the years. Manufacturers evaluate performance on nearly useless wide-spaced measurements. They obviously only want to pass the wider test signal frequency spacing tests, because that is what we look at. Looking at 10kHz or wider tests, we all assume things are getting better. In actual use, most of our problems come from signals nearly on the same frequency, not 10kHz and especially not 50kHz away!

The only truly valid performance test is one where BOTH test signals are within the roofing-filter bandwidth. When close-spaced performance is good, wide-spaced performance is just as good or better. This is true for older radios and modern radios.

We also need to be factual in performance assessments. Too many feelings get in the way of being objective.

Receiver Myths

PIN Diode Mods

Save your money. PIN diode replacement never has changed performance in any receiver I've listened to or measured. PIN diode mods don't change distortion, blocking, noise, or any other parameter. A normal signal diode with proper bias is just as good.

PIN diodes function as "RF switches" or "linear RF resistors" only when the carrier lifetime exceeds the period of an RF cycle by a large margin. Most of the diodes used in PIN mods do not have long enough carrier lifetime to even behave like a low-distortion linear resistance, let along a pure switch. They really aren't any better than manufacturers stock diodes, with the PIN diodes barely being linear at 30MHz let alone at 455kHz. The whole "PIN diode thing" would be laughable if it wasn't costing people money!

If you have even measured a real difference, please e-mail and tell me what radio it was and what the test conditions were. If you are only going by emotion or feeling, don't bother reporting that. If I spent several hours and/or a few hundred dollars changing diodes and could not A-B the change, I'd probably think things got better also.

Drake R4C

The R4C is elevated to a status far above the realities of its actual performance. The R4C, like many DSP radios, has a wide performance variation between close and wide spaced tests. This is

especially true with the early S/N MOSFET mixer R4C's. With a wide-test, all R4C's look "good". That's because a wide test only checks the tube-type RF-amplifier and vacuum tube 1st mixer. Even though the R4C RF amp does better with higher screen voltage and the first mixer suffers from very low injection, the first two stages are still reasonable.

The problem is in the Drake 2nd mixer. The second mixer in the Drake ranges from poor to useless. The MOSFET mixer versions of the R4C are identifiable by looking at the MODE switch. If the MODE switch has CW 1.5, .5, and .25 positions the receiver has a tube (6BE6) 2nd mixer. If it is labeled CW1 and CW2, it has the horrible MOSFET mixer.

A few Drake website claims there is little or no difference between early serial number R4C's using MOSFET 2nd mixers and later R4C's using vacuum tube 2nd mixers. Nothing could be further from the truth. Worse yet, these pages steer people into a wide 6- or 8-kHz roofing filter for weak signal CW work, a very foolish choice as we will see by actual measurements. For serious CW work, the Sherwood Engineering 600-Hz roofing filter is an absolute must! R4C's have far too much filter leakage to be useful on crowded bands using CW with a wide roofing filter.

Roofing Filters

In order to significantly improve close-spaced performance a roofing filter has to attenuate signals on *adjacent* channels. We could have strong and weak signals CW alternating every 500Hz across the dial, assuming our transmitters were cleaned up. In the case of CW a roofing filter, to be useful, needs to be about 500Hz wide or less.

For SSB, we could consider channel width 3kHz. A roofing filer really needs to eliminate the adjacent channel up and/or down, and that means a roofing filter that passes only 3kHz or less.

While the typical wider roofing filters do a good job knocking down problems caused by signals up or down the band several kHz, they don't do anything for signals within the bandwidth of the roofing filter.

Testing Radios

TEST SETUP

My test setup uses two low-noise crystal oscillators. One oscillator is fixed on 1840-kHz, the other oscillator is selectable at 1840.5, 1842, or 1850-kHz (*note*: I now use variable frequency generators). Both oscillators use low-noise CATV transistors, and provide 20dBm output. These oscillators each feed a 1dB per step attenuator, with a total attenuation of 160dB available.

The attenuated outputs are fed to a ["Magic T"](#) combiner. "Magic T's", like every low-loss passive combiner system, are load impedance sensitive. Any mismatch reduces generator port-to-port isolation. In many cases, marginal generator-to-generator isolation can cause IMD in the signal sources. This generator-sourced IMD corrupts readings, changing IMD performance. To reduce combiner mismatch, the output of the magic "T" feeds a small ~3dB attenuator.

Signal level from each individual source can be varied in 1dB steps from +20dBm to -140dBm. +13.5dBm is the maximum level available, after combiner system losses of 6dB are added, for a final receiver signal range of +13.5 to -146.5 dBm. This range is ideal because my own transmitter is typically +15dBm on my closest receive antennas, while my daytime 250Hz bandwidth noise floor is near -140dB. Most of my receivers have sensitivities in the -140dBm range, allowing them to marginally get down to noise floor in the daytime. (Noise increases 10-20dB at night, on quiet nights, because of distant noise sources that propagate via sky-wave.)

Note: *Most setups use a single attenuator after the combiner, I chose not to do that. To reduce generator IMD, I decided to attenuate each signal source with carefully matched attenuators. If a test requires 30dB of attenuation, generator-to-generator isolation will be the sum "magic T" isolation and each attenuator pad isolation. In this case, generator-to-generator isolation would be well over 90dB, far more than I could obtain with the "Magic T" alone.*

The Magic-T combiner also has a 3dB 50-ohm output attenuator. This pad helps stabilize load impedances seen by the "Magic T", insuring return loss is at least 6dB. Total attenuation through the "Magic T", including the internal attenuator pad, is 6.5dB. The port-to-port cross talk of the magic "T" can be nulled with a small trim pot. This adjustment is only necessary when doing tests with near-zero attenuation and a mismatched receiver.

MEASUREMENTS

Since virtually every receiver overloads in stages after the attenuator or pre-amplifier, there probably is no compelling reason to measure receivers with the attenuator on. Adding an attenuator will not increase the dynamic range, it will simply move the raw measurement numbers higher. In other words, we don't care what the absolute numbers are...we can always add or remove gain external to the receiver. What we do care about is *dynamic range* of the system, both for blocking (a strong signal makes the weak signal disappear or get noisy) and intermodulation products. The lower the dynamic range numbers, the worse the receiver will be.

Measurements involve three basic procedures, all measured in dBm (dB milliwatts).

MEASURING MDS

A conventional signal generator is used to measure minimum discernable signal (MDS). This point is where the signal is just clearly audible, about 3dB out of the noise floor.

MEASURING BDR

Blocking Dynamic Range is measured by setting one oscillator (or the signal generator) to a test frequency either 2 or 10 kHz above the interfering signal. This test is equivalent to having a single strong station come on a certain amount away from a very weak station you are trying to copy, and making your receiver lose volume or have a hiss that increases compared to the weak signal's level.

To minimize generator noise, the low-noise crystal oscillators are used for the strong signal. This creates a "perfect" zero bandwidth strong signal with virtually no broadband noise.

The level of the strong signal is adjusted until the slightest detectable change in S/N ratio occurs. The difference between the MDS and the level causing the blocking is the blocking dynamic range. This will be the ratio of the weakest to strongest signal the receiver can handle without losing a noticeable amount of weak-signal sensitivity, if the strong signal is a perfect signal and the weak signal is right at noise-floor.

MEASURING IMDR

Intermodulation dynamic range, or two-tone dynamic range, is measured by running two equal strength signals (from the low-noise oscillators) into the receiver with a certain test spacing. This test is equivalent to having two strong signals very near each other, with just the right spacing to cause a mixing product to fall on top of a noise-floor signal you are trying to copy. When the signal level of the mixing product is just audible above the noise floor, the ratio of the strong signals to the MDS (minimum discernable signal) becomes the IM dynamic range.

Poor IMDR performance shows up as splatter on SSB and as bloops, bleeps, and random musical thumps or phantom signals on CW.

Look at some receiver measurements!

This page has been viewed **1893** since rewritten on June 28, 2004!

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FT-1000 Noise Blanker Mod

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This mod is part of a plug-in package available from me. See the [Keyclick mod](#) page.

I've tested over 15 FT-1000 and 1000D's with this mod. This modification will reduce close-signal IM in every FT1000 or 1000D by a substantial amount!

The IM problem is created because the FT1000 heavily forward-biases the noise-blanker's first FET whenever the blanker is turned **OFF.** This causes the FET hanging on the IF system to have high gain. Strong signals within roofing filter BW saturate the FET's drain, causing mixing products (IM).

These new signals feed directly back into the 8MHz IF section, creating interference to any desired weak signals. The FET's IM creates new "phantom" signals that sound like normal CW signals. On SSB, it creates IM that sounds exactly like splatter!

This modification removes harmful forward FET bias when the noise blanker is off, and prevents the noise-blanker from creating interference to weak signals when the blanker is turned off. The NB functions normally when on.

The Mod

1.) Remove the bottom cover.

2.) Locate the noise blanker on the left rear corner of the IF board near the first group of 8 MHz filters.
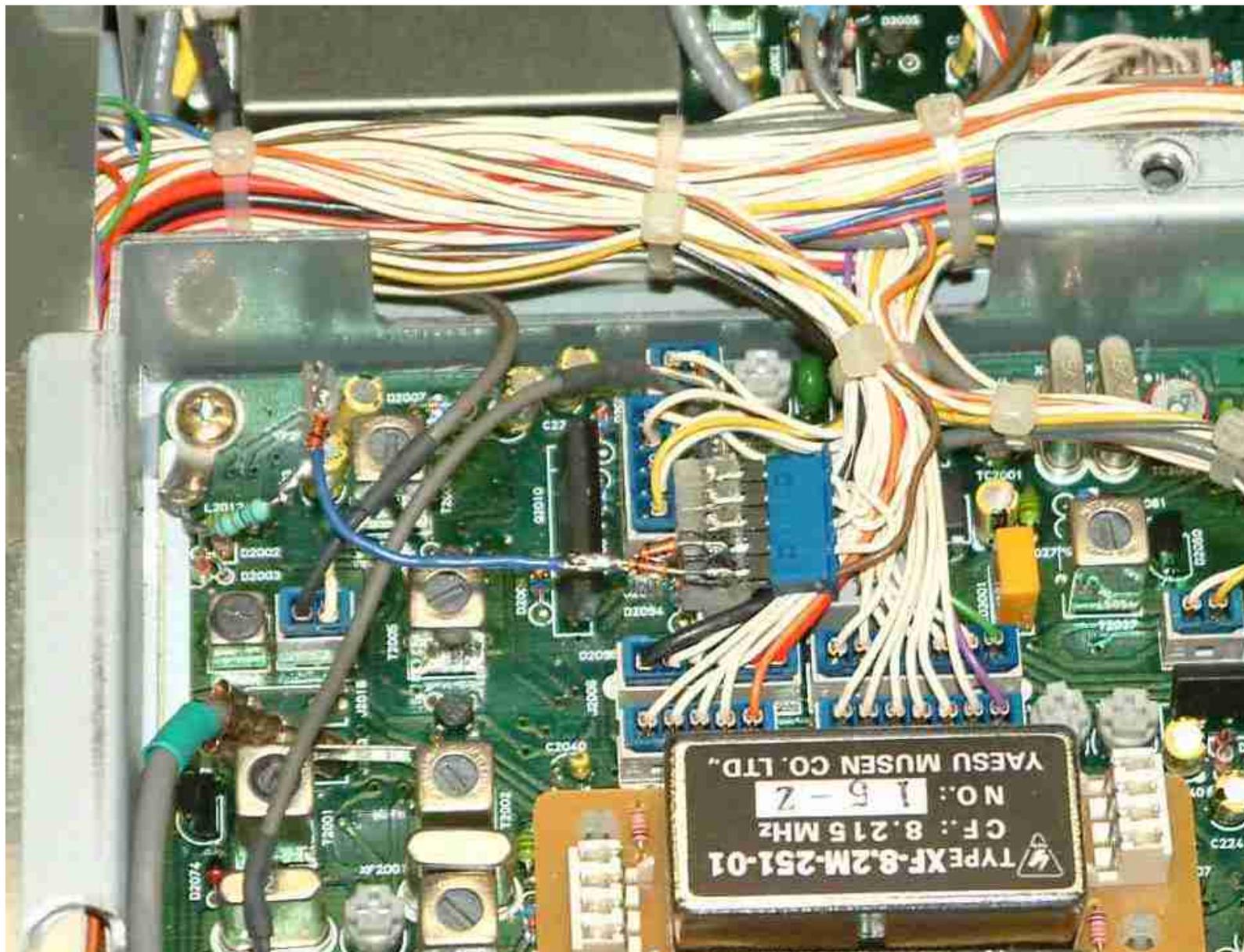
Download Picture

3.) Locate TP-2001.

4.) Connect a small silicon switching diode (1N4148 is one example, although other silicon-type signal or switching diodes will work) through a 10k ohm resistor to ground from TP-2001. The diode cathode (banded end) goes towards the resistor and ground, anode towards TP-2001.

Caution: DO NOT use a high leakage germanium diodes. Use low-current silicon switching or detector diodes.

5.) Construct a pair of diodes with the cathode (banded end) tied together. Leave the leads long for now.

6.) Connect one of this diode pair's anode to pin 1 (the brown wire) of J2001.

7.) Connect the other diode anode to pin 2 of J2001.

8.) Connect the common-cathode of these two diodes to the junction of the 10K resistor and the cathode of the diode installed in step 4.
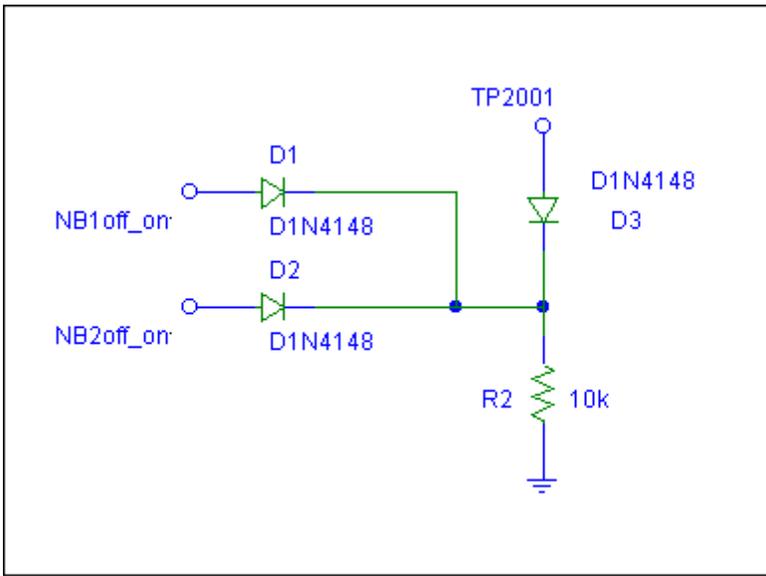
Note: I just pushed the wires down in the pin of the plug (P2001) and tacked them in place with a tiny bit of solder. You can verify these are the correct pins by measuring the voltage when switching the NB and wide NB buttons. Turning either or both NB switches "on" applies about 8 volts to either or both pins. Connect the common to the diode cathode and 10k resistor junction through a short hook-up wire.

This mod removes forward AGC bias from Q2003 and 2004 when the noise blanker is off, and prevents IM distortion fed back via C2041 and foil traces from impacting the receiver when the noise blanker is off. Other than improving IM performance when the blanker is off, it has no effect on NB performance.

Keyclicks_1000D

Revised 1/31/04 at 0130Z

Receiver Improvement for FT1000 and 1000D

The FT-1000(D) runs an essentially unshaped CW waveform into the filters. This harsh signal generates unnecessary key clicks. The clicks are strongest from almost 1 kHz below to 2kHz above the transmit frequency.

The cure:

I now have a plug-and-play 100% tested mod for the FT1000 and 1000D. This mod just plugs in, and includes the receiver NB mod.

Neither mod requires soldering! Only a screwdriver is required. The price for both mods is $45.

View plug in mods.

Locate the seven-pin connector J3024 at the middle-front of the AF board. (The AF board is the one with the audio, PTT, ALC, and key jacks as well as other connectors.)
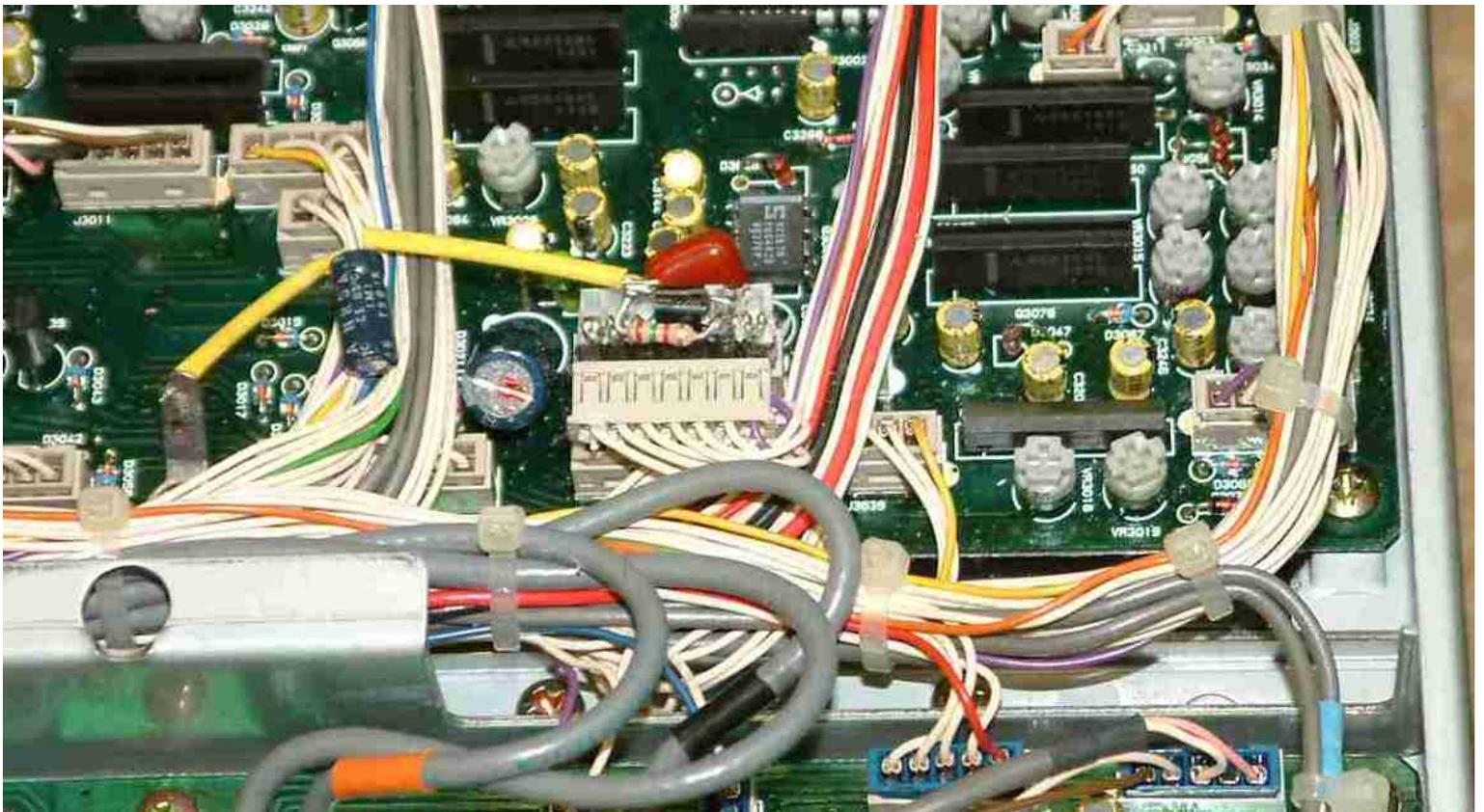
This is a picture with a hand-wired mod in place.

Download picture

Cut the wire going to pin 2 T CNTL). This is the second terminal from the right (front of radio viewing). This is NOT the lead with 9 volts (nine volts is on pin 6). If you are unsure you counted the correct direction, measure the voltage!

Connect the collector of a 2N3904 to the flying harness wire you just cut.

Connect the emitter towards the jack on the wire you just cut. This transistor MUST be a 2N3904.

Connect a 1.5k resistor from collector-to-base on the transistor.

Connect a 10uF, 25-50 volt capacitor from base to ground. A non-polarized capacitor may be better, if you can find one. No one has reported problems with normal electrolytics as long as the positive lead connects to the base of the transistor.
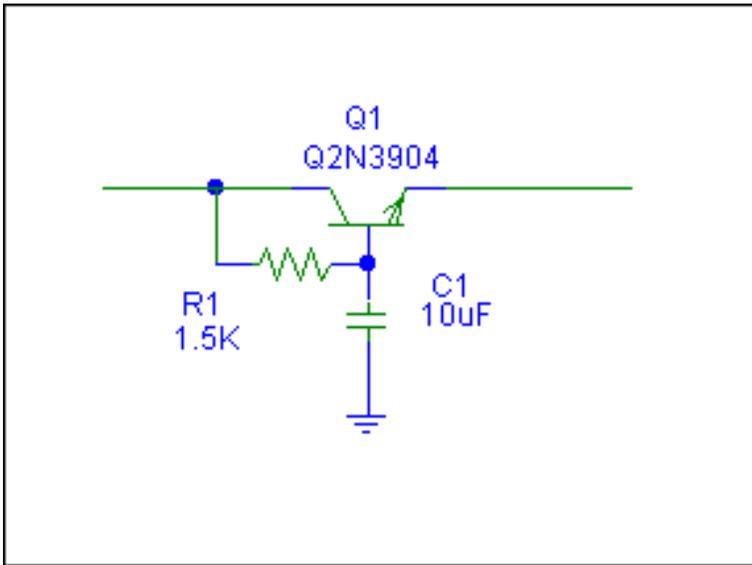
The Results:

This modification slows the waveform rise and fall times at Q3030 and D3009. The result is a normally sloped CW waveshape into the SSB filter.
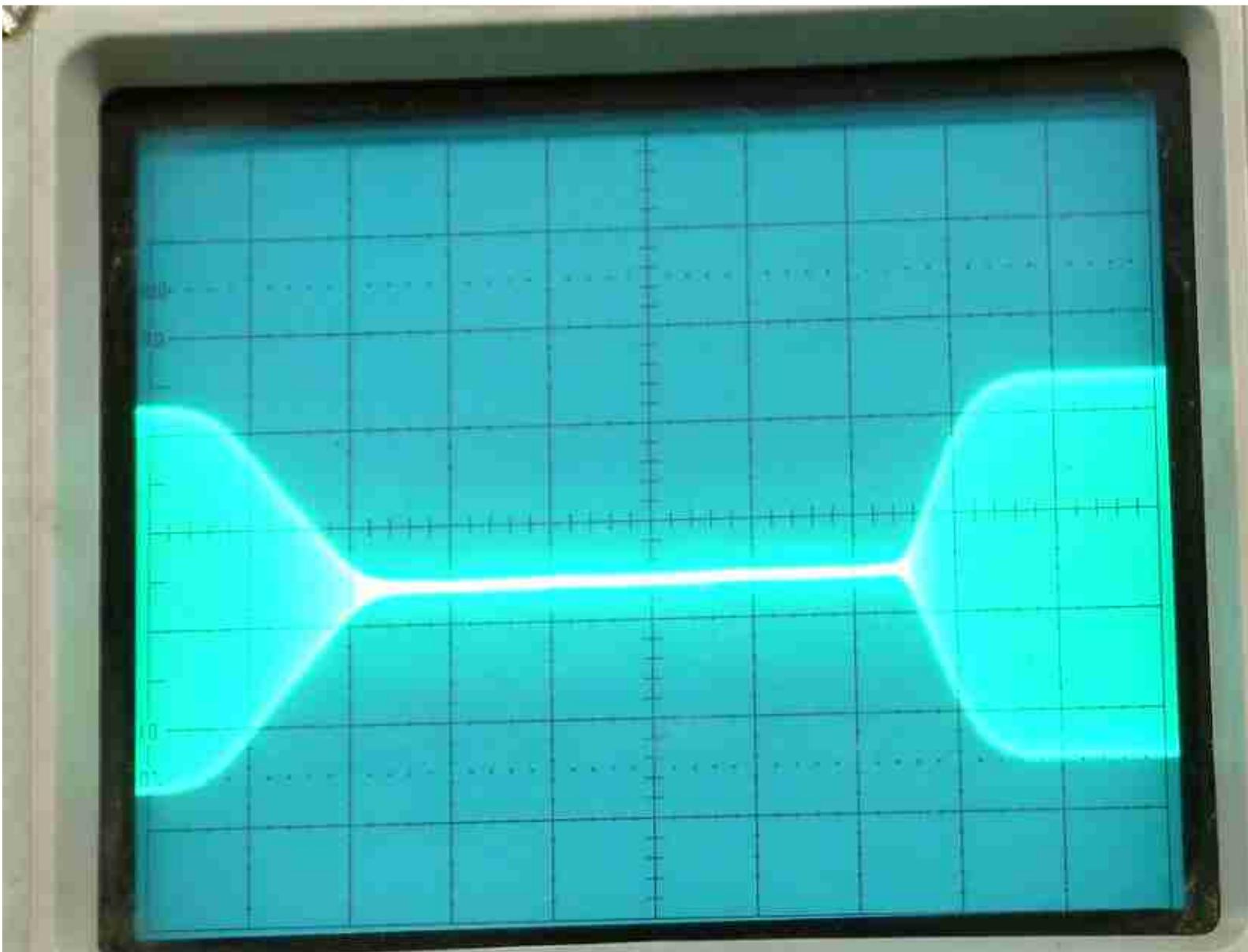
This is a nearly-perfect waveform with about 7mS fall and 4ms rise. The important thing is there are no sharp edges that cause clicks!

Clicks beyond 1 kHz above and below the transmitting frequency move into the composite transmitter noise, and there is about 50 dB reduction in clicks on some radios above the TX frequency!

Clicks 500Hz away are reduced about 20-35dB.

There is almost no change in "keying sound" on the transmitting frequency. The envelope shape is near ideal, and bandwidth reduction is dramatic.

Operating

Warning: This mod is NOT compatible with QSK with the full 10uF in circuit. For QSK use, I suggest adding a SPST switch from the 10uF capacitor to ground. You can place an additional 10uF capacitor across that switch, or simply add an additional 10uF in series with the ground lead from C1 (this gives 5uF total value at the base of Q1) and *not* switch the system. When not operating QSK, I recommend using the full 10uF of C1 for the cleanest signal.

The ideal wave shape occurs when the ALC is at the upper area of the blue ALC line on the FT1000 meter. With full **DRIVE**, adjust for desired power with the **RF PWR** control. After achieving desired power, back the **DRIVE** down until the ALC indication just starts to move away from full blue area (half-scale). Anything from 1/3 to 1/2 scale is a normal waveform.

For lowest clicks, adjust **DRIVE** until ALC until it just registers (perhaps 1/8 to 1/4 scale on ALC).

Keyclicks

[[Home](#)] [[FT1000MK_V](#)] [[Keyclick_MP](#)] [[Keyclicks_1000D](#)] [[What_Causes_Clicks?](#)]

Bandwidth rules [Part 97.307](#)

Link to [W9CF](#). Kevin analyzes key clicks mathematically.

I noticed W2VJN has proposed using absolute signal level at a fixed test spacing for click measurements. While I agree with George's suggestion ([this link](#)) we use an established test method, I strongly disagree that a 15Hz BW signal level measurement of the peaks accumulated in dozens or hundreds of scans is any more useful that listening off-frequency on a known good receiver.

While his mod is certainly better than a stock Yaesu, clicks also can be reduced more than his method provides without affecting ability to work weak signals. We don't need to watch output power and keep duty cycle to 50%, a slight adjustment of weight control would do the same thing.

The real point is that both George and my suggestions are patches, they are not cures. Yaesu should step up to the plate and be responsible, and set an example for others to follow.

Testing

We really need to establish a standard for measurement of transmitters on both SSB and CW, and it should be the effective spurious power level on close adjacent frequencies using typical CW bandwidth. Looking at one spot with slow sweep and peak storage on a spectrum analyzer is not a good test.

It's pretty easy to see using the level (at one specific frequency) of a long-time average of signal peaks with a narrow filter sweeping by (which is what spectrum analyzers do) is a bad idea.

Some radios have a sharp click on one edge, and nothing on the other. This causes peak energy to be higher in proportion to average energy. Radios can have higher peak level and actually do less damage to adjacent channels than a lower level click that hammers on both make and break, like the FT1000MK V does.

If I look on a peak sample and storage device like my spectrum analyzer, a radio could actually

look worse yet bother adjacent CW bandwidth channels LESS than a radio that appears to be better.

A very narrow filter swept over the frequency with average storage of peaks does not account for attenuation slope modulation sidebands. I'd rather have a radio next to me with a steep drop in clicks at 500Hz than one with a single-pole gradually shaped slope even if the gradual sloped radio had less level. A peak comparison at some specified spacing misses too much.

SSB vs. CW

CW shares the same testing problems as FM and SSB. The FCC has changed some commercial voice tests to focus on spurious power level on adjacent channels, rather than using conventional but somewhat useless two-tone tests. A two-tone test shows the very *best* a radio is likely to do, not the *typical performance*. The FCC now requires normal modulation and an adjacent channel *power level measurement* in some cases.

CW and SSB testing for amateur use should be the same way.

I think is is shameful that manufacturers can't manage to spend a few hours of engineering time to select different component values, instead of sticking us with radios that have poor keying and IM characteristics. Frankly, there is NO excuse or justification for selecting wrong component values in a such a simple design area.

Is an excessively clicking radio legal?

Most of us agree right off the bat that some radios are annoying, and MOST people actually want to have the cleanest possible signal. Of course there are a few self-centered people who care less if they bother other people. They generally use the excuse that since a radio was sold it must be legal.

Here is a link that shows the exact 97.307 FCC text.

What do bad and good signals sound like?

For now, you can listen to some recordings and judge the difference.

Key clicks are off-frequency sidebands heard when there is no trace of a tone. If you listen carefully to these recordings, you will see the "hardness" of a properly filtered transmitter on frequency sounds very little different than the hardness of a wide, clicking signal.

Here are some sound files....these files are intended to be educational. While it isn't OUR fault the rigs were incorrectly designed, it is our legal obligation under 97.307 to correct the problems we have been stuck with. This is especially true when big antennas and amplifiers are used.

First, this is a mono recording of a virtually click-less radio. This station is 20dB-over-nine, and my noise is S2. When you listen, notice the "clicks" disappear right when any trace of tone vanishes. You'll notice the CW still sounds hard, and this station regularly broadcasts high-speed CW that hundreds of people copy. This is the *Click-free signal* of W1AW on 160 meters.

If you own a stock FT1000, FT1000D, FT1000MP or FT1000MP MKV (or any rig with a rise and fall faster than 4 or 5mS OR with any sharply rounded corners) you have work to do!

Notice key clicks can be heard long after the tone disappears totally, as we tune across the signal, yet there isn't any trace of receiver desense. S-9 plus 10dB station from England, 40-meters, using *FT1000MP MKV* tuning up 2 and down 2 using 500Hz RX filters. Significant clicks in the USA from England well after sunrise in England when signals are dropping. Imagine a 40 over MKV! This appears to be typical of all MK V's

S-9 +20dB in S-2 noise tuning from 1.5kHz below stock *FT1000D* (sorry, I snipped the above freq tuning to avoid call sign) This is typical of all FT1000(D)'s

40-meter signal S-9 plus 15dB, background noise S-2, 500Hz filters, tuning up and down 2, recording of *FT1000MK V*. This rig cuts a 3kHz wide swath that is S-6 or stronger.

40-meter signal S-9 plus 25dB, S-1.5 noise. Tuning up 2 and down 2, 500Hz filters in RX. *Omni-6* (This might or might not be typical of all Omni-6's. Ten-Tec reports an internal adjustment, when incorrectly set, causes these clicks. I've logged several Omni-6's with this problem, and some users continue to receive complaints after trying suggested mods. The characteristic I seem notice is key clicks extend downwards much more than upwards in frequency.)

160-meter, S-9 plus 15 dB signal, S-1 noise, tuning up and down 2, Inrad-mod in *FT1000MP*

20-meter *Kachina*. S-1 noise, signal S9 plus 10. Which signal is the "real signal"? A snip of a clean signal answering him, about the same signal level, is at the end. Spurs went down 2.4 and up 2.4, and were spaced 600Hz apart. (This might not be typical of all Kachina's)

40-meter signal, S-9 plus 10dB signal S-2 noise, tuning up and down 2kHz, 500Hz filters, stock *FT1000MP* (this is typical of all stock FT1000MP's. If you own one that has not been correctly modified, you can be sure it has grossly excessive key clicks)

160-meter signal, S-9 plus 20dB S-1 noise, tandem 250Hz filters in very high-dynamic range CW-only receiver, tuning up and down 1.2 kHz, stock *IC765*

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FT1000MK_V

[**Home**] [Up]

Preliminary tests show very strong keyclicks +1kHz and -1kHz. The LSB clicks peak 500Hz below the TX frequency, and the USB clicks (in the LSB mode) are very obnoxious because they are at twice the rate as LSB clicks (strong on both make and break).

The FT1000MK V shows a 1mS rise and a 2ms fall with sharp edges and a very poor rise and fall slope.

Keyclick and Noise Blanker Mod for FT1000MP MK V

It's great to see a manufacturer offer improved close-spaced SSB transmit performance. The MK V Yaesu reverses other modern radio's downward spiral of transmitter SSB IM performance. Yaesu included a class-A mode.

Unfortunately close-spaced receiver performance has been neglected, and this affects all modes. The transmitter also has nasty CW clicks. Stock MK V's appear to be as bad as or worse than older 1000 series radios in two important areas:

- Clicks appear to be more severe than those from the notoriously poor 1000MP's
- The noise blanker continues with the same circuit error as the FT1000/1000D/MP series

Since a portion of the keyclick mod requires accessing the same general area as the noise blanker, it would be prudent to fix the receiver and transmitter at the same time. This article offers a combined modification that patches both problems.

Keyclick Problem

Raised- sine rises and falls would provide the fastest possible CW speeds for a given bandwidth. With properly filtered rise and falls, we would hear little or no change or softness when listening on-frequency. Tuning off- frequency, clicks would quickly vanish.

If you want to hear the sound of proper shaping, listen to this click-free signal recording as I tune across the signal. Off- frequency (even a few hundred hertz), we hear no clicks at all. On- frequency the CW is "hard" sounding, allowing copy to 60-WPM or more. When the tone disappears in the deeper receive filter's skirts, clicks also disappear.

In contrast, compare the [MK V recording](#) as I tune past the signal. This signal is from Europe on 40 meters!

There is a day-and-night difference off frequency between the no-click and loud-click signals. On-frequency both signals sound the same.

MK V clicks are caused by excessively fast rise and fall, and very poor shape of the rise and fall.

Rumors sine-shaped waveforms impact tone or readability of signals are false, as are claims clicks rolling of at some "X-dB-per-octave rate" beyond a few hundred Hz are a necessary part of life. Such statements are misleading, likely being based on the incorrect assumption the receiver has very wide bandwidth and the transmitter is filtered through a single stage simple resistor-capacitor click filter.

Unfortunately when we patch poor CW transmitter designs, we can not make perfect corrections. Without major modification we can not modulate the MK V (or most other transceivers) with properly filtered (which also means perfectly shaped) rise and falls.

This modification, like the MP and 1000D click mods, is a patch...not a cure.

Patches Vs Cures

Because mods on existing radios are patches, the radio owner must make a choice. If the user operates speeds faster than 45 or 50 WPM, the rise and fall required for legal close- spaced operation may be less than ideal. This does not mean the ability to work weak signals at modest speeds (up to 30 WPM) would be compromised even the slightest amount. It means high- speed ops (speeds over 45-50 WPM) may find the CW slightly mushy when adequate for close-frequency operation.

If a 2 or 3 millisecond rise and fall is used for operation at very high CW speeds in a single-pole R/C filter (this radio uses a simple RC filter) a transmitter is almost guaranteed to interfere with less strong signals within 1kHz or so. Transmitters with fast rise and fall times should stay at least 1.5kHz away from operators working weak signals, especially when the CW transmitter has 1930-era transmitter CW shaping.

Operators with stock MP MK V's should always try to operate at least 4kHz away from weak or moderate signal- level stations. Part 97 rules prohibiting keyclick emissions that interfere with adjacent frequency operations. The specific rule is 97.307(b) "Emissions outside the necessary bandwidth must not cause splatter or keyclick interference to operations on adjacent frequencies."

The Click Mod

The actual click mod requires changing two stages. The first stage modified is on the IF board. The IF mod slows the rise and fall of mixer transistors Q2033 and Q2038.

Note: This stage is easy to modify, and is located on the same board as the noise blanker. This allows the noise blanker to be corrected at the same time.

By itself, modification of the easy-to-reach IF board is NOT effective for substantially reducing clicks. A later stage on the RF board also has truncated rise-and-fall times with a very poor R/C edge-shaping system. This later stage continues to add clicks even after earlier stages are modified.

RF amplifier stage Q1001 has the fastest rise and fall in the RF section. Q1001 must have proper gate bias shaping and timing to reduce clicks to acceptable levels. Removing D1002 and altering components around Q1002 slightly reduced clicking, but I concluded any effort wasn't worth the result with bias rise and fall more rounded at Q1002.

As designed, Yaesu uses a square wave very rich in harmonics to drive a simple R/C filter. This poorly filtered square wave amplitude modulates the RF and IF sections. The poor basic filtering design, combined with non-linear amplitude response, requires great care in component selection. It also means we never will achieve the optimum bandwidth for any give rise and fall time and ultimate CW speed.

The NB Problem

Signals inside the wide roofing filter BW of the MK V reach the gate of Q2009 through C2043. This point precedes the narrow 8MHz IF filtering, allowing a rather wide swath of unwanted signals to reach the gate of Q2009. Q2009 is left operating even when the noise blanker is turned off, and can have substantial gain depending on bias voltages at TP2001. Bias voltages at TP2001 can be varied by changing menu settings for NB gain, but never fully turn off Q2009.

The accumulated level of all signals reaching the gate of Q2009 produce a large net voltage at the drain of Q2009. This voltage (and resulting net current) causes overload and distortion in the non-linear characteristics of Q2009 and 2010 (my manual has very poor printing, but I believe it says 2010).

New signals created by this distortion and the resulting mixing products are fed back through C2043 to the IF strip. The IM products appear as "phantom splatter" on SSB and " phantom signals" on CW. We can not actually hear the distortion on frequency of strong signals. Instead the IM products appear in the form of artificial interference when we attempt to copy weaker signals within 10kHz of a mixture of signals containing a few moderately strong signals.

The NB Patch

The NB mod is a simple effective mod, and improves close-spaced IM3 dynamic range around 10dB. A simple factory change of moving one foil trace would have made the MK V receiver noticeably better, but fortunately this mod is fairly easy for owners.

The NB correction removes the surface mounted 220-ohm resistor (R2046) from the source of Q2009, replacing it with a 220 ohm resistor connected between Q2009's source at C2027 and Q2016's (2SC4047) collector and the junction of R2049 (220-ohm also).

Making the Mod

This mod is a little more complex than the FT1000MP mod because the chassis of the MK V is a little more complex and unfriendly. Like any service work, having a clean open bench and a spot to separately store screws and other hardware removed in every step in order will make the job smooth and easy. (If someone sends step-by-step text, I'll put it on this page.)

Some may wish to remove and change parts, but I prefer to wire the click-mod to a single terminal strip. This will allow you to customize the mod, switch the mod in and out, or correct any errors without dismantling the entire radio.

You'll need the following parts:

(1) one foot each of two small insulated wires, #20-#26 one (preferably) green and one blue to make connections

(1) four-lug (with ground) terminal strip

(3) .1uF 50 volt disc capacitors (C1-C3)

(1) 22k 1/4w fixed resistor (R2)

(1) 680k 1/4w fixed resistor (R1)

(1) 220-ohm 1/4w fixed resistor

In addition you need a well-lit bench, along with some hand-tools such as soldering pencil and solder, screwdrivers, and cutters and strippers.

Populate the terminal strip as follows:
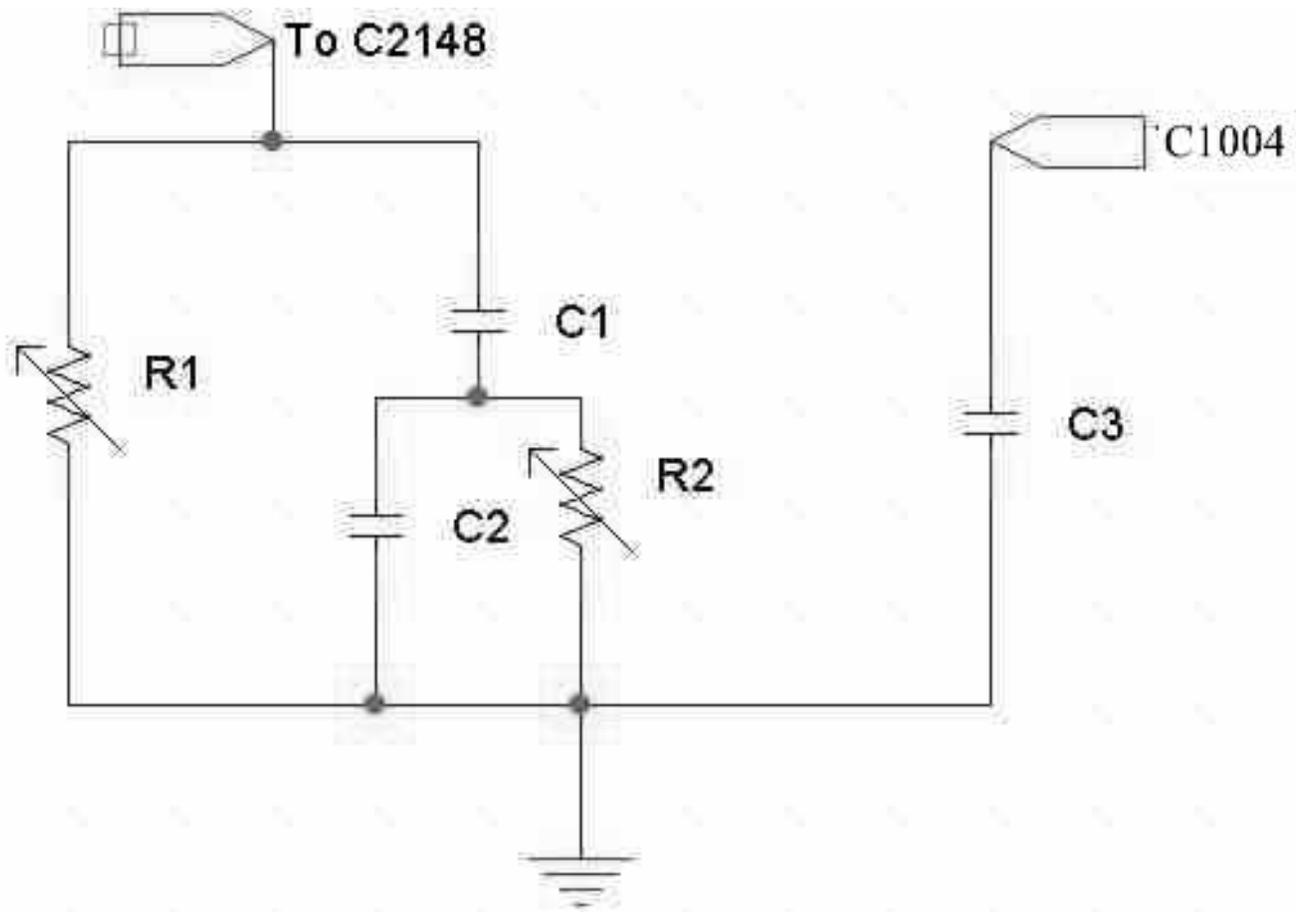
Remove top and bottom covers to gain access to internal circuitry

Remove the screws holding the IF board in place, and the minimal amount of plugs to allow flipping the IF board over. Draw a roadmap of all plugs that must be disconnected. This will help you remember where everything goes!

(click here to download expanded IF board if needed)
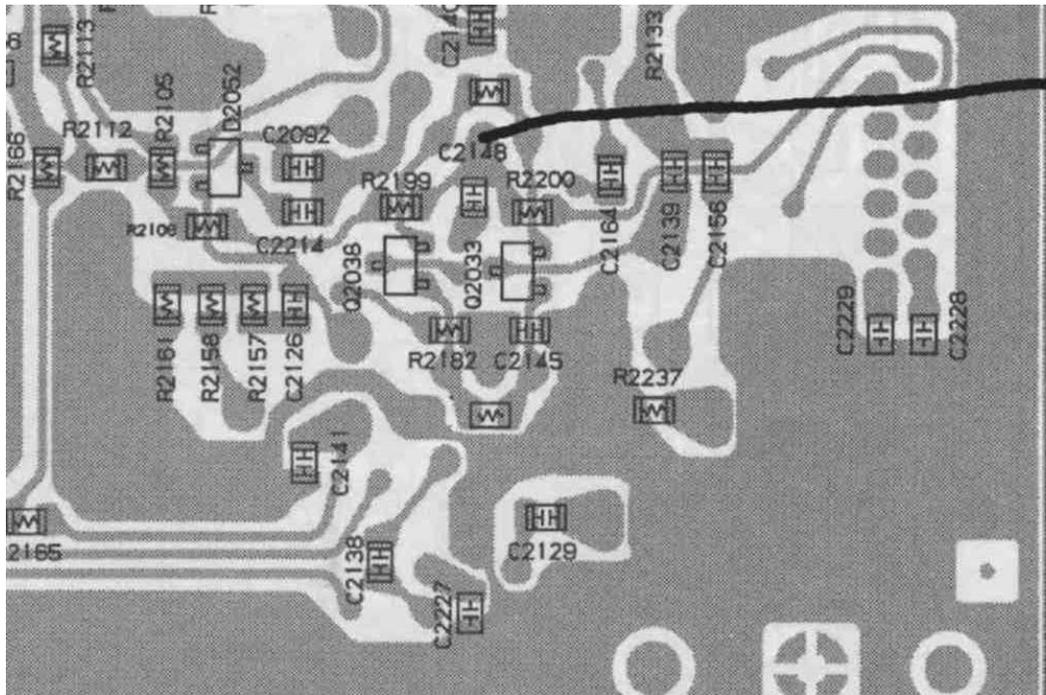
Add wire for click mod

Attach one end of the blue wire to the ungrounded end of C2148, let the other end float
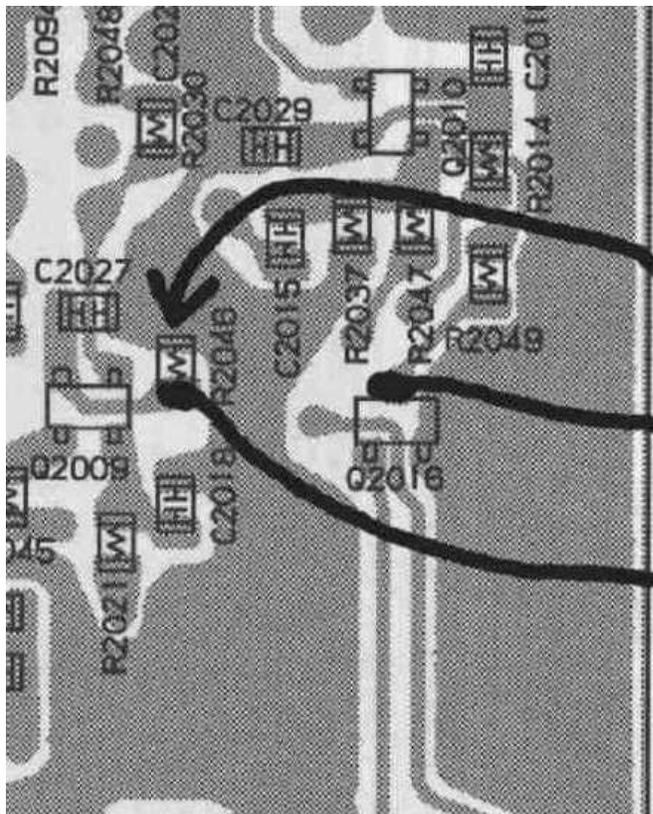
Remove R2046
add 220 ohm resistor
between these points

Locate and remove R2046.

Add the 220-ohm resistor to the Q2009/C2027 source and capacitor connection point

Connect the other lead of this resistor to the junction of Q2016's collector and R2048.

Reinstall the IF board with the flying lead exiting the closest edge of the PC board.

(Take care to avoid pinching any wires.)

Remove all hardware necessary to access the RF board

Remove the RF board

[(Click here to download an expanded view of RF board if needed)](#)
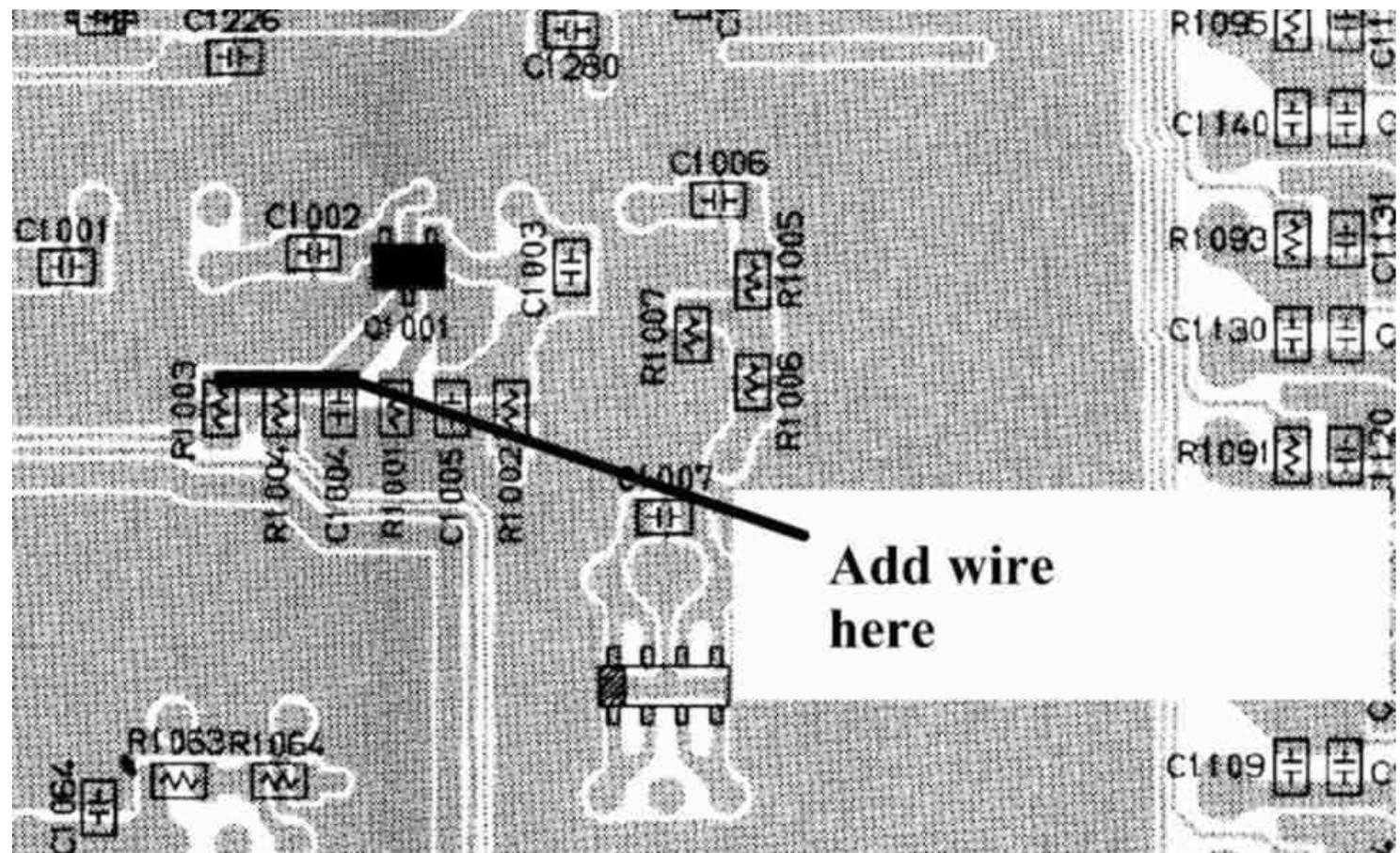
Attach the green wire to the junction of C1004 and R1003/R1004 at the gate of Q1001

Route this wire up to the area of the IF board

Reinstall the RF board and all other hardware taking care to not pinch any wires, and to reconnect all unplugged wires in proper locations

Select a clear area on the IF board and mount the terminal strip under the screws

Connect the green wire (from RF board) to C3, and the blue wire from IF board to the junction of C1 and R1. R1 is a 680k resistor, and R2 is a 22k resistor

Test the radio and reinstall the covers

Remove R2046
add 220 ohm resistor
between these points

Add wire for click mod

Add wire here
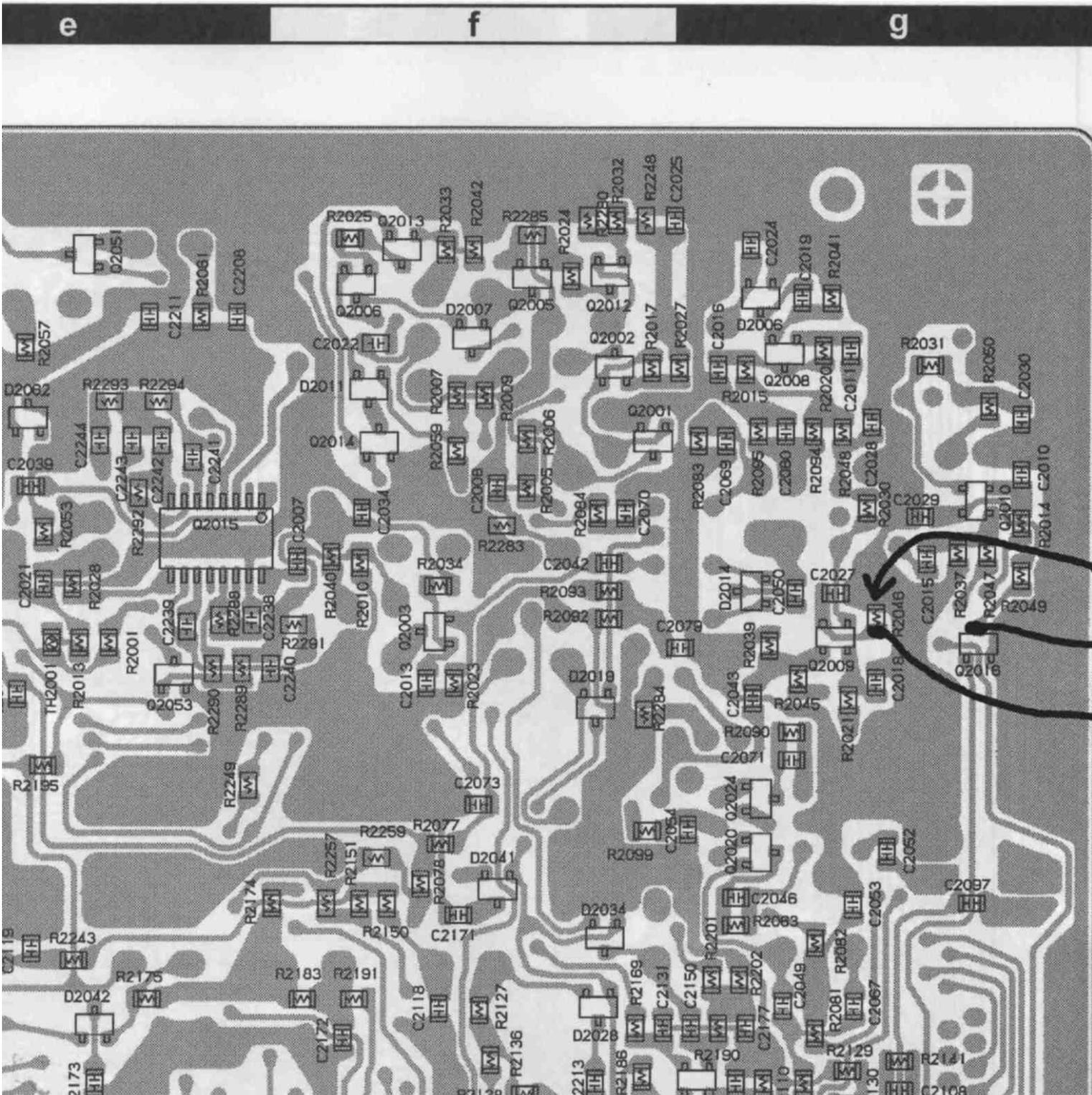
Keyclick_MP

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Curing FT 1000 MP CLICKS

FT-1000MP Measurements

I measured two sample FT-1000MP's (an early and a late model) by operating them into a high power fixed 30 dB attenuator pad. The output of the 30 dB pad was connected through a 3-way splitter to a step attenuator and conventional receiver, a spectrum analyzer, and an oscilloscope. The receiver used a 300 Hz eight-pole filter, the spectrum analyzer used a 50 Hz filter, and the scope was triggered from an external keying signal. Power was measured on a conventional Bird average reading meter.

At 1kHz spacing clicks from the stock FT1000MP's were about 15 dB worse than clicks from my old test bench radio (a well-worn ICOM IC-751A) and more than 20dB worse than the clicks from my click-reduced FT-1000D!

Here is a spectrum display of my stock IC-751A using 30 Hz analyzer bandwidth and ten second sweep:
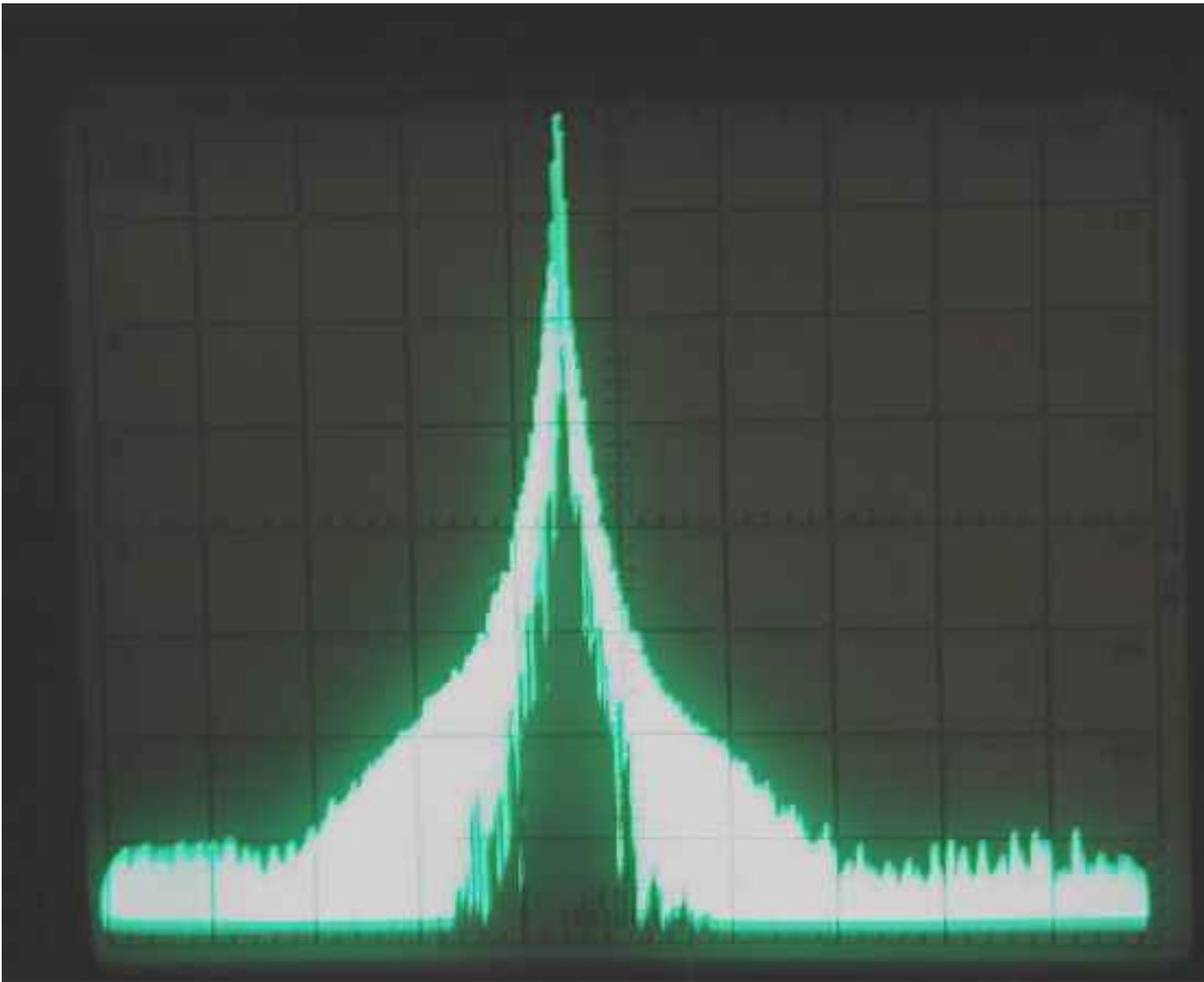
The 751A is approximately -58 dB at 1 kHz, and rolls off smoothly.

Here is the nicely sloped but too fast rise:
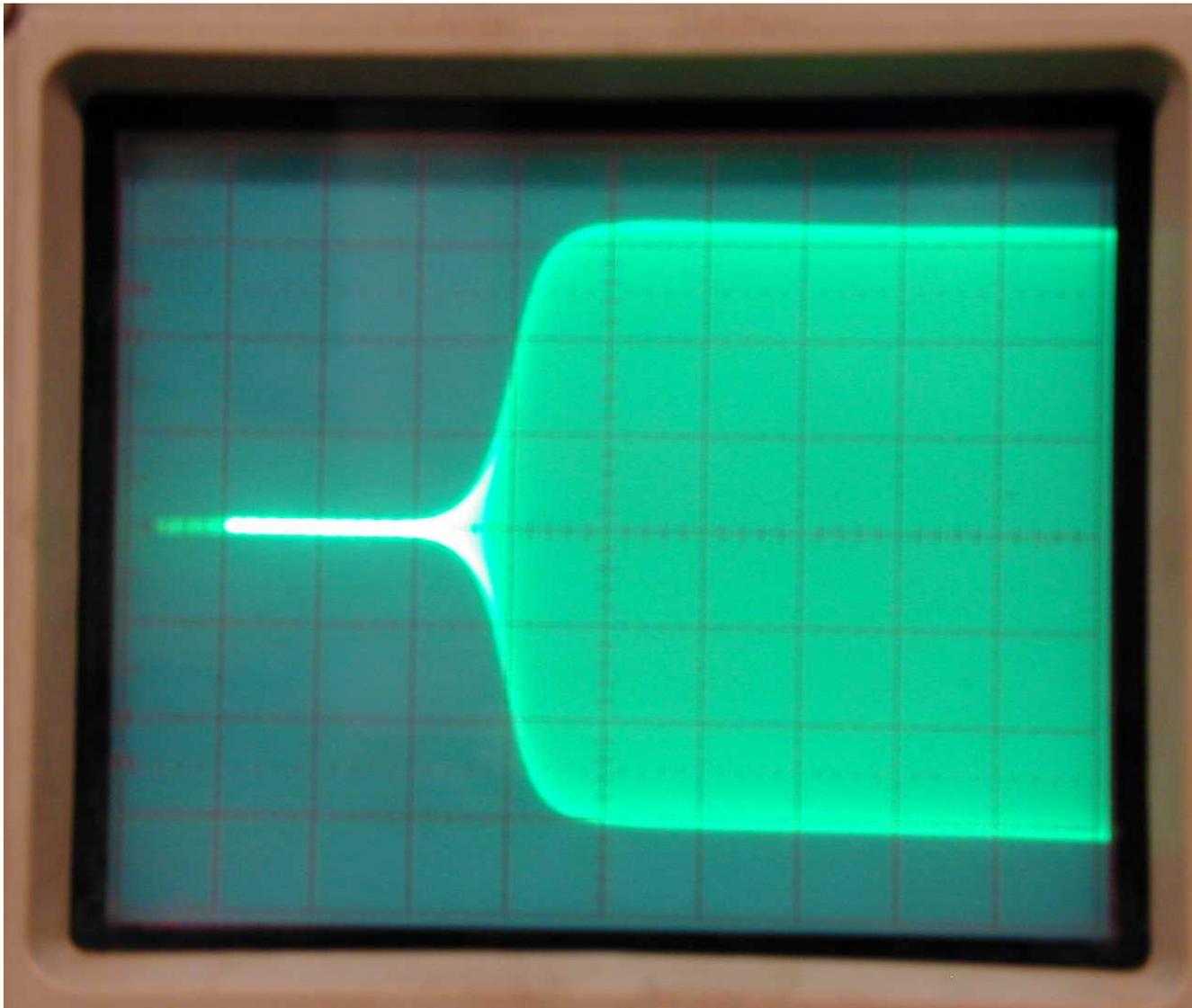
Rise approximately 3 mS....and the slightly sharper at the corner fall:
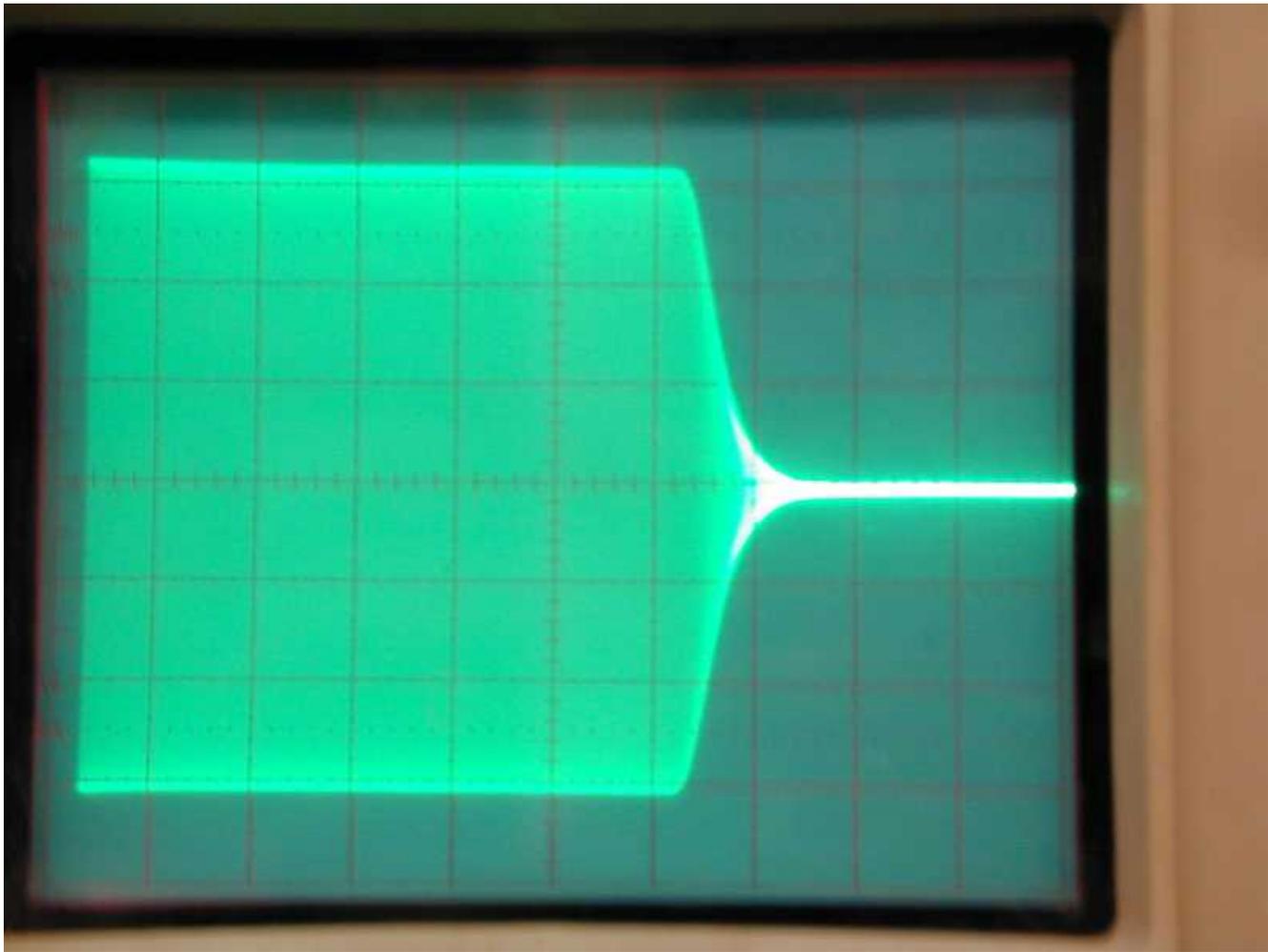
Now the stock 1000MP spectrum:
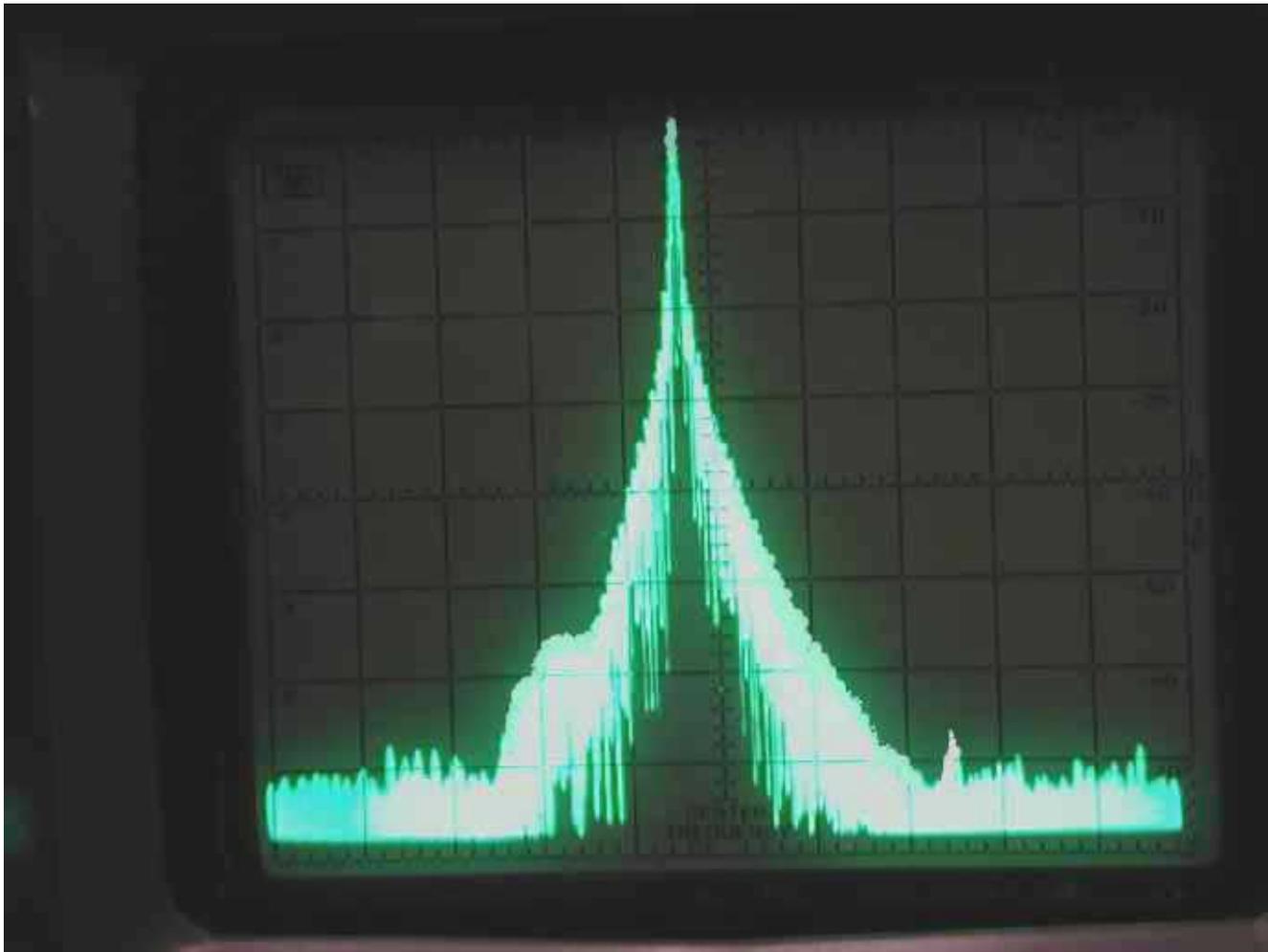
The FT1000-MP is approximately -50dB at 1kHz. It is 8 dB worse than the already "hard" 751A, and has a plateau below the carrier frequency that hovers around -55dB.

In direct comparison, here is a "de-clicked" FT-1000MP:
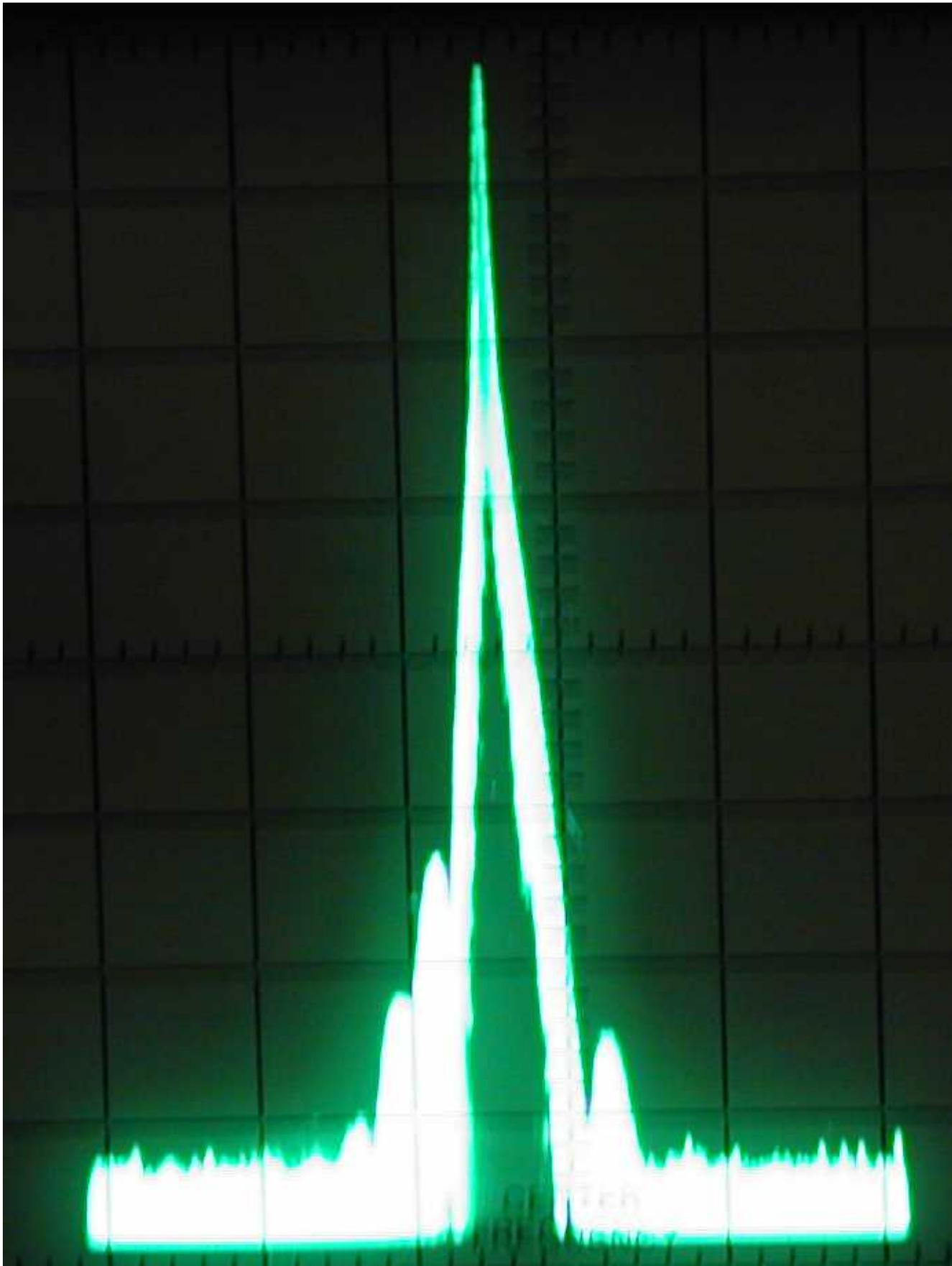
The modified FT-1000MP rivals any of the better radios I have tested, including my "de-clicked" FT-1000(D).

The modified FT-1000MP is around -85 dB at 1 kHz, over 30 dB improvement from the stock MP! Rise time is close to ARRL standards of 5mS, while fall time is around 3 mS. FT1000MP modified rise:
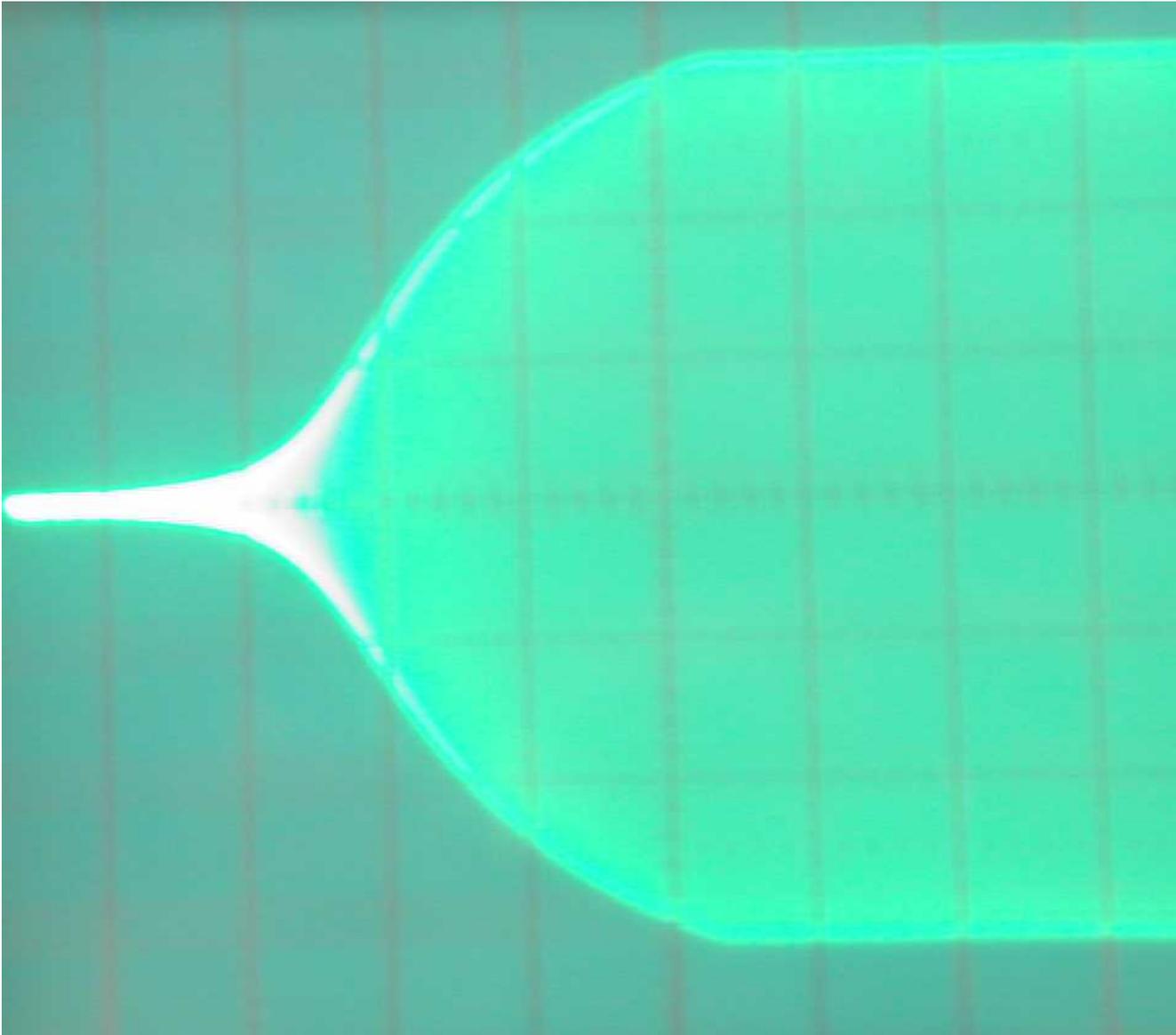
Rise 6 mS. The upper edge is a little sharp, but why worry....clicks are reduced 30dB or more!

Modified MP fall:
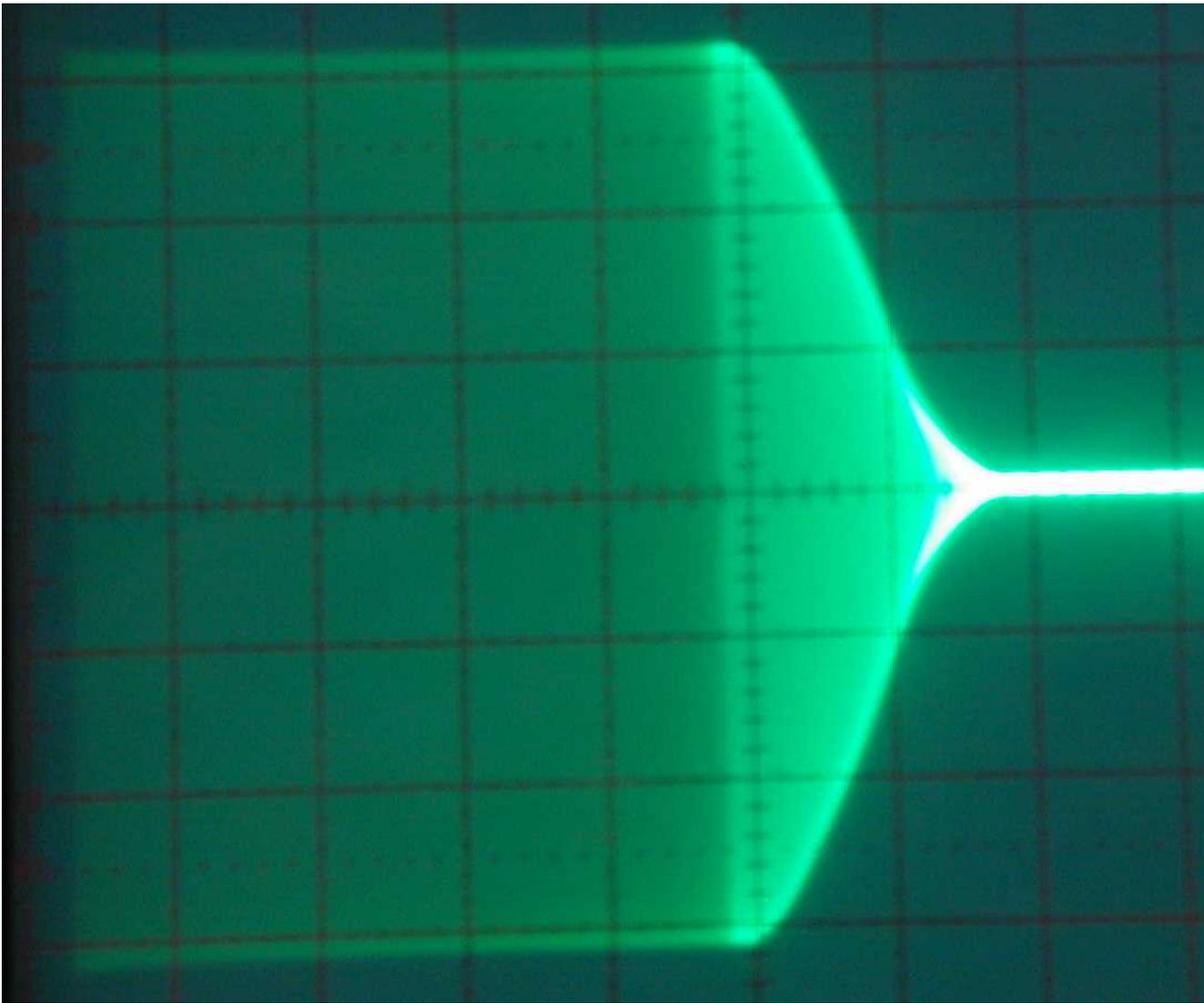
Fall time is around 3mS. While it has much more rounded edges, the slope is still not very "round". Unfortunately we are limited by what is possible to do, and this mod is already difficult enough for laymen.

Some concerned was expressed over the "power" of dots when using long rise and fall times. One simple solution is to turn up the weight control slightly. Keep in mind, even without ANY external weight adjustment, the change in average dot power at 45 WPM was only a few percent! On the air tests with VK3ZL and ZL3REX on 160 meters with fading signals and noise, revealed both could tell absolutely no difference between having the click filter in-line and out-of-line at 40 WPM CW speed. This waveform meets FCC and CCIR specifications for 60-WPM CW modest strength signals, and 35 WPM weak fading signals.

The best method of nulling clicks is by listening on another receiver with a narrow filter. Make sure you are well below overload on the receiver, and set that receiver so the carrier from the MP is just outside the passband of the test radio filter. It is almost impossible to use any other method for proper adjustment, including watching the envelope on an oscilloscope.

 Work in a clear uncluttered location, I like to work with the radio on a clean small carpet on a well-lit bench, and have a container for all the hardware I remove.

 Here's how to make connections to necessary points:

1.) Remove the top and bottom covers of the radio and set them aside.

2.) Invert the radio, so you have the heatsink exposed.
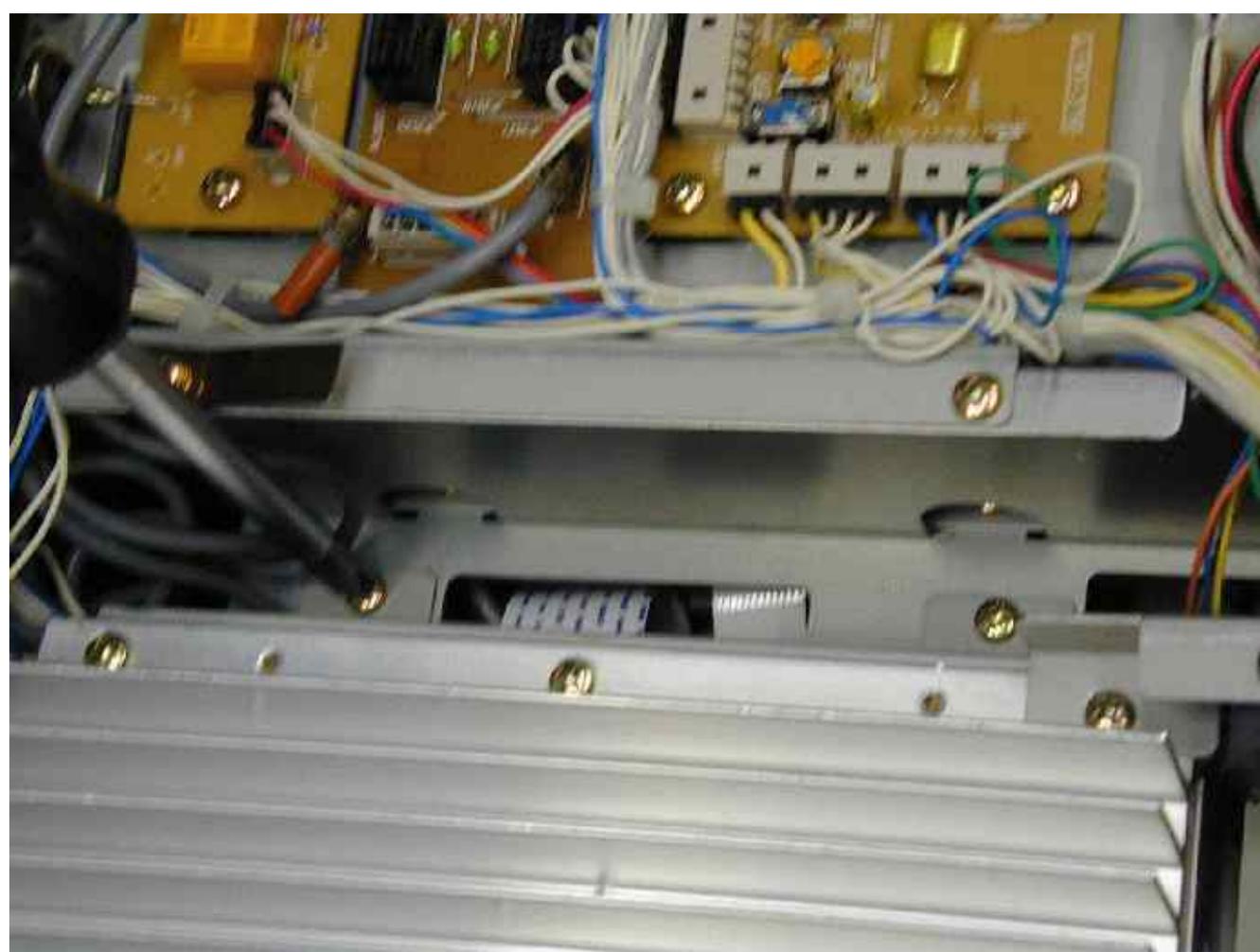
Four main screws hold the heatsink mounting bracket. Two are shown above (one under the screwdriver and one to the right of it). Two other screws are on the side of the radio chassis. You might want to remove the long screws holding the fan bracket, although I got by without doing so.

3.) Lay the fan and PA assembly out of the way, you may have to open some of the wire harness clips to get more wire. Unplug the fan so it is totally free from the unit. It should look like this now:

4.) The RF board is the green-colored board you see above. There are several screws holding that board down, and two screws on a rear panel DIN connector that is mounted on that board. The board will freely move when you remove ALL the screws. Do NOT pry or force the board out, if you have to pry you missed a screw!

5.) Flip the board over, you might have to unplug a wire harness or two...but I managed to work without doing that. The board should look like this:
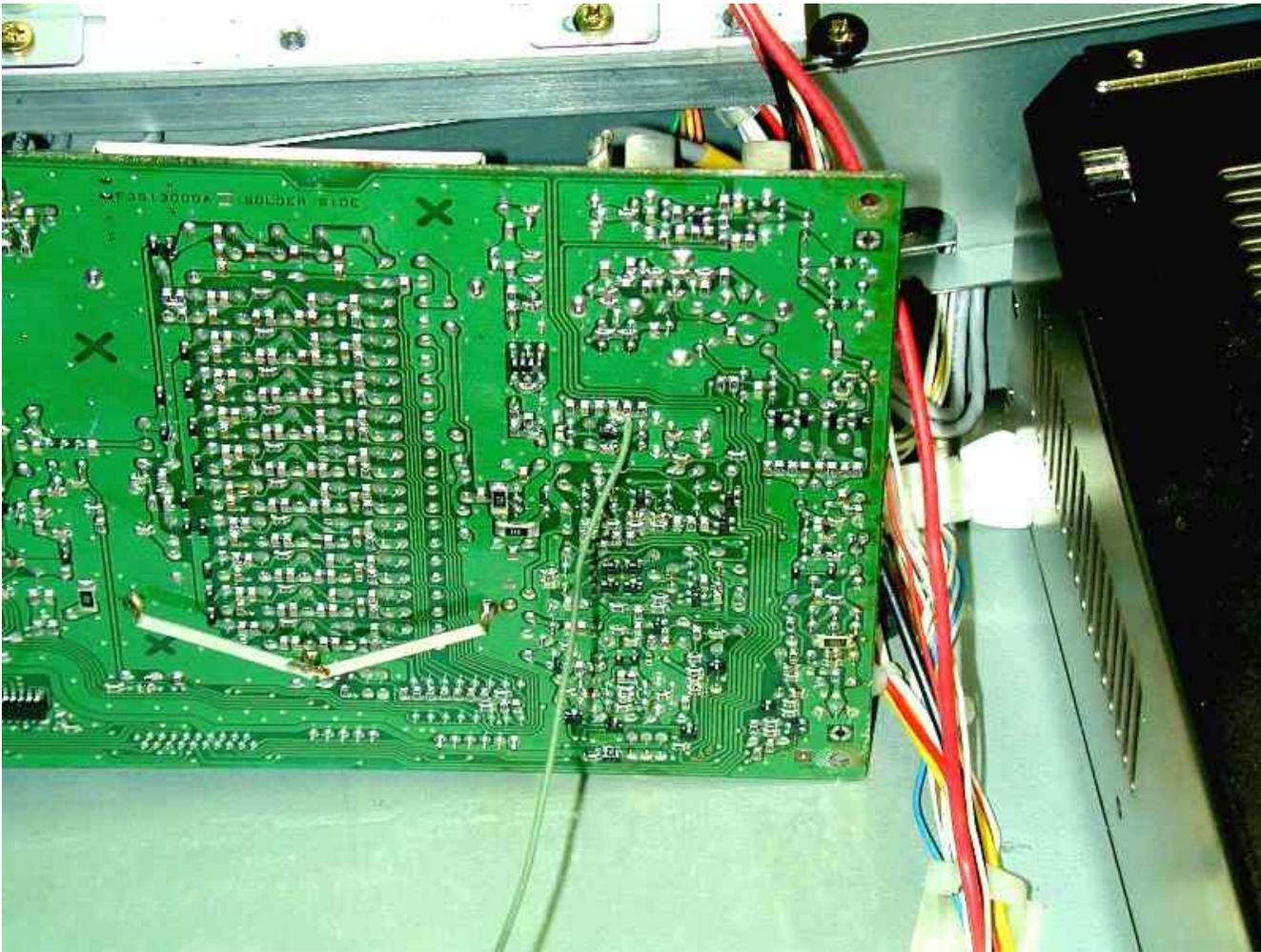
 6.) I added the green wire you see above. I tacked it on a foil pad by laying the wire across the
point where two chip components soldered in, as shown below:

You can see the little black FET (Q1034) and the slightly triangle-shaped foil trace that connects to Q1034. I bent an "L" in the small wire I used. The wire will route topside to a terminal strip, so it needs to be several inches long.

7.) The new wire routes under the RF board to an oval slot in the chassis, where if feeds to the other side and emerges near the IF section unit. Be careful not to pinch any wires when remounting the circuit board. The opening on the right is best for getting the added wire topside:

8.) While re-installing the RF section, fan, and heatsink inspect the wiring carefully. Be sure nothing is touching moving parts of the fan, and be sure no wires are pinched or left unplugged. It might be advisable to check the radio quickly on a dummy load to be sure the transmitter section works properly.

9.) Flip the radio over, and remove the two ribbon cables connecting to the IF section:

10.) Remove the mounting screws and flip the IF board exposing the bottom:
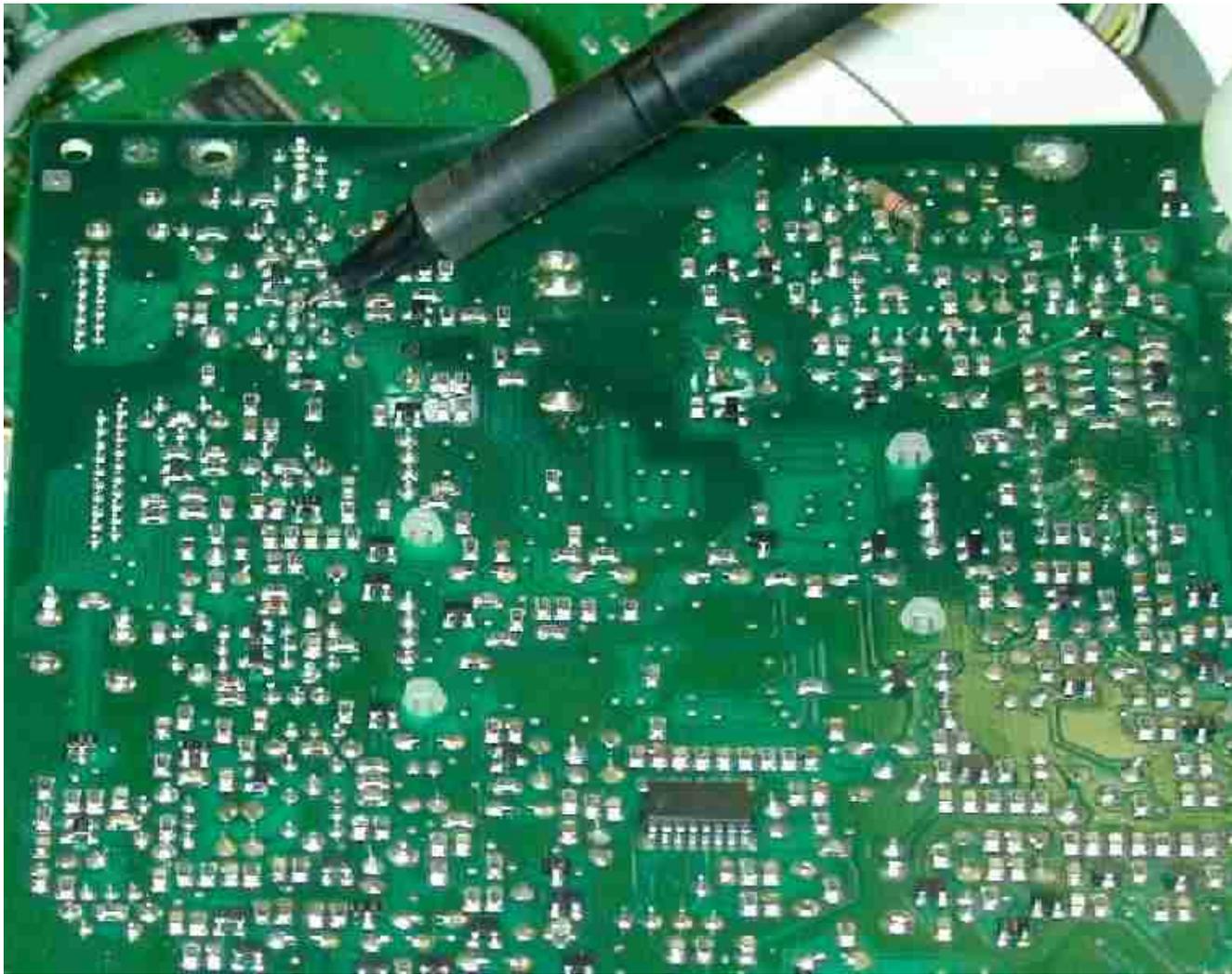
This is the area where the wire attaches to the IF board:
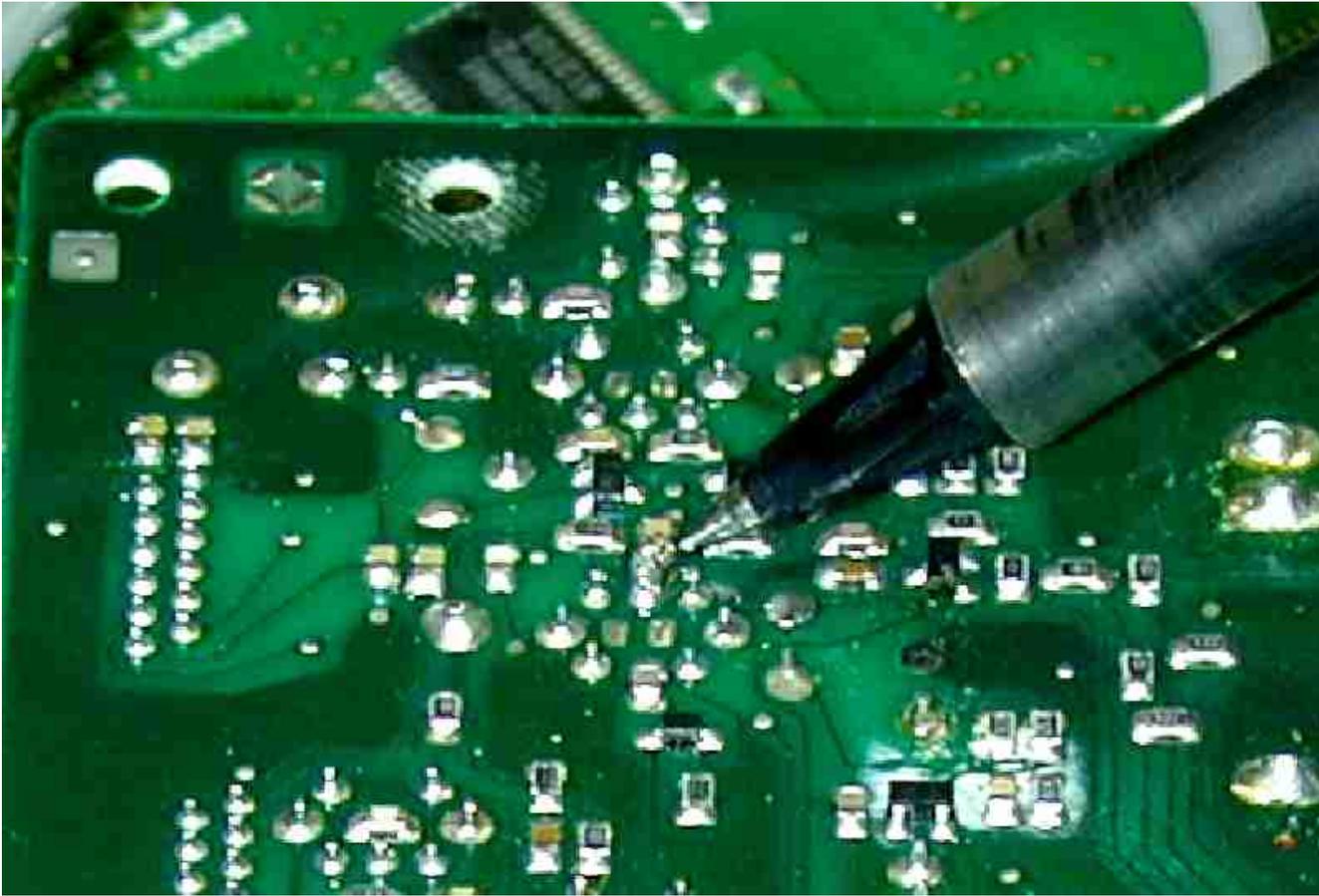
Note the FET above and to the left of the pen, and the IF transformer (two can shield connections and five electrical pin-outs, with an unused pad-set for a surface mount device in the center of the transformer leads) below the pen. The connection point for a new small insulated wire is the chip capacitor (C2148) pad that also connects to the center pin of the top three in-line pins of the IF transformer.

11.) Attach a small wire to the point mentioned above. It will route to the new circuitry.

12.) Re-install the board. Be sure you do not pinch any wires. Be sure you do not forget to plug in the ribbon cables, or any other wires you removed.

I mounted a small four lug (plus ground) terminal strip at the left front corner of this board, and mounted the components on that strip. This allows you the ability to change or adjust the mod if you find it too soft for high-speed operation.

Doing The Mod

The ideal CW radio would use a high-order filter with controlled group delay, and a reasonably linear attenuator or modulator to control the envelope shape. All other stages should be fully on just outside the output window of the CW signal. I initially hoped a CW "modulator" could be

added on, but for now it appears modifying the 1000MP to ideal circuitry would be too involved. My only option was to "hunt and peck" and find a modification that would be reasonable to do, and inexpensive. This is the best solution I could find, reduction of clicks was excellent. The only drawback is two resistors need to be hand-selected, and the radio needs some minor disassembly to reach a connection point on the RF board.

I mounted this mod on a separate terminal strip under a screw on the left front corner of the IF board. This allowed me to experiment with component values while watching bandwidth and other parameters. This is the basic circuit I used:
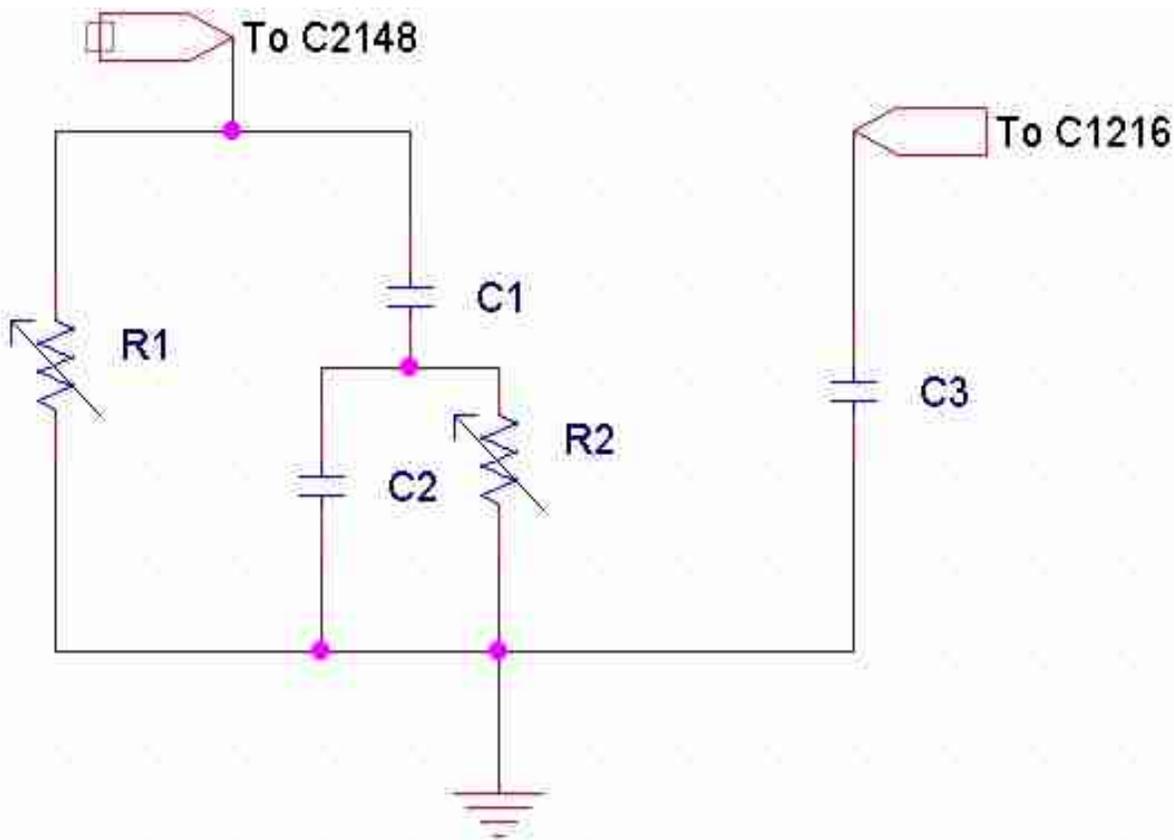
C3 was a .1uF disk capacitor. The value turned out to not be especially critical, it mainly seems to prevent rapid rise and fall of the low-level RF amplifier stage that is driven by a gate. There was no combination of resistance across or in series with this capacitor that reduced clicks in any of the radios I tested.

C1 and C2 are also .1uF disc capacitors. In all units tested, I could find no better combination for reducing clicks.

The only critical components appear to be R1 and R2. R1 ranged from 120k to 470k in the units I tested. R2 ranged from 1k to 10k ohms. I initially clipped in potentiometers, so I could listen to the

output and adjust the clicks at 1kHz spacing. Both pots were adjusted for a null in click amplitude. That null is rather sharp, and turned out to be around 30 dB deep. This takes the 1000MP from being one of the "clickiest" radios I have found to one of the cleanest!
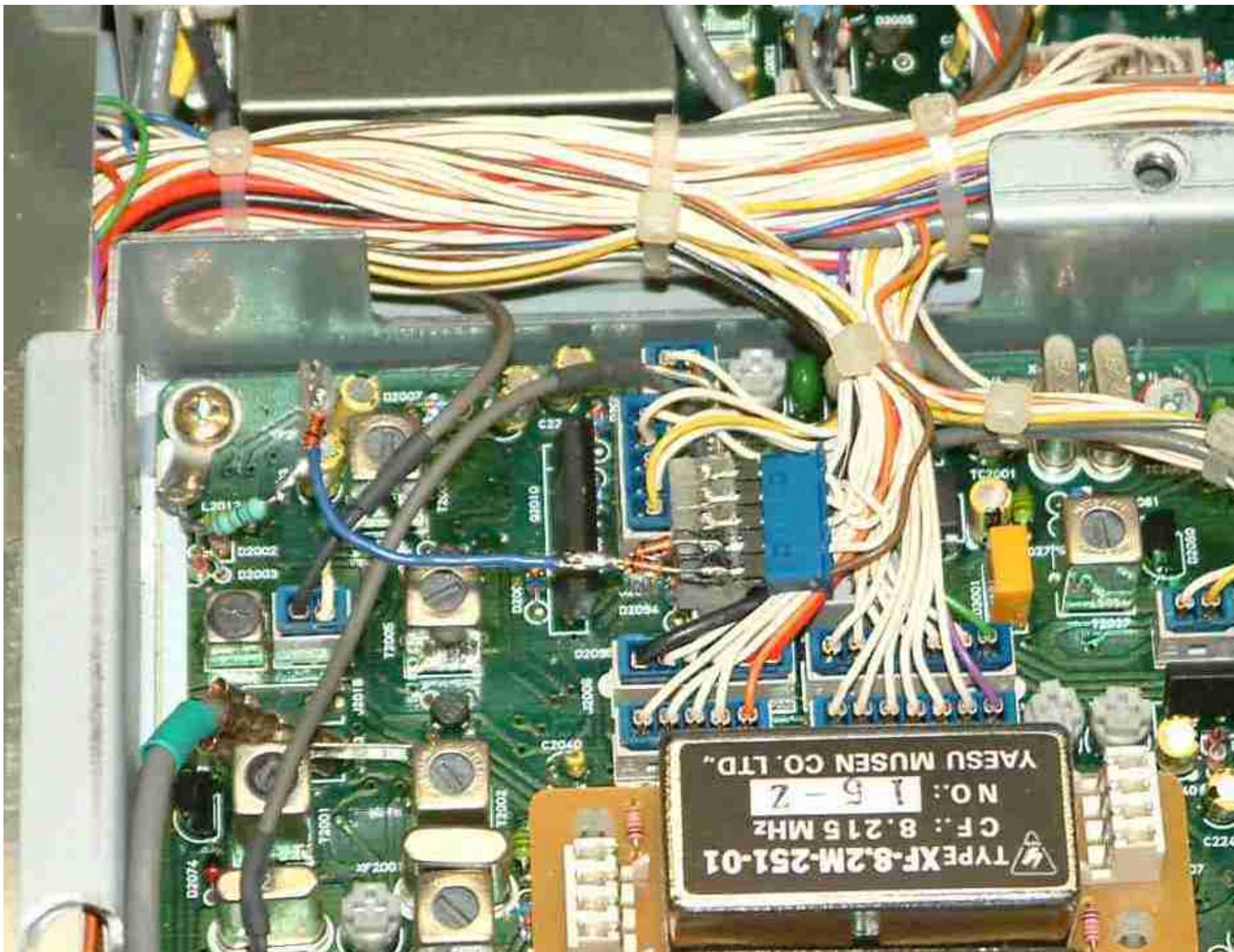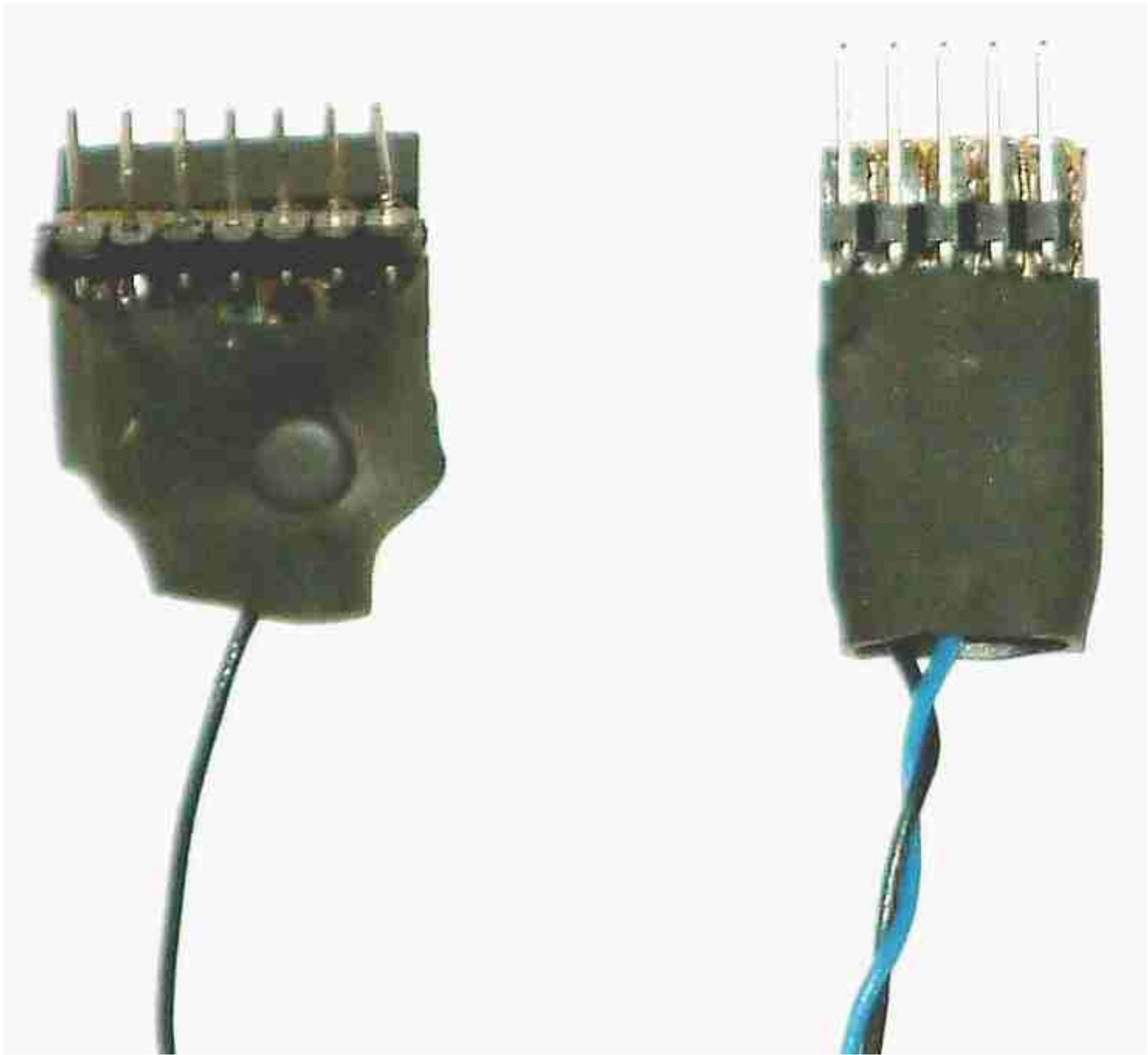
What_Causes_Clicks?

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Bandwidth rules Part 97.307

Clicks are often a problem on congested bands, yet with modern radios they don't need to be a problem. In the past, engineers and designers didn't have the easy ability to generate filtered waveforms. Radios lacked wide 2-3 kHz wide SSB filters, let alone narrow 250-500Hz filters. Op-amps were unheard of, and L-C filters were large, bulky and expensive. Today, every radio manufactured has the ability to be very clean.

Modern Radios

Most modern radios include 500Hz and narrower receiver filters that operate at the same IF frequency as their transmitter section. Transmitted signals are often routed through the SSB filters with intentional TX switching, when they could just as easily be routed through CW filters! We will see later in this page that TX signals could be routed through CW filters to eliminate sidebands, yet manufacturers foolishly use the SSB filters. If you examine the bandwidth of a FT1000-series radio, you will find the clicks disappear at the BW limits of the SSB filter. This is because they run an essentially unshaped CW signal through the SSB filter, and that filter sets the bandwidth of clicks.

The sad thing is once you listen to that signal through a 500Hz filter, it sounds absolutely no different *ON FREQUENCY* than it would if it were nearly click free. The only people who can notice the difference between a clicking rig and a clean rig are the people operating on adjacent frequencies! Claims that certain shapes produce certain "bell sounds" or high readability are not correct, they are certainly not based on engineering or actual blind A-B tests.

If you examine the audio output of your radio with a 500Hz filter in use, you will see ANY waveshape transmitter has the same nearly raised-sine shape output to your ear! That includes wide signals and narrow ones. The speed limit you can copy with a 500Hz filter is the same limit you can transmit with. It makes no difference what end of the path the filter is on, or if there is one at both ends, so far as speed is concerned! (This assumes the filters have reasonably good and easy to achieve group-delay characteristics.) We hear a few mS rise, no matter if it is a sine shape or a square, as a "tick". For demonstration, listen to the pure sine wave on WWV that "tics" every second!

What Could Be Done

At no cost to manufacturers, they could build a click free radio. Every component is in the radio, the problem all centers around poor or careless engineering.

Amplifier stages are reasonably linear (so they can amplify SSB), and virtually every radio contains power control circuitry that could be easily modified to provide wave-shaping. Even without wave-shaping, the transmitter could process transmitted CW though a 250Hz or 500Hz filter.

Sadly, most of the commonly used radios have as bad or worse keying characteristics than old rigs. It's as if the manufactures either don't understand CW, or don't care. The result is we are left with a mess, because many top-of-the-line and very popular rigs have horrible keying sidebands.

On frequency with normal CW filters, we would not be able to tell any difference between the sound of a clicking radio and one that is clean! There is no justification or reason for radios to be 3kHz wide on CW.

How to Identify Click Problems

We hardly notice clicks, and we certainly can not tell a clean rig from a dirty rig, when we are listening right on the CW station's frequency! Even an scope won't tell us much about signal bandwidth, or if the rig has excessive clicks.

In order to check clicks, we must:

1. Be sure the receiver is not overloading
2. Listen with the CW signal outside the receiver filter's bandwidth
3. Listen when the noise is low, and the signal reasonably strong

If we do not follow those three guidelines, we can't tell if a rig is clean or not. If you are testing your own rig, your second receiver must have a narrow filter and be coupled to the rig-under-test through a proper attenuator.

Why Worry About Clicks?

Clicks are most problematic when we try to copy weak signals next to moderately strong signals. If you only operate on empty bands, run low power, and never operate within four or five kHz of weak stations, bandwidth is probably not a concern.

If we contest, work DX, or Ragchew near other QSO's, and especially when we run more than a few hundred watts and have large antennas, we should be mindful of our bandwidth. If you listen to a

recording of a clicking radio, you can hear how devastating clicks are to nearby weaker signals. This signal is from Europe on 40 meters, and it is daylight over half of the path!!

For a mathematical tutorial on clicks, visit W9CF's site. Kevin's analysis deals with bandwidth requirements related ONLY to modulation of the envelope. I'll explain the same thing in verbal form, as I discuss sidebands created by rise and fall times. CW keying is really just 100% AM modulation, as you will see!

There are several INCORRECT but popular misconceptions. They are:

- **A signal has to be clicking and/or wide to send fast CW**
- **Clicks or sharp rises and falls aid in weak signal work**
- **Your CW signal bandwidth changes with the speed you are sending**
- **A certain shape gives a certain on-frequency sound**

What Causes Clicks?

While a fast rise and fall time guarantee excessive bandwidth, *a long rise and fall is no guarantee a radio will be "click-free".* Some radios switch into transmit while the synthesizer (VCO) circuits are still settling to a new frequency. An IC-775DSP I owned was particularly bad about this, and also had VCO leakage problems. The amount of garbage varied with how I used the radio, including "VFO" frequency settings of unused VFO's!

Radios with VCO or synthesizer settling time problems generally produce a loud "thump" on key closure on the second VCO frequency. That thump will be right on the DX station when the operator is working split. If you listen in pileups, you will hear a small percentage of rigs with this problem. If the operator uses QSK, VCO-switching-thumps can be particularly annoying. Thumps will occur every time the VCO moves from the receive frequency to the transmit frequency, sounding like a leading-edge click!

Rise and fall times are also important. *A long rise and fall time does not always result in narrow CW transmitter bandwidth, even though a faster-than-needed rise and fall time almost certainly results in excessive bandwidth.* Many radios have rise and fall times that are much too fast.

How fast is much too fast? For now let's ignore VCO switching problems, and consider envelope shape.

Rise and Fall

The ARRL recommends a 5 mS rise and 5 mS fall time for CW, based on data in section 2.202 of FCC rules and CCIR Radio regulations. According to professional sources, a 5 ms rise and fall time is not harmful to readability at 35 wpm under marginal (fading) conditions, and 60 wpm when signals are reasonably above noise floor. This rise and fall results in a occupied bandwidth of 150 Hz, although unwanted transient energy caused by the shape of the waveform slope may appear at wider bandwidths.

What Limits Bandwidth?

When determining bandwidth of a stable signal (no oscillator problems), two things come into play in.

- The *slope* of the envelope rise and fall at any point controls the band*width* of the keying sidebands
- The *amount of voltage change* during the slope controls the power *level* of the sidebands (clicks).

The slope (bandwidth) and the amount of change in a sloped area (level) combine to determine how offensive the transmitted signal is. *Very subtle changes in envelope shape have a profound effect on key click amplitude and frequency dispersion.* This makes it nearly impossible to tell if our radios are as clean as they could be by looking at envelope shape.

We can be certain sharp transitions will cause problems, especially if we can actually see them on a oscilloscope. We can also be sure that a rise and fall faster than 2 or 3 milliseconds will cause a bandwidth problem.

Reference Data for Radio Engineers, in the section of Radio Noise and Interference, addresses key clicks in a manner the ARRL Handbook does not. They give an example of multi-pole shaping of waveform. The ARRL Handbook seems stuck with the incorrect notion that a single-pole R/C filter provides proper shaping, something doubtless left over from 1940's technology when better filters were expensive, large, and complicated.

Here are the bandwidth curves of three basic envelope shapes, one rectangular (some radios are this bad!), one for a proper single pole R/C filter with slightly rounded shape (The ARRL suggests this shape. Probably because it was practical in the early years and "stuck" even though it is not ideal), and one for a filtered rise and fall (this would be a sine-shaped rise and fall from a multi-pole filter). We can clearly see a large difference in bandwidth in the curves below:
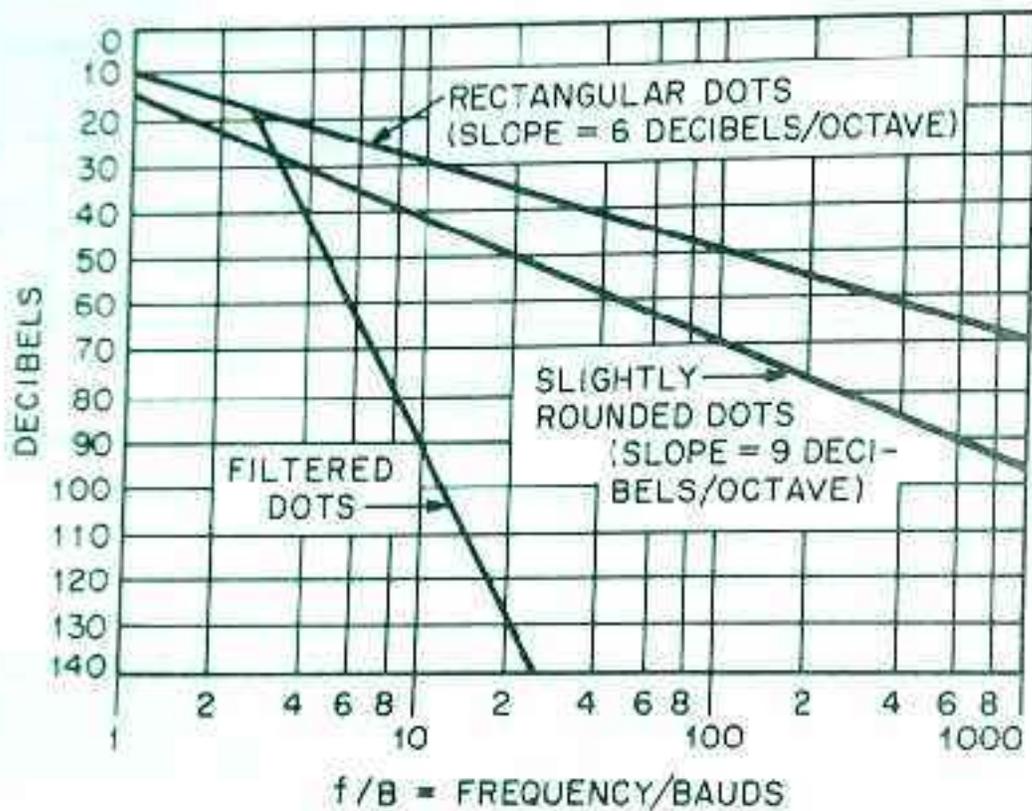
Fig. 12—Curves giving the envelopes for Fourier spectra of the emission resulting from several shapes of a single telegraph dot. For the upper curve the dot is taken to be rectangular and its length is ½ of the period T corresponding to the fundamental dotting frequency. The dotting speed in bauds is $B=1/t=2/T$. The bottom curve would result from the insertion of a filter with a pass band equal to 5 units on the f/B scale, and having a slope of 30 decibels/octave outside of the pass band.

From Ref Data for Radio Engineers 29-10 1977 Edition

Most radios, through poor design, fit in the rectangular to slightly-rounded category!

What Can Manufacturers Do?

Radio manufacturers can certainly do a great deal more than they are. First, they created the problems through poor engineering and design. Why are we stuck fixing them? Did they take our money and run?

All of the parts are there to make radios virtually click-free, yet the only manufacturer who has taken an active interest in this (and who seems to care at all about our signal quality and frequency usage) is Ten-Tec! To date I haven't found any other manufacturer admitting a problem, or even offering technical support for bandwidth problems.

Let me give an example of what could be done with current radios:

Virtually every radio contains a CW filter that operates at the IF frequency of the transmitter, yet nearly every radio transmits CW through the SSB filter! Engineers actually *added circuitry and parts*, in many cases, *to steer the CW through the wider filter on transmit!* If you listen to radios, in particular the FT1000-series, you will notice they have an ultimate click-bandwidth of about the same width as the SSB filter. That's because the poorly-shaped CW waveform with excessively fast rise-and-fall is filtered through the SSB filter.

If these same radios immediately turned on the output stages, and held them on for several mS after the key line was opened, they could send perfect filtered CW through the CW filter. A 500Hz filter would cause a steep roll-off in clicks, even if driven by a relatively "square" and very broad CW signal. The resulting waveform would be a slightly modified raised-sine envelope.

The listener would not be able to tell *any difference* **between the ON FREQUENCY sound of a 500Hz CW-filtered transmitter and an unfiltered signal with excessive bandwidth, if he used a 500Hz or narrower filter in his receiver!** As a matter of fact, I normally transmit through a 250Hz filter in my FT1000D, rather than the 2.4kHz SSB filter Yaesu selected. No one listening on frequency, even DX stations copying my signal near noise level, can tell the difference when I select 2.4KHz or 250Hz bandwidth! The only place transmitter filtering makes a difference is up or down the band from my operating frequency.

This is why we can not tell whether a signal has a proper rise and fall time, sharp level transitions, or any other envelope shape problem when we listen to the actual CW tones through a 500Hz filter. Even a very fast rise-time, with a spiked rise and fall, sounds good (and even looks perfect on a scope connected after the receiver's narrow filter)!

 Claim's that a certain shape rise and fall produce a "pleasing-sound" are not true at all. First, our ears can't identify a sound only 5mS long, and second...the receiver's CW filter (assuming it is under several hundred Hz BW) reshapes the waveform to a proper rise and fall!

Why is any of this our concern? Why do we have to work on radios, and suffer with clicks? Certainly not because of a cost issue! All the parts are in the radios. It is a simple lack of good design-engineering, most likely driven by a lack of concern by manufacturers for providing rigs with good signal quality.

What Can We Do?

First, we can let manufacturers know it is ***their problem***. Let's ask the ARRL to publish useful reviews with bandwidth pictures showing a spectral display of CW (and SSB) bandwidth. Let's ask them to check for VCO problems, and publish any abnormalities. Let's rate radios as poor, fair, good, or excellent so readers don't have to be EE's to understand what they are buying (and using).

Radios are too expensive, too difficult to work on, and last too long for us to ignore this problem. We need to stop these problems at the design phase, instead of out in the field.

FCC 97.307

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Federal Communications Commission

§ 97.307 Emission standards.

Subpart (a) is very clear. If we only have to change the value of three or four components to reduce bandwidth substantially do you think it is "good amateur practice" to not do so?

(a) No amateur station transmission shall occupy more bandwidth than necessary for the information rate and emission type being transmitted, in accordance with good amateur practice.

(b) Emissions resulting from modulation must be confined to the band or segment available to the control operator. Emissions outside the necessary bandwidth must not cause splatter or keyclick interference to operations on adjacent frequencies.

Subpart (b) is also very clear. If the stuff coming out is wider

(c) All spurious emissions from a station transmitter must be reduced to the greatest extent practicable. If any spurious emission, including chassis or power line radiation, causes harmful interference to the reception of an-

than needed, and if you bother someone, you are at fault! Because changing the value of a handful of parts largely corrects this problem, it clearly proves an unmodified radio is illegal to operate near other people.

interference to the reception of another radio station, the licensee of the interfering amateur station is required to take steps to eliminate the interference, in accordance with good engineering practice.

(d) The mean power of any spurious emission from a station transmitter or external RF power amplifier transmitting on a frequency below 30 MHz must not exceed 50 mW and must be at least 40 dB below the mean power of the fundamental emission. For a transmitter of mean power less than 5 W, the attenuation must be at least 30 dB. A transmitter built before April 15, 1977, or first marketed before January 1, 1978, is exempt from this requirement.

Subpart (c) is pretty clear also. It doesn't say "when the manufacturer recalls the radios you should get yours fixed."

Subpart (d) doesn't apply to normal modulation bandwidth issues, but if it did we would be in more trouble. The SSB IM3 and IM5 of most radios fails this criteria! Even if a radio did squeak by, as

soon as it was operated into an amplifier it would again become illegal!

I actually had a radio that violated 97.307(d) when used with even a perfect class-A 400 watt amplifier. The radio manufacturer told me "stop using an amplifier with the radio and you'll be fine".

Orion and IC-7800

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Checking bandwidth with receiver

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Bandwidth rules Part 97.307

Note: Bandwidth measurement dynamic range requirements are based on typical signal-to-noise ratios I have observed over the past several years. They are not the extremes of what I have seen, but rather are typical values. Some information on my receiving system and noise floor is available in (NOISE) and (RECEIVING).

A receiver can be used to check BW if we understand what we are doing! The common mistakes are:

1. **Using a band-scope**
2. **Leaving a noise blanker on (some receivers have a bit of this problem even with the NB off)**
3. **Using excessive (or SSB) bandwidth while tuning to check signal bandwidth**
4. **Using an inferior receiver or a receiver with poor close-spaced strong-signal performance (many DSP-only radios are pretty poor)**
5. **Using excessive gain**
6. **Having excessive background noise or signals**
7. **Relying on an S-meter (most are not linear or accurate).**

1.) Bandscopes

Bandscopes or spectrum analyzers using wide filters can't be used to check signal bandwidth. While such devices are good for monitoring band activity, they have too much bandwidth and are too easily subjected to overload to be useful in determining signal quality or bandwidth. Bandscopes do not have narrow filters and low internal distortion.

You can prove this by looking at a pure unmodulated carrier on the display. The bandwidth of the carrier is nearly zero (it may have some faint noise or jitter), the display should ideally show a perfect single spike of negligible bandwidth.

Bandscopes (and spectrum analyzers with wide filter bandwidths) are all but useless for determining bandwidth or signal defects.

2.) Noise Blankers

Noise blankers must be OFF when checking bandwidth or working close to strong signals.

In order to remove noise, noise blankers add a form of intentional distortion to signals, They do this by turning a switch or gate in the IF amplifiers off and on. An abrupt rise in peak input signal level over average signal level triggers the switch, and shuts the receiver off. The bandwidth of the noise detector is very wide, and this means a strong signal even 5-10 kHz away will activate the NB gate and distort signals.

Some receivers (like Yaesu's) do not fully remove the effects of the NB system, even when the NB is off! In some receivers you have to turn the NB off plus turn the NB gain down, the FT1000MP and FT1000MP MK V are examples of this. In others, like the FT1000D, you actually have to modify internal wiring to correct NB problems. The mechanism is explained in the links to the receiver mods.

3.) Bandwidth

You must select the narrowest filter possible to measure TX BW, certainly less than a 500 Hz filter with good shape factor.

Receiver bandwidth and shape factor directly adds to the transmitter's bandwidth. This means a perfect brick wall 2kHz bandwidth receiver tuning across a perfect 2kHz wide transmitter makes it sound like the actual signal bandwidth is 4kHz. Theoretically it is possible to deduct the receiver bandwidth from apparent measured bandwidth to obtain real bandwidth, but this generally means you have also decreased the dynamic range of the receiver (or spectrum analyzer). In practice, deducting bandwidths often produces unreliable results.

The slope of the receiver (or analyzer) filter is also important. If the receiver response is -6dB at 4kHz and -60dB at 8kHz, you will hear stuff out 8kHz (plus transmitter bandwidth) on very strong signals if you are in a quiet location.

4.) Inferior Receivers

Some radios, in particular DSP only radios, have very poor strong signal performance. They can't be trusted to give accurate BW reports.

Look at tests here, Sherwood Engineering's tests, or ARRL tests of close spaced receiver performance. Many receiver are not all that good. Most Yaesu receivers have a built-in design problem in the noise blanker amplifier that seriously deteriorates close spaced IM performance even

when the noise blanker is OFF.

Even the Sherwood engineering test is too wide for some receivers. The Sherwood test, for example, inflates performance of R4C's with the CF-600/6 filter. This happens because they measured *outside* the passband (2kHz) of the filter (600Hz). The second mixer in the R4C is a horrible design, especially the early FET mixer. Close-spaced tests should always be done inside the bandwidth of the roofing filter, or the roofing filter should be considered the narrowest reliably useable selectivity.

5.) Noise

If the band is noisy you really can't check a signal for low-level IM, clicks, or splatter. The noise will cover up any weak signal defects. There must be at least 50dB headroom between the peak signal level being tested and your noise floor to check bandwidth on SSB, or 80dB of signal to noise headroom to check CW bandwidth if you operate near weak signal areas of the band. For general ragchewing away from weak signal areas 50-60dB headroom is generally enough.

Some bands are a special case because SSB operates near weak signal CW stations. 160 meters is one example. Bandwidth of higher power SSB transmitters operated near weak CW stations can be problematic. I often hear spits from Icom 756 and TS 2000 transmitters on SSB as far as 10kHz away on 160 meters.

Receiving noise floor is probably the single most common source of false "clean signal" reports to what actually are problem transmitters. If the noise or QRM floor is high, you won't hear spurious signals.

Wide-audio operators are particular victims to giving each other false assurances of how narrow and clean they are. They often use "opened-up" receivers that absorb more noise power from wide bandwidth (remember noise power is directly proportional to receiver bandwidth) and they often live in noisy environments. It takes a good weak signal narrow receiver in a quiet location to properly check bandwidth.

Results

The result of the factors above is that some people will report a nasty signal "clean" when it isn't, and some will report a signal "wide" when it isn't.

Understanding how to test will correct problems, and help us use our own equipment better. Receivers make very good measurement devices if used *properly*.

Receiving

[Home] [[Up](Up)] [[Beverage Antenna Construction](Beverage_Antenna_Construction)] [[Echelon-Log Beverages](Echelon-Log_Beverages)]
[K9AY_Flag_Pennant_Ewe] [[Magnetic Loop Antennas Receiving](Magnetic_Loop_Antennas_Receiving)] [Receiving_basics]
[[Slinky and Loaded Beverages](Slinky_and_Loaded_Beverages)] [[Small Vertical Arrays](Small_Vertical_Arrays)] [W8JI_RX_ANTS]

Noise and common mode noise. [Power line and other noise](Power_line_and_other_noise) sources. [Pre-amplifiers.](Pre-amplifiers)

This page has top links to other receiving antennas such as Beverage, loop, and vertical antennas. This area deals primarily with low noise antennas, and discusses effect of antenna directivity on weak-signal reception.

Noise Floor and signal Levels at my location:

My local wintertime 350Hz BW noise (after amplifier) compared to a sample of signals on one night was:

Noise -127dBm

9H1BM -122dBm

OM0WR -95dBm

DF2PY -88dBm

WA8OLN -78dBm

W3GH -60dBm

W4ZV -32dBm

The dynamic range between noise and W4ZV was 95dB! It also illustrates how important antennas, location, and propagation are rather than power. The dB difference between signals from the same area can be profound. Many DX signals run at or near noise floor, while others can be 40dB out of noise floor. The data above is certainly not typical of every night, but it shows how large the signal level variations between weak DX and strong signals can be. Over a period

of time I've found 85 to 90 dB dynamic range about the most that is every needed. In a simple installation with a single Beverage, 80 to 85dB IM3 and blocking DR is probably enough. Receivers with less than 80dB IM3 and BDR probably compromise a reasonably good station's capabilities.

Comparison or Ranking of Receiving Arrays or Antennas

The following is a ranking of receiving antennas based on noise being evenly distributed in all directions. These rankings are most accurate in the frequency range of AM broadcast, 160 or 80 meter bands when:

1.) The receiving location shows a nighttime increase in noise level. In other words the system is not limited by local or internally generated noise, instead being limited by skywave or distant propagated noise.

2.) Thunderstorms or other local noise such as power line noise from specific directions does not dominate the receive system noise floor.

There will be occasional exceptions, but as a general rule the ratio of peak response in the direction of the signal to average response in all directions is the dominant factor in determining how good an antenna works for receiving. In virtually all installations without clearly dominant direction or directions of noise arrival, RDF (receiving directivity factor) very accurately predicts receiving antenna performance.

RDF (directivity) will be an almost perfect indicator of what you can expect from your antenna as long as:

- **Noise is not from the same general direction as the desired signal**
- **Noise field strength is not greater than the ratio of peak antenna response to depth of the pattern in the direction of noise**
- **Noise is not coming from within the antenna's nearfield or Fresnel zone**

In the vast majority of systems, the following RDF table indicates relative performance of antennas:

Antenna Type	RDF (dB)	20-degree forward gain (dBi)	Average Gain (dBi)

1/2wl Beverage	4.52	-20.28	-24.8
Vertical Omni, 60 1/4wl radials	5.05	1.9	-3.15
(Ewe Flag) Pennant	7.39	-36.16	-43.55
K9AY	7.7	-26.23	-33.93
1/2wl end-fire Beverages	7.94	-20.5	-28.44
1-wl Beverage	8.64	-14.31	-22.95
two verts optimum phasing 1/8 wl spacing	9.14	-22.46	-31.6
two 1wl Beverages Echelon 1/8 wl stagger	10.21	-15.45	-25.66
Small 4-square 1/4 wl per side (opt. phase)	10.70	-15.79	-26.49
1-1/2 wl Beverage	10.84	-10.88	-21.72
Small 4-square 1/8wl per side (opt. phase)	10.97	-30.28	-41.52
Single 1.75wl Beverage	11.16	-6.50	-17.66
2 Broadside 1.75wl Beverages .2 wl spacing	11.36	-3.51	-14.87
2 Broadside 1.75wl Beverages .4wl spacing	11.91	-3.50	-15.41
.625wl x .125wl spaced BS/EF vertical array	12.5	-19.5	-32.0
2 Broadside 1.75wl Beverages 5/8 wl spacing	12.98	-3.50	-16.48
2 Broadside 1.75wl Beverages .75wl spacing	13.48	-3.49	-16.97

If antennas are within two dB of each other in RDF, a lesser ranked antenna may occasionally outperform slightly higher RDF antennas. This is because:

1. *Direction and polarization of arriving signals and noise constantly vary, and so the*

relative relationship of each to any individual antenna's response will vary.
2. ***Through various unavoidable errors or omissions, antennas in the real-world may not work precisely as predicted.***

Gain vs. Directivity Myth

One common rumor or myth is that higher antenna gain results in improved reception. Gain is an unreliable way to predict receiving ability on frequencies below upper UHF! A clear example is illustrated above by comparing gain of the single 1.75wl Beverage to the pair of 1.75wl phased Beverages that are spaced .2wl apart.

In this case, the single Beverage has a gain of -6.5dB. The pair of Beverages has a gain of -3.51dB, a gain of about 3 dB. Despite the gain change, antenna directivity and pattern do not change a noticeable amount. RDF only increases 0.2dB, and undetectable difference. Pattern remains essentially the same, so reception remains the same.

Spacing must be at least be 1/2 wl or more for phased Beverages to add noticeable improvement to reception. Even at 3/4 wavelength spacing the directivity improvement falls short of 3dB!

Gain of any spaced pair is about 3dB more than a single Beverage, but reception improves and antenna pattern changes only with relatively wide spacings.

Of nearly equal importance, end-fire arrays actually work better with closer spacing. For an example, compare the 1/8th wl four-square RDF with the 1/4-wl four-square array.

How well does the above hold true?

Over the years, I have had virtually all of the above systems. I always have multiple phase-locked receivers on multiple antennas listening in stereo or a very fast way to "A-B" antennas. When an antenna sits unused most of the time, I replace it with a more useful antenna. My single Beverages are now virtually all eliminated, my last phased loops were in the 80's (when I had four end-fire diamond terminated loops). Even on 80 meters, my large arrays with 300-350 foot spacing almost always beat my single long Beverages. I've migrated towards the bottom end of the chart with all my antennas ***because they actually do receive better.***

If you ask operators who visit for contests, everyone prefers the large vertical or wide-spaced Beverage arrays. Guest operators, given a choice, almost never not use single Beverages or close-spaced Beverages.

You can listen to directivity examples on my [DX Sound files](#) page.

Beverage Antenna Construction

[**Home**] [**Up**]

Link to Beverage Antenna Components I use

I installed my first Beverage Antenna (in the early 1970's). I was delighted to find a large improvement in weak-signal reception from such a simple, inexpensive antenna. Over the years I've continued to use, compare, and refine my Beverage antennas. Despite having very large vertical arrays, Beverages remain my primary DX receiving antennas. There just isn't any antenna that is as simple, as easy to construct and maintain, and as foolproof as a Beverage!

I refine my antenna systems by comparing systems against each other for extended periods of time, usually more than a year. My station has a convenient switching system allowing instant comparison of antenna systems. When an antenna system is almost never used, I abandon that system and try something else. Even though I use engineering tools (books and models), I always compare and measure actual working systems. I presently have over thirty Beverages in three different clusters of arrays, the end result filtered through years of measurements and A-B testing of systems.

A great much has been written about Beverages. Unfortunately much or most information is a repeat of previously published information (and misinformation) from verbal discussions or from other articles or handbooks! It is time to set aside some of the myths that have been handed down and repeated so much they have become "fact".

Types of Beverage Wire

The most commonly used wire types are single conductor hook-up or electrical, electric fence wire, and special antenna wire such as copperweld. The only significant and easily noticed difference between these commonly used wires is in physical properties, such as ease of soldering, strength, and life

Insulated Wire

Sporadic claims have appeared indicating insulation prevents charged droplets of water from making an antenna "noisy". I've never been able to verify that rumor either in A-B tests of actual antennas or through planned experiments. Other reports, many from reliable sources, also seem to discredit this rumor.

One of my experiments was to charge a stream of water (against earth) with an extremely high voltage supply, and spray the water on a wire. Other than corona noise from sharp points, the type of wire made no difference at all in "noise". The water droplets obviously were not hitting the wire like hundreds of random charged capacitors, they generated no noise at all. This is really what we would expect, if we consider that each drop contains only a very miniscule amount of change and also has nearly perfect insulation (distilled water is a very good insulator).

Controlled observations also tend to support the idea corona, and not charges in individual droplets, cause precipitation static.

In Ohio, my long Beverages stretched across open farm fields. Snow would whip across the fields, rain would pelt the wires, yet insulated and bare wire Beverages running in the same direction always had the same noise level. Beverages that picked-up corona (or "p-static") noise were always near or aimed at tall towers. With corona sizzling at 40-over-nine on my tall towers, Beverages (and even small "magnetic" loop antennas) aimed at the towers would "hear" the same precipitation noise.

The same was true for tower-mounted antennas. The largest noise problems came from antennas mounted high on towers, and generally were with antennas that had "sharp" ends jutting out in the air. Lower antennas, even those of identical construction, were either significantly quieter or totally free of precipitation static. This effect was reported many times by contest operators and DX'ers with stacked antennas. They universally switch to low antennas to eliminate or reduce p-static, even though the *same moisture is hitting the lower and upper antennas.* This strongly indicates precipitation static is from corona discharge, and *not* from charges in each individual drop of moisture hitting the antenna.

After my move to Barnesville, Georgia my first antennas were all insulated wire. Hook-up wire was pressed into service in my first group of temporary Beverages. As non-insulated conductors permanent antennas were added, there wasn't any observable change in inclement weather noise. As before, the antennas nearest or aimed at tall towers picked up some p-static noise. Antennas located away from the towers remained free of precipitation static, whether bare or insulated.

There is also some chance, if the antenna wire is not under significant tension, that insulation may sometimes hide a broken conductor.

Insulated wire may reduce leakage currents if a substantial part of the conductor is in contact with resistive paths, such as wet brush or tree branches, but you may be better off trimming back any substantial foliage in contact with the wire.

While insulated wire has no major performance disadvantage, it also has no advantage.

Type of Conductor

Copper wire is a good choice if supports are close. Copper wire lacks the mechanical strength of steel-core wires, but is very easy to work with. It is softer, making it easier to bend. Copper wire can be repeatedly scraped or re-soldered without worries about piercing a thin copper coating and exposing a rust-sensitive steel core, and it is readily available and relatively inexpensive in large quantities.

Copperweld wire is much stronger and has about the same RF resistance as 100% copper. Like copper, it is easy to clean and solder after it has been exposed to the weather as long as you are very careful to not scrape through the thin outer coating of copper. It is considerably more difficult to work with than normal pure copper wires, any kink or sharp bend will substantially weaken the wire.

Most fence wire I've found is cadmium plated, rather than zinc galvanized. Using RF current meters, I have measured increased losses when using zinc or cadmium plated steel wire. Beverages already have substantial current loss due to the close proximity with lossy earth. I've measured about 60% of feed point current remaining (~4.4dB loss) after passing over around 700-feet of electric fence wire, and about 10% more current (~3.1dB loss) using copper-clad steel wire. Steel fence wire would aggravate losses that already limit the benefits of using long Beverage antennas. *In a very long antenna, the small*

additional loss of steel fence wire might slightly reduce performance.

In my Beverages, the important consideration is antenna maintenance. I use copperweld wire or electric fence wire, because strength is a primary concern. With spans exceeding 200 feet, my antennas need a large strength to weight ratio.

Don't use welding wire! It is a very poor material choice. It rusts (and as with aluminum) you'll have connection problems in no time.

Beverage Supports

Some would have us believe we need non-metallic supports for our Beverages, but there is not the slightest technical justification for such suggestions.

The only requirement for the support is it must hold the antenna up, and it can not connect the antenna to ground. A metal pole with a small PVC stub for an insulator is every bit as good as a full non-metallic pole. Trees make good supports, especially if you use nail-type electric-fence insulators for use with wooden posts.

I've never seen a problem allowing a wire to contact a branch, although I do trim out the branches and avoid any contact with trees.

For end supports, I use trees, pressure treated lumber, or landscaping timbers. With a lot of tension, I backstay the poles to a dead-man (generally an old brick) buried in the ground. When I set end-posts with my power auger, I line the hole with copper flashing. That becomes part or all of the feed point (and termination) ground connection.

I never anchor or wrap the Beverage wire around insulators, except at the ends. I always allow the wire to "float" through the insulators. When the wire floats, you can tension the entire antenna from either end. If anything breaks the wire, you can see it at any point! A "floating" wire is much easier to repair if it is damaged, because you only need release tension on one end to splice the wire. Re-tension that same end, and everything is restored. It takes no more tension to support a 1000-foot Beverage with supports every 100-feet than it does to support a 100-foot wire between two rigid supports, but it is a much more difficult to break the longer wire. A longer "floating" wire will often take-up enough slack to remain up after deflecting a large tree branch, where a shorter rigidly-anchored span will almost certainly break either the insulators or wire.

Beverage Insulators

If you expect a long-lasting antenna and have a long antenna, be careful when choosing insulators! Some types of electric fence insulators will not last long. The unreliable types of post insulators have two square folds to hold the wire, a square shaped base, and nail through a small molded plastic angle. The weak points of this insulator are the square retaining tabs, and the molded nail tube at the insulator base. When this type of insulator is mounted horizontally, the wire's weight will stress both the molded nail tube and a single tab. I typically find about 10% of the insulators fail within a few months. After three years, the few dozen installed here have virtually all failed.

Avoid these types!
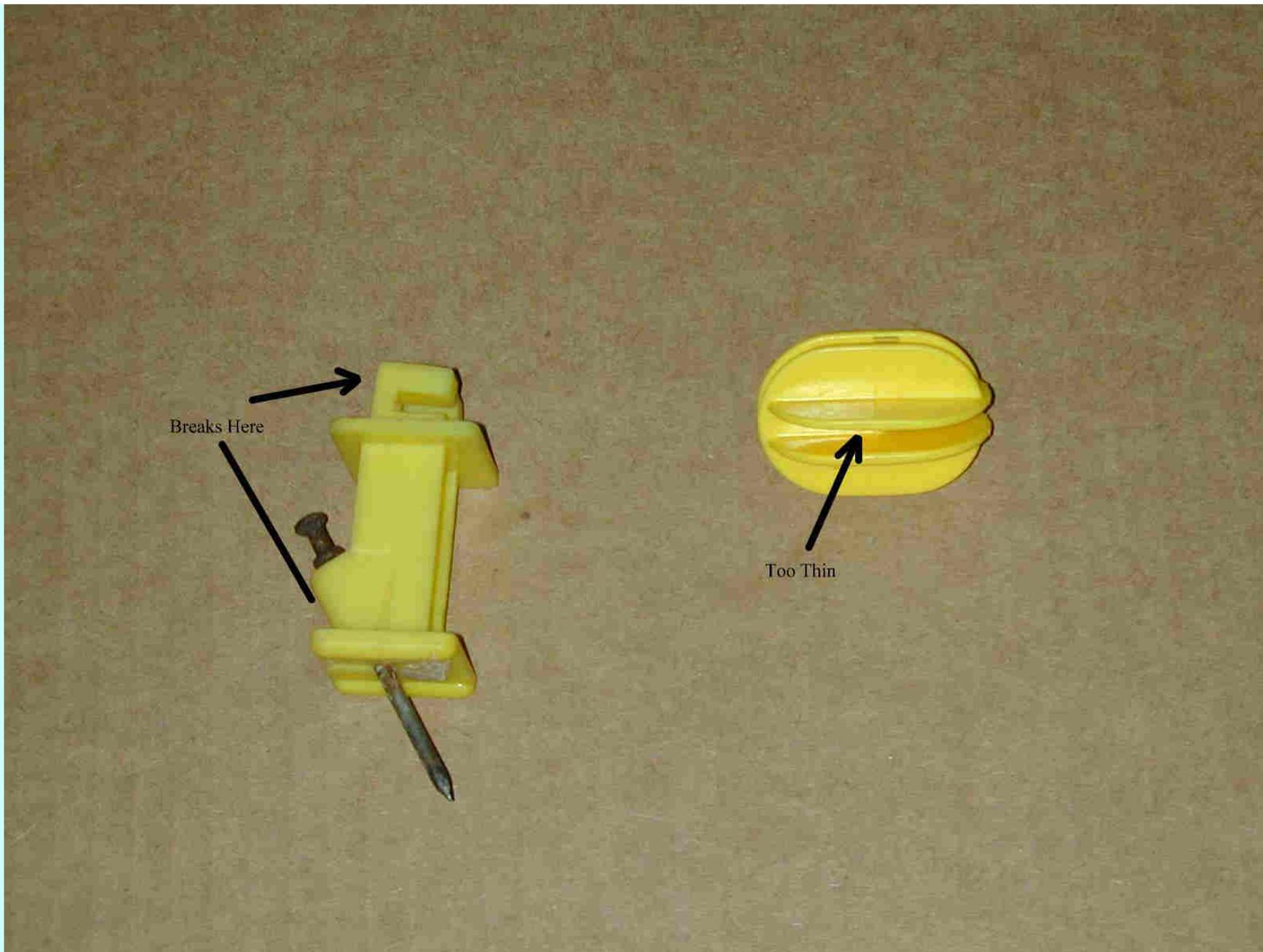
Round yellow or back plastic insulators with the nail going through the center, like the examples below, are much more reliable post insulators

Ceramic post insulators may look great, but they do not allow floating the wire across the insulator. Even if you do manage to find a ceramic insulator that allows floating the wire, the ceramic will quickly wear away at the constantly moving wire. Avoid ceramic insulators, unless you are prepared to "buffer" the wire through a UV resistant soft plastic bushing!

Good end-insulators are becoming difficult to find. I always use compression types, but the material has to be either ceramic or very thick plastic. Some very thin plastic compression insulators will actually cold-flow and allow the wire to pass through the insulation. This is particularly true with thin steel wires that are tensioned over 25 pounds. Heavy-walled egg insulators are much more reliable, and not subject to wire migration through the thin insulation.

My favorite insulators are large these rather thick Fi-Shock yellow plastic insulators. They are slippery enough to allow the dead-end wire or rope to loop over the insulator, and create a 2:1 mechanical advantage when tensioning.

Height

I've found very little performance difference with height, unless the Beverage is more than .05WL high. As the height exceeds .05wl, performance seems to be reduced. Small rolling hills or ravines also seem to make any difference. Follow the contour of gradual slopes, and go straight across ditches or narrow ravines without following the contour.

Sloping Ends

There really isn't a logical reason to slope the ends of a Beverage. After all, six-feet of vertical drop is six feet, no matter if the drop is over 50 feet or straight vertical.

Consider, for example, the K9AY or Pennant antennas. Both have sloped wires, yet virtually all of the response is from the vertical slope of the wire in the antennas. As a matter of fact, the actual shape makes very little difference in the way each antenna works. Why would anyone, knowing how a Pennant or K9AY works, think that a Beverage somehow magically breaks tradition and stops responding to vertical signals in the wires when we slope them a bit? What difference would it make in noise anyway, since the entire antenna responds to vertically polarized signals?

There isn't any possible way, including use of shielding or additional conductors, to prevent the end-wires from having the very small effect they have. Save yourself time and worry, and avoid a needless hazard. Just drop the end-wires vertically right down to earth.

Multiple Antennas Crossing

Crossing of Beverages has little effect if they are not parallel or nearly parallel. Try to cross at an angle of 90 degrees if possible. Even a few inches of spacing is enough for right angle crossing. With shallow angles, assuming they can not be avoided, increase wire spacing to a few feet.

Transformers

Always use isolated transformers for feeding Beverages. It is cheap, simple, easy insurance against unwanted common-mode ingress of noise and signals into the antenna from the feedline shield. See the Common Mode Noise page for an analysis.

I use 73-mix FairRite Products 2873000202 cores (about 1/2 inch square and 1/3 inch thick 73 material) in my transformers. These cores require a *two-turn 50-75 ohm winding.* The high-impedance winding is *5 turns for 75-ohm cables* (6.25:1 Z ratio) or *6 turns for 50-ohm cables* (9:1 Z ratio). Small insulated hookup wire is actually better than enameled wire. The thicker insulation is much less susceptible to developing shorted turns in rough service.
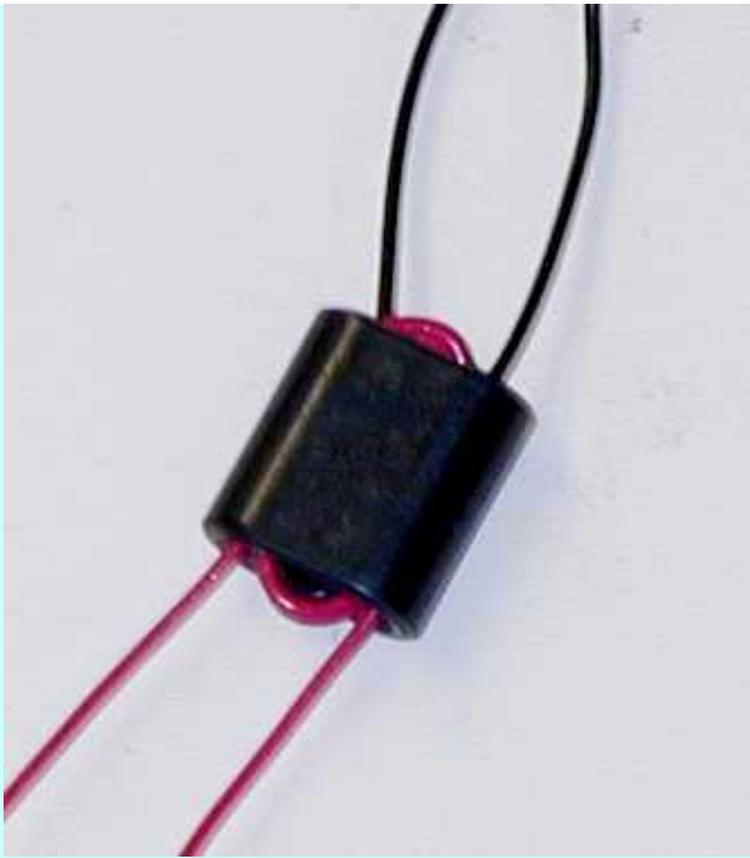
While my early transformers were waterproofed with Krylon and coated with insulation foam, I have finally laid out enclosed transformers and terminations with internal lightning protection. The transformer sections have F-fittings, and all use stainless steel hardware.
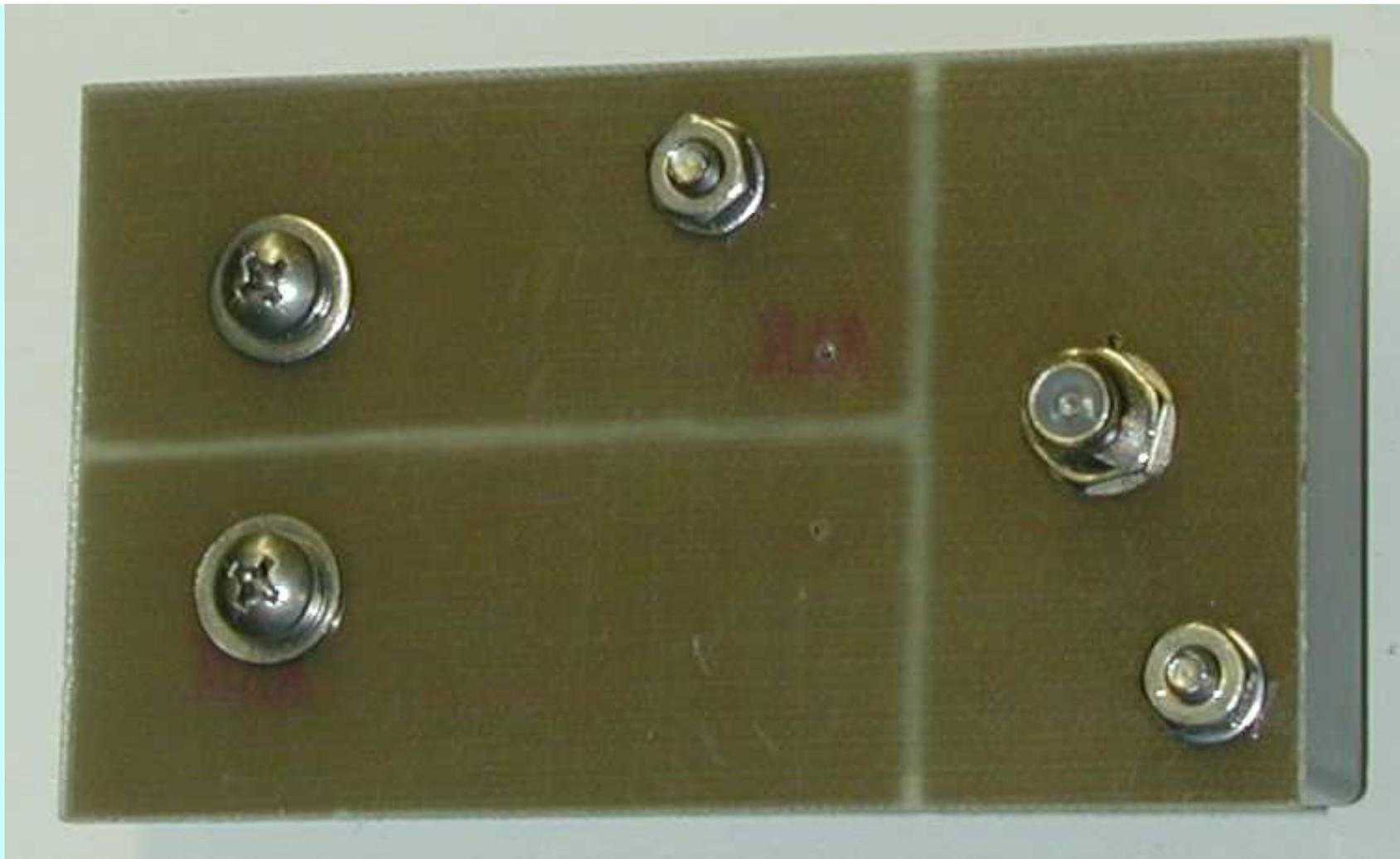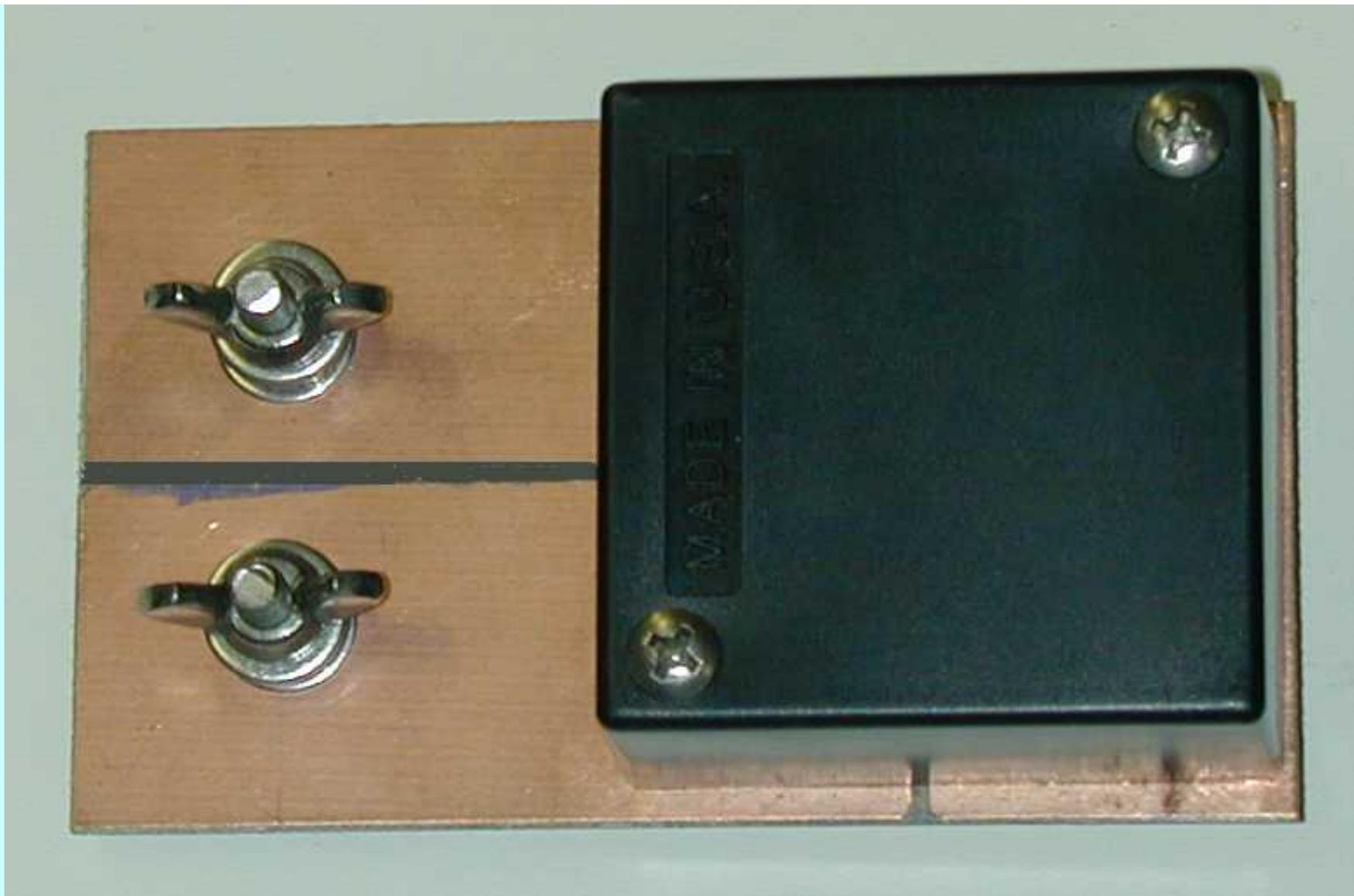
For a Reversible Beverage, I use the following:

Multiple Antennas at One Feedpoint

Never bring multiple antennas to one feedpoint, especially when they share one common ground. I've noticed a definite deterioration in pattern with multiple feedpoints arranged with only ten feet of spacing, even when they had separate ground systems. One set of Beverages installed with 5-10 foot of feedpoint separation has noticeably poorer patterns than other identical length antennas with wide separation at the feedpoint.

Multiple antennas actually may be the only case where a sloped feeder can make a difference, the slope will actually move the effective feedpoints further apart. The best idea, however, is to separate the feedpoints by several times the antenna height.

Termination Value

Having precise termination values isn't necessary, but get as close as you reasonably can. There are some impedance measurement suggestions circulating that absolutely do not work. One is to just use a tuner to match the terminated (or unterminated) antenna, and work backwards with loads to measure tuner impedance ratio after matching. This won't tell you a thing about proper termination, unless you repeat the measurements on dozens of frequencies spread over a wide range!

There are three fast, simple ways to test for proper termination:

With an Antenna SWR Analyzer

1. Connect the antenna analyzer at the Beverage feedpoint through a good matching transformer
2. Sweep the analyzer frequency from 1.8 to 7 MHz (or over a ~4:1 frequency range near the frequency intended for antenna) while watching SWR
3. Adjust termination for minimum SWR *variation* (not minimum SWR, minimum SWR variation!)

When installation (including grounds) and termination is proper, *SWR VALUE* will remain nearly the same regardless of frequency

With an Antenna Impedance Meter

1. Measure the feedpoint impedance (right at the feedpoint) of a roughly terminated antenna at the frequencies of *highest and lowest* resistive impedance. You can do this through a known good transformer by correcting impedance for use of the transformer
2. Multiply the lowest measured impedance by the highest, and then find the square root of that number. This will be the correct termination impedance of the antenna

With a Clamp-on RF Current Meter

(This does not work well with voltage, because of measurement method error problems)

1. Apply a small amount of power from a transmitter, do not exceed antenna system component ratings!
2. Measure current at the termination, and several points up to a distance of at least 1/2 wl from the termination
3. Adjust termination resistance so current shows a smooth current decline as you move the meter towards the termination

Note:

In about 500-800 feet of distance, power loss in a Beverage is around 3dB. This corresponds to a 1/3 reduction in current. If you attempt to adjust for equal currents (or voltages) over any distance, the antenna will be MIS-terminated!

Termination Components

Identifying a Composition Resistor

We commonly assume any brown phenolic resistor is a carbon composition resistor, but that isn't true. Most of the smooth brown-colored phenolic cased resistors manufactured after 1960-1970 have actually been carbon film resistors. There are only a limited number of manufacturers supplying carbon composition resistors. One is Allen-Bradley. They are expensive special-order parts, and the buyer must specify composition types.

COMPOSITION TYPE CARBON

FILM TYPE CARBON

COMPOSITION TYPE METAL

FILM-TYPE

Spiral coating of
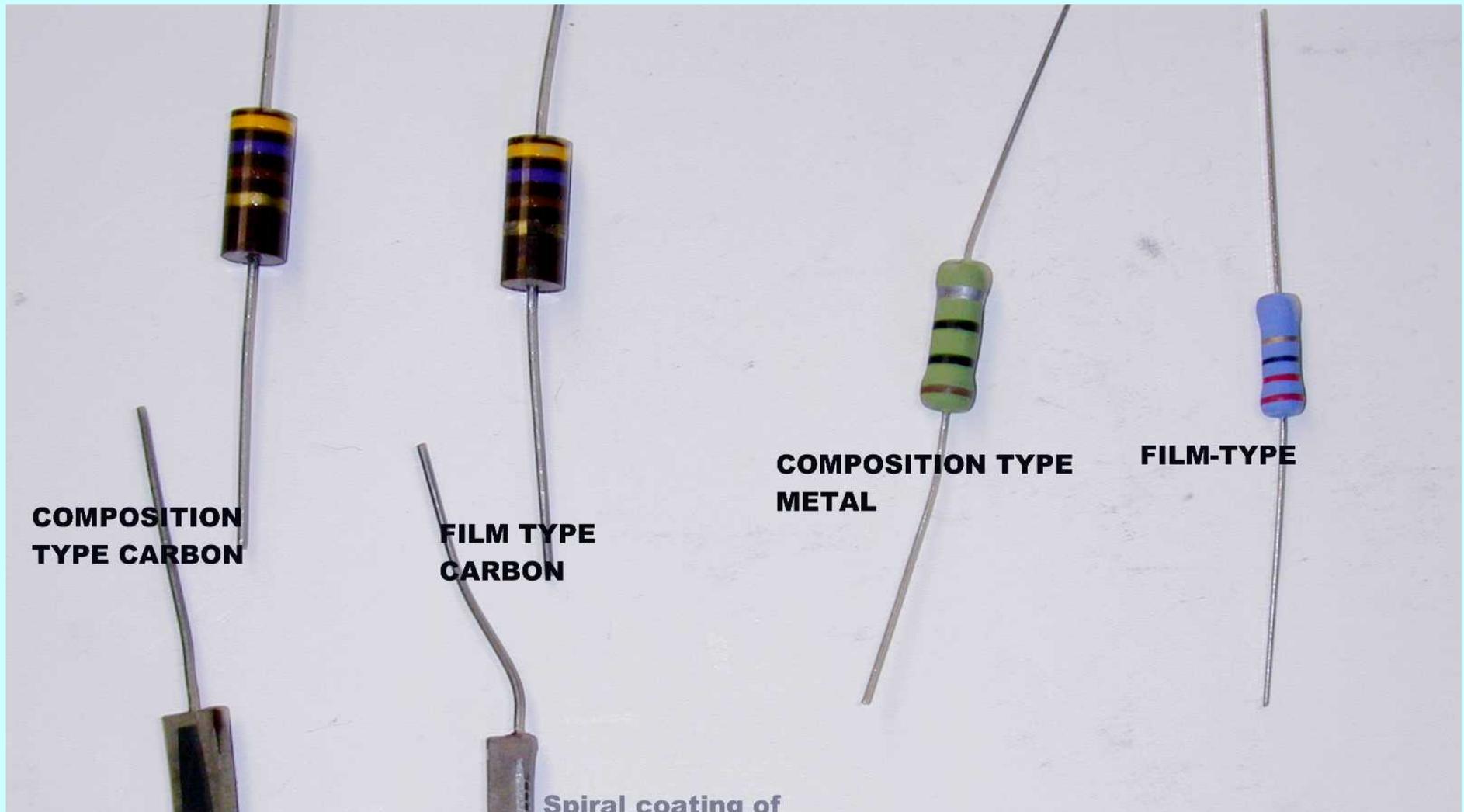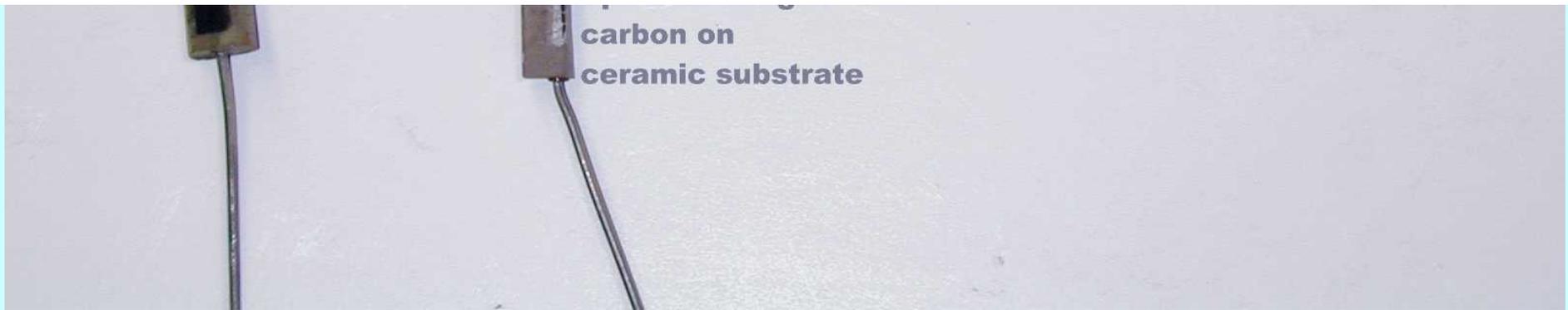
As we see from the photo, it is impossible to identify a composition resistor by external appearance.

The only sure way to identify a resistor, short of ordering it from a reputable source, is through a destructive test. We can, for example, apply a large momentary overload and look for a resistance change. A resistance change indicates a film-type element. We could also cut the resistor open, and look for a non-conductive core. A non-conductive core indicates the resistor is a film style component.

Why Composition Types?

We need composition resistors in any application where the resistor is subjected to very-large very-short overloads, or where the system demands a nearly pure resistance at a very high frequency (F>100MHz).

Obviously, in the case of a Beverage at a few MHz or lower, we could get away with using many styles of wire-wound resistors or spiral-film resistors. A small amount of inductance would not be a major problem, and virtually ALL carbon or metal film resistors (constructed with resistance elements deposited or cut in a spiral on an insulated core) would not have excessive inductance. The thing we can not tolerate is the sensitivity of non-surge rated components to damage from lightning storms, even distant storms.

The life of a carbon or metal film resistor, when used as an antenna termination, is relatively short in most locations. Just a few coulombs of energy, when applied in a few milliseconds, will cause a carbon or metal film resistor to change value. Worse yet, the resistor will not be altered in appearance. (Carbon also has a strong tendency to change value with heat. Even modest operating temperatures, over a period of time, will cause a carbon resistor to change value. Metal resistors are more stable.)

Unless you want to make a full-time career out of testing your antennas and replacing resistors, use a energy absorbing composition type resistor!

I install a small lightning gap of about 1/8th inch across my antenna's ends, both at the feedpoint and the termination. This helps immensely with very close strikes. I use either Ohmite OY-series metal compositions or A-B carbon composition resistors. You can buy metal composition resistors at DX Engineering.

Ground Systems

The ground system mainly provides an RF and lightning ground. Having a very low ground-resistance is not especially important, unless an Autotransformer or Un-un is used! Autotransformers and Un-un's don't isolate the feedline for common-mode. The antenna needs a stable ground, not necessarily a low-resistance ground.

In my tests over the years, a 3/4-inch copper pipe driven five feet or deeper into the soil typically measures between 50-150 ohms of RF resistance on 160-meters. (DC or low frequency AC measurements will NEVER give the correct earth resistance for RF, and they certainly can not tell us ground conductivity.) Unless you have exceptionally poor soil, going deeper than five feet will not reduce RF resistance on frequencies above 1.8 MHz. Skin effect limits the depth of RF current in the soil, so the extra rod depth does nothing. Lower resistance values (about 55 ohms) were obtained in a wet marshy area of NW Ohio, with a very rich black acidic sandy loam soil. The higher resistance were obtained in rocky clay soil typical of the Atlanta, Georgia area.

My present location has rolling pastures and wet clay soils, providing under 100-ohms of RF resistance at 1.8MHz with a five-foot rod.

The general guideline I follow is to use at least two five-foot copper rods (I use 3/4" copper spaced 5 feet apart). If I can not get full depth, or if the soil is particularly poor, I add a few 30-60 foot buried radials. The idea is to obtain a reasonably stable ground, so termination does not change.

CLICK TO LOOK AT ACTUAL MEASURED GROUND TERMINATION RESISTANCES!!!

If you are unsure if you Beverage's ground is adequate, measure the impedance of the beverage with an antenna analyzer with your operating ground systems. Note the reading. Add two temporary radials 1/4 wl long suspended above earth at right angles to the Beverage, and re-measure the impedance. (It is OK to have them there at right angles to the antenna and not have them connected, and them connect them while taking readings.)

You can measure the impedance on the low-Z side of a good transformer. Under almost any condition, the wires would have 100 ohms or less impedance. If you see a very noticeable change in impedance, you probably should consider improving the ground system. Impedance changes of 15% (or larger) indicate a potential ground stability problem, because the ground resistance would be nearly 100 ohms. This test should be done when the ground is dry, or any time you think you might be having a ground problem.

Always remember to keep the shield of the cable isolated from the Beverage ground! Never use un-un or autotransformers.

Length

For length considerations, see the directivity factor text. It is not necessary, nor does it do any good, to go beyond 1-1/2 or 2 WL. By the time the antenna is that long, current is so low any addition length makes the pattern worse. I limit my 160-meter antennas to 800-feet, and use multiple antennas when a sharper pattern is required.

Directivity can actually decrease if a longwire-type array is made too long. This is true with Rhombics and Vee Beams, and it is also true with Beverages.

Zigzagging Wire

While a nice clear straight wire looks great, it does more to make us feel better than hear better! Minor ups and downs in height or dips or valleys don't really seem to have any noticeable impact.

Although it probably is a good idea to keep the wire as straight as possible, it is the overall direction and length that is most important because each small area contributes on a similar small portion to the overall directivity and signal reception.

Verticals and Baluns

[Home]

Related articles at

Balun Test contains model of "perfect" dipole currents.

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Receiving Common Mode Noise shows how lack of a balun can contribute to system noise (it applies to transmitting antennas as well)

Balun and Core selection for transformers and baluns

Transmitting baluns on testing transmitting baluns

RF in the Shack

Many people assume that RF in the shack or (worse yet) RF burns are tied to problems from poor station grounds. With properly operating coaxial feedlines or balanced feedlines, the operating position should have minimal RF even absent a shack ground. Only bringing a longwire or some other single wire feeder directly into the shack should cause RF burns or RF feedback in the shack.

There is one exception to the above. Direct radiation from the antenna into station wiring could cause high levels of RF to appear on equipment or wiring, but even in that case the station ground is not the problem or cure!

Contrary to popular belief that only dipoles need baluns, verticals and longwires can require baluns (more correctly called common-mode chokes or isolators in this application) also. This article shows why baluns (or common mode chokes) might be required, and what the balun can do.

What causes common-mode current?

Current flows because there is a voltage difference between two parts of a system along with a return path for the current flow. The path can be "special currents" without actual electrons flowing called

"displacement currents", or it can be actual charges moving through conductors. Displacement currents flow through the dielectric of a capacitor, between a vertical or single wire fed antenna and the "ground" for that antenna, or currents that flow between a mobile's antenna and a car body.

Displacement currents commonly complete the entire current path in antennas. They are the sole reason current in physically large coils can vary from end-to-end, and are especially problematic in mobile antenna installations. They are the reason an open-ended antenna like a dipole, longwire, or vertical is able to have current flow!

When we force charges up into a Marconi vertical or longwire antenna (making current flow), we have to move an equal number of charges out of some ground system or counterpoise into the feedpoint. The ground can be a single conductor or many dozens of wires, and it can (and often does) involve equipment in the house and/or the coaxial feedline shield. The bottom line is we always must have the same current coming back to the feedpoint as moves up into the antenna! There is no way around that rule.

This creates two problems:

1. **We may not be able to handle all the charge displacement with a few radials without having significant voltage driving those radials**
2. **The feedline has to connect to the antenna, the outside of the shield may be excited by this voltage, and become part of the return path for "collecting" displacement current**

Another way to view this is the feedline has to have something to *push against* to force current into the antenna. It is like pushing a car. If you have very poor footing, your feet will move and slide. The same is true for a ground system, as feedline power "forces" current up into the antenna the other terminal of the feedline has to be held steady.

The Feedline

If you are not familiar with how coaxial cables work, you might want to look at a simple explanation on this site or one of the ARRL Handbooks.

In order for a conductor like the outside of the shield to *not* have current flow at Radio Frequencies, it must have the same electrical potential and phase all along the length. If it has a high series impedance or if the potential difference along the conductor is low, very little current will flow. As seen in coaxial cable operational descriptions, any coaxial feedline can have unwanted common mode currents.

Does a vertical or longwire present common mode voltages to the feeder that can cause common mode currents? You bet! The only vertical (or longwire) that would not cause such problems is one with a very good or nearly perfect ground system, and that means something that looks like a large infinite groundplane. Even then, the cable must exit *below* that groundplane to be "shield current free".

Aren't Four Elevated Radials Perfect?

Here is a model of a groundplane with four radials:
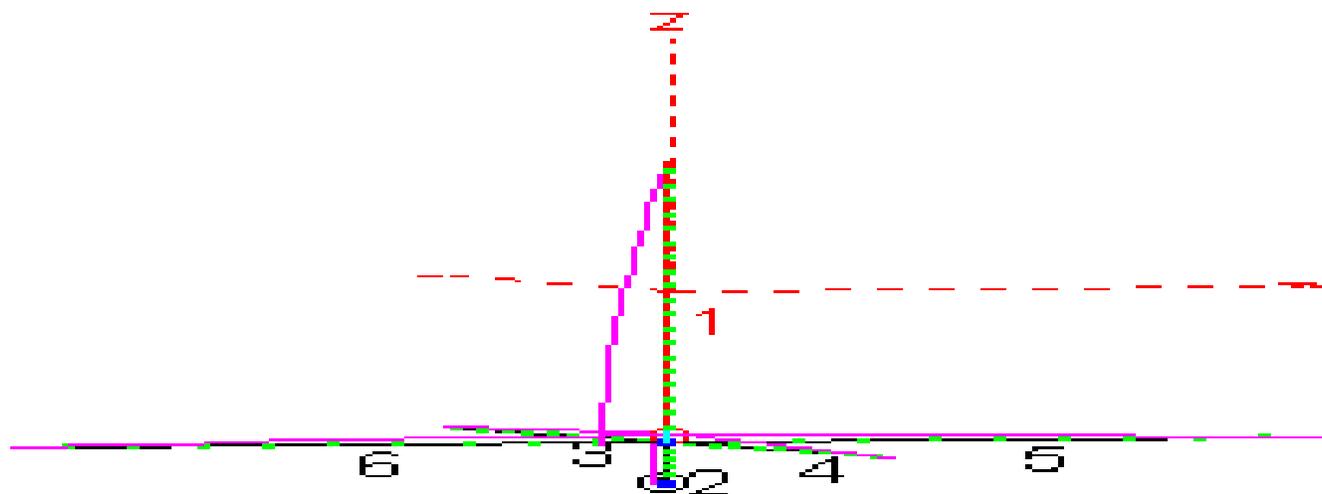
EZNEC ver. 3.0
Balun 80 vertical 1/3/04 7:19:05 PM
---------------- CURRENT DATA ---------------
Frequency = 3.6 MHz.
Wire No. 1:   6.700
Wire No. 2:   1.359 (This is your feedline or mast)
Wire No. 3:   1.985 (These are the radials)
Wire No. 4:   1.985
Wire No. 5:   1.985
Wire No. 6:   1.985

We can see significant current flows over wire 2, which would be the coax shield, a mast, or both.

There is a trick with Eznec. By inserting an additional source in the mast or feedline and setting current to zero, we can observe the radial common point to earth voltage required across a balun to force current to zero. In this case the voltage across the balun would be:

Source 2 Voltage = 145.5 V. at 67.97 deg.
Current = 0 A. at 0.0 deg.
Impedance is infinite

Amazing isn't it? At 1500 watts the ground common point for the radials actually wants to have ***145.5 volts to earth*** to prevent current flow and unbalance of the system!!! If we elevate the common point to 145.5 volts at 68 degrees phase angle, we now have the following currents at 1500 watts:

Wire No. 1: 6.4 A
Wire No. 2: 0 A (coax shield or mast)
Radials: 1.58 A each

Other Systems

With fewer radials the situation becomes much worse! (As a matter of fact, this is a good reason to use as many radials as we can even if the radials are resonant.)

Claims that four elevated radials form a "perfect ground better than 120 radials" are obvious nonsense! If it was a perfect ground, there would be no potential difference to earth and no common mode current flowing to "real ground" !

If the antenna has a high base impedance, it will have less current at the feed connection. (Sorry, but we cannot do this by using a folded unipole!)

Longwire and Windom antennas really aren't much different than verticals. They are a form of Marconi antennas, and require a counterpoise or ground of some sort. As with verticals, common mode current flowing into the antenna must be balanced by current flowing into a ground system.

Instead of bringing the longwire directly to an antenna tuner, a better solution is using an RF ground system independent of the station safety ground, and keeping that ground isolated from the station safety ground. That can be accomplished by adding a good heavy-duty 1:1 choke or current balun a few feet from the tuner, and connecting the RF ground to one output terminal and the antenna to the other terminal. With a two-wire feed Windom (really an OCF dipole), the two wire should connect to the choke balun.

The balun MUST be a current balun, rather than a voltage balun.

Summary

The cure for common mode problems caused by less-than-perfect grounds is inserting a 1:1 choke balun in the system at the antenna feedpoint. The coax should also be kept away from the radials as it exits the area of the radials and the antenna. An antenna with a poor ground using few radials cannot have a support mast grounded to the radial common point (at least it shouldn't if designed properly). There is no exception to this!

For comparisons of baluns see balun test.

Transmitting Baluns

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Testing Transmitting Baluns

One of the most popular Ham-lore rumors is a balun's performance can be tested or evaluated by grabbing the coax and watching for an SWR change. This is probably one of the worse test-rumors circulating!

An observable SWR change when touching, grounding, or altering length of a feedline (with the same Zo at the test meter) indicates severe common-mode current problems.

The opposite response, however, means nothing. Lack of SWR change does NOT prove the balun adequate and the system free from common-mode currents.

I was first exposed to this wild idea when asked to evaluate a problem with a commercially manufactured antenna. The antenna's design engineer decided a bead balun was adequate, based on this "grab and touch test". The antenna/balun combination passed the "grab test" with flying colors, even though the system had gross common-mode current levels. Despite extending the bead balun from the original foot to almost three-feet of beads, I couldn't remove common mode excitation of

the feedline by the grossly unbalanced antenna.

If we think about it, we would never consider it likely that our hand would have significantly lower impedance than the shield on a long length of coaxial cable! We also would never expect, just by dumb luck, to always grab the high impedance point of a random wire like the feedline.

A hand-grab-test will cause an SWR change only when the impedance of your hand is very low compared to the common-mode impedance of the cable at the point where you are "grabbing" the cable. The portion of impedance attributed to feedline radiation (compared to overall antenna impedance) must be a significant portion of antenna terminal impedance or the SWR will not change.

It is possible, if we alter antenna feed-cable lengths, to observe SWR changes as an indicator of common mode currents from poor system design. But such methods are never nearly as accurate as an absolute current measurement, and are actually only a little bit better than the useless "hand test". They also can give false answers if the feedline impedance is not the same as the SWR measurement device impedance.

An Accurate Test

The most simple and accurate way to test the effectiveness of a balun is to actually measure the common mode currents on the feedline with a suitable RF current meter that does not perturb the system. Such meters are inexpensive to purchase and are also easily manufactured at home. The meter would consist of a closed core that snaps around the cable, a terminated winding on that core, and a meter that measures RF voltage across that winding. The coil's load resistor should be low enough in value that the impedance of the cable shield is not perturbed.

By sliding the meter along the feedline, we could get a good idea of the common mode currents. The meter should be moved along at least 1/4wl of feedline. It normally would be sufficient to sample just four points along that distance.

Feedline radiation, like antenna radiation, is a function of the linear ampere-feet of conductor. The important distance is the spatial or linear distance, not the length of a coiled conductor. It is the charge acceleration over a linear distance of space that accounts for the radiation. A one foot box packed with 500 feet of cable is still just a one-foot antenna!

There are many cases where adding a balun will actually INCREASE common mode currents. Be sure you either fully understand the system or have modeled the feedline and antenna correctly.

Power Testing

A choke or current balun can be tested by inverting the ground on either end, and using a dummy load. Actual operation may produce more stress than this test, or less stress, but such a test gives a reasonable idea of the power rating under reasonable conditions.
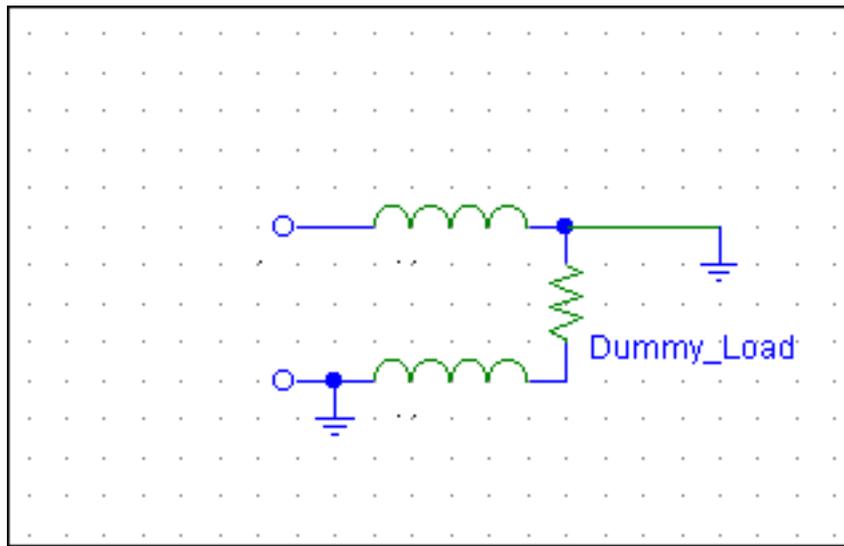

The balun should not overheat or change impedance with application of full power for the expected duty cycle in the expected maximum operating temperature (be sure to include sunlight heating).

Impedance Testing

Using the above test configuration, SWR can be measured. SWR should be flat, and should not change when the ground by the dummy load is moved from side-to-side. This should be true at full sustained operating power.

Testing Voltage Baluns

While voltage baluns are the least desirable types in most applications, they still have some useful applications. Using this configuration:
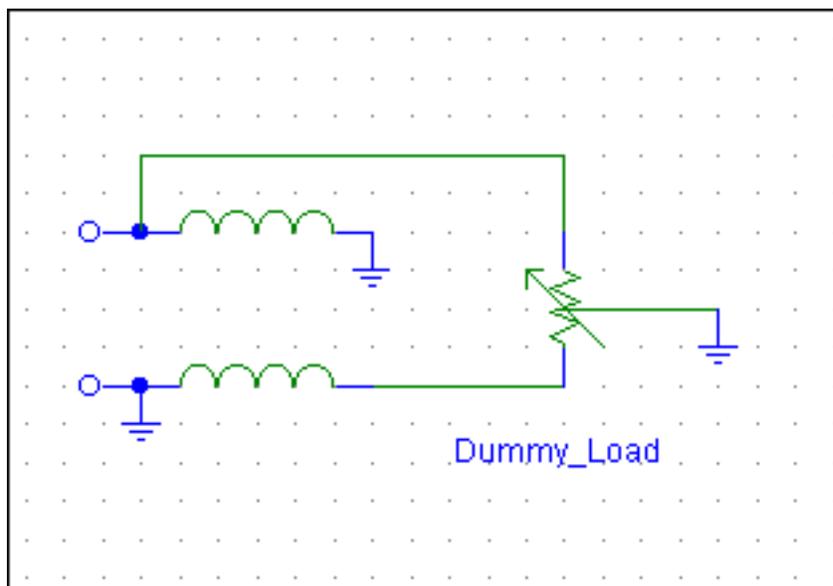
By moving the tap point along the dummy resistance, an idea of balance can be obtained. The balun should show perfect performance with the tap-point at the load center, and should show an SWR increase as the tap moves to the top or bottom.

Crossed Split Windings on Toroids

Some articles and handbooks show a Split winding method. This method is supposed to reduce winding capacitance by moving the ends of windings further apart. The proposed theory is by reducing shunt capacitance that "leaks RF around the balun", balun performance is enhanced.

This method very often hurts the performance! Check this article

toroid balun winding

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Toroid Chokes and Baluns

Commonly published information tells us winding a choke or 1:1 balun on a toroid with this special winding technique increases common mode impedance (choking isolation) of an 11-turn balun on a pair of #65 material ferrite cores:
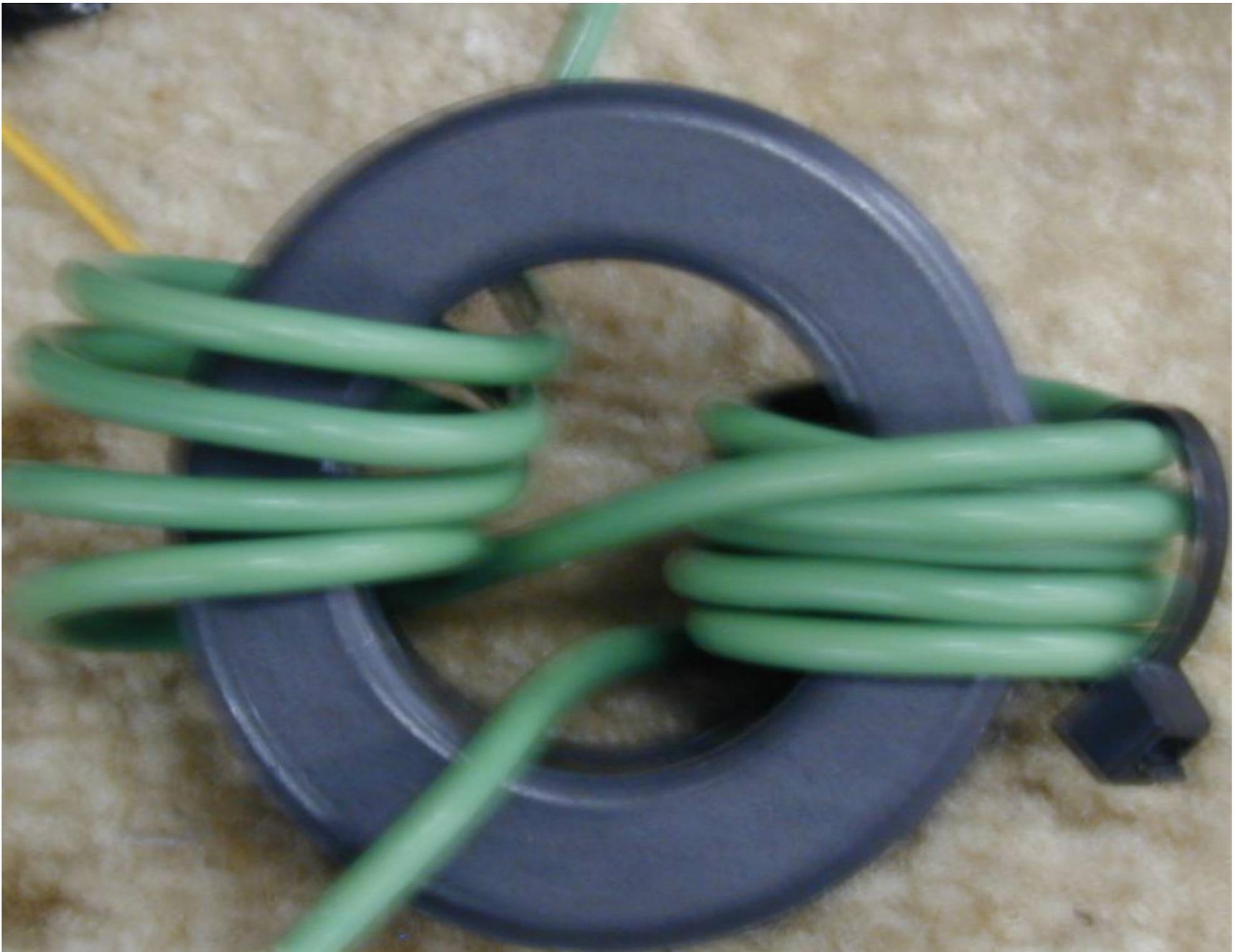
The common mode rejection sweep, assuming a 50-ohm common mode source and load is:

(pardon the camera jitter)
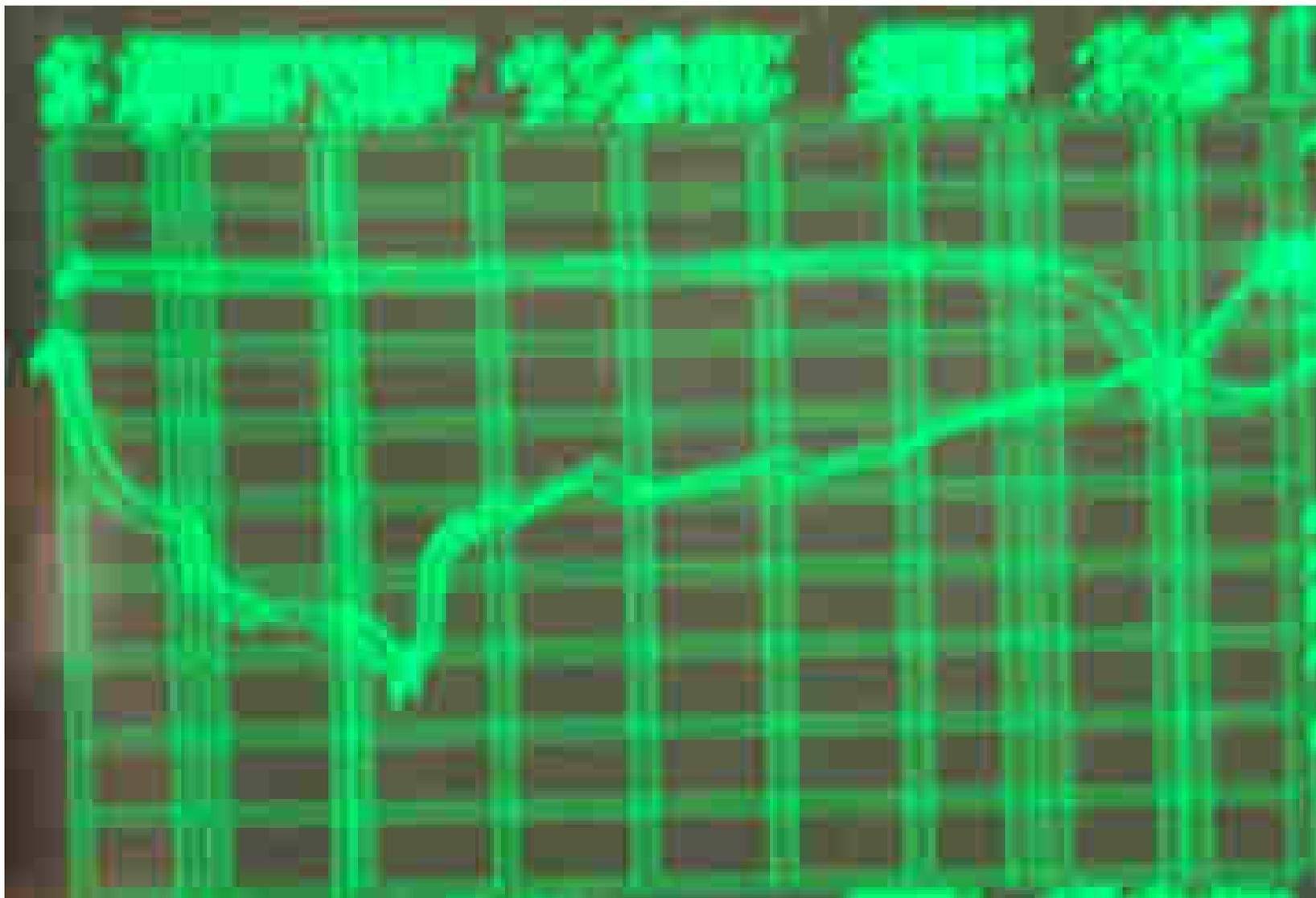
This sweep shows the following suppression using split winding method:

15.71 MHz	37.1MHz	134.8MHz	157 MHz	
-33dB	-46dB (peak performance)	-19dB	-3dB (minimum rejection)	

The same core was rewound using a normal parallel winding:

The sweep looked like this:
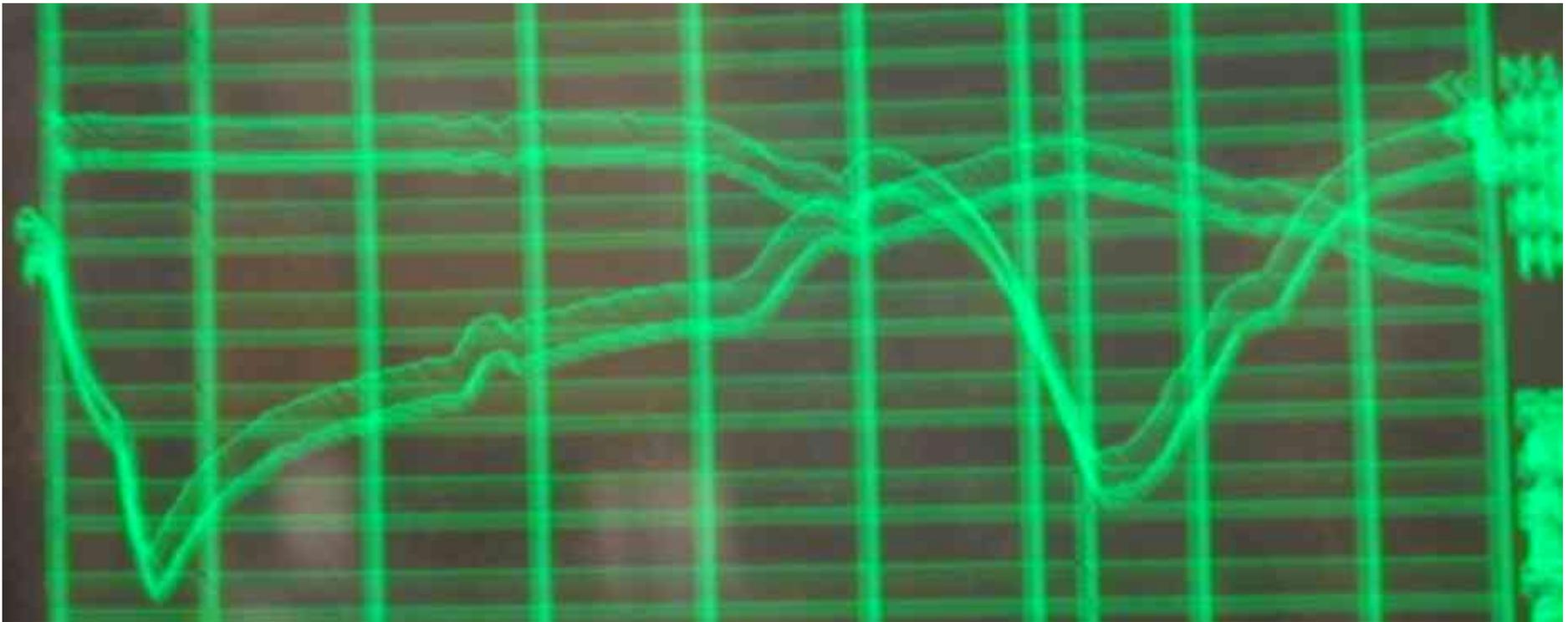
Common mode rejection, compared to the more complex winding, was:

Frequency	15.71 MHz	37.1MHz	134.8MHz	157 MHz
Split winding	-33dB	-46dB (peak performance)	-19dB	-3dB (minimum performance)
Normal winding	-47dB (peak performance)	-41dB	-4.6dB (minimum rejection)	-11dB

Conclusion

The ONLY change by using the more-difficult split winding is peak performance points moved higher in frequency. Performance in the HF range and lower actually decreased, so the extra effort resulting in a poorer HF balun.

This follows what many of us have been saying, and what anyone familiar with resonant circuits already understands intuitively. Common mode impedance increases with a slight amount of parallel capacitance is adding. The capacitance does NOT, as some claim, reduce performance by "leaking" RF around the choke.

The rule is simple. With any parallel L and C (neglecting losses), impedance always increases unless the parallel capacitive reactance (in ohms) is less than half the value of the inductance (in ohms).

With a choke winding impedance of 300j with an added parallel -150j capacitive reactance, the combined impedance would be -300j. The isolation would be the same! Any value of capacitance reactance higher than that (less capacitance value) would actually increase choke isolation!

There are many "improvements" we just blindly accept from reading articles, because on the surface they seem to make sense. This dispels one of the myths about baluns.

Another common myth is the widely accepted claim moving a choke balun to a lower common-mode impedance point of a circuit, such as the input of a tuner, improves balance or balun performance in the system. If you think baluns at the input of tuners are a good idea, go to this W7EL link and read the section on current baluns. You might be surprised!

Inductors and Loading Coil Current (Mobile and Loaded Antennas)

Related pages:

Mobile antenna FS comparisons   Loading Inductors   Inductor Spice Model

W7EL's Measurements and comments   Constructing an RF Current Meter

Much of the data below also applies to inductors in equipment, such as tank circuit.

If you arrived here from a link from K3BU, welcome to W8JI.com I hope you enjoy your visit. If you are not linked here from loading coil current articles on another web site, please skip the blue-text immediately below. It is meaningless.

If you arrived here from a link from K3BU, I want to apologize for the personal nature of that page and this response. Why is this response necessary? None of us like to see false or out of context claims or personal attack articles. Many of the comments from below were removed from context and quoted to mean something else in another internet article. It is important to read everything IN CONTEXT.

How Does an Inductor or Loading Coil Work?

The most basic answers are:

What does the coil do? A loading coil does not replace a missing fraction of a wavelength. The coil simply inserts a series inductive reactance that cancels capacitive antenna reactance. When a 150 ohm reactance inductor is inserted in series with a 150 ohm capacitive load (like an antenna), only the resistive parts remain.

What determines current distribution in a loading coil? The capacitance to the outside world and the impedance above the loading coil. The current in any inductor would be equal at each end except for displacement currents, which are "imaginary currents" that flow through capacitance.

How much difference is there in loading coil current entering the coil and loading coil current exiting the far end? If the antenna beyond the coil has a low self-impedance compared to the impedance of the shunting capacitance from the coil to "ground", the currents at each end of the coil will be essentially equal. It has NOTHING to do with electrical degrees the coil compensates for. In other words if the portion of antenna above or beyond the loading coil is long or has a large area compared to the physical size of the coil, current is essentially equal throughout the coil.

What does significant current taper in the loading coil indicate? It generally indicates a poor antenna design, where the loading coil (or any other application using an inductor) has high stray capacitance to other areas of the antenna system (like the groundplane) compared to the capacitive reactance of the antenna beyond the inductor. Significant current taper indicates a poor loading coil or poor antenna design.

click to view typical installation measurements

The Difference Between a Loading Coil and a Normal Inductor

There really isn't any difference, except the location where the coil is used. Mobile loading coils or loading coils for short antennas often have very high reactance. They have small amounts of capacitance at the end, and so stray capacitances are more of a concern. Stray capacitance from turn-to-

turn increases circulating currents and has the effect of increasing inductance and effective resistance at the expense of reducing bandwidth and Q. Stray capacitance to the outside world causes the coil to behave like an L-network, and transform impedances instead of providing a series reactance. This is why the optimum form factor of a coil becomes longer compared to the diameter with any inductor having very high reactance.

Inductors with low reactance are less critical of stray capacitances. Optimum form-factor in a low reactance inductor leans towards a short coil with the diameter nearly equaling length. In tank circuits or loading coils used with longer antennas or with capacitance hats, an optimum inductor is shorter and larger in diameter.

Most optimized inductors fall between 1:1 and 4:1 length to diameter ratios, the exact value depending on the terminating reactances. A coil is a coil, they behave the same way regardless if used as a loading coil or a tank inductor.

Common myths about inductor behavior:

One common myth is loading coil current is reduced as it passes through the coil. There are two reasons cited for this. One idea is the current is reduced because the loading coil replaces a certain amount of "electrical degrees" of antenna area, like the current taper in an antenna. The other idea is that series loss resistance causes a current reduction.

We often find inexperienced builders of 5/8th wl antennas think the "loading coil" needs to contain 1/8th wavelength of wire in order to make a "3/4wl antenna" and thus cause a low feed impedance. In other cases, some claim a half-wave of wire wound on a form causes a 180-degree delay, and is useful for phasing in a collinear array.

The basic flaw is the above ideas do not account for what happens in a coil with mutual coupling between turns. The flawed viewpoint is that current goes in one end, winds its way around through the physical length of wire in the coil, and after a time delay comes out the other end. There is a physical mechanism that prevents this, as we might intuitively think, from actually happening . The mechanism is the magnetic field in the coil!

What Really Happens

When current flows in the transmitter-end of the coil, a magnetic field is created. This magnetic field causes charges in the other turns to instantly move. This effect ripples through the coil at light-speed, just over 186,000 miles per second. As long as the magnetic flux coupling is high, the delay through the coil is the speed of light over the physical length of the coil. The electrical delay is the physical length of expressed the coil expressed in degrees at the operating frequency.

(Another interesting effect occurs. The increasing magnetic field sets up an "opposing voltage" as it cuts across conductors. This opposing voltage, created as the field expands, is what causes the current to rise slower than the applied voltage. If the exciting voltage is decreased the field collapses, and now the voltage changes polarity and aids current flow! If we don't allow the current to flow, the voltage will rise until it does. This is what causes the kick in a relay coil when we open the relay coil path, or the spark in an ignition when the points abruptly open.)

In an RF system, the physical size of the coil actually does add some "antenna effect". For example, on 160 meters the wavelength is about 550 feet. 1.5 feet is about one electrical degree. A skinny one foot tall coil, with negligible stray capacitance, would have about 0.67 electrical degrees phase delay. This delay occurs because to coil occupies a physical length of .67 degrees. Current at each end would be almost perfectly equal, the taper would be about what we would expect for a fractional-degree-long coil.

(*In the real world, all components have some stray capacitance and flux leakage, so they have a different amount of electrical length and current taper than the "negligible capacitance" case. In good coil designs, the capacitance and leakage is small and can be ignored. I'll show you measurements later to prove this.*)

Now let's look at an extreme case. If the entire antenna is "coil", like a helically wound antenna with no top hat or stinger, current would be reduced to nearly zero at the open end. This is because distributed capacitance over the length of the antenna is fairly high, the shunting capacitance has a low impedance compared to impedance at the end of the antenna, and current is diverted to ground in the form of displacement currents.

Compact loading coils are another matter. In many cases phase delay is negligible or immeasurable by normal methods...flux coupling is nearly perfect. A good example would be a relatively compact toroid or a compact nearly-square L/D ratio loading inductor. I've found it impossible to measure the current taper in a toroid and very difficult to measure in a compact air-core loading coil. (The opposite extreme would be a perfectly straight wire with no folds or bends or the helical antenna described above.)

In the case of the toroid or compact coil, the behavior would be such that doubling the turns nearly perfectly quadruples inductance. If we doubled turns and inductance simply doubled or increased at a much faster rate, we should know the coil is in a mode other than a pure inductor mode. This is a *strong* indicator inductor operating Q is less than optimum, and the inductor might behave less than ideally in critical applications.

As a matter of fact, observing inductance change while adding turns can be an excellent test for flaws or shortfalls in system design. A linear increase in inductance when adding conductor length indicates design problems.

A perfect impedance squaring effect indicates minimal electrical phase-delay, or "antenna length" of an inductor. Impedance squaring as turns are doubled indicates the undesired inductor stray capacitance has a high reactance compared to the antenna system beyond the loading coil. Of course there can be exceptions, but it is a good general rule that large current taper indicates the loading system is much less efficient than necessary.

Making a Delay Line

It's certainly possible to make a delay line from a coil without opposing flux, but doing so requires stray capacitive reactance to be significant compared to the value of distributed inductance in the coil. This would occur in a very long helice, a very large diameter helice or loop, or an inductor near a large metal counterpoise or ground plane.

It's important to remember unless a coil is "stretched out" or "expanded" a great deal, the phase delay will not even be close to the physical conductor length. (The exception could be if you had so much capacitance the inductor acted like a series connected string of L/C/L networks as shown below).
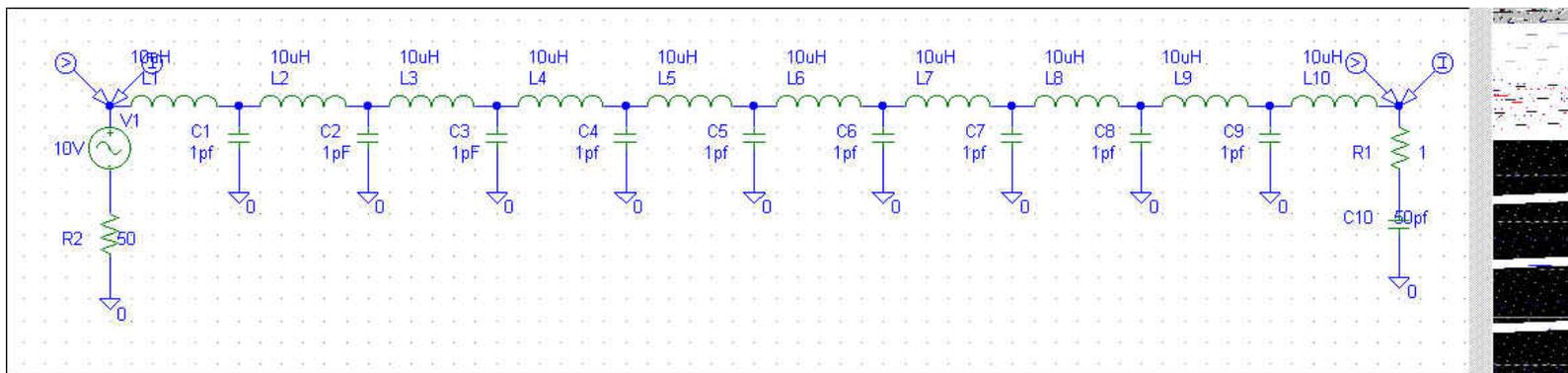
In any case while this effect might be good in a collinear antenna or plate choke (assuming you do it right) it is a BAD effect in a short loaded antenna!

Inductor E/I Phase shift

An inductor delays the flow of current in relationship to applied voltages as the magnetic field inside the coil expands. Voltage increases before current starts to flow. This phase relationship between voltage and current is often confused with time-delay phase in the inductor. Say we have this simple circuit:
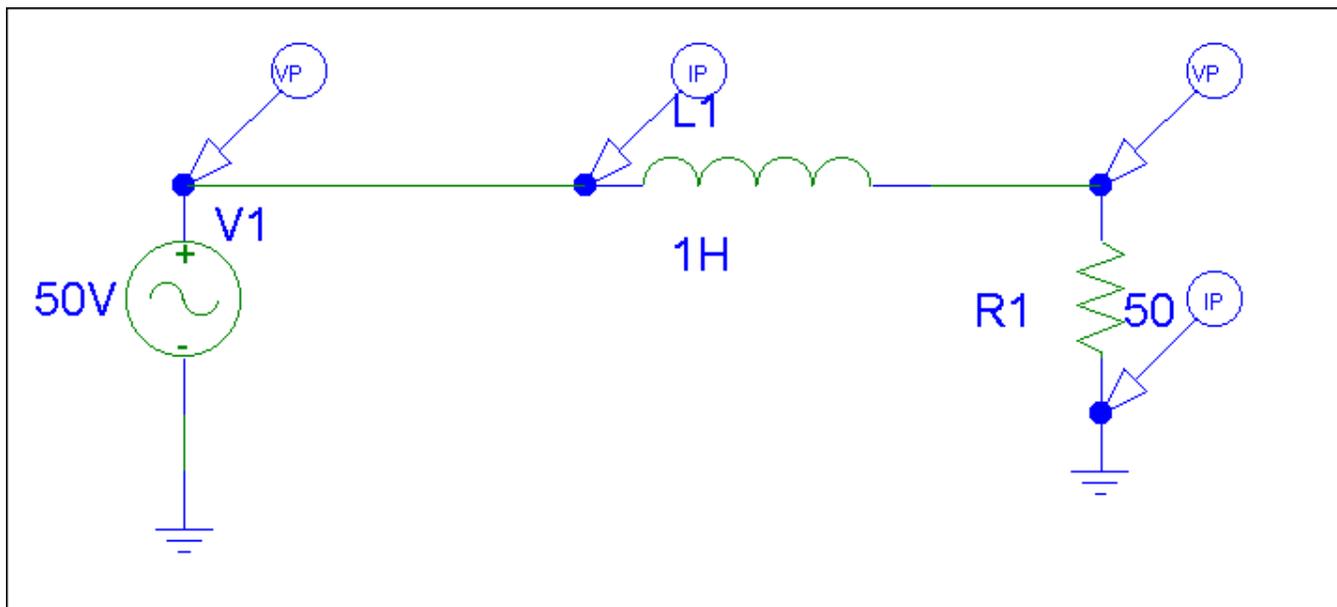
Current and voltage at V1 will be out-of-phase by the effect of L1 "charging" with magnetic flux. Current appears AFTER voltage rises, and falls after voltage falls. Current in R1, however, is exactly in phase with voltage across R1. That's because the voltage across R1 is always E=I*R.

Every component must follow the Laws or electrical rules established for that component.

The current in R1 is delayed from VOLTAGE rise in V1 by the voltage to current phase delay of L1. This does result in a time delay in relationship to voltage rise at V1, but there is NO current time delay through L1! V1, L1, and R1 all have the same peak current at the same time!!!

The notion that antenna loading coil delay current by the same time as they delay response to increased voltage is obviously nonsense.

Here is a graph of phase delays in the above system:
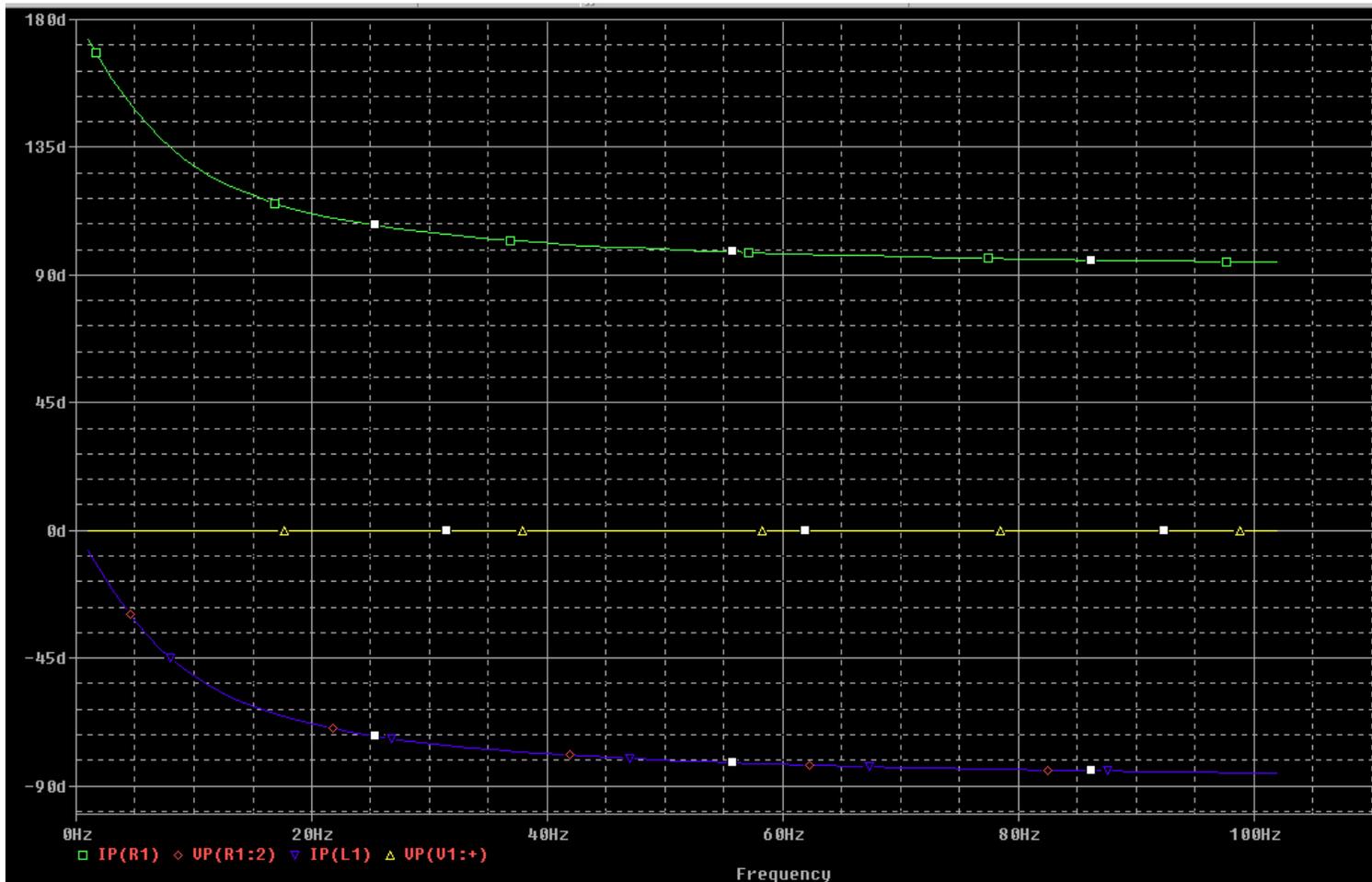
Current in the load, generator, and inductor all exactly track in the same relationship from dc up. There is no "phase delay". The generator voltage is a straight line different than current, and this indicates the generator sees a "reactive load".

The Misplaced Notion

Proponents of the idea that coils replace "antenna length" so far have been unable to define a set of rules or logical reasons why a current reduction and electrical-degree phase delay related to antenna area "replaced" would occur in a two-terminal component. While a long inductor with poor flux linkage from end-to-end or an inductor with low values of stray capacitive reactance to a groundplane compared to series impedance can cause SOME current inequalities or phase delays, the amount is normally immeasurable with normal thermal current meters with a reasonable coil form factor and termination above the coil. The amount of current taper actually rivals the disturbance of the system by adding the measurement device, unless we are very careful in how we construct the measurement device.

Both W7EL and myself made independent measurements that show the fact the coil or inductor is in an antenna has nothing to do with current distribution in the coil. Some people have actually incorrectly reported W7EL's data! Here's what he had to post on rec.radio.amateur.antenna to correct misrepresented claims.

The Need for a Measurement

An article on E-ham claimed measurements proved a new concept about loading coil current. The E-ham article put forth an idea that current disappears as it moves through a loading inductor without a mechanism like displacement currents providing a path. This claim conflicts with established component behavior, so it would indeed be fascinating if it were true! One of the claims supporting the idea that coils in antenna work differently than coils in circuits was that a non-radiating toroid loading inductor showed a current taper when used in an antenna.

I recently constructed a calibrated current meter that slips over whip antennas and masts, and is for all practical purposes totally immune to variations in voltage in the system. It also is mostly plastic, and has minimal effect on stray capacitance of the antenna. The resonant frequency and currents are not significantly perturbed by measurements with this meter. When I added a similar meter used in the other tests, resonant frequency shifted significantly! This is a sure sign the meter's capacitance or inductance is affecting the system.

In late December 2003 and early January 2004, I made additional measurements of loading inductor currents. The results clearly agree with the analysis that had been presented on this page since early 2003. Without displacement currents, currents into and out of a loading inductor *are equal.* That is a hard rule, it agrees with theories defined by people much smarter than me, and I believe it is unbendable unless the works of Faraday, Maxwell, Ohm, and Kirchoff were incorrect.

A sample of measurements above and below the loading coil with various antenna above the coil (current as percent of reference) follow:

	Toroid with hat	Small 2x2" coil with 24" hat up 24"	Long 12x3" coil with 24" hat up 24"	Long 12x3"coil with 6' whip
Current below	100%	100	100	100
Current above	100	94.4	73	76%

	Toroid with whip	long 12x1.5" coil with 6' whip	long 12x1.5" coil with 24" hat	Small 2x2" coil with 6' whip
Current below	100%	100	100	100
Current above coil	100%	79%	75%	96%
Current in whip 1ft above top of coil		73%		

The most revealing thing was how noticeably small changes in stray capacitance near the middle and top of the loading coil affect current distribution. It was quite evident hanging a large meter on each end of the coil would greatly perturb the system.

Clearly we do NOT want:

- **A large hat just above a large coil**
- **A long large coil and a short whip**
- **A coil near large sheetmetal**

More data along with photographs will appear on a new page over the next month or so. Until then, I can assure everyone the conventional theories presented below are accurate, and the theory that "electrical length" the coil "replaces" is incorrect. Loading coils indeed behave like any other inductor in the world.

Independent measurements by a reliable engineer have agreed with my measurements above. Anyone doubting my data need only read the following e-mail from W7EL.

On to the old text in this page:

The Incorrect Assumption

Another commonly misconception is, since voltage increases at the far end of the loading coil, current must logically decrease. After all, we have a fixed amount of power and voltage has increased. The assumption is:

1.) We multiply voltage times current to get power.

2.) If voltage increases current must decrease.

Unfortunately, this is **not** correct in reactive systems! Simple P (power) = I (current) times E (voltage) only works when the system is non-reactive. This condition only occurs at resonance, and only *below* the loading coil at the antenna feedpoint!

In a reactive system, like in a mobile whip above a loading coil, voltage and current are no longer in phase. As a matter of fact, voltage and current can closely approach being 90 degrees out-of-phase when the whip is electrically very short. Since the antenna area above the loading coil is highly reactive (voltage is not in phase with current), we can not multiply voltage times current without considering phase differences.

You may have heard the term "reactive power" or VAR (volt-amperes-reactive). Reactive power is voltage times current without consideration of phase angle. We can have kilowatts of VAR power with only a low power transmitter, and that is what we actually have in the reactive part of the small antenna.

Coil Q and Changes in Efficiency

Current taper or reduction has been cited as a reason coil "Q" has little effect on signal level in mobile systems. Speculation is only the first few turns of the loading coil carry significant current because the coil "leaks" magnetic fields and radiates, and this is why the coil Q has little effect.

Another idea proposes the loading coil "makes up" a certain missing part of the antenna. It goes on to conclude the loading coil can be accounted for in "electrical degrees", making up the "missing difference" in antenna degrees. This isn't true either. The inductor doesn't know where it is and suddenly change from "x" ohms reactance to electrical degrees! It responds to AC currents and voltages as any inductor in any circuit does. It doesn't suddenly change measurement units.

As an example of this, try to define a 45-degree electrical length inductor at 1.8 MHz. That would mean it is a capacitor at 3.61MHz, where it is over 90-degrees long! How many turns at what length and diameter is a 45-degree inductor?? Where is a formula that allows converting a given size inductor to electrical degrees? This shows how useless and meaningless that definition is!

The inductor adds a certain amount of series reactance, that's all. A 300uH inductor is not 20-degrees long, nor is it 80-degrees long, so far as radiation goes unless it is really that long **physically**. It is a certain number of ohms reactance at a certain frequency, or a certain number of units called Henries. It is not "electrical degrees" that it adds, it is a non-dissipative reactance (in combination with a loss resistance because of finite quality) at a certain frequency!

A loading inductor can "insert" a large amount of phase shift, but the phase shift is between voltage and current. The only exception to this would be if the inductor had considerable distributed shunt capacitance to the outside world, and acted like a string of series inductors (with the antenna) and shunt capacitors (shunting to the ground system). In that case we could expect coil Q to be extremely low, since it would be the electrical equivalent of a lossy transmission line. That's either an awful loading coil, or it is a less-efficient helical loaded antenna!

The Correct View

Another group of people don't argue against established and proven circuit theory. They understand charges flowing into one end of the loading coil must have someplace to branch off (a virtual third terminal), or they must flow out the other end. Without that additional "virtual" path, charges flowing into the coil would always equal charges flowing out. This is true regardless of radiation, losses, or induction fields.

This makes perfect sense when we think of any dc circuit, antenna, or RF system. Electrical rules are satisfied, the system behaves as it does in the real world.

There is very little change in current, unless the coil is physically very long compared to the rest of the antenna above the coil or unless the coil is laid right against "grounded" conductors and the whip above the coil is very short. This fits perfectly with helical verticals, where the coil is "stretched out" over the length of the antenna.

It also agrees with base loaded antennas, which have nearly as much current into the antenna above the coil as at the feedpoint. It agrees with center loaded antennas, where current below the coil is essentially uniform and the whip above has triangular distribution.

Current *can* be different in various areas of an inductor, but *only* if shunting capacitances (impedances) to the outside world are significant compared to load capacitance (impedance). Another condition where current can vary substantially is with operation near the condition of self-resonance in what is normally considered or defined as a "series-resonant" mode. This would be a very poor and inefficient loading inductor, such as when a 160-meter antenna is used at a secondary resonant frequency in upper HF.

Circuit Model of a Mobile Antenna
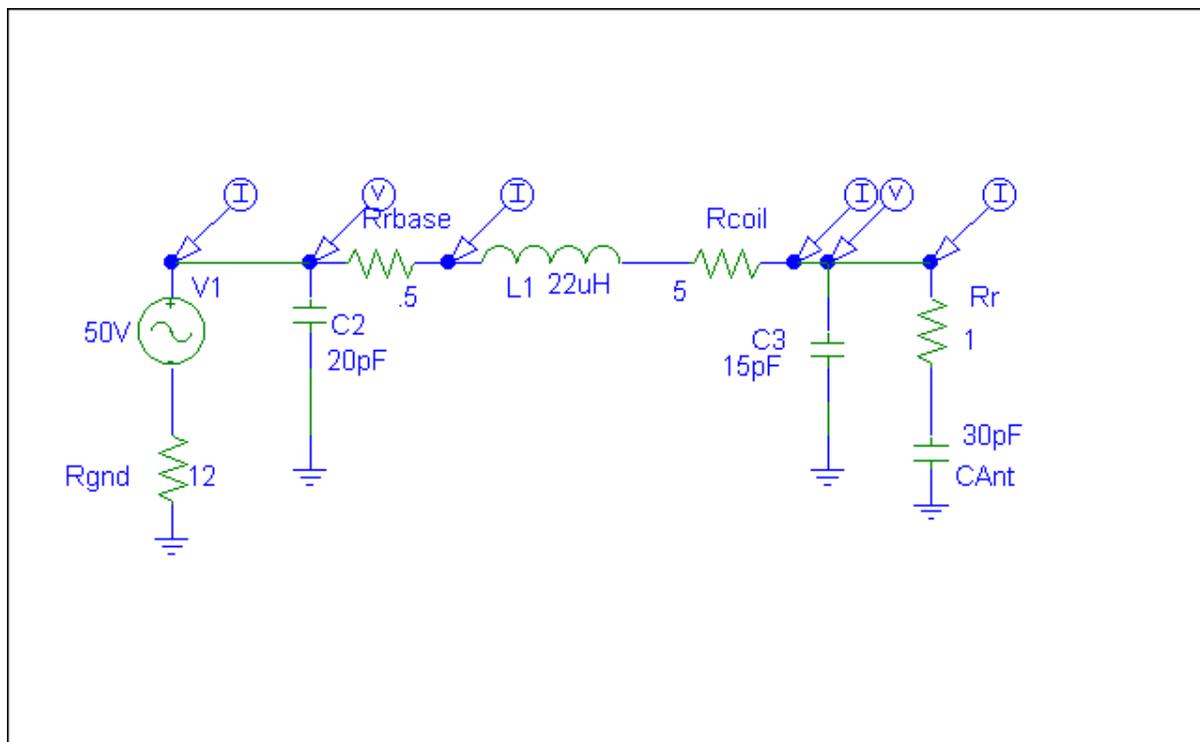
The above model shows what might be a typical mobile antenna installation.

1. Rgnd ground resistance of vehicle normalized to feedpoint
2. V1 coaxial feedline
3. C2 base capacitance
4. Rrbase [Radiation resistance](#) of the base area of the antenna
5. L1 loading coil
6. Rcoil coil equivalent series loss resistance
7. C3 coil shunt C to ground
8. Rr top area [radiation resistance](#)
9. Cant equivalent antenna capacitance above coil area

My 160 Mobile Antenna

I've worked all continents except Africa while 160-meter mobile. I have CW contacts at over 10,000 miles, and SSB as far as Europe (4000 miles). My mobile antenna consists of an eight- foot antenna with a six-foot hat (made from surplus Ford or Chrysler car antennas) at the top. The loading coil is at 5 feet.

This antenna has been on the truck for thousands of miles, without mechanical failure. It is mounted at the upper left corner of the truck bed, about one foot back from the cab.

The following is a model of the current antenna system on my Ford F-250 HD long bed super cab truck:

Mobile antennas, short verticals, loading coil loss,and loading coil current

EZNEC
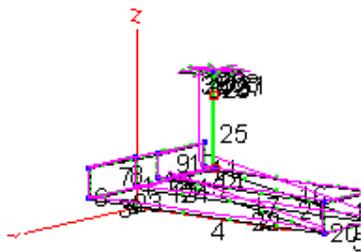
The base impedance in the model is:

Frequency = 1.854 MHz.
Source 1 Voltage = 24 V. at 2.35 deg.
Current = 1 A. at 0.0 deg.
Impedance = 23.98 + J 0.9853 ohms
Power = 23.98 watts
SWR (50 ohm system) = 2.087

Actual measurement at my Johnstonville, GA farm in open flat pastures on August 17 at 8AM. Wet soil 25.8 ohms 0j base impedance, pretty close agreement to EZNEC model and earlier data! (I did have to adjust the model for very low ground conductivity, otherwise the resistance was far too low. It's my belief that NEC-2 underestimates ground losses in small radial or counterpoise systems that are close to earth.) Earlier text shows a base resistance of *28 ohms*, that was dry soil with a slightly different loading coil and antenna.

The modeled current distribution for 1-ampere applied at the base (in 1-foot intervals) is:

1ft= 1.0031

2 ft= 1.0091

3ft= 1.0178

4ft= 1.0318

<Coil>

5ft= 1.0175

6ft= .97512

7ft= .92984

8ft = .89522

Measuring the current into and out of the loading coil with a small thermocouple RF meter, I detect no difference This is in close agreement with the model.

The efficiency of this antenna knowing coil Q, radiation resistance, and base resistance calculates just under 1 percent. The model indicates about 1/3 percent efficiency. This is reasonably close.

Removing the hat (in the model only) shows the following changes:

1ft = 1.0043
2ft = 1.0133
3 ft= 1.0279
4 ft= 1.0566

<coil>

5 ft= .95508
6 ft= .72232
7 ft= .27813
8 ft = open

I haven't tested the above, but with the same loading coil loss resistance the model says efficiency is now around 3dB worse. Removing the hat, with NO change in coil resistance, shows nearly loss nearly doubles. Of course the coil resistance would increase, because the loading coil nearly quadruples in size. Bandwidth is less and efficiency is less, even if I could maintain the same coil resistance.

Examples of Unequal Current

In the above models, we see that current into and out of the one-foot long coil is about the same. There is only about *2% change in current* even though the coil occupies 12% of antenna length in the "hat-loaded" antenna, but in fairness I couldn't resolve that change with a reasonably good RF current meter.

The model predicts *10% change in a non-hat antenna*, but I never measured that antenna to confirm it.

Clearly there is no basis to the claim current is high only in the first few turns of an inductor, or that current tapers in relationship to "electrical degrees". The most accurate way to state the effect would be to say: "When the loading coil is short and the capacitance of the antenna beyond the coil is reasonable (in this case 3000 ohms Xc or less), there is an immeasurable reduction in current in the coil. When the required loading reactance is very high (in this case 8000 ohms), the reduction in current is about what we would expect for an equivalent length of antenna replacing the coil."

Degrees Vs Radiation Resistance

This upper four feet of this antenna resonates near 24 MHz with the hat. We can assume it is 90 degrees long at 24 MHz, which would translate to 6.9 degrees on 1.85 MHz. Following that same logic, this would mean the loading coil would be about 83 degrees long electrically. Using the incorrect logic proposed by others where the loading coil "makes up the difference in electrical degrees", there would be almost no current past the loading coil. Obviously this is not the case, the loading coil has very little "electrical length". As a matter of fact, the electrical length is about equivalent to the physical length!

This goes back to radiation theory, and my favorite saying: "Five hundred feet of wire in a one foot long tube is still one foot of antenna". Some CB manufactures sell antennas to consumers with the claim they use 5/8 or 3/4 wavelength of wire in an eight-foot fiberglass whip, so the antenna has more gain. Obviously this is not true. Let's not let such silly claims spread into amateur radio!

Related topics:

[Inductors](#)

The spice inductor model shows one example of how unequal current is created. The model demonstrates a coil having significant distributed capacitance to the point of current return in the system compared to terminating impedance of the coil. In a monopole this return path would be to the groundplane, or anything closer to the potential of the groundplane than the area above the loading coil's position in the antenna system.

Another Practical Antenna Example

Let's assume we have a lossless 15.3 foot long 0.2 inch diameter conductor over a perfect groundplane. Eznec gives the 1.821 MHz base impedance as .3004 -2169j. In other words, the antenna "looks like" .3004 ohms of load resistance in series with 40.32pF on 1821kHz. The return path for current is through the .3004 ohm resistance and 40.32pF capacitance, back to the ground of the antenna (it is a Marconi antenna).

Such a termination (load) would require a series inductance of 2169j (189.57μH) to cancel feedpoint capacitive reactance. A typical 190μH inductor

would be rather large, requiring somewhere around 53 turns when using a 4" by 4" form factor. One would expect a physically large inductor to have noticeable but very small displacement currents to the groundplane, when the small stray coil capacitance is compared to the 40.32pF termination capacitance. This raises two very important design guidelines:

- *When installing a loading coil of substantial inductance in an electrically short antenna, sheetmetal and dielectrics should be kept away from the coil and areas of antenna above the loading coil. This would include dielectrics on or near the inductor, since the presence of dielectrics would increase undesirable capacitance.*
- *When inductive reactance requirements are large, as when short thin "stingers" without hats are used above a coil, the coil form factor should lean more towards long and thin. Capacitances near the open end of the coil (high voltage end) should be minimized. This would be true even when the coil length increase results in a small reduction in mutual turns coupling, since the stray capacitance may result in a larger loss penalty than the slight increase in accumulated resistance from additional wire length.*

Efficiency

Efficiency in any antenna near earth is almost always dominated by ground related losses, short-height Marconi antennas are no exception. The overall effect of loading inductor Q and matching system losses are "diluted" or "swamped-out" by ground losses. Ground losses cause most systems to have greatly reduced sensitivity to inductor design.

The only consistently predictable factor in efficiency in fractional wavelength Marconi antennas with limited size ground systems is radiation resistance. Efficiency increases almost directly in proportion to radiation resistance.

Radiation Resistance and Power Radiated

Radiation resistance is probably the most poorly defined term used with antennas. The lack of clear definition creates errors and misjudgments when predicting antenna performance. If you wish more detailed information, this page contains information on radiation resistance. For the purposes of this discussion and to avoid pitfalls associated with using feedpoint impedance as radiation resistance, I'll use the same definitions Jasik, Balmain, and others have used. This definition is based on the IRE definition of radiation resistance being equal to the net or effective current causing radiation squared divided by the power radiated as EM energy, or $Rr=Pr/I^2$.

Using this definition, a folded dipole has a radiation resistance identical to a conventional dipole of the same physical dimensions (~70 ohms).

Radiation is caused by charge acceleration, there is no magic. The only thing affecting radiation resistance in a short vertical antenna near ground is current distribution over the linear area occupied by the radiation portion of the antenna. The general rules are:

Radiation resistance of a Marconi vertical in the maximum possible radiation resistance case for a given height (this is the case where current is uniform throughout the structure) is equal to $1580*(H/L)^2$ where H equals height and L equals wavelength and both are expressed in the same units. Using degrees, we see a 10-degree tall antenna has a maximum possible radiation resistance of $1580*(10/360)^2$ or $1580*.000772 = 1.22$ ohms. This would apply even if the antenna is a vertical, DDRR, Fractal, or folded unipole with considerable top loading.

If current is triangular, radiation resistance would decrease by a factor of four to 0.305 ohms.

Power radiated is given by I^2*Rr

With 100-watts applied to a 10-degree tall antenna, net current in a lossless antenna with uniform current distribution would be 9.05 amperes. With triangular distribution, such as appears in a small diameter short base loaded whip, current would be approximately 18.1 amperes. We are in serious problems if the inductor reduces current along its length, since the only possible way to radiate 100 watts would be to have somewhere around 9 amperes of effective current integrated over the 10-degree vertical area of space for the radiator!

Ground Losses

All current flowing (or displaced) vertically into the antenna must equal current flowing out of the ground or counterpoise system. Even though ground losses are distributed losses, we must normalize all losses to the feedpoint in order to compare systems. There are cases where this will not always occur, causing us to falsely assume we have lower losses than really exist.

In this tutorial and comparison, I have normalized ground losses to the same point where radiation resistance is considered.

System Losses

Mobile antennas, short verticals, loading coil loss,and loading coil current

(Measured data below of actual antenna given below was from 1995 data taken at a different location near Atlanta with a slightly different loading coil and antenna. There is a slight disagreement with current data. I left this all in so you can see the departure from measurements and models using 8 year old data.)

Base Loaded (Triangular Antenna Current Distribution) with no ground loss

Assuming we have a base-loaded antenna, and the operating frequency has a wavelength of 550 feet (around the 160-meter band), a 15.3 foot vertical would fit the above 10-degree value. Interestingly enough when we compare Eznec to formulas available in older (1950 vintage) engineering textbooks, we find radiation resistance predicted by Eznec is .3003 ohms while the triangular current estimate for the same height radiator is .305 ohms! This is an amazing degree of agreement, illustrating what we could do before modeling programs became available. (With perfect top loading, both Eznec and longhand calculations show approximately 1.2 ohms of radiation resistance.)

Assuming our 15.3 foot tall (10-degree) base-loaded antenna uses a coil Q of 200, the coil has 10.845 ohms of ESR. Total resistance with a perfect ground would be 10.85+.3= 11.15 ohms. Current into this system with 100 watts applied would be around 3 amperes, resulting in ~2.7 watts radiated and ~97.3 watts lost as heat in the inductor.

Doubling coil Q (400) would provide 5.73 ohms of base resistance with 4.18 amperes. Power radiated would be 5.2 watts, power lost as heat would be 94.8 watts. Efficiency does not quite double, changing from 2.7 to 5.2%. This results in a 2.8dB change in signal level.

Top Loaded (with no ground loss)

If we added a four-wire hat with 15-foot wires, current would no longer be triangular. While we wouldn't quite reach the optimum uniform distribution, current at the top would be about 78% of current at the antenna base. Feedpoint impedance would become 0.97 -551j, and the antenna would look like 0.97 ohms in series with 159pF.

Using a coil Q of 200, we would now have 2.76 ohms of inductor loss. Current becomes 5.18 amperes. Radiated power is 26 watts, while power lost as heat becomes 74 watts. Even in the perfect ground case, the change in efficiency caused by top loading is large. Top loading (with only the hat) results in 9.8 dB change in signal level when compared to the base loaded case when coil Q remains 200. Efficiency is 26%. The coil remains at ground level for easy matching and frequency change.

In this case current at each terminal of the loading coil would be essentially the same regardless of poor coil mounting techniques. In order to have significant current taper in the coil or in the bottom of the mast, shunt capacitance would have to be a significant compared to 160pF. The antenna's high input capacitance relaxes inductor and antenna mounting electrical requirements.

Base Loaded (high ground loss)

My F-250HD Super Cab pickup truck, when parked over open medium quality pasture land, has a ground resistance of about 20 ohms (normalized to the feedpoint) on 160 meters. Applying this ground loss to the base loaded antenna, the system has a feedpoint resistance of 20+.3=20.3 ohms. (This is reasonably close to actual feedpoint resistances measured with a similar operating antenna.) Adding coil losses, the system has 20.3+10.85=31.15 ohms. (**NOTE: Current coil is ~8 ohms ESR, 10.85 ohms is from ~8 year old data**) Current is sqrt (100/31.15) or 1.79 amperes.

This results in .96 watts radiated, and 99.04 watts lost as heat. Efficiency is now around .96%.

Substitution of a coil with a Q of 400 results in 25.7 ohms feed resistance, or 1.97 amperes antenna current at 100 watts. In this case efficiency is now 1.16% for 1.16 watts radiated. The change caused by doubling coil Q with high system ground losses is about 0.8dB, compared to almost 3dB in the perfect ground case! With a poor ground (in this case typical of a very large vehicle), a large change in coil Q produces little change in system efficiency.

Another Top Loaded (high ground loss) System Example (made prior to the EZNEC model above)

Using a large hat isn't practical in a moving mobile, although it could apply to fixed stations suffering with poor ground systems. When the hat is smaller, such as a mobile requires, the loading inductor can be moved higher in the system. Such a move would produce uniform current below the loading coil, with a current shape above the coil dictated by the construction of the upper portion of the antenna. My own mobile uses a six-foot diameter hat manufactured from stainless steel automobile antennas arranged in a spoke. I have no problems with wind or occasional obstructions. While unsightly, a modest hat is workable.

In order to keep the systems comparable I'll use the same radiation resistance provided by a large hat, but intentionally add high ground loss as a lumped resistance. This model ignores field losses near the antenna.

In this case we have 0.97 -551j as the inductor termination presented by the antenna. With ground losses normalized at 20 ohms and an inductor Q of

200, we have 20+2.76+.97 = 23.73 ohms of feedpoint resistance. Current is 2.05 amperes, and power radiated is 4.1 watts. Power lost is 95.9 watts.

Efficiency is 4.1%, a 6.3dB increase over a base-loaded triangular current system with the same lossy ground. This system is 8dB down from the same "top-loaded" distribution using a perfect ground.

When the system has significant fixed losses, increasing radiation resistance four times by top loading provides a similar dividend in system efficiency. At the same time a substantial increase in coil Q provides only minimal change in field strength.

Current Through Coil

Related pages:

Inductor spice model

There has been some speculation that current is high only in the first few turns of a loading inductor. Radiation comes solely from charge acceleration or current over spatial (in line) distance.

If any loading inductor shows substantial decrease in current over the length of the inductor, it is an absolute certainty that the inductor is poorly designed and that the system above the loading inductor is not contributing to system efficiency. The reason for this is very simple and straight forward. Any two-terminal component (even considering wire as a "component" applies) MUST have equal charges flowing into and out of each terminal. Voltages to other reference points can be different, but for every charge moving into one terminal a like number of charges MUST move out of the other terminal. Radiation, induction fields, and loss resistances have no influence on this rule.

In order to have any change in current, there must be an additional path or paths for charges. This path can be through leakage resistances, or through fictitious currents called displacement currents. Whatever the path, the total charge movements must be reconcilable. We simply can not have current "disappear".

The normal path upsetting "unbalancing" current into and out of each terminal in an inductor is provided by displacement currents through electric fields. As with any system, the amount of current flow is proportional to potential difference and impedance of the path. In order to shunt a substantial current out of an inductor, the potential difference between the ends of the path has to be high compared to the impedance of the path. The impedance of the stray path must also be reasonably low compared to the normal desired path.

Current diversion is problematic in very large inductors operated at (or very near) internal self resonance, when the self resonance is what we typically refer to as a "series-resonant" condition. This condition is common in plate chokes used in vacuum tube power amplifiers, where the system operates over many octaves of frequency range.

"Series resonances" inside components occur when distributed inductance forms a pair (or multiples of pairs) of "L" networks. The large series inductance from each end of a winding reacts with the small stray capacitance at the center, and forms a very high impedance transformation L network. The electrical potential at the center of the system becomes extremely high, and even the smallest amount of capacitance to surrounding objects will carry a substantial displacement current. The large displacement currents cause the terminal impedances to drop, and allow considerable current to concentrate in small areas of the component. At the same time, considerable voltage can be present. The normal result is arcing or destruction of the component, or failure of the system depending on the choke to operate.

Series resonance always occurs at a frequency higher than the self *parallel* resonant frequency of the component. A loading coil operating under such conditions would be required to have serious design errors to fall into this category, since the end termination capacitances should always be substantially higher than stray capacitance throughout the component. Failure to follow this rule would result in needless loss and reduced SWR bandwidth in an antenna.

The speculation or supposition that the first few turns of a loading coil carry most of the current is clearly untrue. In order to shunt current off, high series impedances would have to exist along with high stray shunting capacitance to areas removed from the radiator. Additionally, the remaining coil area connected to the top area of the antenna above the loading coil would have to present a high impedance to the area where current reduction occurs. This would never be the case, unless the top area of the antenna and loading coil are not resonant near the operating frequency.

A reasonable test for proper inductor and system design would be to remove the antenna above the loading coil, measuring system resonance. If resonance does not change substantially, the area above the coil is not correctly terminating the system. First-order self resonance of the inductor (parallel resonance), when removed from the system, should also be far above the operating frequency of the system. If self-resonance comes within three or four times the operating frequency range, the loading coil almost certainly will have needless performance shortfalls.

Conclusion

Mobile antennas, short verticals, loading coil loss,and loading coil current

A normally functioning inductor has essentially equal currents throughout the inductor, loading coils are no exception. Any current difference requires a substantial current flow through undesired stray capacitances or leakage currents. Neither radiation or induction could change this, it is a basic rule of circuitry.

In a reasonably well-designed system, current into and out of the loading inductor should be substantially equal. Differences in current would indicate excessive and problematic undesired stray capacitance in the loading system design, or measurement errors.

Reduced sensitivity to coil Q is primarily a function of additional losses in the system, not reduction of current through the coil.

This page has had `13558` **visits since February 11, 2004.**

Independent Measurements

[[Home](#)] [[Up](#)] [[Roy's Measurements](#)]

Tom,

Thanks very much for sending the copy of the posting. I'd appreciate it very much if you'd post the following on the same reflector for me:

It was brought to my attention that the following statement was recently posted on this reflector by Yuri:

"I will leave it here, as the rest of it. W7EL, author of EZNEC measured toroid coil and found that it HAS different current at its ends, roughly proportional to the part of antenna that it replaces."

This is not true. I assume the statement was made due to careless reading of the postings I made on rec.radio.amateur.antenna, rather than a deliberate distortion, so I'll very briefly describe the results again here. I made measurements of the current into and out of two different toroids in series with an antenna at its base. The first measurement was made using a 33 foot vertical with seven ground radials, whose feedpoint impedance measured 35 - j185 ohms at 3.8 MHz, and a toroidal inductor with reactance of 193 ohms and Q a bit over 300. I found that the difference in current between input and output of the inductor was 3.1% in magnitude and with no measurable phase shift, despite the short antenna. The 3.1% current difference between input and output can be explained simply by 6.8 pF of stray capacitance between the coil output and ground and/or current probe, or a bit greater value distributed over the coil. I repeated the measurement on the bench, with a series resistor and capacitor in place of the antenna, and measured 2.3% difference between input and output current. This is within measurement error of being the same as when connected to the antenna. (One would expect the stray C to be a bit different, also.) So in the one case I did the measurement with the inductor "replacing" a very significant

part of an antenna and in the other with the inductor replacing no antenna at all (or an entire antenna, if you choose), and got essentially the same result. How this can be interpreted as my having drawn the conclusion stated by Yuri is beyond me.

Unfortunately, the fact that the antenna is spaced only about 1/4" from a 4 foot mounting pipe (which altered its input impedance), in addition to the abbreviated ground system, left a great deal of wiggle room for people proposing alternate theories, so I did a second test with a more idealized antenna. For this test I constructed a vertical antenna that was 33 feet high, made of #16 insulated wire, and strung 23 radials out on the surface of the wet ground. The feedpoint impedance of this antenna at 3.8 MHz was measured as 15.8 - j437 ohms. This is close to the theoretical impedance, assuming about 8 ohms of ground loss resistance. The inductor had a reactance of 387 ohms. As described in my posting on rraa of November 11, the inductor "replaces" about 33 electrical degrees of the antenna.

The result from the second test was a current difference of 5.4%, again with no measurable phase shift. And again, this small difference can be explained by about the same amount of stray capacitance. It's nowhere near the 16+% that the "cosine rule" (that the output current equals the cosine of the "replaced" antenna) proposed by Yuri and others would predict. In no way did my measurement validate his theory -- 5.4% isn't even "roughly" more than 16%.

Frankly, I'm a bit embarrassed to have bothered to make these measurements at all -- it's a lot like making careful measurements to validate Ohm's law in order to refute someone's measurements that "prove" it wrong. As it is, it's turned out to be even worse than just a waste of time, since the results are now being distorted to support the very theory they clearly refute. I hope this will clarify just what measurements I made and what they showed.

For anyone who can stomach the waffling, backpedaling, and insults which characterized the original discussion, look up the thread "Re: Current in antenna loading coils controversy" and its variants in the google archives of rec.radio.amateur.antenna. Postings describing my measurements were made between November 8 and 11, 2003. If you read the thread, particularly the reaction to my posting of the measurements, you'll see why I've given up on that newsgroup as a forum for rational communication.

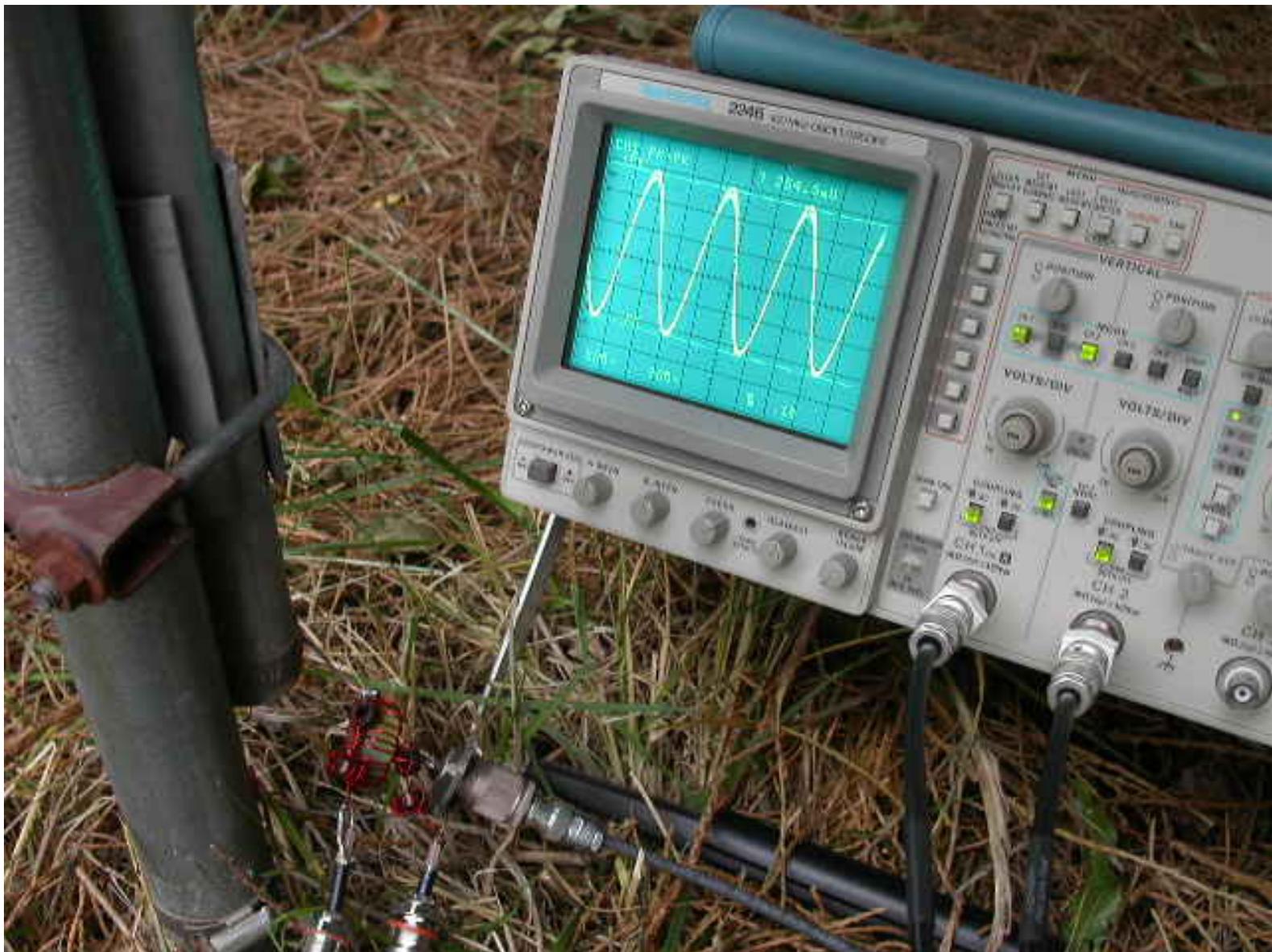Many thanks to Tom, W8JI for bringing the misleading quote to my

attention and for posting this for me.

Roy Lewallen, W7EL

Roy's Measurements

[[Home](#)] [[Up](#)]

W7EL's Antenna current measurements:

This is the overall measurement setup described in the rec.radio.amateur.antenna newsgroup on and about November 9, 2003, showing both traces on the oscilloscope. Signal input is from an HP 8640B signal generator.

This is a closeup of the inductor and current probes. The upper pipe is the vertical antenna. The one to its left is the mounting, which extends about four feet above ground. The two are spaced 1/4" apart. The yellow toroid is the inductor being measured. The smaller black cores are the current probes. One core of each probe is a transformer, with the wire carrying the current to be measured passed through the core's hole. The second core is a common mode choke or "current balun", which serves two functions. One is to insure that no significant current flows on the outside of the coax cables connecting the probes to the oscilloscope. The second is to isolate the probes and reduce any possible capacitive coupling that might disturb the measurement. The probe outputs are terminated with 50 ohms at the scope.

Roy's Measurements

Inductor operation

The following is a copy of RRAA postings by Roy Lewallen W7EL about inductors. I think the technical portion of these posts are useful for understanding current in loading coils, and how we need to cross check theories properly. I've included all of the text, even though much text is not technical. This really is educational, so read it if you have time!!!

I think the goal should be education, not conversation, so I enlarged important text size and changed color to blue in what I think are important areas.

Comments of mine are in RED.

W8JI Tom

The following are verbatim copies of postings made on rec.radio.amateur.antenna, in the thread "Re: Current in antenna loading coils controversy" (first posting), "Re: Current in antenna loading coils controversy - new measurement" (last posting), and "Re: Current in antenna loading coils controversy (long)" (remainder of postings). I and others made many additional postings on the topic, which should be available at http://www.groups.google.com.

Roy Lewallen

January 11, 2004
Posted on 11-4-03:
No, I will make one more comment. After a bit of reflection, I think this might be at the core of some people's problem in envisioning a lumped inductor.

When a current flows into an inductor, it doesn't go round and round and round the turns, taking its time to get to the other end. An inductor wound with 100 feet of wire behaves nothing like a 100 foot wire. Why? It's because when the current begins flowing, it creates a magnetic field. This field couples to, or links with, the other turns. The portion of the field from one turn that links with the others is the measurable quantity called the coefficient of coupling. For a good HF toroid, it's commonly 99% or better; solenoids are lower, and vary with aspect ratio. The field from the input turn creates a voltage all along the wire in the other turns which, in turn, produce an output current (presuming there's a load to sustain current flow). Consequently, the current at the input appears nearly instantaneously at the output. Those who are physics oriented can have lots of fun, I'm sure, debating just how long it takes. The field travels at near the speed of light, but the ability of the current to change rapidly is limited by other factors.

So please flush your minds of the image of current whirling around the coil, turn by turn, wending its way from one end to the other. It doesn't work at all like that. The coupling of fields from turn to turn or region to region is what brings about the property of inductance in the first place.

That was a great explanation of how an inductor works. I often learn new or better ways to view things when reading what Roy writes!

Radiation is another issue, and provides a path for current, via displacement current, to free space. (I can see it now in Weekly World News: WORLD FILLING WITH COULOMBS! DISASTER LOOMS!) For a component to fit the lumped element model, radiation has to be negligible. And, for the same reason, it can't be allowed to interact with external fields as a receiver, either.

This is very fundamental stuff. You can find a lot more about the topic in any elementary circuit analysis or physics text. If you don't believe what you read there, just killfile my postings -- you won't believe me, either, and reading what I post will be a waste of time for both of us.

Real inductors, of course, are neither zero length nor do they have a perfect coefficient of coupling. And they do radiate. The essence of engineering is to understand the principles well enough to

realize which imperfections are important enough to affect the outcome in a particular situation. We simplify the problem by putting aside the inconsequential effects, but don't oversimplify by ignoring factors that are important for the job at hand. Those who insist on using only the simplest model for all applications will often get invalid results. And those who use only the most complex model for all applications (as is often done in computer circuit modeling), often lose track of what's really going on -- they become good analysts but poor designers. I've seen people capable of only those approaches struggle, and fail, to become competent design engineers.

And with that, I'm outta here. Hope my postings have been helpful.

Posted 11-8-03:
Here are some preliminary details about the inductor current measurement I made.

My antenna isn't nearly as ideal as the one Yuri described. (But if my results are different from the ones reported at the web site Yuri referenced, I'll be eager to hear why.) It's about 33 feet high, and has only 7 buried radials. The feedpoint impedance indicates a loss of about 25 ohms at 7 MHz, so I'd expect it to be a bit more at 3.8. It's bolted to a galvanized fence line post which protrudes nearly four feet from the ground, with spacing between the antenna and the post of about 1/4". This mounting has only a minor effect on the feedpoint impedance at 7 MHz, which is the antenna's intended frequency of use. It's quite profound at 3.8 MHz, though. The expected 370 or so ohms of capacitive reactance is transformed to 185, while the feedpoint R is 35 ohms, not far from the expected value. So the overall feedpoint Z is 35 - j185 ohms at 3.8 MHz, measured with a GR 1606A impedance meter. (I found that my MFJ 269 was about right with the X, but measured R as zero -- apparently the combination of low frequency and large X is a problem for it in resolving the R.) So I built an inductor with measured impedance of 0.6 + j193 ohms. It's 26 turns on a T-106-6 toroid core. Q is a bit over 300. This was placed in series at the antenna feedpoint.

For current measurements, I made two identical current probes. Each one consists of 10 turns wound on an FT-37-73B ferrite core. The two leads from the winding are twisted and wound in bifilar fashion on another FT-37-73B core, 10 turns. This is then connected to an oscilloscope input via a two-foot (approx.) piece of RG-58. A 50 ohm termination is also at the scope input. This gives the probe a theoretical insertion impedance of 0.5 ohm. While making the measurements, I moved, grabbed, and re-oriented the coax cables, with no noticeable effect. This gave me confidence that the outsides of the coax weren't carrying any significant current.

One probe went to each channel of the scope. I left the two scope inputs in the cal position, put both probes on the wire at the input end of the inductor, and recorded the p-p values with the scope's digital measurement feature. Then I reversed the order of the probes and remeasured. I found a slight prejudice toward the probe closest to the source -- 1.2% in one ordering, and 2.1% in the other. Averaging the two channels, though, showed them to be the same within less than 1%. (Each probe was always connected to the same scope channel, so this is a test of the probe-scope channel combinations.)

Then I moved one probe to the output side of the inductor, and measured input and output current.

And I reversed the probe positions on inductor input and output. The ratio of output to input current in the two tests differed by only 1.4%.

When I encounter an astrologist, they invariably ask what "sign" I "am", then proceed to tell me how my personality meets their expectations. So what I do instead is to have them tell *me* what "sign" I "am" *first* -- which they should easily be able to do, based on my personality. Well, they don't find that to be fair, for some reason (although I certainly find it amusing). And so, I doubt if the following challenge will be regarded to be fair, for much the same reason. Those with alternative rules for solving circuit problems are challenged to predict what the ratio of output current to input current will be. I'm particularly targeting Cecil, and others who have bandied about a lot of pseudo-analysis about electrical length, reflections, and the like. And, Richard (Harrison), who said something like "an inductor without phase shift is like". . . I don't recall. . .hot dog without ketchup or something. Pull out your theories, and calculate it, like any competent engineer should be able to do. For cryin' out loud, it's a simple series circuit (except for Cecil, where it's some kind of distributed thing).

First post your answers, then I'll post the result of my measurements.
Roy Lewallen, W7EL

Posted on 11-11-03:
It sounds like the predictions are in. Among the several people who believe that the current out of a small inductor doesn't equal the current in, only Yuri was able to calculate a predicted value for the test, of 2.5 - 5% reduction in current at the output compared to the input, with a phase shift of about 18 degrees.

You see above Yuri "predicted" a current value based on a cosign formula that relates to antenna area. One way to test a theory is to change the conditions or parameters of the test, and see if the predictions apply in multiple cases or if it was just a "guess" or dumb-luck. This is ALWAYS a good idea. Watch what happened when Roy changed the antenna length and used the same formula!

What I measured was a 3.1% reduction in magnitude from input to output, with no discernible phase shift. The 3.1% is an average of two readings, with the input and output probes exchanged. The output was smaller than the input in both measurements, about 2% and 4%. So I believe there's a real difference between output and input current, although with the accuracy of my measurements, I only have reasonable confidence it's somewhere between 1 and 5%. I can resolve about 2-3 degrees of phase shift, though, and I couldn't discern any at all. (Yes, the scope trigger was from one channel, not

alternating.) So I have very high confidence that Yuri's prediction of 18 degrees is **incorrect.**

I don't subscribe to the notion that the current out of a very small inductor should be different than the current in due to some magical property it acquires when connected to an antenna. My working hypothesis is that the currrent difference I did see was due to stray capacitance, either from the probes or simply to the Earth and other objects. It would take an equivalent of 6.8 pF at the output of the coil (that is, between the coil output and the current probe) to get 3% reduction, and only about 1/3 that amount to see the minimum value of reduction of 1% I estimate was actually present.

Here is where Roy changed the test. Now the coil isn't in an antenna, and the "sine rule formula" should not apply. If the sine rule was correct, the current differential should have changed significantly.

I repeated the test on the bench, with a 36 ohm resistor in series with a 220 pF capacitor substituting for the antenna. The result was a 2.3% output:input reduction, again with no discernible phase shift. This is within the measurement error of being the same result. This is what should be expected -- except for unintentional coupling to the antenna's field, the inductor's environment is the same on the bench as at the antenna base, in these single frequency, steady state tests. (That also contradicts what some newsgroup participants have been claiming.)

So, although the small output:input current reduction was within Yuri's prediction, the phase shift certainly wasn't. If time permits, I'll make a more idealized antenna and repeat the measurements with a larger inductor at the base of a more reactive antenna. I'll predict in advance that if I double the amount of loading L, I'll approximately double the amount of current magnitude attenuation -- that is, to somewhere around 6%. That's what should be expected if the cause of the attenuation is stray C or a similar phenomenon.

I've added a picture to the http://eznec.com/rraa/Inductor_Current_Measurement.html page, showing the overall setup including the scope. It gives a little better perspective on the relative sizes of various objects.

Roy Lewallen, W7EL

Posted on 11-11-03:
Today's project was to construct and measure a more idealized antenna.

The antenna is 33 feet high, made of #16 insulated wire. I put out 23 radials on the surface of the

wet ground. Radials were of various lengths, most about 30 feet long. The feedpoint impedance of the antenna, measured with a GR bridge, was 15.8 - j437 ohms at 3.8 MHz. Allowing 3% lengthening effect for insulation, EZNEC says a lossless vertical of that height and diameter should have an input Z of 7.5 - j478. 8.3 ohms loss resistance is reasonable for that number of radials, and the somewhat lower than predicted reactance is likely due to the fact that the radial wires were grouped together as they came up a few inches to the antenna base, and not immediately coming in contact with the ground. That would add a bit of inductive reactance.

I wound an inductor on a T-106-6 core as before, but with more turns, for a measured Z of 1.3 + j387 ohms. After putting it in series with the antenna at the base, the base impedance measured 17.1 - j54 ohms. This is only 4 ohms from the expected reactance, and spot on the expected resistance, so measurements are consistent.

Analyzing verticals with EZNEC, made from #16 wire at 3.8 MHz, shows that:

-- An antenna 63.2' high is resonant.
-- An antenna 35.9' high has a feedpoint reactance of -j437 ohms.
-- An antenna 59.35' high has a feepoint reactance of -j54 ohms.

With a resonant height of 63.2', you could say that 63.2' is "90 electrical degrees" as far as the antenna is concerned. So you might say that my inductor has "replaced 33.4 electrical degrees" of the antenna.

Here again the test is modified to test Yuri's prediction.

Using Yuri's cosine rule, we should then expect the inductor output current to be cos(33.4 deg) times the input current, or 16.5% less. Also, we should expect to see those 33 degrees of "replaced antenna" as phase shift from the input to the output of the inductor. That is, the current change from the input to output of the inductor is the same as it would be for the portion of the antenna it "replaces". (I think Jim Kelley subscribes to this theory also, but I'm not sure.)

In contrast, conventional electrical circuit theory predicts no current difference between the input and output for a physically very small inductor with no radiation or stray coupling. I saw about 3% in the previous measurement, which I believe can be attributed to stray capacitance. So I predicted that we should see about twice that amount with the higher valued inductor used for this experiment (387 vs 192 ohms reactance). I didn't see

any measurable phase shift between input and output before, so I didn't expect to see it this time.

So for this test, there's quite a difference in predictions for output:input current --

**Yuri's method predicts a reduction of output current magnitude of 16.5% and a phase shift of 33 degrees.

**I predict around 6% magnitude reduction (due to stray C) and no measurable phase shift (less than 2 or 3 degrees).

I have very high confidence that my measurements are good enough to resolve the difference between these two possibilities.
Would anyone care to comment before I post the measurement results? And, Yuri, please correct me if I've misinterpreted your theory.
Roy Lewallen, W7EL

Here are the test results, still using the same formula. If the results of the Cos theory and formula are correct, the results would be 16.5% current reduction and 33 degrees phase shift. This shows why it is important to test new formulas in a variety of applications.

Posted on 11-11-03:
Ok,
For anyone who cares, the magnitude of the current out of the inductor in the later test measured 5.4% less than the current in. No phase shift was discernible. An analytical person could build on this information to investigate the properties of longer inductors placed elsewhere in the antenna.

Thank you for the comments, Cecil, Yuri, Richards, Art, and others. I've learned a good lesson from this -- that this isn't an appropriate forum or appropriate audience for the sort of quantitative analysis and reasoning I'm familiar and comfortable with. And that the considerable time and effort required to make careful measurements is really of very little benefit -- certainly not anywhere near enough to justify it.
With a great sigh of relief from everyone, I'm sure, I'll now turn this thread back over to Yuri, Cecil, et al.
My apologies to everyone for taking up so much bandwidth.
73,
Roy Lewallen, W7EL

So there we have it. The Cosine theory that states an inductor replaces part of

the antenna and has the same or similar phase shift and current taper as that part of the antenna proved false once the system was perturbed. It is quite typical for a formula that is incorrect to work at one point just by sheer luck.

When we have a theory, the test is to apply that theory to various systems and see if it works in all cases. If it does not work in all cases, the theory is incorrect.

While I agree it should not be necessary to prove conventional circuit theory, I think Roy's test (and my test) might shed some useful insight into how short and heavily loaded antennas actually work. A secondary benefit is we see how an incorrect idea or concept can actually work once in a while just by sheer luck, and why we need peer review.

Many people have the misplaced notion a wire in an inductor replaces a certain "electrical length" of the antenna. That isn't true. A loading inductor simply cancels reactance (corrects power factor). Current change across the inductor actually indicates a problem with the inductor, unless it is intentional.

Mobile Antenna Current Measurements at W8JI

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A sample of measurements above and below the loading coil with various antenna above the coil (current as percent of reference) follow:

	Toroid with hat	Small 2x2" coil with 24" hat up 24"	Long 12x3" coil with 24" hat up 24"	Long 12x3"coil with 6' whip
Current below	100%	100	100	100
Current above	100	94.4	73	76%

	Toroid with whip	long 12x1.5" coil with 6' whip	long 12x1.5" coil with 24" hat	Small 2x2" coil with 6' whip
Current below	100%	100	100	100
Current above coil	100%	79%	75%	96%
Current in whip 1ft above top of coil		73%		

The most revealing thing was how noticeably small changes in stray capacitance near the middle and top of the loading coil affect current distribution. It was quite evident hanging a large meter on each end of the coil would greatly perturb the system.

Mobile Antenna FS Comparisons

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The hat used in these tests was the DX Engineering Hot Rodz system.

Type	Height Base	Top Length	Hat dia	Field Strength	Frequency	Relative FS	
Reference				-18.5			best tested antenna
Large air	36"	84"	0	-18.50	7.2	0.00	3"dia #12
Large Air	36'	27"	47"	-18.70	7.2	-0.20	3" dia #12 47" hat
Small air	36"	27"	47"	-19.90	7.2	-1.40	1.5"dia #16 47" hat
TarHeel	42"	27"	47"	-21.00	7.2	-2.50	Tar with 47" dia
TarHeel	43"	84"	0	-21.10	7.2	-2.60	Tar with 7' whip
RM-20	36"	27"	52"	-21.20	7.2	-2.70	Hustler RM-20 on 40m
TarHeel	43"	27"	23"	-22.00	7.2	-3.50	Tar with 23" dia
TarHeel	61"	84"	0	-19.80	3.5		
TarHeel	59"	27"	47"	-19.70	3.5		47" DXE

The most striking things are:

- **The range between the best and worse system was only 3.5dB**
- **A large hat at the TOP of the antenna allows considerable shortening with no loss in efficiency**
- **For a given height antenna, efficiency is increased with a hat at the top**

Mobile and Loaded Antenna

Inductor Spice Model

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Spice model of inductor including stray capacitances:
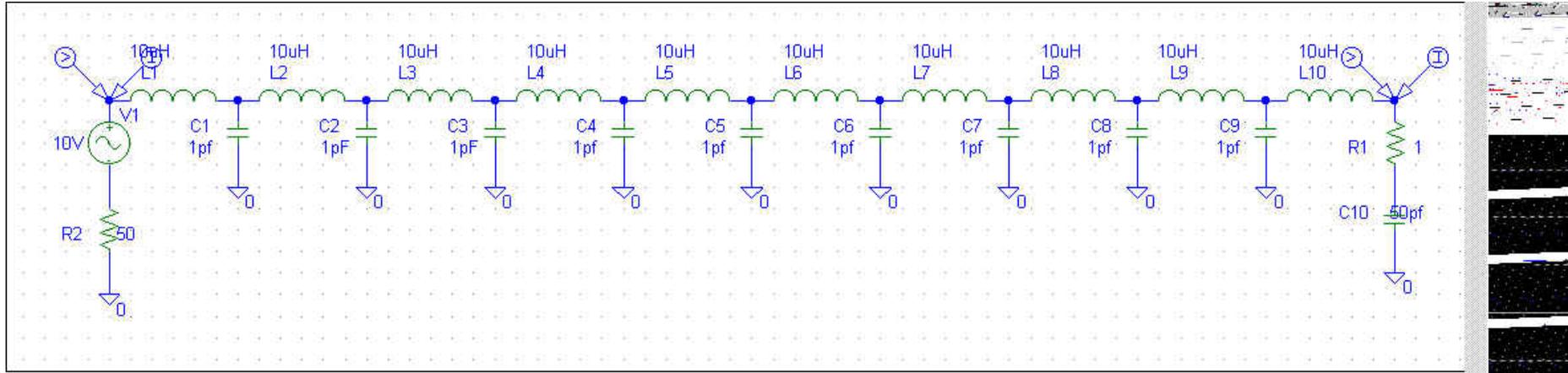
Simulation.
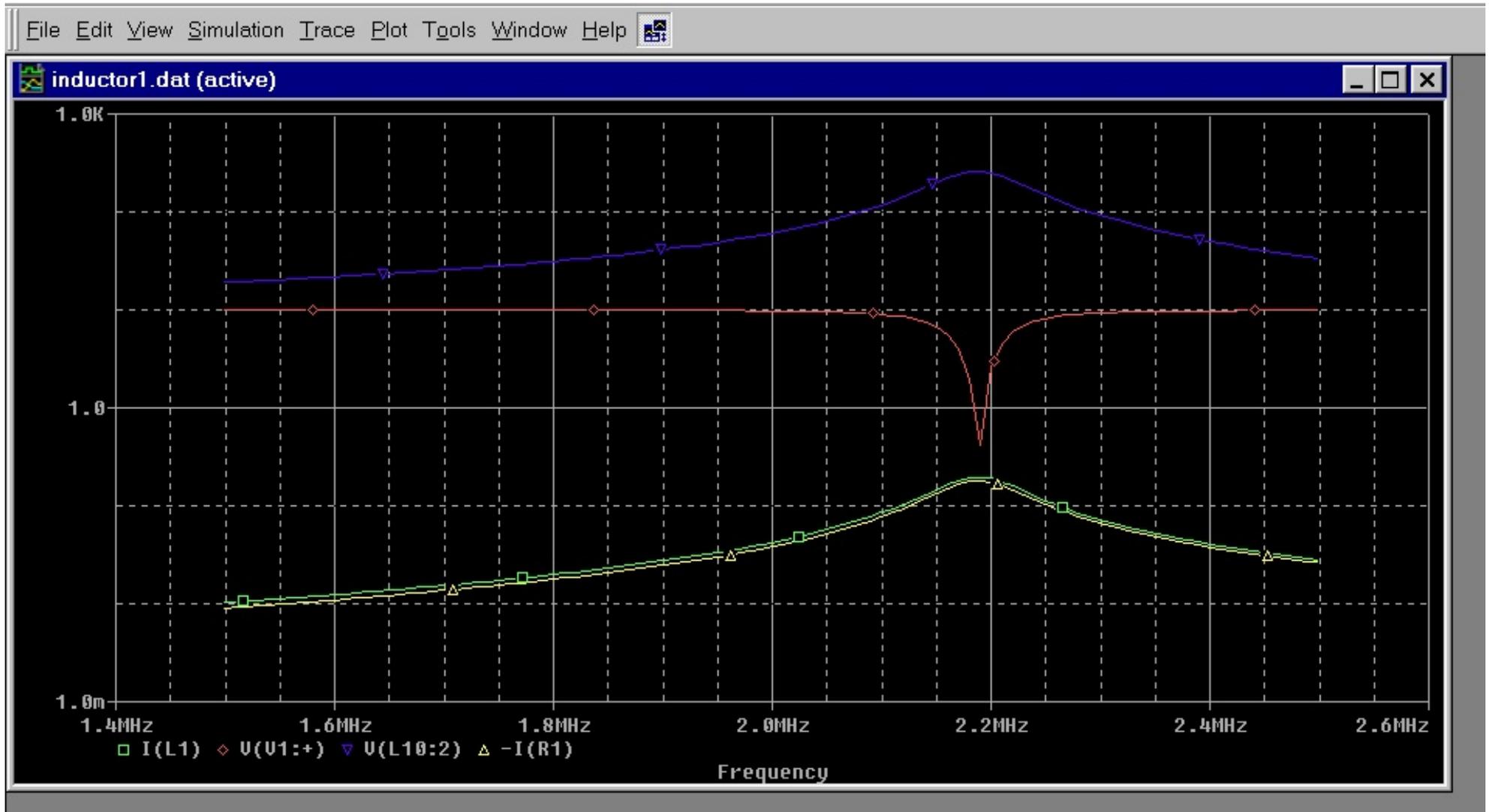
Red=input voltage

Green=input current

Blue=load voltage

Yellow=load current

   This system (including reasonable leakage capacitance) representing a 50 ohm source driving a 100uH inductor with stray capacitance of 9pF and a 50pF load (small

whip), shows input and output currents are essentially equal, while output voltage greatly increases at resonance.

In the following SPICE plot, increasing stray capacitance shows expected departure from equal input and output currents. Distributed capacitance to ground is now at 45pF, nearly equal to the 50pF antenna capacitance:
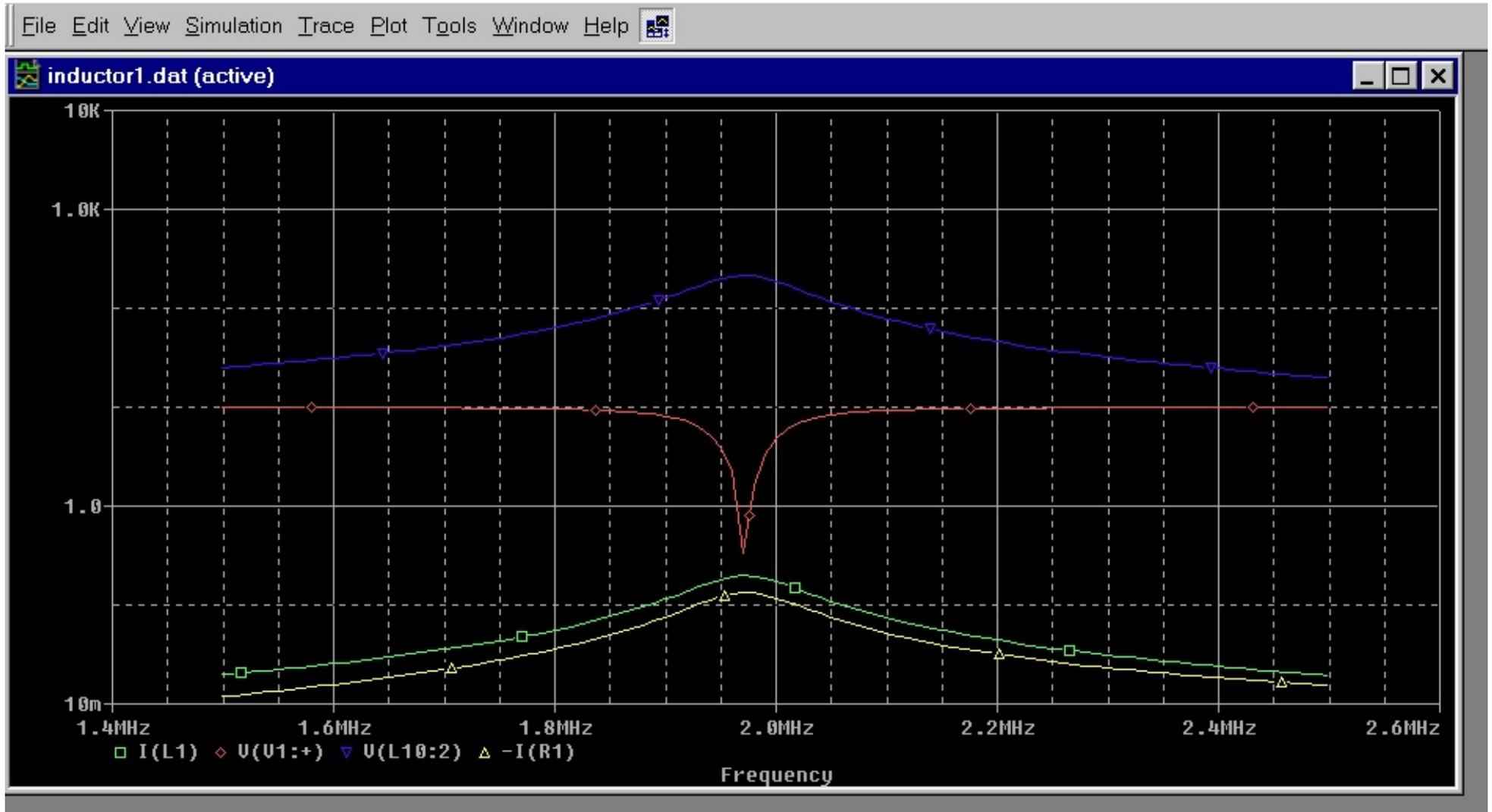
Again despite stray capacitance nearly equaling load capacitance, current levels at input and output are reasonably similar:

Green=source current

Yellow=load current

Red=source voltage

Blue=load voltage

If we multiply output voltage times current, the apparent power is many dozens of times the input power. The reason is simple, phase relationship between voltage and current along the inductor changes even while current remains essentially uniform.

Although casual experimenters often assume a loading coil "replaces" the missing length in an antenna and has sinking voltage with rising current, this is not the case unless stray capacitance is extreme or the inductor is incorrectly and inefficiently operated at or near self-(series) resonance. The loading inductor really just corrects power factor, and bringing voltage and current into phase at resonance. It does replace missing length by simulating an antenna.

Linear loading is no different, it simply behaves like a very lossy low-Q coil and does not add radiation to the system. Unless of course we are talking about heat radiation.

Loading Inductors

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Optimizing Inductors

While intended mainly for antenna loading coils, the text below also applies to other resonant systems, such as amplifier tank circuits.

Related pages:

[Inductor spice model](#)

[Mobile and loaded verticals](#)

Optimum form factor of any inductor depends heavily on the actual end use of the component. Current through and voltage across an inductor as well as the required reactance influences both materials and form factor of an inductor. In high impedance systems, optimum form factor becomes longer with a smaller diameter. Even the best insulation materials will have a deleterious effect on component Q when impedances are high, and dielectric constant often becomes more important than dissipation factor in insulating materials. With low impedances, optimum form factors become more square and insulation has a much less noticeable effect.

Weight, size, and cost often require use of less-than-ideal materials and construction, but careful design generally results in a compromise that doesn't noticeably compromise system performance.

As a further complication, very simple systems might work quite differently than we intuitively believe. Many experimenters fail to consider what the inductor does in the system, and how the inductor behaves internally. For example many people assume, with a fixed amount of applied power, as voltage increases current must proportionally decrease. This isn't true in reactive systems, where voltage and current are not exactly in-phase or 180-degrees out-of-phase.

It is relatively common for programs and formulas to greatly overestimate maximum obtainable inductor Q. Confirming measurements are often flawed, either being made far from the operating frequency or with inadequate methods or equipment. This is especially true in high impedance high frequency systems.

Worse of all, antenna manufacturers and builders tend to seriously understate or estimate loss in other forms of loading, specifically linear-loading.

There are five common errors we should avoid:

- Building for excessive Q, when the reduced ESR will not noticeably improve system efficiency
- In antennas, considering inductor ESR directly as a portion of loss resistance at the point where radiation resistance is taken, rather than normalizing ESR to the point where radiation resistance is taken
- Believing programs or articles predicting Q's in the range of 1000 or more for inductors
- Thinking one optimum form factor (L to D ratio) always provides optimum performance
- Misapplying radiation or loss resistance formulas
- Believing claims that loading reactance obtained from stub or linear-loading provides lower loss than well-designed lumped loading

Before doing anything with information in this article or any other article related to loss and efficiency in antennas, please read the radiation resistance article on this site!

Range of Inductor Form Factor

There are two critical form factor dimensions, diameter and length. The ratio of diameter to length has two limits. The first limit occurs when the inductor occupies only one wire diameter as the length. The other limit occurs when the inductor is one wire diameter in diameter. The first condition would be met by a single layer pancake coil, the second by a linear conductor such as the inductance of a single wire transmission line. Optimum form factor occurs between these two extremes, and varies with the exact application.

Length-to-diameter ratio is important for two reasons:

- Shorter lengths and larger diameters increase capacitance across the inductor. Capacitance across any inductor carrying time-varying current increases circulating currents in the inductor, increasing loss while simultaneously reducing system bandwidth.
- Longer lengths and smaller coil diameters reduce mutual coupling between turns and increase leakage flux. This results in use of increased conductor length for a given inductance, increasing wire resistance.

These two situations are obviously in direct conflict, a balance must be achieved. Optimum balance between conflicting L/D effects listed above depend heavily on external circuit capacitance and operating frequency.

There is actually only one nearly constant parameter in design of high-Q RF solenoid inductors,

turn-to-turn spacing. Optimum turn-to-turn spacing occurs when the spacing or gap between turns is about the same diameter as the wire. If the turn-to-turn gap is filled (even partially) with insulation, optimum *conductor spacing* increases.

For the purposes of this article, the following terms are used:

- D=diameter
- d=turn diameter
- L=coil length

As a general rule, Q in a RF inductor peaks with a form factor (L/D) between 1 and 4.

The size and shape of the conductor used in the coil sets the optimum diameter, larger conductors require larger diameters.

Lower optimum L/D ratios (near unity) appear in systems where higher amounts of external capacitance load the system. Two examples would be amplifier tank circuits or large antennas with considerable loading capacitance beyond the coil. Another way to view this is by resonant frequencies. Form factor moves closer to 1:1 when an inductor is operated far below its natural self-resonant frequency.

Higher optimum L/D ratios (up to 4:1) occur when capacitance values external to the coil are reduced. Small mobile antennas without hats, especially top-loaded antennas, require longer form factors. Such systems operate the inductor closer to its self (parallel) resonant frequency.

The Reason Optimum Form Factors Vary

The underlying reason for change in optimum form factor with external circuit impedance rests almost entirely on inductor stray capacitance and mutual coupling between turns.

With high external capacitance, any reasonable amount of internal stray capacitance shunting the inductor causes a very small change in circulating current in the inductor. The external circuit, in effect, determines circulating currents. In this case, inductor Q is set mostly by flux leakage and conductor resistance in the inductor.

In such a system designers can place turns closer together, increasing mutual coupling or flux linkage from turn-to-turn. Since external capacitance causes most of the circulating currents, any increase in inductor distributed capacitance has little effect on the system. It becomes most important to reduce wire resistance by minimizing wire length. Dielectrics around the conductors have little effect on Q, because increases in capacitance caused by replacing air with a dielectric has little effect on the overall circulating currents.

As the system's external capacitance is reduced, circulating currents inside an inductor become increasingly influenced by stray capacitance. This includes capacitance within the inductor, as well as capacitance between the inductor and objects surrounding the inductor.

If inductor design or location is poor, and if system impedances are high, current can actually vary significantly along the length of an inductor. In properly designed systems, this will not occur.

When external capacitance is reduced, the coil ends must be increasingly separated from each other. The form factor chosen must reduce coil diameter while increasing length. In high impedance (reactance) systems, reducing capacitance improves component Q in spite of the resistance penalty resulting from increased conductor length required in long form factors.

We also must avoid using dielectrics near or in the inductor, especially any dielectric coating or between turns. Dielectrics other than air or vacuum, even low dissipation factor dielectrics, increase stray capacitance. Anything that increases capacitance will reduce component Q, and ALL dielectrics (other than air or vacuum) increase capacitance. The most noticeable effects in high reactance systems often come from dielectrics increasing capacitance, rather than actual dielectric losses!

The increase in loss can be directly proportional to the increase in capacitance, even when required turns are reduced. Low-loss Teflon or Polyethylene dielectrics can be nearly as detrimental as higher dissipation factor materials like fiberglass or Delrin.

Inductor Modeling Programs

Many inductor modeling programs fail to consider two important effects:

- They ignore capacitance across the inductor
- They ignore "current pushing" or bunching caused by strong magnetic fields

The first effect causes Q to peak well below the self-resonant frequency of the inductor. The second effect causes a decrease in Q as frequency is increased or as turns are brought closer together. The second effect occurs because current flows in a smaller and smaller cross section of conductor with increasing frequency.

If a model, prediction, or estimate does not show Q dropping drastically as first order (parallel) self-resonance is approached, the results almost certainly contain significant Q errors.

I've corresponded with some program writers who claim to have verifying measurements, and found their test equipment doesn't reliably operate (or operate at all) at the operating frequency of the inductors! Verifying inductor Q at frequencies far below the operating frequency in the model

does NOT provide any assurance the model or predictions are correct.

Optimum Q

There is a strong tendency to overkill the size of inductors, in an effort to reach unrealistic Q factors. Examples are commonly found in high-performance mobile antenna systems, where ground loss and other system resistances dominate the system. We often find high performance inductors with Q's in the several hundreds (at the upper practical limit of Q) and very low ESR's used in systems where overall loss resistances normalized to the feedpoint are very high.

Even though electrical problems are *NOT* created when using the highest possible Q, there is a point where the end improvement in signal level does not justify the physical size and cost to obtain "excessive" Q.

One example can be found in my Trap Measurements article, where differences in trap Rp (parallel resistance) when #10 AWG and copper tubing are compared, with unnoticeable changes in performance.

Another example appears in my mobile and loaded antennas article.

Inductor Placement in Antennas

The optimum location of an inductor varies with ground resistance and overall length of the antenna. Fortunately efficiency changes are smooth and gradual changes. Minor errors in placement generally do not result in noticeable efficiency changes.

Radiation Resistance and Mobile and Loaded Antennas articles on this site give some perspective of how load placement affects radiation resistance.

Q Ranges

The highest Q HF inductors I have measured, at least when operated away from self-resonance, are copper tubing coils and edge-wound inductors, such as those commonly used in high power tank coils. The highest Q I have measured in very large inductors of optimum form factor in the HF range has been near 1000.

Miniductor-type coils have a surprising amount of Q for the wire size, and maintain Q better as self-resonance is approached than larger coils.

These are the typical ranges of peak Q I have measured:

	HF Q Peak	Q at 80% of self-resonant freq
Copper Tubing Coils	600-1100	400-600
Edge-wound inductors	600-900	400-600
#8 miniductors	500-700	300-500
#12 miniductors	300-500	200-400
#16 miniductors	250-350	200-300
Large #2 mix iron on 1.8MHz	500-600	
enameled wire close-wound	200	100

As I measure inductors in the future, I'll include pictures and impedance data here.

Final Comments

We should keep the following in mind:

- Optimum form factor varies with application.
- Q peaks at some frequency significantly lower than self-resonant frequency, at self-resonance Q is zero (the coil appears as a pure resistance to any external circuit). Above that frequency inductor becomes the electrical equivalent of a low-Q capacitor.
- Linear Loading is really nothing other than a poor form-factor inductor. The radiation from the linear loading does NOT change the radiation resistance of the antenna except as the effective position of the load might change from the direction of fold. In all cases, a proper form-factor inductor would have less loss, and provide the same radiation resistance.
- Most inductor Q calculation programs overestimate Q.
- Any metal around an inductor decreases Q. Copper or steel, it often has nearly the same effect.
- Any dielectric (even low dissipation dielectrics) decreases Q because the dielectric increases shunt capacitance. This increases circulating currents. The effect is most pronounced as self-resonance is approached.

Radiation Resistance

[[Home](#)] [[Up](#)]

Related pages on [Antennas](#) , [radiation and fields](#) , [mobile and short verticals](#)

My 2004 Dayton Hamvention Power Point presentation on Small Verticals can be downloaded here... [DAYTON 2004](#).

The main points when dealing with small antennas are:

There is no magic bullet or magic cure to make a small antenna act like a large one. It all comes down to current distribution over linear distance.

Small antennas require extraordinary care to obtain high efficiency.

How do we make a *small* antenna as efficient as possible?

- First, we make current as uniform as possible over the length of the antenna.
- Second, we use low loss loading such as optimum form (size, length, and diameter) [loading coils.](#)
- Third, we make the antenna as large and straight as possible in a line. We don't fold, bend, zigzag, or curve the antenna especially in the high current areas.
- We keep the high voltage points (the open ends) away form other things (like lossy earth), and the high current areas away from other large lossy conductors.
- Most important, we keep current as high as possible throughout the length of the antenna by using as much capacitance as possible at the antenna ends.

What this does is maximize radiation resistance (while at the same time minimizing loss). The text below explains how radiation resistance and loss interact.

Radiation Resistance

 Radiation resistance is probably the most abused and misused term in antennas. The reason it is so often misused is the lack of clear definition. When a term has several nebulous meanings and uses, it is only natural that misuses will appear. The lack of a firm well-accepted definition allows meanings to slip from one application into another, where a totally erroneous conclusion can be drawn from what otherwise would be a good formula!

Common Uses

There are two commonly used "correct" meanings of radiation resistance and one totally incorrect use. The "correct" uses are:

- The resistive part of an antenna's feedpoint impedance that is caused by radiation from the antenna
- The total EM power radiated in all directions divided by the square of net current causing the radiation

Neither of the above definitions include loss resistances of any type! The moment loss resistance is included, we have the third commonly used (but totally useless) definition. This definition, which includes losses, must be considered incorrect because it is comprised of resistances that have nothing to do with radiation. The misused definition is:

- The real (or resistive) part of an antenna's feedpoint impedance including loss related resistances.

The correct name for the third "radiation resistance is actually the antenna *feedpoint resistance*, not radiation resistance!

Of the above *good* definitions, the first definition is the most commonly abused through mistake. The second definition is an IRE definition (albeit a good one that never caught on). In every case I've seen, it is the second good definition that always provides the most direct and useful answer.

Examples of Misuse

Folded Monopoles

Folded monopoles probably provide the best examples of misuse of the term radiation resistance. Quite often, in discussions of folded monopoles, claims are made that multiple drops increase the radiation resistance and lower losses. The justification for this incorrect claim is the folding raises

radiation resistance, and % eff = 100 * Rrad/(Rrad + Rloss) .

What folded monopole fanatics forget is that all losses must be normalized to the point where radiation resistance is taken, otherwise the efficiency formula won't mean a thing!

Let's look at what actually happens in a folded element, and use it to understand how the poor definition of radiation resistance causes the misunderstanding.
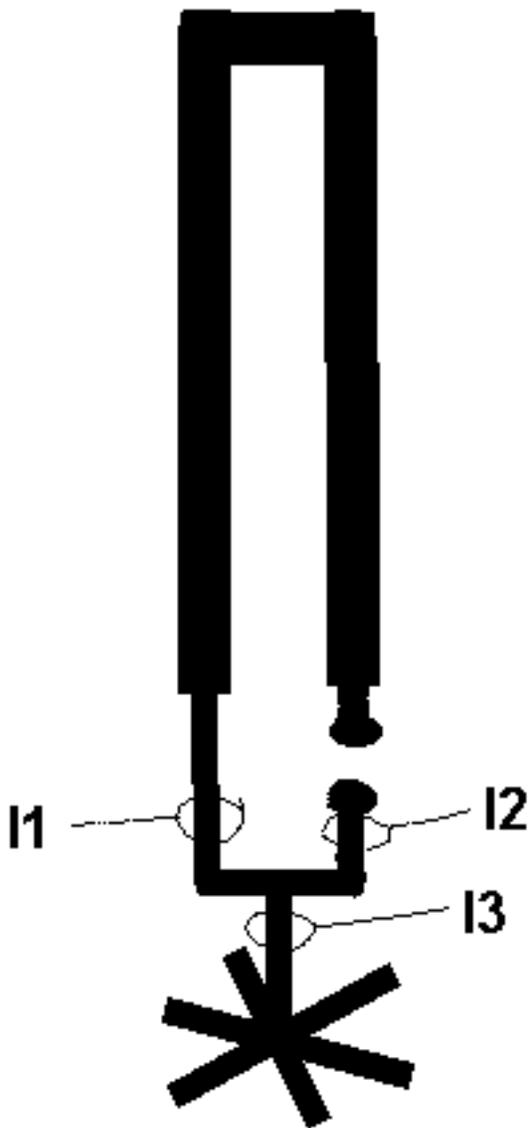
Consider the unipole above. Lets assume we short the open terminals, and feed it as a normal Marconi vertical with a feedpoint at the point where we measure I3. I3 is ALWAYS the vector sum or in-phase combination of currents I1 and I2. With continuity through each leg, I1 and I2 share all of the ground current. This happens regardless of where the feedpoint is located in the lower portions of the antenna.

With 1/4wl height and a reasonable element diameter, the radiation resistance (fed as a traditional monopole) would be about 36-ohms.

Assume ground loss, normalized to the point where we measure I3, could be represented by 14 ohms. Applying 500 watts would make current I3 equal 3.16 amperes. **Power loss in ground resistance would be I^2R, or 3.16^2 times 14, about 140 watts.** Feedpoint resistance would be 50 ohms. Feedpoint power, as a check, would be 3.16^2 times 50 ohms or 500 watts. With equal diameter legs, that current would divide and 1.58 amperes would flow in each upper leg.

Let's use the formula %eff = 100*Rrad/(Rrad+Rloss). We have 36/36+14 = .72 so the result is 72% efficiency, 28% loss. **28% loss times 500 watts is 140 watts in ground losses.** This matches the other method just above.

Opening the terminals and feeding as a folded unipole, half of the radiator current is in I1 while the other half is in I2. Current is halved to 1.58 amperes at the feedpoint and power remains the same. The feedpoint resistance now becomes 200 ohms. We can confirm this with I squared R, or 1.58^2 *200=500 watts. It all works out great so far!

Now let's MIS-use the same efficiency formula, like Orr does in his Handbook and others do in various places. We would have 200/200+14 = .9346 or 93.46 % efficiency.

We know we still have 3.16 amperes flowing as I3, and we know ground resistance is still 14 ohms (normalized to the point where I3 is measured). I-squared-R losses are 3.16^2 * 14 = 140 watts! We have exactly the same power loss.

Let's transform the ground loss value that was normalized at 14-ohms where I3 is measured to the feedpoint by the same impedance multiplication as the feed resistance, or 1:4. We'd now think ground resistance would be 4*14 = 56 ohms. 56 ohms of the feedpoint resistance is loss. Trying that same efficiency formula, we get:

144/144+56 = .72, or 72% efficiency!!! Now everything checks out fine.

The Common Mistake

Orr and others have used the first definition of radiation resistance, the portion of the terminal

resistance of the feedpoint responsible for radiation. Unfortunately ***they failed to normalize ground losses to the same point where the radiation resistance was taken!***

We can not use a formula that is based on everything being normalized to one point, unless we actually do that for every term in the formula!

There is no change in efficiency when the NET radiator current remains the same, and when ground current remains the same. It is pathological engineering to think otherwise.

Using The Second Definition

If we use the second IRE definition of radiation resistance, where the effective current causing radiation is compared to power radiated, we find nothing changes. A folded dipole or monopole has the same radiation resistance as a regular dipole or monopole the same size, and a small loop has the same radiation resistance regardless of turns.

The magic vanishes along with the incorrect definitions and perceptions.

You can read about this in textbooks. The "Antenna Engineering Handbook" by Jasik in 3-13, 19-3, and in other sections uses correct definitions and descriptions.

Quad's and other Loops

We find the same efficiency misconceptions in articles about small loops and large quads. Authors sometimes assume, incorrectly, radiation resistance changes in a favorable proportion to loss resistance as we make the feed impedance increase. What we really are doing is placing the feedpoint in series with a smaller portion of NET current causing radiation.

With a large full-size quad element the pattern under some conditions will slightly change, but efficiency remains basically the same. With a small loop antenna, losses can actually increase with more turns!

Terminated Folded Elements

Another abuse of radiation resistance is found in terminated antennas. Some manufacturers and authors claim a resistance can be inserted in series with one leg of a folded monopole or folded dipole, and the other leg fed. The usual arm-waving claim is the antenna isn't really resistor loaded, and that efficiency is very good because radiation resistance is high.

That claim is absolute nonsense!

A large terminated Rhombic is well-known to have poor efficiency. Rhombic gain is actually low compared to other antennas having the same sin/sin x antenna pattern, because Rhombic efficiency is generally less than 50%. At least half of the power is consumed in termination and ground losses below the antenna. The actual gain may be reasonably high compared to a dipole, but not to other efficient antennas with the same half-power beam width.

The typical manufacturing buzz-word is that terminated monopole and dipole antennas are "traveling wave antennas" and by some magic (that even large terminated Rhombic antennas can't achieve) have broad bandwidth and high efficiency.

A Rhombic focuses energy (that is not transformed into heat) into a narrow beam that has considerable gain, but if it sprayed the radiation around in a non-focused pattern, a regular dipole would win hands down. Throw a resistor on that dipole to smooth SWR variations, or on a vertical, and efficiency suffers.

I listened to a station on 75 meters 600 miles away testing a Sommer T-25 vertical. He was 30 over nine using a dipole, and dropped to S6 with the vertical. The European he was working reported a similar change. By removing the termination resistor and base-loading the same vertical, a local Ham gained almost 25dB on 80 meters!

When we abuse or misuse radiation resistance, we can invent all kinds of magical antennas. We can have CFA's, E-H antennas, terminated dipoles, small magnetic loops, and verticals with all sorts of magical claims. Few, if any, of the claims are ever correct. Any time we see a claim that efficiency changes a *large amount* because of feed method change, it should be a red flag.

Increasing Radiation Resistance

Radiation resistance, at least under the useful IRE definition, can be defined by the following formula:

$$Rrad := 160 \cdot \pi^2 \cdot \left(\frac{He}{\lambda}\right)^2$$

which would translate to:

$$Rrad := 1580 \cdot \left(\frac{He}{\lambda}\right)^2$$

Where *He* is the effective height center of accelerating charges that cause radiation. In other words,

He is the effective height, expressed in fractions of a wavelength, of the distributed common-mode current in the structure.

(Common-mode current is the vector sum of all currents, or the effective in-phase current at any point, or the current we would measure if we placed a giant clamp-on current probe around ALL of the conductors at that any given height.)

He and λ must both be in the same units, either given as degrees or decimal fractions of a wavelength.

As an example, a uniform current single conductor antenna has an actual physical height of 15.19 feet on 1.8 MHz, where one wavelength is 546.67 feet, the effective height is:

15.19/546.67= 0.0278 wl

Since charges are distributed evenly throughout the structure the full height is used. The effective height is .027 wl, the same as the physical height. The height in electrical degrees is .0278 * 360 = 10 degrees

We have a radiation resistance of:

$$He := .0278 \qquad\qquad \lambda := 1$$

$$Rrad := 160 \cdot \pi^2 \cdot \left(\frac{He}{\lambda}\right)^2$$

$$Rrad = 1.22$$

We can express this graphically in a chart, such as one found in the Antenna Engineering Handbook by Jasik:

Finding 10-degree height on the graph above, and following that line until we reach the crossing for

unity current ratio, we see the ~1.27 ohm radiation resistance is in agreement.

Notice that the number of vertical conductors does NOT enter into the equation! This is the absolute maximum possible radiation resistance we can obtain for a given radiator height.

Non-uniform Current

Radiation resistance is purely a function of the effective current distribution and height of the radiator, and is limited by height (spatial length)! Current throughout the antenna will not remain uniform if we reduce the size of the flat-top or hat.

Current will become zero at the very top with no hat, and 100% base loading. In this case, with no change in height, radiation resistance will be approximately 1/4th the value of the uniform current example. The result is exactly like a 50% reduction in effective element height.

If we follow the 10-degree line to the intersection point with 0 top current, we find radiation resistance to be around .32 ohms. 1.27 ohms, the radiation resistance for uniform current, becomes 1.27/4 or .3175 ohms.

If we stay on the uniform current line, we find that .3175 ohms would be the radiation resistance of a 5-degree monopole with uniform current.

Efficiency

It often helps to look at the extremes, so we can get a feel for the effect of changes.

Let's look at the poor ground extreme and assume we have system losses, normalized to the current maximum, that are many times the radiation resistance. This would be the case for a short 160 or 80 meter mobile antenna.

In such a system radiation resistance would dominate any change that would affect efficiency. Current distribution would mean everything to efficiency.

Assume we have a base loading coil, either good or poor, and a thin mobile whip above the coil. Efficiency would increase by a factor of approximately four times by installing a capacitance hat with several times the distributed capacitance of antenna conductors below the hat.

Moving the coil would have little or no effect on efficiency.

A six-foot antenna with a large hat would be electrically equal to a 12-foot antenna without a hat.

This is why very poor inductors used on antennas in mobile shootouts, with large hats, equal or beat very large high-Q coils in similar height antennas that do not have large hats. One case in mind was a Hamstick lash-up in a mobile. The Hamstick, a notoriously poor efficiency antenna for 75-meters, soundly trounced Bugcatcher antennas when a large hat was added to the Hamstick.

Moving the coil up on the antenna has the effect of making current below the coil uniform, but without a hat current above the coil is a triangular taper that reaches zero at the element tip. The effective height of the area above the coil is 50% of actual height.

If we add a large hat at the bottom of the whip, current in the whip is actually reduced! At the same time, we change nothing below the coil. The effect of adding a large hat below the whip is to reduce the effective height of the antenna, when considered as a percentage of physical height. Radiation resistance and efficiency is generally reduced by adding a hat just above a coil, even if the hat allows us to use a smaller coil!

Adding a large hat at either end of a coil also reduces coil Q, since a large portion of the hat capacitance directly shunts the inductor.

Conclusion

We can reach the following conclusions:

- Radiation resistance, or at least the useful definition of radiation resistance, is limited by spatial area (or height in the case of a vertical) any antenna occupies.
- Radiation resistance is maximized by making current as large as possible over the entire spatial area of the antenna.
- Surrounding objects generally reduce radiation resistance and efficiency, even when they are NOT resonant, because they reduce effective height! This includes dielectrics that increase capacitance of the antenna to ground, since any increase in capacitance appearing well below the top of an antenna reduces effective height.
- Radiation comes from charge acceleration, nothing else. The longer the linear spatial distance we move charges in, the fewer charges we need to move at any point for the same amount of EM radiation. This is just another way of saying radiation resistance is higher in physically longer structures, especially when they carry uniform current.
- Any antenna (including Linear Loading, Helical Loading, Folded Monopoles, Fractal shapes, CFA, E-H, and so on) claiming to increase radiation resistance beyond the limits outline above is based on misunderstandings or untrue distortions of basic antenna principles.

Radiation and Fields

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Related pages [Radiation Resistance](#)

We tend to think of electric, magnetic, and radiation fields as physical "things"...much like an element would be in chemistry. This can cause problems!

One popular (but incorrect) assumption is we can combine, mix or blend fields into another field. After all.... we have electric fields...we have magnetic fields...surely we can mix them and create an electromagnetic field. This would have a distinct advantage, because we all know nearfield losses are higher than farfield losses. We know the farfield field ratio (or field impedance) is set by the media the wave propagates through. With reasonably dry air or vacuum, at normal radio frequencies, we know the impedance is 377 ohms. Why not just mix the two fields and create the third field without all that needless nearfield loss?

If, in our mind's eyes, we can mix fields the next logical step would be to conclude we can separate, filter, or sort them out whenever handy. If we have a problem with electrical interference, we can just eliminate the electric field. If we have a problem with field intensity (defined as volts-per-meter) causing problems with the neighbor's VCR, we can reduce the electric field and cure the problem. After all, it is electrical interference, isn't it?

We logically might assume, just by name, we can mix the various fields and make a new field. We logically might assume we can split the combined fields back apart if we don't like what is happening to one of them.

Unfortunately, none of the above is true! The above conceptual problems started at the very beginning, because we assumed fields were physical things like building blocks. They aren't.

Fields are really just mathematical descriptions of forces between charges. We have three simple conditions that create physical forces between charges. They are electric, magnetic, and electromagnetic (radiation) forces. They are all created by distinctly different physical actions in a system. We can't mix the various names for effects resulting from a certain physical condition and create a new cause! Fields describe effects, they don't create the causes of those effects!!!

Electric Field

Electric field describes a force created by uneven charge distribution. Nature wants charges to be evenly distributed, she can only take so much piling up of charges in one spot! The force between charges, caused by nature trying to balance or even the distribution of charges, is called an ***electric field***.

Uneven charge distribution goes hand-in-hand with a voltage difference between two physical points. We can obviously can have a difference in charge distribution in insulators as well as conductors. A comb, "charged" by running through our hair (if we have any left), can have an electric field. The force of this field can pick up tiny bits of

paper, as nature tries to equalize the charge distribution. The terminals of a battery have an electric field between them, and when we place a conductor in that field the charges try to equalize. Another example would be an antenna, where a voltage difference (uneven charge distribution) between two points creates an electric field.

Any difference in *charge distribution*, whether the charges are "moving" or standing, causes a physical force. We name this effect an **electric field,** and describe it mathematically in volts over a given distance. This field (or force) decreases rapidly with distance.

Magnetic Field

Magnetic field describes a force created by moving charges. When charges are moving, they exert a force on all other charges around them. We call this effect the **magnetic field.**

One example of a magnetic field is a conductor carrying current. Perhaps it is a wire connected between two terminals of a battery. Another example would be a RF-current carrying conductor in an antenna.

Any movement of charges causes a magnetic field, and somewhere rooted in the creation of that magnetic field is an uneven distribution of charges causing an electric field! Once again, this field (or force) decreases rapidly with distance.

Electromagnetic Field

An **electromagnetic field** is created whenever charges are accelerated. Acceleration occurs whenever a charge changes direction or velocity. When a charge accelerates, all the other charges in the universe feel a force trying to make them move. The only thing that stops that force from going on forever is if another charge (or combination of charges) accelerate to create an equal but opposing force. We call the velocity at which this force ripples through the universe the *speed of light*.

One example of electromagnetic fields is in an AC current carrying conductor, like a power line. The time-varying voltage causes charges to move back and forth, and the change in velocity and direction causes the effect called electromagnetic radiation.

It is plain to see why our antennas have all three fields, and why we can communicate so well over large distances with low power. While the electric and magnetic induction fields drop off rapidly, the (initially) much weaker electromagnetic radiation field goes on until something cancels it. It is the radiation field that allows us to communicate, not the electric or magnetic induction fields!

Near the Antenna

Near any antenna fields are a complex mixture or " soup" of various effects from charges. Charges are almost always moving in multiple directions and over various distances, when viewed from distances close to the antenna. It isn't always easy to understand what actually is happening very near antennas (where the area of the antenna is large compared to the distance from which we observe the effects of charges). Near the antenna, pattern and field impedance is generally nothing like we might intuitively imagine!

It is the response in this area, generally within $\lambda/10$ distance from the antenna, that small "magnetic loop" and

"electric dipole" antennas get their names.

Very close to the high-current area of a small loop antenna (but not near the open ends, because that is where the electric field dominates), the magnetic field dominates. Magnetic fields are mathematical descriptions of forces derived from moving charges, or current flow. This effect, when large compared to the electric field, is sometimes described by saying the "*field impedance*" is "*low*".

Conversely, near a small dipole or monopole with high voltage and little current, the electric field dominates. The largest force is from the very high open-end voltages, and very uneven charge distribution. We might say such an antenna has a " *high field impedance*" in the area where the electric field dominates any forces cause by moving charges.

In all of these cases, if the antenna is electrically small, the dominant fields apply only within approximately λ/10 distance from the antenna!

As we move out further the weaker radiation field, because it is attenuated less with distance, starts to have a noticeable contribution to the charge forces. Because the phase of the fields (fields are a way of describing effects) is different at the antenna, the sum of the effects is different with distance. At some distance the low field impedance of a small loop becomes high, and the high field impedance of a small dipole becomes low!
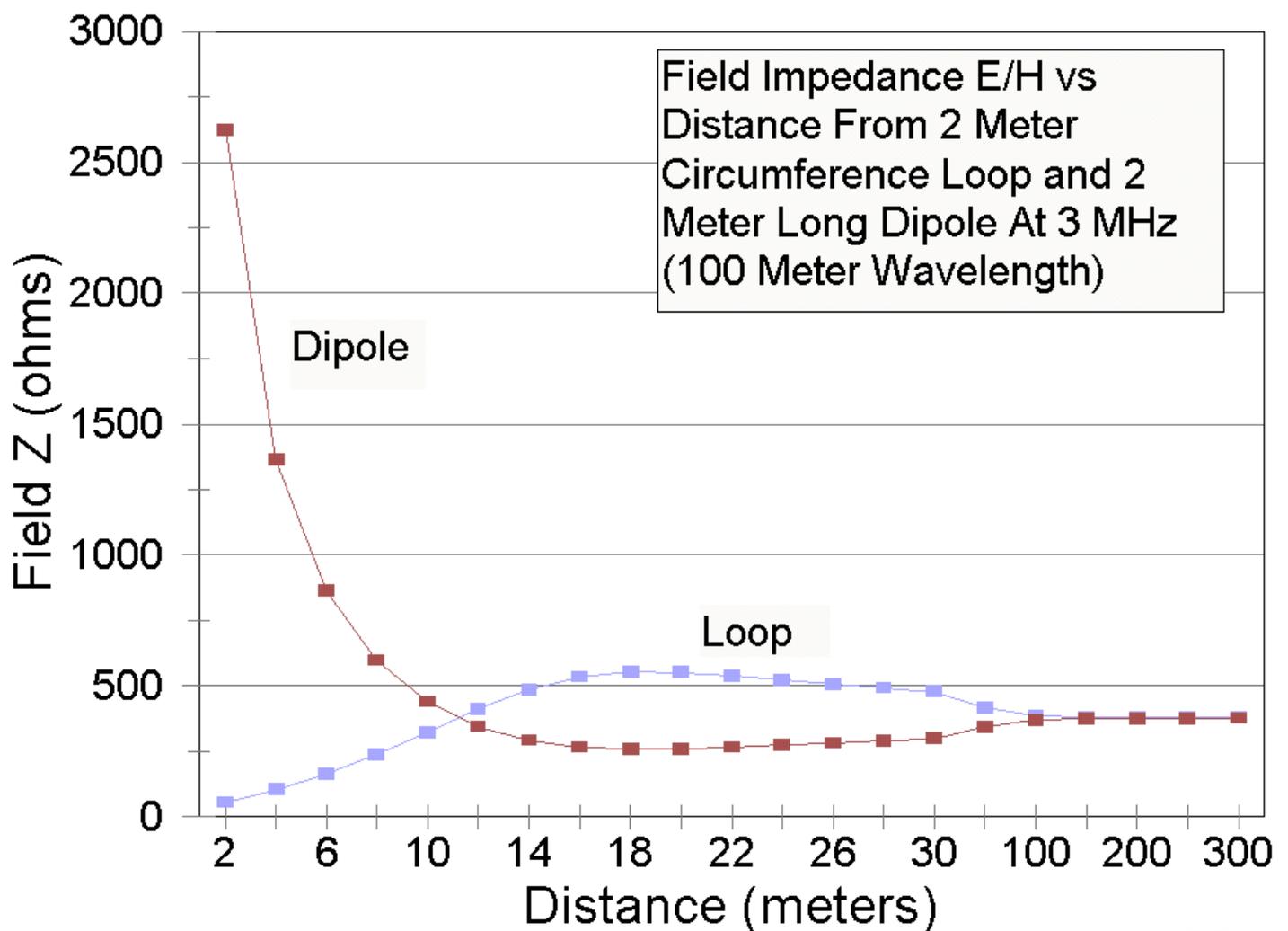
Since the distance of a wavelength in the above graph (thanks W7EL) is 100 meters, we can also considered the bottom scale as a percentage of a wavelength. We can see at about 11 percent of a wavelength (which would be about 50 feet on 160 meters), there is no difference in field impedance between a small loop and a small dipole. At distances beyond 50 on 160 meters, the loop actually has a *higher* field impedance than a dipole.

Losses

None of this has anything to do with losses being high or low in small antennas. Losses are directly related to the field density, and when we are close to any antenna the fields are very intense. Losses are not a field ratio problem, they are a field intensity related. In very small antennas, virtually ALL of the losses are related to reactance canceling and resistive losses in the antenna and any lossy media around the antenna!

We also must be mindful of the painful truth that we can not take either field to zero, or all radiation stops. By definition, radiation is an electromagnetic wave. We can't modify the field impedance of an antenna without changing the voltage and current distribution of the antenna.

Nearfield

The nearfield area is an area where the ultimate pattern is not fully formed, and where induction fields (from charge distribution and charge movement) have a noticeable effect.

It is possible, with large arrays of small elements, to be out of the induction field region but still find the area called the "nearfield" area or zone. Let's consider individual groups of elements as "cells", and the array a combination of small directional cells occupying a very large physical area.

Each cell has formed a radiation field. Depending on the size and type of radiator in each cell, the induction fields that charge distribution plays a role in may be attenuated so much as to be negligible, yet the radiation pattern of the entire array may not be totally formed.

This is the case with my phased Beverages and phased verticals, where the individual antennas making up the array are so distant that the effects of charge distribution (electric induction field, sometimes called the electrostatic field) or steady movement (considered at one infinitely brief instant of time, or magnetic induction field) have no effect. For example, at about 1 wavelength distance the electric and magnetic induction fields are negligible from either my circle of eight verticals or 780-foot Beverages, yet the pattern of the overall array established by the phasing of multiple cells is not fully formed. The pattern would only be fully formed several wavelengths from each array, where the distance between cells or elements is a small fraction of the viewing distance.

The total pattern of two 780-foot long Beverages spaced 350 feet apart is not fully formed even at distances of several thousand feet, yet nearfield induction effects are totally gone at much shorter distances. The field impedance is established, yet the antenna pattern is not.

The nearfield generally refers to or includes the area where "static" or induction fields still have a noticeable influence.

Fresnel Zone

The Fresnel (fre-nel, no "S" sound) zone is the area where pattern is still being formed. It may or may not include induction field areas.

Physically large arrays almost always have a physically large Fresnel zone. Even simple omni-verticals have a Fresnel zone extending out a few wavelengths. The field impedance may or may not have already been established in the Fresnel zone.

You may have heard about Fresnel zones during discussions of vertical antenna loss at low wave angles, or Fresnel lenses for lighthouses or other beacon lights.

Farfield

The farfield is the area where any changes in distance result in no noticeable change in pattern or field impedance. Losses are lower in the farfield area because field density is lower, not because we call it farfield.

Summary

There really isn't a clear distance where certain effects just abruptly stop. Transitions are smooth and gradual, because transitions are the result of distance....nothing else. These regions of effects are all distance related, and caused by the three distinct effects of charges moving, charge distribution, and charge acceleration.

Even if we could somehow make the field-impedance a certain value or electric and magnetic effects a ratio equal to farfield ratio, it would not eliminate or in the slightest amount reduce nearfield losses!

Losses are only related to the density of fields in any given media or environment, and nearfield or Fresnel zone losses decrease with distance because of the lower field intensity in any given lossy area, or wider cross-sectional area of lossy media carrying the same energy.

Larger antennas, in general, have larger boundary areas for electric and magnetic fields. With wider boundary areas, the system has less-concentrated fields and charge effects. Very small antennas of any type obviously have very concentrated fields, and the high field density or concentration of current or voltage in conductors are at the root of increased loss. The concentration of energy in a small area is the real reason losses are generally much higher in smaller antennas, and this has little or nothing to do with field impedance!

We can not mix fields and create a different field and make a small antenna become an "artificial large antenna", nor can we "filter" fields and remove the electric field to reduce electric noise.

This page has been visited **893** times since June 12, 2004

Traps

This article is from a talk I gave about antenna traps. It contains measurements of traps, performance of trap antennas using models of traps, and ideas on how to make trap antennas more efficient.

Try taking this Trap-Q test! Be honest.

1.)   Is it best to make the trap resonant close to the desired operating frequency?

2.)   Does bandwidth decrease with increasing trap Q?

3.)   Do traps create noticeable loss, perhaps one dB per trap typically?

4.)   Does higher trap operating Q always mean lower loss?

Coaxial Trap Designer by VE6YP (Tony Fields)
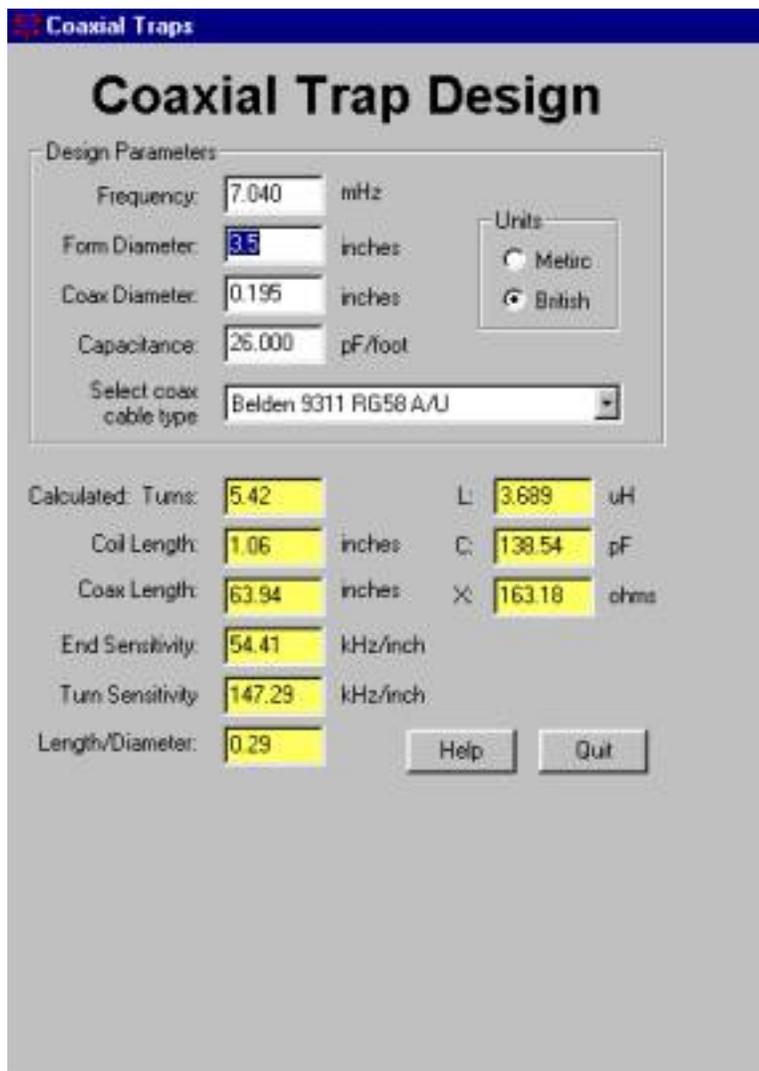
This is a good program to get you in the ballpark with a trap design. It was available as freeware. (Unfortunately coaxial traps are relatively lossy on the trapped frequency compared to other types.)

The software is available at **http://members.shaw.ca./ve6yp/index.html**

7.04 MHz   3.5 inch diameter form RG58/U into the VE6YP program yields calculated values of:

Calculated **Actual Measurement**

L= 3.689μH **3.116μH**

C= 138.5 pF **164 pF**

64 inches **59 inches**

Using the program TLA by N6BV (from ARRL), we would estimate capacitance of a 59" RG-58/U cable as:

R .22 –j143.61 or about 157 pF (Q=650)

Measuring a real-world stub, capacitance was 164pF (Q=590).

While that Q seems high, remember a typical transmitting-type air-variable capacitor has a Q of several thousand!

Coaxial Trap Articles and Programs use capacitance/ft multiplied times length….

26 pF * 4.917 feet = 127.84 pF in trap program

C164 pF measured. This error, 36pF low from 164pF, occurs because the transmission line making up the "coaxial capacitor" is not actually treated as a transmission line in the modeling program.

Fortunately the error is in a useful direction, because we can shorten the cable! Coaxial capacitors are really open stubs, and should be treated that way once they are more than a few degrees long.

CONCLUSION: The difference between TLA and an actual measurement was around 4%. This is very close, but the result has significant difference from the coaxial trap program since it only considers pF per foot as the capacitance. A longer cable (in fraction of a wavelength) results in greater error by using pF per foot. The error comes because a coaxial cable capacitor is really a stub, NOT a pure capacitor!!

traps antenas

Trap Measurements (at resonance)

Type	F MHz	R parallel	X
Coax RG-58	7.034	17,800	0
UT-141-75 semi-rigid	7.045	45,330	0
100pF 7.5kV & #12 wire	7.040	99,850	0
60pF 15 kV & #10 wire	7.040	250,000	0
60 pF vac & Copper tubing	7.040	300,000	0
Coax RG-58	3.700	23,200	0
Coaxial with fixed mica capacitor	7.040	21,660	0

Highest R parallel equivalent is best!! Lower Rp means more loss.

Trap Measurement summary:

- **Coaxial trap poorest**
- **Once #10AWG wire is used, not much improvement**
- **Space-wound bare wire makes best inductor**
- **Transmitting-type capacitors noticeably better than capacitors made from coax**

10 Meter (Tribander) Traps

Type	Freq	R parallel	X
Coax RG-58	29.00	13,800	0
Mosley TA-33	30.64	43,100	0
Mosley Pro-57	27.46	66,080	0

Cushcraft A3	**28.78**	**110,000**	**0**
HyGain TH-3	**29.67**	**140,200**	**0**

Traps are not all that bad when you plug them into models.

15 Meter (Tribander) Traps

Type	Freq	R parallel	X
Coax RG-58	**21.00**	**13,980**	**0**
Cushcraft A3	**21.43**	**76,270**	**0**
Mosley TA-33	**21.68**	**79,000**	**0**
HyGain TH-6	**22.23**	**142,000**	**0**

Trap Model
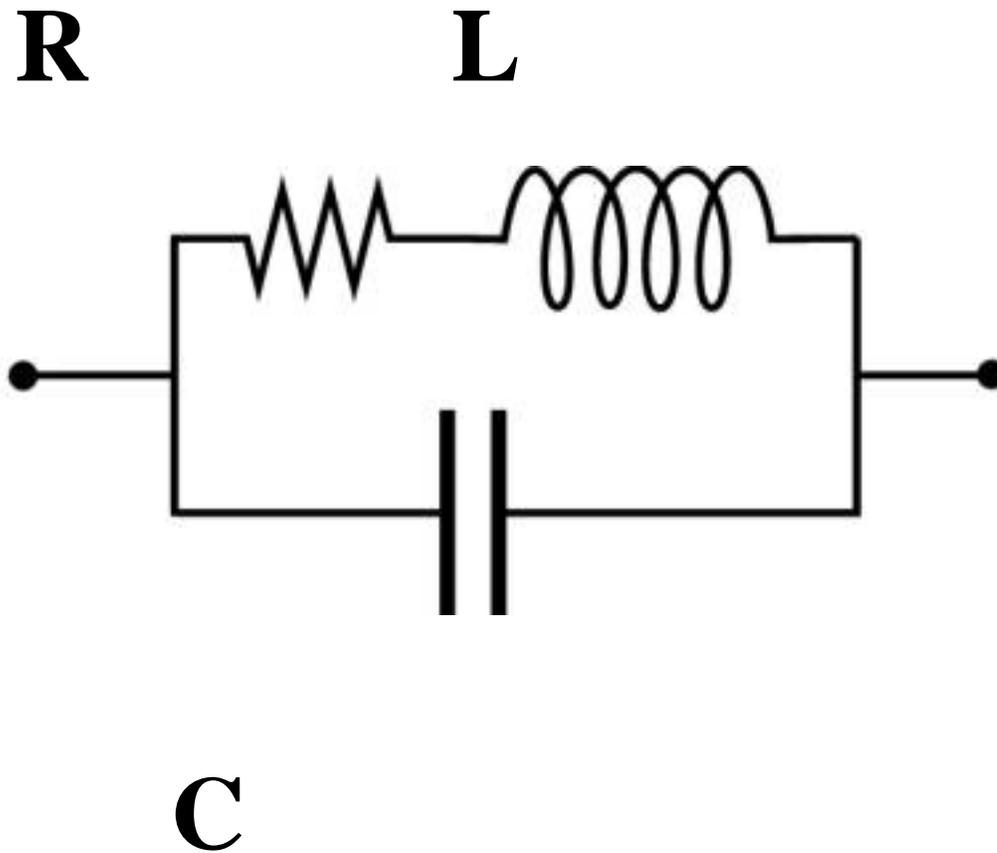
Measured Values Coax 7 MHz Trap

Freq	Imp	R	Xc	Q	L uH	C pF
7.04	17,800 j0	1.03	138	134	3.114	164

3.7	1.1 j 97	.6	283	88	3.114	152

Measured Values L/C 7 MHz Trap

Freq	Imp	R	Xc	Q	L uH	C pF
7.04	99,850 j0	.36	215	465	4.92	105
3.7	0.53 j 156	.25	409	294	4.92	105

SWR Bandwidth

Freq	3.7 MHz		Source #	1
SWR	1.14		Z0	75 ohms
Z	85.54 + j 2.565 ohms			
Refl Coeff	0.06754 at 12.77 deg.			
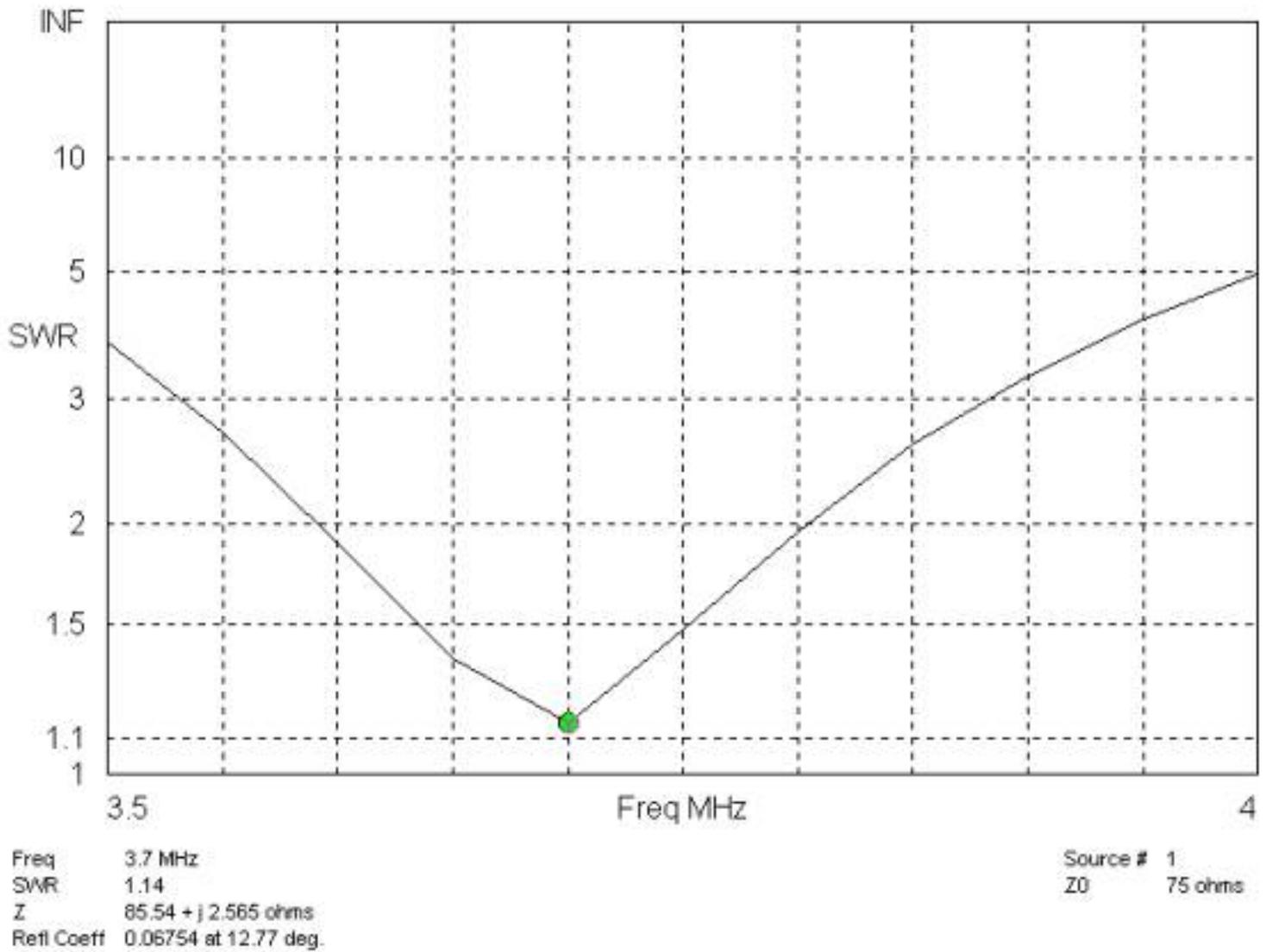
7 MHz RG-58 TRAP

80 m 75 ohm VSWR

EZNEC #12AWG dipole

Coax trap 80m 2:1 VSWR   ~210 kHz

**Total trap loss = 0.05 dB**

RG-58 TRAP, 75 ohm VSWR, 40 METERS

VSWR BW

Freq   7.05 MHz   Source #   1
SWR   1.32   Z0   75 ohms
Z   99.11 + j 0.5849 ohms
Refl Coeff 0.1385 at 1.2 deg.
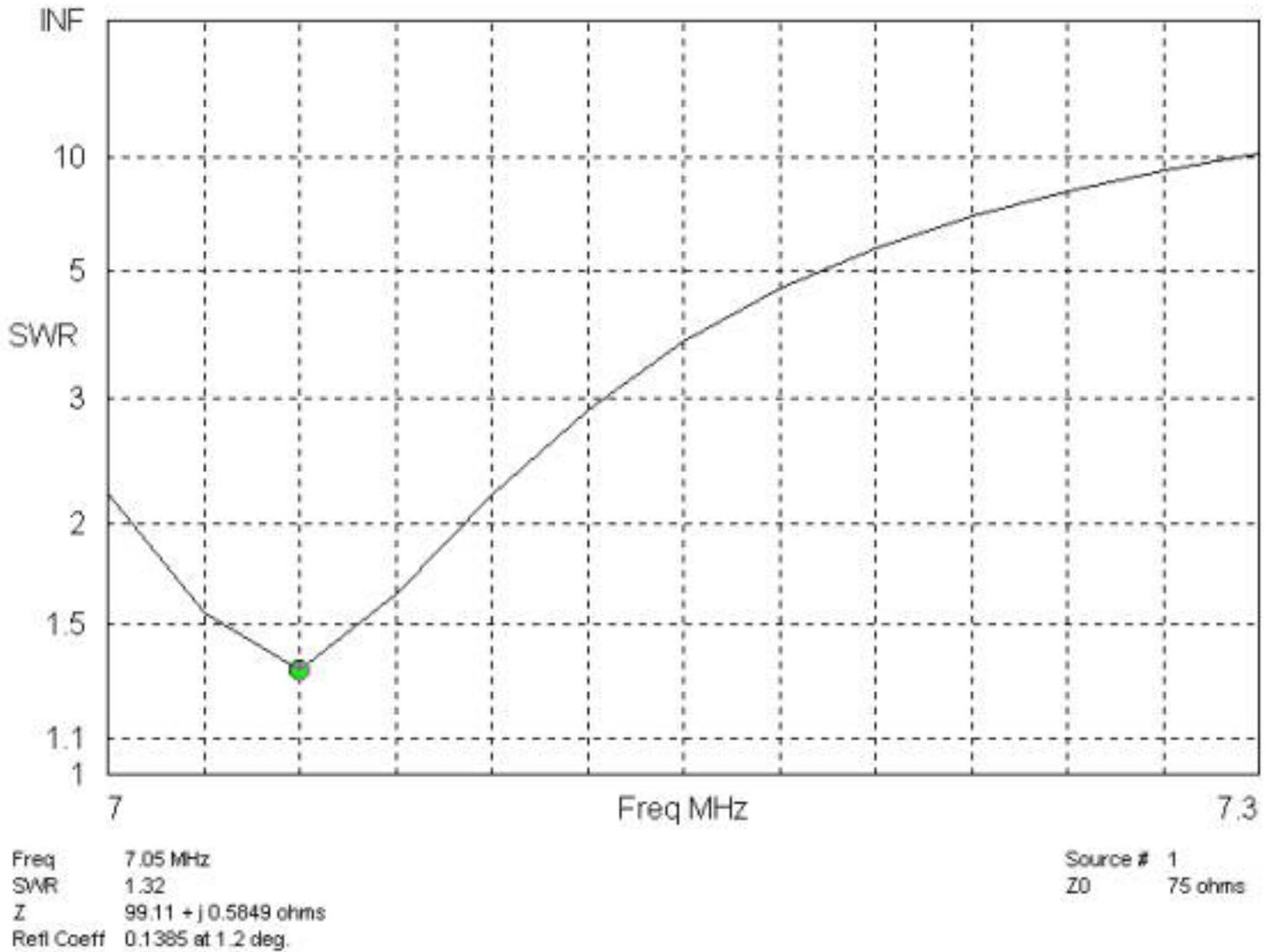
Coax trap 40 meter 2:1 VSWR ~ 80 kHz

Total coaxial trap loss at resonance on 40m= 1.6 dB

Total coaxial trap loss 100kHz off-resonance (at 7.15 MHz)= 1.06 dB

Note that loss is maximum at trap resonance!!!

Never make a trap resonant on the desired operating frequency!!

W2LH ARRL Handbook Trap Design

100pF #12awg Miniductor trap
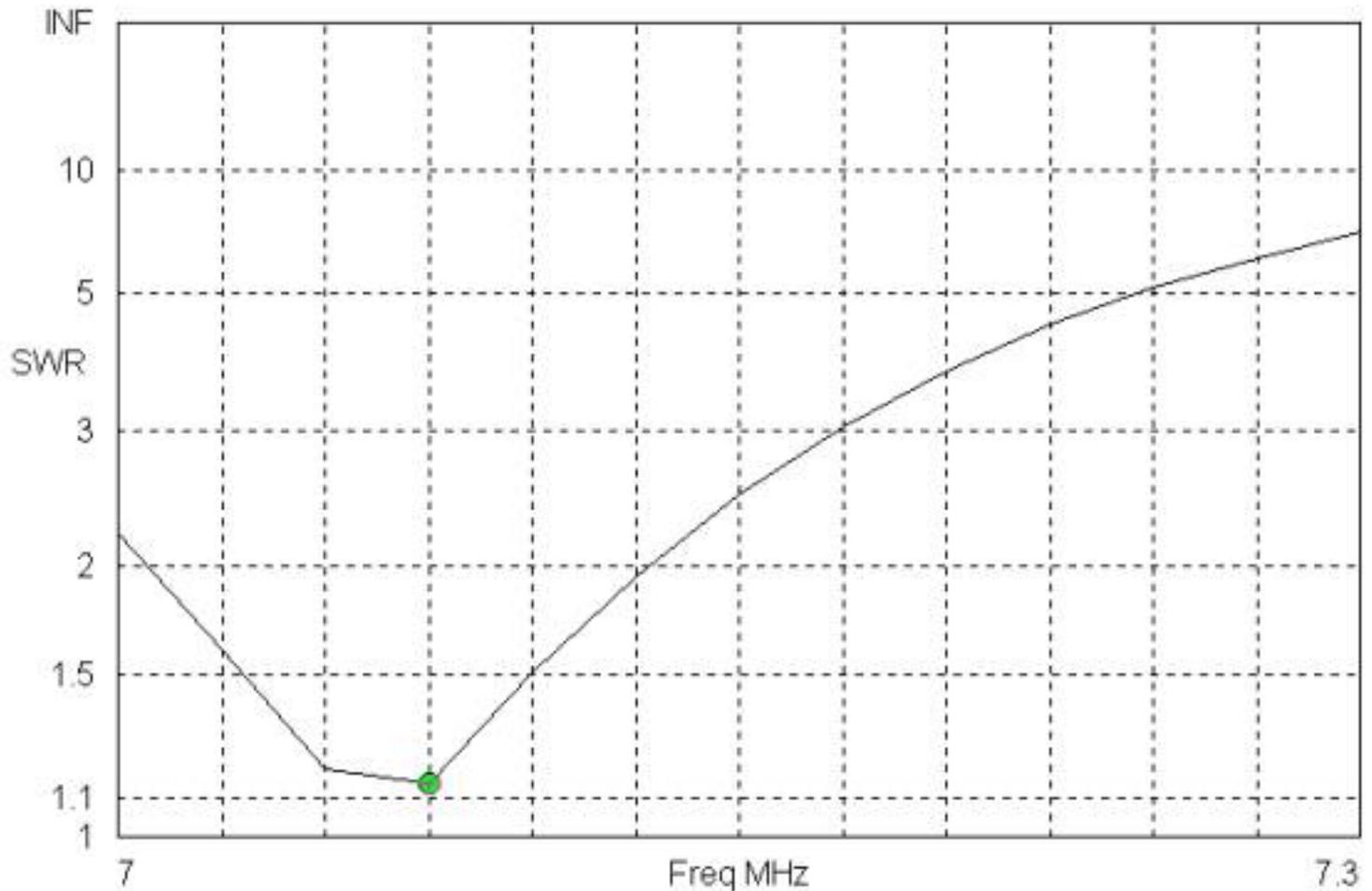
Freq	7.075 MHz		Source #	1
SWR	1.14		Z0	75 ohms
Z	77.73 + j 9.703 ohms			
Refl Coeff	0.06587 at 70.64 deg.			

40m 2:1 VSWR ~120 kHz

Total loss = 0.24 dB

W2LH ARRL HANDBOOK TRAP 80m

VSWR


```
Freq        3.65 MHz                          Source #   1
SWR         1.17                              Z0         75 ohms
Z           79.53 - j 11.05 ohms
Refl Coeff  0.07707 at -63.59 deg.
```
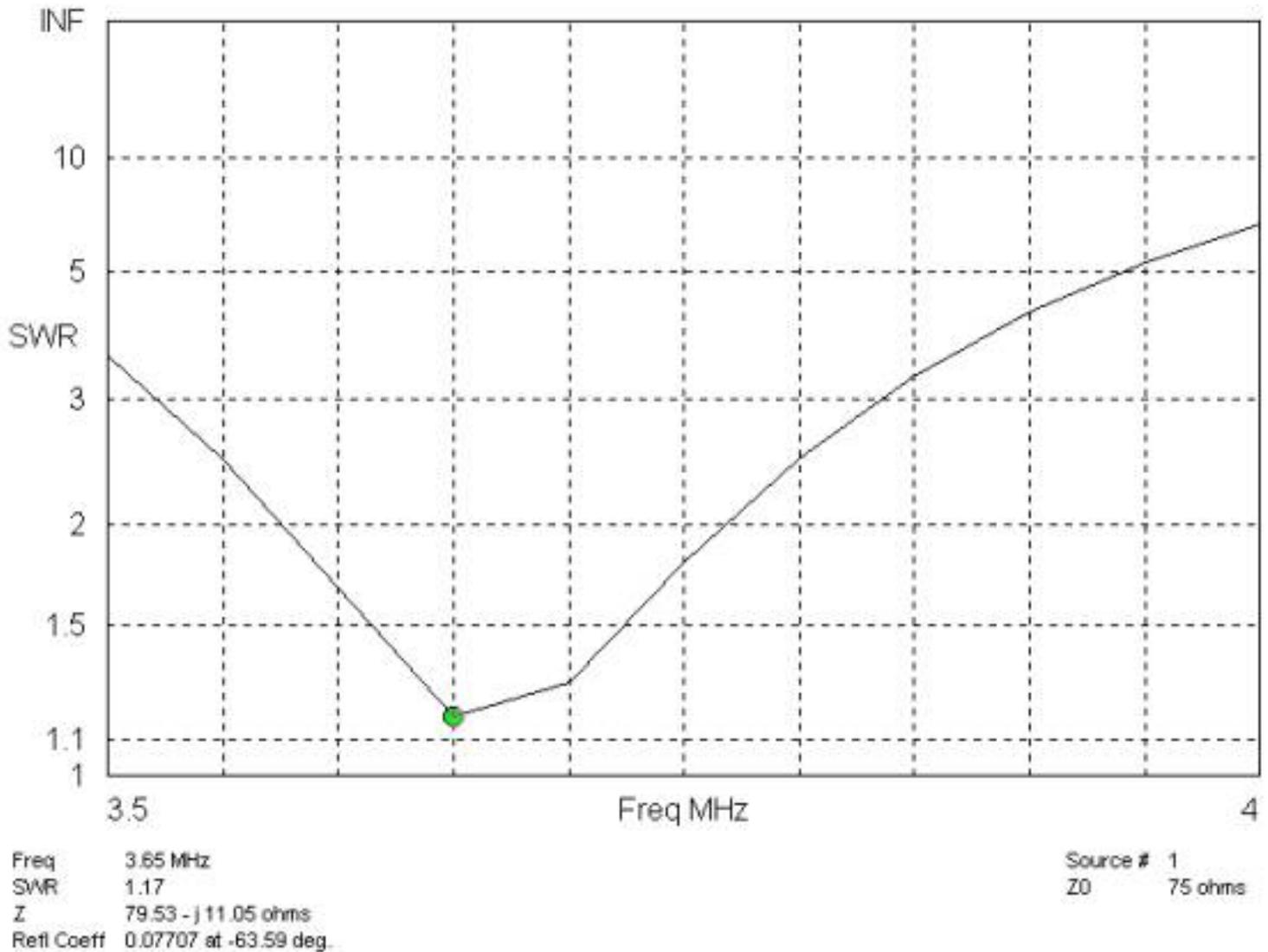
80m 2:1 VSWR BW ~ 200 kHz

Total trap loss = 0.026 dB

What happens if trap is not in band?

VSWR Bandwidth of 6.51MHz trap in 80/40 dipole

Freq   7.10001 MHz
SWR   1.047
Z   78.32 + j 1.237 ohms
Refl Coeff 0.02312 at 19.97 deg.

Source #   1
Z0   75 ohms
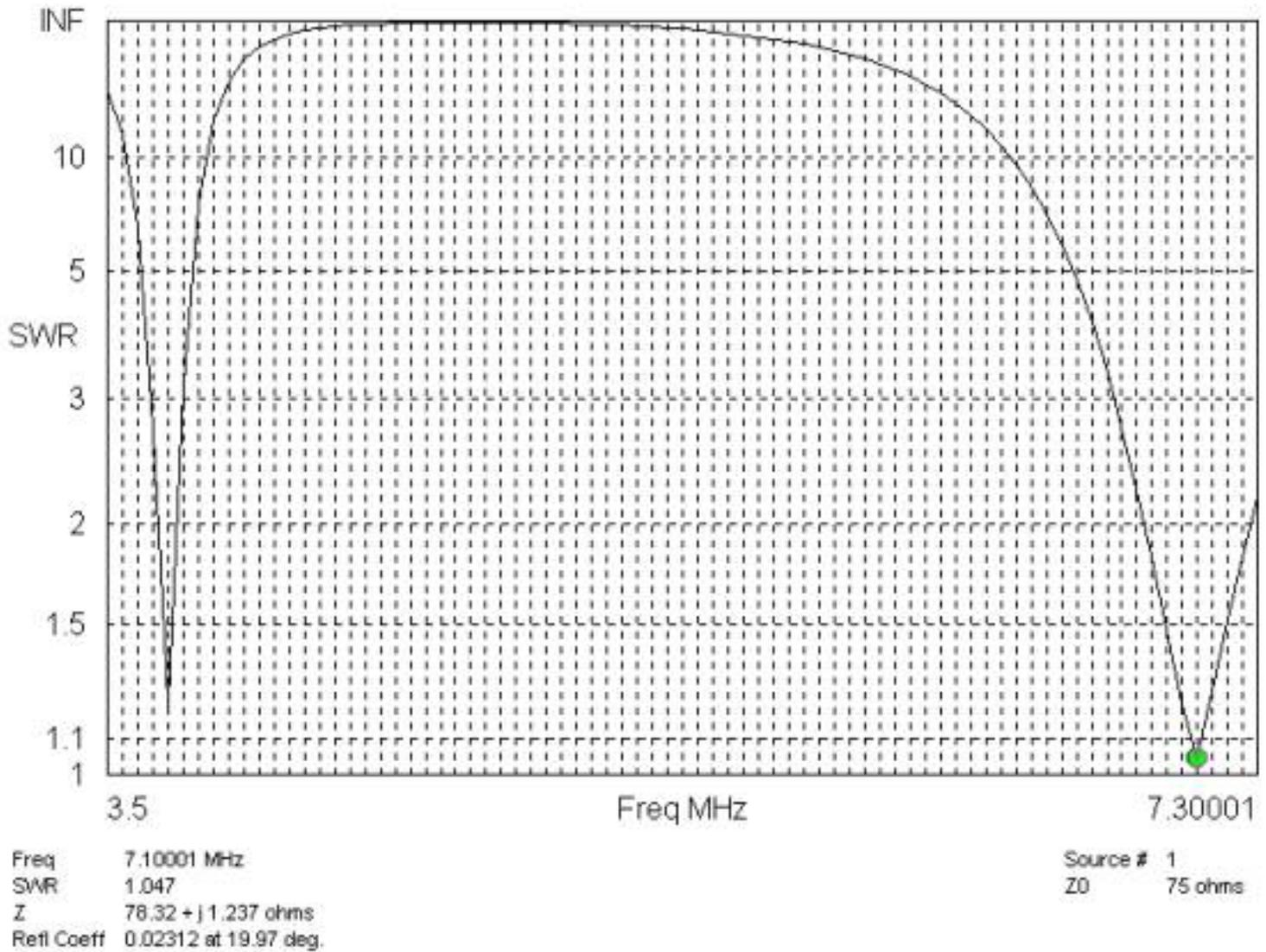
Trap at 6.51 MHz Q=130

Loss at 7.15 MHz = 0.314 dB

Loss at 3.7 MHz   = 0.324 dB

This is a 104-foot long antenna, with very poor Q traps, and loss is less than .4dB! The reason loss is low is we have moved the trap slightly out-of-band.

6.15 MHz Q=130 TRAP 40m VSWR

Freq	7 MHz		Source #	1
SWR	1.6		Z0	75 ohms
Z	78.7 - j 36.25 ohms			
Refl Coeff	0.2307 at -70.9 deg.			
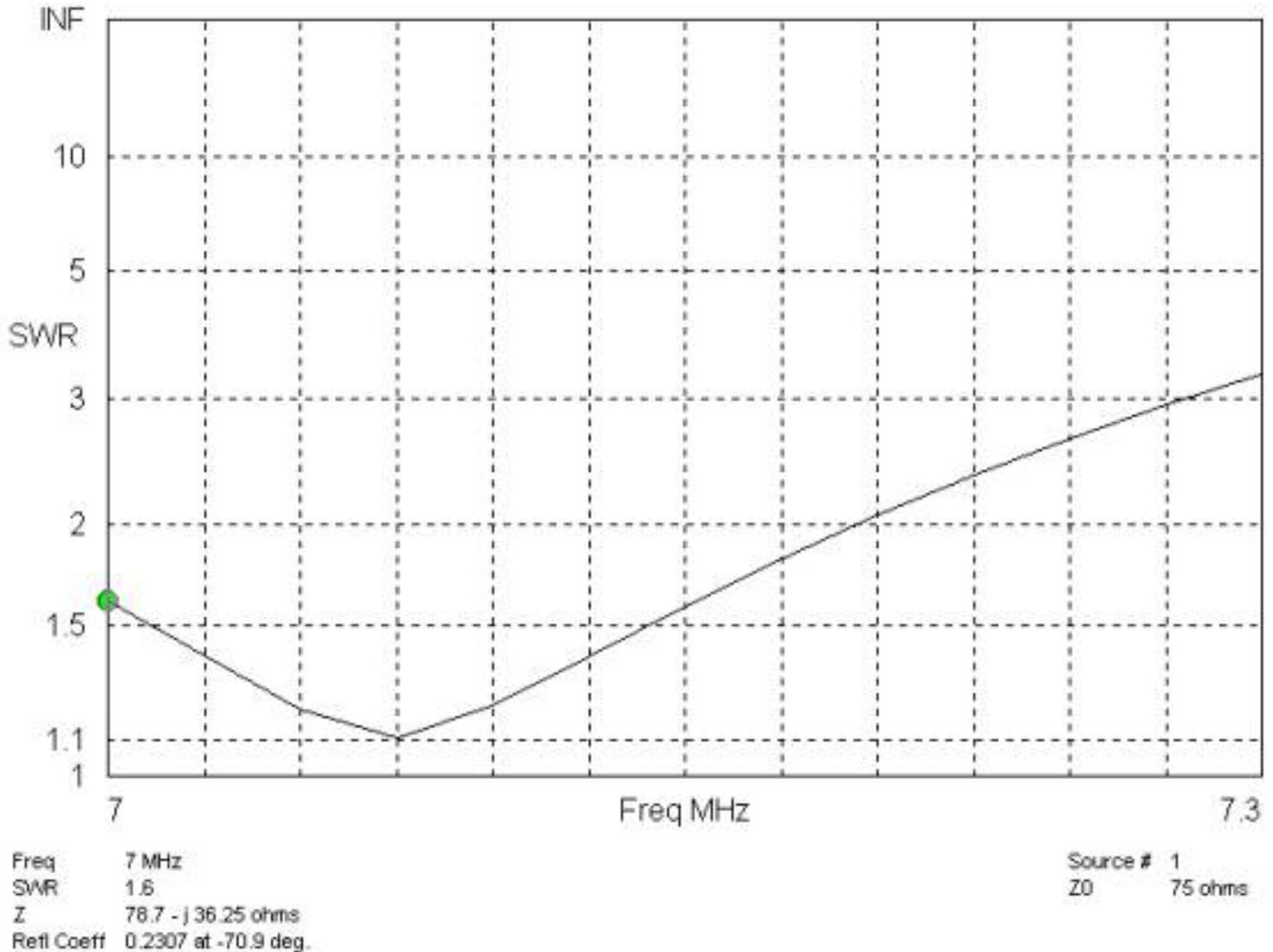
7 MHz 2:1 VSWR BANDWIDTH ~200kHz

Trap Q at resonance = 130   7 MHz loss ~ .3 dB

6.15 MHz Q=130 TRAP 80m VSWR

Freq	3.775 MHz		Source #	1
SWR	1.084		Z0	75 ohms
Z	78 - j 5.387 ohms			
Refl Coeff	0.04026 at -58.91 deg.			
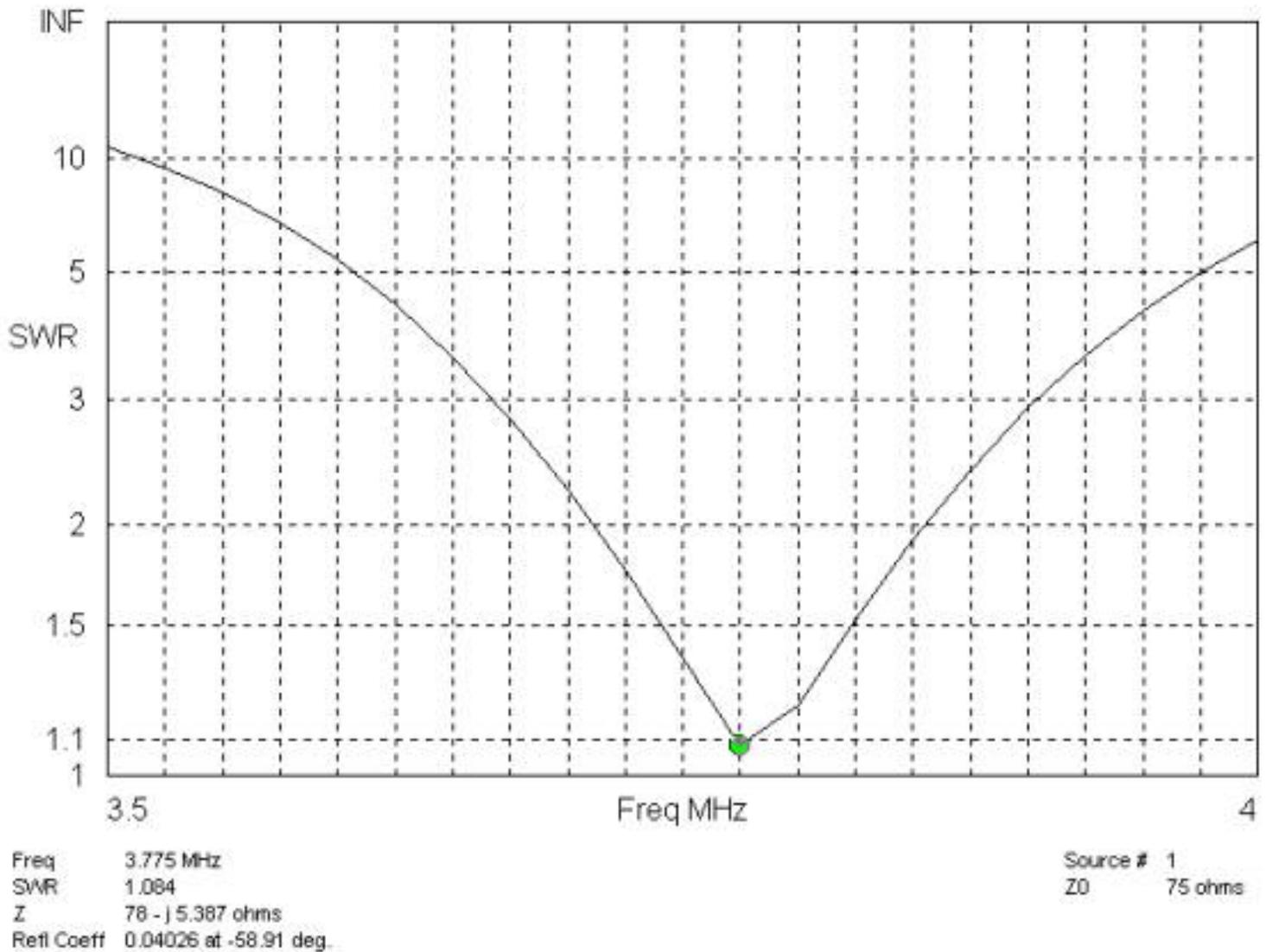
80M 2:1 VSWR BW ~130 kHz

Loss at 3.7 MHz = 0.324 dB

1.) Is it best to make the trap resonant close to the desired operating frequency?

NO! Loss is highest when the trap is resonant at the operating frequency!

2.) Does bandwidth decrease with increasing trap Q?

NO! Bandwidth is a function of many variables, trap Q actually has one of the smallest influences on BW.

3.)   Do traps create noticeable loss, perhaps one dB per trap typically?

NO! Even the worse traps (coaxial traps) in the worse possible condition of operation are only 1.6dB loss for BOTH traps!

4.)   Does higher trap operating Q always mean lower loss?

NO! Loss depends on many factors, including trap resonant frequency.

Conclusions:

- Trap loss has been greatly exaggerated by advertising hype
- Traps should not be resonant at the actual planned operating frequency
- Coaxial traps are more lossy than articles conclude
- Coaxial stubs used as capacitors can not be calculated using pF/ft unless the stub is a very small fraction of a wavelength long (less than a few electrical degrees)
- Coaxial stubs have low Q (are relatively lossy) compared to normal lumped components.

Building a Current Meter

[Home]

It takes a special meter to measure current accurately in the presence of strong electric fields, and to not disturb low-capacitance high impedance systems. The meter described below is suitable for measuring RF current accurately in high voltage applications, such as short antennas. It is also accurate in normal applications

Another current meter project link (*not* suitable for mobile antenna measurements, however!)

http://www.ifwtech.co.uk/g3sek/clip-on/clip-on.htm

Low Capacitance, Voltage Immune, RF Current Meter

The meter is constructed with a 100uA all plastic case meter. The meter scale is plastic. The movement and all metallic areas are small. The lack of large metallic components minimizes stray capacitance added to the circuit under test by the proximity of the current meter.

Low capacitance ensures the meter has the least possible effect on the circuit being tested.

This meter has a 1.5" toroid current transformer on the rear. Just behind the meter is a calibration pot and all the circuitry. The toroid and circuitry is hot-melt glued to the meter with only the screwdriver calibration pot exposed.

I did not add a Faraday shield because the shield would increase the capacitance. Since this entire meter floats above ground, there is no need for a shield anyway, balance is not critical.

The meter has a low threshold Schottky diode detector, a series resistor for calibration, and standard .1uF bypass capacitors.
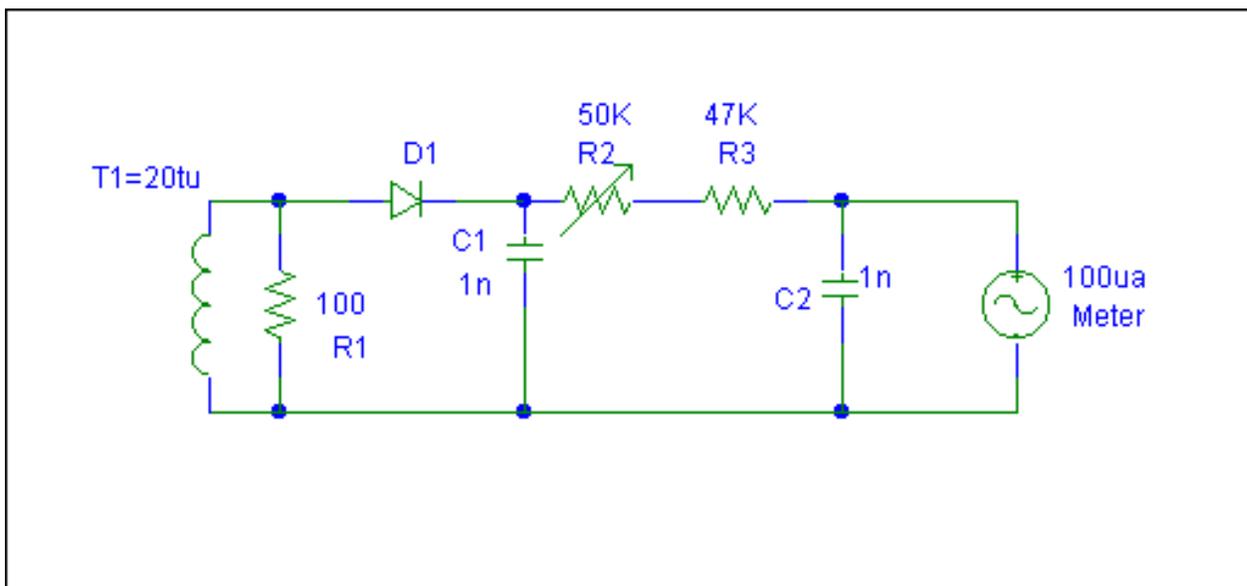
T1 is a current transformer. When the single turn primary (a whip or mast) has 1 ampere, the secondary will have

.05 amperes (inverse of the turns ratio). R1 flattens the response and limits the voltage. With 100 ohms we have .05*100=5 volts RMS. The peak dc voltage is 1.414 times 5= ~7 volts. C1 is a filter capacitor for the RF pulses, R2 and R3 set the FS range. With 100uA meter the resistance is 10,000 ohms/volt. 7 volts requires 70K ohms, which will be approximately midway on R2.

Dissipation in R1 is .25 watts from .05 amperes (T1 current) times 5-Vrms (secondary voltage of 5v with .05a flowing through 100 ohms).

Calibration

The meter is tracked in a test fixture for linearity. The actual fixture used is shown below.
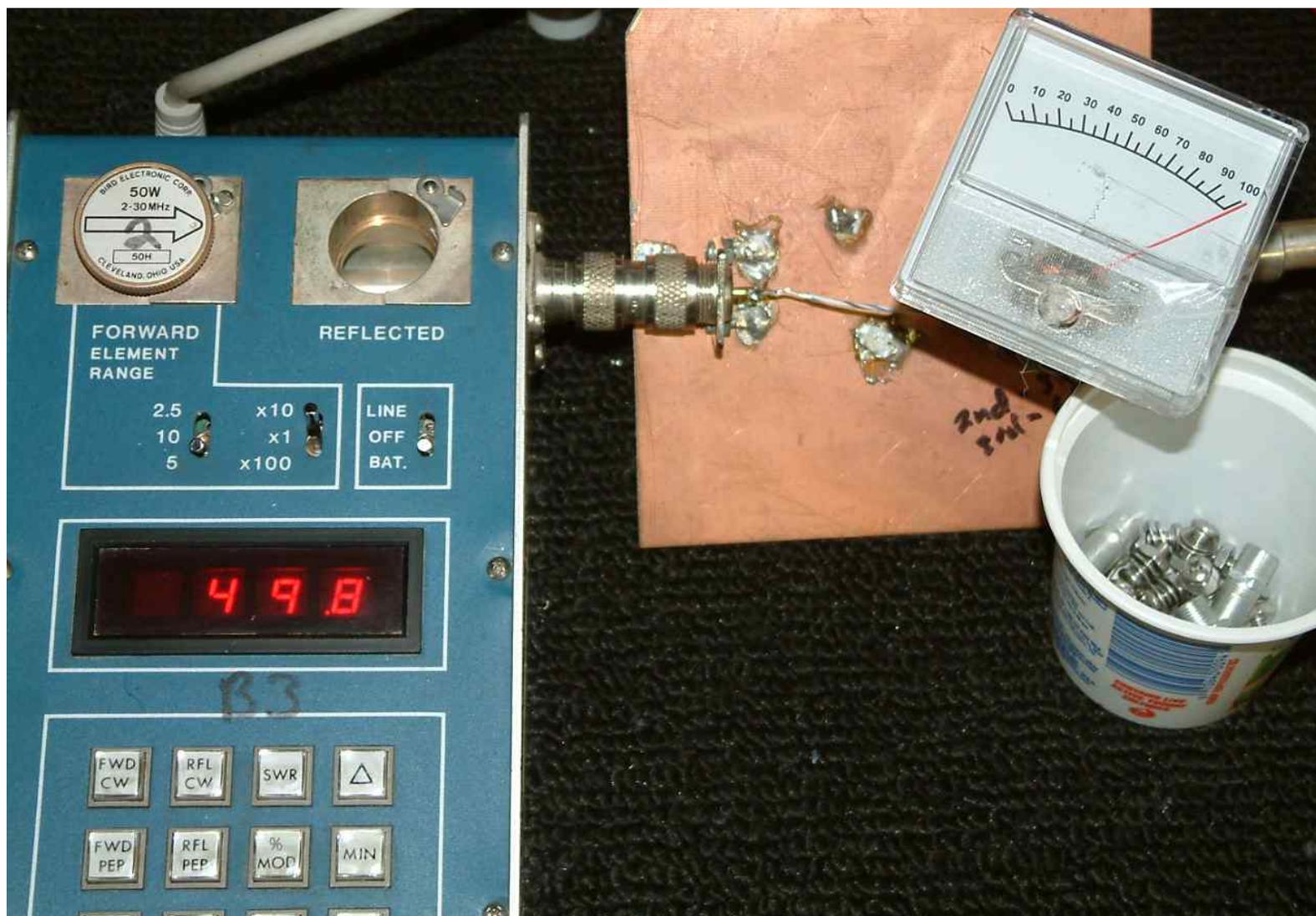
The test fixture consists of two UHF female connectors soldered to a sheet of PC board. There is a single #16 Teflon wire running between the center pins.

Note the numbers "2" and "B3" on the meters. My elements and meters have calibration charts that correct for linearity errors in tracking and readings. In this case I've applied 50 watts to a precision 50 ohm load, making wire current 1-ampere.

The meter is calibrated and scale was tracked from 1 ampere down to .2 amperes.

FS accuracy is not required in comparison measurements, since the meter references against itself. Linearity within a few percent is important.

This type of meter is much more reliable and linear than thermocouple RF ammeters, and perturbs systems much less. Stray capacitance added to the system being tested is very small, because only the proximity of the meter and the compact wiring area. Contrast that to actually connecting a meter in the line with the associated lead lengths and capacitance of the meter itself, and the advantages of a transformer coupled meter become apparent.

[Back to Mobile and loaded antennas.](#)

RRAA_post

[ ]

I really begrudge the necessity of posting once more on this newsgroup, and particularly on this thread. However, I'm dismayed and disgusted by postings being made by Yuri and Cecil in other forums in which they're claiming that measurements I made agree with theories and predictions they allegedly made, and that my measurements therefore validate their theories. (A quick scan of this thread shows that they even made the false claims here, after I had quit posting.)

One of the postings is the following, made by Yuri on the eHam TowerTalk group, Nov. 21, topic "Trap Resonance":

"Why don't you mention what W5DXP came up as explanation for Tom's errors, why don't you mention what W7EL measured and that it was what I predicted based on available information and was right on - the 5% difference for the base loading coil?"

and this one, posted by Yuri on the EHam.net forum topic "Re: Current in Antenna Loading Coils" on Jan. 7:

"I will leave it here, as the rest of it. W7EL, author of EZNEC measured toroid coil and found that it HAS different current at its ends, roughly proportional to the part of antenna that it replaces."

and this one, posted by Cecil on that same group on Jan. 12:

"Roy's data clearly illustrates the phase shift through the coil. ARCCOS(Iout/Iin) gives an estimate of the phase shift (assuming forward current and reflected current are of equal magnitude). In Roy's experiment, Iout/Iin was about 0.95. ARCCOS(0.95) equals 18 degrees, an approximation for the phase shift through the coil."

As you'll see below (or by looking up the original thread), the first of my two measurements, for an antenna shortened an equivalent of about 18 degrees, resulted in 3% current attenuation across the coil (not 5%), and zero phase shift (not 18 degrees). The second test, where the

antenna was shortened more than 33 degrees, measured 5% current reduction and no phase shift. The method used in the above quote predicts more than 16% amplitude reduction and 33 degrees of phase shift for the second test. There's no way my data "clearly illustrates" Cecil's explanation. To say that it does is a pure fabrication.

I feel compelled to respond to these fabrications, and put the record straight. I'll do it here, since this is where my measurements were originally posted.

I made two sets of measurements of the current into and out of a toroidal inductor at the base of a vertical antenna. The details of the measurement method and the measurement results were posted here, on this newsgroup, on this thread. Pictures of the setup were posted on my web site, with a link posted here.

Before I posted each set of measurements, I asked for predictions of the results, so that alternate theories could be tested. (I was criticized for doing this -- it seems that the preferred method of testing a theory is to look at the results first, then adjust the theory to fit.) Yuri made a prediction (actually, two different ones) for the first set of measurements that didn't accurately predict the results. When I calculated the predicted result for the second set of measurements using the same method he had used for the first prediction, he retracted any claim that the method would be valid. (Exact quotes are below.) He didn't make any prediction at all for the second set of measurements.

Cecil made a number of vague predictions which he later contradicted or retracted. At the time the second set of measurements were posted, he had made no prediction at all.

Perhaps Yuri and Cecil have, after the measurements were posted, developed theories to explain the results. As of the time the measurements were posted, they hadn't. I highly recommend that anyone considering their alternative theories to find where they have calculated the results which agree with my measurements (particularly the second one, which was designed to produce a testable difference), and how they derived the equations used for the calculation.

Following is a summary of some of the exchanges between Yuri and me on this thread last November. The entire thread, "Re: Current in antenna loading coils controversy" and variants, is available for viewing at groups.google.com.

------ Summary ------

Here's what really happened. The following quotes are directly from the
google archives of the rraa thread.

I made two sets of measurements. The first had the inductor connected at
the base of a 33 foot vertical. But the vertical was mounted about 1/4"
from a four foot pipe, which reduced the base reactance.

Here, I was asking for predictions for my FIRST measurement -- the one
with the vertical mounted on the pipe.

Yuri posted on Nov. 9:

"In that case,
If the feedpoint current was at 0 deg of the radiator length, and coil
replaces 18 deg of wire, the cos 18 deg = 0.951 which should make
difference, drop in the coil current 5% (or half, 2.5 deg?)
Providing current maximum is exactly at the bottom end of the coil."

and later on Nov. 9:

">Incidentally, I take it that your prediction for the setup I did measure
 >includes an 18 degree phase shift of current from input to output of the
 >inductor?
 >
 >Roy Lewallen, W7EL
 >

Yes, I used Cecil estimate/calculation and taking
cos 18 = 0.951056516 which is 4.8943483%"

So now we have his prediction, using the "cosine rule". The measurement
I made showed about 3% current reduction from input to output, but with
about 2% (the same amount within measurement error) also occurring when
the antenna was replaced by a series resistor and capacitor -- that is,
no antenna at all. So the 5% prediction was wrong. His prediction of 18
degrees phase shift, which wasn't present, was also wrong. When asked
for the justification for the "cosine rule", he never offered any, so
its origin remains obscure.

However, I saw that the value was too small to be convincing, which is
why I devised the second test. The second test used a more ideal
antenna, with more of the antenna being "replaced" by the inductor. The
"cosine rule" would predict more than 16% reduction, and more than 33

degrees of phase shift.

Before I gave the results from the second measurement, I posted the predictions which had been made, as I understood them. Since Yuri had invoked the "cosine rule" for the first test, I naturally assumed it would also apply to the second. (This is simply applying the equation Yuri used in his Nov. 9 posting to the second antenna setup. It's also the equation now being used by Cecil, as shown in his quote from the eHam group.) So in my posting I said:

(Quote from my posting on Nov. 11):

"**Yuri's method predicts a reduction of output current magnitude of 16.5% and a phase shift of 33 degrees."

to which Yuri responded, also on Nov. 11:

"It is not my theory. My argument with W8JI and his followers: is the current in typical loading coil in quarter wave radiator same at both ends or does it drop with distance from the feedpoint. I have made temperature observations, W9UCW measured the difference, W5DXP provided some explanation. Based on Cecils analysis of data you provided, and on my understanding of the phenomena I guestimated drop in current in your setup. No theory, no mathematical procedure (yet) just attempt (using degrees replaced by coil in a radiator) at explanation of what is happening. I will measure things myself, try to verify previous measurements and then come up with conclusions and "theory". So far Cecils (and ON4UN book) theory seems to be closest to the truth. . . "

So now, Yuri has disclaimed the "cosine rule". /He made no other prediction of the results of the second test./

In summary, Yuri first stated that the "cosine rule" can be used to calculate the current drop. That would have predicted over 16% current reduction in the second test. Then he retracted his claim that that theory would work, before the results from the second test were posted, and never made any other prediction. He never predicted the 5% result which was measured, as he's now claiming.

And if you can find a numerical prediction anywhere in the thread which Cecil made and stayed with, my hat's off to you. $100 goes to the first person who can point to any prediction made by either Yuri or Cecil before the second measurement results were posted that predicted second measurement results of 5% magnitude and zero phase shift. (In the case

of Cecil, this would have to be a prediction that wasn't later modified or retracted before the second set of results were posted.)

My measurement results are consistent with the fact that the currents into and out of a physically small inductor are equal. The small magnitude difference I measured can be explained by stray capacitance on the order of 7 pF from the output to ground and/or the probe -- not an unreasonable amount to expect. In no way do my measurements support the odd theories being proposed by Cecil and Yuri, and any statement that they do is completely false.

Roy Lewallen, W7EL

Coaxial line and shielded wires

[[Home](#)]

related pages at [antennas](#)

Concentric Transmission Lines (Coaxial Cable)

When the shield is more than several skin depths thick coaxial cables, single-wire shielded cables, or concentric lines become the electrical equivalent of three wire concentric transmission lines. Electrical rules require the center conductor and *inside* of the shield to always carry equal and opposite RF currents.
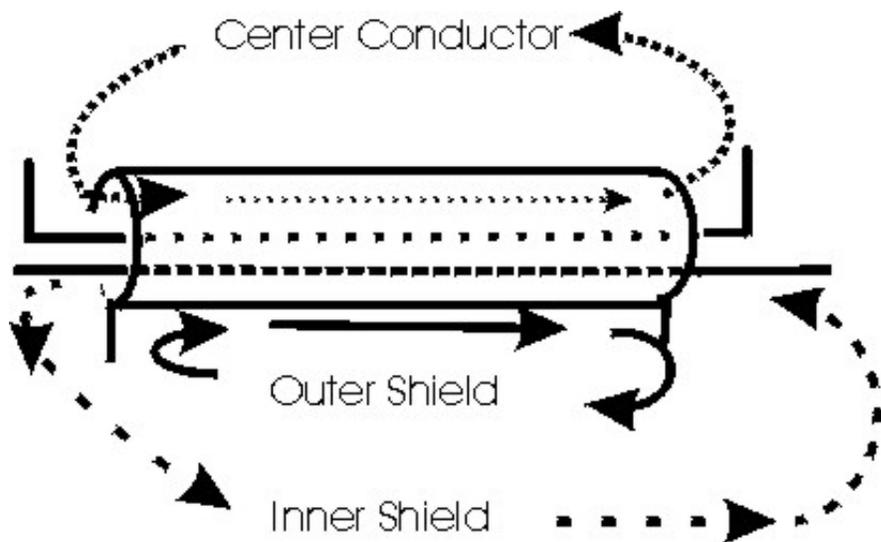
Notice in the drawing above the center and inside of the shield carry equal and opposite direction RF currents. This ALWAYS is the case when the shield is several skin depths thick. We cannot force anything else to happen!

In the drawing above and below, the outside of the shield is isolated by skin effect. It behaves like a separate transmission outer conductor. Skin effect prevents any current, voltage, or field (even magnetic) from penetrating the shield when the shield is many skin depths thick. Only the breaks at the ends connect the inner and outer shield conduction layers.
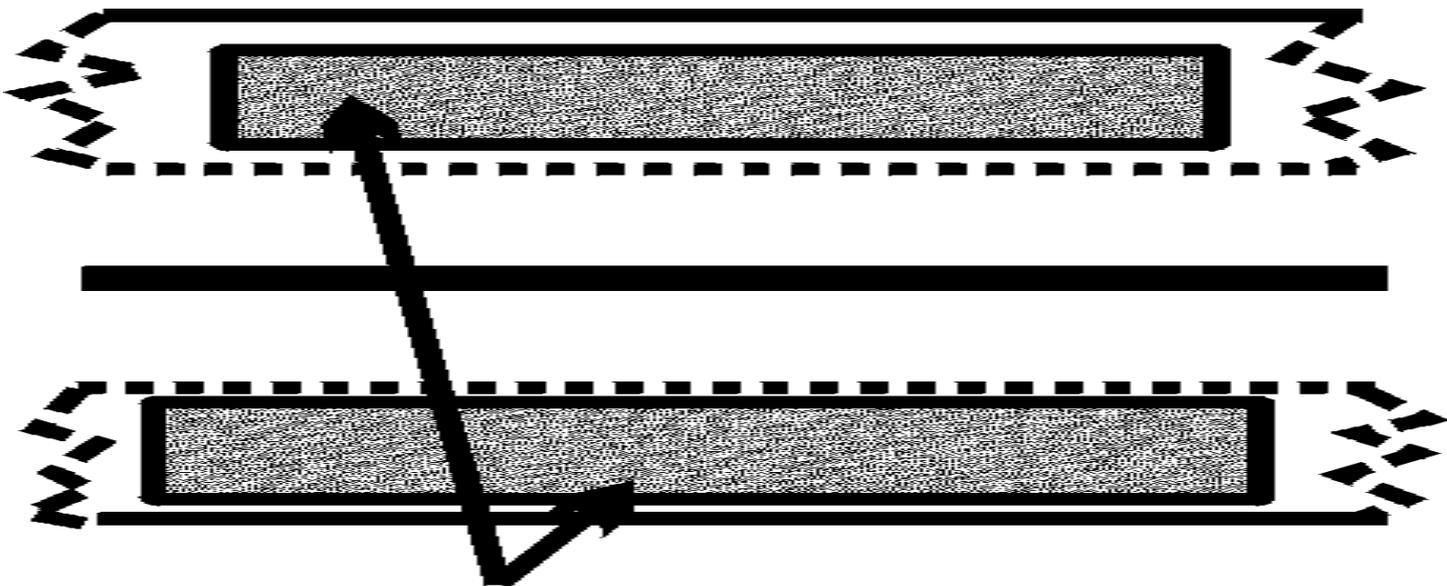
SKIN DEPTH AREA. Nothing passes.

The only egress or ingress points for voltage and current are where discontinuity in the shield might occur through an intentional or accidental break in the shield.

Radiation In or Out of Shielded Lines

To operate without radiation, coaxial cables require equal and opposite currents in the shield and the center conductor. The outside of the shield must not have an RF potential difference along it's length. We cannot have voltage gradients (electrical fields) along the length, or the outer conductor layer will conduct currents. This means both ends of the cable should have zero volts to earth or the environment the cable operates in.

If we feed exactly equal and opposite current into the coax, currents have no choice but to flow *INSIDE* the cable shield! The outermost conductor can be treated as a single wire, since everything else is "hidden inside". There will be no "extra" current to flow over the outside of the shield. Because of this rule, we do not need to ground the shield to prevent radiation. The trick here is we need to have equal and opposite currents. When the shield has potential difference along the length we can greatly reduce outside currents by making the outside of the coax have a high impedance. We can do this by selecting a proper cable length, by adding sleeves of soft-iron magnetic materials or winding the cable in a coil with or without a core to form a choke. *Skin depth prevents the inside of the cable from "seeing" what is done on the outside.*

Most interesting in all of this is the claim or idea the "shield" of a [shielded loop](#) allows magnetic fields to pass while filtering electric fields. The same rules apply as with coaxial lines. The shield actually becomes the antenna, and it couples to the inside only at shield gaps. If the gaps aren't exactly opposite the grounded part of the loop and inner conductors don't exit exactly at that grounded point, the shield actually UNbalances the loop!

Magnetic Loop Antennas Receiving

[[Home](#)] [[Up](#)]

(Please read the [Radiation and Fields](#) page)

Small Receiving Loops

Small loops are often referred to as "magnetic radiators". Folklore claims a small "shielded" loop antenna behaves like a sieve, sorting "good magnetic signals" from "bad electrical noise".

Nothing is further from the truth! At relatively small distances a small magnetic loop is more sensitive to electric fields than a small electric field probe.

The ratio of electric to magnetic field sensitivity is sometimes called the "field impedance". An antenna with *high field impedance* has a dominant electric field while an antenna with *low field impedance* has a dominant magnetic field when compared to free space field ratios at a large distance from the antenna.

Although fields have different ratios close to the antenna, at distances of *about $\lambda/2$ the field impedances of all but very large antennas are virtually indistinguishable from each other*. Once we understand the basics, we might picture our antennas differently. We might find antennas don't respond to signals and noise like we thought, but at least we won't be as susceptible to the "noise" caused by rumors and folklore!

Loop Antenna Fields

It is the response in within $\lambda/10$ distance from the antenna that gives small "magnetic loop" and "electric dipole" antennas their names.

Very close to a small loop antenna (but not necessarily near the open ends of the small loop where the tuning capacitor is) the magnetic field dominates. Current is essentially uniform all around the loop circumference, while voltage has a nearly straight-line increase as we move to the tuning capacitor area. Since most of the area has uniform current and only a small concentrated area has the highest voltages, the magnetic field clearly dominates over most of the loop area.

Magnetic fields are effects derived from the actual moving of charges (current flow). The magnetic effect is related to current, and if it is proportionally large compared to voltage the field is described as having a "*low*" impedance. This is similar to the description used in circuits, where a system with high current and low voltage

is said to have "low impedance".

Small Dipole Fields

Near a small dipole or monopole the electric field dominates. A short antenna has to have very high voltages all along it's length. It normally has highest current only near the feedpoint, with current tapering down in a straight line to zero at the antennas open end. The dominant effect we deal with is from the very high voltages that appear all along the antenna. We say the antenna has a "*high field impedance*".

In the electrically small antennas described above, the dominant field descriptions only apply within $\lambda/10$ distance!

Radiation

The induction fields described above are NOT what radiates or receives at a distance. The induction effects are a necessary byproduct of:

1. Uneven charge distribution (a difference in voltage) causing a physical force on other charges. We call this effect an **electric field.**
2. Moving charges (a current) causing a force on other charges. We call this effect a **magnetic field.**

Accelerating charges causing a force on other charges. We call this effect **electromagnetic radiation. It is a totally different effect independent of the induction fields. This is the only effect that works at a distance.**
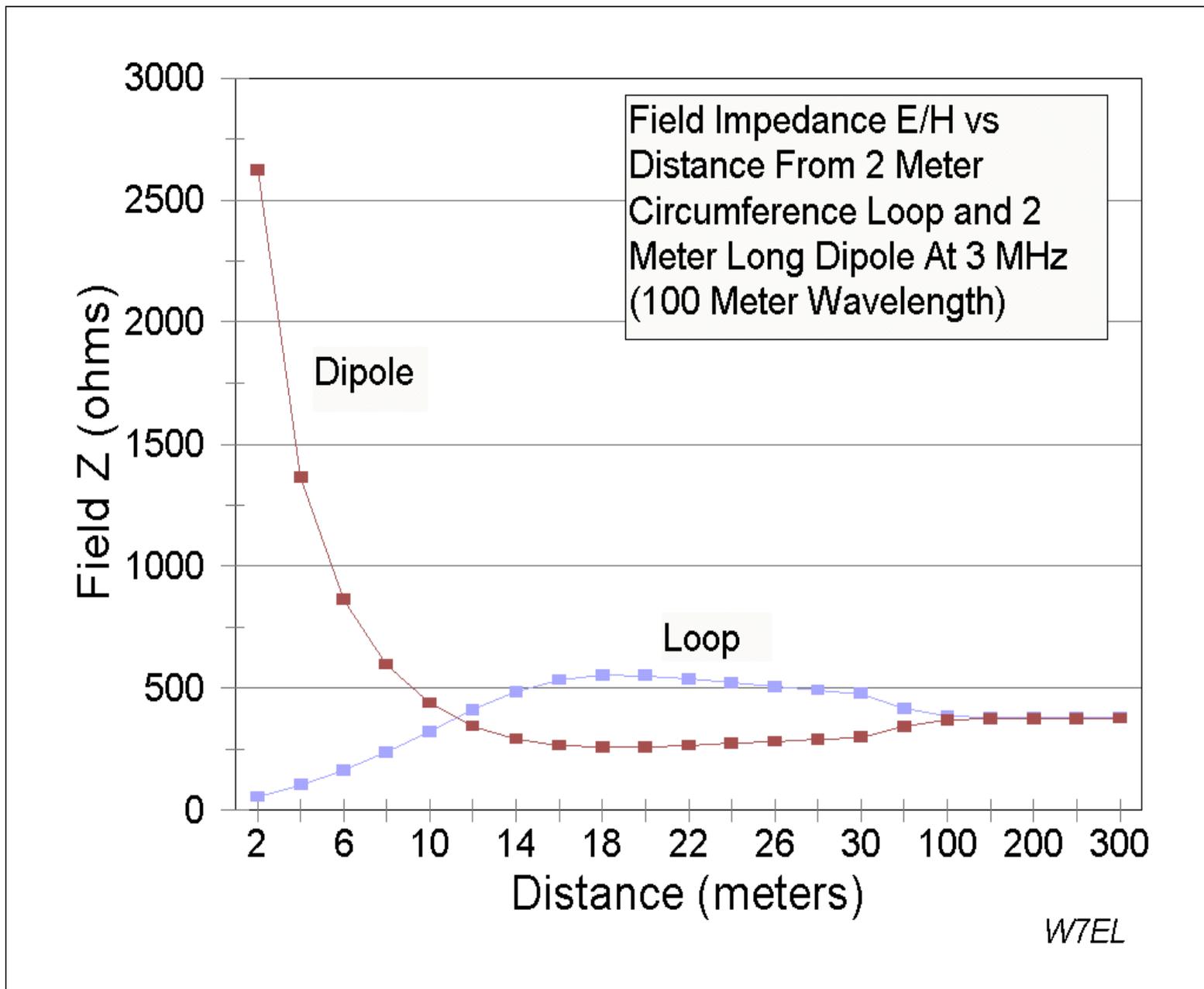
Since the distance of a wavelength in the above graph is 100 meters, we can considered the bottom scale as the percentage of a wavelength. We can see at about 11 percent of a wavelength (which would be about 50 feet on 160 meters), there is no difference in field impedance between a small loop and a small dipole. At distances beyond 50 meters on 160 meters, the magnetic loop actually has a *higher* field impedance than a dipole.

So much for the myth that a receiving antenna can sort good signals from bad signals (noise) by virtue of being "magnetic"! We not only don't have the response we might have imagined, we also almost certainly have no idea if close-by unwanted noise or signal sources are radiated from electric or magnetic field dominant sources. Successful noise reduction by virtue of by antenna "style" would mostly be a matter of hitting a lucky combination through careful experimentation.

Loop Shielding and Balance

Loop shields do not sort noise out, nor do they prevent electric fields from affecting the antenna. For a description of how shields work, look at the [Concentric and Coaxial Transmission lines](#) page. From that page, you will see the shield actually becomes the antenna in a "shielded" loop.

There are many construction articles about small loop antenna available. I'm not going to add one more to the group, but I can tell you what to watch for. It is _VERY_ important that all conductors exit the loop at the ground point of the shield, and that the loop is grounded exactly at the electrical center of the shield. The loop must also be symmetrical, and you must mount the loop so the feedline and any metallic supports leave the center area of the loop with maximum symmetry. If you DON'T do this, the loop can actually use the feedline as the antenna. This can greatly increase sensitivity of the loop to conducted noises!

Remember the following guidelines:

- The shield is the actual antenna
- The shield must be perfectly symmetrical away from the inner wire exit point
- The gap in the shield must be exactly opposite the grounded point
- The ground must be at the inner wire exit point
- The shield will not make an unshielded loop that is properly balanced any quieter
- ***The shield only is a tool to help you balance the system IF the shield is properly implemented***

I hope this article is useful in helping you select a GOOD design!

E-H Antenna

[[Home](#)] [[Up](#)]

Background

Many antennas function because of common mode currents, rather than working in spite of them. Two popular examples are CFA and EH antennas. Another recent example, appearing in Antennex's compact antenna articles, is a thick stub "vertical" with no counterpoise.

All of these antennas become significantly poorer radiators if [common-mode currents](#) on feedlines are eliminated. Why? Because the feedline is the actual radiator, ***NOT*** *the tiny thing they call the antenna.*

Misunderstanding or misapplying Maxwell's equations and the principles behind radiation, in combination with missing some very key points of conventional circuit theory, causes problems. Some of us have unwittingly attributed increased feedpoint resistance and/or seemingly disproportionate amounts of radiation from very small structures to new methods of radiating EM waves. Reviewing these antennas and the theoretical or technical mistakes surrounding them will help us understand how antennas and transmission lines work. With that knowledge, we can build better antenna systems. The fastest and best way to learn is often to look in detail at mistakes!

What They Claim

Articles and user reports of CFA, EH, and thick stub verticals (without groundplanes) appearing in Antennex and other internet publications have one common thread, the operational descriptions almost always include strong indicators of problems with feedline common-mode current.

Authors commonly warn users to NEVER choke feedlines with baluns and to "be sure the feedline is straight and in the clear"! Authors lay blame for RF burns from feedlines or shack equipment on the antenna's "high radiation efficiency", claiming these small magical antennas radiate so efficiently they naturally excite the feedline and equipment more than full-size antennas.

For example, Ted Hart (of EH Antenna fame) claims the following at:

  http://eh-antenna.com/documents/RF_on_the_Coax.pdf

"RF on the Coax

Due to the large radiation at the EH Antenna, there will be some RF coupling to the coax. Whether this is a problem is dependent on the radio you use. Some are subject to RF coupling into the audio system, which causes severe distortion while transmitting. On some field day setups with 100 watt transmitters we have had so much RF on the radio you can get an RF burn. Below we have suggested ways to eliminate the RF coupling problem."

In the above statement, Ted actually acknowledges current on the feedline shield and RF-voltage-on-chassis problems. The problem must be severe when a low-power 100-watt radio causes a burn. Like any good salesman, he turns a design shortfall into a feature! According to the author, unwanted RF on the feedline doesn't come from a feedpoint or antenna design problem like it does on other antennas, in this case it appears because the antenna works so well!

Here's what actually causes RF to appear on a coax shield and radio chassis. RF can only appear on the radio chassis through two methods:

1.) The antenna, from poor feedline or feedpoint design, can couple to the radio chassis through external wiring or cables attached to the radio.

2.) The radio chassis itself, being large in terms of the wavelength, can actually become an antenna and receive energy from actual desired "over the air" signals.

(Many of us have these problems. Click on this link to see one reason why.)

In this case, we can probably rule out reason two above. It is unlikely the chassis is a large portion of a wavelength long on 20-meters and that the antenna field is suddenly so strong it is "lighting up the house" with RF. After all, if power is radiating effectively it will all be going out to distant stations and NOT cooking you or the radio gear in your house! That only leaves reason one, poor feedpoint and antenna design, as the cause of common-mode feedline or wiring currents that excite the coax shield and eventually the radio chassis, as described in method one.

Actually someone has measured this, and posted it to a web page!

http://www.home.earthlink.net/~calvinf15/_technical/

When the time-varying current from the transmitter flows in any conductor, we will have charges accelerating in the cable. The outside of a shield is no exception. A feedline's shield will radiate proportionally by the ampere-feet of the cable. just as any other conductor will. Of course the antenna element will radiate also, but there is something else to consider.

A very small current flowing unopposed over a large linear distance will radiate quite a strong signal, because radiation resistance of a long antenna is generally very high compared to very short antennas. You can find this explained in the [Radiation&Fields](#) and [Radiation Resistance](#) articles on this site, and in engineering textbooks such as those written by Jasik, Kraus, and Jordan-Balmain. From all of this, we know the shield radiates.

The inventor of the E-H antenna goes on to say:

"If you use RF beads, since the coax shield is not a magnetic shield, the beads affect both the inner and outer conductors. Therefore, most of the transmitter power will be converted to heat. Not good."

Not a very knowledgeable statement at all, at least from the standpoint of how shields work!

Time-varying fields can not pass through a shield that is more than several skin depths thick. The inner part of the shield and outer part are isolated by the skin effect, and nothing passes through. The ARRL Handbook, Maxwell's book "Reflections", Reference Data for Radio Engineers, and dozens of other amateur radio and engineering texts describe this effect correctly. If we bring a time-varying electric field to zero in a system, the time-varying magnetic field is also by definition zero. The shield DOES isolate the center of the cable from time-varying magnetic fields on the outside of the cable!

If that is true, why then does a shielded cable passed through a current transformer used in a directional coupler appear to pass RF magnetic fields? Why does the RF magnetic field seem to "pass through" the shield of a shielded receiving loop antenna? The answer is quite simple. There is a gap or intentional break in the shield.

Current on the inside of the shield "spills over" the edge of the shield where the shield is broken, and causes a current on the outside of the shield. There is also a voltage across the gap at each end of the shield. We have both our time-varying voltage and current, via a circuitous path to the shield ends! Our Amateur texts explain that effect, as do all of the engineering texts dealing with shields on transmission lines. If the gap is closed and the shield's ends are shorted together, making voltage across the shield gap zero, the magnetic field no longer "seems to" penetrate the shield!

If a shield did not behave this way, we'd be in serious trouble in the radio world. Without the shield stopping both magnetic and electric time-varying fields, we could not shield our radios. We also could not shield our microwave ovens, with non-magnetic materials!

If anyone thinks a ferrite core affects the impedance inside a coaxial cable that does NOT have significant common mode current, they only need to slip some beads over a cable on a working normal "cold-for-RF" feedline. You will find absolutely no difference in system performance when

beads are added, proving Ted does not "have it right" in the text above that appeared on the E-H antenna web site.

The author and inventor of the E-H goes on to say:

"Use of a small choke made of several turns of the coax is good. We find that a wire connected to the ground side of the coax at the antenna and connected to either a ground rod or a wire laying on the ground will eliminate RF problems - in most cases. For some radios we also need to add a ground wire to the radio."

Of course adding a ground wire might help! The ground wire becomes the path for common-mode currents, or at least a portion of them. The additional wire to ground becomes part of a long-wire antenna (made by the cable shield) that actually does most of the radiating!

"A preferred method is to run the coax to ground then back to the radio. Near ground, connect the shield of the coax to a ground rod or radial. Another method is to connect a wire from the radio to ground. If the radio is very far from ground you will need to add a series resonant circuit in the ground wire to effectively cancel the inductive reactance."

It is understandable why this is a preferred method! The outside of the coax shield can remain the primary antenna, saving us the bother of installing the additional ground wire that becomes the antenna in the previous suggestion!

"It may take one or more of the above to solve your problem. Remember that if you have a good ground on the antenna, you have also minimized problems with lightning."

In the above text, we can see every solution carefully avoids installing an effective choke balun on the feedline. A properly designed and effective choke balun has no effect on a coaxial feeder or system SWR, unless the feedline is radiating!

Let's look at another E-H antenna experimenter and former proponent of the E-H antennas test of a 160-meter E-H antenna. You can read the text directly at:

http://www.qsl.net/iz7ath/web/02_brew/18_eh/english/pag11_eng.htm

Steve writes:

"With my short (and easy) tests I deduce the EH lose about 10 dB versus my short 10 mt vertical with capacitive hat in the top;
10 db is a big amount of power, but remember we are using 2 small cylinder of 37,5 cm on 160 mt band;
my vertical performed better than a 40 mt long sloper, so, may be, loss versus full size dipole is not too much."

Steve claims the E-H antenna looses about 10dB compared to a 1/4wl sloper, which has an unpredictable efficiency. A typical sloper 1/4wl sloper is likely only around 50% efficient in the best situations, and more likely much worse! The actual range can be from 10 to nearly 100% efficiency. Steve's data repudiates Ted's claim and the CFA inventors claim that "crossed-field" antennas provide high efficiency.

"There seems to be something good;
May be optimizing something we can have more gain (or, better, we can lose less dB);
This antenna can be useful on the low band, expecially on 137 KHz (the EH cylinder should be 6 mt high with a 2mt diameter).
On the web I've seen an other similar antenna (ISOTRON), wich seems caming (as EH Antenna) from CFA antennas;
On the italian Radio Rivista (1995-1996) there were same informations about an "Antenna Toroidale" wich in same way remember the EH: will be interesting build and test one."

Steve seems to be saying something we all agree with. A short antenna will radiate, but not nearly as well as a full-size antenna.

"By the way, I think that my EH is not for Dxing; it will be useful for that radioamateur who have no possibility to install "long" antennas for low band, but want just to have local QSO or few contest-contact."

I agree! 10-20dB loss from a full size antenna would not make a good DX antenna! Now here come the current-on-the-feedline problems:

"Other point: tuning;
I think it's not as simple as others say; I've used an MFJ259B which tells me all about the antenna, but the tuning was critical. On the other side construction is easy;
I want suggest to you , when tuning, to connect the MFJ259B to the shack ground (if there's one). If not, EH will probably resonate higher (20-30 KHz)."

If grounding and ungrounding an antenna analyzer or any other piece of equipment connected to a coaxial cable causes resonant frequency or SWR of an antenna to change, the system has severe feedline radiation problems. See my article on testing baluns.

"Coaxial cable influence: inserting more coax cable, the resonance seems to vary a little; I suppose that's normal;

I don't know for certain if the line radiates or not; I've added two iron-powder coil forms back my TX and EH performance was unchanged. Then later I've added 2 big ferrite beads with a lot of turns on the roof (at the feed point of the EH) and all has changed. My signal was 3 S point weaker than before and S.W.R was very high.
I don't know if that loss is due to the changed resonance in EH (as I said before, outside the bandwidth performance goes down quickly)."

We see again that any attempt at reducing feedline radiation results in an antenna that does not radiate very well. With the feedline choked, Steve's EH antenna dropped 3 S-units in addition to the original 10dB from his sloper, or maybe 25dB or more!

There surely is a hidden message in all of the above contradictions!

How the E-H Antenna *Really* Works

There are many examples where designers intentionally use common-mode currents. Examples are found in textbooks, such as the "Antenna Engineering Handbook" by Jasik on and around page 22-6.

The antennas at the right, copied from Jasik's textbook, outline the derivation of a skirt collinear antenna from a simple feedline with the open end terminated into a conductor. (It could be a ground

Two or even three skirts may be added to the mast as shown in Fig. 22-5c, but
reduced excitation of the lower sections diminishes the effectiveness of each additic
skirt.[6] Thus, while the multiple-skirt coaxial antenna resembles a collinear arra
in-phase half-wave elements, its gain is not as great.   Typical gain values are ta
lated in Table 22-2, but since no test data are available for this type of antenna,
values given are merely engineering estimates.

FIG. 22-5. Evolution of multiple-skirt coaxial antenna.

The relatively large diameter-to-length ratio of a practical skirt produces an
effect which requires that the exterior length of the skirt be reduced by a factor of
0.8 to 0.9 and the interior length by from 0.95 to 0.98.[6] This is accomplishe
cutting the skirt to the proper exterior length and inserting a dielectric slug ir
skirt to increase its electrical length.

rod or an antenna, like a Beverage or large loop, the antenna does not have to be an "open circuit".

Looking at (a), we find by hanging any conductor from the end of a coaxial cable the shield is excited (on the outside) with common-mode current. The electrical equivalent of the OUTSIDE of the shield is just as if a generator located at the end of the shield was driving the outside of the shield as longwire antenna. This goes along with Kirchoff's Laws, that tell us the sum of currents entering a point must equal the sum of currents leaving that point. For any current to flow up into the antenna, and equal current must flow back down over the outside of the shield.

With one ampere flowing up the center conductor into the "stinger" at the coax's end-point, the same level of current flows back over the outside of the shield. (The shield's inside and outside are isolated by the skin-depth of the current at the operating frequency, an can be treated as two independent conductors that are connected over the open edge of the shield.) We *MUST* have this current simply because this is how coaxial cables work, the current on the inside of the shield is *ALWAYS* equal and opposite to current in the center conductor. There has to be some place for that shield current to flow, so it makes the bend over the end of the cable and flows back down the outside.

This is also why, when we use a cable's shield as a ground lead the center conductor and inside of the shield do nothing to reduce resistance. Any current that flows down the center conductor is cancelled by current flowing on the inner wall of the shield, the result is no current at all flows down the center conductor as long as the shield is several skin depths thick.

Many antennas intentionally and unintentionally use this principle, two examples are shown in (b) and (c).

A recent Antennex Article on a "magical" ultra-compact antennas claims an identical system, using a loaded fat cylinder, has an extremely high radiation resistance and excellent performance because some magical field-trickery increases the radiation resistance of a thick cylinder at the antenna end. Certainly radiation resistance is somewhat high...but not for the reason something magical or special is happening!

The small coil-loaded cylinder is actually only a fraction of the antenna length, and being so short has a very small radiation resistance. The point missed is the shield of the cable is very long, and is in series with that short section. Since the shield is long, it also has a reasonably high impedance both from radiation resistance and loss resistances. Shortening the length of the end-stub results in an insignificant reduction in radiation resistance, because the overall length of the radiating system is very long! We have a simple off-center fed dipole, with one very long leg and one very short leg!

The main radiator is the outside of the feedline shield, not the tiny thing being called an antenna!

Unless we make the coaxial shield an infinite length or pass it through what amounts to an infinite groundplane with zero resistance, current continues on down the cable shield. Looking at (c), we find even multiple sleeves appearing as parallel tuned high-impedance circuits do not fully decouple the

shield. Many collinear antennas work on this principle, yet E-H antennas and others attribute it to some form of electro-voodoo.

Baluns and verticals

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Ground resistance measurements

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160-meter Ground Connection Resistance at W8JI

Clay soil, dry conditions, 20-foot tall hat element resonant on 160 meters. All radials 50 ft long except for test radial, which is 1/4 wl electrical insulated counterpoise.

Ground	Ground Resistance Ω	Antenna Resonance MHz	System change with resonant test radial	System change adding one additional rod
Two 5' rods	85	1.944	Large over 50%	Rod has large effect
One radial NE	80	1.905	Large	Rod has large effect
One radial SE	80	1.905	Large	Rod has large effect
One radial NW	77	1.931	Large	Rod has large effect
Four radials	42	1.819	Moderate 50%	Rod has large effect
Three radials and two rods	50	1.819	Moderate 50%	Rod has moderate effect
Four radials three rods	54	1.820	Low 25%	Rod has noticeable but small effect

Seven radials three rods	29	1.875	Very low <10%	Rod makes very little difference
Eight radials three rods	27	1.880	Very low	Rod makes very little difference

Performance Indications of Test

Ground systems are reactive. Poor grounds can add reactance. This is shown by the resonant frequency shift as the ground system is altered.

In slightly moist clay pastureland soil it takes about 8 short (1/8th wl) radials to stabilize the ground impedance.

In slightly moist clay pastureland soil, two or three rods make a poor RF ground for low impedance antennas.

If you use four short radials on a vertical, it is almost like no ground at all.

With small radial systems, ground rods can actually increase feed resistance. Use this particular phenomena with some caution, since I did not sweep frequency!

Echelon-Log Beverages

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Enhanced F/B Beverages

Two Beverages are paralleled a short distance apart, and stepped one in front of the other, can be phased to remove unwanted rearward response. This really just amounts to end-fire phasing two antennas. It is effective with long antennas, and also increases F/B ratio (and directivity) with very short beverages (even those approaching 1/2 wl).

Many or most previously published systems use frequency-selective phasing systems. With such systems, operation on more than one band requires switching systems and many components. "Conventional thinking" will make the system more complex than needed, and actually provide less performance! This is unfortunate, because it is possible to design a simple phasing system with almost no components that will cover octaves of bandwidth...no switching required!!

The two basic schemes I've used are very simple. One system is totally independent of grounds at either antenna end, the other system requires grounds but is more forgiving of dimensional errors in construction. I've described a model of a loaded Beverage, for those who would like to experiment with a small-space Beverage at [Slinky and Loaded Beverages.](#)

Bandwidth of Directivity

In high-efficiency unidirectional ***transmitting*** phased arrays, element impedances change significantly with frequency and phase angle changes. In unidirectional high-efficiency transmitting arrays, mutual coupling causes each element in every end-fire element group to have greatly different feedpoint impedances. This seriously complicates the design of ***proper phasing and current distribution*** as frequency is varied. It is very difficult to cover a few percent of the operating frequency, let alone multiple bands, with a high efficiency array.

Most amateur receiving antennas copy what we do with transmitting, and many of our transmitting systems copy narrow-band single-frequency systems. This is very unfortunate!

The key to broadband performance is having frequency-independent stable element impedances, and similar element impedances, throughout the system. This requires thinking "outside the box", not copying of transmitting phasing systems (often TX systems that are not that well-thought-out or planned).

Requirements for Broadband Directivity

There are three ways to obtain stable predictable impedances or broadband directive performance:

- Multiple elements can be used, such as those in log periodic or Fishbone arrays
- Active elements that are far from resonance can be used (short ***non-***resonant verticals or small ***non-***resonant loops with amplifiers)
- Very lossy elements (even large elements) can be used, with losses swamping out mutual coupling effects

While all three methods work, the simplest systems generally use lossy elements that are large enough to provide sufficient external signal levels to overcome receiver system hardware noise.

Broadband phasing systems are easily implemented in systems where feedpoint impedance is stabilized through intentional or "natural" loss mechanisms. Losses are made large enough to "swamp-out" or dilute mutual coupling and resonance effects, antenna feedpoint impedance remains stable and predictable even when elements are end-fire phased with a unidirectional pattern and close spacing!

Beverages As Elements

In the case of Beverages, *radiation resistance* is very low (in the order of an ohm or two). Most of the resistance we see at the feedpoint of a Beverage is from ***dissipative losses*** in soil below the antenna, not from losses associated with radiation! In addition, termination resistance adds another source of loss (perhaps 30% of overall system loss). Overall, loss resistance in a Beverage is very high (several hundred ohms). This means the Beverage's feedpoint impedance is stable, even if mutual coupling radically changes radiation related impedances.

Note: Mutual coupling still remains, since mutual coupling is a function of element spacing and position. Even though mutual coupling still exists, it only affects the ***radiation resistance*** portion of element impedance. Since the large loss resistance (mostly from ground losses below the antenna) dilutes or swamps out mutual coupling effects, mutual impedances can be ignored.

Not only is a Beverage antenna's feedpoint impedance "immune" to mutual coupling effects when phased or placed near another Beverage, terminated Beverages offer a relatively constant feedpoint impedance over very wide frequency excursions. This makes arrays of Beverages ideal candidates for wide-bandwidth phasing systems, eliminating complex phasing and/or switching systems.

Types Of Phasing Systems

In order to better understand my antenna phasing systems, it is necessary to understand antenna phasing systems.

W8JI Parallelogram Array

Although it looks like a large horizontal loop, this antenna actually is two ground-connection-independent end-fire staggered (or stepped) Beverages. The short end wires form two single-wire feedlines at each end. One end-wire is series terminated with exactly twice the resistance of a normal Beverage, while the short wire at the opposite end is the feedpoint.

Stagger and spacing determines feed and termination location, the offset at the two ends being mirror images. A top view, looking down from straight above the antenna, looks like this:
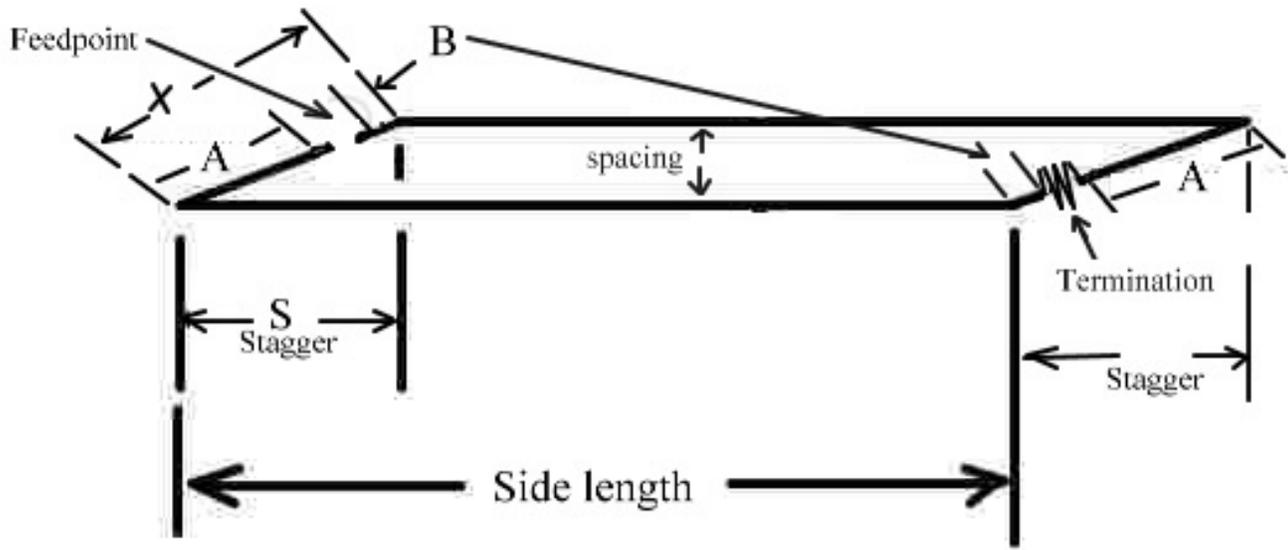
This is an exactly "top down" view, looking down at the antenna from right above the antenna!

Click here for an Eznec file.

Notice the feedpoint is offset towards the termination end (front) of the antenna, and away from the null direction. This is typical for crossfire phased arrays. Crossfire arrays respond *away* from the delay line direction, exactly opposite conventional arrays. The termination is offset the same amount, but moves *towards the feedpoint end* of the antenna.

The feedpoint terminals, being floated (push-pull), provide 180-degree phasing between the two elements. The extra line length to the forward (left and front) element provides the "Stagger" delay.

Consider the actual wire length of a "short side" called "X" (which is the same as A+B). This length is the same on both ends of the antenna. The difference between A and B must equal or be slightly less than stagger (S).

To determine the offset of feedpoint and load:

1. Measure length of the end-wire, length "X".
2. Measure stagger in the end-fire direction, "S"
3. (X-S)/2=B
4. A=X-B

You may want to slightly offset the feed by making B longer and A shorter. This will move the null upwards, forming a cone. It is best to model the results.

Let's review a system, assuming S is 60ft and X is 100ft. We have X-S=40 divided by 2 for 20.

B is 20, A is then 80. The difference is 60, and that is the phase delay. Assuming 1.5 ft per degree that is 60/1.5 = 40 degrees delay. S is 60/1.5, or 40 degrees also.

We have a 180 shift at the push-pull feedpoint, so -40 rotates to +140. We have a 140 lead in the forward element, with a 40 degree spatial array delay. In the forward direction (towards termination) phase is (-40) + 140= 100 degrees out-of-phase. This results in nearly the voltage of one element alone, when the two element outputs of the long sides are summed. Towards the null (feedpoint) phase is +40 + 140 = 180 for a sin180=0 or zero voltage, a perfect 180 null.

On the second harmonic forward array feed system phase is -80 rotated to +100. Spatial array phase is now 80 degrees, or -80 towards termination. The result is (-80)+ 100= +20 in the forward direction. The result is nearly twice the voltage of one element in the forward direction. In the reverse direction, array phase is +80 + 100 or 180 degrees out-of-phase. We once again have zero back-fire response!

The general pattern holds true for any length of S less than 180 degrees, although grating lobes would make the pattern useless. The array, with 60ft stagger S, is usable from about 5 MHz down to VLF.

Summary, W8JI Parallelogram Array

Placing the feedpoint and termination centered on the end-wires is like using 180-degree phasing.

Placing the feedpoint and termination at the stagger minus side length distance is like using S-180 degrees phasing on any frequency with one exception, the antenna fires towards the feedpoint offset. This always results in a perfect backfire null, regardless of frequency.

Here is a wire table for a short 160-meter W8JI Parallelogram Array:

		Wires							
No.	End 1				End 2				
	X (ft)	Y (ft)	Z (ft)	Conn	X (ft)	Y (ft)	Z (ft)	Conn	
1	0	0	6	W2E1	0	300	6	W4E1	
2	0	0	6	W1E1	17	70	6	W3E1	
3	17	70	6	W2E2	17	370	6	W4E2	
4	0	300	6	W1E2	17	370	6	W3E2	

In a 370-foot length, this antenna has the following 20-degree wave angle pattern:
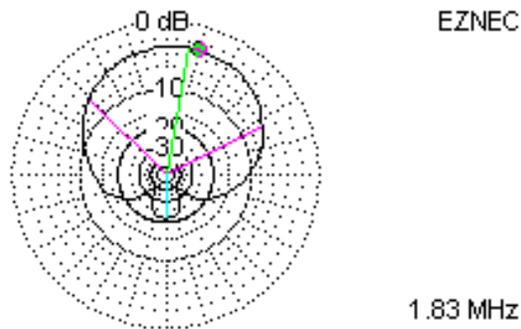
```
                    0 dB                    EZNEC
                      10
                    30


                                           1.83 MHz

Azimuth Plot                Cursor Az    75.0 deg.
Elevation Angle  20.0 deg.  Gain        -18.92 dBi
Outer Ring       -15.99dBi              -2.93 dBmax

3D Max Gain      -15.99 dBi
Slice Max Gain   -18.91 dBi @ Az Angle = 80.0 deg.
Front/Back       19.2
Beamwidth        108.2 deg.; -3dB @ 26.9, 135.1 deg.
Sidelobe Gain    -37.54 dBi @ Az Angle = 270.0 deg.
Front/Sidelobe   18.63 dB
```

Average gain is -25dB, making the RDF about 6dB. In comparison a typical Beverage 370 feet long has this pattern:
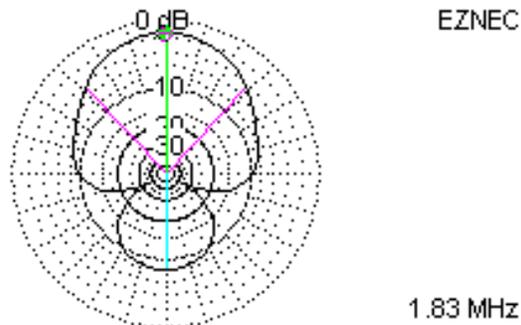

```
                    0 dB                    EZNEC
                      10
                    30


                                           1.83 MHz

Azimuth Plot                Cursor Az    90.0 deg.
Elevation Angle  20.0 deg.  Gain        -16.7 dBi
Outer Ring       -15.25dBi              -1.45 dBmax

3D Max Gain      -15.25 dBi
Slice Max Gain   -16.7 dBi @ Az Angle = 90.0 deg.
Front/Back       6.49
Beamwidth        84.8 deg.; -3dB @ 47.6, 132.4 deg.
Sidelobe Gain    -23.19 dBi @ Az Angle = 270.0 deg.
Front/Sidelobe   6.49 dB
```

If you wish to remove signals from the rear, the W8JI Parallelogram Array is a good broadband solution. It not only requires one transformer, one termination, and no ground systems at either end....it also is useful on several bands without switching anything!

Remember the above example is a reasonably short antenna, performance improves as side-length increases. There is no reason why this antenna can't be used from a band where it is only 1/2-wl long to bands where it is

several wavelengths long, as long as you properly choose stagger and width.

Crossfire-Echelon Beverage

This antenna is another frequency-independent antenna. It requires no band switching to work multiple bands. Because it uses cross fire phasing, rather than conventional phasing or hybrids, one phasing and delay line system covers several octaves of bandwidth.

A second feed method involves installing a pair of typical Beverages using isolation-type matching transformers, and inverting the phase of one antenna. Instead of being fed through two equal lengths of transmission line with a conventional splitter or combiner, the rearward stepped antenna (null direction) has an additional electrical delay equal-to or slightly-less-than the system's stagger distance in degrees. The delay line calculation can be done at any frequency, since stagger and delay change in step with frequency!

Performance is nearly identical to the above array, even very short antenna lengths (down to 1/2 wavelength) provide good F/B ratio. Of course the antenna will work better if length is increased. This arrangement is slightly more forgiving of dimensional errors than other systems, because each antenna is independent.

Remember you need a 180-degree flip at one feedpoint!! When you do this, the electrical delay of the antenna located towards the NULL must be S or slightly LESS than S. Using slightly less than S elevates the null and forms a cone, improving null usefulness and RDF of the array.
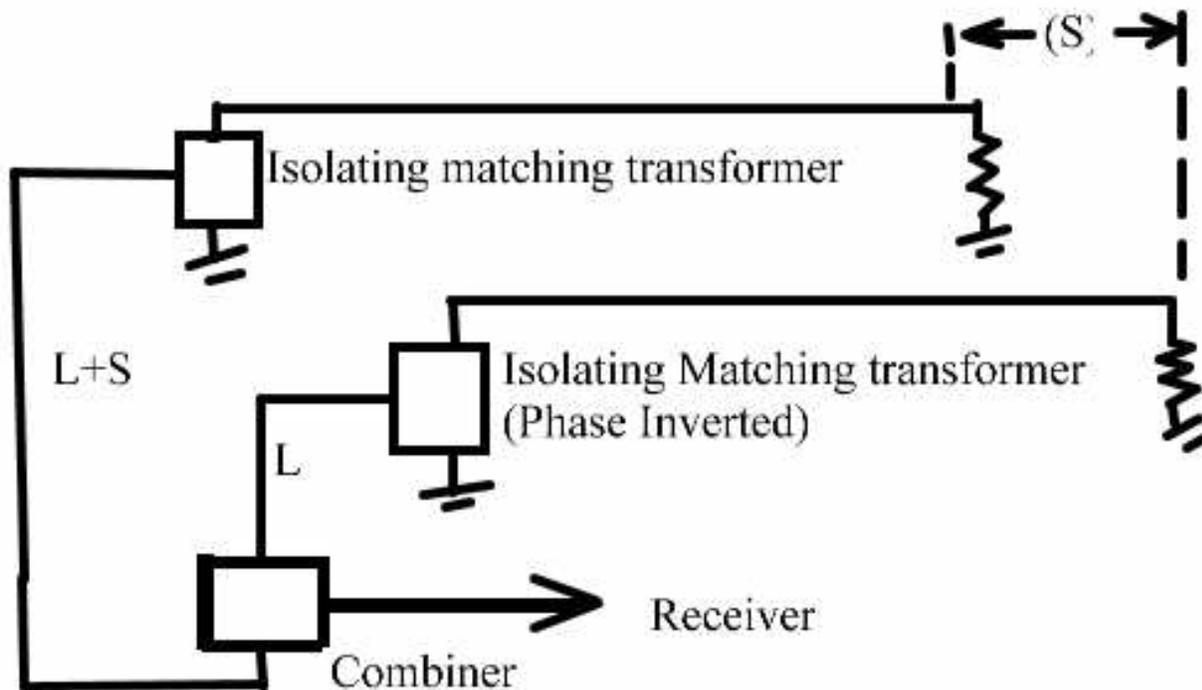
It is important to makes sure antenna feedpoint impedance is correctly matched to the transmission line. Be sure to measure SWR with an antenna analyzer!

You can also use a balanced wire feed, here is an EZnec model of an array using a combination W8JI parallelogram and Echelon-Crossfire feed.

Slinky and Loaded Beverages

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Slinky and Loaded Beverages

When space is limited and a Beverage antenna is too short for normal operation, increasing inductance (or capacitance) along the length of a "Beverage" will increase directivity. Inductance must be adding in series with the antenna, and doing so will increase the antenna's surge impedance.

Capacitance must be added in shunt with the antenna, and the extra capacitance reduces surge impedance. Additional shunt capacitance is very difficult to implement because each capacitor would need a ground, or you would need to surround the antenna with a slow velocity factor *very* thick dielectric. (Most of the air between the antenna and ground would have to be replaced with the dielectric.) There aren't any practical or easily implemented methods of slowing Vp to anything near the amount required through increased capacitance (at least none I could think of), so I will set that method aside.

There are three practical ways to slow velocity of propagation through inductance:

1. Use a spiral or zig-zag element
2. Use lumped inductor loading
3. A string of beads

How They Work

Contrary to what we might intuitively think, loaded antennas *DO NOT* work because we pack 800-ft of wire in a 200-ft area! They exhibit improved directivity because wave velocity is slowed, altering phase shift along the length of the antenna. Slinky (helical) and loaded Beverages show increased directivity because velocity of propagation along the antenna is *decreased*.

As Vp is slowed, the antenna shows increased end-fire response and a gradually narrowing pattern. End-fire arrays fire in the direction of lagging phase, and the slower Vp causes more optimum phasing but only within certain limits. If phase is retarded too much, it actually starts to bring the signal back out-of-phase. With too much delay, the array tries to fire in the reverse direction.

Because of the reverse firing effect, there is a definite limit in phase delay (or Vp slowing) a system will tolerate. If the design goes beyond the optimum value, the antenna pattern tends to reverse direction, causing directivity to decrease. I've found optimum Vp for a 1/2 wl wire is generally around 0.5 times freespace velocity.

Slinky users should be particularly cautious to extend the coils an optimum amount. With 1/2 wl of distance, you would want somewhat less than 1wl of total conductor length. Too many "turns-per-foot" and the slinky, like any loaded Beverage, will try to fire backwards (towards the feedpoint).

Optimum termination impedance will always be somewhat higher than a conventional Beverage antenna. Surge impedance of the antenna is increased by the additional series inductance.

Helices vs. Lumped

For all practical purposes helices and lumped antennas are electrically identical. It doesn't matter at all if the antenna uses a string of soft-iron beads, a series of lumped inductors, zig-zags, stubs, or helices (a Slinky). The only criteria is the spacing between lumped loads must be a small fraction of a wavelength. For all practical purposes, 1/8th wl or less load spacing distance will make a lumped system perform identical to continuously loaded systems (slinkies).

Lumped reactances, whether in the form of a bead, stub, or conventional coil should have at the minimum a somewhat modest Q. This is just another way of saying the inductances should not add excessive series resistance. *Total* reactance of *ALL* loads should be between two and fives times the surge impedance of the antenna. With nine inductors along a 1/2 wl wire, I've found optimum reactance to range between 150 and 250 ohms. Surge impedance was about 700-1100 ohms in systems I tested. A Q of 20 or more would be acceptable in each inductor.

Attached Model

I've added an Eznec file of a loaded Beverage. A few things to note are I used radials to avoid a connection to high-accuracy ground at the termination. The feedpoint isn't critical, since it does not greatly affect directivity in this antenna. Unlike many antenna models, it appears a Mininec ground can be used without harming results.

Using a variety of ground-types I've found RDF remains at about 6dB. This places this antenna slightly below most elongated terminated loops, but well ahead of conventional short beverages.

loaded_bev

EZNEC
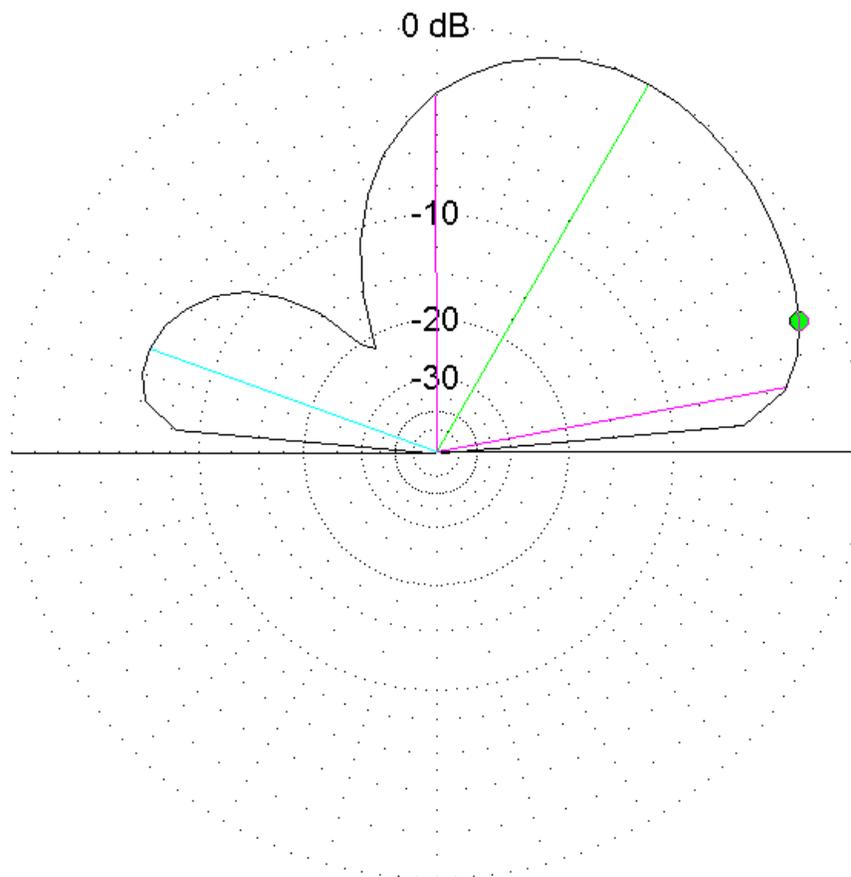
1.8 MHz

Elevation Plot		
Azimuth Angle	90.0 deg.	
Outer Ring	-24.68dBi	
3D Max Gain	-24.68 dBi	
Slice Max Gain	-24.68 dBi @ Elev Angle = 60.0 deg.	
Beamwidth	79.6 deg.; -3dB @ 10.6, 90.2 deg.	
Sidelobe Gain	-30.36 dBi @ Elev Angle = 160.0 deg.	
Front/Sidelobe	5.68 dB	

Cursor Elev 20.0 deg.
Gain -26.33 dBi
-1.65 dBmax

EZNEC

0 dB

-10

-20

-30
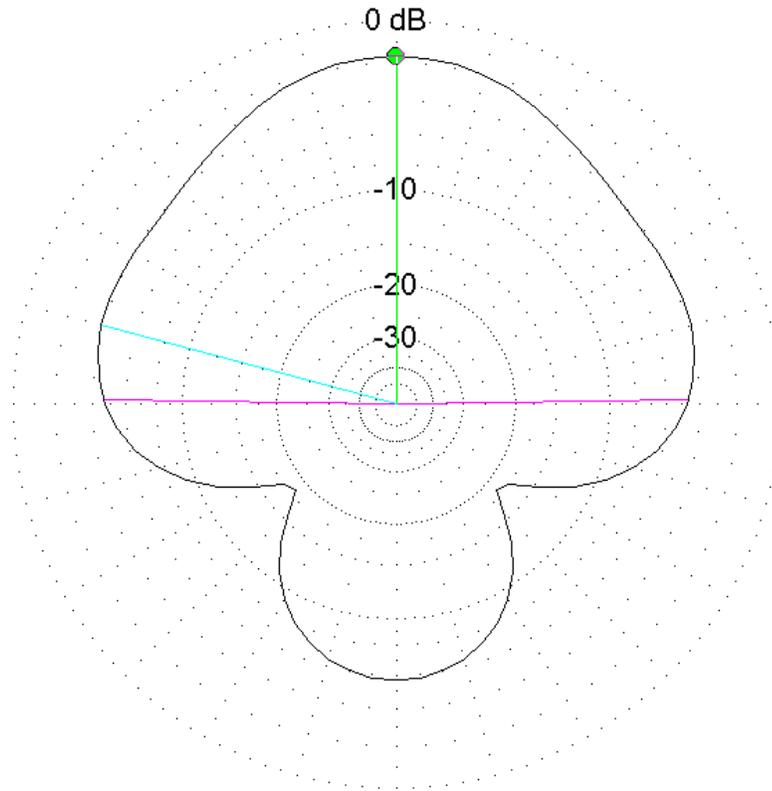
1.8 MHz

Azimuth Plot
Elevation Angle   20.0 deg.
Outer Ring   -24.68dBi

3D Max Gain   -24.68 dBi
Slice Max Gain   -26.33 dBi @ Az Angle = 90.0 deg.
Front/Back   4.03
Beamwidth   178.0 deg.; -3dB @ 1.0, 179.0 deg.
Sidelobe Gain   -28.58 dBi @ Az Angle = 165.0 deg.
Front/Sidelobe   2.25 dB

Cursor Az   90.0 deg.
Gain   -26.33 dBi
             -1.65 dBmax

Phasing Systems

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Standard Delay Line Systems

Many phasing systems use standard ***180-Spacing=Delay line*** phasing. In other words, the end-fire distance in degrees is subtracted from 180-degrees, and the result is the delay line length. This is an acceptable method in ***single frequency or narrow-band*** systems.

Conventional delay line systems have the following problems when used in endfire unidirectional arrays:

- Phase shift almost always generally changes the wrong direction as frequency is varied
- Phase shift only equals line length when the line is perfectly terminated or when the line is lossless and an exact multiple of 1/4wl
- Voltages or currents at both ends of the line are equal only when the line is perfectly terminated or an exact multiple of 1/2 wl and has negligible loss
- Currents at the load end of an odd-quarter wave (or multiple) line are only equal when the lines are fed from the same voltage and have negligible loss

Hybrids

Some systems use ***90-degree Hybrids or L/C phasing systems.*** Hybrids offer ideal distribution of power and provide the expected phase-shift only when perfectly constructed and terminated. While there is some tendency to self-compensate phase, they still suffer bandwidth limitations. Hybrids are very useful in electronic systems, amplifiers are one example. In an amplifier system, we might want a constant 90-degree phase shift despite slight frequency shifts.

Unidirectional antenna systems are never optimized when phase shift is fixed at some arbitrary value that remains constant as frequency is varied. ALL *unidirectional endfire* arrays require phase to track with element spacing change in degrees, as frequency is varied. Lumped component systems and Hybrids might save space, but they do not enhance array performance over the proper choice and design of transmission line phasing systems. This is true in both transmitting and receiving antenna systems!

If hybrids are so poor, why do we see so many of them in antennas? There are a few reasons

authors and manufacturers use hybrids. They often think:

- Hybrids assist in maintaining proper currents in elements
- The dump resistor only absorbs a portion of reflected power that would be wasted anyway
- Power is wasted only when the load SWR is high
- The association with Collins engineering must mean the system works well in any application
- Well-tuned Hi-Q components must work better than transmission lines

In reality, none of the above are true in a broad sense. Hybrids have their place, but it certainly is not in wide-bandwidth phased arrays.

Transmission-line phasing systems are a bit more tolerant than Hybrids. For example, a 90-degree long transmission line has zero degree phase error, even when grossly misterminated. A 90-degree phase delay transmission line has less phase error than a quadrature (90-degree) lumped component Hybrid when mismatched. Phase error peaks in misterminated transmission lines when the line is any odd-multiple of 1/8 wavelength, and is minimum with lines any multiple (even or odd) of 1/4-wl. For example, a 45-degree long transmission line provides 27 degrees phase lag when terminated in 25 ohms...not the 45 degrees people often expect! (Remember this when you see phasing designs that just throw a certain length of cable in series with a mismatched impedance!)

With all systems, amplitude errors are a problem. There isn't any passive system that provides correct phase ***and*** amplitude when load impedance changes, especially over a wide frequency range.

Crossfire Phasing

I prefer ***[cross-fire phasing](#)***, rather than the conventional narrow-band phasing methods discussed above. Cross-fire phasing, when designed properly, ensures phasing is always correct regardless of frequency. When elements (in this article Beverages) offer a near-constant impedance that is almost entirely resistive over a wide frequency range, cross-fire phasing can function perfectly from VLF to LF all in one antenna. Phase and amplitude can be designed for a back-fire null, with the upper limit in frequency set by element spacing and the lower limit set by array sensitivity. It is possible to design cross-fire receiving arrays that maintain the same basic directional response over several octaves of bandwidth.

Crossfire Phasing

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Crossfire Phasing

Most amateur antennas are copies of the ideas used in single frequency systems, like AM BC stations. Most of my work has been in HF systems in and around amateur bands. Generally these systems require wide bandwidth, rather than a certain pattern on one single frequency.

The proper way to extend bandwidth is to use cross-fire phasing, where one element's phase is rotated 180-degrees and the delay is set approximately at the spacing. Here's how it works:

We have two elements with spacing "s" at X and Y. The signals come from direction 1 or 2.

1-----X-----s------Y------2

 Let's assume X and Y are 90-degrees apart on 80 meters.

A traditional array would delay the array feed to X and Y by 180-s = 90 degrees. Assuming X lags Y by 90 degrees, the phase of a signal arriving at X from direction 1 would be 0 degrees for the reference. The signal continues on to arrive at Y 90 degrees later because of the physical distance in space. Since X lags Y in the array phasing, the phase at the common point of the feed system is now X's phase of zero plus -90 delay is -90. The additional signal delay in space to Y is also -90. The signals are in phase, and the result is the in-phase sum of the two signals. Our receiving level has doubled over one element.

For a signal coming from 2, the phase of element Y is zero. Element X receives the signal 90-degrees later for -90 phase. That -90 adds to a -90 in the delay lines, and the result is -180 at the common point. The signals subtract exactly to zero, and there is no response.

Now assume we move the array to 160 meters, where our fixed length delay line and spacing are both 45-degrees long. The in-phase direction of 1 is still the direct sum of the two signals, but the out-of-phase direction becomes (-45) + (-45) degrees or -90 degrees at the common point. The signal vector addition is now unity. We have a very poor null!

Now let's go back to 80 meters and invert one element 180-degrees someplace in the system. In this case, we have:

From 1 we have X=0 degrees plus -90 though air is -90 at Y. Assume Y is inverted (it also works if we invert only X). The phase at Y is -90 but rotated to the opposite side of a "phase circle" by the mirror flip of 180 in element Y's inverted driving system. A -90 with a 180 flip is now +90. This combines with the -90 delay from element X through it's phasing line delay, and the sum of two equal signals of -90 and +90 is zero. -90 is exactly 180 degrees from +90, and the array now has no response towards 1.

When the signal arrives from 2, the phase is now inverted the same way and the result is perfect addition. We have Y =0 plus invert 180 is -180. Space delay is -90 plus -90 in the delay line for a net delay of -180. The result is signals from 2 are now in phase from both elements. The only change by adding a 180-degree element flip is the array inverts direction!

Here is where it gets interesting, and I am amazed so many amateur antenna designers miss this. Y is still inverted 180. When we move to 160 meters, the delay from direction 1 is:

X=0 The delay line adds -45, the same as the distance through space to Y. Since the delay line and space are the same, the delays effectively cancel. Since element Y is inverted, the combined phase at the common point is now exactly 180-degrees out of phase. the result is zero response towards 1!!!

More amazing, this is true for any frequency! There is always a perfect null towards 1 as long as the signal levels are even and impedances are matched from the two antennas, and we use a physical delay line rather than lumped components.

From direction 2 we have:

Y=0 plus a 180 flip for -180. At X we have -45 space plus the delay line of -45 for -90. We have -90 difference. The signal is the same as the signal from one element.

The only effect, as frequency is reduced, is sensitivity of the array drops. We have less signal, but we would have that anyway even if we used the narrow band phasing systems commonly used!

Why do amateurs, who almost always move around in frequency, use single-frequency 180-s phasing systems? Probably for the same reason we use 90 and 180-degree shift in a four square. We started out wrong, and just kept doing the same thing. While this won't provide octaves of bandwidth on transmitting, it does reduce phase errors across a single band substantially. It also provides a phasing method that allows us to build receiving arrays covering octaves without any loss of null direction or depth!

Combiner and Splitters

Combiners and Splitters

The magic "T" is a very useful device. It can provide equal voltages, equal current, or equal power to matched or unmatched loads. It is not a magic bullet.

The basic element looks like this:
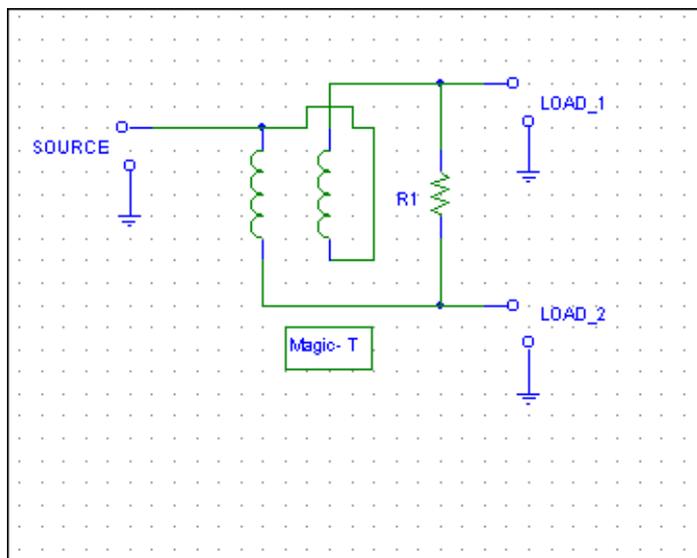
For 0.1 to 30MHz applications, the transformer is a twisted pair of small (number 18 to 26 gauge) enameled wires with five to seven passes through a 73 material one-inch binocular core. The source impedance is parallel combination of the two load impedances, and generally is stepped back up through a 5:7 turn ratio (1:2 impedance ratio) transformer. R1 is twice the load impedance in splitter/combiner applications.

Like all passive splitters and combiners, this device is less than ideal. Isolation is only maximum when at least one port, and often two, are properly terminated.

This system does have an interesting characteristic, it can be used to force equal current into two different load impedances (or equal power by adjusting R1 to some value between zero and infinity). If R1 is open, each load is forced to have equal currents. If R1 is shorted, the loads naturally see equal voltages (T1 can be removed and the entire circuit configured as a T connector).

By forcing equal currents, a 1/2 wl transmission line can be used to feed a directional array instead of the more typical odd-1/4 wl line. With a 2:1 load resistance unbalance, I measured less than 0.1dB current error in such an application.

Receiving Splitters

Virtually all splitters are based on the Magic T with a matching transformer on the input. This splitters are certainly less than ideal, but they are simple and many times (but not always) better than a direct parallel connection.
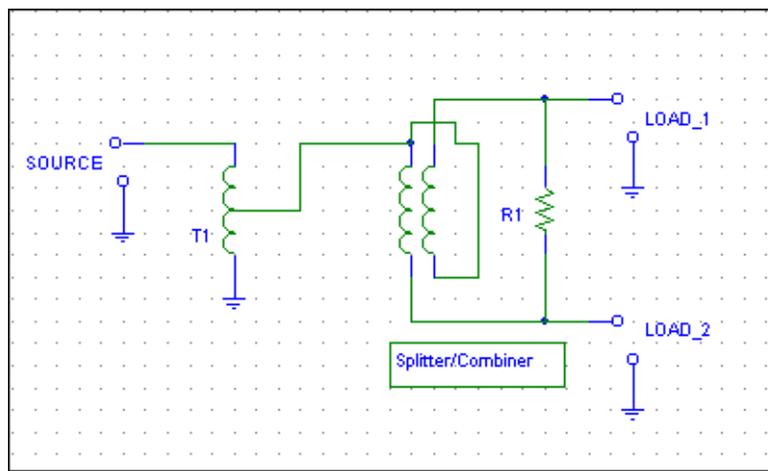
T1 is a 7-turn transformer tapped at 5 turns (1.4:1 turns ratio, 2:1 impedance ratio) step-down transformer. 73 material binocular cores are ideal for 100kHZ to 30MHz.

The magic T transformer is 5 to 10 turns of twisted pair wire through a 73 material binocular core. Configured as a center-tapped winding.

R1 is twice the expected load impedance. For 50- ohm systems use a 100 ohm resistor.

Losses

Typical loss of the Magic T is 3dB per port. The only other significant loss, when properly terminated, is a slight loss in T1. Normal loss of the entire system to either output port is 3 to 3.5dB. When combining in-phase signals, loss is only the loss in T1, there is no "combining loss".

When the Magic T splitter or combiner is open-circuited on one port, the electrical equivalent is:
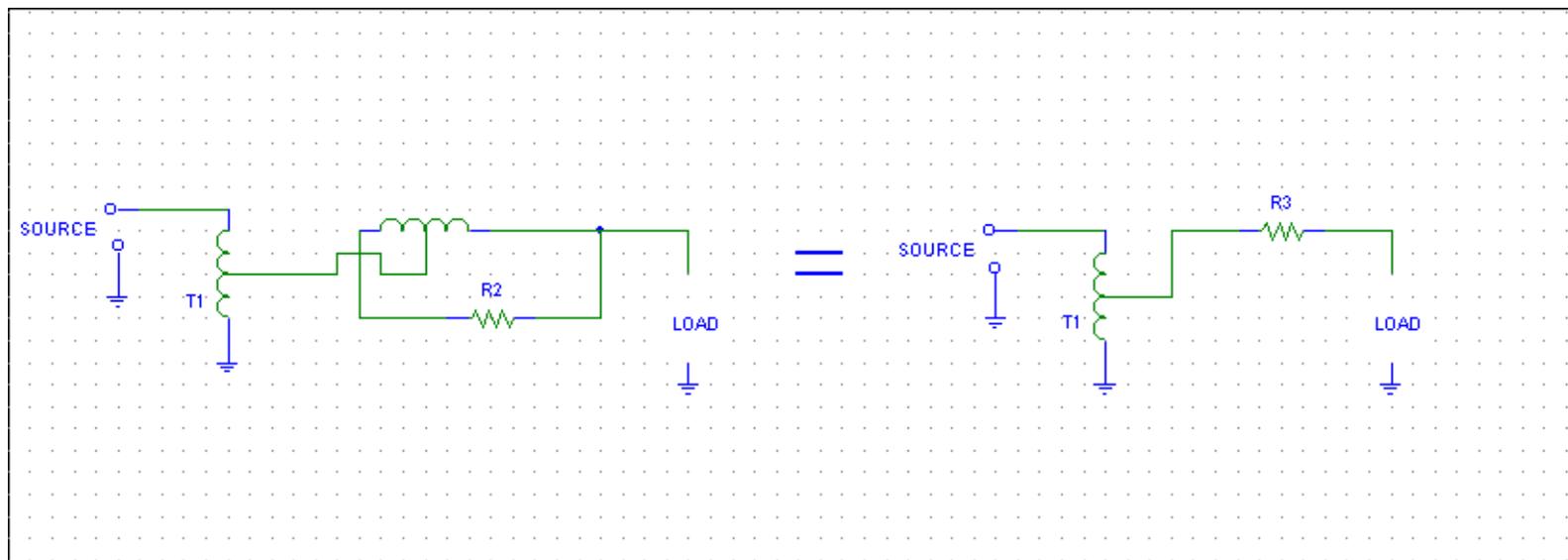
R2, normally twice the value of any port, is stepped down by the impedance ratio of 4:1 in the Magic T transformer when a load port is opened. In a 50-ohm system, this effectively results in 25 ohms (R3) in series with the load impedance of 50 ohms. The resulting impedance of 75 ohms is doubled by T1 to 150 ohms. The equivalent input SWR, when open on one port, will be 3:1.

Loss will be 1.25dB from mistermination of the source by the 3:1 VSWR (assuming it is a 50 ohm source) with an additional loss of 1.7dB as power divides between R3 and the output port. The **theoretical open circuit loss (including transformer losses) is about 3 dB, the same as if a load were connected.**

The Magic-T behaves in a similar manner if one output port is shorted, with the exception R2 appears directly across the load. This is the equivalent of a 100-ohm resistor shunting a 50-ohm load. This impedance (33.34 ohms) is quartered by the center tapped winding of the Magic T

to 8.3 ohms. T1 steps this impedance up to 16.6 ohms. The result is, once more, a 3:1 VSWR.

This means signal level at port 1 theoretically does not change when port 2 is open or shorted, or anything in between.

Like many systems, these losses are based on source and load impedances being ideal. This is almost never the case, and the optimum value of R2 is often significantly different from being twice the load resistance. Because of that, our actual systems may show significantly more level change than the theoretical value (which is no change) when one port is incorrectly terminated.

If the system does not divide signal properly, the systems requires additional gain. This gain comes at the expense of lower IM performance. This is particularly true if the low-headroom amplifiers typical of amateur radio "preamplifiers" are used.

Even with high dynamic push-pull CATV transistor amplifiers, I've often found it necessary to optimize the splitting system. For example, stock my R4C's typically have ~20-ohm input impedances on 160-meters. Using R4C's on one port and higher-impedance receivers (or worse yet an open-circuit) on the other port results in starving the already low-sensitivity R4C receivers for signal (despite the use of a "splitter") if they are used in conjunction with higher sensitivity higher impedance receivers like FT1000's. The solution is to make every load (all the receivers) look like 50-ohms, readjust R1 to a new value, or use emitter-follower "active splitters".

I opted to make all of my receivers "look like" 50 ohms, but your choice may be different.

Amplifiers

Caution must be used in selecting amplifiers. Virtually all commercial amplifiers I have tested have very limited dynamic range. They are fine for amplifying high-loss antennas like Flags and EWE's, but terrible when connected to large Beverage systems. Adding an additional 6dB or more gain to the output of a Beverage just to overcome splitting losses (plus problems associated with uneven power division) almost always pushes these amplifiers "over the top". If you are going to use single-ended amplifiers with small low-power devices, be prepared to use multiple amplifiers (one for each receiver) and place them after the splitter system.

It is far better to buy or build one good amplifier, but it should be a very high dynamic range device!

K9AY_Flag_Pennant_Ewe

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Elongated Terminated Loops

Terminated loops have only recently become popular, even though such systems were available commercially in the 1960's and earlier. Sometime in the early 1970's, I received an advertisement from the late W1BB. That advertising brochure described arrays of receiving loop antennas marketed by a Canadian manufacturer for commercial and military use. The brochure had pictures of diamond-shaped terminated loops, individual small loops, and loop arrays that looked like a number of Flag antennas strung end-to-end where each rectangular "loop cell" fed the next loop. The brochure gave directional patterns of the arrays, and the patterns looked much better than those of other antennas. Unfortunately the price, as I recall, was in the tens of thousand dollars for even the least expensive array! Although detailed technical information was withheld, it wasn't difficult to figure out the electrical design of each array.

That early brochure started me thinking in terms of antennas other than Beverages. I experimented with small loop antennas, and eventually installed an array of eight in-line loop cells. That system allowed me to be the first eastern USA station to work Japan and other Asian countries through interference from multiple multi-KW LORAN transmitters, and as the system was refined I could work JA's on a regular basis long before the broad LORAN pulse transmitters were removed from 1900 kHz. This is a testimony to the clean pattern and wide deep nulls possible using arrays other than traditional Beverage antennas.

Are they quiet because they are loops? Not at all!

One of the most damaging rumors to good antenna science are claims that loop antennas have a mystical property that somehow rejects noise. Nothing is further from the truth, with one exception. Corona.

Loops do not have sharp points sticking out into air at high impedance points, and that can be an advantage in limited cases.

Any sharp point extended from the area of anything, even wooden ship's masts in old sailing vessels, is more subject to corona in inclement weather. Sailors feared St. Elmo's Fire long before anyone knew what electricity was, and amateurs fear it today when trying to receive. We often just don't know what to call it, and can't see it in our brighter skies, so we call it "precipitation static".

The sharp point in a dipole where the wire or tubing ends not only promotes corona by virtue of the

fact it is "out in space by itself", it almost always has the disadvantage of being at a high impedance point in the system. The tiny random charge movements with little current and very high voltage (high impedance source) inherent in conductors around the corona discharge are driving a high impedance part of the system, an ideal situation to maximize transfer of tiny amounts of noise power into the system!

That is almost certainly where the rumor that loops reject noise comes from. We forget it is a function of the sharpness and impedance in the area where the antenna is sharply shaped, and instead fixate on the fact it is "a quiet loop".

Are terminated loops actually loops?

Terminated loops do not behave like small conventional loop antennas. They do not carry uniform in-phase currents throughout the area of the antenna. They don't behave like directional couplers either, they are antennas...not nearfield coupling devices coupled to other conductors (we hope). Terminated loops are really just short verticals, with the phasing inherent in the long horizontal wires. The loop's vertical areas receive the desired signals, while the horizontal conductors merely serve to act as transmission/phasing lines for the vertical (or sloping vertical) ends. A small elongated terminated loop acts like a simple two-element vertical array with integrated phasing lines.

This is true even in the K9AY Loop and Pennant antennas, which have sloped conductors. The sloped conductor behaves as a vertical (think about this when people follow the mistaken advice that sloping the last few feet of a Beverage "stops noise pickup of vertically polarized signals"), ten feet of vertical drop is still ten feet of vertical conductor exposed to vertically polarized signals. The fact the antennas work so well is testimony to how sensitive a sloped wire is to vertical polarization. If sloped wires weren't sensitive to vertical signals, the Pennant and K9AY Loop wouldn't work!

Elongated terminated loops, and arrays of elongated terminated loops, are a special form of short phased verticals where phasing and feed systems are an integral part of the antenna. This integrated system of elements, feed system, and phasing solves construction problems associated with arrays using more recognizable vertical elements. With any system there are tradeoffs, elongated loop antennas only allow very limited control of phase and unwanted high-angle response. We can't obtain optimum patterns (although a properly placed series capacitance will help) because this system only allows limited control of current distribution and phase shift.

These antennas "want to" be verticals because the earth below them and somewhat closer spacing prevents the horizontal components from being efficiently coupled to space while the earth simultaneously enhanced the vertical radiation component of the gradually sloped conductors making up the antenna. The horizontal sections act as a wide-spaced air dielectric transmission line.

These antennas work best when propagation time of signals along the horizontal wires matches propagation time of the wave in space around the antenna, and when the earth or a ground system below the antenna is good. Suppression of high-angle horizontally polarized radiation, and maintaining velocity of propagation near unity, are why EWE antennas (and other forms of this family of antennas) thrive on good ground systems or good earth below the antenna. The desired earth effects are opposite those desired with Beverage antennas, an ideal situation for highly conductive soil!

Why use terminated loops?

It is easy to understand why "loop" systems, even very small loop systems, have become popular. Arrays of terminated loops vertical elements produce effective low-angle receiving performance along with a relatively clean pattern. EWE's, Flags, Pennants, and K9AY loops are effective methods of building two element vertical arrays. They are easy, small, and inexpensive! They are noticeably less directive than two phased verticals would be, because the horizontal components are not totally cancelled by ground effects and the opposing wire, but they certainly are easier to construct than phased verticals (with all their loading, grounds, and coaxial cables). It's all the standard old tradeoff we just can't seem to get away from, we always must balance complexity against performance.

Many people are working with the various arrays of elongated loop antennas, so there are few contributions I can make other than describing how or why they work. I would like to suggest it is possible to extend the arrays end-to-end for some distance without external feed systems, and well-placed reactances can be used to modify patterns. Very little work has been done in that area. I'd suggest experimenting with series capacitors, perhaps placed mid-way in phasing (the horizontal wires) areas to increase velocity of propagation through elongated loop arrays and increase directivity.

Are small loops hyper-sensitive to vertical masts?

There is no compelling evidence that any of these antennas are more sensitive to vertical metal masts than any other antenna would be. As a matter of fact, the only basis for such claims appears to come from models that fail to pass simple recommended tests for model accuracy and stability. If we build a model that is flawed and oversensitive to changes in things like the number of segments used in the model, we can expect it to be sensitive to nearly any change!

Everyone is free to say and do what they like, but other than keeping a short mast a few feet away from vertical wires and NOT connecting that mast directly to the antenna or feedline, I wouldn't hesitate for a second using metal supports. My large arrays of loops in the early 70's used metal masts, my arrays in the 80's did, and as have commercial arrays.

Feedpoint Matching

Great care must be taken in decoupling the feedline from the antenna in the balanced versions of these antennas, although the EWE (being unbalanced) is relatively immune to such problems. Keep in mind the antenna generally looks capacitive as a common-mode structure, so inductive decoupling (i.e. a choke coil of coax) can actually *increase* system problems. The best common-mode isolation system would be an isolated winding transformer designed with minimal capacitance between the antenna winding and the rest of the system. I use a small transformer with stacked 73 material binocular cores in feeding some of my high impedance "log-Beverage" arrays, and similar transformers should work with ~1000 ohm impedance elongated loop antennas.

Because this transformer only has a single turn primary (two turns with the balancing pass), I'm able to reduce stray capacitance to a dozen pF or less. It has excellent balance, low SWR over a wide bandwidth, and very low loss. The reasonably low transformer capacitance, when used in concert with proper feedline grounding and routing, should make the system relatively immune to common mode problems. I'd route the feedline horizontally directly away from the end of the antenna for a few dozen feet (but never a distance approaching 1/10th wavelength or longer), and then drop the feedline to ground earthing the feedline shield at that point. Decoupling beads or sleeves belong on the receiver side of the shield grounding point, not between the ground and the antenna!
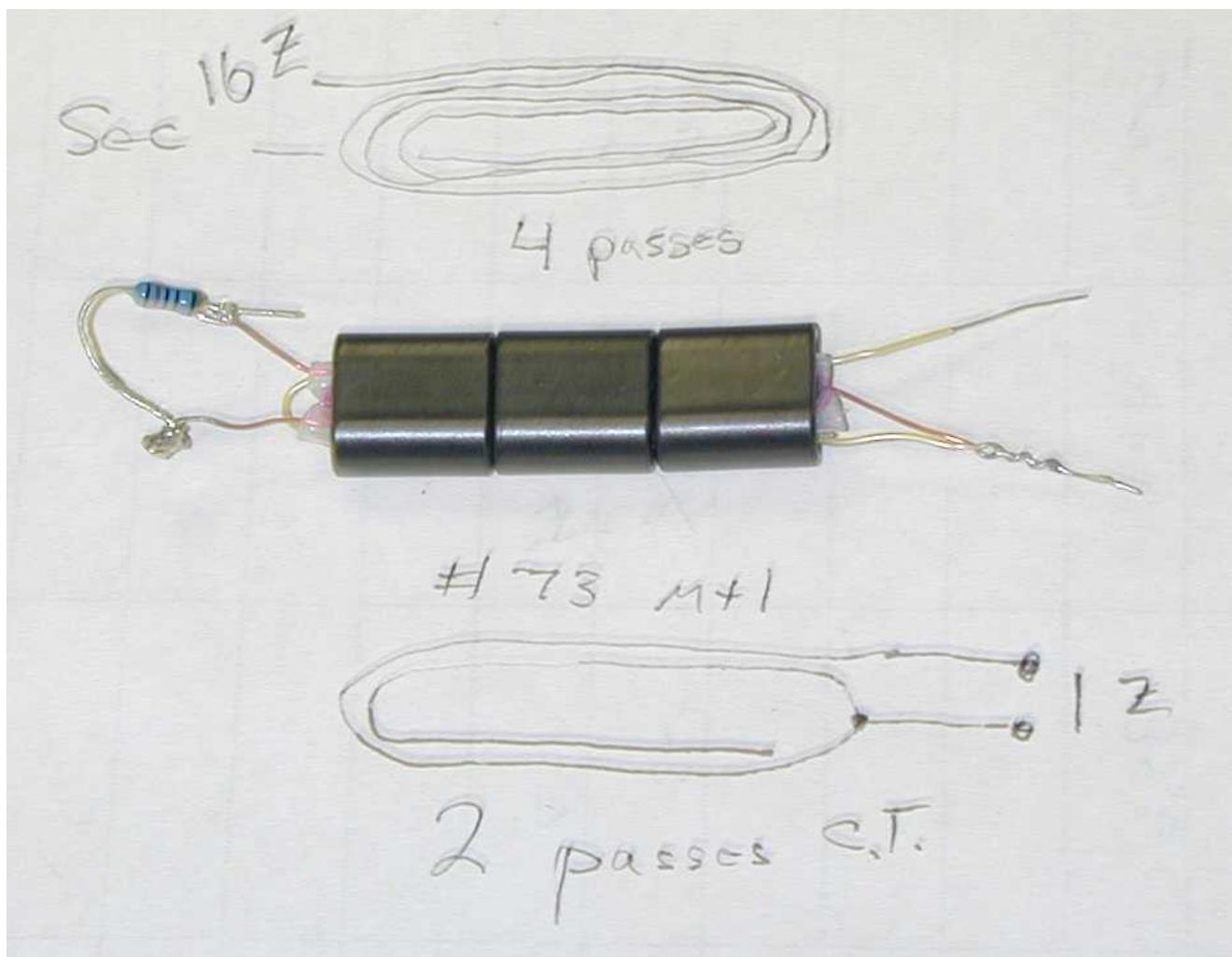
I wind this transformer on three Fair Rite Products 2873000202 cores (about 1/2 inch square and 1/3 inch thick 73 material). The high impedance (secondary) winding is #26 enameled wire through Teflon tubing, while the primary is Teflon coated wire-wrap wire wound outside the tubing. The small extra pass that "dangles" on the low-Z primary winding helps balance the system, even though it adds a few pF of capacitance.

By the way, a Faraday shield will only make things worse. It will increase unwanted stray capacitance and might deteriorate the high impedance winding's balance if it is not properly grounded. The proper grounding point for a Faraday shielded primary is *opposite* the exit point of the primary winding, or on the secondary winding's exit side of the transformer. Most Faraday shields described for Beverage and other transformers are not only useless, they are often incorrectly grounded and actually increase unwanted coupling!

Eznec file courtesy of Roy Lewallen.

[K9AY_Lewallen](#)

Receiving_basics

Receiving Antenna Design

My local wintertime 350Hz BW noise (after amplifier) compared to a sample of signals on one night was:

Noise -127dBm

9H1BM -122dBm

OM0WR -95dBm

DF2PY -88dBm

WA8OLN -78dBm

W3GH -60dBm

W4ZV -32dBm

The dynamic range between noise and W4ZV was 95dB! It also illustrates how important antennas, location, and propagation are rather than power. The dB difference between signals from the same area is profound. Many signals run at or near noise floor. This may not be typical of every night, but it shows how large the signal level variations between weak DX and strong signals are.

Before talking about receiving antennas for lower frequencies, it is important to understand a few basics. We all understand the primary reason we use special receiving antenna systems is to improve signal-to-noise ratio. On the surface this sounds like the same reason we use directional transmitting antennas, but there are some very important differences between transmitting and receiving applications.

One thing we often hear is how a 7dBd gain yagi will improve receiving by a similar amount over a dipole, the logic behind that idea is the gain works both ways. I've read that time and again as justification for using a high gain transmitting antenna. On the surface, it does sound reasonable and correct. Yet it's really not correct in most (if not all) cases! A 7dBd gain antenna with 20 dB nulls can improve receiving an amount from nothing to 30 dB or more, depending on many things we often don't consider.

When transmitting, we can not tolerate much power loss. The most important antenna characteristic, other than a reasonable SWR and having the antenna remain in the air for a while, is having maximum possible field strength in a given direction or directions. In general, the single most important parameter is having maximum gain in all the desired directions and at the most useful wave angles. We know, for a given polarization and wave angle, the highest gain antenna will deliver the strongest signal to our target area. Transmitting antenna choice is reasonably simple, we want properly positioned gain. Take-off angle, side-lobe strength, and F/B ratio take a back seat to maximum gain at the desired wave angle and direction. We can generally model the antenna and confirm the results with a few simple

measurements, and be confident we have made a good choice. By the way, if you hear anyone selecting an antenna based on take-off angle help him out. Take off angle is not important, what we actually need is maximum possible gain at the desired angle and direction. After all, we don't care where the peak is as long as the antenna we pick has more signal (gain) at the desired spot than other antenna choices!

When receiving, the system has a much different design priority. If we consider gain, we can be totally misled. It's easy and somewhat common to make poor choices when picking receiving antennas, mostly because we often don't know what to look for. What we look for can also be different depending on the actual application. A system that works extremely well in some applications can be a real dud in other applications. I'll explain how to make the best choice and what the pitfalls are, but first we have to understand some basics. *If you would like to hear a demonstration of how excellent directivity reduces noise, listen to QRN_JA2ZJW on the [DX Signal Recording](#) page.*

Antennas and Fields

 If we are going to pick the best receiving antenna, it is important we understand how antennas respond to "signals" and "noise". Unfortunately, this is probably the single most misunderstood part of receiving antenna design. This a problem mostly because we are constantly bombarded with folklore giving us the wrong impression of how antennas work.

One example that comes to mind is small loops, often referred to as "magnetic radiators". Folklore often considers the radiation or response of these antennas as somewhat magical, with the antenna behaving like a sieve that sorts "good magnetic signals" from "bad electrical noise". Nothing is further from the truth! At a relatively small distance the small magnetic loop is more sensitive to electric fields than a small electric field probe. The ratio of electric to magnetic fields are sometimes called the "field impedance" of the antenna. An antenna with a high field impedance has a dominant electric field, and an antenna with a low field impedance has a dominant magnetic field, when both are compared to freespace ratios at a large distance from the antenna.

At distances of about $\lambda/2$, the field impedances of all by physically large antennas are almost indistinguishable from each other, no matter what the antenna type. Strange things happen at closer distances, as we will soon see. Once we understand the basics, we might picture our antennas differently. We might find antennas don't respond to signals and noise like we thought, but at least we won't be as susceptible to the "noise" caused by rumors and folklore!

Induction field

The induction-field area involves the energy storage area around the antenna, and involves both electric and magnetic fields. These energy storage fields dominate the system near the antenna, although the radiation still exists. Contrary to some bizarre claims (i.e. CFA or CTHA Antennas), there is no possible way to "mix" induction fields and create a radiation field. The effects that cause each are different, and they follow different rules. The induction fields fall off much more rapidly with distance than radiation does, even though we "measure" or talk about each in terms of magnetic and electric effects.

We can move the fields around, concentrate them or spread them out, and even change the ratio of these fields...but every antenna has both electric and magnetic induction fields. *Whenever we have a time-varying electric field, it must be accompanied by a time-varying magnetic field.*

We might consider induction fields part of direct capacitive or inductive coupling between areas of the antenna and objects or space around the antenna. Close to the antenna, field intensity might not fall off as the rules normally predict with distance. Very near an antenna, fields are an accumulation of effects from sources at multiple directions and distances. It isn't always easy to understand what actually is happening very near an antenna, where the area of the

antenna is large compared to the distance where we are observing the effects. Near the antenna, pattern and field impedance is generally nothing like we might intuitively imagine.

It is the response in this area, generally within λ/10 distance from the antenna, that small "magnetic loop" and "electric dipole" antennas get their names.

Very close to a small loop antenna (but not necessarily very near the open ends), the magnetic field dominates. We could describe that effect by saying say the field impedance is "low". Conversely, near a small dipole or monopole the electric field dominates, the antenna has a high field impedance. Remember, this generally applies only with λ/10 distance from the antenna.
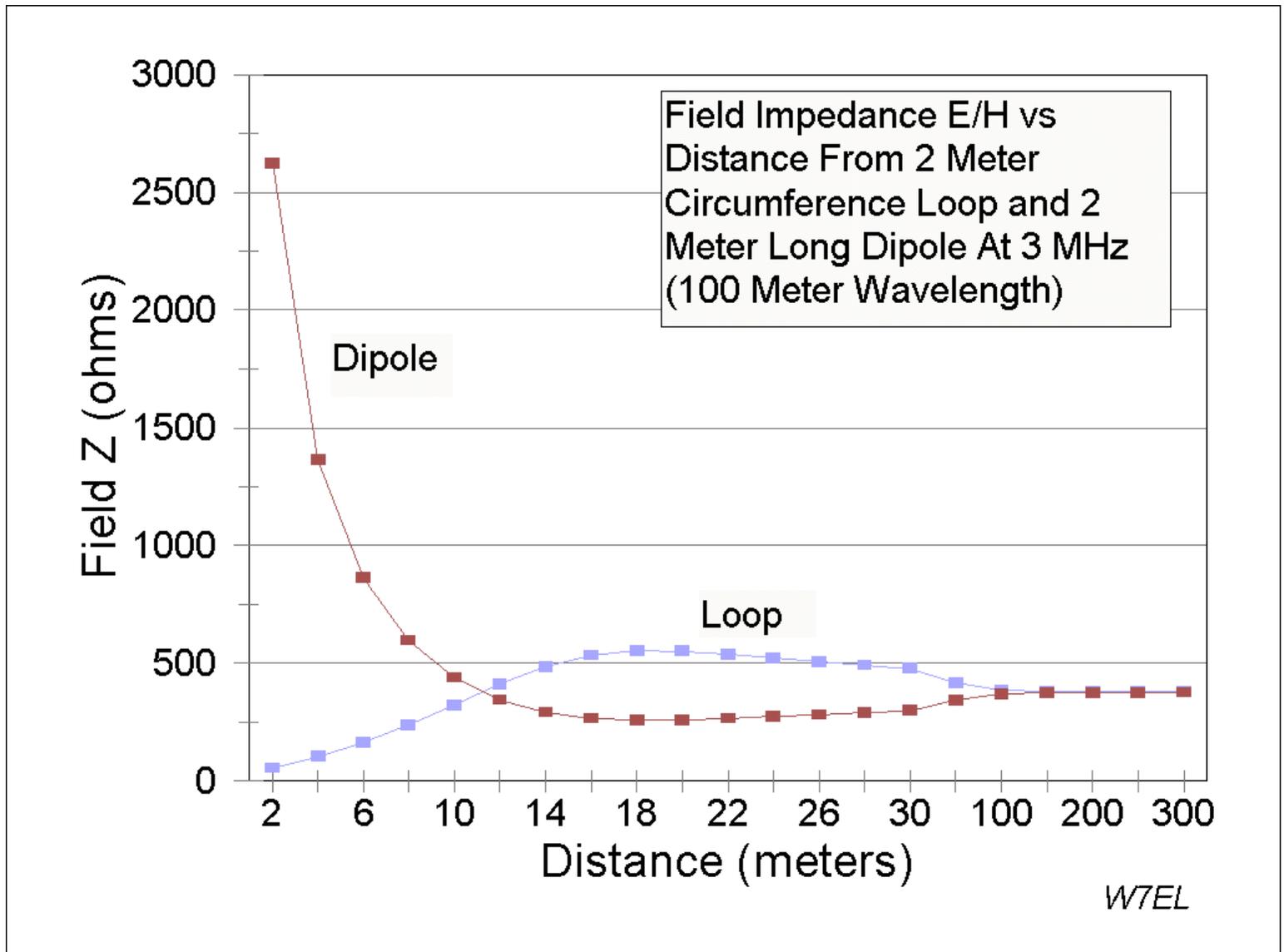
The since the distance of a wavelength in the above graph (thanks W7EL) is 100 meters, we can also considered the bottom scale as a percentage of a wavelength. We can see at about 11 percent of a wavelength (which would be about 50 feet on 160 meters), there is no difference in field impedance between a small loop and a small dipole. At distances beyond 50 on 160 meters, the loop actually has a *higher* field impedance than a dipole.

So much for the myth that a receiving antenna can sort good signals from bad signals (noise) by virtue of being "magnetic"! We not only don't have the response we might imagine, we probably have no idea if the close-by unwanted signal source is radiated from a source that is electric or magnetic field dominant. Successful noise reduction by virtue of

by antenna "style" would mostly be a matter of hitting a lucky combination through careful experimentation.

Near Field

The near field area is an area where the ultimate pattern is not fully formed. It is possible, with large arrays of small elements, to be out of the induction field region but still well within the near field area. Let's consider individual groups of elements as "cells", and the array a combination of small directional cells occupying a very large physical area.

This occurs in some of my broadside/end-fire combination arrays because physically small antennas forming individual end-fire cells are combined forming an array more than one wavelength across. The array is physically large, and at a distances of several wavelengths the combined field clearly originates from multiple sources that vary in distance or direction. The pattern is not perfectly formed because the signal from closer elements is less attenuated by distance, and the phase difference between the cells at a given azimuth is not the same as the ultimate phase further away at the same azimuth.

This applies to utility line radiation, where distant noise sources are conducted along the entire line. To the receiving array, it might appear that a single arc on an insulator a mile away is actually radiating from multiple directions with many phase relationships. The field the receiving antenna "sees" might appear to arrive from multiple directions, making it difficult or impossible to null the noise with a simple directional cell or more complicated array of directional cells. This would be a good application for a device like an antenna noise canceller, because the noise at the antenna terminals will be the vector sum of response to multiple radiated fields reaching the receiving antenna from a single noise source. In this situation a sample of noise taken near the power line could be combined out-of-phase with the sum of radiation points along the power line to null the noise.

Clearly a noise source or sources might also be outside the induction-field dominant area, and still not have a fully formed pattern. Antennas must be in the far field of the noise source, and the noise source must be in the far field of the antenna if patterns are to be predictable when nulling noise. Besides the fact larger distances provide greater attenuation of noise, it is also very desirable to locate our receiving antennas in the far field of noise sources, and noise sources in the far field of the antenna. This makes a good case for placing receiving antennas as far from other conductors (such as power lines) as possible.

This doesn't means urban or suburban dwellers should give up and move to rural areas, it just means the results will be less predictable and actual experimentation is likely required. Results aren't always predictable, unless the antenna is a few wavelengths from other potential unwanted signal radiators or objects that might affect antenna pattern.

Noise

The noise that limits our ability to hear a weak signal on the lower bands is almost always an accumulation of many signal sources. Below 18 MHz, the noise we hear on our receivers (even at the quietest sites) comes from terrestrial sources. Receiver noise is generally a mixture of local groundwave and ionosphere propagated noise sources, although some of us suffer with dominant noise sources located very close to our antennas.

Our locations fall into three basic "radio" categories that may or may not be related to our actual communities:

Note: noise levels quoted in this text are the average of three independent studies by Bell Labs, FCC Land Mobile Advisory committee, and the Institute for Telecommunication Sciences. Rural data are actual measurements of summer noontime and winter midnight noise at my location, several miles from high voltage transmission lines and far from any industrial or suburban populations.

Urban

In urban-type noise situations, noise arrives from multiple random sources through direct and groundwave propagation from local sources. One or more sources can actually be the induction-field zone of our antennas (in most cases the induction field dominates at distances less than $1/2\lambda$). Urban locations are the least desirable locations because typical noise floors average 16dB higher than suburban locations. There is often no evidence of winter night noise increase on 160 meters, since ionosphere-propagated noises are swamped out by the combined noise power of multiple local noise sources. Much of the noise sources are utility distribution lines, because of the large amount of hardware required to serve multiple users. Other noise sources are switching power supplies, arcing signs, and other unintentional man-made noise transmitters.

Suburban

Suburban locations average about 16 dB quieter than urban locations, and are about 20 dB noisier than rural locations. Noise generally is directional, arriving mostly from areas of densest population or the most noise-offensive power lines. Utility high-voltage transmission lines are often problematic at distances greater than a mile, and occasionally distribution lines can be problems. The recent influx of computers and switching power supplies has added a new dimension to suburban noise.

There is often a small increase in nighttime winter noise at exceptionally quiet suburban locations. This increase occurs when propagated terrestrial noise equals or exceeds local noise sources.

Rural

Rural locations, especially those miles from any population center, offer the quietest environment for low-band receiving. Daytime 160 meter noise levels are typically around 35-50 dB quieter than urban, more than 20 dB quieter than suburban locations. Nighttime brings a dramatic increase in low-band noise, as noise propagates in via the ionosphere from multiple distant sources.

Primary local noise sources are electric fences, switching power supplies, and utility lines. I can measure a 3 to 5dB daytime noise increase in the direction of two population centers, Barnesville (population 7500, distance 6 miles) and Forsyth (population 10,000, distance 6 miles) Georgia.

Typical daytime noise levels, measured on a 200 foot omni-directional vertical, are around -130 dBm with a 350 Hz bandwidth (noise power is directly proportional to receiver bandwidth) . Noise power increases about 5 to15 dB at night, when the band "opens". As in the case of suburban systems, directional antennas reduce noise power.

Nighttime is the "big equalizer", reducing the advantage of location as distant noises increase with improved propagation.

How Important is Gain?

When choosing one receiving antenna over another, never select an antenna by comparing antenna gain. Gain only affects receiving performance when the receiving system noise floor is determined by internal system noise. If we hear a definite decrease in noise when replacing the antenna with a dummy load of the same impedance, gain by itself is not a factor.

 Noise not only varies with the hour of day, it also varies with direction and season. Further complicating matters, noise power is directly proportion to receiver bandwidth. Going from a 2.5kHz filter to a 250 Hz filter with the same basic shape-factor reduces noise by 10 dB. When determining if system gain (for a given directivity) is adequate to establish system noise floor, tests must be made at the quietest operating time using the narrowest selectivity. In general, we should set system gain to allow a comfortable increase in noise when the antenna is connected at a time when propagation is poor but adequate to allow some contacts. This sometimes results in excessive gain when propagation is more favorable, but it is simple to remove gain (either by adding an attenuator or removing an amplifier) when using wider bandwidths or if conditions are noisier.

My receiving antennas have high dynamic range amplifiers using push-pull 2N5109 transistors. Amplifier gain is approximately 15 dB, and noise figure is approximately 5 dB. Gain compression occurs at approximately 30 dBm (1 watt) output. The blocking dynamic range of the amplifiers (using a 250 Hz bandwidth on the receiver) is around 180 dB, making my receivers the limiting factor. The amplifiers have internal bypass relays, allowing them to be easily bypassed when required by removing amplifier power.

Each of these amplifiers drives a four-way splitter, allowing any one-of-four receivers to connect to any antenna. Multiple receivers can select the same antenna simultaneously. Splitter system loss is 7 dB. Typical net system gain is around 8 dB to each receiver. Some amplifiers are designed to have slightly more gain and some less gain, because some antenna systems have a lower noise floor than others.
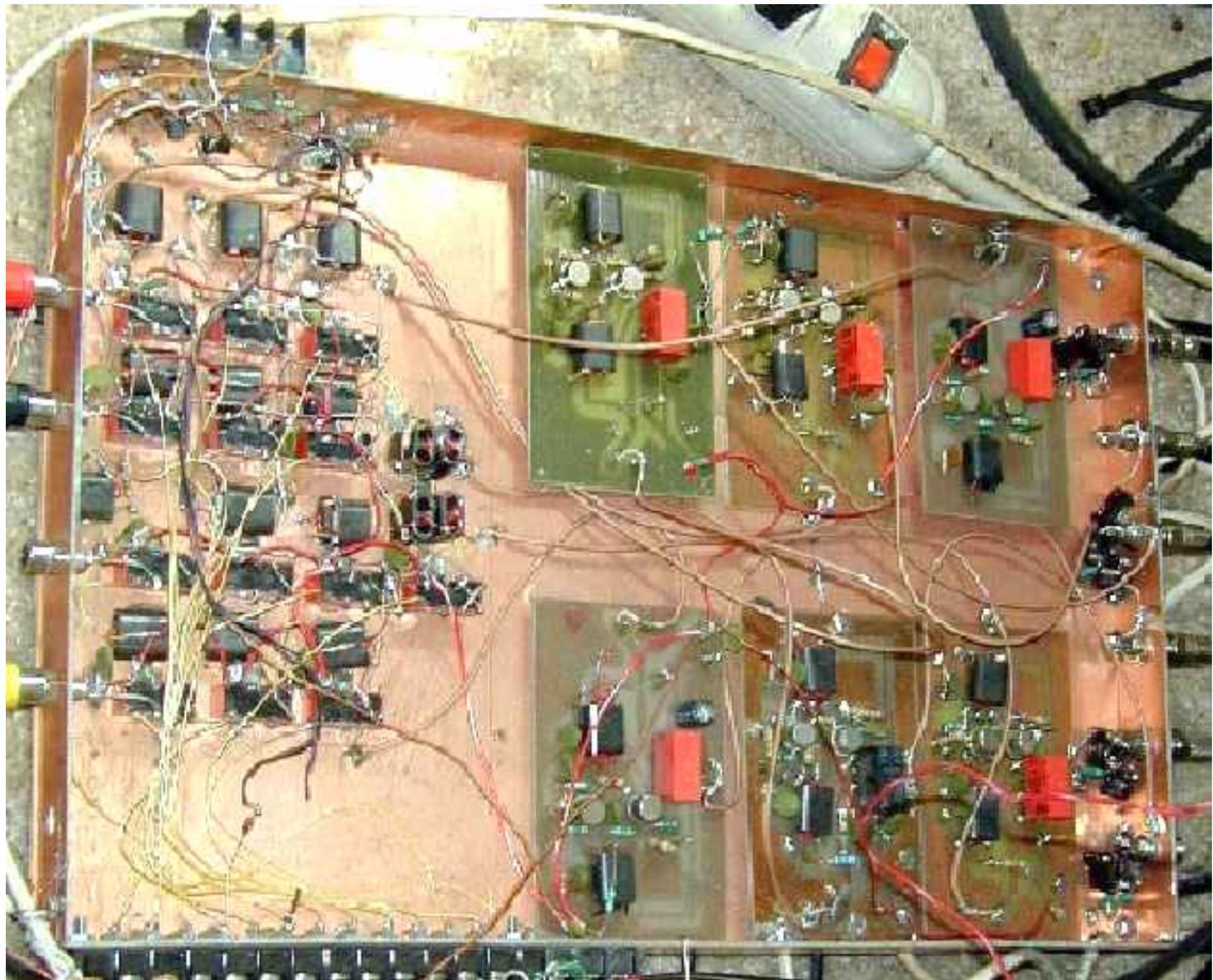
What Antenna Parameter is Important?

Once any receiving system limits on *external* noise, antenna pattern becomes the only thing that matters. Here are the patterns of two Beverage antenna systems, each with different gains. One antenna uses two side-by-side Beverages, while the other is a single Beverage. I'll include Eznec wire tables so you can model the antennas, and see the mistake we often make.

First, the wire table can look like this:

No.	End 1				End 2				Diameter	Segs
	X (ft)	Y (ft)	Z (ft)	Conn	X (ft)	Y (ft)	Z (ft)	Conn	(in)	
1	0	0	0	Ground	0	0	7	W2E1	#16	5
2	0	0	7	W1E2	0	780	7	W4E2	#16	50
3	70	0	7	W5E2	70	780	7	W6E2	#16	50
4	0	780	0	Ground	0	780	7	W2E2	#16	5
5	70	0	0	Ground	70	0	7	W3E1	#16	5
6	70	780	0	Ground	70	780	7	W3E2	#16	5

This is two Beverages 70 feet apart. If we terminate both of them and set the source of one at zero (so only one Beverage is active), the following pattern results:
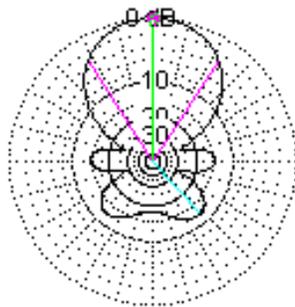
EZNEC

1.83 MHz

Azimuth Plot	
Elevation Angle	20.0 deg.
Outer Ring	-11.24dBi

Cursor Az	90.0 deg.
Gain	-11.24 dBi
	0.0 dBmax

Slice Max Gain	-11.24 dBi @ Az Angle = 90.0 deg.
Front/Back	18.11
Beamwidth	65.7 deg.; -3dB @ 56.8, 122.5 deg.
Sidelobe Gain	-23.33 dBi @ Az Angle = 312.0 deg.
Front/Sidelobe	12.09 dB

Now we add the second source, so both parallel Beverages are active. The pattern and gain is now:
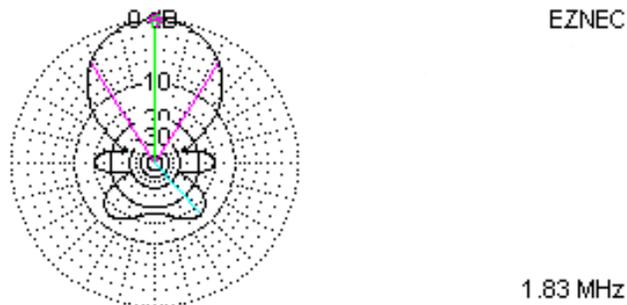
EZNEC

1.83 MHz

Azimuth Plot		Cursor Az	90.0 deg.
Elevation Angle	20.0 deg.	Gain	-8.2 dBi
Outer Ring	-8.2dBi		0.0 dBmax

Slice Max Gain	-8.2 dBi @ Az Angle = 90.0 deg.
Front/Back	18.11
Beamwidth	63.8 deg.; -3dB @ 58.1, 121.9 deg.
Sidelobe Gain	-20.76 dBi @ Az Angle = 312.0 deg.
Front/Sidelobe	12.56 dB

Gain is now -8.2dBi, about 3 dB better than the single wire's -11.24dBi. Did we gain any signal/noise ratio, for all the work of adding a second antenna? *Not a bit...the S/N ratio is exactly the same because the directivity is the same.* Since unwanted noise is external to the antenna, unless we have a change in pattern we can not have a change in S/N ratio. It is a common mistake, and absolutely incorrect, to compare gain and use gain to estimate receiving ability.

Eznec version 3.0 allows us to estimate directivity by modeling the antenna in three dimensions, and looking at the pattern. Here is the first Beverage modeled that way:
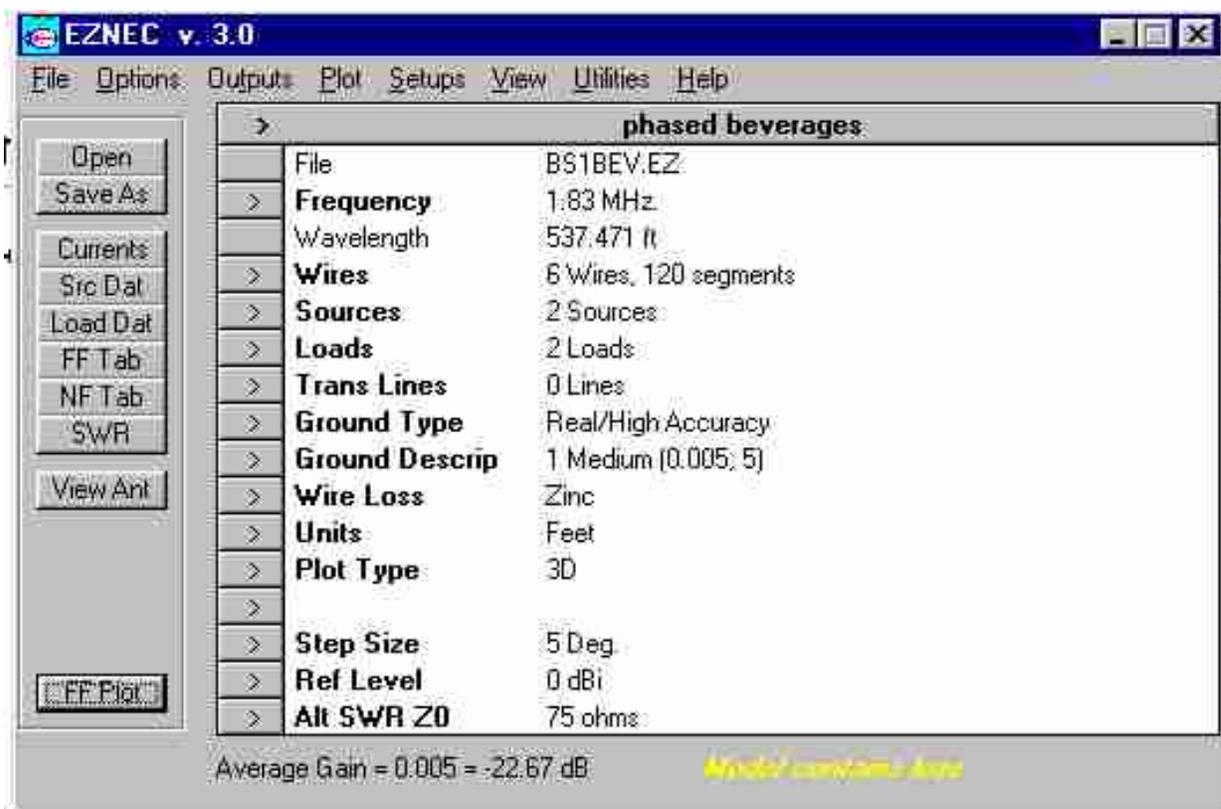
One source is set to 0 amperes, turning off one wire. Notice the "Average Gain" that appears in the lower area, -22.67 dB. If we subtract the average gain from the -11.24dBi gain at 20 degrees, we will have the ratio of response in the desired direction (from a two dimensional plot) to the overall gain of the antenna. This will be a positive number. The result is what we can consider a "receiving directivity factor" of the antenna expressed in dB. In this case, the RPF is 11.43dB.

Now let's look at the two wire antenna, using the same method:
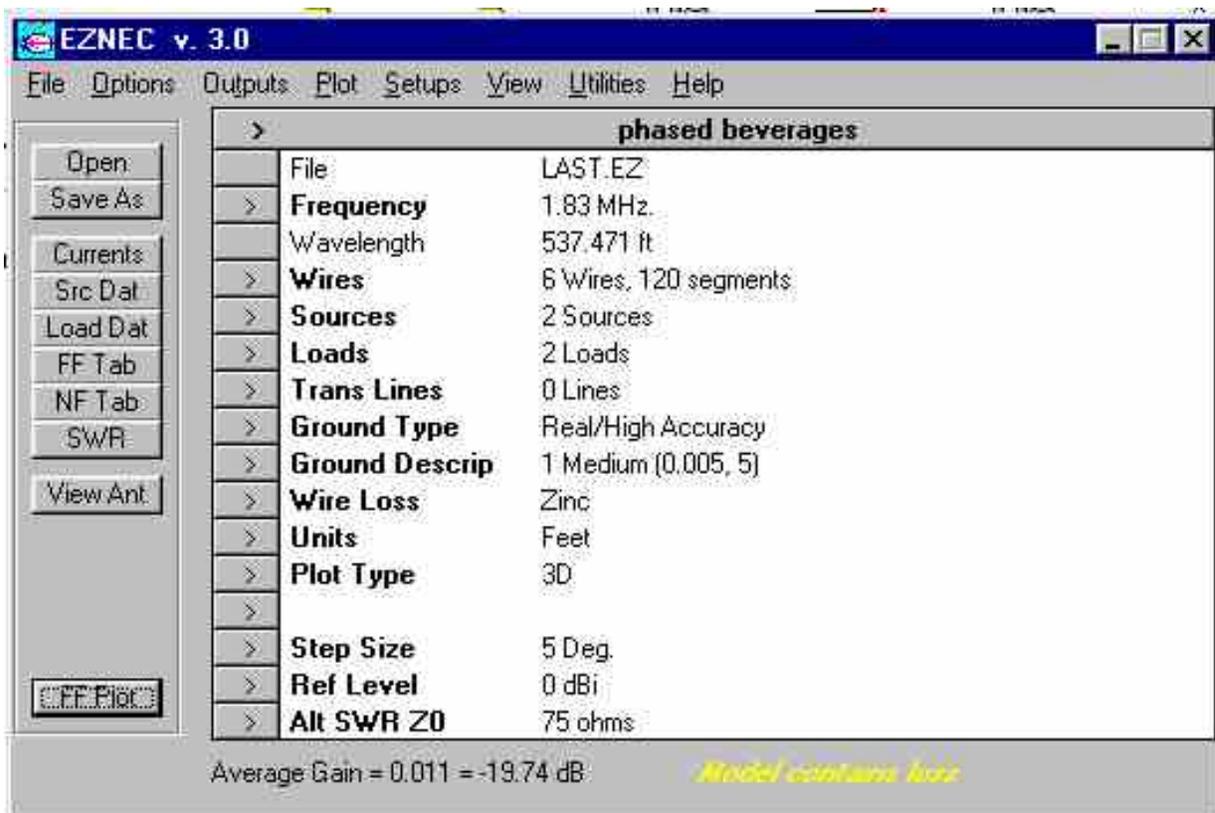
Notice I modeled this with both sources at 1 ampere, so both Beverages are active. Using the same process we used for the single source (which was the same as only having one Beverage) we see the RDF at 20 degrees is now (-8.2)-(-19.74) = 11.54db!

The effective receiving performance only increased 0.11dB, not worth the effort of adding a second antenna!

Changing spacing to 350 feet, we have the following two dimensional pattern:
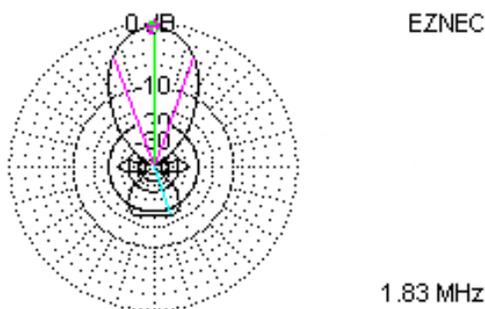
Azimuth Plot		Cursor Az	90.0 deg.
Elevation Angle	20.0 deg.	Gain	-7.86 dBi
Outer Ring	-7.44dBi		-0.42 dBmax

3D Max Gain	-7.44 dBi
Slice Max Gain	-7.86 dBi @ Az Angle = 90.0 deg.
Front/Back	17.83
Beamwidth	40.0 deg.; -3dB @ 70.0, 110.0 deg.
Sidelobe Gain	-25.1 dBi @ Az Angle = 290.0 deg.
Front/Sidelobe	17.24 dB

.....and the following data in the main menu of Eznec after modeling a three dimension plot.....
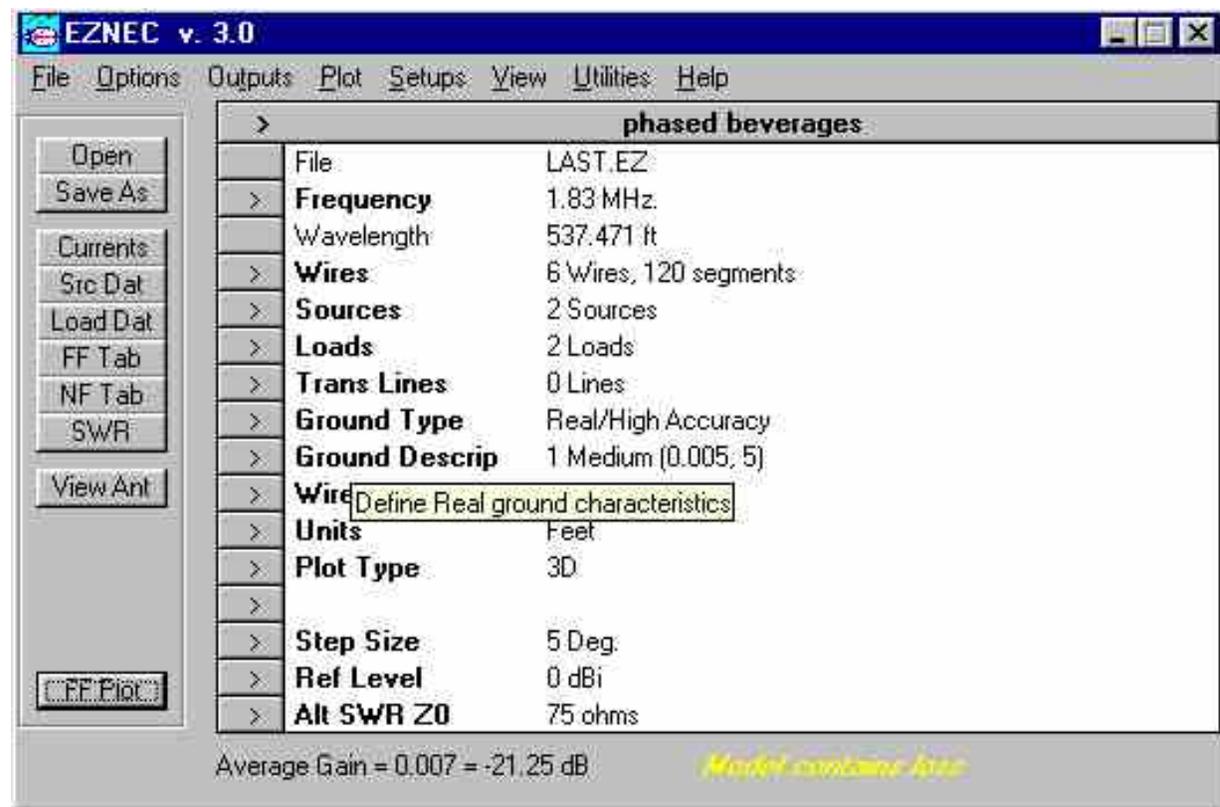
Taking (-7.44)-(-21.25) we now have a RDF of 13.81dB at a wave angle of 20 degrees. Comparing the three systems:

Antenna	Gain @ desired elev & azmth	Average gain	Rcvg Directivity Factor
single Beverage	-11.24dBi	-22.67dB	11.43dB
70 ft spaced Beverages	-8.2dBi	-19.74dB	11.54dB (+.11dB RDF)
350 ft spaced Beverages	-7.44dBi	-21.25dB	13.81dB (+2.38dB RDF)

Summary

Until Eznec's Windows version came along, it was very difficult to estimate performance. We often made mistakes, because we only compared gain. Using the above method, we can estimate antenna performance for evenly distributed noise. If there is no change in the Receiving Directivity Factor for similar pattern antennas, we can be sure we haven't changed overall performance for distant signal and noise sources.

There are a few things we have to keep in mind:

1.) If noise is not evenly distributed (which is often the case) performance will depend on the gain difference between desired signal direction (azimuth and elevation) to gain in the direction of noise(s). A 20dB null on the noise compared to gain in the desired signal direction will actually improve S/N ratio by 20 dB, if the noise from the null direction totally dominates all other noises.

2.) If noise arrives primarily from the same direction and angle as the desired signal (and assuming polarization of signals and noise is the same), there will be no S/N improvement.

3.) If noise originates in the near field of the antenna, all bets are off. Anything can happen.

All in all, the above method gives us a rough idea how various antennas compare in a environment with reasonably even noise distribution.

DX sound files

[Home]

Topband and DX Recordings and Sound Files

K1UO to JA on 160. The JA's are weak here!

These are mono and stereo recordings using either a FT1000D or phase-locked modified-R4C's with dual antenna systems for stereo diversity. This gives you an idea of how DX signals at this location under various conditions.

Clear band Beverage MONO directivity demonstration (W1AW-MONO-FT1000D). Every 5 seconds I switch directions. The first direction is NW, and I move around E, SE, S, SW, W, and NW. W1AW

Demo using storms in Texas and south. Every 5 seconds switch (right channel beverages, left channel verticals) NW, W, and SW. QRN

JA2 through noise, 10 seconds NW, five W, five SW. Receiver right channel phased Beverages, left channel Vertical Array QRN_ja2zjw

DX Stations On 160 Meters:

XU7ACB Long path (in evening) xu7acb121001_LP.wav

XU7ACB short path (my sunrise) xu7acb_am.wav

JT1/K4ZW short path (morning) jt1_k4zw112801.wav

VK9ML short path (morning) vk9ml041702_am.wav

BQ9P 160m (not on peak) Oct 11, 2003 around 1040Z

JA opening on good day Oct 12, 2003

Big file of Europe from east path during severe Aurora on Oct 30,2003

9V1GO from NW path (unusual path, 9V is normally skewed SW) Nov 03, 2003

On 80 meters:

A51B SW path 80-meters Feb 19 2003 1200-1215Z

Small Vertical Arrays

Small Vertical Arrays

While this is mainly intended as a guide to help install arrays, I have stock of all components in these antennas. If you are really stuck, I can provide turn-key systems that are tested and tuned, include phase-matched cables and phasing systems. I can supply custom arrays of up to eight directions using up to 16-elements from stock components and assemblies, tested on reliable commercial test equipment.

Never Pick Gain, F/B, or Take-off Angle as a Parameter for Weak Signal Receiving!

Arrays of small verticals provide excellent receiving performance when systems are designed and installed properly. As pointed out in other articles, there are key differences between receiving and transmitting systems. Parameters considered important in large transmitting systems are sometimes far from optimum in receiving applications, especially systems using small antennas. One key item is gain. Contrary to common opinion, more gain does not translate to better receiving, once the receiver is limited by external noise reaching the receiver. The key parameter is directivity, which may or may not have a parallel relationship with gain. Gain includes efficiency, directivity excludes efficiency. This important consideration applies to systems discussed below.

Noise rarely comes from one direction, or a narrow range of directions. With that in mind, antennas will be compared by a receiving directivity factor calculated with Eznec's (version 3) average gain. Removing efficiency from the equation allows direct comparison of receiving systems since directivity, not gain, is the determining factor in selecting an HF or LF receiving array. S/N ratio is very dependent on nulling or rejecting unwanted signals or noise.

It is important to locate receiving antennas as far as possible from radiators or re-radiators of unwanted signals and noise. Always remember noise has exactly the same characteristics, so far as an antenna is concerned, as signals from *intentional* transmitters. There is no way to sort "good signals" from "bad noise" except through the directional characteristics of your receiving antenna. Noise is not electric field dominant. Desired signals are not magnetic field dominant. The field impedances are all the same except near the antenna or source, and near the antenna or source coupling to multiple unknown sources is largely unpredictable. The truth is, it is anyone's guess what field impedance is actually best!

A small loop antenna, at a distance of a few meters, is magnetic field dominant. Here is an important fact few people, outside of those who work with nearfield systems know. At a distance of an eighth wave and larger a small magnetic loop becomes electric field dominant! Conversely, a small voltage probe becomes magnetic field dominant at about the same distance! The fields reverse dominance because of phase shift between the fields as the radiation fields start to overtake the induction fields.

Susceptibility to unwanted near-field and induction field coupling between receiving antennas and large transmitting antennas or noise sources is obviously largely unpredictable, although many problems can be corrected through changes in antenna placement or detuning structures and/or canceling the radiation from surrounding structures. When dealing with nulls, a modest amount of re-radiation from surrounding conductors can make a large difference in system performance, but the key is to watch overall directivity. Anything that reduces directivity will reduce the S/N ratio of a receiving system. The reduction is directly by the amount of null reduction *ONLY* when noise comes very predominantly from within the area encompassed by the deeper areas of null. *We always, unless we have noise from a specific direction and angle all of the time, want a wider more modest depth null in favor of having a sharp point with a deep null and a wide-nosed response.*

One of my best arrays on Europe is only a few hundred feet from a transmitting four-square, clearly in the near field of the four-square. There is no detectable influence on this array when it "looks away" from the four-square, although there is a quite noticeable reduction of F/B ratio when beaming back into my transmitting antennas. The null to the SW is very deep, in excess of 35dB, regardless of four-square tuning. The converse is not true, the null NE when looking SW is only 10-15dB deep unless I detune the antennas. Yet the lower null depth barely causes a detectable noise increase, because the directivity does not change much. It is only when a very dominant noise (or QRM) arrives from the NE that this array becomes almost useless (compared to other arrays with deeper nulls in the NE direction). Of course detuning the transmitting antennas completely restores southwest performance, even though spacing is close.

Building Blocks For Arrays Using Verticals

There are four key areas overlooked in most published receiving arrays using small elements. Common oversights in element and phasing system design cause the antenna to be more critical to adjust, less stable, and provide a poorer pattern even if array elements happen to be working.

The most aspects are:

1.) Elements must be very low Q (wide bandwidth). They should have little reactance change with frequency or weather.

2.) Elements must be heavily swamped with loss. Mutual coupling effects must not change

element impedance.

3.) The phasing system must be designed for the impedances that actually appear at the phasing system. Transmitting-type boxes, because the systems have low loss, must have high mutual coupling effects and radically different element impedances. By definition, any given system _must_ be seriously flawed in one application or the other!

4.) The phasing system must be stable and have very broad bandwidth characteristics.

Before building an array, we must select an element style that we can live with. Eznec and other programs have made this process simple.

The Basic Element

Length

A receiving array element should be as short as possible but still maintain sufficient sensitivity (gain) to ensure external noise exceeds receiver noise. My 160 Meter Band elements are about 20-foot vertical height. I've found all of my arrays with elements that height have overall sensitivities (gain) on par with my Beverage antennas, and that the signal levels are very easy to deal with.

Keep in mind that our systems require more gain as receiver selectivity is decreased. The noise floor drops in direct proportion to selectivity increase, and a change from 2.5 kHz selectivity to 250 Hz selectivity reduces noise voltage or power by 10dB. Signal level, however, remains the same for the same transmitted power within our receiver's bandwidth.

20-foot tall elements with reasonable element spacing always provide more than enough signal to operate through nearly 1/2 mile of F-11 CATV cable (similar to RG-11 or RG-8 cables in size) at my very quiet rural location *before amplification.* 3-5dB noise figure amplifiers are adequate to establish S/N ratio by arriving noise even when placed after the signals travel through those long cable lengths. To gauge my noise floor, a standard FT1000 with preamplifier "on" on my 200-foot vertical has less than S3 noise in the SSB position at mid-day. Unless you have less noon-time noise than that, you will certainly not require an amplifier at the antenna! The only exception is if you have very close element spacing, because close spacing decreases antenna sensitivity (gain).

Mechanical and Electrical Concerns

I use two basic mechanical configurations of elements. One system uses steel electrical conduit on 1/2-inch fiberglass rods (rods driven directly into the ground) with four "loading" wires, while my other system uses stronger chain link fence top-rail mounted on wooden posts. Both systems handled weather from ice storms (where the antennas were coated with almost a radial inch of ice) to high winds without problems. I have had no electrical connection problems, and no weather

detuning problems. ***It is not necessary to "insulate" the antennas mounted on wooden posts because system Q is very low and impedances are modest.*** Wet posts will have no deleterious effect on performance, although I would always place the loading system (and the base of the verticals) above snow depth. ***It is not necessary to use high-Q loading inductors.*** The only requirement is that inductors remain relatively stable in characteristics with climatic changes.

My systems are normalized at 75- ohms for several reasons:

1.) 75-ohm feed produces a wider VSWR bandwidth than 50-ohm feed systems, the swamping resistive losses are about 50% higher.

2.) CATV cable suitable for direct burial is inexpensive and connectors are inexpensive, reliable, and easy to install.

I use F6 flooded CATV cable for local cables in arrays, and F11 (RG-11 size) or 5/8" flooded CATV cable for trunk leads. It is NOT necessary to use double or triple shielded cable, you gain nothing. But you do want to use good quality cable that will last years without weather changes.

Practical Verticals

My rectangular arrays use elements with four 20 foot long #16 loading wires, insulated by fishing line used to support the wires. I terminate the guying 20 feet out from the base of the antenna. The entire structure is self-resonant on 80 meters. The large "hat" makes current essentially uniform throughout the vertical element while minimizing unwanted sensitivity to high angle radiation, and also supports the elements.

No wires selected
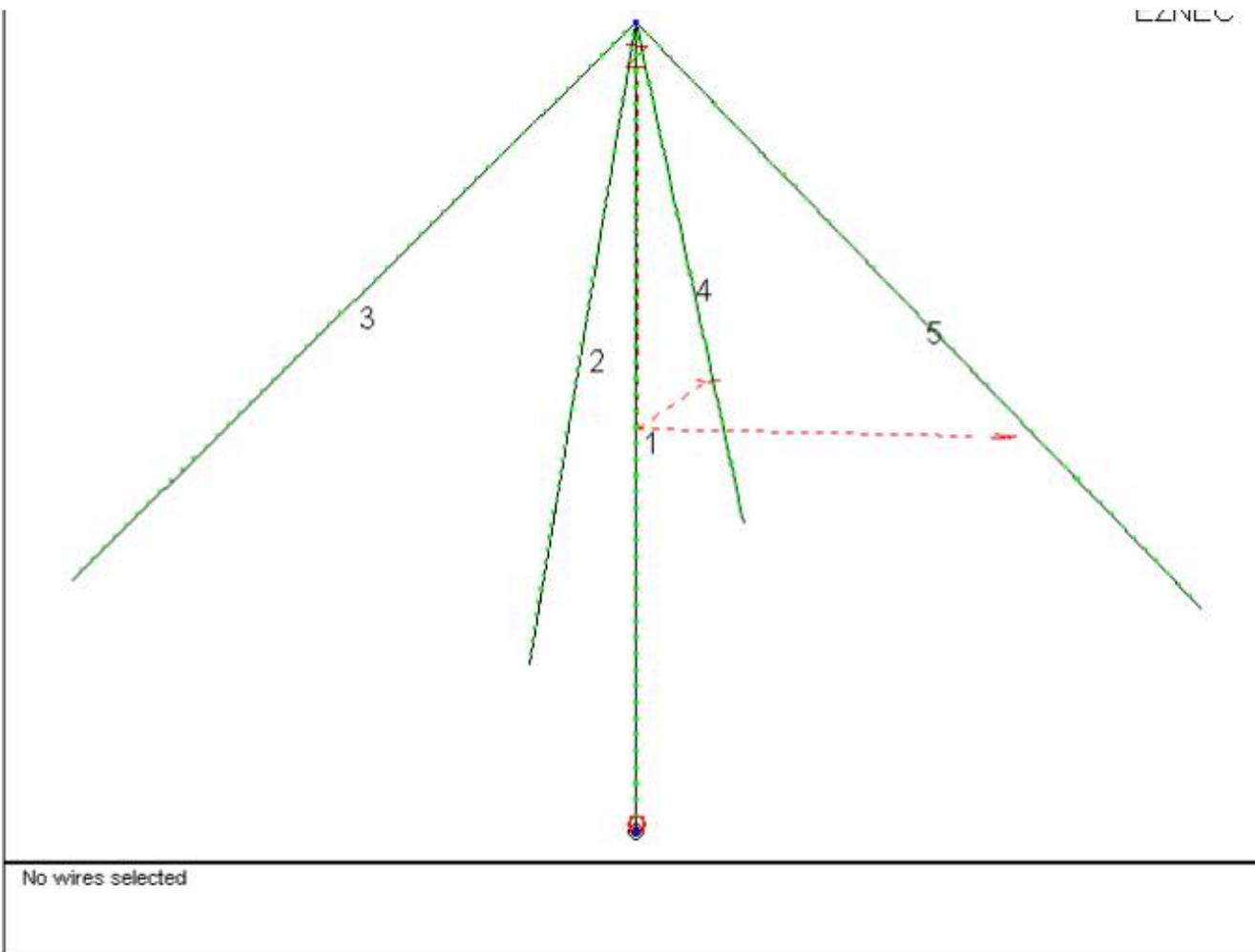
Eznec Download

Eznec [RXvrhat](#)

This structure is base loaded with a series L/R combination of approximately 30uH and contains a *total* loss resistance of 75 ohms. This resistance includes resistive losses related to inductor Q, as well as ground system loss resistance. I used small molded choke inductors, although other components will work. My system requires only 56 ohms of lumped resistance to bring base resistance to 75 ohms.

Each element requires a *stable* ground system. Ground loss is not important, but long and short term loss stability with climatic changes is very important. I use a minimum of four buried radials, each 1/8- to 1/4-wl long, on each element. Always place radials directly under each hat wire. Do *NOT* use small elevated radial systems or grossly non-symmetrical radial systems! Elevated radials will reduce VSWR bandwidth of the array, introducing unwanted phase shift. They also make the system susceptible to high angle signals, and are more susceptible to common-mode noise on feedlines and other conductors around the antenna than buried or earthed radials. It is not necessary to bury radials, but if the radials aren't buried multiple ground rods are a good idea. The feedline should also be buried or if laid on the ground "choked" with high permeability ferrite beads near

each element.

Feedline Length

Unlike transmitting arrays, it is not necessary to use odd-quarter wave lines. It might be a tiny bit better if you use exact multiples of 1/4 wl, but even 1/2 wl lines work perfectly fine. Multiples of 1/4 wl work better in cases where you might fail to match antenna impedances to the transmission lines correctly. The reason of this is that phase shift in a transmission line is independent of line SWR when the feedline is ANY multiple of 90-degrees. With ANY phasing system having standing waves on the feedline, you can properly feed the system by supplying equal currents to any feedline an even multiple of 1/2 wl. Any feedline having odd multiples of 1/4 wl requires equal voltages feeding the line. The phasing systems I use, unlike transmitting systems, are designed to supply either equal voltages or equal currents! The proper ratio adjustments are easily made. You will, however, have slightly less phase error if you use any multiple of 1/4 wl when the lines are mismatched.

Beware that foam cables are NOT .82 or any other standard velocity factor. They range from the .70 range up to around .92 in velocity factor, depending on the ratio of gas to material in the dielectric. Only solid dielectric cable are predictable without measuring the cable.

Tuning the Elements

After careful planning and selecting the type of array, you should install the elements and the array's internal feedlines. Each element must be evaluated with an antenna analyzer that measures resistance and reactance. Connect the analyzer at the element's feedpoint, and follow these steps:
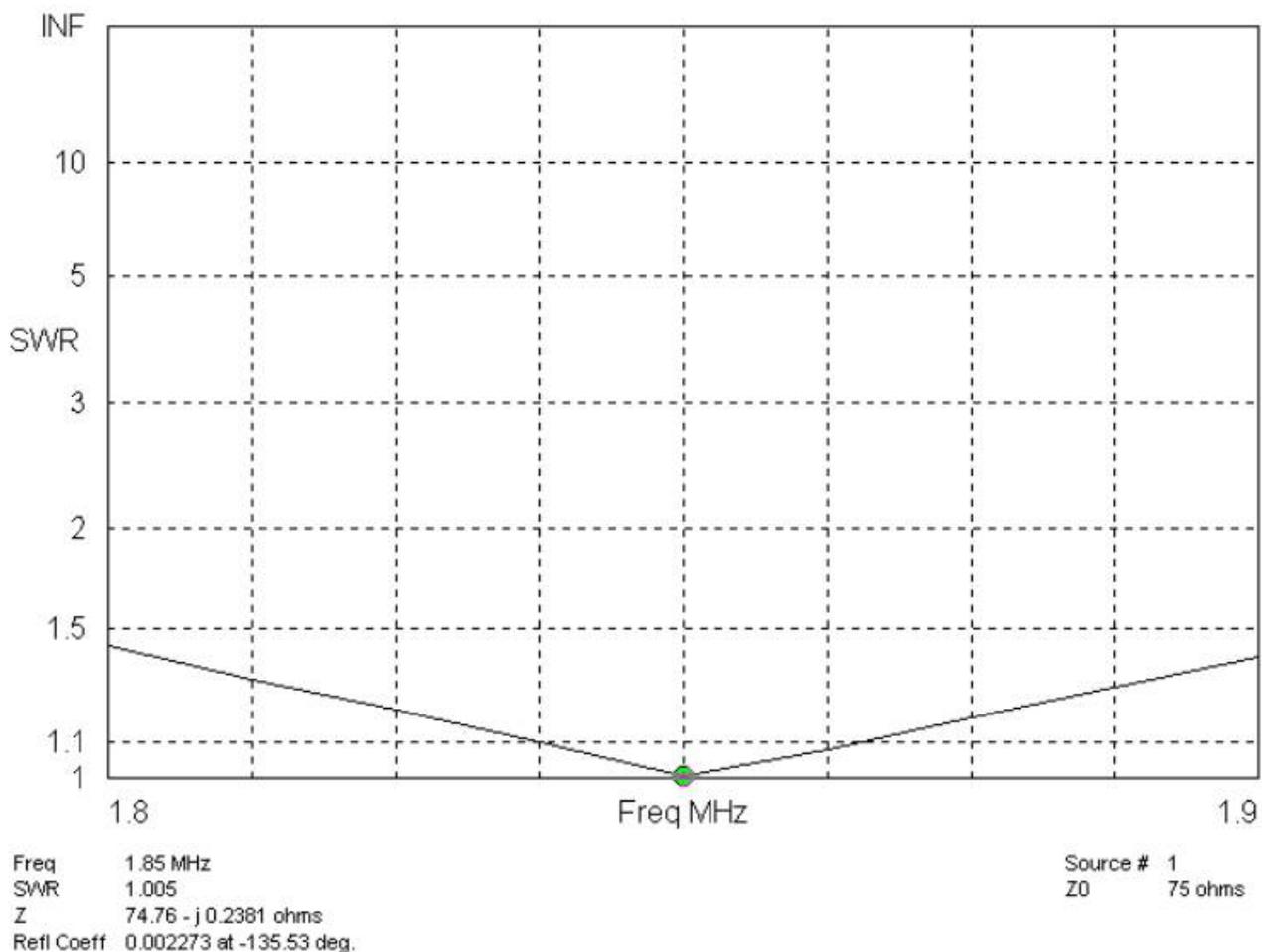
1.) Using a two foot long or less jumper cable, measure the antenna without a loading coil. Check the resonance for predicted values. It should be within several percent of the modeled self-resonant frequency.

2.) Touch the shield of the feedline in the array to the case of the analyzer or the vertical's ground. If impedance changes more than five percent, you need a better ground system.

3.) Install the loading inductance predicted, and sweep the desired frequency range for lowest SWR and zero reactance. If your analyzer is working correctly the lowest SWR will be at zero reactance, or very close to that frequency.

4.) Fine tune the inductance to make the antenna resonant at the desired frequency. You do this by adding or removing small inductors in series with a main inductor, or by adding or removing turns. Fine adjustments can be made by squeezing or spreading turns on the main loading coil's form, if it is a toroid or non-potted construction.

5.) Add enough series resistance to bring antenna impedance to 75 ohms at resonance (assuming you use 75 ohm cables).

6.) Check the feedpoint again for stability by connecting and disconnecting the shield of the unused array transmission line from the case of the analyzer or the ground system connection point. Again it should remain within 5%, or you need to improve the ground. If you can not improve the ground, you will have to isolate the ground by using a choke balun on the feedline.

Once one element is tested and proven, you should be able to duplicate that element with near-perfect results. I remove the matching system and take it to the test bench, and find a series C/R combination that produces the same resonant frequency. Multiple networks can be constructed on the test bench, and then moved to each element. My elements normally fall within 20 kHz of each other, any large difference in impedance or resonant frequency is a sign of potential performance problems.

Here is the 75 ohm SWR plot of this element:

Freq	1.85 MHz		Source #	1
SWR	1.005		Z0	75 ohms
Z	74.76 - j 0.2381 ohms			
Refl Coeff	0.002273 at -135.53 deg.			

Bandwidth is excellent, and sensitivity (gain) including all losses is -13 dBi. This element is almost perfect for use in small receiving arrays, since signal level and bandwidth are very good. Because of

the large amount of capacitance and the resistance loading, the element will not significantly change phase or sensitivity with frequency over the entire 160-meter band! In addition, it has very little response to high angle signals (and noise).

In circular arrays, hat wires can be extended to ~35 feet with only three loading wires used. My 350-foot diameter arrays position two loading wires (using tarred nylon fishing net string for insulation) in line with the perimeter of the antennas, while the third wire on each element is used to "pull out" from the array center. This tensions the perimeter wires and guys the entire structure. Even though hat wires are not spaced exactly 120 degrees apart, the effects on sensitivity to high angle signals are insignificant.

Arrays of Elements

Arrays of short verticals have both advantages and disadvantages compared to other antennas, such as Beverages and elongated loops. The comparison is:

	Elongated Loops	Beverages	Verticals and Wide-spaced Beverages
Area required	small <3/8-wl	large >1-wl	1/2- to 3-wl
Effort	modest	minimal	modest to large
Pattern	worse than 2-el vert, poorest	about same as 4-sq, middle	middle to best
Bandwidth	multiple bands	multiple bands	single/multiple band
Signal level	Low	Modest	Modest

The primary advantage of arrays of short vertical arrays are excellent pattern and reasonable signal levels. Unlike balanced elongated loop systems, they are non-critical for feedline routing (other than keeping the feedline on the ground), matching transformers (transformers are not even required!), and earth conditions around the antenna. They have the same approximate output as simple Beverage arrays, and can still be made to work over very wide frequency ranges. The disadvantage is they are more complex, and require bandswitching to work on two bands.

Note: I am finishing the layout of an FET amplified array that will allow ten-foot non-hat verticals that work from VLF to 80-meters and above, with the same basic directional characteristics over that range without switching.

Broadside-Endfire Array (soon to be an

article)

The most directive four-element antenna possible is a broadside-endfire array. I use 70 feet endfire spacing and 330 feet broadside spacing between endfire cells, and get a pattern like this:

EZNEC

1.85 MHz

Azimuth Plot
Elevation Angle   14.0 deg.
Outer Ring   -10.7dBi

Slice Max Gain   -10.7 dBi @ Az Angle = 90.0 deg.
Front/Back   22.22
Beamwidth   49.4 deg.; -3dB @ 65.3, 114.7 deg.
Sidelobe Gain   -29.94 dBi @ Az Angle = 174.0 deg.
Front/Sidelobe   19.24 dB

Cursor Az   7.0 deg.
Gain   -29.94 dBi
             -19.24 dBmax
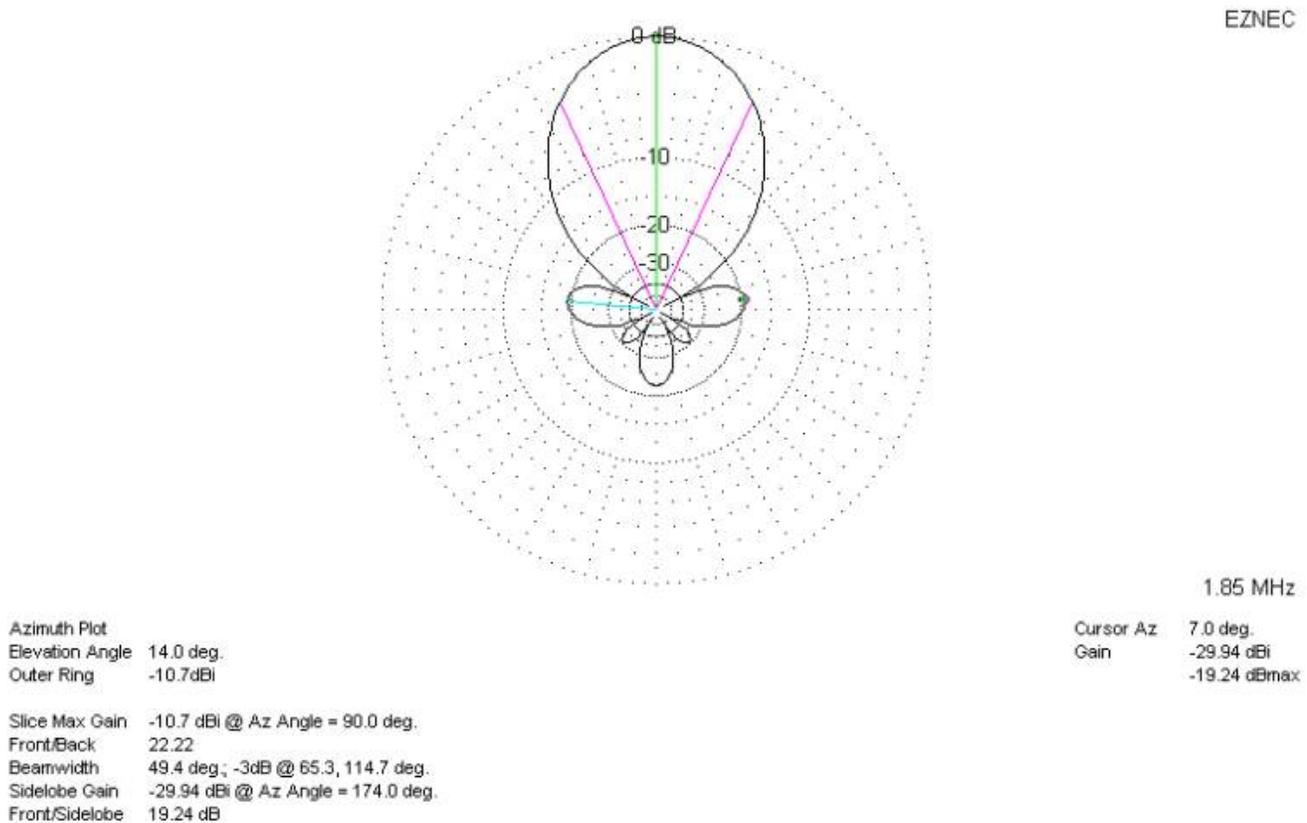
This array has an RDF of 13dB (two dB more than the large 4-square), and a HPBW of 47 degrees! The performance is similar to a circular array with eight elements in a 350 foot diameter circle, that allows directional selection every 45 degrees. The circular array uses techniques similar to the broadside-endfire, except the relay system is modified to switch eight antennas.

My next planned array is an array of four four-squares modified to fire eight directions, with 500 foot center point separation between each four-square array. I've found that 1000 feet is about the maximum physical separation allowing signals to be reliably combined on 160-meters. Arrays occupying areas larger than 2 wavelengths have far too much random phase and amplitude shift, preventing reliable combining of signals.

Receiving 4-Square

[[Home](#)] [[Up](#)]

Receiving 4-Square Array

Four small elements in a square layout, between 1/8- and 1/4-wl side length, can be used to form a four- direction array. Performance will be approximately equal to a 1 wl Beverage. The elements can be similar to those described in [small vertical arrays](#).

As mentioned earlier, conventional transmitting phasing systems might be easy to obtain, but they are *NOT correct* when used in receiving arrays with short elements. Since each element has the same impedance in a lossy receiving array, the phasing unit "sees" 75-ohms from each element. When the antenna presents equal impedances at each port, the phasing system should source equal power at each port (which is also equal voltage and equal current)!

Use of a single 180-degree phase-inversion transformer for the center element along with delay lines approximately equal to effective element spacing create a very wide bandwidth phasing system. ***The phasing system is usable over at least one octave with appropriate element spacing and design!***

Upper and Lower Frequency Limits

Assuming you have ***broadband active elements*** of good design, the array will be useful when array side-length ranges from ~1/16wl spacing to 1/3 wl. The phasing system and delay lines are usable at any frequency, but the array spatial delay will provide useful end-fire patterns where array side-length is approximately 1/3-wl or longer. The appearance of grating lobes and dimpling of the front lobe limits upper frequency usefulness.

As frequency is decreased, ***array sensitivity*** eventually drops to unusable levels. The effect comes from two effects:

- Elements become electrically shorter, reducing element and array sensitivity.
- Element spacing becomes smaller in electrical degrees, reducing array sensitivity.

Shorter elements, in electrical degrees, mean less sensitivity. Sensitivity obviously must drop with frequency, for a fixed element height.

Element nulling always fully subtracts with reduced spacing. With any fixed spacing distance, a lower frequency array sensitivity limit appears when "forward" voltage phase vectors become too close to -180 degrees.

Element and array phasing-sensitivity eventually push the antenna into the receiver system's noise-floor, and this becomes the lower frequency cutoff. Pattern is maintained down to that cut-off, assuming the elements remain matched to the feedlines.

Even without broadband active elements, multiband or broadband arrays are practical with ONE fixed element group and a single unchanged phasing system. Element resonance and resistance would be the only parameter that would require switching.

Pattern with 70- foot per side spacing, on 1.8 MHz is:

EZNEC

1.8 MHz
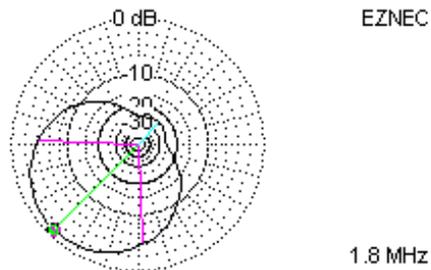
Azimuth Plot		Cursor Az	225.0 deg.
Elevation Angle	20.0 deg.	Gain	-11.51 dBi
Outer Ring	-10.59dBi		-0.92 dBmax

3D Max Gain	-10.59 dBi
Slice Max Gain	-11.51 dBi @ Az Angle = 225.0 deg.
Front/Back	23.19
Beamwidth	95.0 deg.; -3dB @ 177.5, 272.5 deg.
Sidelobe Gain	-34.7 dBi @ Az Angle = 45.0 deg.
Front/Sidelobe	23.19 dB

There is a slight distribution problem with smaller size arrays because of mutual coupling. This can be compensated if 1/4-wl feedlines are used on the lowest band with the phasing coupler shown. RDF (receiving directivity factor) is around 10dB. This is about 3dB higher than the best single Pennant antennas, and on par with a very long Beverage.

If the array is 140 foot per side, the pattern is:

EZNEC

1.8 MHz
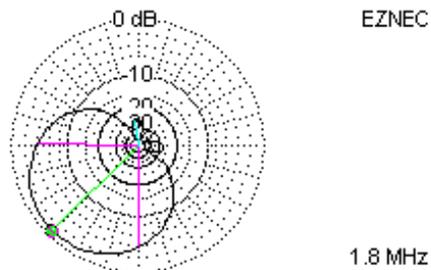
Azimuth Plot		Cursor Az	225.0 deg.
Elevation Angle	20.0 deg.	Gain	-6.88 dBi
Outer Ring	-6.14dBi		-0.74 dBmax

3D Max Gain	-6.14 dBi
Slice Max Gain	-6.88 dBi @ Az Angle = 225.0 deg.
Front/Back	31.17
Beamwidth	91.2 deg.; -3dB @ 179.4, 270.6 deg.
Sidelobe Gain	-33.16 dBi @ Az Angle = 95.0 deg.
Front/Sidelobe	26.28 dB

Doubling array size only provides about 1dB more directivity (note that HPBW only changes a few degrees), although signal level increases about 4dB! This array would make an ideal portable antenna for limited-space two-band operation, a SPDT relay at each antenna could be used to switch loading networks for each band.

The phasing system should look like this:

W8JI RX 4-Square Phasing
copyright 1/02/2001 W8JI
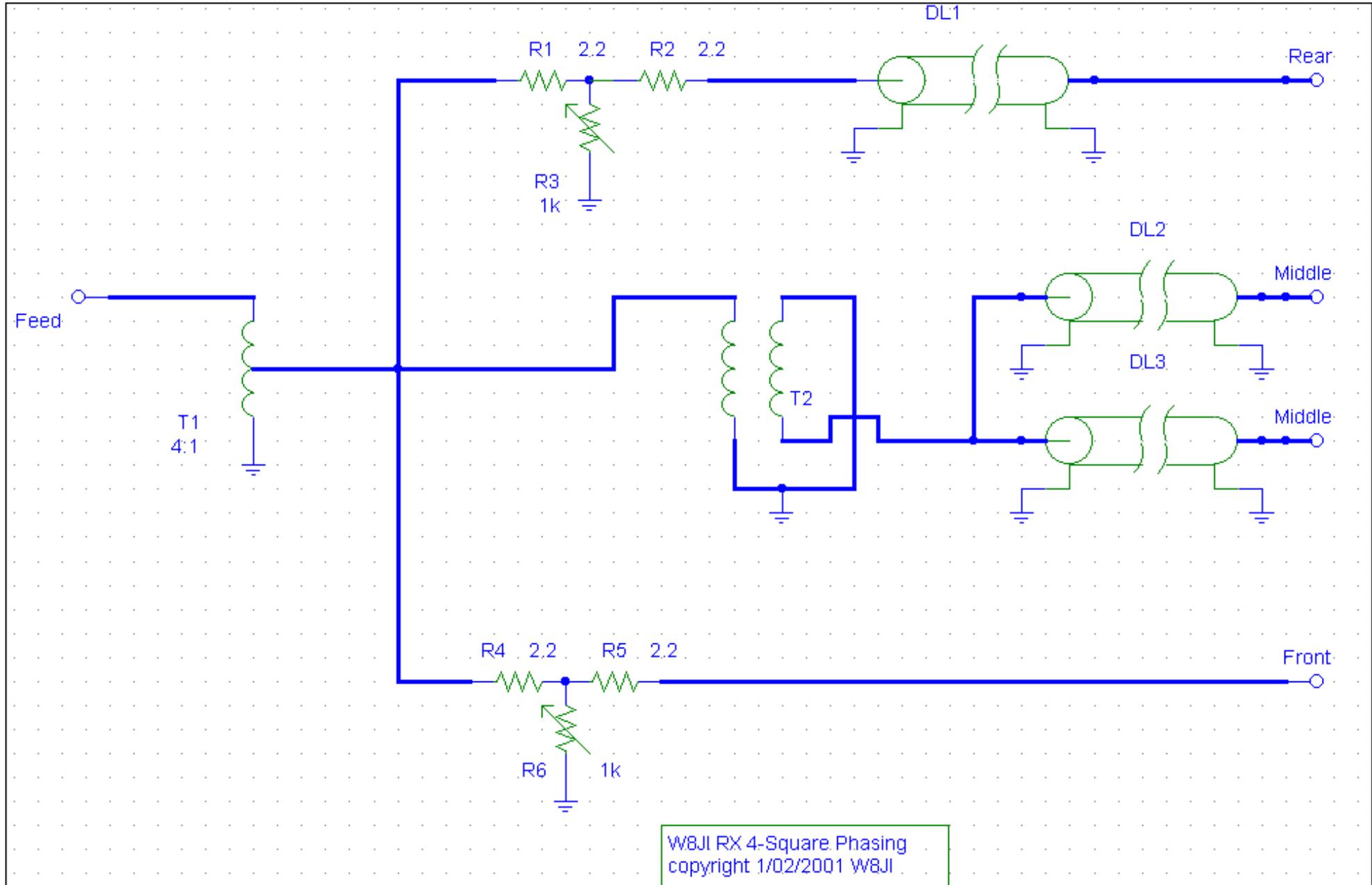
T1 is a 4:1 impedance ratio (2:1 turns ratio) auto transformer. A suitable 100 kHz to 20 MHz transformer would use 5 to 10 turns of #22 wire (wire size is not critical) in a twisted pair, connected with opposite ends in series. A Fair Rite Products 2873000202 core is recommended, and available from me if you can not find one. (A turn is one complete pass of the twisted wires through both holes in the binocular core). This transformer steps the 18.75-ohm impedance, produced by paralleling four 75-ohm lines, up to the feedline output impedance of 75-ohms.

T2 is a 1:1 ratio phase inverting transformer. Wire size is not critical as long as the winding fits the core. A good choice is a twisted pair of #22 wires, making five to ten passes through a Fair Rite Products 2873000202 core. This is the proper core! This is the same transformer used in T1, with a different circuit connection. Again, I will help you with complete tested transformers if you can not locate materials.

A low attenuation value ~ 75-ohm attenuator is inserted in the line to the front antenna and rear antennas. This attenuator compensates for slightly different losses in transformers and delay lines. Although it absolutely isn't a requirement when using wider spacings (1/4 wl) with proper transformer and element construction and materials, they can be used to "tweak" the array to extremely high F/B ratios. The attenuators become more and more important as the array becomes smaller or less perfectly constructed and tuned.

DL1 should be slightly shorter electrically than the (around .9 times the) corner-to-corner distance across the array (diagonal distance). This line MUST be the same impedance as the element feedlines.

DL2 and DL3 are exactly half the length of DL1. These lines must be the same impedance as the array's element feedlines.

NOTE: The delay lines are slightly shorter than element spacing. This elevates the null a small amount above the horizon directly off the rear, forming a cone-shaped null reaching the ground an equal number of degrees either side of the rear. I generally use about 10-20 degrees less than the element physical separation distance of a line running between diagonal corners of the array for delay lines on the highest band.

If you model this antenna, remember phase is inverted. Effective phase shift from the delay line actually becomes a phase lead that is 180- degrees around a circle from the delay you might expect. For example, a 90- degree delay line actually provides 90- degrees phase advance (or -270 delay) because of the inversion, while a 60- degree delay line provides 120- degrees advance (which is the same as 240-degree delay) .

This "trick" causes phase-shift to track changes in element spacing with frequency, allowing the same delay line to work on multiple bands. The delay always is the proper length, which is slightly shorter than element spacing. It also causes the array to fire in the direction of the common-point element, since the feedline attachment point is the direction of lagging phase.

See the Crossfire phasing article.

Why Are the Lines nearly equal to the spacing for the elements they serve?

For an explanation, see the Cross-fire phasing page on this site!

The following system is useful for electrically rotating the array:
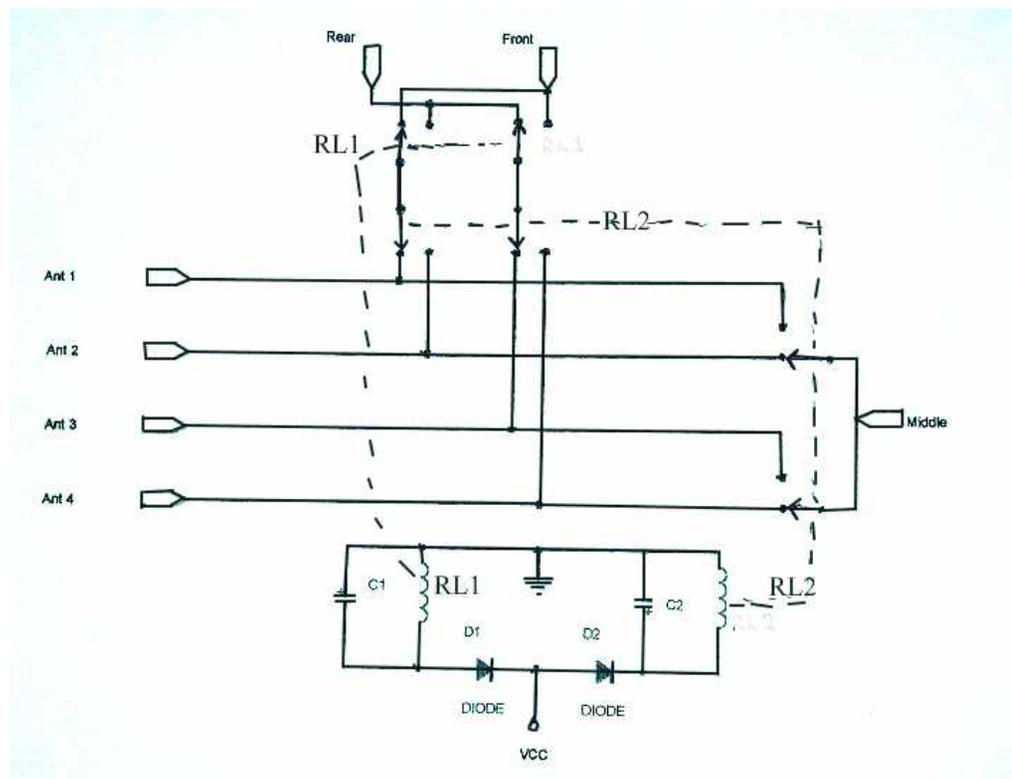
The antenna ports on the left from top to bottom start at the front of the antenna and from a top view go clockwise around the antenna. Zero voltage is position one at zero degrees, + is 90 degrees CW, - is 180 degrees, and AC is 270 degrees.

RL2 can be two separate DPDT relays with coils in parallel. I use 12V DIP relays with 220 uF 25v capacitors across relay coils. Any common silicon power rectifier will work for the diodes. If you have high RF levels near the control line, bypass it with a .1uF disc 50v capacitor. I route control signals through the feedline, but you must not have water in the feedline or poor shield connections to do that!

If you can not find parts, I have parts. I'm in the process of ordering commercial PC boards right now, to supply turn-key components kits or systems. I'd be happy to help with anything you need, as these projects move along.

W8JI_RX_ANTS

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W8JI's Receiving Antennas

Conventional Beverage's ~580 feet long. Currently only two left, all others removed over the past two years due to lack of performance.

Specifications @ 15 degree wave angle:

Every 45 degrees from NE to NW. Directivity factor 7.29 dB, Half-power Beamwidth 78 degrees
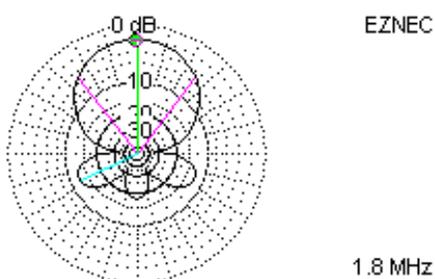
Arrays of Broadside Beverages 780 x 350 feet. Currently two duplicate systems (spaced ~1500 feet apart with antennas NE, E, SE, S, SW, W, and NW (28 individual Beverages make up this system):

Specifications @ 15 degree wave angle:

Europe, Africa, Asia/Japan, South America, ZL/South Pacific, VK/Pacific.

Directivity factor 13.8 dB, Half-power Beamwidth 42 degrees
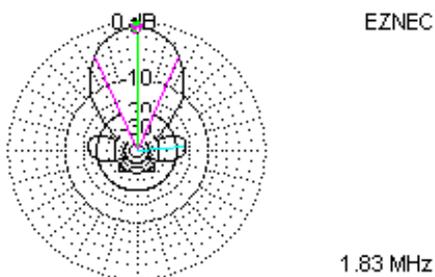
Arrays of Verticals 70 x 330 feet. One system:

Specifications @ 15 degree wave angle:

ZL/South Pacific, Europe.

Directivity factor 13 dB, Half-power Beamwidth 47 degrees

EZNEC

1.83 MHz
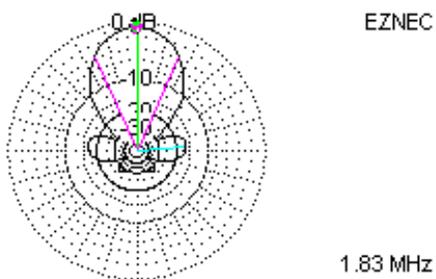
350-foot diameter circular eight-element array. One system:

This array covers eight directions, spaced every 45 degrees around the compass. It has 22-foot tall top and base loaded elements. Over 1/4 mile of 75-ohm cable is used in this array (and 2500 feet of cable to reach the house).

Each antenna element is made from 1-1/2 inch chain link fence top rail. The fence rail hinges on these wooden posts with a bolt through the bottom of the vertical rail. The top of the wooden post has a steel pin that fits in a hole in the metal railing, and it is lashed in place.

Specifications @ 15 degree wave angle:

All compass directions. Electrically rotatable.

Directivity factor 12.1 dB, Half-power Beamwidth 53 degrees
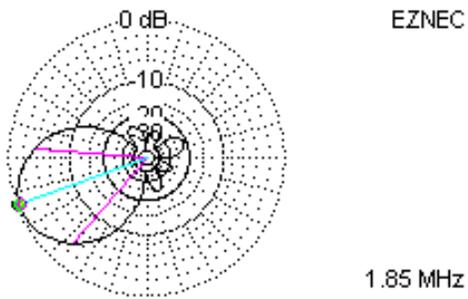
RADIATED and CONDUCTED NOISE

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Switching mode power supplies, light dimmers, computer networking systems, poor connections that arc, and other "accidental transmitters" that either switch or spark can create considerable RF energy on wiring. It is helpful, when attempting to reduce such noise, that we understand how the noise travels from the source into the receiving system.

Definition of a Source

Let's consider the problematic device a "thing in a black-box", and call it the "*source"*. It doesn't matter if it is a computer, the spark in an electric fence, a light dimmer, or a doorbell transformer. The smallest area we can isolate creating the unwanted energy will be the *source*.

Radiation from a Source

There is very little radiation from a source. As a matter of fact, even a very powerful single-terminal source can't transmit at any distance. In order to be an effective radiator, the source must have two or more connections to the outside world. The multiple connections (two or more) could be the two (or three) wires in a power cord, or a insertion point in a single wire. It could also be a connection to two totally different sources. Let's look at examples of each.

Arcing Splice

Consider a poor slice in a conductor carrying more than a few volts can cause a RFI generating arc. Two of the most common examples are an electric fence or power line.

The noise results when an insulated area with an electrical open circuit at low voltages breaks down as sufficient voltage appears across the gap. This ionizes air in the gap, and the resulting plasma conducts heavily. Such arcs can be particularly noisy at radio frequencies, because resistance can fall faster than voltage increases creating a negative-resistance. Negative resistances created in spark plasma were actually used as the amplifying element in crude oscillators. The plasma excited resonant circuits in the very first "single-frequency" transmitters, and at times those transmitters spanned distances of thousands of miles to very poor receivers using what were very poor antennas by today's standards!

Each conductor direction leaving the arc forms half of a large antenna. The entire system, from the source out in both directions, radiates. There isn't any effective way to cure or substantially reduce the unwanted radiation except by stopping the arc. Poor slices can become quieter in heavy rain, and can be broken up by wind or wire movement.

This type of arc excites the conductor entering the splice from one direction as one terminal, and the wire exiting the splice as the other terminal! The coupling mechanism is like a dipole feedpoint, and is very efficient at radiating over very wide frequency ranges.

Arcing to Another Conductor

A poor insulator or a close-spaced gap can break down and generate noise.

Once again, the noise results when an insulated area with an electrical open circuit at lower voltages breaks down when sufficient voltage appears across the gap. This ionizes air in the gap, and the resulting plasma conducts heavily. Such arcs can be particularly noisy at higher radio frequencies, because resistance can fall faster than voltage increases creating a negative-resistance.

Shunting-arcs generally have poor low-frequency energy, because the gaps are almost always connected to a very high impedance at low frequencies. This makes a very poor antenna connection for the arc, and reduces low-frequency response. If the gap had a path to a low resistance, the plasma would form a dead-short in the path. This would quickly cause a catastrophic system failure in systems able to supply sufficient current.

Non-arcing RF Generators

Many devices contain high-speed switching systems. These devices include but are not limited to televisions and monitors, computers, lighting systems with dimmers, and low-voltage lighting systems using switch-mode power supplies.

These non-arcing sources are similar to arcing sources, in that they require at least two terminal paths to become effective radiators. They generally are frequency selective or frequency periodic, and produce a broad buzzing signal that drifts around.

Common Characteristics

No matter what source is at work, we can be sure more than one conductor is at work in coupling from the source. If we disturb that path by isolating one conductor with a high RF impedance, we can reduce the interference. But by far the most effective way to remove or reduce interference, other than by removing the actual offending source, is by "shorting" the RF path with a bypass capacitor. As we will see later, this is almost always much more effective than adding series

impedance.

Common Noise Sources

Arcing Powerline Insulator or Hardware

The bell-insulator in a power line is probably the most common noise source. Older bell insulators has two interlocking but well-insulated metal posts. These metal posts have considerable area, because of the required mechanical strength. This results in considerable capacitance, in the order of a few dozen picofarads or greater. The ends of each bell terminal are held by a loose-fitting pin assembly. This pin assembly is the root of most powerline noise problems.

Bell insulators are often misused by power line crews. They are sometimes used to terminate short spans between poles, where the correct insulator would be a rigid post type or a polymer insulated fiberglass rod. The bell insulator, used in a short span, often is installed without sufficient tension. The lack of tension allows corrosion to set-up on the loose-fitting pins, and the capacitance of the insulator creates a voltage divider. Since the pin capacitive reactance is very high compared to the insulators leakage resistance and capacitance, a very large voltage can appear across the pin. This is true even when the pin is connected to grounded hardware! These loose short spans are called "slack spans", and bells should NEVER be used on slack spans! Slack spans with bell insulators generally become quieter during damp weather or in or just after rain, and can often be broken up by wire movement.

As the sine-wave peaks, the corrosion or oxides in the joint of the loose-fitting pin break down and arc. This causes an unintentional electrical noise to be generated, and the two terminals for the transmitter become the power line conductor and the grounded pole or hardware on the pole. With a wooden pole, the entire upper area of the can be excited with leakage currents from leakage capacitance and leakage resistances. Even if an insulator isn't arcing, a loose metal staple over a grounded wire or other poorly connected metal-to-metal joints will arc. Even a very tiny arc can excite the ground wire running down the pole and the power line wires as two (or more) terminals for the source.

To a much lesser extent transformers, lightning arrestors, fuse assemblies, and disconnects can have internal arcs.

Trees and foliage may also contact HV power lines and "burn" against the wire. Generally foliage noises are not extremely loud sources at low frequencies because one terminal of the source is through the tree, but occasionally they can be strong at lower frequencies.

As a general rule, but not always:

- Series-connection arcs from loose or corroded metallic hardware are reduced by wet weather. Water, contaminated by the corroded metals, forms a path around the poor connection and stops or reduces the arc.
- Shunt connection arcs from poor insulation are often reduced and may disappear in dry weather. They are caused by dielectric (insulation) failures.
- Corona noises are stronger at very high radio frequencies
- Hardware arcs and bells are generally strongest at higher radio frequencies
- Hard arcs like splices, clamps, transformer insulation, and switches are generally worse at lower radio frequencies
- The arc can be a very long distance from the point where you hear it on LOWER radio frequencies

Electric Fences

Older electric fences had a few-second timed off-and-on buzz. Newer fences have a "popping" noise, because they charge the fence in "tics". Fence problems are similar to power line problems, except of course the hardware is much smaller! Most severe problems are related to poor insulators, bad splices, or loose hardware.

Computers, TV's, and Switching Systems

In older times, TV sets created problems from sweep circuits. The horizontal sweep circuit operates with a saw tooth shaped waveform that was rich in harmonics. Early video systems overpowered the horizontal sweep system, and used the extra energy to supply high tension for the CRT. This resulted in many point-to-point wires carrying high power harmonic-rich 15kHz RF energy. A typical 25-inch television would use about 15 watts of the horizontal sweep system power for sweep, and perhaps 30 watts for powering the CRT second anode!

Wasted energy from the 50-75 watt sweep system was often coupled to the outside world a push-pull signal between the antenna leads and the power cord.

Modern computers and TV's mostly create noise-problems from internal switching power supplies, rather than sweep systems. The sweep systems, if ever even required, now generally only power the sweep. The HV often comes from separate HV power supplies.

Switch mode power supplies, whether in a TV, computer, telephone, video, or lighting system are rapidly becoming the most common source of modern RFI problems. Most of this has to do with poor testing techniques, lack of good standards, poor enforcement of the few poor standards we have, and a lack of skill or knowledge in power supply and device design and installation.

Fortunately most problems can be corrected by external filtering, but it generally needs to be done at the offending device.

Curing RF Egress or Ingress

Once the outside world is reached, the push-pull nature of the source can appear as a common-mode signal following a group of wires for many miles. We can think of this as one very long wire with the earth as the return path. The signal either directly radiates into our antennas, or it excites our antennas via common mode paths along cable shields.

While power lines and high-tension electric fences must be cured by removing the arc at the source, lower voltage systems can almost always be cured through proper external bypassing.

As a first response, we often like to throw a few ferrite beads at the problem. In more sophisticated approach, we might use multiple-turn series inductors. What these approaches miss is the great difficulty in obtaining adequate series impedance. When we consider the series impedance has to be totally *ISOLATED* for each conductor, and further consider core saturation from low-frequency operating currents we quickly realize series impedance is ***NOT*** the most effective method. It is helpful and may reduce RFI to acceptable levels in mild cases, but it certainly is an ineffective cure. Even if we are successful at one frequency range, it becomes virtually impossible to have a high series impedance over very wide frequency ranges.

The best approach is to add some RF series impedance and to augment it with excellent bypassing of all leads entering or leaving a device to one common point. That common point generally is connected to a safety ground, available in the USA at the third ground terminal of every modern-code electrical outlet.

Coaxial or audio cables should all leave the device through a bulkhead plate with all shields grounded. The power cord should pass through that bulkhead, and be bypassed to the bulkhead with properly selected UL/CSA/VDE approved line-bypass capacitors. Any chokes should be installed between the source device and the bulkhead. When this happens, the device becomes an isolated entity all by itself. It can not excite outside world conductors with unwanted RF.

It is indeed fortunate that the very same things that cure RFI often lead to greatly improved lightning protection.

Pre-amplifiers

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For the summary page click here [pre amp_summary](#)

The following is a comparison of HF low noise receiving preamplifiers. More amplifiers will be added as they become available to test. All instruments used for these measurements are currently calibrated newer-production laboratory grade instruments.

Detailed graphs will only be added if specifically required. The .jpg pictures on this page are examples of how measurements are displayed on the instruments used in tests.

Gain Compression Tests

Pre-amplifier gain compression is measured on an HP (Agilent) Vector Network Analyzer and S-parameter test set using the power level sweep function at 2 MHz. A gain compression test shows loss of gain as input power is increased, or the input level where the amplifier starts to saturate.

This test indicates how much overall signal power the amplifier will handle before gain is reduced by non-linearity or saturation. This is power or gain non-linearity, and really does not directly represent distortion level vs. input or output signal power. A correction is applied to results for amplifier gain and equipment protection attenuators.

Note: *Third-order intercept is a better demonstration of signal level capacity. Third-order intercept better indicates immunity of an amplifier to producing unwanted signals (mixing products) from multiple strong signals.*
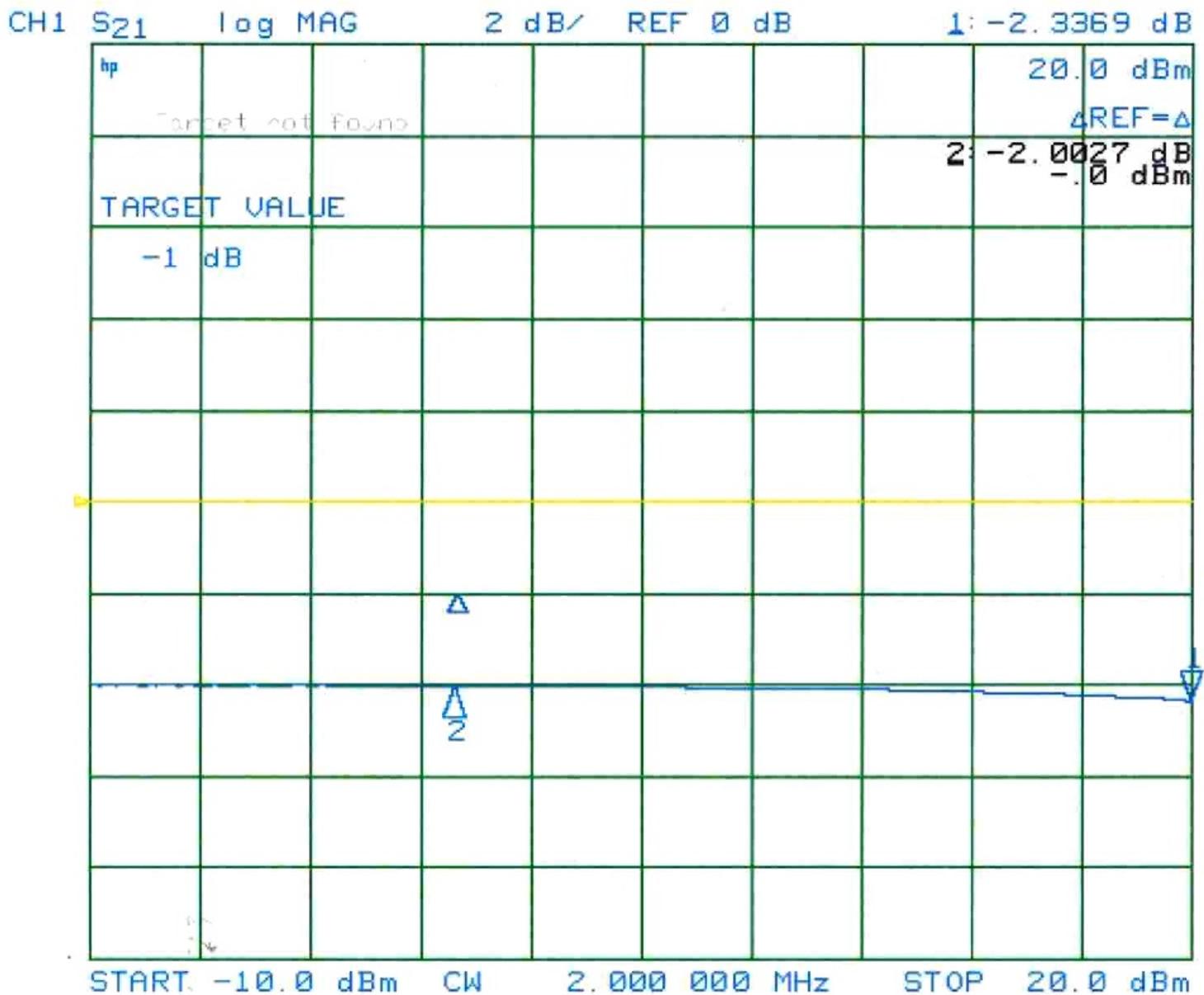
ARR 1-30
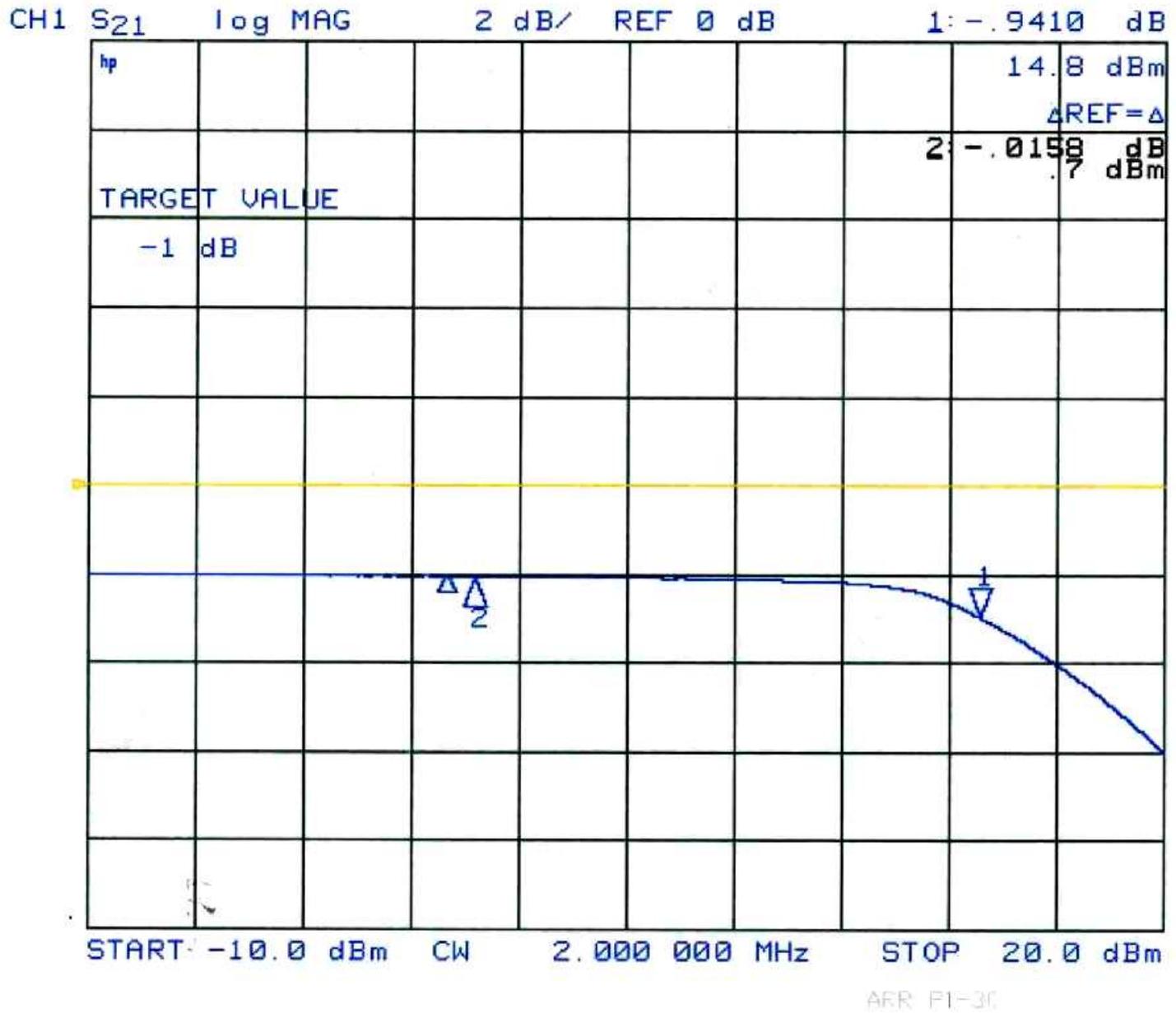
Amplifier input power (horizontal sweep) is 3dB per division. Measured 1dB compression is corrected to **21.5dBm output power @1dB compression**.

ARR GAsFET (1.8-2 MHz model)

CH1 S21   log MAG   2 dB∕   REF -.982 dB   1:-1.0629 dB

Marker 1=-2.6dBm

START -10.0 dBm   CW   2.000 000 MHz   STOP   20.0 dBm

3db/div Horiz

1dB compression is corrected to 8.1dBm output power.

DX Engineering Push-pull HF preamplifier

```
CH1  S21    log MAG      2 dB/  REF 0 dB          1:-2.3369 dB
    hp                                              20.0 dBm
        Target not found                          ⊿REF=⊿
                                                2:-2.0027 dB
    TARGET VALUE                                    -.0 dBm
      -1 dB




                            ⊿
                            ⊿
                            2



  START -10.0 dBm   CW    2.000 000 MHz    STOP  20.0 dBm
```

Gain compression beyond power output limit of analyzer. In other words, my test equipment does not have enough output power to drive the RPA-1 preamplifier into compression. Actual corrected gain compression (manual test) is approximately +26dBm (400-milliwatts output).

Comparison of Third Order Intercept

Third-order intercept tells us the immunity of the amplifier to generating spurious signals in the presence of multiple strong signals. This test was at 1.85MHz using the automatic IM3 test measurement. A stepped attenuator pad at the amplifier output was initially used to prevent analyzer damage during testing. Because of this "protection" attenuator, actual TOI must be increased by adding the attenuator value to the analyzer display reading.
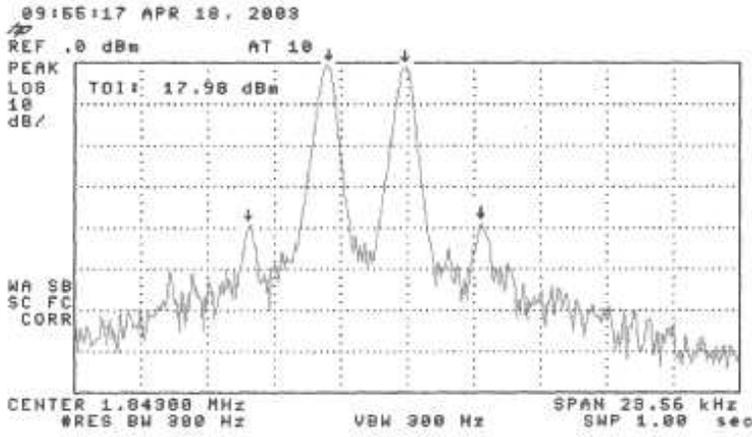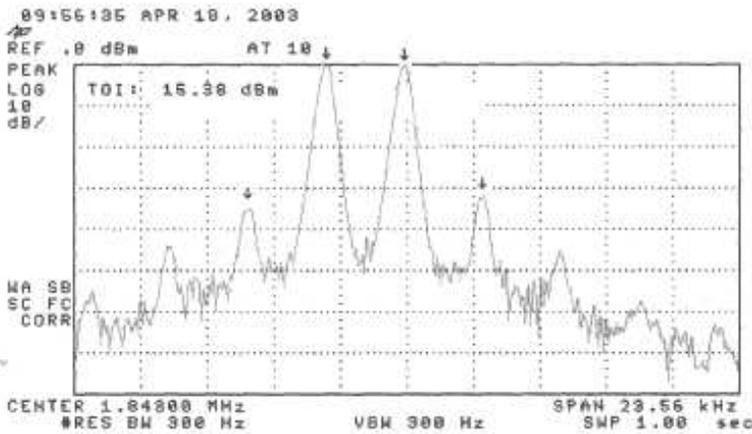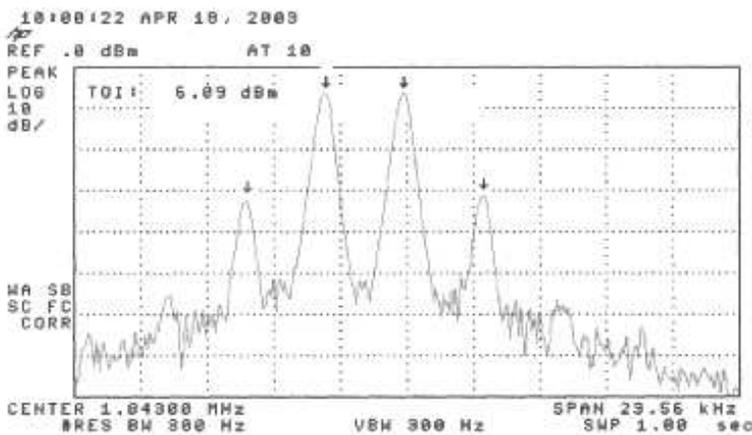
Third-order IM3 levels occur at the following single tone (of an equal two-tone test):

DXE (early model) -0dBm input power (instrument TOI 17.98+20.2dB pad= 38.18dB TOI)

Note: Later model RPA-1 measured 43dB third order intercept

ARR P1-30   -5dBm input power (instrument TOI 15.38dB+15dB pad= 30.38dB TOI)

ARR GAsFET -16dBm input power (instrument TOI 6.09dB+5.3dB pad= 11.39dB TOI)

If you look at the above measurements, you will see third-order intercept in the DX Engineering amplifier occurs almost 8dB higher than in the better performing of the two ARR amplifiers that were tested. Not only is the DXE 3rd order intercept higher, higher order products are substantially less than the other two amplifiers produced.

DXE-RPA-1 Test

This is a later model DXE preamplifier model RPA-1. This measurement has no external attenuator, TOI in this case is a direct reading.
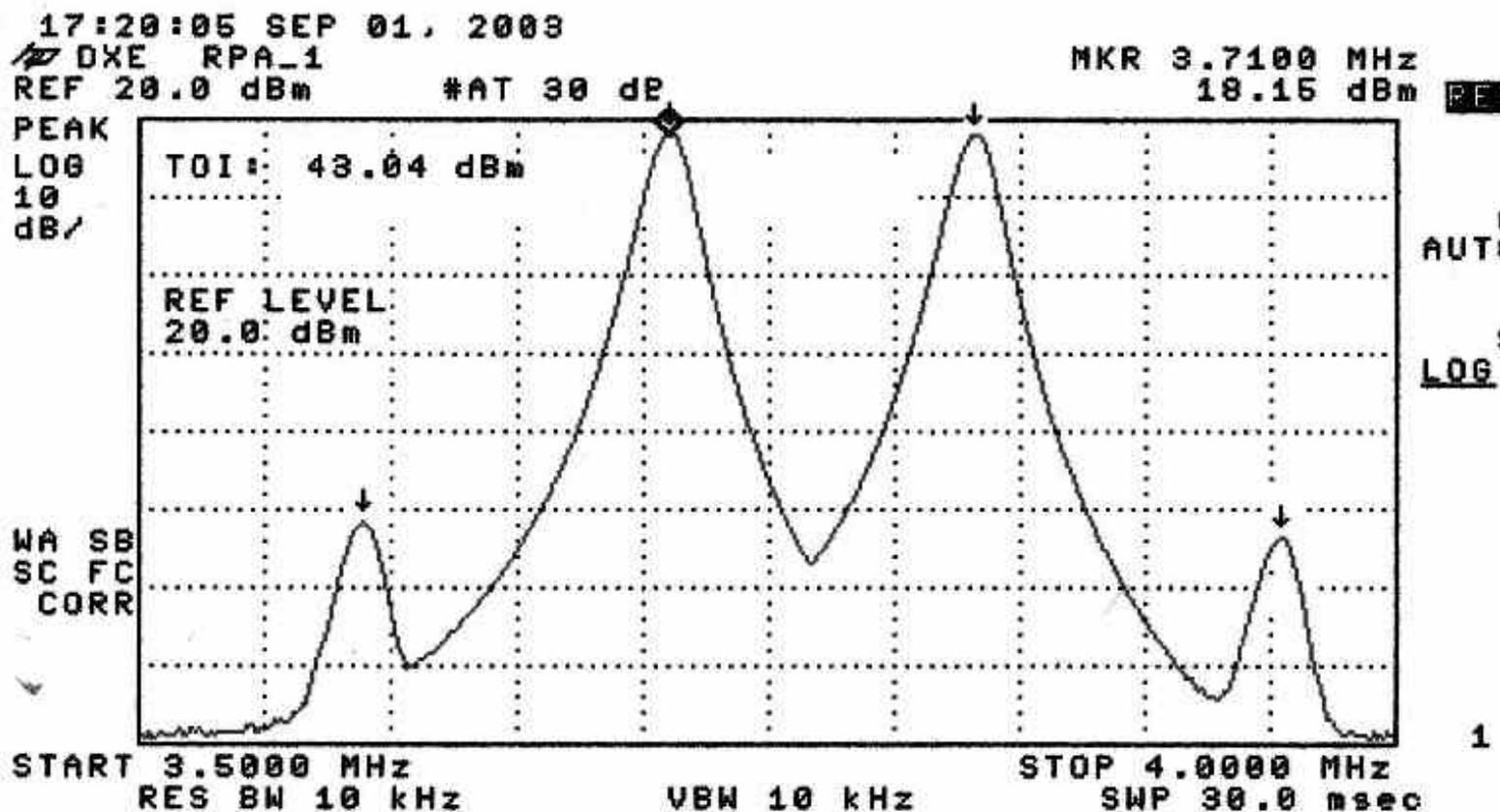

Preamp Noise Figure

 I don't have this data saved to a file yet, but the noise figure of the GAsFET amplifier and the DXE amplifier are about the same. The ARR 1-30 had the highest noise figure, about 5.5dB. All of these amplifiers are acceptable for HF receiving in most applications so far as noise figure is concerned.

Input Impedance and Gain

So far, the DXE pre-amplifier has the closest match to 50-ohms over the range of 300kHz to 30 MHz. Gain is also reasonably flat in the DXE amplifier.

DX Engineering Preamp
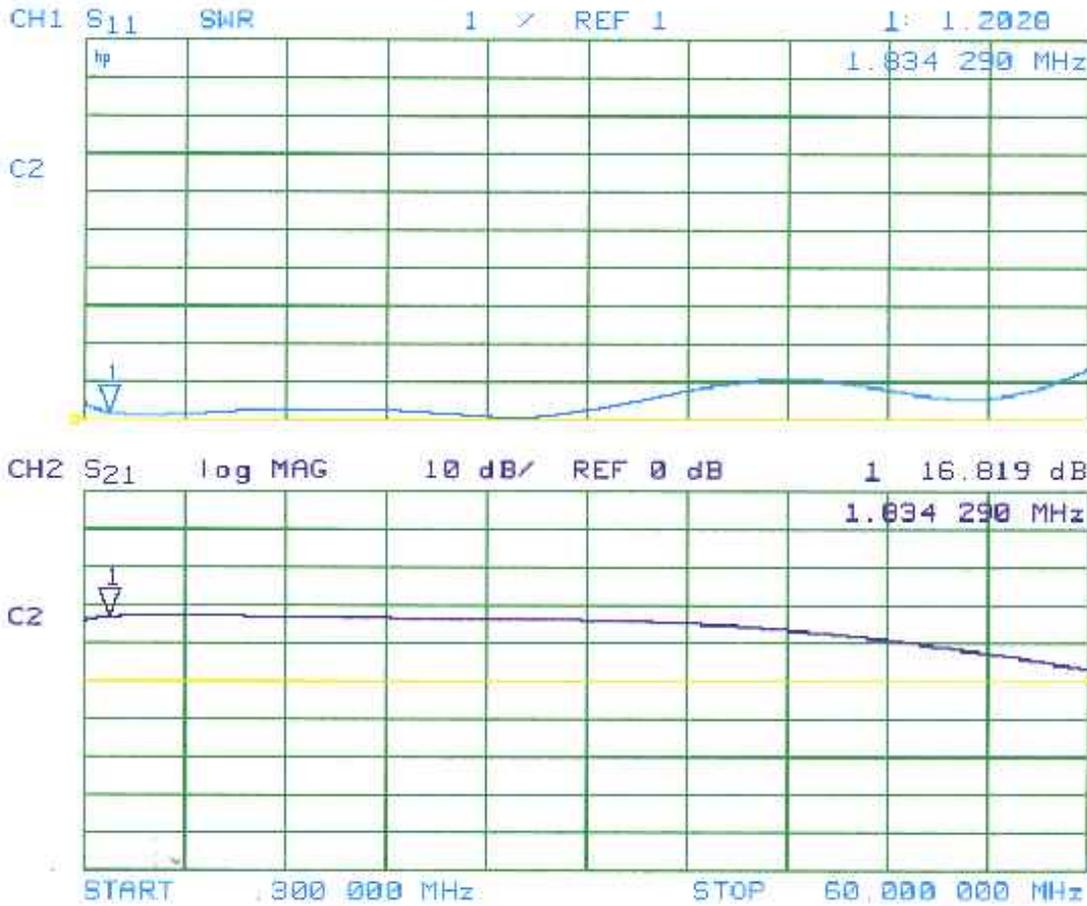
DXE preamplifier SWR (top)
and gain vs. frequency (bot)

The display above shows input SWR of the amplifier on the top graph, and gain on the lower graph. The marker is at 1.834290 MHz. The vertical graph lines are frequency at about 6 MHz per division. Gain at marker frequency is indicated as 16.819dB, SWR is 1.2028:1.

It is also possible to display input impedance on a Smith Chart:

CH1 S₁₁   1 U. FS   1: 62.984 Ω   −4.1387 Ω   31.849 nF
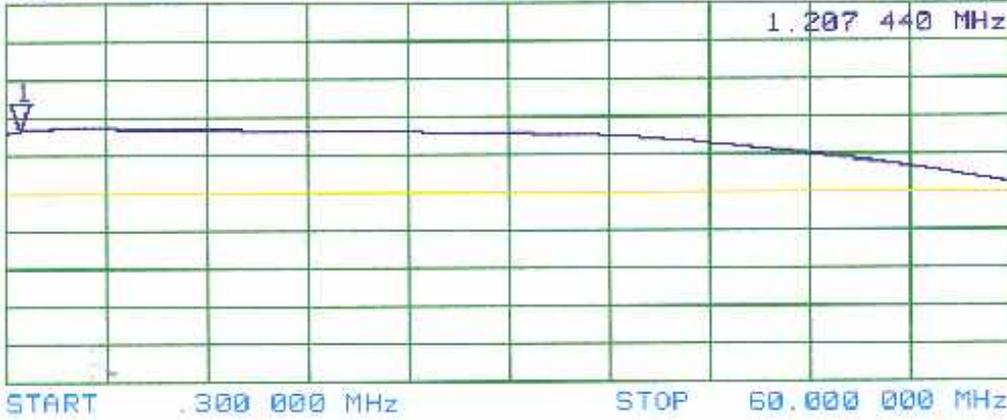 hp   1.207 440 MHz

C2?
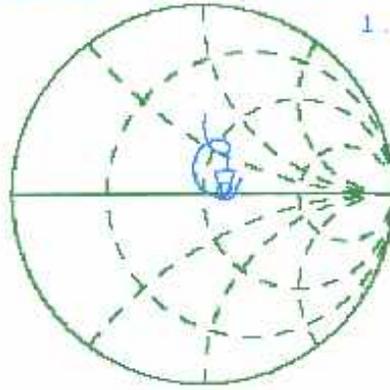
CH2 S₂₁   log MAG   10 dB/   REF 0 dB   1   16.524 dB
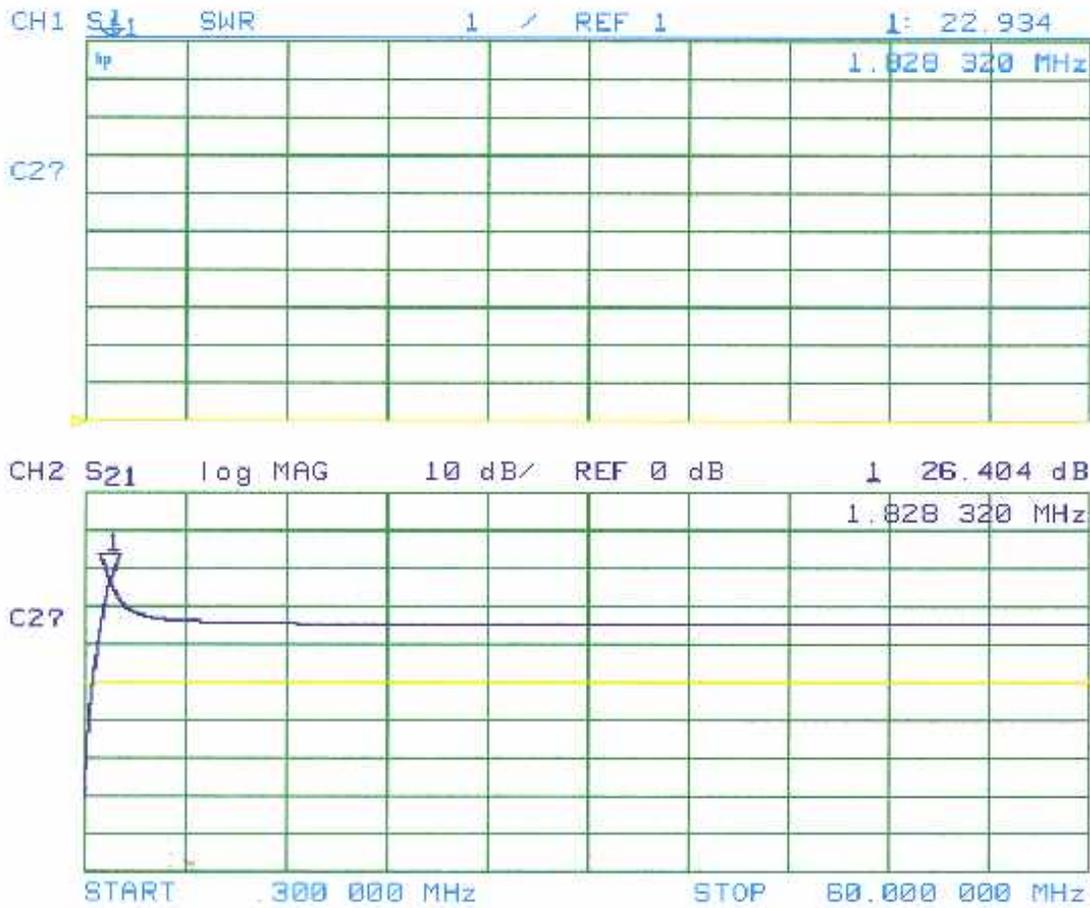   1.207 440 MHz

C2?

START   .300 000 MHz   STOP   60.000 000 MHz

DXE Smith Chart (top) Gain (bot)
marker at 1.210 MHz.

This is the gain and input impedance at 1.207440 MHz.

ARR GAsFET PreAmplifier

This amplifier is intended to be a 160-meter only amplifier. It has high gain despite of the very poor match on any frequency. It also has substantial gain on all frequencies swept, but a strong peak on 1.8 MHz.

CH1 S11   SWR   1 / REF 1   1: 22.934
                                                1.828 320 MHz
C27

CH2 S21   log MAG   10 dB/ REF 0 dB   1   26.404 dB
                                                1.828 320 MHz
C27

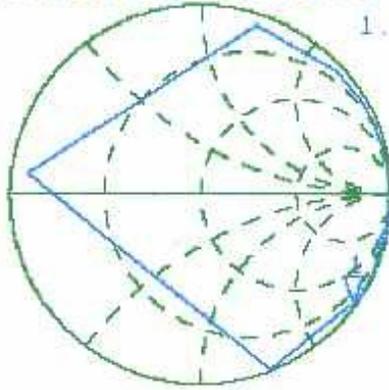START   .300 000 MHz   STOP   60.000 000 MHz

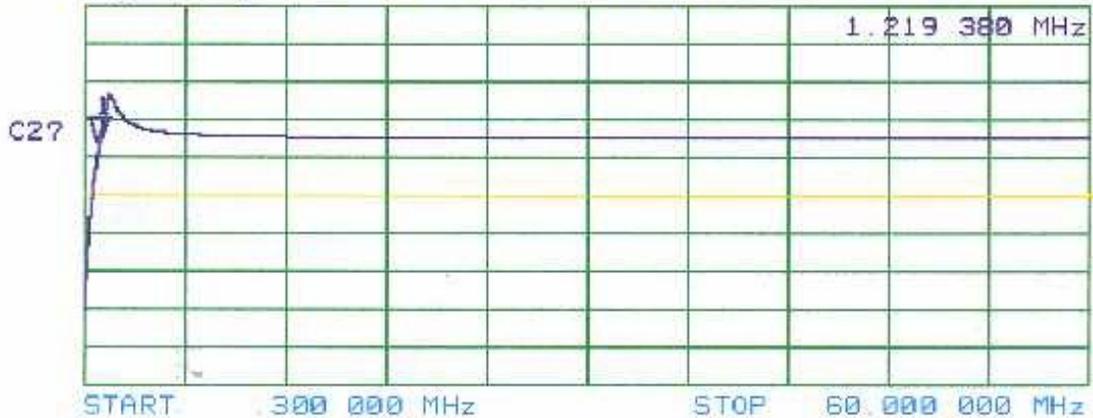ARR GAsFET SWR (top) Gain
(bot) marker at 1.828 MHz,

Input SWR is 23:1, gain is 26.4dB. Unfortunately noise figure is not that good for a GAsFET, probably because of the very large input

```
CH1 S11     1 U FS        1: 04.703 Ω    -157.13 Ω    830.68 pF
     by                                      1.219 380 MHz

C2?
```

```
CH2 S21    log MAG    10 dB/  REF 0 dB       1   13.006 dB
                                             1.219 380 MHz

C2?

START   .300 000 MHz              STOP   60.000 000 MHz
```

mismatch.

The above data is a Smith Chart display of the ARR GAsFET at 1.22 MHz. Gain is still 13dB, input impedance is 5 ohms and very reactive (this measurement is in the BCB). Despite being a "160-only" amplifier this amp has substantial gain (~15dB) at 60MHz.

Other Data

If you have a commercial amplifier or *commonly available* homebrew amplifier design you would like tested, please contact me. I'd be happy to measure your amplifier (time permitting) and post the results here.

See (soon to be added) summary page for text comparisons.

Amplifiers

[[Home](#)] [[Amplifier Power Parts](#)] [[Bandswitch Failures](#)] [[Demonstration](#)] [[Pre-amplifiers](#)]
[[Tank Circuit](#)] [[SB-221](#)] [[Vacuum Tubes](#)] [[Vacuum Tube Amplifiers](#)] [[VHF Stability](#)]
[[FL2100 Problems](#)] [[HF Stability](#)]

[SB-221 Tank Voltages and the parasitic myth](#)

[Tuning or loading an amplifier](#)

Preamplifiers

[Receiving amplifier measurements](#)

Vacuum Tubes

[Vacuum Tube Amplifiers](#) how the PA vacuum tube converts dc to RF

[Safe working model](#) for understanding how PA RF voltages are created and how amplifiers behave under various loading and drive conditions

[Vacuum Tubes](#), arcing, vacuum tube failures, advantages and disadvantages of each type of power-grid vacuum tube, vacuum tube grid failures, vacuum tube anode failures, maximizing vacuum tube life, and quite a bit about how power grid vacuum tubes actually function

Components

[Tank Circuits](#) two models showing tank voltages, currents, loss, and phase shift

[Bandswitch Failures](#) discusses the many common reasons for amplifier tank system switch failures

T/R relays and relay timing, including amplifier relay sequencing that will work with vacuum relays

Inductors while this deals mainly with loading coils, it is also true for amplifiers

Stability

HF Stability HF instability or instability near the operating frequency of an amplifier is by far the most damaging for of oscillation. This problem plagues amplifiers like the Collins 30L1, Dentron Clipperton L, and Yaesu FL2100.

VHF Stability Parasitics discusses causes of instability. Has spice models, pictures of resistors internal construction. Your replacement carbon composition resistor might not be a carbon composition after all!

Arcing Causes

Antenna and antenna system failures that cause power amplifier arcs

Poor or improper relay sequencing causing power amplifier arcs. This is a common problem in TL-922 Kenwood amplifiers

Improper Drive and Amplifier Tuning how incorrect drive and loading adjustments cause power amplifier arcs, especially with exciter power overshoot

Amplifier Tube Arcs causes and cures including real reasons why grid chokes fail. This is a common problem in the SB-220 SB-221, TL-922 and other amplifiers.

Future Articles

Choke selection will describe the proper selection and construction of RF chokes
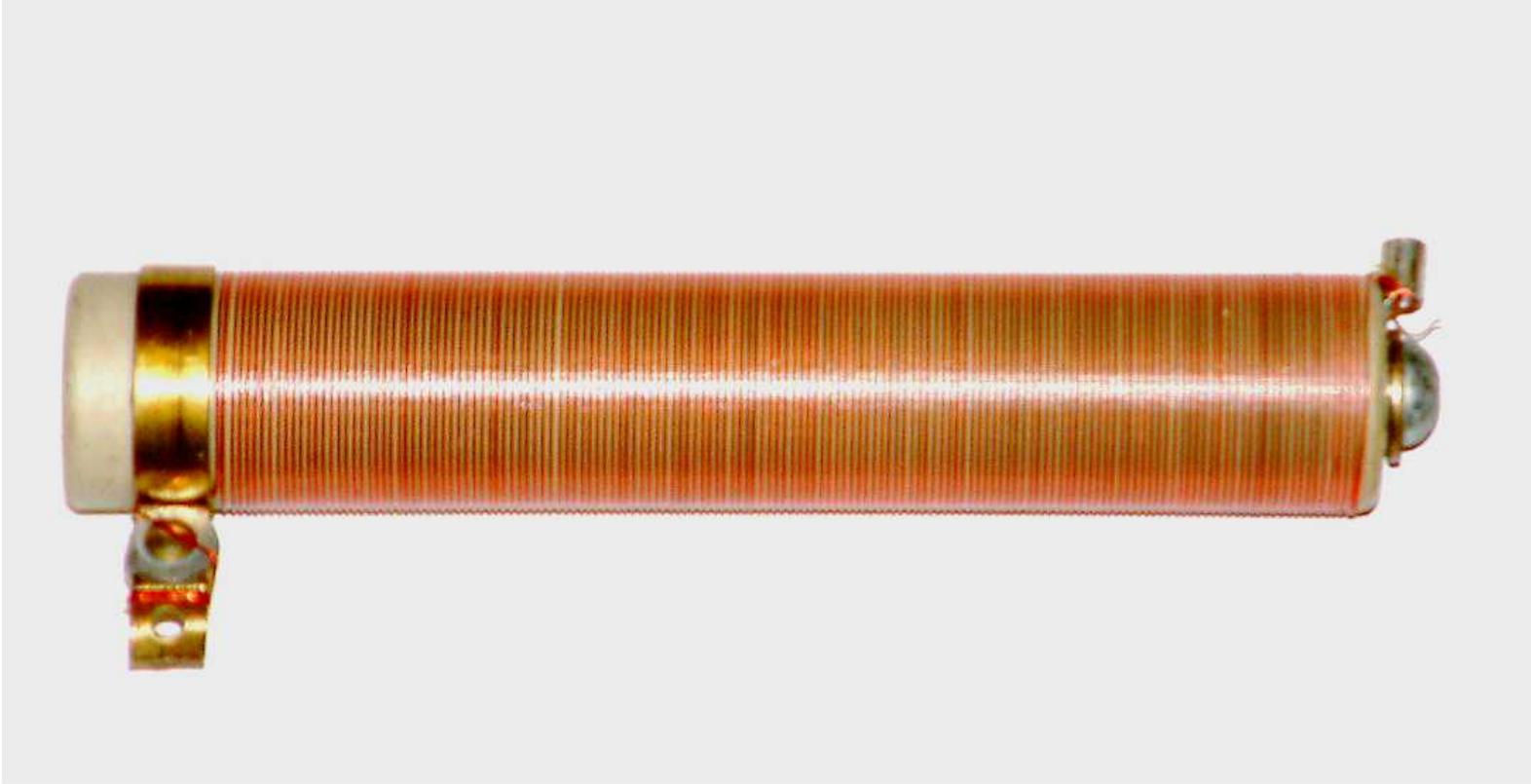
More pages soon......

visits since January 2004

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Amplifier Power Parts

RF Plate Choke

142uH @ 1.8 MHz. 3 amperes

Single layer space wound

Ceramic form

Hi-Q insulated

1" diameter by 5.5" long.

1/4" x 20 female thread both ends

Series resonances at 17, 25, and 32MHz. (Typically unusable at +-5% these frequencies)

Brand-new (similar Heathkit SB1000 and Ameritron)

$15.95 + 3.00 S&H in USA

Vacuum Tubes

I have a large stock of new vacuum tubes, including high power tubes.

Vacuum Tubes For Sale

[**Home**] [**Up**]

Transmitting Tubes

Various tube types from 100TH through 4CX15000 including hard to find 4CX1600B tubes.

List posted soon

Receiving Tubes

TUBE	QTY	BOX	TUBE	QTY	BOX	TUBE	QTY	BOX	TUBE	QTY	BOX	TUBES	QTY	BOX
35Z3	1	1	556	2	2	5687	1	3	1258	1	4	655	1	5
12SW7	1	1	1U5	1	2	6AF4A	11	3	10DX8	2	4	12A6	1	5
25L6	1	1	1X2C	1	2	6AG5	2	3	10GK6	2	4	12DQ6B	1	5
2A5	1	1	2AF4B	1	2	6AK5	1	3	10KR8	2	4	12SA7	1	5
50L6GT	1	1	2AV2	1	2	6AL5	1	3	11CY7	2	4	12SA7GT	2	5
5Y3WGTA	2	1	2CY5	3	2	6AQ5	3	3	11KV8	1	4	12SQ7	2	5
6JB6A	4	1	2FS5	1	2	6AS8	1	3	11LT8	1	4	1N5GT	1	5
6BG6A	1	1	2HA5	2	2	6AT6	1	3	12AB5	2	4	25L6GT	1	5
6BQ6GTB	2	1	3A2	1	2	6AT8A	2	3	12AD6	1	4	35Z5GT	2	5
6BX7	1	1	3AL5	4	2	6AW8	2	3	12AE7	2	4	50A5	1	5
6EW7	1	1	3AU6	3	2	6BC5	1	3	12AL5	2	4	50L6GT	2	5
6K6GT	1	1	3AV6	2	2	6BC7	1	3	12AT6	1	4	6BG6	1	5
6SG7Y	1	1	3BC5	2	2	6BC8	1	3	12AU6	4	4	6BY5GA	1	5
6SL7WGT	2	1	3BE6	1	2	6BE6	2	3	12AV6	31	4	6DQ5	1	5
6V6GTY	2	1	3BY6	1	2	6BF5	1	3	12AV7	3	4	6JE6A	1	5
6X5GT	1	1	3BZ6	1	2	6BH8	1	3	12AW6	1	4	6KN6	1	5
7C5	1	1	3CB6	2	2	6BJ8	2	3	12BA6	6	4	6KN6	1	5
6N060T	1	1	3CS6	2	2	6BK5	3	3	12BF6	2	4	6S57	1	5
1613 (JAN Number)	2	1	3CY5	1	2	6BR5	1	3	12BH7	1	4	6SA7	1	5

			Name	Qty		Name	Qty		Name	Qty		Name	Qty	
			3DT6A	2	2	6BS8	2	3	12DT5	2	4	6SA7	1	5
			3EH7	1	2	6BW8	1	3	12F8	1	4	6SC7	1	5
			3GK5	1	2	6BY8	3	3	12FX5	1	4	6SF5	2	5
			3HA5	1	2	6BZ7	2	3	12HL7	1	4	6SK7	1	5
			3V4	3	2	6C4	3	3	15HB6	1	4	6SK7	3	5
			4BQ7A	2	2	6CM7	1	3	17C9	4	4	6SN6	1	5
			4GK5	3	2	6CQ8	1	3	18FW6A	1	4	6SN7	1	5
			4HS8	1	2	6CS6	1	3	18FX6A	1	4	6SQ7	3	5
			4LJ8	1	2	6CS7	1	3	18FY6A	2	4	6V6GT	7	5
			5AM8	1	2	6CS7	1	3	19J6	4	4	6X5GT	1	5
			5AN8	2	2	6CW5	1	3	19T8	2	4	O24	2	5
			5AQ5	1	2	6CX8	1	3	20EZ7	1	4			
			5AT8	3	2	6DE6	1	3	25BK5	1	4			
			5BK7	1	2	6DE7	1	3	35B5	1	4			
			5BR8	3	2	6DK6	1	3	35W4	5	4			
			5BT8	1	2	6DT6A	2	3	50C5	1	4			
			5CG8	4	2	6EB8	1	3	50DC4	1	4			
			5CL8A	2	2	6EH5	1	3	50EH	5	4			
			5FV8	2	2	6EH7	1	3	684A	1	4			
			5KE8	1	2	6EH8	1	3	6JCA	4	4			
			5T8	1	2	6ES5	1	3	6JU8A	1	4			
			5U8	1	2	6ES8	1	3	6LC8	3	4			
						6EU8	3	3	6LE8	3	4			
						6FG7	2	3	6LN8	2	4			
						6FQ7	4	3	6LX8	3	4			
						6FV6	2	3	6T8	1	4			
						6GK5	1	3	6X4	1	4			
						6GK6	1	3	6X9	3	4			
						6GN8	1	3	6Y9	1	4			
						6GW8	1	3	8BQ5	2	4			

6HZ6	1	3	8KA8	2	4
6JC6A	2	3			

TUBES	QTY	BOX	TUBES	QTY	BOX	TUBES	QTY	BOX	TUBES	QTY	BOX
12AU7	1	6	32	1	7	80	1	8	0B2	1	9
12AU7 used	31	6	43	1	7	3223	1	8	10DE7	3	9
12BA6	8	6	45	1	7	5687	1	8	12AT7	5	9
12BF6	1	6	2526	1	7	5692	1	8	12SQ7	1	9
12X4	1	6	5744	2	7	5842	1	8	12V6	1	9
1K5	1	6	5933	1	7	6080	1	8	25C6	1	9
1S5	1	6	6087	1	7	6100	1	8	3Q4	1	9
1T4	1	6	117Z6	1	7	6136	1	8	50C5	1	9
1U4	1	6	12AU7	1	7	6201	2	8	5Z4	1	9
25BK5	1	6	12AV5GA	1	7	7061	1	8	6AK5	2	9
3Q4	1	6	12AX7	1	7	7061	1	8	6AK6	1	9
6AF4	1	6	12SH7	2	7	7586	2	8	6AQ5	3	9
6AG5	1	6	12SH7	1	7	7868	2	8	6AS6	1	9
6AH6	1	6	12SH7	1	7	10-4C BALLAST	1	8	6AU6	4	9
6AK6	1	6	14A7	1	7	12A56	1	8	6AV5	1	9
6AL5	2	6	14AF7	1	7	12AU7	8	8	6BA6	3	9
6AM8A	1	6	17D4	1	7	12AV6	1	8	6BE6	3	9
6AN5WA	1	6	25CD6GB	1	7	12B4A	2	8	6BH6	1	9
6AU6	2	6	3525GT	1	7	12BA6	1	8	6BN6	1	9
6AV6	5	6	50L6GT	1	7	12BE6	2	8	6BY5	1	9
6BF6	1	6	5651A	1	7	12BH7	1	8	6BZ6	1	9

Type	Qty		Type	Qty		Type	Qty		Type	Qty	
6BN6	1	6	5V4	1	7	12DQ6	2	8	6CG8	5	9
6CG7	1	6	6080WA	1	7	12GN7	1	8	6CM7	5	9
6CM6	5	6	65U7	1	7	12SH7	2	8	6CM8	5	9
6EW6	10	6	6AS7	1	7	12SR7	1	8	6EJ7	1	9
6FG7	1	6	6AU6	2	7	12V6	1	8	6EU7	7	9
6GW6	5	6	6AU8A	1	7	17AX4GTA	1	8	6EU7	5	9
6J6	1	6	6BG6GA	1	7	17EW8	1	8	6J6A	2	9
6SU8	5	6	6BH6	1	7	1V2	1	8	6K6	1	9
			6BJ6	1	7	25BQ6	1	8	6LW6	1	9
			6CA7	4	7	2D21	1	8	6Q7	1	9
			6CB5A	2	7	35L6	1	8	6S57	1	9
			6CD6	1	7	35W4	1	8	6SN7	1	9
			6SK7	1	7	35W4	1	8	6T8	1	9
			6V6	1	7	3CY5	1	8	6V6	2	9
			7F7	3	7	3DF3	1	8			
			DELAY RELAY 26 V 300 SEC		7	50C5	1	8			
			DELAY RELAY E-1215		7	5749W	1	8			
			DELAY RELAY X-35		7	5U4	1	8			
			117 V DELAY RELAY 300 SEC	1	1	5Y3	2	8			
						5Z4	1	8			
						6140/423A	2	8			
						6AG5	1	8			
						6AG5	1	8			
						6AL5	4	8			

			6AN8A	2	8			
			6AQ5	1	8			
			6AQ5	2	8			
			6AQ6	1	8			
			6AV6	1	8			
			6AW8	1	8			

TUBES	QTY	BOX
1AD2	3	10
1AY2A	3	10
2AH2	2	10
2AS2	4	10
3A3C	2	10
3BW2	1	10
3DB3	2	10
3DG4	1	10
5DJ4	2	10

2747

Bandswitch Failures

Bandswitch Failures

Bandswitch failures are caused by:

1.) Improper installation or wiring of switch

    a.) Improper Wiring Techniques

    b.) Improper contact Alignment

    c.) Contamination-of or damage-to switch

2.) Improper relay timing or defective antenna relay

    a.) Transmit contact closes *after* input contact closes

    b.) Transmit contact opens *before* input contact opens

3.) Failures in antenna systems

    a.) Intermittent arc, open, or short in a tuner

    b.) Intermittent arc, open, or short in an antenna

c.) Intermittent arc, open, or short in external lightning protection devices

d.) Intermittent arc, open, or short in a feedline or connector

e.) Sudden large change in either SWR or antenna impedance

4.) [Improper loading of amplifier](#)

a.) Loading control set too far meshed (loading too light)

b.) SWR or antenna impedance beyond matching range of amplifier

5.) [Excessive drive or drive power transients](#)

a.) Exciter has transients or leading edge overshoot

b.) Operation power peaks higher than tuning peaks

6.) [Excessive current](#)

a.) Improper loading (loading too light)

b.) Improper connections

c.) Improper switch ratings

Failures_in_antenna_systems

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Any of the problems below can result in destruction of a bandswitch or other amplifier components.

Tuners

Antenna tuners are a common problem. Many tuners are over-advertised, and will not handle the power claimed. Some models also have poor or unreliable connections in inductors. Many tuners have poorly designed power meters with peak meters that don't indicate true peak power. If you look at the antenna tuner review in the February 2003 issue of QST, you will see power rating problems exist with many tuners.

User adjustments also greatly control tuner power ratings. People adjusting "T" network tuners often forget to use the maximum possible capacitance required to match the load. Using less than maximum capacitance will reduce tuner power handling capability.

A very brief arc or open circuit in a tuner can cause a momentary very large change in antenna tuner input impedance. This large impedance change reflects back into the amplifier, and if the tank is momentarily unloaded a catastrophic bandswitch or capacitor arc will occur.

The worse possible change is one that results in a condition of high impedance at the amplifier tank input, in which cases tank voltages can reach many times the normal operating values.

Always be sure your tuner has reliable internal connections, and is tuned correctly (maximum capacitance that allows matching the load). Be sure the tuner can handle the peak transient power from your exciter-amplifier combination.

At one point, I owned an ICOM 775DSP radio. The peak transient power from that radio, even when set to 50 watts, was over 300 watts! While that pulse would not show on a normal power meter, it clearly showed on peak storage meters. My Bird digital storage, a special Coaxial Dynamics peak meter for medical applications,   When I drove an 8877 amplifier with that radio, the few milliseconds of peak driving pulse would trip the very fast grid-current-fault protection system in my amplifier.

When I advanced the loading to prevent the grid trip from activating, the transient peak amplifier out was over 5kW. This would cause my antenna tuner, a large old KW Johnson Matchbox, to momentarily flash-over! When the tuner would arc, the amplifier tank would arc.

The entire system would have to be able to handle the exciter's 300-350 watt peak transient (amplified through the PA gain) in order to not have reliability problems!

Antenna and Feedline Failures

Antennas can have arcing problems from bad connections or other defects. If an antenna fails to accept power or changes impedance intermittently, even if for a few milliseconds, it can trigger an arc elsewhere in the system. That point may even be in an amplifier tank system.

Traps can arc, as can other components like gamma capacitors or matching networks. Be sure antenna wires are not able to brush up against other objects in the wind.

Be sure all connectors and hardware are tight, and connections are dry and well-insulated. Normal UHF connectors, even though rated less, handle much more power than type N connectors. UHF connectors are much less likely to develop intermittent connections, and actually have LESS loss below 50MHz. The only potential drawback for HF use is UHF connectors are less water resistant than properly installed N connectors, but that can be cured with proper waterproofing.

Antenna problems will show if you use a peak-storage meter, or a meter with a long hang-time. The only amateur radio products I'm aware of that have enough response time and hang time are the Ameritron meters used in the AWM-30 meter and ATR-30 and 20 tuners. Any momentary faults in my antennas immediately show on the reflected power of those meters, when they are in the peak mode.

Loading_amplifier

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Improper Loading of Amplifier

As far as tank voltages are concerned, there really isn't any difference between excessive drive, load faults, or improper adjustment of loading. Improper tank adjustment and excessive drive are equally harmful to component life, and either can create splatter (and in extreme cases cause clicks) on adjacent frequencies.

Most Common Tuning Error

The most common tuning error is adjusting the amplifier at low power, rather than tuning at maximum obtainable drive.

The last few steps in tuning should be:

- Load the amplifier to maximum obtainable output at full exciter drive
- After that, advance the loading control slightly beyond (towards less capacitance).

ALWAYS load your amplifier for maximum obtainable power, and reduce drive to rated or desired power levels! This ensures minimum voltage and current in the tank, and maximum possible linearity for the best signal quality. High grid current is a good indicator of excessively light loading in grounded grid amplifiers.

Exciter Transients or Power Overshoot

 Maximum available carrier drive might not result in sufficient drive when an exciter has transients or power overshoot. Transients or overshoot only appears when going from zero power towards full operating power levels, a condition met at the leading edge of speech or CW transmissions. Once the ALC comes up, transients and/or overshoot will generally disappear. With transients or overshoot, it is impossible to tune your amplifier properly with either a carrier, tuning-pulser, or speech waveform. We can not just tune for maximum output and expect the amplifier to be properly loaded when the exciter has transients!

Loading_amplifier

Transients and overshoot, being of short duration and infrequently occurring, make it impossible to tune correctly at maximum drive.

Let's assume your exciter is rated to deliver 100 watts, but has momentary peaks or transients of 160 watts while the ALC or power control loop "takes hold". Power surges of 160 watts, too short to register on normal power meters, occur at the start of ever transmission. Of course, if you don't run the exciter wide open, and reduce power to 50 watts, this problem is rally aggravated!

In this case, the maximum power setting of the exciter should be used while tuning. If you suspect or know the exciter has transient problems, the loading control should be advanced a reasonable amount beyond (further open) the maximum output power setting. This will allow the amplifier tank system to handle transients without arcing or component failure.

Easy-Drive Linear Amplifiers

Some people and manufacturers tout low drive as an advantage, claiming it offers "cleaner signals". Nothing is further from the truth.

Exciters almost always provide the best IM performance when operated at an average peak power a reasonable amount below full output, rather than very low levels. At low levels, exciter performance is dominated by cross-over distortion caused by low-level operation of stages optimized for higher power levels. ALC also bias back early stages, often increasing distortion in those stages. At very high levels, gain compression or bias shift becomes an issue. Exciters typically do best when operated in the area of 60-80% of rated power.

Worse yet, low drive amplifiers are especially susceptible to damage from exciter overshoot or transient problems. Transients and overshoot remains almost the same level regardless of exciter power control settings. As exciter power levels are reduced, the percent of power overshoot becomes worse.

The most undesirable situation is one where exciter power greatly exceeds an amplifier's normal drive power limit. Not only does this reduce system IM performance, amplifier drive transients are aggravated. Amplifiers should be designed or selected to match the exciter's maximum power output, or an external attenuator used to bring the amplifier's drive requirement up to the exciter's full power level.

Demonstration

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A Practical Demonstration

This test circuit is easy to duplicate, and demonstrates the extreme voltages that appear in an amplifier when the load is improper for the amount of drive power applied.

Years ago, almost every amplifier had a output transformer. The transformer was similar to tanks circuits in RF PA's, in that it transformed or matched impedances. Even though Q was very low, the transformer could still store and release energy.

Experienced people knew if the volume was turned up too high in a transformer-coupled output stage, and if the load was absent, voltage across the transformer would soar to many times the normal operating voltage. No load or grossly mismatched loads often resulted in damaged transformers, blown output devices, or other output circuit component failures.

Failures induced by load or matching faults would occur in conservative amplifier designs, where components would last years in continuous *proper* operation. An audiophile would NEVER think of operating his expensive tube-type amplifier at anywhere near full volume into an open load, let alone a load where grossly improper impedances are selected in amplifier output-transformer taps.

Blame was never placed on stability. Everyone knew and understood conservatively designed well-constructed amplifiers with energy storage systems of any type, even very low Q systems, would still produce over-voltage failures when grossly overdriven or misterminated.

TV manufacturers used this well-known effect to advantage in horizontal output sections, where a flyback transformer with moderately low Q would produce many times the actual turns ratio in peak voltage, because of energy storage. Even modern switching supplies and our automobiles depend on energy storage to produce entirely new voltages, far above supply voltage, without requiring parasitics or high Q.

RF Systems

RF systems are certainly no more immune to mismatch than audio amplifiers, they very often are much worse. RF circuits are generally single-ended, and tube-type amplifiers have moderately high-Q (efficient) energy storage tank systems. Single-ended amplifiers with conduction angles under 360-degrees almost always contain intentionally designed "fly-back" systems, where tank circuit Q re-creates the missing portion of a sine wave from the half-cycle (or less) tug of the single-ended output device.

Somehow we have forgotten all this, and allowed ourselves to be misled into believing it takes a circuit or design flaw producing an oscillation to cause an arc or component failure.

This article demonstrates how easy it is to produce very high voltages from normal perfectly stable PA's with normal tank systems.

Demonstration Circuit

It isn't safe to poke around in a high-power vacuum tube amplifier while looking at voltages, but a simple demonstration circuit can be constructed.
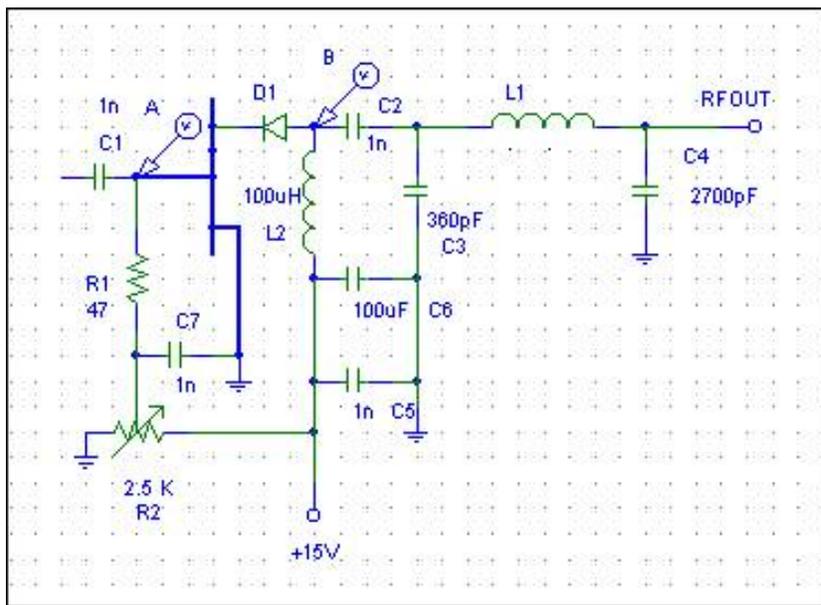
C1 is driven with a signal generator, L1 is selected to match the FET output to a 50-ohm load. The FET is operated at low current and has a series fast-switching diode, to simulate a one-way conducting vacuum tube.
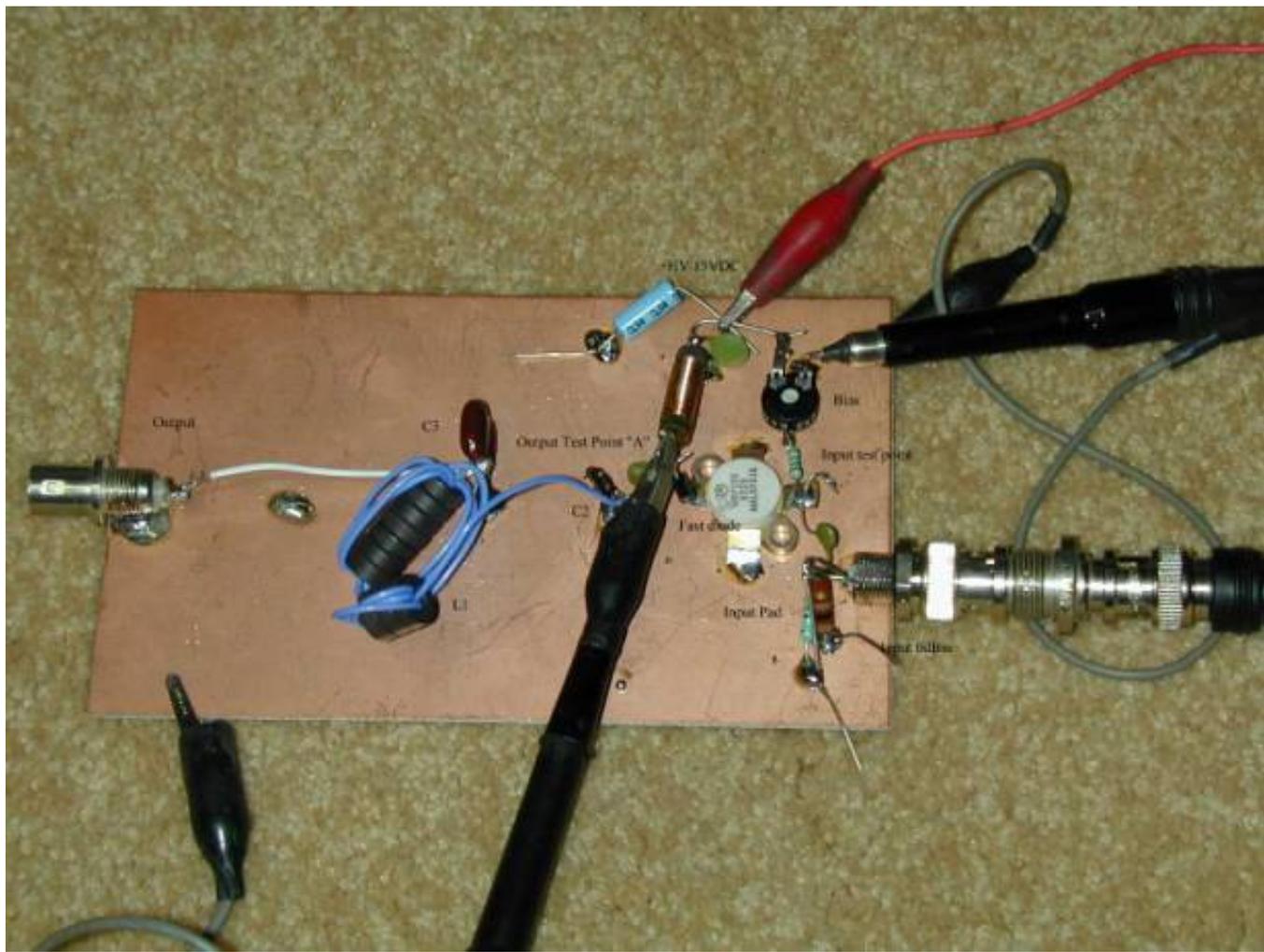
This system was matched at 1.8MHz using the return-loss function of a network analyzer.

Demonstration

0dBm drive was applied with a signal generator, and the resulting waveforms appeared:

http://www.w8ji.com/demonstation.htm (4 of 7) [9/6/2004 9:26:11 PM]
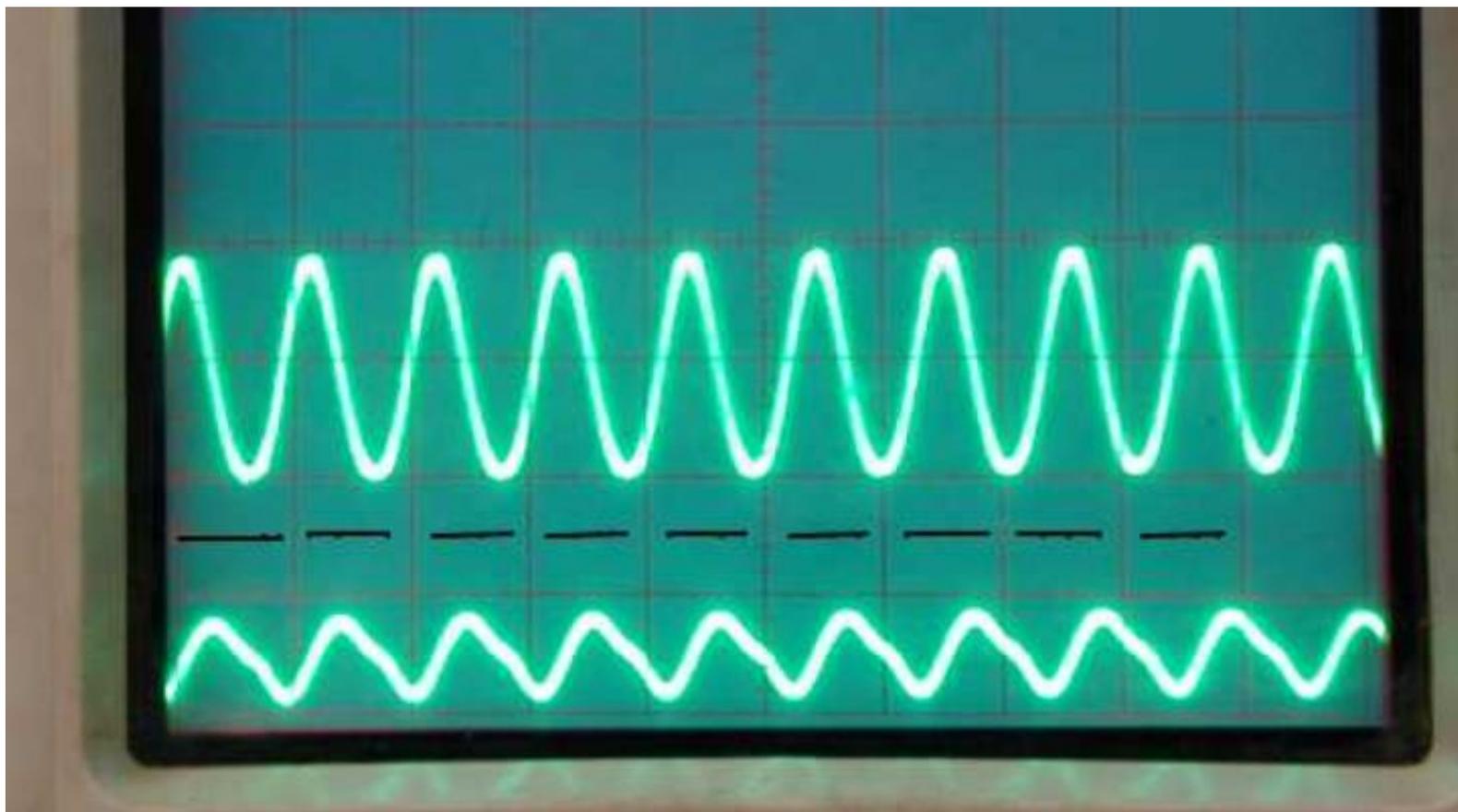
The upper trace is the output at D1 anode (point "B"). The black-dashed line was set at zero-volts. The scale is 10v/div.

The lower trace is drive voltage (point "A"). The scale is .1volt per division.

From this we see peak drain voltage swings approximately 20 volts, from around 5 volts to around 25 volts. This would be normal operation of a PA stage.

If we increase drive power and overdrive the PA, we get the following voltages:
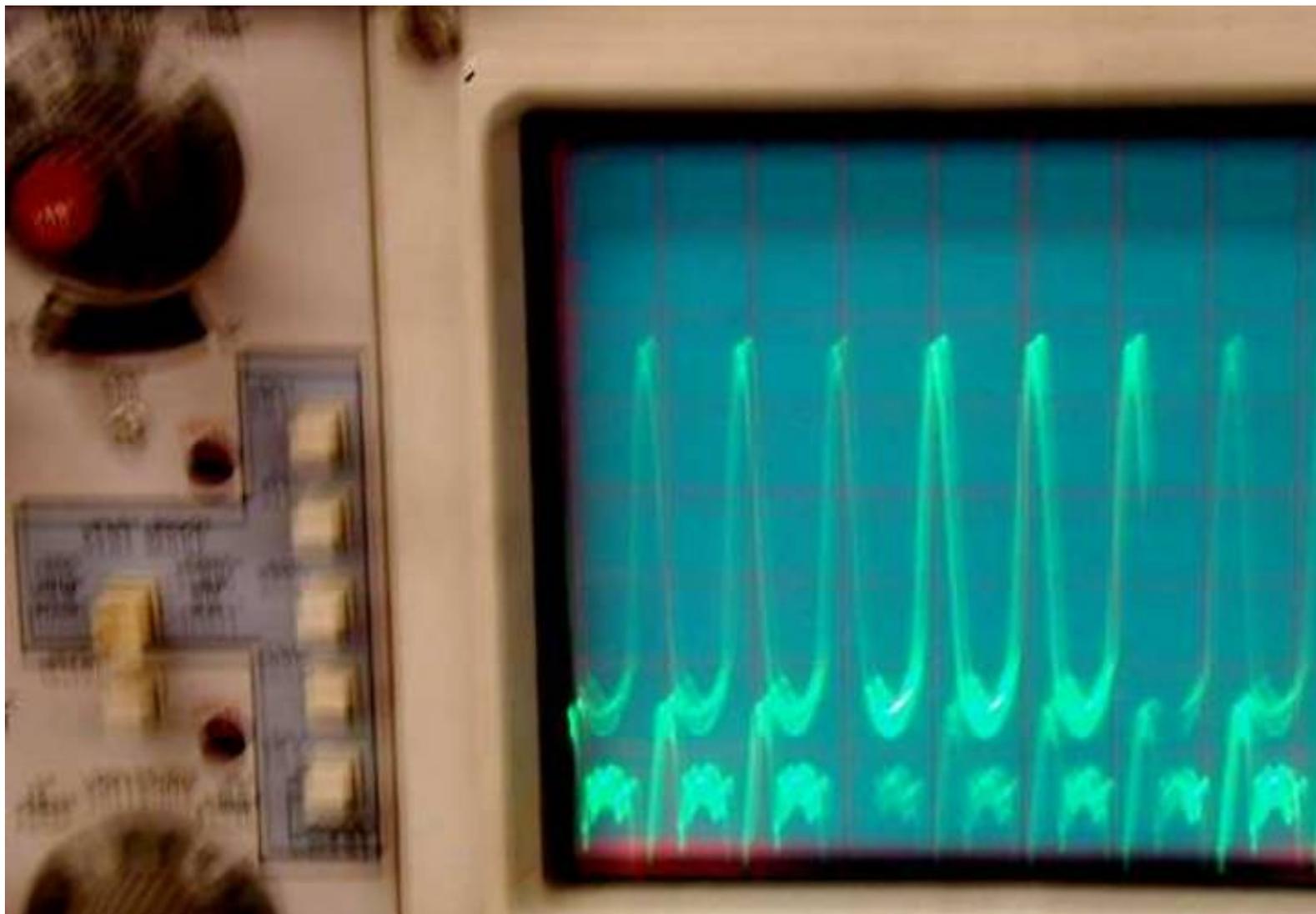
We now see the peak drain voltage is almost 50-volts from a 15Vdc supply! In a 3000V PA stage, this would be 10,000V peak! It should make sense that amplifiers arc from grossly excessive drive power.

The next question would be what makes a normally driven amplifier arc. Often is when the load is inadvertently disconnected, either through poor relay timing, a bad cable or connection, or perhaps a failure in a component between the antenna and the amplifier.

Here is a scope picture with normal drive, but the load disconnected:
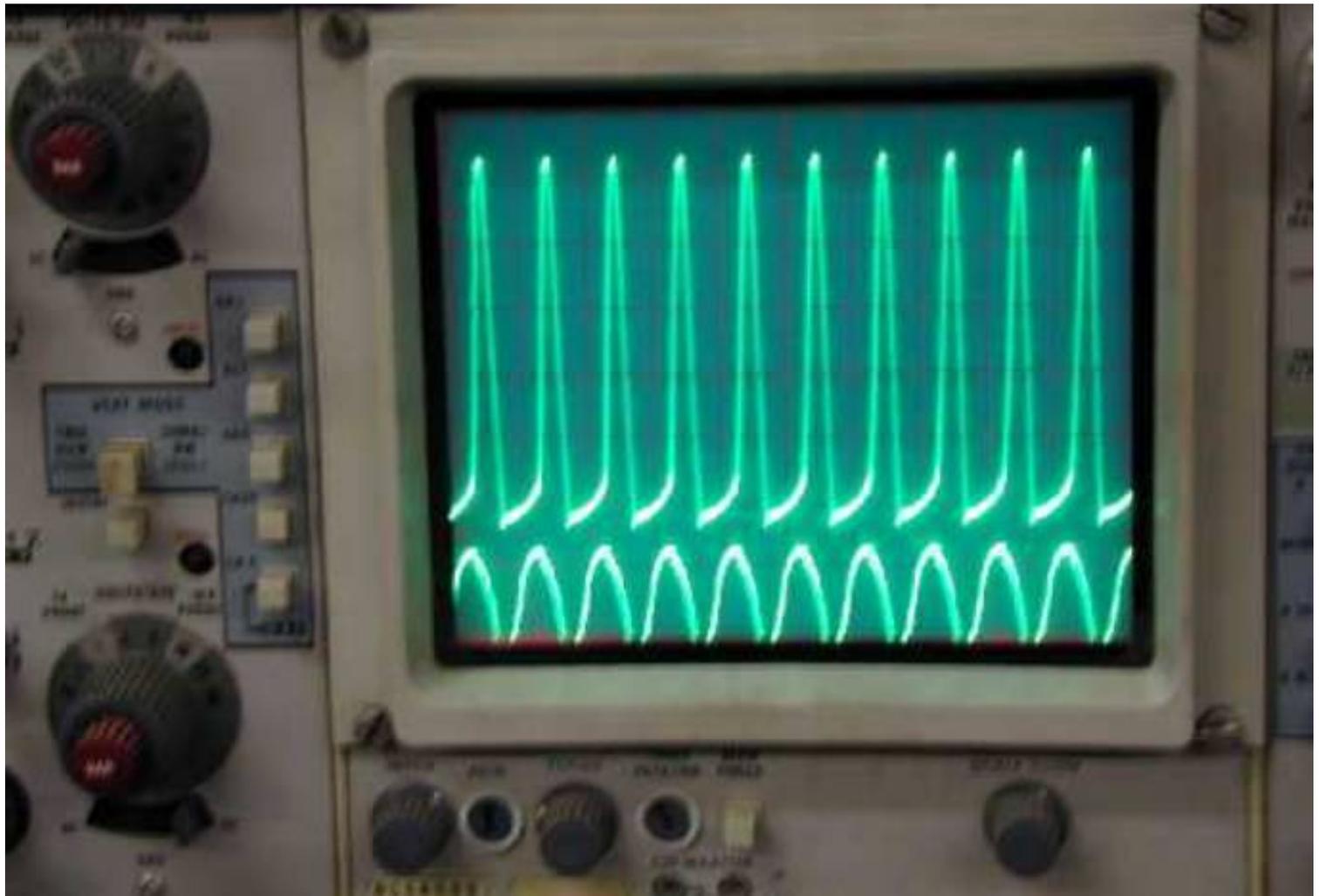
The drive has increased slightly because of feedback through the FET, now that the drain is swinging wildly almost 70 volts, all from a 15-volt supply. This would be the electrical equivalent of 14kV on the anode of a 3-500Z operating from 3000Vdc!

Conclusion

It's easy to see why perfectly stable HF amplifiers, if overdriven or operated at moderate drive levels under conditions of a load fault, can be damaged by severe arcing.

Virtually all PA arcs (other than those caused by component failures or reduced voltage breakdown from dust or contaminates) occur when the load is interrupted or mismatched, and the PA no longer transfers energy to a proper load.

The vast majority of PA failures are caused by improper operation or defective components, not by "strange events" that are unpredictable and non-measurable.

Tank Circuit

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Amplifier tank circuit

The following is a model of an amplifier tank circuit with all parameters shown:

Qa = 0.0

Overall Phase Shift = -130.0 deg

overall effective Q = 11.1

Phase Shift = -130.2 deg

0.6 Amps

1552 Volts   Middle R = 24.4 Ohms   4.4 Amps

-84.3 deg   flip-phase = +/-81.9 deg   -45.7 deg   219 Volts

1000 Watts

Q1 = 10.1   Q2 = 1.0   962 Watts

2407.90 - j17.63 Ohms   50.00 + j0.00 Ohms

swr 1.04   R3 = 0.90 Ohms   50.00 Ohms at 0.0 Degs

X3 = 270.8 Ohms

R1 = 0.05 Ohms   R2 = 0.01 Ohms

X1 = -248.3 Ohms   11.65 uH   X2 = -48.8 Ohms

173.3 pF   6.3 Amps   880.7 pF

6.3 Amps   1701 Volts   4.5 Amps

1.9 Watts   35.6 Watts   0.2 Watts

loss = 37.7 Watts = 3.8 percent = 0.17 dB

Notice the following:

Voltage across the tank coil exceeds input voltage

Current in capacitors exceeds input current

Phase shift across the inductor is -130 degrees

Increasing the loaded Q from ~11 to ~15 results in:

Qa = 0.0

Overall Phase Shift = -145.0 deg

overall effective Q = 15.4

Phase Shift = -145.2 deg

Middle R = 13.1 Ohms

flip-phase = +/-81.9 deg

0.6 Amps

1540 Volts

1000 Watts

-85.8 deg

Q1 = 13.8

4.4 Amps

218 Volts

948 Watts

-59.2 deg

Q2 = 1.7

2372.42 - j17.88 Ohms

swr 1.05

50.00 + j0.00 Ohms

50.00 Ohms at 0.0 Degs

R3 = 0.68 Ohms

X3 = 202.8 Ohms

8.72 uH

8.5 Amps

1723 Volts

48.8 Watts

R1 = 0.04 Ohms

X1 = -181.7 Ohms

236.7 pF

8.5 Amps

2.6 Watts

R2 = 0.01 Ohms

X2 = -29.9 Ohms

1440.8 pF

7.3 Amps

0.3 Watts

loss = 51.7 Watts = 5.2 percent = 0.23 dB

Notice the following:

1. Voltage across the inductor increases
2. Currents increase
3. Loss increases
4. Phase shift increases

Conclusions

(In all cases Q refers to loaded or operating Q)

1. Voltage across the inductor is always more than tank input voltage
2. Voltage increases with increasing Q
3. Phase shift increases with increasing Q
4. Currents increase with increasing Q

Voltages across a bandswitch contact can exceed tank input voltage, if the contact is near the input of the network. This, coupled with the high energy storage, can result in seemingly conservative components failing when the amplifier is lightly loaded for two reasons:

1. Tank circuit operating Q is higher when the tank is lightly loaded
2. The output device pumps the tank, and all of the energy may not be removed. This allows voltages at the input to greatly increase.
   See practical demonstration

Relay Timing

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Related page at **Relay Timing**

The text below covers electronic circuit to sequence and speed separate relays (like dual vacuum relays) and describes how to sequence mechanically ganged contacts like those in open frame relays.

Relay Timing and Sequencing

All amplifiers must be operated with a load on the output any time drive is applied to the input. Without this, the amplifier can be damaged by arcs or excessive internal voltages or currents. After all, all that power in the PA has to go somewhere! If there is no antenna connected, the energy builds up in the tank circuit or other energy storage system in the amplifier until something absorbs the excess energy that would have made it to the antenna.

In cases of unstable amplifiers, such as those using non-neutralized tubes with high internal feed-through capacitance (i.e. 572B's or 811A's), the output relay must be closed before any idling current is applied. (See stability)

The normal process of engagement sequencing is:

- The operator "keys" the radio (zero time)
- Amplifier output contact fully transfers (time "a" 1-10mS)
- The amplifier input contact fully or partially transfers (time "a" plus 1-3mS=time "b")
- Amplifier operating bias or cathode return path contact closes, applying normal operating bias (time "b" plus 0-2mS=time "c")
- Radio output delay, RF appears at the output of the amplifier (time "c" plus 0-10mS)

This entire sequence can take up to 15mS with large open-frame relays.

The release sequence must be:

- Transmitter ceases output of RF
- Bias drops
- Input relay drops
- Output relay drops

If this process is not followed the bandswitch, capacitors, or other tank components can be damaged from intermittent arcing!

Even though all radios should have a transmitter inhibit delay, many radio designs do not include such delays. The radio output delay primarily prevents shutting down of the radio during leading edges while relays transfer, and annoying broadband clicks or spits at the start of VOX closure or each transmitting cycle.

Some radios have incorporated delays, but incorrectly applied the delays. For example, some early Kenwood radios actually delayed the turn-on and turn-off of amplifier relays, forcing amplifiers to "hot-switch". Such radios are often disastrous to amplifier component life.

Increasing Relay Speed with proper sequencing/ Dual Relay System

Relays can be made faster by operating them from a higher-than-normal supply voltage, and using external current limiting to provide a constant current at the relay's rating. For a single relay, omit everything associated with RL2. R3 becomes a jumper.

A two-relay circuit would look like this:
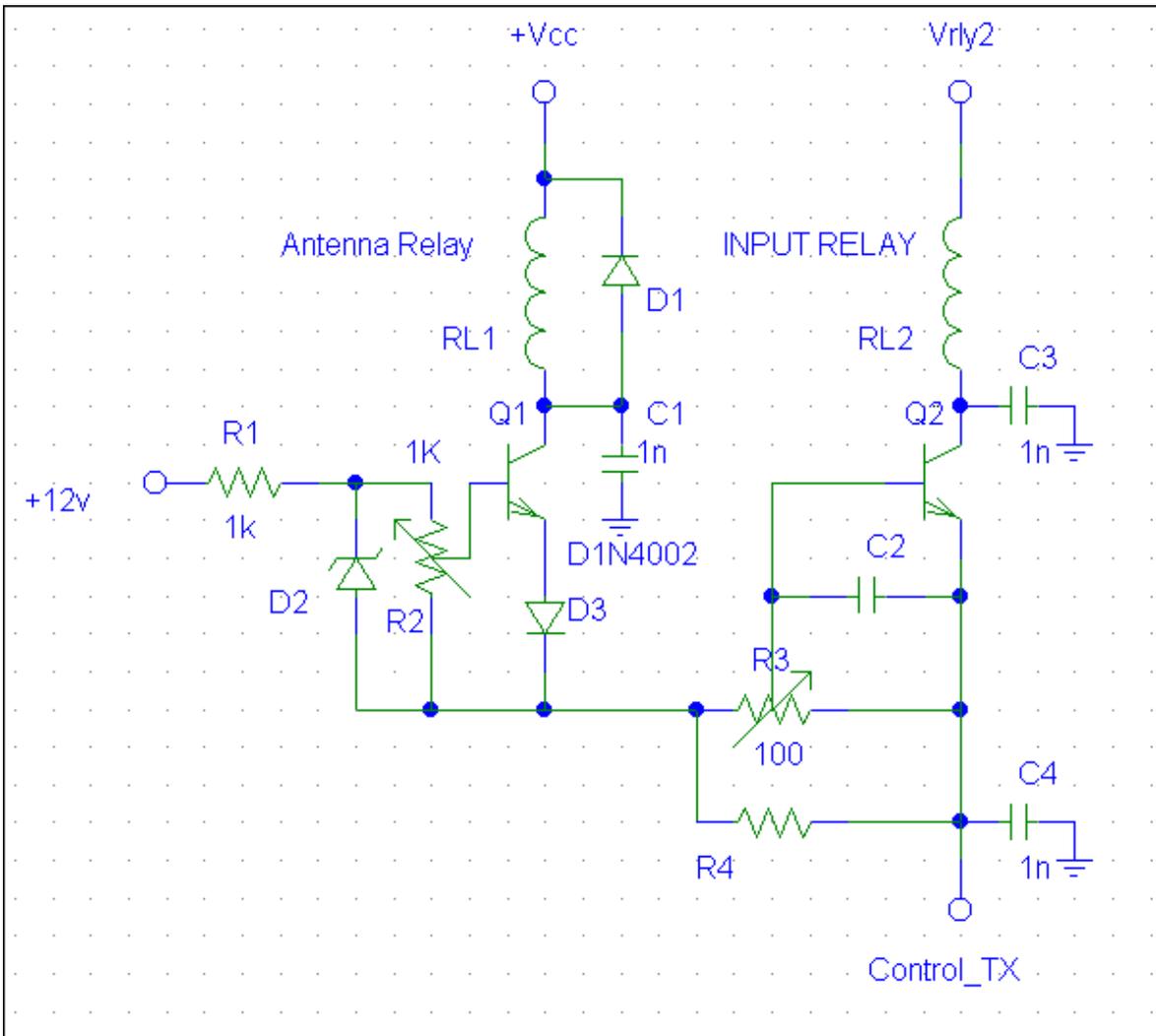
figure 1

R1 1k 1-watt

R2 1k 1/4-watt

R3 100-ohm 1/4-watt

R4 found by the following, where Irl1 is the antenna relay current:

$$R := \frac{.7}{\left(\dfrac{Irl1}{4}\right)}$$

For example:

$$Irl1 := .25$$

$$R := \frac{.7}{\left(\frac{Irl1}{4}\right)}$$

$$R = 11.2$$

Power rating of R4 is:

$$P := I^2 \cdot R$$ where I is RL1 current and R is R4 resistance.

C2 input time delay cap, normally 10-50uFd 10v

D1 1N4001-1N4007

D2 3-4V small zener (.25w-1w)

Q1,Q2 NPN power transistor. Vceo rating must exceed Vcc supply voltage, plus safety margin. Imax must exceed relay current ten times or more. Watch dissipation! Use hfe above 40 for best current stability.

Power Source:

- Vcc supply must exceed RL1 relay voltage by 2-3 times.
- Vrly2 supply should approximately equal RL2 rated voltage. This voltage must be obtained from a reasonably stable supply voltage +- 20% regulation no load to full load. Do not use a large dropping resistor, or you will unintentionally speed up the input relay RL2.

R2 sets current through RL1 (antenna relay). The adjustment range of R2 is from zero current to a maximum where only RL1 resistance limits relay current. R2 must be adjusted for rated current through RL1.

R3 should be set for proper delay of RL2 (input relay).

D1 adds delay to release of RL1 (antenna relay). If release of antenna is too slow, add small value of series resistance until RL1 opens just after RL2 opens.

The above circuit has the following features:

- Control V open circuit will never exceed +12V
- No back-pulse or transient is generated
- Delays are easily tailored to any relay
- Extremely high or dangerous voltages are not present.
- Speed is faster than systems using dropping resistors for speedup
- Relay speed is maximized without exceeding rated relay current

How It Works

Relay pull-in is always slowed by the inductance of the armature coils. Since RL1 is fed by a voltage higher than normal, current will reach pull-in strength much sooner than with normal supply voltages. Q1 is a constant current source, adjusted by R2. This circuit prevents RL1 from being overheated, or having excessive steady-state operating current. Q1 also limits the voltage appearing at Control_TX.

After RL1 (antenna relay) current reaches a value that allows contact transfer, Q2 will conduct. This applies current to RL2, the input relay. It pulls in last, time delay is set by R4, R3, and C2.

Upon release of Control_TX, RL2 releases with only internal relay delays. RL1 is held on by the current loop through D1 as the field tries to

collapse. This causes RL1 to have a longer release time than RL2, so the antenna remains connected longer.

No back-pulse voltages make it to the exciter, because of the circuit configuration.

This is a simple, safe, adjustable relay system.

Mechanically Ganged Relay Contacts

Large multiple-contact open-frame relays using a common armature can be mechanically sequenced by slightly bending each contact carrier, the associated normally open contact, or both.

By carefully pushing on the armature center:

- The output contact should very obviously close first
- The input contact should close second
- The cathode or bias contact should close last, but nearly at the same time as the input contact

Relay Sequencing open frame relay
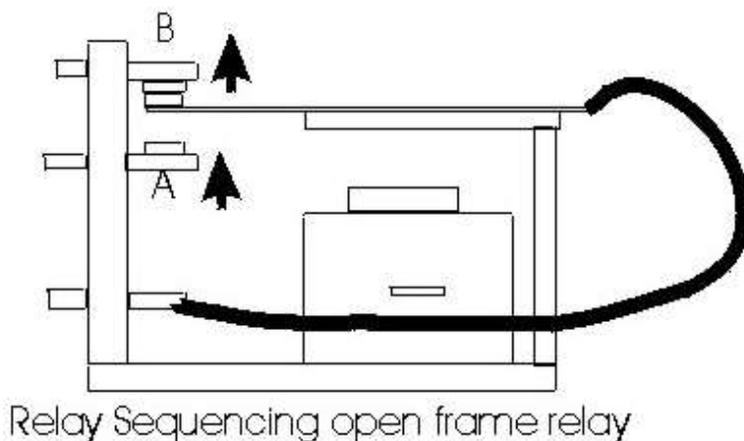
1.) Reference the pictorial above, use needle nose pliers to bend the unused normally OPEN contact at the amplifier tank circuit connection UPWARDS. Do this by positioning one jaw on the top of the bakelite frame and the other under the outer tip of "A" and squeezing until "A" is just slightly bent upwards towards the moving contact. *DO THIS ONLY ON THE OUTPUT CONTACT!* This will force the TANK CIRCUIT of the amplifier to connect to the antenna *BEFORE* the input circuit connects. **This change prevents arcs in tank components caused by improper relay timing.**

2.) Locate the unused normally closed contact for bias switching. Using the same technique above, bend "B" (the normally closed contact) upwards until B no longer contacts the moving contact. **This change puts 1/3 more contact pressure on the amplifier bypass contacts, and decreases receiver dropout when the amplifier is on bypass.**

When properly modified, you should see the *output contact* make slightly before any other contact on the relay when the armature is manually closed. You should see the UNUSED bias or cathode switching normally closed contact have a slight gap when the relay is NOT energized.

Separate Relays

Separate relays offer a unique problem. First, some relays can actually be dangerous to use.

NEVER USE REED RELAYS, UNLESS YOU ARE ABSOLUTELY SURE EACH RELAY HAS A SINGLE REED THAT MOVES BETWEEN THE NORMALLY OPEN AND CLOSED STATIONARY CONTACTS. MANY REED RELAYS HAVE A NO and NC REED IN A SINGLE COIL. THEY WILL RUIN EXPENSIVE RADIOS IF A REED WELDS SHUT!

NEVER USE COMBINATIONS OF SINGLE THROW RELAYS TO MAKE A DOUBLE THROW RELAY!!

DUAL REED RELAYS OR SEPARATE SINGLE-THROW RELAYS (like a SP NO and SP NC combined to make a SPDT section) CAN CAUSE SEVERE DAMAGE TO YOUR AMPLIFIER OR RADIO.

Relays must have an absolutely fail-proof positive transfer between open and closed poles. The normally closed poles should also be series-connected so the path is through both normally-closed contacts. This will prevent any possibility of accidental connection between amplifier output and input in the event of a relay or component failure.

Amplifier and Radio Relay Sequencing

[[Home](#)]

Related Page on [Relay Timing](#)

One problem on SSB and CW is relay time delay, especially when multiple devices are connected. Some radios have inherent RF output to external relay control timing issues. The result of improper timing is often a very wide leading edge "pop" or "click". The pop or click repeats at every relay closure. Slow VOX or semi-break in causes less frequent pops than QSK or fast VOX operation.

Required Sequence

The required sequence is:

1. Key closure or voice start
2. Unwanted delays in keying, VOX, and radio circuits
3. Amplifier relay or accessory (like front-end saver) signal activation
4. Desired RF hold off delay that is slightly greater than longest possible relay pull in delay for all switched devices
5. RF Appears
6. Keying or voice stops
7. RF stops (must include fall time of RF envelope)
8. Desired delay for complete signal fall before relay release

Time Delays

Most small to medium relays for amplifier switching are 5-15 milliseconds, with small enclosed and vacuum relays running around 5mS. Typical open-frame relays are in the 10-20mS range.

- **Relay closing time delay increases greatly as voltage across the coil is reduced. Decrease in speed can be as much as 300% as minimum operating voltage is approached.**
- **Relay opening time is delayed by excessively high relay holding current and back-pulse canceling diodes. The slowing can typically be as much as 200%. If you work QSK, use a design that eliminates back-pulse canceling diodes.**

PIN Diode switches can close and open in a few milliseconds or less.

Some very small reed relays rival PIN diodes in time. There are some low power reed relays that operate in less than 1mS!

Common Errors

The most common errors are:

- **Lack of high operating voltage across the relay coil on pull-in. This is caused by excessive voltage drop in switching components or low relay supply voltage.**
- **Use of relays to buffer or isolate amplifiers**
- **Use of accessories that add time delay to relay lines**

Measuring Time Delays

Measurement of time delay requires a dual channel oscilloscope with separate trigger input, and a low frequency signal generator.

I normally use this keying test setup:
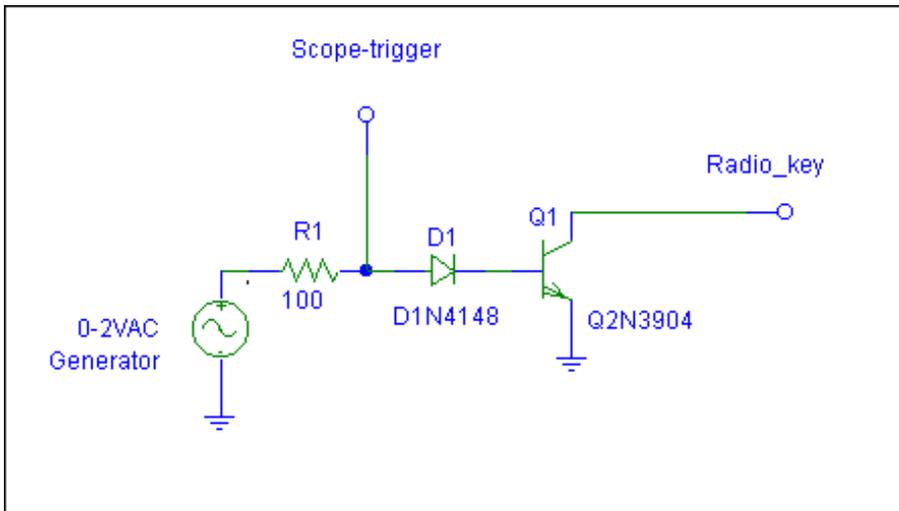
- The level of the generator controls weight or duty cycle (ratio of on-to-off time).
- The trigger is to the scope external trigger. The level threshold of trigger sets scope sweep starting time.
- The scope probes can be moved to any points to compare timing or sequence delays.

What to Watch

The radio must have enough RF output delay AFTER the radio low-amplifier control signal appears so the amplifier relay does not truncate

leading edge of RF

The radio or amplifier must hold the relay long enough so the falling edge is not truncated.

Problem Areas

- **Any device added between the control line output of the radio and the amplifier**
- **Any device that switches on the output of the radio**
- **Front end savers and amplifier relay buffers are a particular problem, often causing excessive control line delay times**

W8JI Control Line Interface

Some systems have multiple control lines from different devices that must be "keyed" in parallel. With direct parallel connections, one device can load the system and stop other devices from operating properly. This usually happens when a device is off but still connected. Simple isolation diodes work to prevent that, but the voltage and current at the radio is the highest voltage of any device and current is the sum of all the device currents. Even worse, the radio output control has to discharge the RF bypass capacitors of each connected device.

My 160 station consists of a solid-state homebrew amplifier with 16 MRF-150's, an AL1200, an 8877 amplifier and either an FT1000D or an Orion. I almost never listen through the FT1000 or Orion when working CW, so I also have at least two outboard receivers that must be muted.

My homebrew 8877 amplifier uses a high-voltage relay system, the solid state amplifier has 5-volt TTL level switching, the TX antenna sampling line uses a 28-volt vacuum relay, and the receivers have 5 volt muting lines that mute when pulled to ground. The Ten-Tec Orion requires an isolation relay on the receive antenna port, and I occasionally switch other devices when transmitting.

All of this is driven by a microprocessor controlled antenna switching system that prevents changing antenna directions or antenna while transmitting.

Any attempt to directly parallel devices would be a disaster!

The circuit below interfaces my radio to multiple amplifiers, receivers, and antenna control systems. Multiple devices can be connected in parallel, such as 28 volt relay control lines and 5 volt control lines, as long as they are all positive voltage. This system pulls within 1 volt of ground.

It is necessary to be careful with some radios. There is no standard interface voltage. TenTec, for example, uses interface systems that have a threshold around 1 volt. TenTec's handshake lines require buffering with a small very fast reed relay, many transistors and IC's will not pull low enough to activate the TT handshake line. If you need relays of any type between the amplifier keying system and the radio, always be sure you test the entire system for proper relay sequencing.
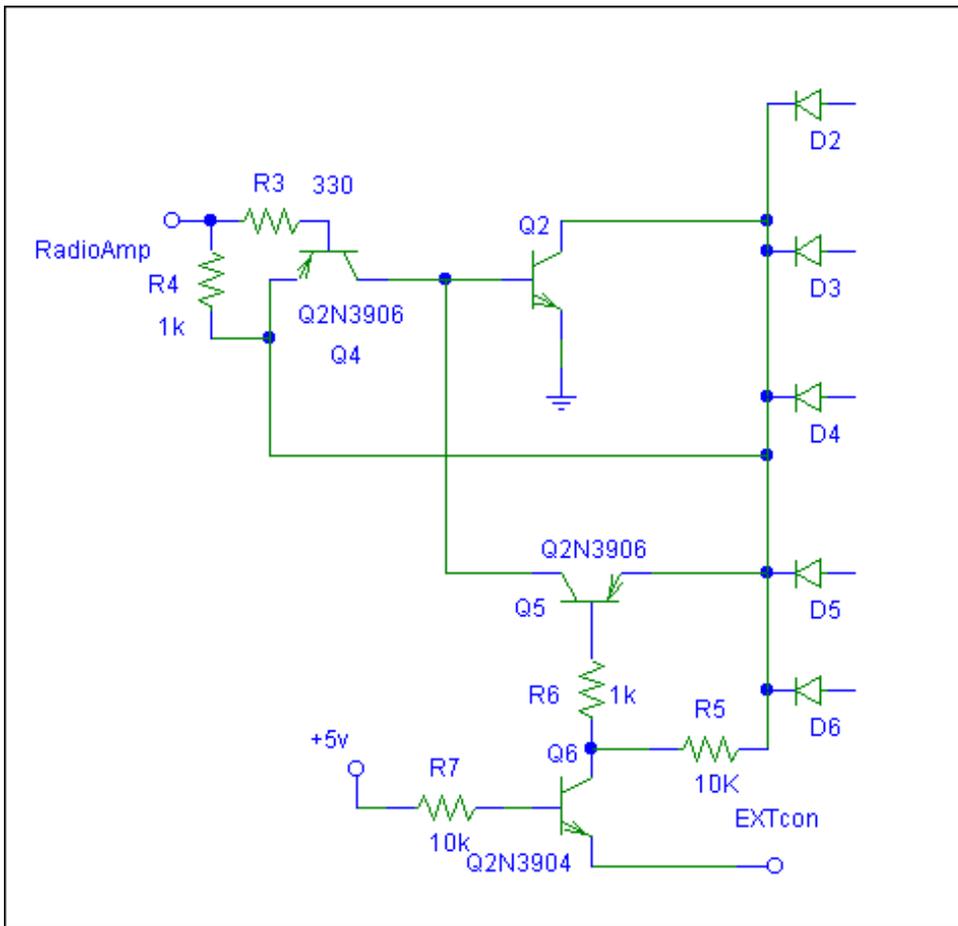
RadioAmp goes to radio's amplifier keying control output.

EXTcon goes to computer or other keying device that only handles low-voltage low-current.

D2-D6 go to devices that require control.

+5V sets the threshold of EXTcon trigger. EXTcon open circuit voltage is also limited to less than 5 volts.

The radio is exposed to the highest voltage from the keyed loads, but the current is limited to roughly V/500 where V is highest voltage device. A 28 volt amplifier would require the radio to key about 14mA peak current during relay transition, although average current would be lower. The exact current depends on beta of Q2 and Q4, and if EXTcon is used. A device like Q6 can be added to the radio line to limit radio voltage to 5 volts or less regardless of amplifier relay voltages.

HF Stability

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HF instability, or instability at the operating frequency of the amplifier, is the most damaging type of problem. This is because the primary energy storage system, the tank circuit, is resonant with high Q at or near the operating frequency.

If you look at [network analyzer swept measurements](#) of a tank circuit, you'll see the maximum voltage distributed through the tank occurs at or near the operating frequency.

Some amplifiers have designed-in HF stability problems. The [Yaesu FL2100](#) is one example of an amplifier with poor RF design. Un-neutralized amplifiers using tubes with high feedthrough capacitance are another source of problems. Any amplifier with high feedthrough (or feedback) capacitance is a candidate for HF instability if the load is removed while the amplifier tube or tubes are drawing quiescent current. Dentron amplifiers with four 811A and four 572B tubes, as well as the Collins 30L1 with four 811A's, are examples of relatively unstable amplifiers (along with the FL2100).

The cure is to remove intentional undesired feedback or to neutralize the PA!

SB-221

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SB221 Tank Circuit Voltage Distribution Vs. Frequency

Arcs in the bandswitch, tuning capacitor, and other tank components are often blamed on parasitics. No one ever can catch the amplifier in such a state of oscillation, so we are just left with conjecture and speculation that the arc was caused by a VHF parasitic.

It should be easy to prove if a VHF oscillation, or oscillation or drive on any frequency, could cause an arc. All we need to do is excite the anode of the PA tube with RF, and measure the voltage distribution in the tank with a very high impedance detector probe. This would then tell us at what frequencies the highest voltages occur.

While every amplifier manufactured is claimed to suffer from parasitics, I have limited time to measure and document the results. It would be impossible to test every PA, so I picked a PA that is commonly rumored to have parasitics.

The purpose of the series of measurements below is to see if it is possible for resonances in the tank to INCREASE anode voltage to the point where an arc will occur, or if it just an imagined problem or a problem "invented" to sell a specific cure.

Test Method

 It is not practical to drive an HF PA from frequencies between 1 and 300MHz, so we have to excite the tank another way. The best method is to use a constant voltage source injected at a tube anode, and measure the voltage throughout the tank without disturbing anything. This can only be accomplished with a high input impedance detector probe, and a voltage-leveled source. Fortunately I have both available.

In order to be accurate, the tank must be adjusted normally. I did this by tuning the PA with a 100 watt exciter and a 50 ohm load. After tuning, the amplifier is disconnected from power mains and from the exciter.

The tube is purported to be the source of unwanted VHF oscillation energy. By injecting the anode with a known voltage, and sweeping the tank slowly over a wide frequency range, we can determine voltage distribution at any point in the tank system. By looking at the relationship to anode voltage, we can determine if it is even possible to arc a tank at frequencies far removed from the operating frequency.

Of course this would not prove an arc would occur, it would simply prove if it is possible for an arc to occur and at what frequency. The higher the voltage at a given frequency, the more likely an arc would occur. If the voltage is much lower, and arc at that frequency would be impossible.

The Measurements

The anode is excited with a swept power-leveled source feeding a bridging pad and the anode. The sweep is from 1 MHz to 300 MHz. The small coaxial cable and multicolored wire are from the high impedance J-FET source follower probe used to feed the second RF port of an Agilent network analyzer. In this condition the network analyzer acts like a sweep generator with a voltage detector.

- HP Network Analyzer source connected to tune anode at tube. Constant power source leveled at 10dBm at injection point load.
- 100k-ohm 3pF active voltage detector probe
- Transmission loss calibrated at zero dBm for mid-scale into the 100k probe as detector, across 2700-ohm resistor.
- All displays 10dB/div voltage, sweep is 1MHz to 300MHz.
- 1600 data points, 5 second sweep time.
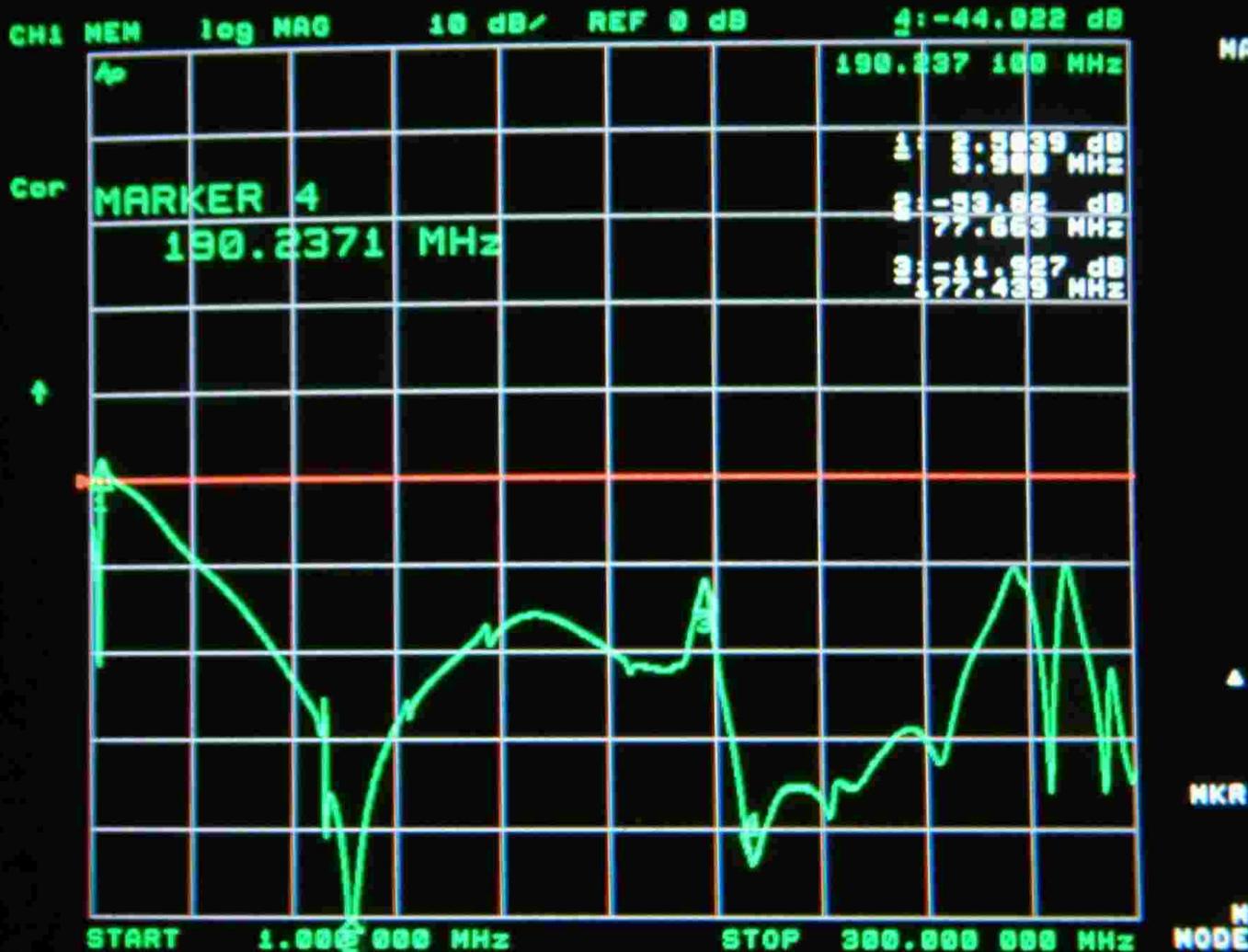
Figure 1. Tank adjusted for normal operation at 3.8MHz. The above sweep is voltage across the tuning capacitor, which is +2.5dB at plate tune capacitor. +2.5dB becomes reference for comparing HF voltages to parasitic voltages. With the anode at 3000 volts peak swing at 1300 watts on 80 meters, the reference level would be 2.5dB. A 1300-watt parasitic would develop the same starting voltage swing and it would be attenuated through the tank by the same amount as the displays indicate.

In this case at the tuning capacitor, a 180MHz parasitic would be 14.4dB down in level (marker 3) compared to the HF voltage (marker 1). This is 19% of the HF voltage level, or 570 volts peak with 3000 volts of anode voltage swing. A VHF arc at 180MHz would be impossible. As a matter of fact, there isn't any frequency very far from the operating frequency where an arc could occur!

A grid dip meter would show a strong dip at 180MHz, because it would be loaded by the tank. This is why a grid dip meter can not locate a "problem frequency", it might actually be locating a good frequency!
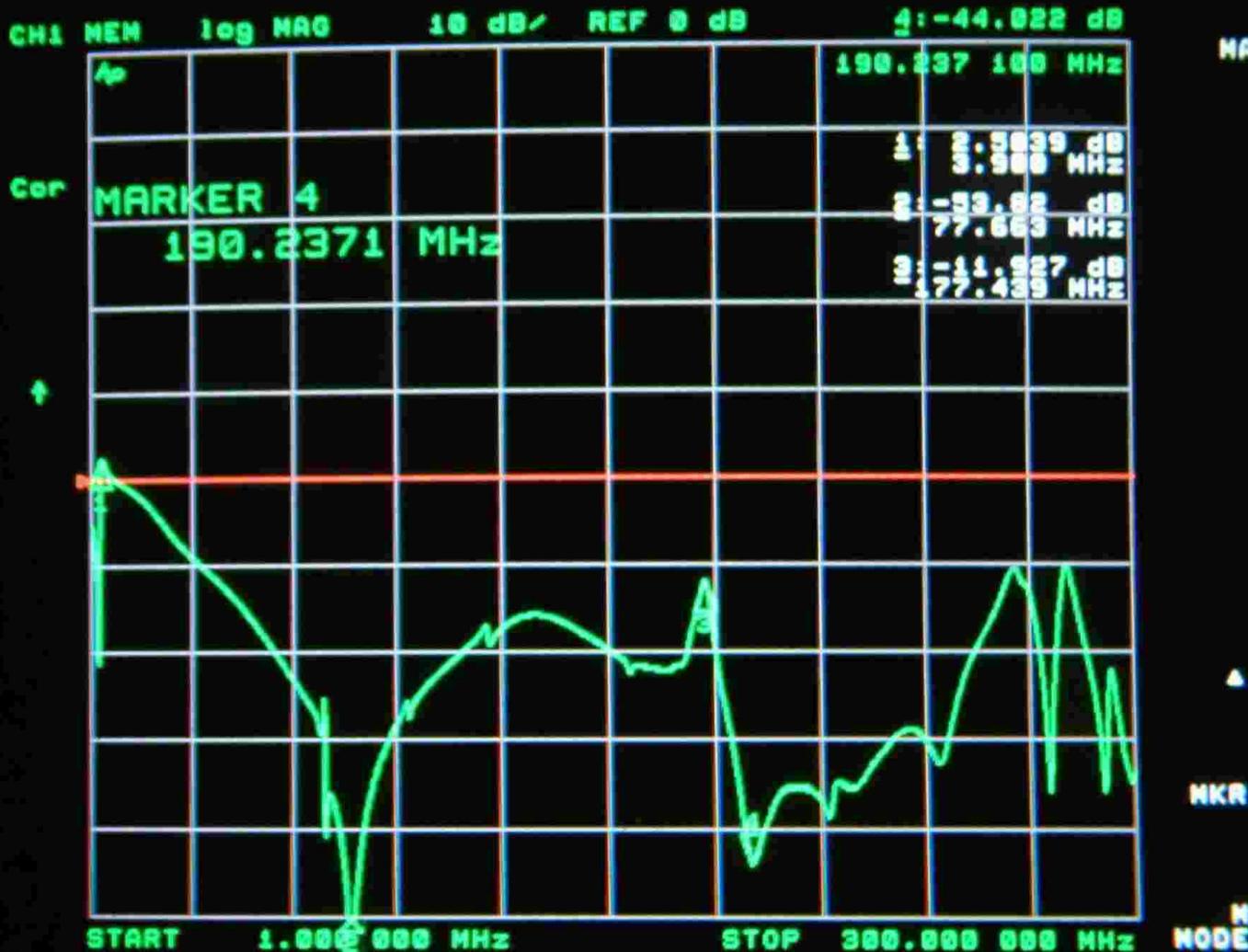
Figure 2. 80m position. The above graph indicates voltage at the 15 meter tap to closest arc point on switch. Highest voltage at frequency marker 3 is -12dB at 177 MHz. This is -14.5dB anode voltage, or 19% of anode voltage! 19% of 3000 would be only 570 volts, making it impossible to arc the 15M tap at VHF frequencies. After all, that tap survives constant exposure to over 3000v at the fundamental frequency when operating 80 meters!
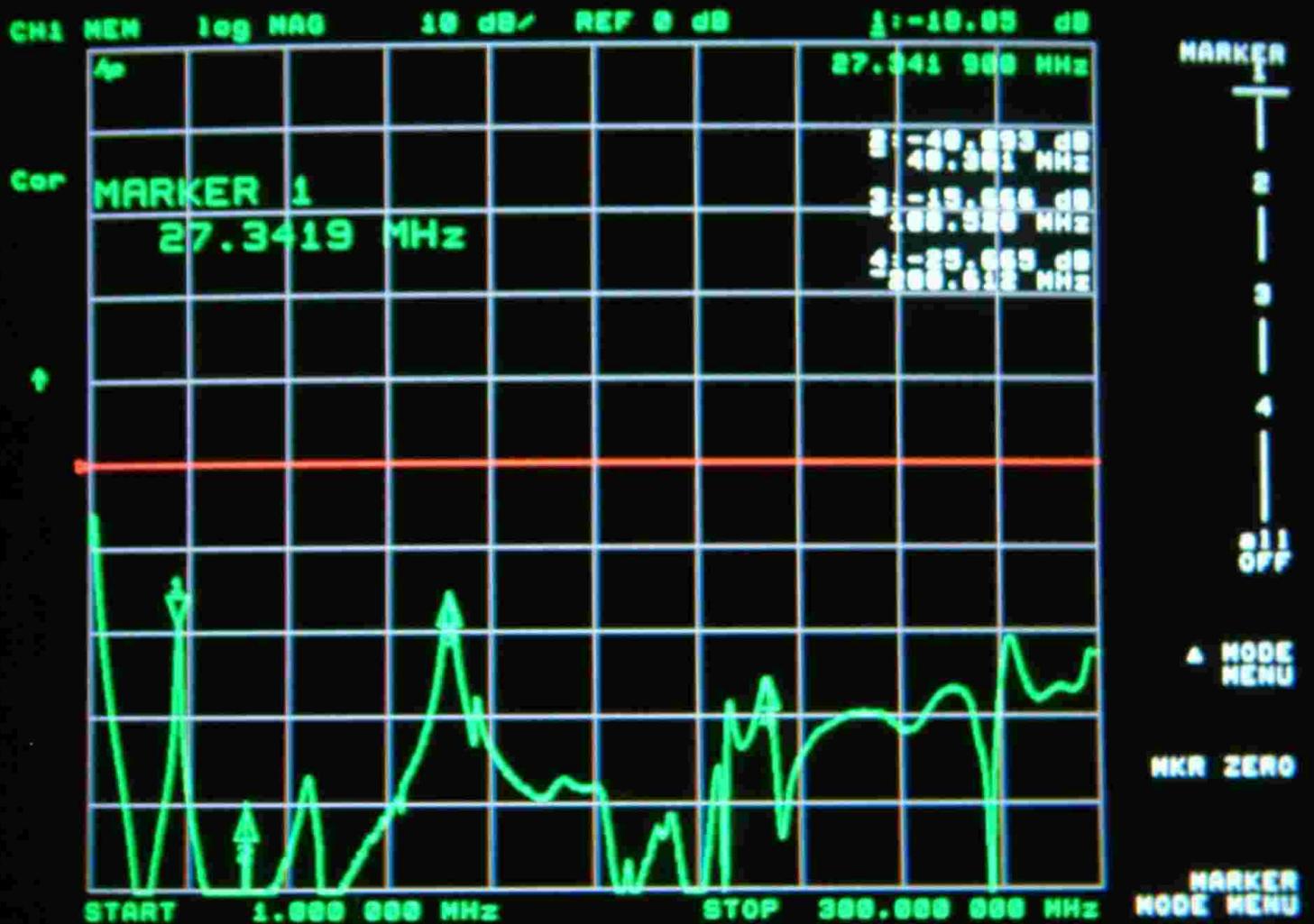
Figure 3. 80m position. Voltage across loading capacitor. All spurious response voltages are at least 20dB down from the anode voltage. No VHF arcs possible here either.
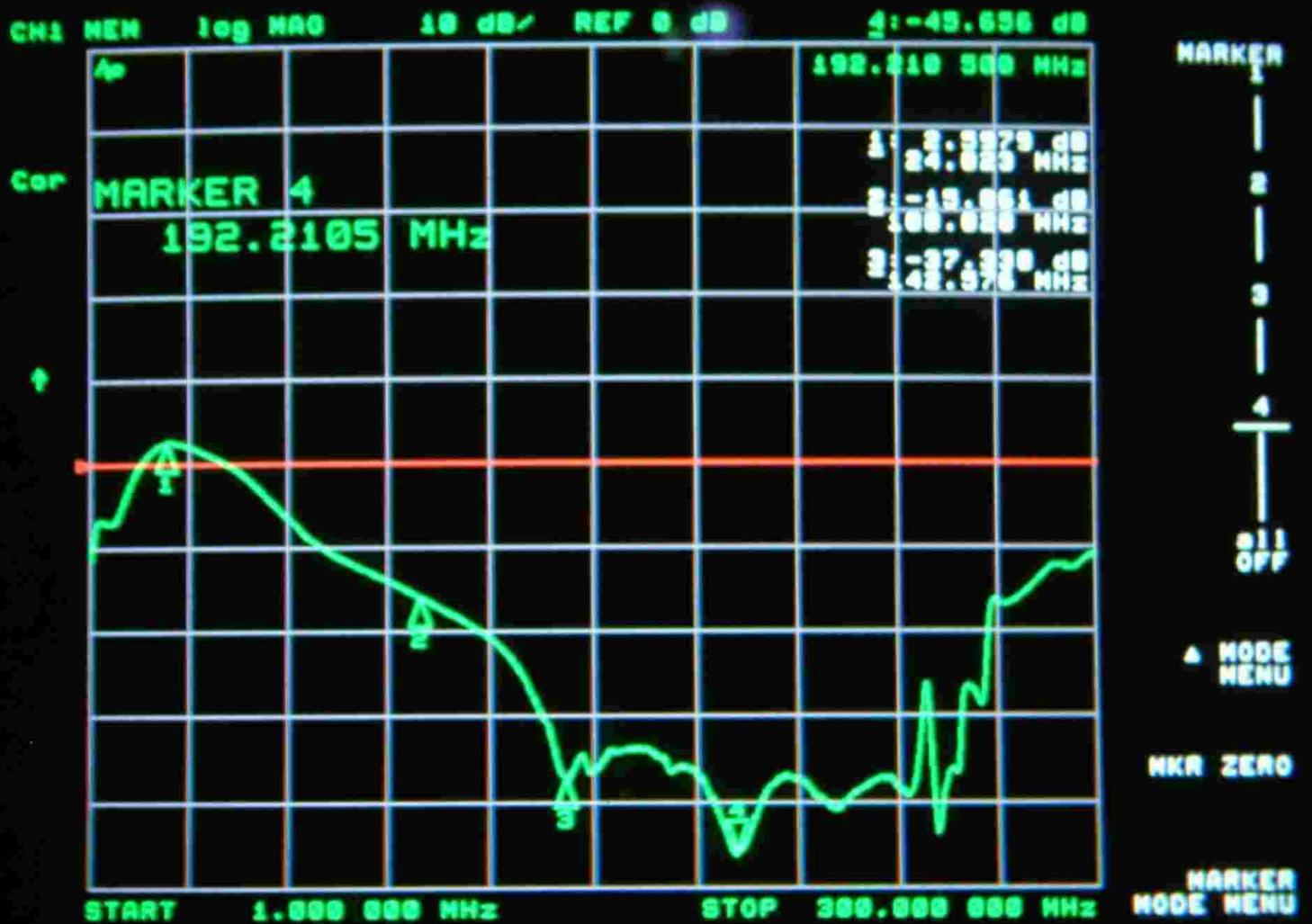
Figure 4. Tank normally loaded on 15-meters. Voltage across tuning capacitor. Anode injection voltage reference at +2.6 dB.
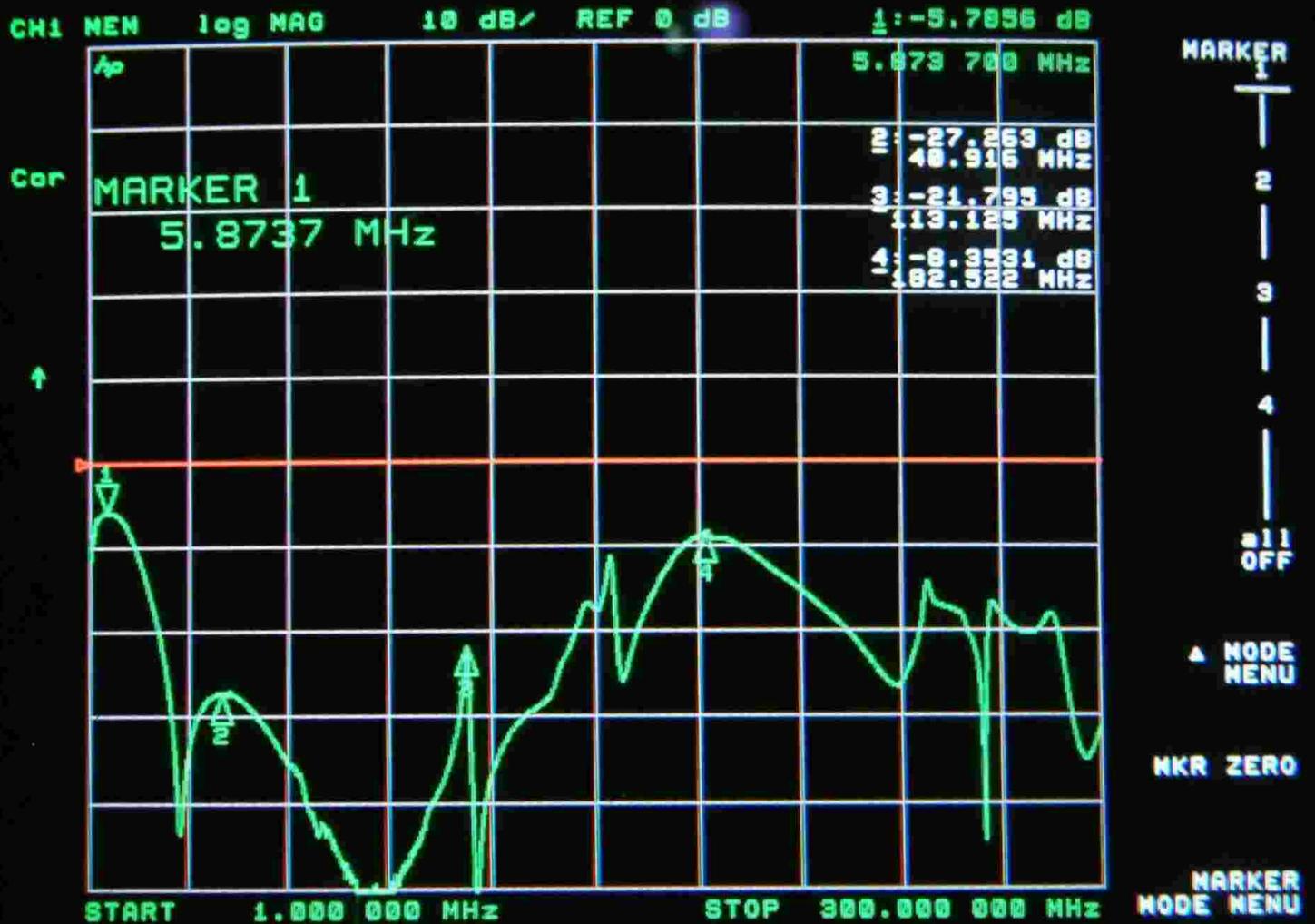
Figure 5. 15m operation. Voltage at highest VHF voltage point in system past the parasitic suppressors. This voltage was at plate blocking capacitor on plate choke side. Point 4 is 11 dB down, or 27% of anode voltage.
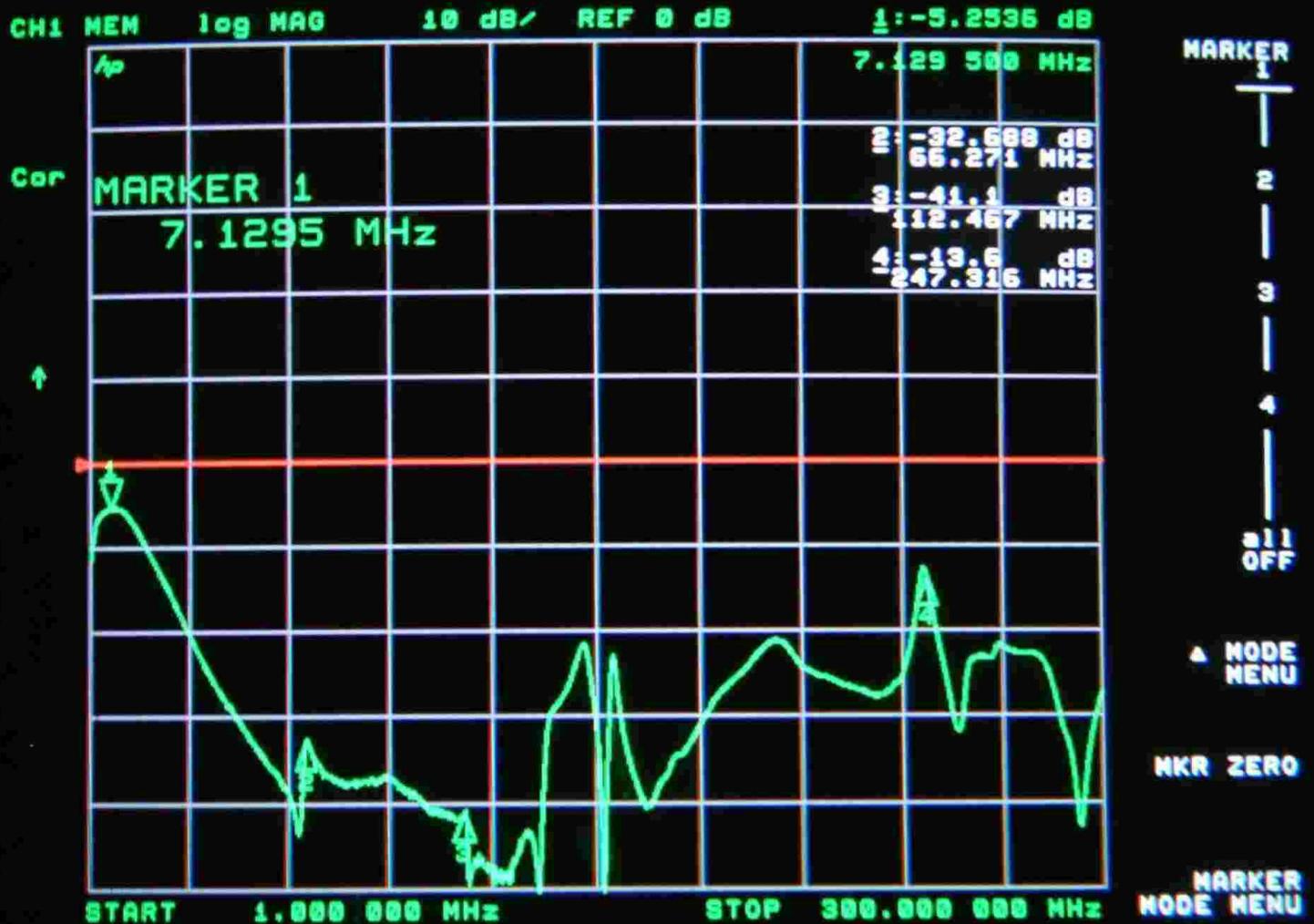
Figure 6. Voltage across loading capacitor at tank-end stud to chassis. Peak at 247MHz is -16dB, or 16% of anode voltage (480v at this point with 3000v at anode). This peak is caused by layout, and the loading cap coupling directly to tubes through stray capacitance from anode system. Impossible to have a VHF arc here, but it could aggravate TVI at USA channel 13.

Conclusion

From the above swept data, we can see VHF voltage would always be much less than HF voltage. The only likely candidate for arcing is improper operation at or near the operating frequency.

What causes arcs? Click here.

More SB221 will be added at a later date. Including performance improvement modifications.

FL2100 Problems

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The reason Svetlana tubes oscillate in the FL2100 is two-fold:

1. Bias is inadequate for Svetlana tubes. Svetlana tubes have LOWER mu (not higher as some people claim) and draw more current for a given amount of bias.
2. The FL2100 has a very poor attempt at neutralization or feedback. It also has a poor bias system.

Bias Error

The FL2100 only applies ~15 volts negative bias to the 572B grids for cutoff. This in not adequate for some brands of 572B tubes, especially those with slightly lower mu.

Feedback Error

The FL2100 design uses a capacitor from the OUTPUT of the pi network back to the input. One should NEVER do that, because the pi transforms phase along with load impedance and feedback voltage depending on control settings and load impedance!

A good design would never neutralize or add feedback from the OUTPUT of a tank circuit with a varying load back to the input.

What the design errors do in the Yaesu FL2100

When the antenna relay open, the LOAD is removed from the pi network. This INCREASES the amount of feedback by several times. The tubes, having less mu and requiring more bias, are no longer in cutoff with the relay open and draw current. This allows the tubes to amplify, and the greatly increased feedback causes the FL2100 to oscillate on whatever frequency the tank is tuned to. With no load, the voltage soars to many kV until the bandswitch arcs.

Changing tube brands in even a reasonably design amplifier is no problem at all. Changing tube brands (especially when a brand does not have the same characteristics) in an amplifier with multiple major design errors can be a big problem.

The Svetlana "fix" of increasing FL2100 bias is really not a " fix".

Corrections

Feedback

The most important "fix" would involve removing the capacitor from the output of the tank back to the tube input. This would disable the very poor feedback system.

If the amp is unstable at upper HF, a neutralization system similar to that used in the AL-811H Ameritron should be used.

Bias

The FL2100 bias system should be rewired so the relay opens the filament CT return. Normal operating bias should be left on the grids. This would FULLY cut-off any tube brand.

Switch Installation

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Switch Installation Problems

Wiring

Wiring must not have sharp points or protrusions. Solder connections should be smooth and rounded, the solder should generally for a small rounded ball that coves all sharp edges. The last motion in switch wiring to each contact should be soldering, not clipping! Clipping of protruding wires results in sharp edges. Sharp edges concentrate the electric field in a small area. This greatly decreases voltage breakdown. The difference in voltage breakdown can be profound, more than 50% reduction in voltage breakdown may result from changing a rounded connection to a sharp point.

In some cases, and intentional anti-corona ring is added near the highest-voltage contacts. Any large smooth rounded metallic object connected to the contact will disperse the electric field and reduce voltage gradient between that contact and other nearby connection or hardware. This will greatly increase switch voltage hold-off.

Contact Alignment

Switches normally have a detent located at the front of the switch. The detent causes the switch to "snap" into each band position, keeping the contacts centered.

The detent may consist of one or more balls held with spring pressure against a ball-race with recesses or dimples, or a spring-loaded cam follower that presses against a notched wheel attached to the shaft.

When the mounting nuts of supports are loosened, clearance in wafer holes and mounting brackets allows wafers to rotate a few degrees. This allows the contacts of each wafer rotor to be centered in the mating wafer stator contact. Pick-up-and-hold contacts are normally adjusted so they do not extend beyond the edge of the leading-edge stator contact.

Contact alignment is normally performed in the middle of rotation, checking by rotating in from each direction. The center position is selected, and the wafer mounting screws loosened. Each switch wafer contact group is centered or aligned, and the mounting screws retightened. After

adjustment, each position of rotation should be checked for contact centering or alignment.

Contamination

Dirt and dust are deleterious to voltage breakdown of components. They are most harmful under high humidity, when moisture and salts in the contaminant may increase conductivity. Contaminants mainly reduce voltage breakdown because they fill air gaps with slightly conductive particles.

Well-designed switches normally have much closer contact spacing than dielectric paths, because contact angles move the ends near the rotor close together. Even so, contamination of the (normally) ceramic dielectric can be an issue in conditions of severe dust buildup.

Never use contact cleaners or other lubricants on high-voltage high-current switches or relays. Many contact cleaners have carbon-based chemicals that will speed the formation of carbon tracks when corona or arcing forms. Other lubricants are very poor RF dielectrics, reducing voltage ratings.

Cleaners should be pure light hydrocarbons that fully evaporate, leaving no residual contaminants.

Preamp Summary

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When comparing low-band preamplifiers the important parameter is almost always output TOI.

Output third order intercept is a measurement standard of how well the amplifier stands up to strong signals with creating new "phantom" signals through unwanted mixing.

Type	Advertised range	NF	Gain peak	OUTPUT TOI (third order intercept)	1dB gain compression (output)
ARR 1-30	1-30MHz	5.43dB	20.4dB@13MHz	30.38dBm	21.5dBm
ARR GAsFET	single band	3.2dB	26.4dB	11.39dBm	8.1dBm
DXE RPA-1	200kHz-40MHz	3.4dB	17.3@5MHz	43.04 dBm	26dBm
ICE 124A		4.2dB		14.05 dBm	
KD9SV	160M/80M	4.9dB		14.02@full gain 0.14 dBm@12dB below full gain	

Note: *These are measurements of a single new random sample that appear to be working properly, with no noticeable defects in components or wiring.*

Vacuum Tubes

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Power Grid Tubes

Power-grid vacuum tubes cover a wide range of applications, and a wide variety of styles have evolved over the years. Every design is a compromise of some type, designed to better fit a few specific applications. For example, there are dozens of grid and cathode (electron emitter) designs. There is a wide variety of anode styles and types. Even the heaters or filaments have multitudes of shapes, materials, and physical dimensions.

The big enemy of old vacuum tubes or tubes that have sat a long time is gas. The big enemy of high gain power grid tubes (or valves) using metal oxide cathodes is grid current and excessive cathode current. Many common problems are discussed below.

Numbering System

While manufacturers use a variety of numbering systems, most Eimac tubes use an easily decoded system.

First number = number of active elements in the tube

- 2=diode 3=triode 4=tetrode 5=pentode

First letter = envelope type

- nothing or - = glass
- C = ceramic

Second letter = anode type

- nothing or - = internal
- X = external
- W = water

- V = vapor

Number group = anode dissipation (actual dissipation is much higher than number is some tubes, so check the books)

Last letter = Base or connection type

- Blank = pins or tabs
- A or Z = coaxial or pins
- F = flying leads

Last number = mu, where 1 = lowest mu through   7 = highest mu

- 1 through 5 are best suited for grid driven applications
- 7's are best for cathode driven

From this we see a 3CX3000F7 is a triode with ceramic envelope and external anode, has approximately 3000 watts dissipation (in fact it is 4000w), has flying leads, and highest mu (most suitable for grounded grid).

In addition, some tubes have additional letters inserted in the type number. Examples are the 4PR1000 or 3CPX800A7. The P generally signifies pulse rating, and the R more rugged construction.

Electron Emitters

Power-grid vacuum tubes operate by a thermionic emission process. In a vacuum tube, we commonly call the electron emitter a "cathode". The cathode can be directly heated or a filament-cathode. Another type is the indirectly heated, or heater-cathode type of electron emitter.

The cathode, regardless of sub-category, uses a material easily able to "give-up" electrons with heat. The cathode is heated in a nearly perfect vacuum to temperatures above 1000K (we would see that as a red glow) to as 2600K (a yellowish-white glow). This heat supplies energy to loosely bound electrons at the cathodes surface, causing some electrons to escape the material. These electrons form a cloud suspended around the cathode.

Commonly Used Cathode Materials

Material	Operating Temp Range K	Emission mA/watt	Emission mA per cm^2	Special Features

Pure Tungsten	2400-2600	2-10	100-1000	Immune to positive ions, can be operated at highest anode voltages. Instant on.
Thoriated Tungsten	1800-2000	50-100	700-3000	Much higher emission, but less immune to positive ions. Can be operated at reduced filament voltage or operated with peak currents over emission limits with little effect on life. Instant on.
Oxide-filaments	1000-1100	200-1000	400-3000	Very high emission per watt of filament power, but extremely sensitive to positive ions. Emission failures occur from low temperature operation if electron cloud is depleted. Must be thoroughly warmed up before any current is drawn, but warm-up time is short.
Oxide-cathodes	1000-1100	10-200	1000-3000	High emission in small cathode. Long warm-up time. High peak emission for pulse applications.

Positive ions are formed through ionization of residual gasses. High anode voltages speed formation of positive ions. The space charge (electron cloud) surrounding the cathode repels positive ions, and keeps the ions from poisoning the cathode. When positive ions contaminate the cathode, electron emission is reduced.

Tungsten has the highest melting temperature of any metal, above 3600K. Tungsten has the ability to operate at high temperatures for many hours without evaporation of materials becoming a major problem. Tungsten is one of the few materials able to withstand total depletion of the electron cloud in a vacuum tube. Tungsten emitters are not damaged by stripping away of electron clouds, and the resulting cathode bombardment by positive ions. This immunity to damage means emission life is

not shortened by excessive current, operation before full temperature is reached, or low filament voltage.

If a cathode material sensitive to positive ions is operated with excessive HV, or has excessive cathode current for the cathode operating temperature, it will suffer emission failure. This is why indirectly-heated tubes or oxide-cathode tubes must have long controlled warm-up times before any cathode current is drawn. We must *NEVER* operate the 3CX1500A7/ 8877, 3CX800A7, or other oxide-cathode tubes at reduced heater voltage! This would allow the cathode's protective electron cloud to be depleted, and nothing would prevent positive ions from striking and poisoning the cathode.

Anodes

The anode is operated with a positive potential. This creates a strong electric field that attracts electrons from the cathode. The electrons are accelerated (or impeded) as the pass through the opening in the grids, and eventually strike the anode at reasonably high velocity. Since the electrons have mass and speed, they have kinetic energy. This energy produces heat, as well as dislodging other electrons from the anode. The amount of heat is proportional to the velocity and number of electrons. The power converted to heat is the anode voltage times anode current, at any instant of time.

The actual plate dissipation is the time-integrated amount of anode current and voltage. In an amplifier with steady current, anode dissipation is found quite easily by multiplying the accelerating voltage (anode-cathode potential) times anode current. In an amplifier system, dissipation is a complex function of the constantly changing anode voltage and current. In such systems, a very close approximation of anode dissipation is given by multiplying average anode current by average anode voltage, and deducting the useful power extracted by the load.

Internal Anodes

Many lower power tubes use internal anodes. Internal anode tubes can use any electron emitter or grid configuration, although the most common large transmitting types use thoriated-tungsten filament/cathodes. Internal anode tubes are cooled by infrared radiation, by direct thermal conduction to the envelope, and through external connections via leads exiting the envelope.

The anode often serves three very important functions in a transmitting tube:

- The anode collects most of the electron current from the electron emitter in the tube (cathode)
- The anode must dissipate heat produced from kinetic energy of electrons that strike the anode

- The anode often degasses the tube

The material used in the anode varies with tube type and manufacturer. Most transmitting tubes use molybdenum, tantalum, graphite (carbon), or other high temperature materials in the anode. The material must be mechanically and electrically stable, even while operating with very high temperatures.

In transmitting tubes, virtually all anode heating comes from the kinetic energy of electrons striking the anode. The actual resistance of materials in the anode (and other elements in the tube) is very low, and the relatively small amount of $I^2 \cdot R$ heating is dwarfed by the hat caused by electrons impacting the surface of materials in the tube.

Higher power internal anode tubes almost always have a gettering material coated directly on the anode. The getter acts as a sponge, soaking up any gas molecules inside the tube. This is necessary because any gas creates positive ions, reducing tube life. Not only that, even the slightest amount of gas greatly reduces breakdown voltage through a vacuum.

Note: Peak anode voltage in a properly tuned and properly driven amplifier approaches two times the dc anode potential, and under conditions of mistuning, load faults, or excessive drive anode voltage can be several times the dc anode potential. It is important that any tube have an anode breakdown voltage at least four or more times the expected highest dc anode potential, or occasional anode to grid (and cathode) arcing may occur.

The most common gettering material is zirconium. For example, zirconium is used on the outside of graphite or molybdenum anodes in 3-500Z and other tubes. It is the dull gray powdery or grainy texture coating you see on the surface of 3-500Z, 811A, and 572B anodes.

Zirconium getters best at about 1000 degrees C, this is why large metal anode transmitting tubes like the 4-400A, 4-1000A, and 3-500Z must be operated with a dull red to red anode color. Zirconium also releases some gasses and absorbs other gasses at various temperatures. The varying temperature across the length of the anode (and as the anode heats and cools) allows the gettering agent to absorb a wide variety of gasses.

The quickest way to ruin a 3-500Z, or other glass power grid tube, is to never show anode color over a prolonged period of time! Storing a 3-500Z for many years without operation almost guarantees a flash-over will occur at the first application of high peak anode voltage.

Gas, either from poor manufacturing processing, out-gassing from elements, or seal leakage is the primary failure mode of glass tubes. Secondary to high-vacuum (gas) arcs are problems like bad welds or grid or cathode materials that fall out of alignment due to thermal or mechanical stresses.

External Anodes

We commonly associate external anodes with expensive tubes having fragile low-dissipation grids, long warm-up time oxide cathodes, and poorer reliability of tubes like the 3CX1500A7/8877, 8874, or 3CPX5000A7.

External anode tubes, just like *internal* anode tubes, can use any type of grid and cathode structure.

3CX1200's, 3CX3000's, and 3CX10,000's are popular tubes using thoriated-tungsten emitters. They have very long life, nearly instant warm-up, and very rugged grids.

External anode construction provides three primary advantages, all of which center around size reduction:

- More dissipation (power) can be handled in a smaller package
- Compact size allows better operation at higher frequencies
- Gain is higher, because electrons are more focused and better controlled by the electrostatic field of grids

Only oxide-cathode tubes enjoy the last two advantages. Oxide cathodes provide a very compact high-current emitter. Normally cathode oxide is deposited in bands or rings, and grid wires are aligned directly over the gaps in oxide bands. The 8877, for example, has fairly large diameter cathode (over 1"). More than 100 concentric bands of oxide are deposited on the cathode, and a grid wire is aligned just outside the area of each emission band.

This construction allows the control grid's electrostatic field to control emission, yet keep grid wires away from the primary cathode-to-anode electron streams. The grid is placed very close to the cathode (thousandth's of an inch), while the anode is much further away.

The result is very low grid current (grid intercept) from wire placement out of the electron stream, and very high gain from the very high ratio of electric fields (from distant anode and very close grid) reaching the cathode electron cloud.

Unfortunately the very same things that create very high gain also cause manufacturing and potential operating problems. The close spacing of cathode and grid increases the chance of grid-cathode shorts. Critical placement of grid wires outside of the many cathode bands makes the grid susceptible to alignment problems. If every grid wire, out of hundreds of wires, is not perfectly aligned outside the electron stream electrons will impact one or more wires and cause gold to migrate off the grid. The mechanism is much like water evaporation, rather than a catastrophic event like boiling or melting. The gold will slowly evaporate and redeposit elsewhere in the tube, either causing arcs, shorts, or poisoning of cathode oxide. This is the most common failure of metal-

oxide cathode tubes.

Ceramic tubes with thoriated tungsten electron emitters have much longer life and higher reliability than any other type of power grid tube, as a general rule. They combine the best of both worlds, having the low seal leakage and low out-gassing typical of ceramic external anode tubes and the rugged wider-spaced high dissipation grids.

The only disadvantage of thoriated-tungsten ceramic tubes is gain and frequency response is generally less than provided by oxide-cathode tubes.

The most common failures in thoriated-tungsten ceramic tubes relate to old age, and loss of emission. Ceramic tubes have very good shelf life, unlike large glass tubes.

Grids

Grid control the movement of electrons inside the tube through electric fields surrounding the grid wires. The electric field extends beyond the many wires making up the grid, almost completely dominating the high electric field created by the anode.

The anode and grid normally operate at cooler temperatures than those where robust thermionic emission of electrons occur. With little or no thermionic emission, the electric potential difference (strong electric field gradient) between the anode and grid(s) results in minimal current flow (electron movement) between anode and grid. Most of the current that does flow between anode and grid occurs because of stay gas molecules creating ions or electrons dislodged from the anode by kinetic energy of cathode-to-anode electron flow as electrons impact the anode.

Elevated grid temperatures, primarily caused by kinetic energy of cathode electrons striking the grid, may eventually reach magnitudes where the grid actually starts to show thermionic emission. Many people assume the grid dissipate rating of a tube is a value set by grid failure, but that is incorrect. The dissipation rating is actually a power related value, indicating the power required to heat the grid enough to start thermionic emission.

Eimac determines rated grid dissipation of a tube by increasing long-term grid dissipation while periodically checking at very short intervals for thermionic emission from the grid.

In actuality, tungsten grids in tubes like 3-500Z's can be operated at temperatures where they show color without permanent damage. While tube operation at such temperatures is compromised, the grid and tube often suffers no permanent damage.

The same is not true for gold-plated grids, like those in metal-oxide cathode tubes. Grids plated with soft poorly-bound materials like gold suffer gradual long-term deterioration even with

moderately low levels of grid current. Heat the grid of a metal-oxide cathode tube to temperatures even well below incandescence, and the result is instant irreversible failure.

This is why ALL metal-oxide cathode tubes should have a fast-acting electronic grid protection systems, and why fuses and worse yet resistors intended as fuses offer no protection at all to grids in metal-oxide cathode amplifiers.

Just as anode dissipation can not be determined by simply multiplying Ip times Ep in a working amplifier, control grid dissipation can not be determined by simply multiplying voltage by current in a sub-class 2 amplifier (i.e. AB2). Beware of any article or author who tells you RF grid voltage can simply be multiplied by average grid current to determine grid dissipation.

Grid dissipation, like anode dissipation, is a time-integrated function of instantaneous dissipation throughout the RF cycle. Computer models have made the difficult task of calculating grid current simple and accurate. Short of that you need to do an actual complex analysis of the grid system, such as in a Chaffee analysis.

In grid driven stages, assuming total power applied at the grid is accurately known, grid dissipation is determined by deducting bias power from RF grid power. This is very similar to the short-form determination of anode dissipation from anode power and RF power output. This does not easily work in cathode-driven PA's, because drive power is mixed with output power.

Grids are normally supported only at one or two places, and use very small wires. Normally the base material forming the grid is selected to be very hard, and to have minimal movement over wide temperature changes. The normal base metal in high power tubes is tungsten, just as used in filaments. Metal-oxide tubes require a gold overlay, to prevent contamination of materials.

Grid failures occur from four primary causes:

- Poor alignment of the grid causes some areas to "hot spot" without overall current being high
- Excessive grid current in metal-oxide tubes causes gold to migrate
- Gold plating is not always bonded properly, resulting in sputtering or flaking of gold
- Poor welds or material problems result in grids warping or moving into the cathode

Non tube-manufacturer related grid failures, like anode failures, are generally the result of a series of somewhat lengthy abuse spread over long periods of time.

Using Old Tubes

Tubes that have "set up" for a while often collect gas. This gas either comes from slow leakage through tube seals or outgassing of tube elements. This is the number one problem with old vacuum

tubes. The section below describes how to remove gas in old tubes.

Gettering and Arcing

Gettering is very important, since even miniscule amounts of gas will cause a low-resistance arc from anode to grid or cathode. The normal results of such arcs are blown grid chokes, collapsed anode chokes, damaged meter shunts, and other problems cause by high fault currents. While a few people blame high fault currents on parasitics, it is actually impossible for a parasitic to create such arcs. All the anode and grid can do is deplete the electron cloud from the area of the cathode, and the available current even with a parasitic is limited by the available emission. Uncontrolled arcs are always the result of gas or element alignment in the tube, rather than excessive current from oscillations.

Anode systems should have series-resistance to limit peak current in the event of a tube arc or failure. That series resistance should always be in the anode lead between the filter capacitors and RF choke. A diode clamp should be installed to protect meters, especially the grid meter since the grid is in the normal path of any internal tube arc.

Ceramic tubes, because of low anode operating temperatures, have the gettering agent applied to the cathode or filament assembly. This is the only area inside the tube that heats enough to activate most common gettering materials. **Ceramic tubes without internal flaws or broken seals can generally be gettered by running the filament at rated voltage for an extended period of time before application of any high voltage. The normal time for gettering is between one hour and one full day. If the tube does not getter within a day it is most likely never going to be restored to a operational relatively pure vacuum.**

Glass internal anode tubes generally have the gettering material coated on anodes, which must be operated a high temperatures to activate the getter. Glass tubes have a propensity for seal leakage and element out-gassing, both of which lead to a short self life for large tubes.

Under some conditions a glass tube can be restored to operation by running low anode voltages and positive bias on the grid. This will sometimes allow full operating anode temperatures to be reached, and the tube can be "cooked" for several hours. I've had about a 50% success rate restoring old 3-500Z's that have sat for years without use, and that arced when started at full voltage, by cooking them this way.

Most important, an arc by itself will break down and getter gas inside the tube. This is why an amplifier with a gassy tube will sometimes operate without problem after a sudden tube arc.

Intentional arcing and overheating, while pumping down a tube, are often a normal part of tube manufacturing processes.

Filaments

Filament life in vacuum tubes, like filament life in light bulbs is a complex function of operating hours, hot and cold cycles (thermal shock), mechanical shock, and temperature.

Other than avoiding physical shocks to the tube, or cycling the filament off and on needlessly, temperature is the only variable we can control. We should be sure the filament is operated at or below the maximum recommended voltage.

In amateur service, filaments and heaters rarely (if ever) are operated enough hours to have operating-time related failures. Most failures are mechanical failures, more related to constant on-and-off cycling of the tube rather than wear.

Mechanical issues such as material or assembly defects can combine with close element spacings inside the tube, and cause failures. With spacings in thousands of inches, even the slightest change in physical shape or positioning filaments or cathodes can create a grid-to-filament (or cathode) shorts. Material quality and manufacturing techniques by tube manufacturers are critical to tube life, and sometimes mistakes are made.

Some claims are made that oscillations can cause filaments to bend, but there isn't anything that even remotely supports such claims.

For example, the normal peak filament current of a sine-wave powered 3-500Z filament is about 1.414 times 15-amperes, or 21 amperes. The total force on the filament helice is 11 grams distributed over a 7 cm area for this 21-ampere current in the typical helical structural dimensions of the filament.

The small additional force of a fully saturated emission condition (which requires around 1000 volts of positive grid voltage) results in only 30% increase in total bending force, hardly significant in any filament robust enough to last years in normal operation. Not only can the peak current not reach fatal levels though any type of emission, the grid-cathode potential is not able to reach levels required to saturate filaments.

Why Does The Tube Go Bang?

In order for any tube under any condition to draw more than the peak saturated current, operating potential has to exceed the breakdown voltage of the tube.

There are two reasons excessive current might flow between elements in a tube:

- The peak anode supply voltage exceeds the hold-off voltage of the tube, causing a good tube

to arc
- The tube may be defective either through gas or an incorrectly located anode or grid

Either of these conditions would allow an internal arc, with fault current limited mostly by external circuit resistances. Oscillations, like normal drive power, can only take a healthy tube up to the saturated emission limits. Even then it would take hundreds or thousands of grid-cathode volts to saturate the emission, an impossible condition unless someone drives a 3-500Z mistakenly with a 4CX5000 or some other large PA!

There is a special condition where tubes can fail, and that is where excessively high voltage components are used in tank systems. Under load fault conditions, when nothing absorbs energy supplied to the tank and tank components do not saturate or breakdown, the tube peak anode voltage can reach levels where an arc occurs in a healthy tube. See the load faults and tuning section, and the practical demonstration areas of this website.

Summary

Tubes are relatively robust components, but they are one of the primary points of failure in any electronic device that uses tubes. While many of us would like to believe we can make small circuit changes to prolong tube life, the fact is most tube failures are related to material or construction problems within the tube. In amateur service, most failures relate to manufacturing problems in the tube.

Out-gassing and seal leakage are particularly troublesome in glass tubes, since porous anodes and imperfect glass-to-metal seals allow miniscule amounts of gas to enter the vacuum.

To insure maximum life in Amateur Service using thoriated tungsten tubes:

- Keep seals below rated temperature
- Avoid long periods of excessive anode or grid dissipation
- Avoid extended periods of non-operation
- Avoid filament voltage OVER the manufacturer's rating
- Do not excessively cycle the filament
- Do not subject the tube to shock or vibration
- Use some form of reliable fault-current limiting in the anode

To insure maximum life in Amateur Service using metal-oxide cathode tubes:

- Keep seals below rated temperature
- Avoid long periods of excessive anode dissipation
- Avoid even very short periods of excessive grid dissipation by including a fast electronic

grid-trip circuit

- Avoid filament voltage OVER or UNDER the manufacturer's rating
- Do not excessively cycle the filament
- Always allow full warm-up before allowing any cathode current to flow
- Never apply excessive high voltage, as it may strip or poison the cathode
- Do not subject the tube to shock or vibration
- Use some form of reliable fault-current limiting in the anode

Vacuum Tube Amplifiers

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How the Tube Converts DC to RF Power

A typical vacuum tube radio-frequency amplifier has a high voltage power source. This power source supplies the energy for the RF output power. The vacuum tube acts like a non-linear variable resistance in series with a diode. The electrical equivalent of the anode and output system is:
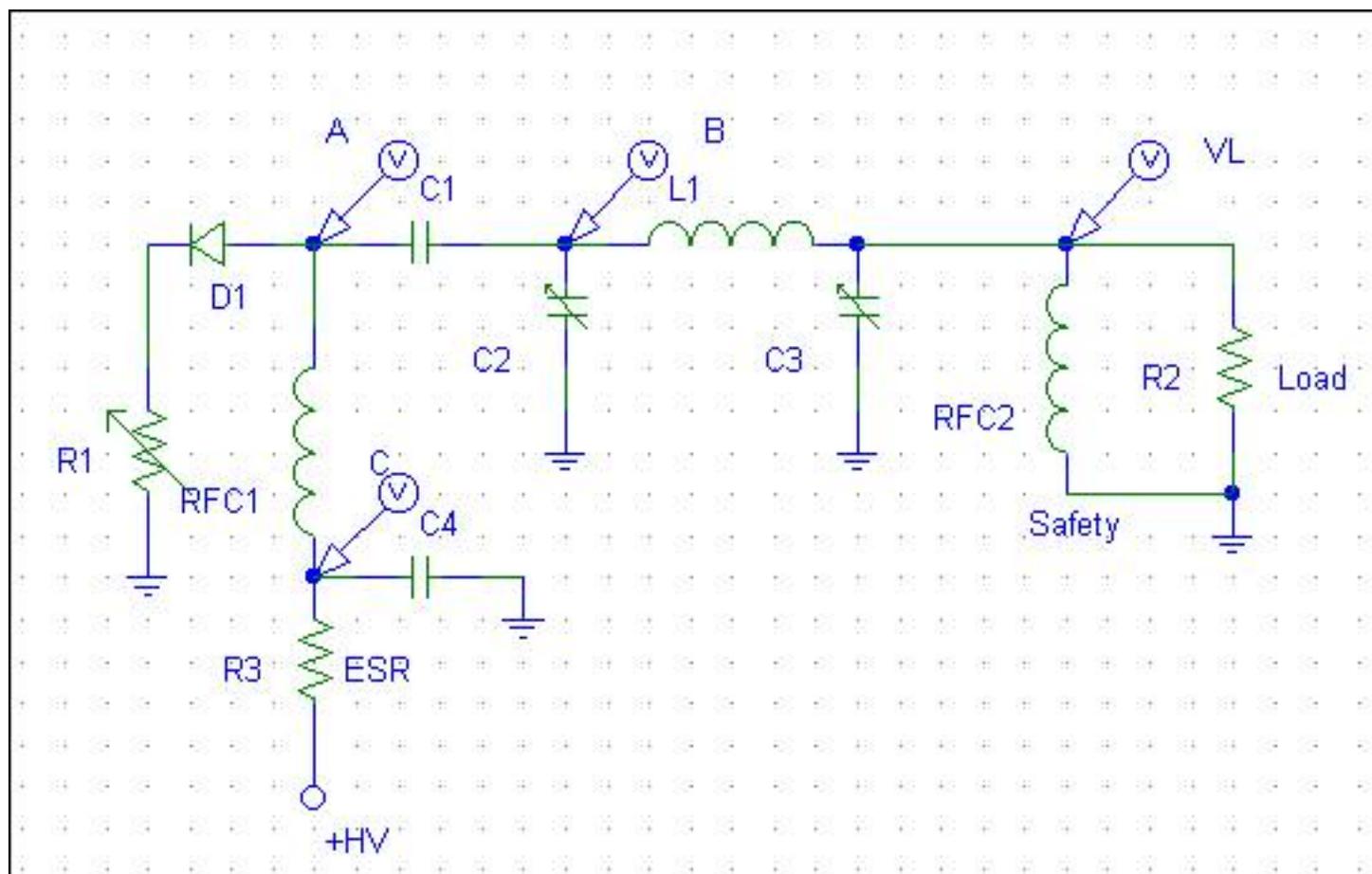
R1 represents the time-varying anode to cathode resistance, as grid-cathode and anode-cathode voltage changes, R1's value changes.

 D1 represents the directional characteristics of the tube anode-to-cathode path. It is always "on" when the voltage at "A" is more positive than ground. D1 would also behave as a current limiter, the limit being somewhat less than the saturated emission current of the cathode.

Let's examine how this circuit converts DC to RF power. We will assume +HV is 3000 volts.

In the initial state, C1 has been charged through the following DC path:
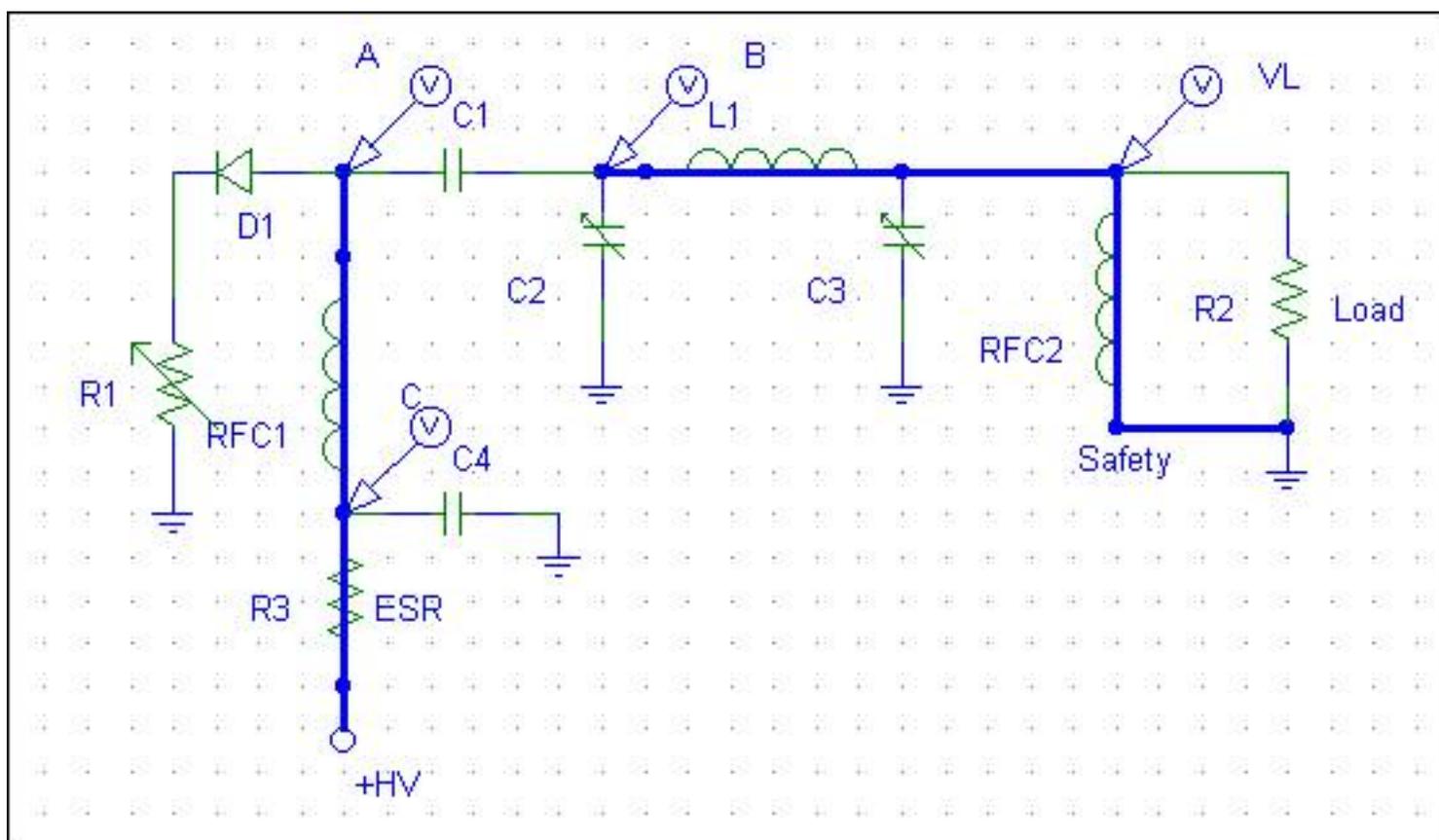
Hopefully the designer was smart enough to include RFC2, to prevent charging or discharging C1 into the load, which might be a sensitive piece of gear or the operator!

The voltage across C1 is the anode-to-ground voltage at A. Everything past A, at B and VL, is zero

volts in the resting state because of the low-dc resistance of RFC2 and the nearly infinite resistance of C1. RFC1 establishes a fixed magnetic field that is set by the quiescent current through R1.

As a positive-going grid-to-cathode voltage perturbs R1, the magnetic flux in RFC1 tries to hold supply current steady. RFC1 does this by increasing the voltage across its terminals. Energy stored in the electric field of C1 tries to hold the voltage steady, and it sources any additional current the tube (R1) requires.

The anode, in effect, "tugs" or pulls point A towards ground. The initial peak current can be quite high, because C1 has a low-reactance path to ground through C2. L1 also builds up a magnetic field, as C2 develops voltage from the current flowing through C2 and C1 back to the tube. On the initial RF cycle, the voltages hardly change before the grid is moving back negative.

At a time interval equal to 1/(4F) (where F is the RF frequency), the positive grid voltage has peaked. The grid begins to swing negative (or less positive) with respect to the cathode. Anode-cathode resistance (R1) decreases.

The collapsing field in L1 (and to a much lesser extent RFC1) tries to hold current the same, and in doing so the collapsing fields supply a small amount of additional anode voltage. Since there is no forced upward swing (the tube cannot source anything), the tank components are left to pull the small voltage change back up to the original anode voltage and slightly beyond. L1 (and to a lesser extent RFC1) actually pull the anode voltage higher by an amount nearly equal to the amount the than the tube pulled the voltage down. ***The actual amount the tank pulls this voltage higher than +HV depends on the amount of energy transferred to the load compared to energy stored in the tank system!*** This is VERY important in understanding where arcs come from!

This cycle repeats over and over, and assuming the tank system at C2, L1, and C3 is resonant, the voltage at point A increases in swing while the peak current through R1 (the tube) is gradually reduced. Eventually***, in a fully loaded amplifier,*** equilibrium is reached. In a ***properly tuned amplifier*** running at maximum available power, equilibrium occurs when point "A" swings up and down an amount just under twice the anode DC voltage.

Consider a class AB 3-500Z using a 3000-volt supply. When the amplifier is tuned into the load properly at full rated power, and driven to full power, anode voltage will swing between 5500 volts maximum and 500 volts minimum at point A. This would be a ***total anode voltage swing of 5000 volts***. Peak anode current in continuous carrier operation would typically be 1.12 amperes, with a minimum R1 value of 446 ohms. This is typical of continuous CW power out of 750 watts with a single 3-500Z tube with a 190-degree conduction angle running at 400mA anode current.

Let's assume (we actually CAN have this condition, if the amplifier and exciter relays are not sequenced correctly) we have an envelope that rises instantly. At the peak of first positive grid-cathode cycle, the anode resistance would drop to someplace well under 400 ohms (anode

resistance is non-linear with anode voltage, and is lower with higher anode voltages). The saturated anode current would easily reach 7.5 amperes, if tube emission permits.

(Emission in a directly heated thoriated-tungsten tube is typically in the range of 50-100mA per watt of heater power, large transmitting tubes being at the upper end of that range. A full-emission 3-500Z has a saturated emission current of about 7.5 amperes.)

If we have infinitely fast envelope rise and fall times from the exciter, amplifier RF-envelope rise and fall times are determined by the operating Q of the tank system. C2/L1/C3 dominate the high frequency energy storage. RFC1 and C1, being larger values, dominate lower-frequency energy storage, while stored energy in the power supply dominates long-term energy demands. The voltage across C2, upon initial application of RF drive, is a rapidly expanding sine wave. It reaches maximum steady-state swing many dozens or hundreds of RF cycles later than the initial tube excitation (and decaying in a similar fashion).

The peak voltage across C2 is equal to the peak anode voltage swing, and is slightly less than twice the HV supply voltage in NORMAL operation. The voltage across C3 is a function of the load resistance, and power delivered to the load.

Incorrect Loading or Load

We know the tank circuit stores energy. We now understand the conversion process where DC is converted to AC (or RF) power. We also must understand *energy must be transferred out of the tank at a rate equal to or exceeding the rate at which it is supplied by the downward "pull" of the tube. If we do not remove energy at a sufficient rate, voltages and currents increase until a new point of equilibrium is reached.* Voltage at point "A" can actually swing well beyond twice +HV on upward excursions, and below zero volts (becoming negative) on downward excursions.

The maximum voltage with a load or drive fault can be tremendously higher than the typical working voltage of the tank system when energy is being removed at the proper rate.

We have the same basic tank system as discussed earlier, but *with light loading compared to drive level* the tank is pulled down very hard by the tube. Minimum Rp is reached early in the cycle, before the tank voltage reaches it's minimum swing. This is easy to do, since the tube only pulls the system down and the tank stores the energy of that downward tug.

As the plate voltage swings below zero (negative) from the tank energy, the tube is already cutting off. Nothing clamps or prevents point A from going negative. D1 effectively takes the tube out of the circuit. By the time the tank reaches it's minimum, the tube's grid-cathode voltage is already on its way positive. The tank free wheels positive, and can overshoot the +HV supply by several times the supply voltage. If loading is light enough and Q is high enough, this continues until the energy stored in the tank reaches equilibrium with energy transferred to the load, or a component fails and the arc dampens the tanks gyrations.

See the practical demonstration page.

VHF Stability

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rev PM Feb23/03

HF Amplifier Stability at VHF

A great deal of empirical engineering surrounds stabilizing amplifiers. While other forms of troubleshooting and engineering follow logical steps, it seems very few builders actually follow planned steps to test for amplifier stability. It is difficult to find detailed information describing suppressors and how they function.

Cause of VHF Oscillations

As in most planned oscillators, the most sensitive control element in the tube generally has the largest influence in determining oscillation frequency. The normal mode of VHF oscillation in HF PA's is at a frequency where the tube becomes a tuned-plate tuned grid oscillator. The control grid to anode path generally has the highest possible gain in the amplifier system, and that is why this part of the system is (by far) the most problematic area of the amplifier system.

The control grid system behaves like it is connected to a parallel tuned circuit. The stray capacitance is primarily between the grid element inside the tube and ground, generally via the filament and other connections. The inductance is via the grid leads inside the tube through the socket to the actual chassis connection. At some frequency, the grid capacitance will parallel-resonate the total inductance in the grid to ground path.
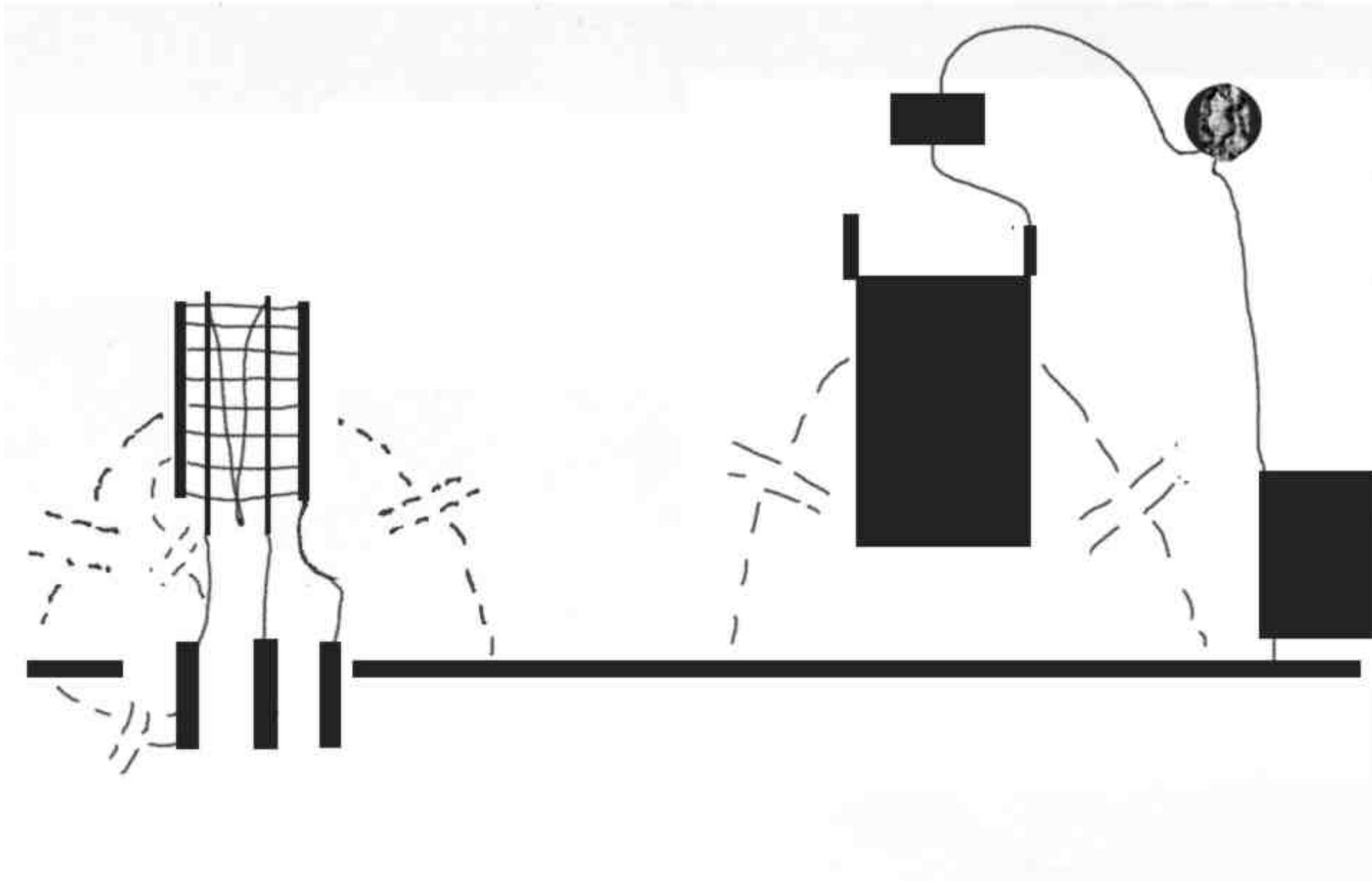
The anode is the second most common problem area. If the anode also has stray capacitance to ground. The path from the anode to chassis has inductance. At some frequency, anode capacitance parallel tunes the path to ground. This resonance greatly increases anode impedance at some very high frequency.

The grid also has capacitance to the anode, and this is the feedback path.

With all of this, the circuit has everything needed to become a tuned-plate tuned-grid oscillator.

If feedback loss (attenuation) from anode-to-grid is less than tube gain at some problem frequency, the tube *may* oscillate. The final requirement is the phase of unwanted feedback must be a value that causes regeneration, or positive feedback. These requirements are the same in any oscillator.

Once again, the conditions required for instability are:

- Gain must exceed attenuation in the feedback path
- The grid must have a sufficiently high impedance for the amount of available feedback to cause a stability problem,
- The anode or other element involved in the oscillation process must have a sufficiently high impedance at the same frequency as the grid to cause a stability problem
- Feedback phase must be within the correct range to obtain *positive* feedback

If any one of these four requirements are not met, the tube will not oscillate! This is true no matter how high or low Q is in any individual path, or if the tube has suppressors or not.

Claims have been made that tubes will remain stable for years, and a "sudden event" will make the tube break into an uncontrolled oscillation. That absolutely can not happen, unless one or more of the four important system parameters above change. Once one or more parameters change, the tube will oscillate continuously until operating voltages are removed. Quite often, in fact most of the time, oscillations are not damaging. The most common effect of VHF oscillations are creation of spurious signals; *not* bangs, pops, or arcing bandswitches. Bangs and pops are caused by gassy tubes or other problems, while arced bandswitches (if caused by an oscillation) are generally caused by oscillations at or near the desired operating frequency!

Location of Suppressor

Suppressors are normally found in anode systems, even though outer locations will work. A VHF suppressor must be located between the tube element and a low-impedance path to ground at VHF. This is because the suppressor must be able to load or "de-Q" one or more portions of the unwanted oscillator circuit. The actual working circuit causing a VHF oscillation is almost always entirely different than what appears on the actual component-based schematic. The cathode, an element commonly involved in low-frequency instability, is rarely involved in VHF oscillations, other than supplying electrons and stray capacitance to ground.

A VHF oscillation, if it happens to occur in an HF PA, is almost always rooted in the system behaving like a "tuned-plate/tuned-grid" oscillator.

Most of our modern PA's are grounded grid (cathode driven). Cathode driven operation requires one or more grids be directly grounded to the chassis (at least for RF) with the lowest impedance possible. *This is necessary to shield the output from the input, and assure operating frequency stability and purity of emissions.*

Using the anode for suppression generally works best because the grid or grids can remain well-grounded for RF, provided the best operating frequency performance.

The Most Unstable Tubes

The most problematic tubes for VHF oscillation have relatively large elements and long thin leads. Tubes of this type have low gain or are unusable at VHF, because elements in the tube (combined with lead inductances) are actually resonant at VHF.

Leads are a particular problem. The longer and thinner the leads, the less stable a tube becomes. Long thin leads move the self-resonance lower in frequency while increasing element impedance. This allows even a tiny amount of anode-to-grid feedback to cause oscillations.

Tubes most troublesome are 811A's, 572B's, 833's, 4-1000A's, 3CX1200A7's and 3CX1200D7's.

Tubes of moderate instability are 3-500Z, 3-1000Z, and 4-400A's.

Tubes having virtually unconditional VHF stability are the 3CX800, 3CX1200Z7, 3CX1500/8877, 3CX3000, and 3CX5000/3CPX5000/YU-156 series.

Looking at the above tubes, it is the tubes with the thinnest and longest leads that are most troublesome. These also are tubes with the poorest VHF performance when used in amplifiers intended to operate at VHF.

The most troublesome tubes above tend to oscillate in the lower-VHF range, between 30 and 100 MHz. The typical instability frequency of an 811A or 572B is around 80-100MHz, assuming all leads are short.

Moderately stable tubes tend to oscillate at 100-200MHz. 3-500Z's, for example, generally are most unstable from 180-200 MHz.

Anode Circuit Layout

Anode circuit layout can contribute to VHF instability. Long thin leads from the tube anode connector to the chassis at VHF are a problem. Problems can occur when thin (and long) plate blocking capacitor leads, thin and/or long wiring, and poor mounting of the plate tuning capacitor are used. Remember, this is a VHF path also, even if the amplifier only *intentionally* operates on HF.

To maximize stability:

- Use wide anode circuit leads from the tube to the tuning capacitor
- Mount the tuning capacitor directly on the chassis, or on a large metallic groundplane area that is thoroughly bonded to the chassis at many points
- Use a low-inductance plate blocking capacitor
- Keep all leads as short as possible, even if it is at the expense of having wiring "look pretty" with all perfectly aligned 90-degree angles
- Use the chassis as a groundplane and an input/output shield, not a front panel

Grid Circuit Layout

The grid circuit layout is probably the single most important area for insuring a stable design. Long thin leads from the tube grid connector to chassis at VHF are a problem. This can be from thin (and long) grid bypass capacitors, thin and/or needlessly long wiring, and failure to ground grids directly to the chassis by mounting ground lugs directly on the chassis near the grid pins.

To maximize stability:

- Use wide low-inductance grid leads from the tube socket directly to the chassis, Connect the ground leads at the closest possible point, using ground lugs right at the grid pins (rather than using socket mounting screws) for grounding
- Use low-pass Pi-network or parallel tuned networks as input matching circuits
- Mount any swamping or load resistors near the tube, with short leads
- Mount the low-pass or bandpass input matching system near the tube, or use exceptionally low-impedance transmission lines to reach the input matching system
- Keep all grid connections as short as possible, even if it is at the expense of having wiring "look pretty" with all perfectly aligned 90-degree angles

A Bad Grid Idea

One of the very worse things in modern grounded-grid triode PA's is the inane engineering claiming floating grids on capacitors adds useful negative feedback. This is similar to what Collins did in their 811A amplifier, and Japanese manufacturers copied the bad idea into their power amplifiers. Heathkit was also a victim of this engineering gaff.

When I was designing PA's in the late 70's and early 80's, an employee of Eimac and author of many articles and a handbook put considerable pressure on me to float the grids of 3-500Z PA's through small mica capacitors.

The theory presented was pretty simple. He claimed floating grids through small mica capacitors added negative feedback, making the amplifier "work better". The basic idea was that filament-to-grid capacitance formed one part of a capacitive voltage divider, the grid-to-ground capacitors forming the other half of the divider.

The alleged "idea" was this capacitive divider would float the grid partially up from ground, and reduce grid-to-cathode (grid-to-filament) driving voltage. I quickly concluded that no one ever actually measured or calculated feedback over a wide range of operating frequencies and grid currents. Since this is a C1/C2 divider, the sampled feedback should be constant in both amplitude and phase regardless of frequency, power levels, and tuning.

The basic circuit is:
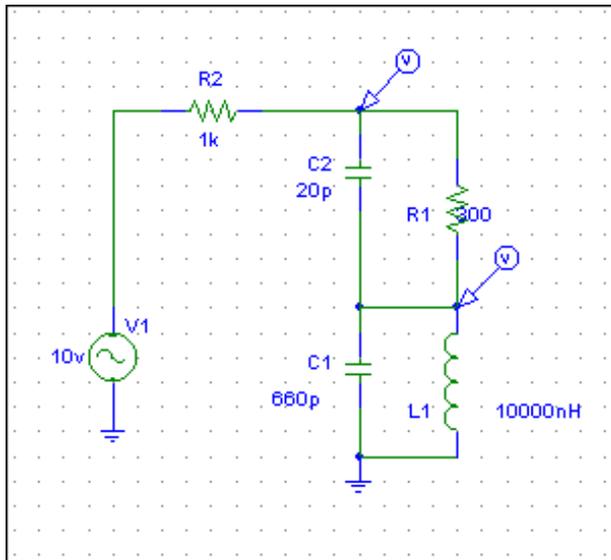
The grid connects at the junction of C1 and C2, while the cathode connects to the top of C2.

C2 is the internal stray GK capacitance of the tube

R1 is the time-varying grid impedance

R2 is added to allow us to see the input impedance change of the divider on a probe model.

Sweeping the system from 100KHz to 30MHz shows us the following:
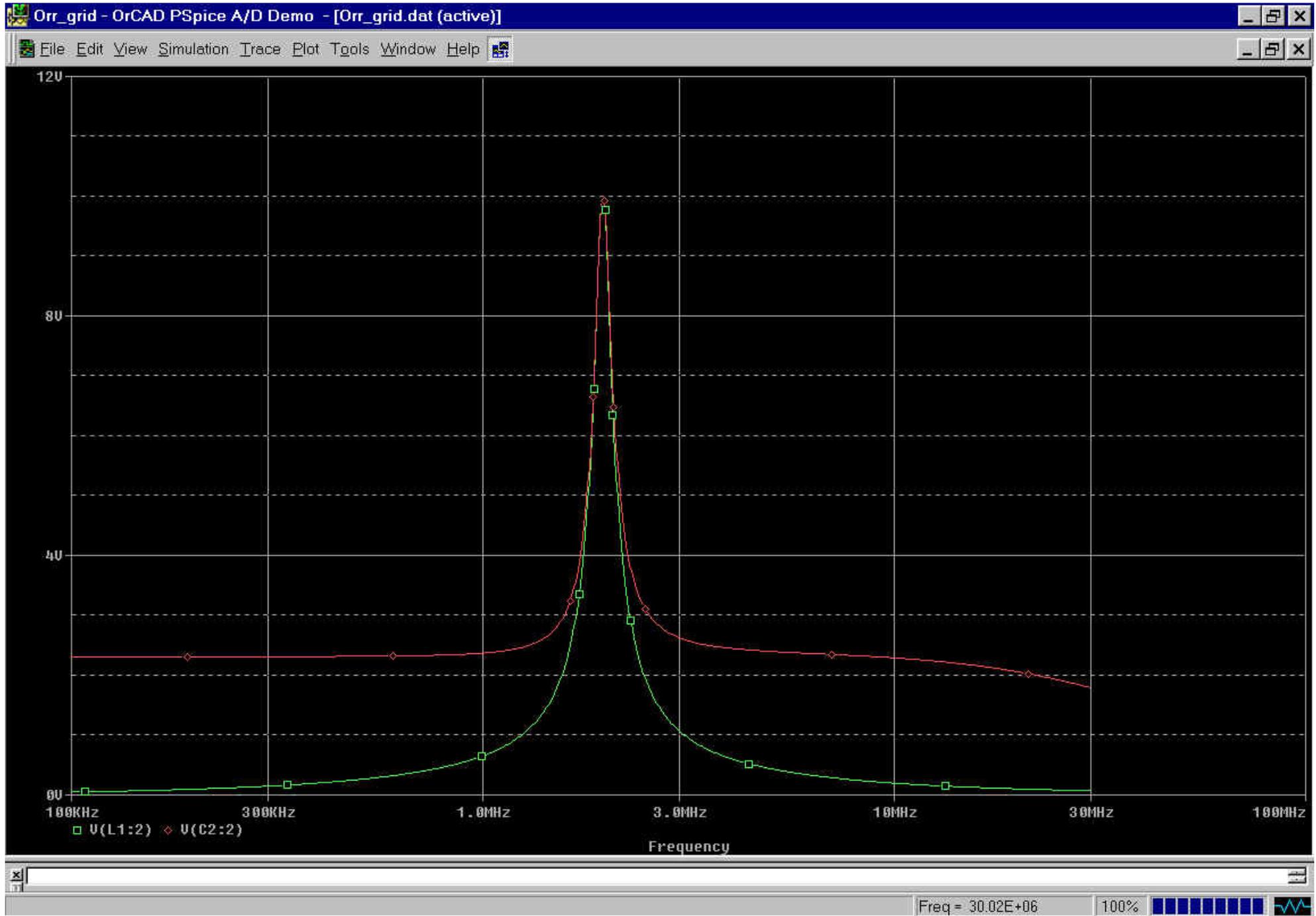
What we see is a huge spike in grid-to-ground impedance at 2MHz, and very uneven response above that range. We did not even include the time-varying grid resistance and phase error,

since this would take up to much space on my web site.

By manipulating the value of L1 (the grid chokes) we can move the spike around, but we are ALWAYS left with some frequency where the grid isn't grounded! This is a serious violation of good engineering practices in a grounded-grid PA, and is actually at the root of stability problems in some PA's. Collins, for example, had a series of field modifications to the 30L1 grid system. The best idea would have been to abandon the bad idea that this system adds controlled negative feedback, and changed to a true grounded grid.

If they wanted negative feedback, the PROPER method would have been adding a resistor in series with the cathode by placing a series resistance immediately at the drive point to the cathode!

There are obviously major flaws with the super-cathode drive concept, when it uses a capacitor divider. Grid current causes grid-to-cathode impedance to constantly vary. When grid current is absent, the grid-to-cathode impedance is nearly an open circuit. Grid-to-cathode capacitance dominates the upper half of the divider, and everything appears to work as planned.

Unfortunately, a problem appears whenever the grid draws current. Even the tiniest amount of grid current causes grid-to-cathode impedance to decreases rapidly. With only a few dozen milliamperes of grid current, grid impedance drops to a few hundred ohms or less. As grid current is drawn, the decreasing grid impedance dominates the upper leg of the voltage division circuit!

There are also new potentially destabilizing resonances added in the grid path.

This system causes three major problems:

- Grid drive is effectively reduced as operating frequency becomes higher, just where we need the gain to flatten normal gain roll-off
- Feedback starts to show significant phase-lag with increased drive and reduced operating frequency
- Grid-to-chassis inductance at VHF and LF is increased, making the amplifier less stable

When I tested several amplifiers with and without this alleged "super-cathode" system, I found IMD performance decreased significantly under some operating conditions. Stability also significantly decreased.

Ground the grids either directly with short heavy leads or a low-inductance high-value capacitor with very *short leads* in any grounded-grid PA!

What Does the Parasitic Suppressor Do?

The parasitic suppressor normally has two components in parallel, a resistor and an inductor. At low frequencies, the path through the inductor dominates the system. At very high frequencies, the resistor dominates the system (assuming it is a low-inductance resistor).

One common problem is people assume brown carbon resistors are non-inductive. That isn't the case. For an example, look at the following resistors:

COMPOSITION
TYPE CARBON

FILM TYPE
CARBON

COMPOSITION TYPE
METAL

FILM-TYPE

Spiral coating of
carbon on
ceramic substrate
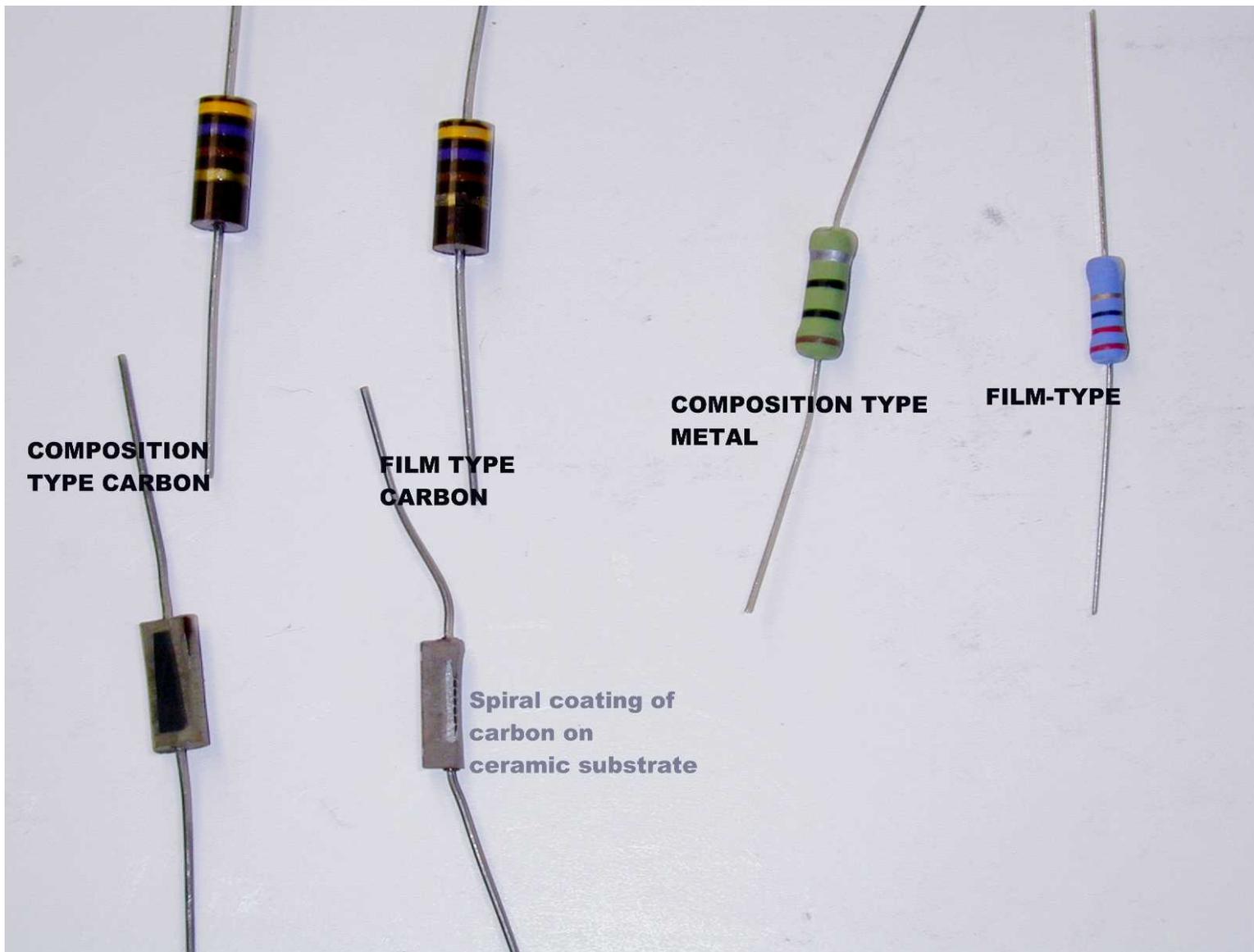
All of the spiral-conductor resistors above have significant inductance at VHF, and make very ineffective suppressors unless the reactance is cancelled. Only the true carbon composition resistors are useful in non-resonant standard suppressors.

This is a typical suppressor system, including inductance of the anode lead:
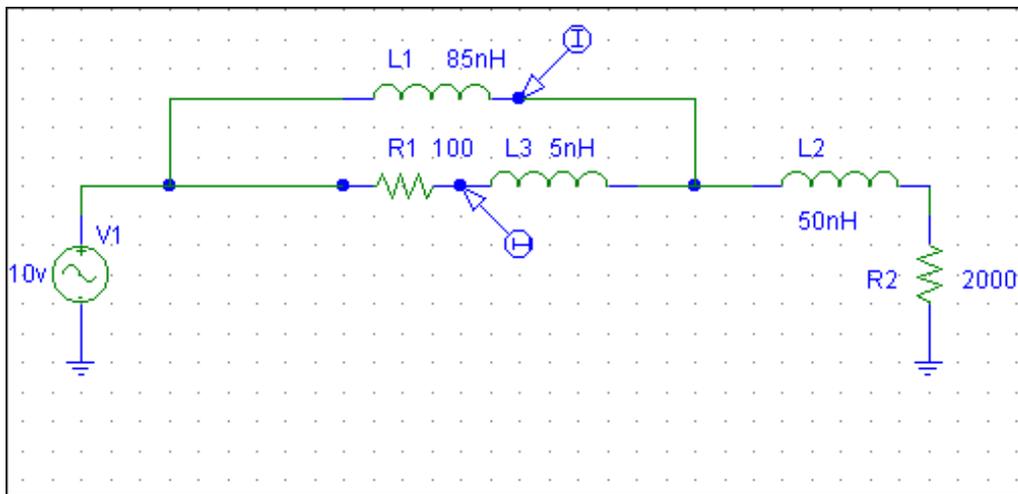
In this case V1 represents the tube. The following is a simulation of currents in the suppressor:

```
         0Hz        50MHz        100MHz        150MHz        200MHz        250MHz        300MHz
          □ -I(L1)  ◇ -I(R1)
                                               Frequency
```

```
×|                                                                                        ⊒
                                          Freq = 300.0E+06    100% ████████ ⩘
```
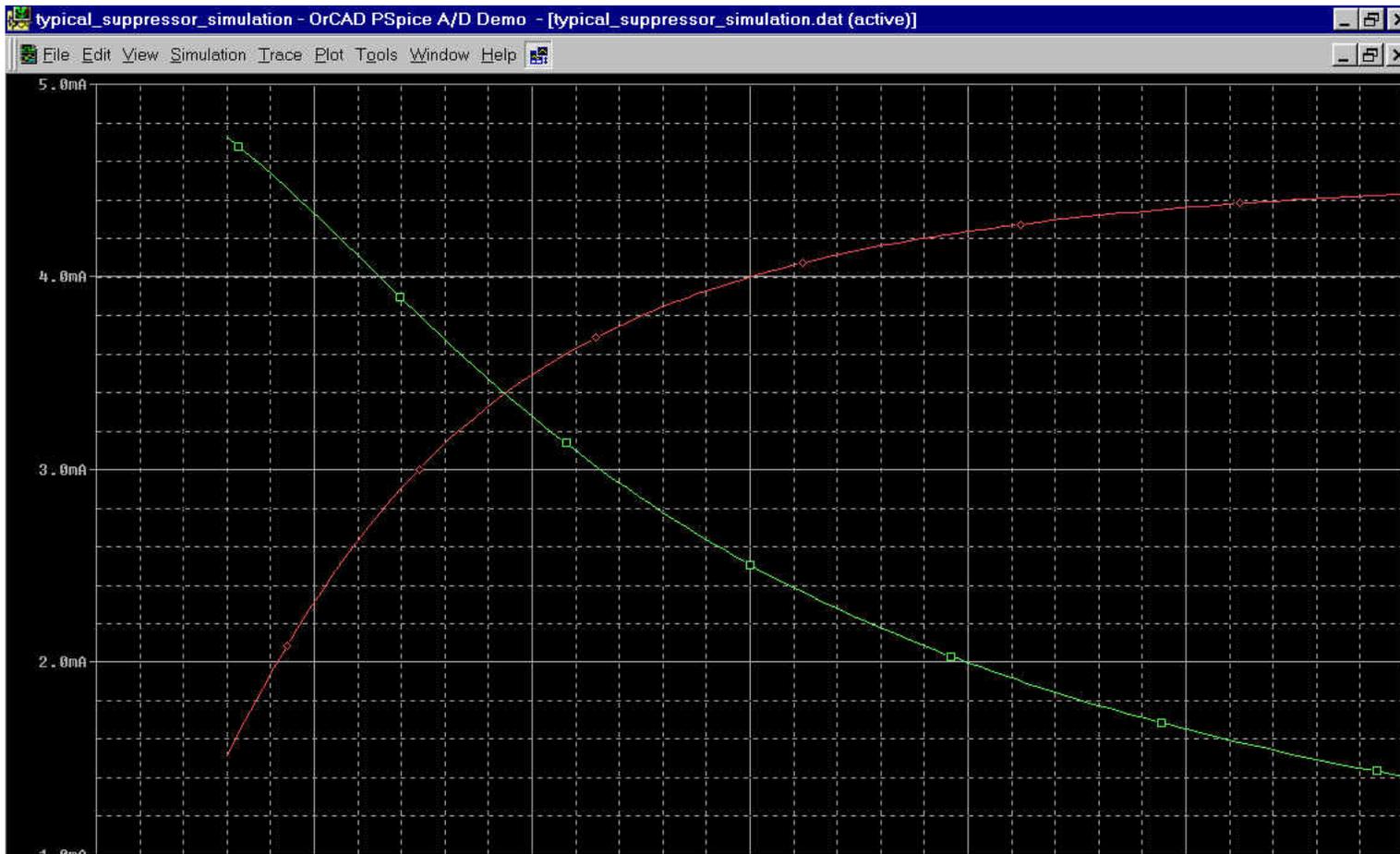
Starting at 30MHz, the ratio of current in the inductor to current in the resistor is:

Frequency	-I(L1)	-I(R1)
30MHz	0.0047	0.0015
60	0.0041	0.0026
90	0.0034	0.0034
120	0.0029	0.0037
160	0.0024	0.0041
190	0.0021	0.0042
220	0.0018	0.0043

This tells us something very important. The INDUCTOR dominates only at low frequencies. At 30MHz, current in the inductor is three times current in the resistor.

At 190MHz, in the range of the instability frequency of a 3-500Z, the resistor has twice the current as the inductor.

This tells us any changes in INDUCTOR design or inductor Q (such as use of nichrome wire) mainly lowers low frequency Q. It would have virtually no effect on very high frequency Q of the system.

- The dominant factor in controlling VHF Q is the resistor value, and any reactance in the resistor path
- The dominate factor in determining HF Q and performance is the inductor value, and any changes in inductor Q

This has been my point all along with the Measure's nichrome suppressor. Measures claims, incorrectly, his suppressors provide lower VHF Q while, in fact, they do exactly the opposite! A typical Measures hairpin suppressor actually produced significantly *higher* system Q in the anode of a 3-500Z (nearly twice the VHF Q), because the equivalent Rp of the suppressor in *series* with the anode lead was *lower*!

The reasons HF PA's arc are explained at other pages of this site, and include incorrect relay sequencing, load faults, as well as improper tuning and exciter transients.

Reducing VHF Q

If we want a lower VHF Q, while maintaining high LF Q and efficiency, the system must shift current into the resistor faster as frequency increases. The suppressor must also have higher Rp, so it dominates the anode path inductance that is in series with the suppressor.

While Measures openly touts his "low-Rp suppressor", the fact is a low Rp suppressor results in higher anode system Q!

A Truly Improved Parasitic Suppressor

In order to reduce VHF Q, we must have a resistance dominate the anode system. This means, in a frequency sweep simulation, the ratio of currents in the resistance to

current in the inductance must be as high as possible. Let's call that slope *the rate of transfer.*

The rate of transfer can be increased by adding a small value of capacitance in series with the resistor:
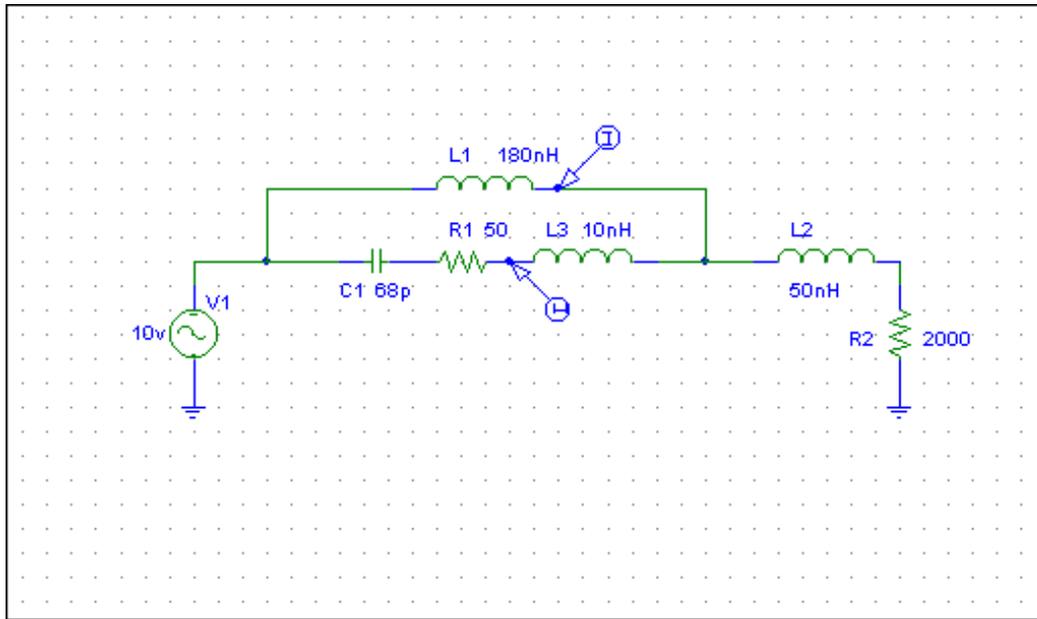
The old suppressor was:

Frequency	-I(L1)	-I(R1)	Ratio
30MHz	0.0047	0.0015	3
60	0.0041	0.0026	1.6
90	0.0034	0.0034	1
120	0.0029	0.0037	.78
160	0.0024	0.0041	.58
190	0.0021	0.0042	.5
220	0.0018	0.0043	.42

The new one:

Frequency	-I(L1)	-I(R1)	Ratio
30MHz	0.0069	0.0026	2.6
60	0.0050	0.0055	.9
90	0.0027	0.0052	.52
120	0.0019	0.0050	.38
160	0.0013	0.0048	.27
190	0.0011	0.0047	.23
220	0.0009	0.0047	.19

Graphically we see the currents are:
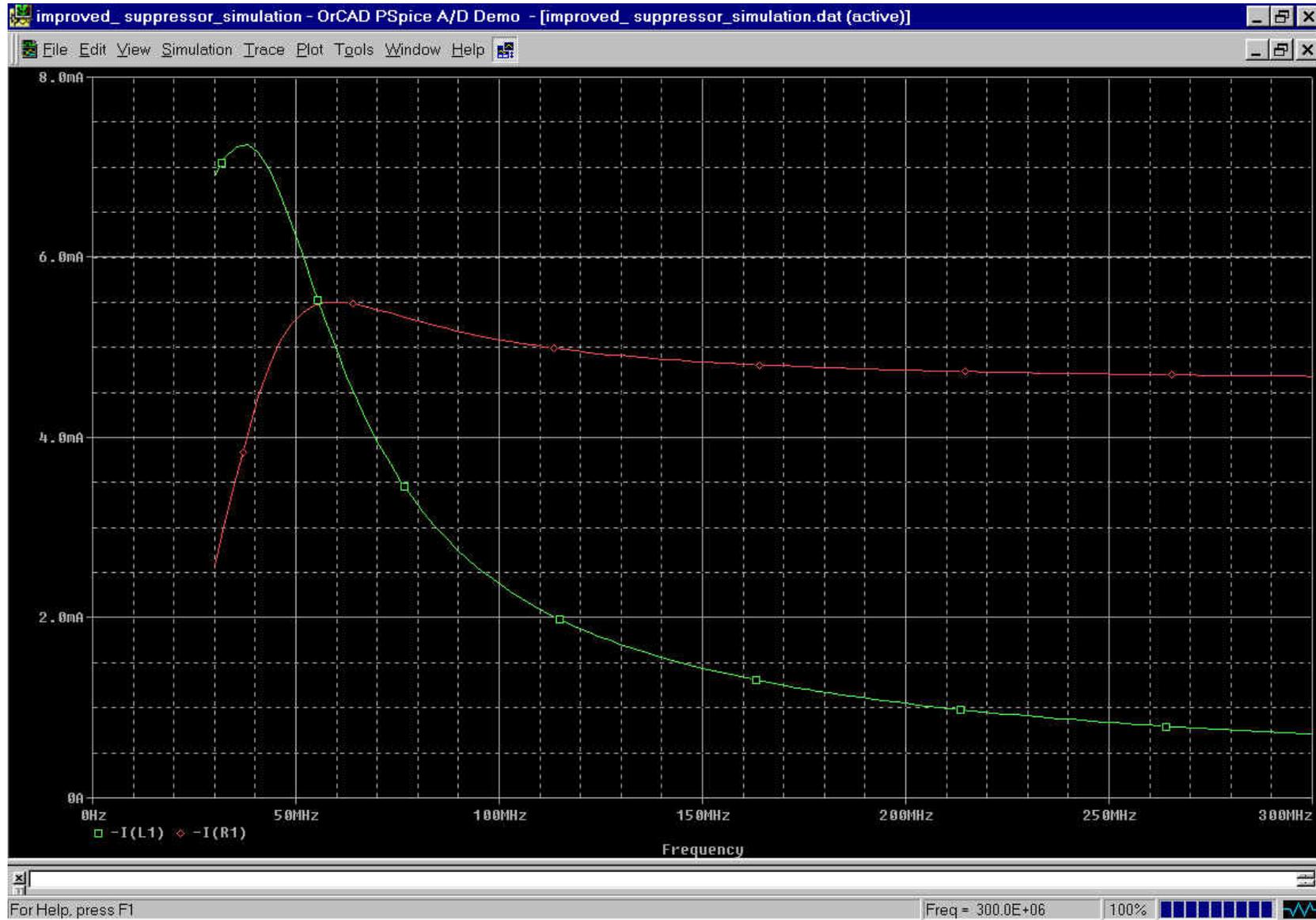
The green curve is current through the inductor, the red curve shows current through the resistor. Notice how flat current is in the resistor, and how sharp roll off of current in the inductor becomes.

This means we will have very low *anode SYSTEM* Q starting at a low VHF frequency of 50-60MHz, and continuing up to UHF. Dissipation in the resistor is still reasonable at HF, efficiency and tank Q at the operating frequency remain high, yet VHF suppression is greatly improved.

Selecting Component Values

Optimum resistor value can actually be determined by measurement, or determine empirically.

If the anode path is long and thin, the impedance will be high. A high anode path impedance (thin or long leads) requires higher values of resistance, because we want the resistor to dominate the anode system impedance. The best value for a resistor is generally one that is approximately equal to, or slightly higher than, the anode path reactance at the frequency of instability.

That impedance can be measured on an impedance test set, or other ways by creative engineers or technicians, but as a general rule long, thin anode leads like 811A's require 100-150 ohms of resistance while shorter thicker anode leads like those in 3-500Z tubes require 50-100 ohms of resistance. Stable tubes with external anodes often can just use anode lead resistance, using brass or other materials, to adequately dampen anode path reactance.

The inductance has to present a significantly higher reactance than the suppression resistor value at the frequency of instability. This causes the majority of current to flow through the resistance at the very high frequency, and not the inductor.

If you look at amplifier designs, you will find tubes like 811A's generally have higher value resistors and many turns of wire in the suppressor. Tubes like 3-500Z's have significantly fewer turns, especially when grid leads are kept very short and direct to the chassis, and lower value resistors.

The more unstable the amplifier tube, the larger the inductor and resistor must be.

One way to view this is to consider the frequency response of a Hi-fi amplifier. Larger values of plate load resistors in amplifier stages reduce higher-frequency gain. The same is true in HF PA's.

Lower frequencies of instability require larger inductors, so the RF path is shifted over to the resistor at a lower frequency.

Uses For Improved Suppressors

Series-resonant suppressors are used with slightly inductive resistor paths, and larger-than- normal shunt inductors. A small capacitor is placed in series with the inductive resistor path, and this capacitor series-tunes the resistor path. This results in a very rapid shift of current into the resistor as frequency is increased. This works well with amplifiers operating at 1/3 to 1/2 the instability frequency, minimizing resistor heat while providing perfect stability.

Typical applications are 3CX1200A7 and D7 tubes, 572B tubes, and 811A tubes.

Shunt suppressors with series-resonant tuning are also sometimes used, the normal application is very high power stages with substantial anode-to-tank currents. These suppressors consist of a series R/L/C system, where the C is normally just stray capacitance to the tube anode. Sometimes these suppressors take the form of a ferrite block placed between the anode and chassis. The inductance of the block series-tunes stray capacitance, and the losses act like a damping resistance in series with that path. I've stabilized 50-100kW VHF transmitter designs using shunt suppression.

Other Instability

Some PA systems are prone to oscillation at low frequencies. Yaesu and Dentron amplifiers using 572B's, and the Collins amplifier using 811A's are good examples of production amplifiers with stability problems.

These amplifiers tend to oscillate NEAR the operating frequency.

All of these amplifiers, except the Yaesu, use tubes with high anode-to-grid feed-through capacitance and no neutralization. Worse, the Collins floats the grids for RF, reducing the already poor isolation of anode-to-cathode feedback path in the 811A.

Yaesu uses one of the poorest engineered feedback systems of all, with a capacitor from the output of the pi section back to the cathode! Phase shift in that path would vary wildly with tank circuit tuning and load impedance on the PA, as would the amount of feedback!

The Yaesu amplifier is a particular problem with Chinese 572B tubes, because grid mu is lower. Negative grid bias has LESS of an effect on cathode current, so the Chinese (and Russian) tubes draw extra quiescent current when the antenna relay is open. This additional current allows the tube to amplify while the amp is in standby. Since the antenna and input source are removed in standby, and the improperly designed feedback path to the tank output remains in place, the PA oscillates near the operating frequency with no load! Voltage in the tank builds up to many thousands of volts, because no energy is extracted to a load. The fact the oscillation is at a low frequency allows the bandswitch to see the full voltage, and it fails.

Amplifiers can create extremely large voltages when RF is applied and a load is not present!

All of the amplifiers discussed above would be greatly improved by:

- Adding a proper bridge neutralization circuit like Heathkit, Ameritron, and Gonset used in 811 amplifiers.
- Grounding the grids either directly or through low reactance very-short-lead capacitors, directly between the socket's grid pin and chassis.
- Using the improved suppressor outlined above to de-Q the amp at lower VHF.

Conclusion

I hope this information is useful, and helps people understand what really goes on in a parasitic suppression system. As time permits, I add more articles about curing unique problems in amplifiers, and diagnosing amplifier failures. I hope these pages are a good start.

Please pass this web address along to others.

COMPOSITION
TYPE CARBON

FILM TYPE
CARBON

COMPOSITION TYPE
METAL

FILM-TYPE

Spiral coating of
carbon on
ceramic substrate
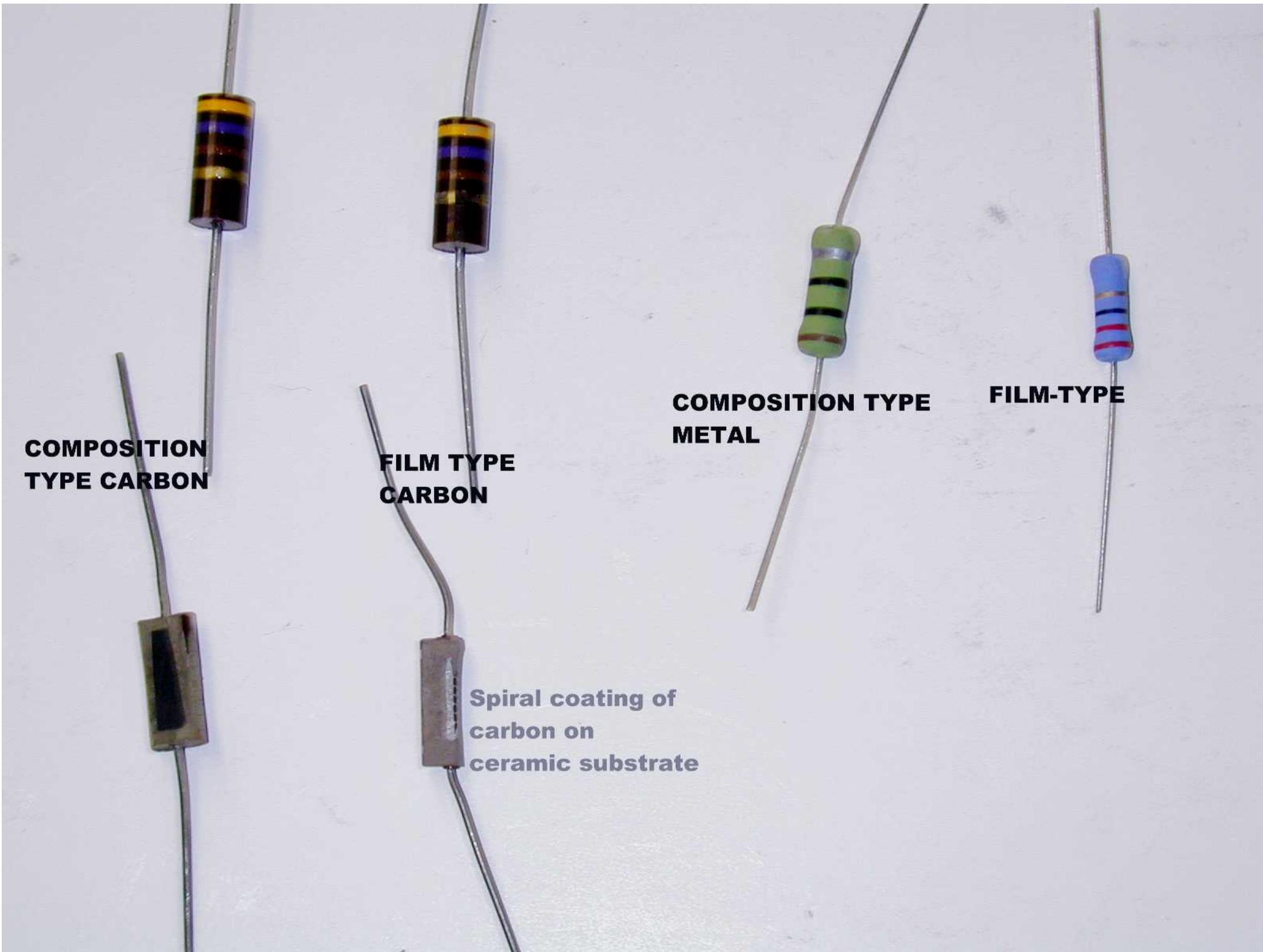
Sorted by IM3

[**Home**] [Up]

Sorted by Intermodulation. Best receiver first.

This is a measure of how badly a close-spaced signal will trash your receiving with spurious signals or splatter sounds, and you might blame it on the other guy thinking your receiver is just great. Remember that even in my very quiet rural location where I have to "beam" through the east coast to hear Europe, 80dB or so IM3 is probably more than enough. The reason my Drakes are so good is because they were built to duplex on the same band with my own transmitters running!! Unless you are in a situation like that, anything over 80dB is likely wasted.

Most transmitters are really the source of problems. The exceptions to transmitters being the primary source of problems occurs when:

1.) You use a noise blanker. Most noise blankers kill receiver IM and blocking performance.

2.) You don't have good narrow IF filters.

3.) You use too much gain in the receiver front end.

4.) You have a receiver with a design problem (less than 80-85dB close-spaced IM3 DR) .

Beyond the above, most transmitters are the real problem. Look at transmitter IM tests on SSB and read about keyclicks.

Poor Designs

Poor designs include:

1.) Many different Yaesu's without noise blanker mods. NO, changing the diodes doesn't do a thing! Been there, done that. The hissing noise comes from a FET following the narrow filters, the IM comes from the silly Yaesu noise-blanker design that hangs an FET right on a wide IF line with the potential for that FET to be at full gain even when the NB is OFF!!!

2.) DSP filter only radios. They aren't that good to date, and some of them are downright crummy.

If you have a DSP radio without a roofing filter that matches the mode, you probably have a ***poor*** receiver.

3.) Radios with spurious response problems. The JRC JST-245's I've tested and used all test good, but they have so many spurious synthesizer responses on 160 meters I'd never consider using one for serious weak signal work. I don't know how they are on other bands, but they sure stink on 160.

Here's how a few receivers stack up. Look at how much the FET in the Yaesu hurts the performance. Look at how poor a stock early R4C really is. Notice almost anything is good at 10kHz, even some very poor receivers.

	Model (in some cases serial included)	MDS	2kHz BDR	10kHz BDR	2kHz IMDR	10kHz IMDR
1	R4C heavy mod. Experimental. See note 5	-143	127	131	118	119
2	R4C 17816 med mod. See note 4	-140	96	115	92	103
3	TenTec Orion	-129	98	100	92	92
4	FT1000D/W8JI mods	-136	73	87	89	88
5	TS-870 Inrad 400Hz IF filter mod (AIP)	-139 (-125)	74 (82)	94 (98)	86 (81)	90 (91)
6	FT1000MK V W8JI mods	-131	80	89	85	90
7	IC-751A	-140	83.5	98	79	91
8	IC-7800	-127	80	98	78	88@5 kHz

					62* 75.5** note 1	70.5* 80** note 1
9	FT1000MP 8N370165	-135	73.5	84		
10	FT1000D	-134	68	85	72	81
11	R4C S/N 22291 stock	-139	80	116	68	86 note3
12	IC-756 PRO	-133	N/A	78	67	77
13	FT1000MK V stock	-131	74	86	60 note1A	77 note 1A
14	R4C S/N 17816 stock	-132	57	109	48	82 note2

Note 1: The FT1000MP has the same inherent problem as the FT1000 and FT1000D. **A noise-blanker amplifier FET is left on** *even when the noise blanker is turned off with front panel NB and NB-W switches*. This FET connects directly to the IF, and overloads when strong signals are present. Unlike the FT1000 and FT1000D, the front panel noise blanker level control affects the FT1000MP IMD performance. * is with the NB gain control fully advanced, ** is with the NB gain control fully off. ___In ALL cases the NB off-on switch is OFF!!___

Note 1A: Yaesu copied the same basic NB from the FT1000D, and so this radio also has a NB problem even when the NB is switched off. This test was with NB gain menu set for minimum gain. The radio is worse when NB gain is increased in the NB gain menu.

Note 2: This is a MOSFET 2nd mixer R4C, aligned and in good shape but with no changes. This receiver is almost useless for weak signal CW work of any kind because of poor IM performance and filter blow-by.

Note 3: This is a tube 2nd mixer R4C, aligned and in good shape but with no changes. This receiver is almost useless for contesting or crowded bands because of poor IM performance and bad filter blow-by.

Note 4: This is the same receiver in note 2, but with the following changes:

1. Increased screen voltage on RF amplifier to 130 volts.
2. 6J6 first mixer with oscillator injection increased to 5 volts.
3. Diode double-balanced 2nd mixer with +10dBm injection.
4. 10dB of gain removed from first mixer and first IF and redistributed behind the narrow filter.
5. Sherwood Engineering 600Hz roofing filter.
6. New AGC system. After narrow filter only, no AGC on front end.

Note 5: Gutted and rebuilt with solid-state double balanced high-level mixers. No vacuum tubes in signal path. Sherwood 600Hz roofing filter. Experimental receiver for full-duplex operation on 160 meters. This receiver allows me to hear weak DX while I am transmitting, as long as they are a few kHz off frequency.

Detuning Towers

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Detuning Towers

The goal of detuning is not to avoid resonance, but rather to minimize re-radiation and/or current in the interfering structure.

When large enough, structures can re-radiate and cause severe pattern distortion even when they are NOT resonant. Consider, for example, how effective a large non-resonant screen or solid disk is at reflecting signals even when it is nowhere near resonance.

We can minimize re-radiation by making an area or areas of the structure "electrically vanish". We often call this "de-tuning", even though it is more correctly electrical trapping or sectionalizing of a structure.

Most structures or towers, when detuned, have a section adjusted to represent a *parallel tuned* circuit.

Section C

Section A
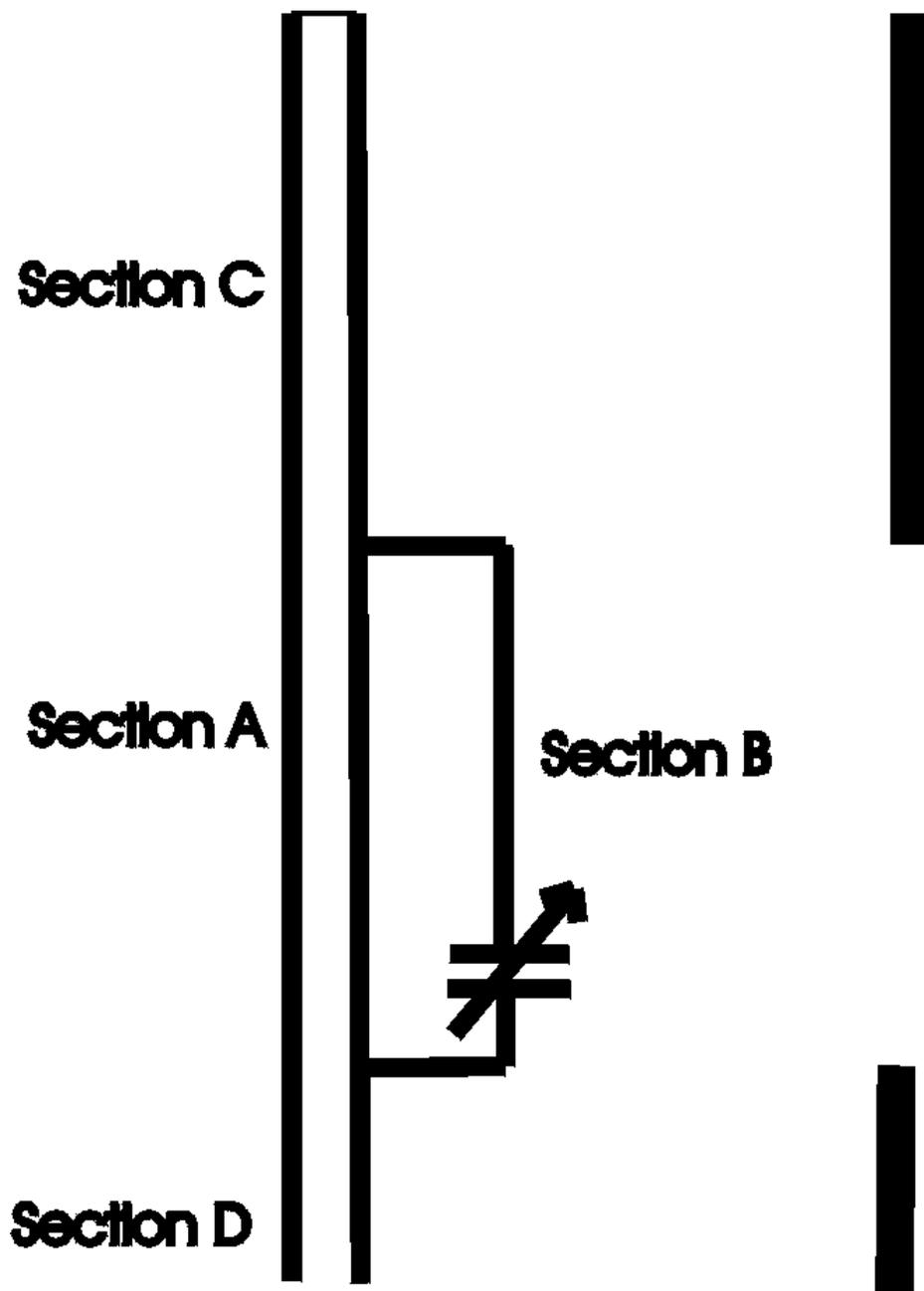
Section B

Section D

Section A and B carry out-of-phase currents. Picture the current flowing upwards in A. It must then flow downwards in B. Since it is a closed loop, these out-of-phase currents are equal and flow in *opposite directions* at resonance. The result is radiation from sections A and B cancel each other. When section A and B are exactly resonant, sections D and C are isolated by a high impedance. The high impedance is caused by or related to the high current though the capacitor and the inductance of section A. When current is maximum, voltage drop is maximum.

This results in the electrical structure on the right, with section A and B removed!!

In effect, we have created a trap much like the trap in a dipole. As in the trap dipole, current is maximum in the trap at the trap's resonant frequency.

NOTE: *The condition of proper tuning occurs with MAXIMUM current in section B, NOT minimum current!! To electrically sectionalize the tower and isolate C and D (and minimize radiation from A) section B must be tuned for MAXIMUM current!*

As either section C or D approach resonance by themselves, the tuning condition will change. This would occur when D is grounded and near 1/4wl or an odd multiple of 1/4wl long, or when C (with whatever is mounted on it) is self-resonant with section A removed!!

Under this condition, you would either need to sectionalize and detune C or D with additional detuning, or move the location of sections A and B to a new point that (when isolated) prevents resonance in C and/or D.

A few general rules apply. Pay attention to these guidelines to insure best results:

- Never parallel-tune a large area. Certainly not an area over 3/16th wavelength long
- The detuning "loop" must have a good solid connection to the structure being detuned. Don't connect the detuning wire out to a separate object or earth stake
- We want to adjust for MAXIMUM current in section B, the exception being when that would cause resonance in C or D
- We can not have any electrically large structures or wires hanging from the tower in the area being detuned.
- Ideally any cables passing the detuning area should be grounded to the tower at the top and bottom of the detuning area, or pass through that area in the center of the tower or mast. At the very least, cable shields should be bonded to the tower at the top and bottom of the cable run and unshielded cables placed inside the tower.
- Tuning is fairly narrow. ~5% total BW is about all that can be expected in most cases, but this varies greatly with the system including distances to the other affected antennas and the amount of pattern distortion tolerated

Note: I added the section about cables on Feb 17,2003. I'm surprised cables are often not grounded at the top and bottom of tall towers, and that unshielded control cables are not passed through the inside of towers. Cables should always be treated that way for lightning protection if for no other reason!

Capacitor Size

The amount of capacitance and the voltage rating of the capacitor is not easy to predict. The size depends on unwanted power levels that excite the detuned structure, the electrical characteristics of the detuned structure, and the Q of the detuning section. Capacitance values will be fairly high with short sections on lower bands like 160, almost certainly in the range of a few thousand pF for ~20 foot long sections. The exact value would depend heavily on dimensions of the A to B loop.

Voltages across the capacitor are generally not high, although they can be at times. The "loop Q" of A and B affects voltage, as does the amount of excitation and load presented by the impedances of C and D.

MFJ sells a clamp-on calibrated current meter that will not perturb the system. It is a cheap version of a current meter I designed. This is a calibrated meter with internal amplifier that measures current from a few mA to 3 amperes, not the uncalibrated RF-sniffer commonly sold. Some RF-sniffers, including those by MFJ, actually change the impedance and resonant frequency of the system because the pick-up transformers are not properly designed and terminated current transformers. Avoid loop-stick type current meters, since they measure ANY external field and can provide misleading results. Use a current meter that is directly inserted in line B, or clamps around line B with the closed core of a terminated current transformer. Use a meter that does NOT perturb the system when removed!

Lacking a current meter, it is possible to tune this system with a grid dip meter, by forming a small one or two turn coupling loop. As an alternative, the loop can can be broken at any point near the capacitor and a MFJ-259 or similar antenna analyzer connected in series. Proper adjustment is at the point were minimum impedance occurs. If that impedance is not low, you probably are not effectively detuning the structure.

Multiple Stacked Antennas or Tall Structures

When multiple stacked antennas are used, especially on a fairly tall tower, it may be necessary to sectionalize multiple points. Individual sections between antennas can be resonant, or appear electrically long.

If the tower or structure or any part of the structure or tower becomes resonant when section A is tuned to present a high impedance, then we need to move section A or tune it to some condition other than maximum current (resonance). Adjustments under this condition can only be made two ways:

- A sampling loop can be mounted on the structure 1/10th wl or more above or below section B and adjusted for minimum terminal voltage
- Field strength of the pattern can be plotted, and the structure tuned for minimum pattern distortion
- Never detune an area that contains large yagis or other electrically large objects, like long conductive guy lines, dipoles, or cables leaving the tower

Omega and Gama Matching

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Omega's and Gamma's

Impedance Limits

The best matching system for feeding a grounded Marconi element, unless we are very lucky or very careful in planning the installation, is a simple gamma capacitor and shunt feed conductor with the tap point adjusted to find 50 ohms resistive.

While losses are *not* significantly increased when using an Omega match with reasonably sized components, matching range is not significantly extended either! At best, the Omega match saves us a few trips up and down the tower while searching for the precise shorting strap position between the gamma wire or cage and the tower. If we are already at the top of a tower with the gamma tap point, and if at that point we find resistance (after reactance is cancelled by adjustment of the gamma) is too high, an Omega will not help. The Omega can only match loads *LESS* than 50 ohms resistive by stepping the resistance up! It can NOT step or transform antenna feed resistance downwards when using capacitors, and neither the Omega or Gamma can match capacitive antenna loads.

By adding an inductor, we can greatly extend the matching range. The extra time, expense, and complexity of adding a high-Q inductor would not offset the slight reduction in effort to simply find the correct shorting point for gamma and omega capacitor systems.

Gamma Match

The gamma match capacitor can only cancel reactance, it can not modify the "real part" (resistance) presented to the feedline. It is the most simple form of matching, and has the lowest operating Q and loss of any system (when it is useable). Adjustment of resistance (real part) requires adjusting the diameter, spacing, or length of the gamma section.

The voltage rating of the capacitor is simple to find. It is Y*I*X where Y is the safety factor and adjustment for peak voltage (RMS*1.414), I is the RMS current from the feedline, and X is the reactance (in ohms) of the capacitor. Y is normally a factor of four to allow for peak conversion, SWR, component flaws, and transients.

The current rating is simply the RMS current plus a safety factor. Component physical size as well as Q control the current rating. See component ratings.
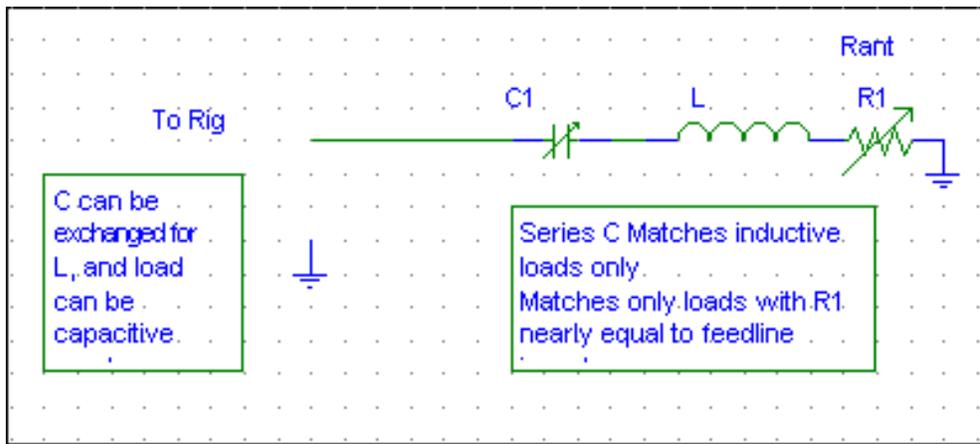
Looking at a Smith Chart, we see a single line (yellow) following a portion of the resistance circle with a resistance value of 1 (1 is normalized to 50 ohms). The gamma match can not deviate from that resistance circle, and so the resistance must be adjusted or transformed by other means. We normally adjust the gamma system's tap point by using a movable shorting strap to change R. Just moving a short is enough, it is not necessary to remove the extra "unused" portion above or beyond the shorting strap.
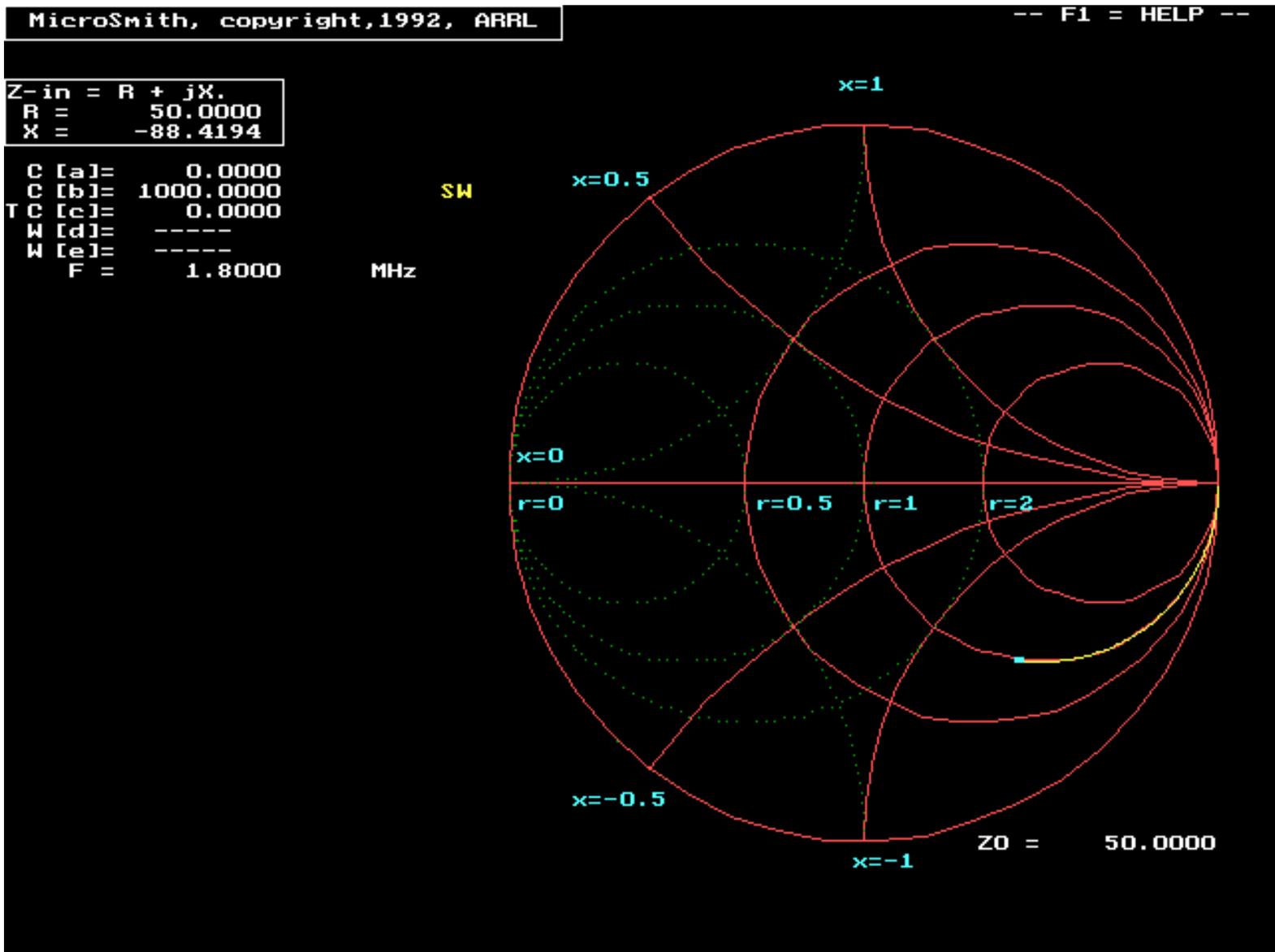
Let's consider a 1500-watt transmitter and 50-ohm transmission line, with a 1:1 VSWR. The matched line current is found using $I := \sqrt{\dfrac{P}{R}}$. The result is 5.5a RMS current.

With 1500 watts and 50 ohms we have 5.5 amperes RMS. Assuming the match requires 300pF at 1.8MHz we have: $Xc := \dfrac{1000000}{2 \cdot \pi \cdot F \cdot C}$ $Xc = 294.731$

4*295* 5.5 = 6490 volts. It's easy to see why arcing is a problem when capacitance is low! Very few air variables would take such a high voltage, and insulators (especially when exposed to moisture) also become a problem. This is typical of a shunt-fed element when the matching section is a thin long wire.

We can decrease reactance by increasing the effective diameter of gamma conductors, and this will increase required capacitance value. One way to decrease reactance is to use a cage of wires spaced as far apart as possible. *Halving the reactance (doubling the capacitance value) will cut peak voltage in half, with no other changes. This has the effect of quadrupling the power rating when using the same breakdown voltage ratings in components!*

A thicker gamma conductor also lowers operating (loaded) Q. This is a series-resonant system, operating Q is set by the amount of reactance in series with the load resistance (50 ohms). Reducing operating Q has the effect of increasing bandwidth while simultaneously reducing loss and increasing power rating.

Note that bandwidth increased while efficiency increased. This happens in many cases. Popular folklore tells us narrow antennas are efficient, but that is true only in a few specific cases. In most cases, bandwidth by itself tells us nothing about system efficiency!

One example of a conundrum with gamma capacitors is found in yagi antennas. If we increase capacitor dielectric thickness by decreasing inner rod size, the voltage across the capacitor can actually increase at a faster rate than insulation thickness increases! The same is true if we increase gamma capacitor length in an effort to restore capacitance after increasing outer conductor diameter. The effect thicker insulation and higher voltage rating, if done incorrectly, can be to make little change or perhaps even *decrease* gamma capacitor power rating!! Always use a large diameter gamma rod, and increase the dielectric thickness ONLY by increasing the outer rod to accommodate dielectric thickness changes. At the same time, the inner rod should be increased to avoid any requirement of making the gamma capacitor longer. Avoid sharp points or edges on the rods, just as you would avoid sharp points in any HV system....especially inside the "capacitor".

The above example of decreased power rating is especially important to Amateurs using coaxial cables as capacitors. Voltage is NOT constant along the length of a long coaxial capacitor. Maximum voltage in the component is always *HIGHER* than the actual voltage across the terminals of the "capacitor", and it is higher than the voltage calculated by the current through the capacitor! Coaxial capacitors or linear stubs used as reactive elements always have significantly lower operating Q, higher power loss, and operate under more electrical stress than a well-designed lumped component. Stubs and linear loading does have the advantage of spreading heat out. You won't notice the heat as much, even though there is a lot more heat energy! Just don't let the smaller temperature rise fool you into thinking the system has less power loss.

Omega Match

The Omega Match is really just a form of the simple L-network.

Traditional Cs/Cp Omega Match
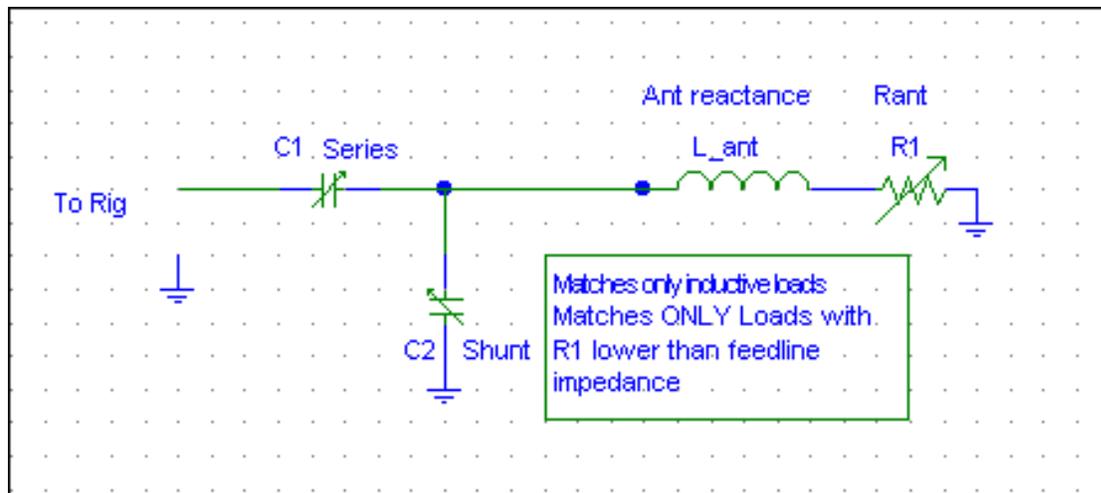
In this system, C2 parallels the shunt terminal and primarily sets the resistance transformation C1 is in series with the feedline, and primarily cancels reactance. If we look at this network on a Smith Chart, we see the resistance range is somewhat expanded over a conventional Gamma Match.
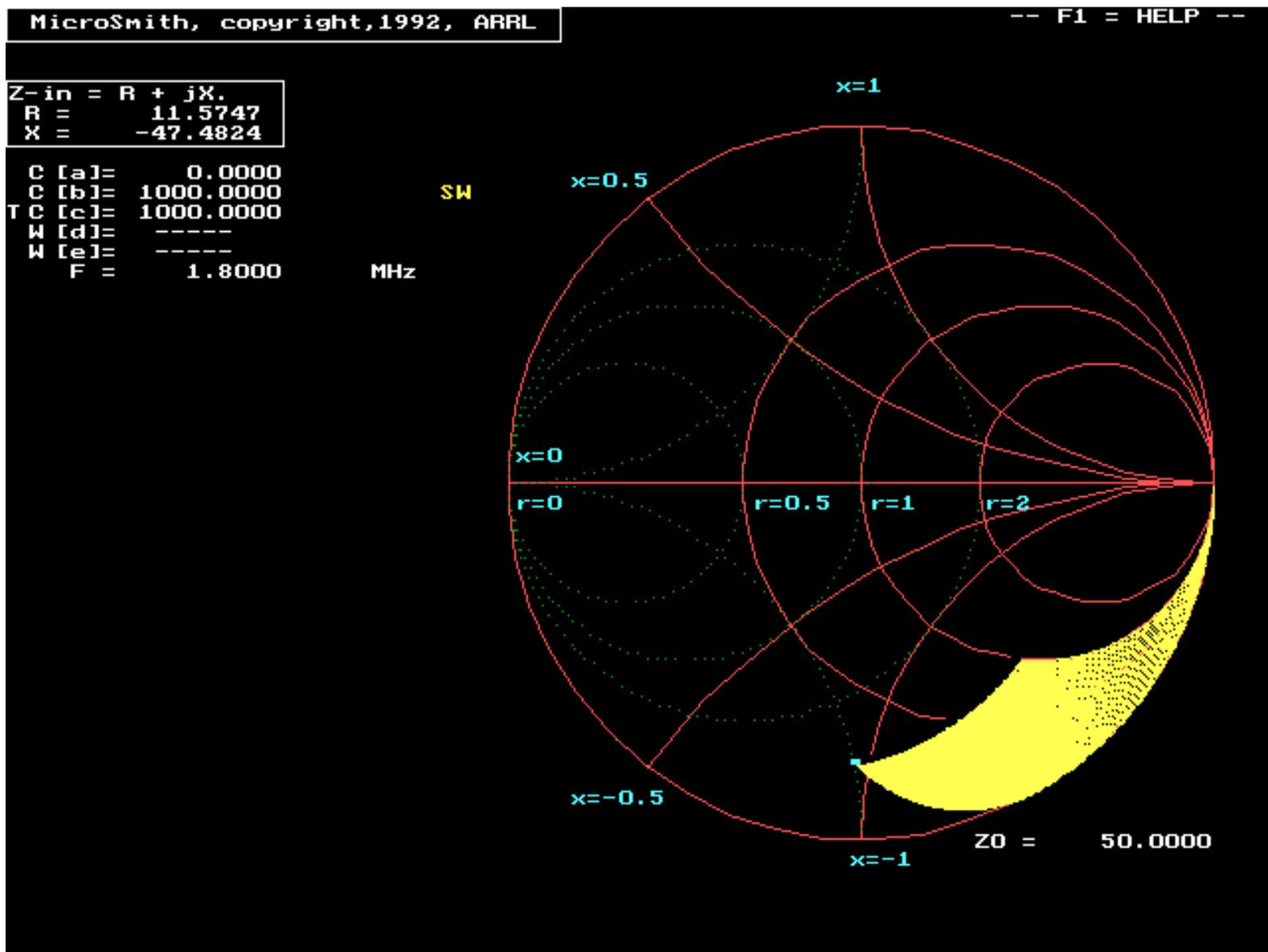
We also can see it is not the panacea often claimed. Look at the "r" circles on the above chart. "r=1" is normalized to 50 ohms. With that in mind "r=2" is 100 ohms, while "r=0.5" becomes 25 ohms. The Omega's resistive matching range, using 1000pF vacuum capacitors, would only from 50 ohms *downwards*. Voltage ratings are still a problem when the system requires capacitor C1 to have a small capacitance value (high reactance), and now we have two capacitors that have to handle essentially the same high voltages when C1's operating capacitance value is small!

This arrangement once again requires antenna shunt wire impedance to be *lower than feedline impedance*. The network, since it only uses capacitors, *must be used with inductive loads* or the feedline terminal will be reactive.

One useful application of this circuit is in matching very short non-resonant vertical element, sometimes called a "Hairpin Monopole". Well-designed capacitors, such as vacuum capacitors, have such high Q they are essentially lossless. With vacuum capacitors and very good connections, losses in the matching system are very low even with very high currents. We shouldn't be misled into thinking we have a "free-lunch" magical short antenna! We "pay for our meal" in greatly increased conductor losses in the antenna conductor. The conductors not only carry common-mode radiation currents, they also must carry significant circulating currents involved with the feed-system. These high currents causes distributed conductor losses to be much higher than conventional

loading systems using a reasonable-Q lumped inductor. Bandwidth and radiation resistance for a given element size is often significantly less than conventional series-feed when using a hairpin monopole with Omega loading and matching, although the system has the advantage of being easily tuned to new frequencies. The Omega/hairpin vertical also allows the feedpoint resistance to be adjusted to nearly any value we might require.

C1 is primarily for power factor (reactance) canceling. It is rated exactly as in the gamma, voltage and current being determined by the required amount of operating reactance and the line current.

C2 primarily sets resistance transformation. The component's electrical requirements are found only by knowing the load impedance and the reactive voltage and current in that branch of the system. It handles a portion of the input current, as well as a portion of the shunt current, so the current in C2 is always higher than the current from the source flowing through C1. The required voltage rating of C2 is also always higher than C1.

Assuming we have 10-1000pF variables, the normal matching range covers inductive reactance ranging from 80 to 4000 ohms with a resistance between 1 and 50 ohms. Lower reactances are only able to be tuned when the resistance is near 50 ohms. As you can see, resistive matching range is wide but limited to values below 50 ohms.

Cp/Cs Omega Match

The parallel-C input, series-C output is also a form of the L-network. The output, once again, must be the lower impedance while the feedline must have the higher impedance. Once again, this system only matches inductive loads.

It is not a very useful circuit, except in specific applications. The only advantage of this arrangement is C2 operates at significantly less voltage. The voltage across C2 is always no greater than the transmission line voltage, larger receiving-type air variables will operate at 1500-watt and higher power levels. Because C2 is under significantly less electrical stress, It is the better choice if you are very close to having a match and only need adjust resistance upwards slightly for perfect match.
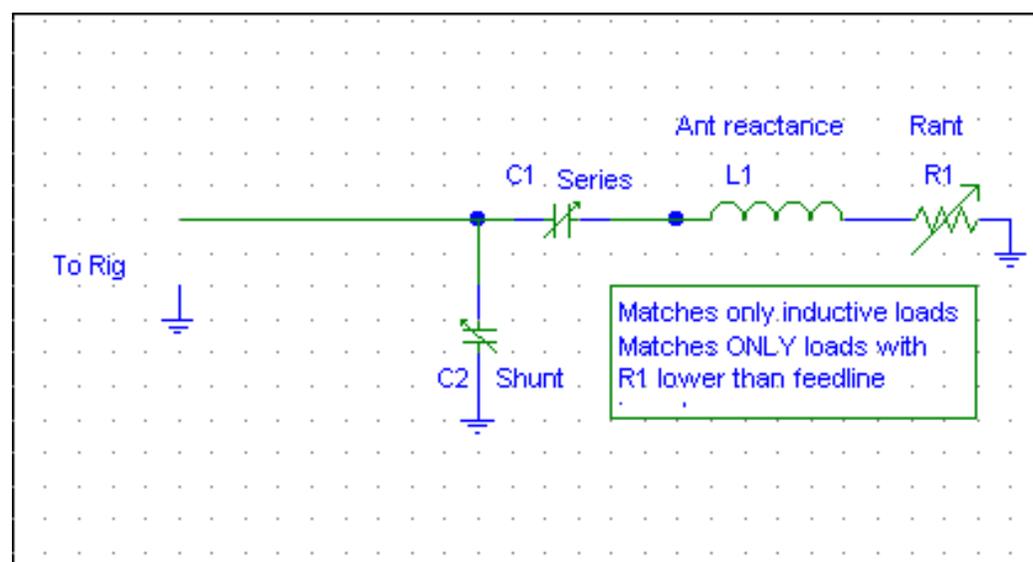
C1 primarily sets reactance, while C2 primarily sets resistance.
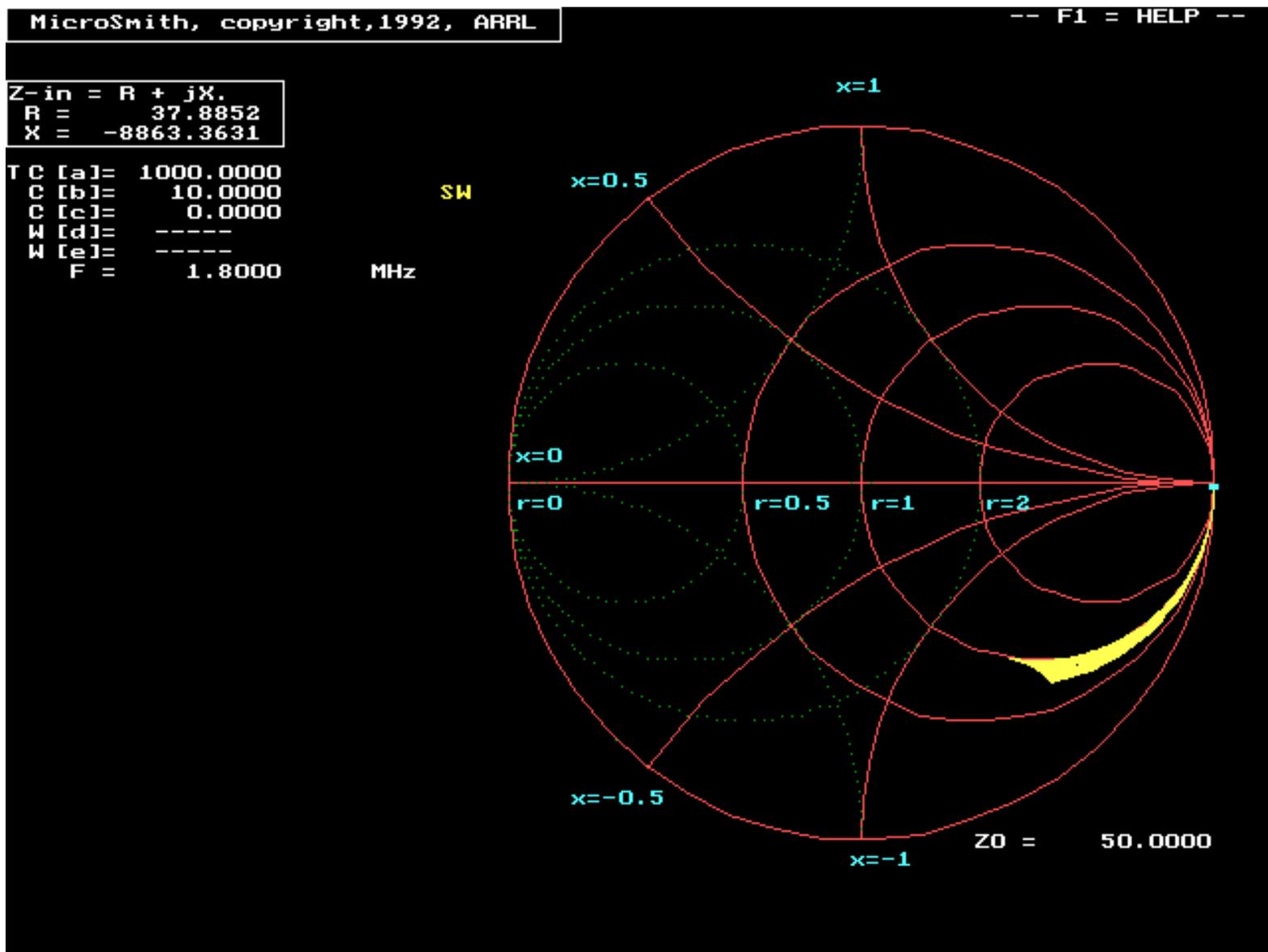
Assuming we have 10-1000pF variables, the normal matching range is reactances up 8000 ohms inductive and resistances between 38 and 50 ohms. As you can see, resistive matching range is very limited. C2 would need to be a very large capacitor (very low reactance) to have a wider resistance range.

Pi Matching

In cases where we can not find a 50-ohm tap point, we can add an inductor. In this case C1 can be a fixed or variable capacitor capable of handling modest voltage and current, such as a snubber-mica or receiving-type air variable.
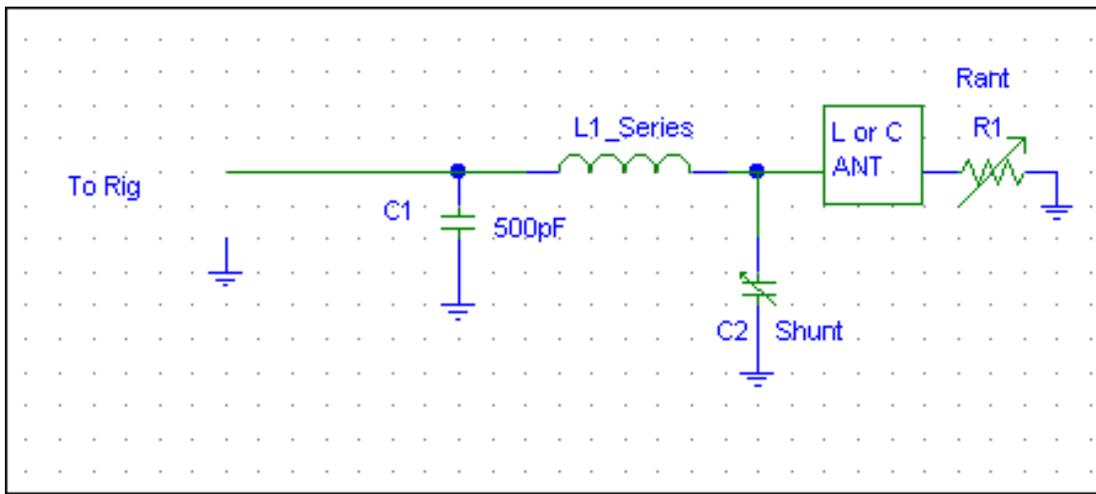
L1 has to be fairly large to match high resistance and reactance ratios, at least 100uH on 160 meters for the extremes shown in the Smith Chart below. Assuming a 10-1000pF capacitor for C2, a 500pF capacitor for C1, and a 100uH adjustable inductor for L1 we would have the following matching range:
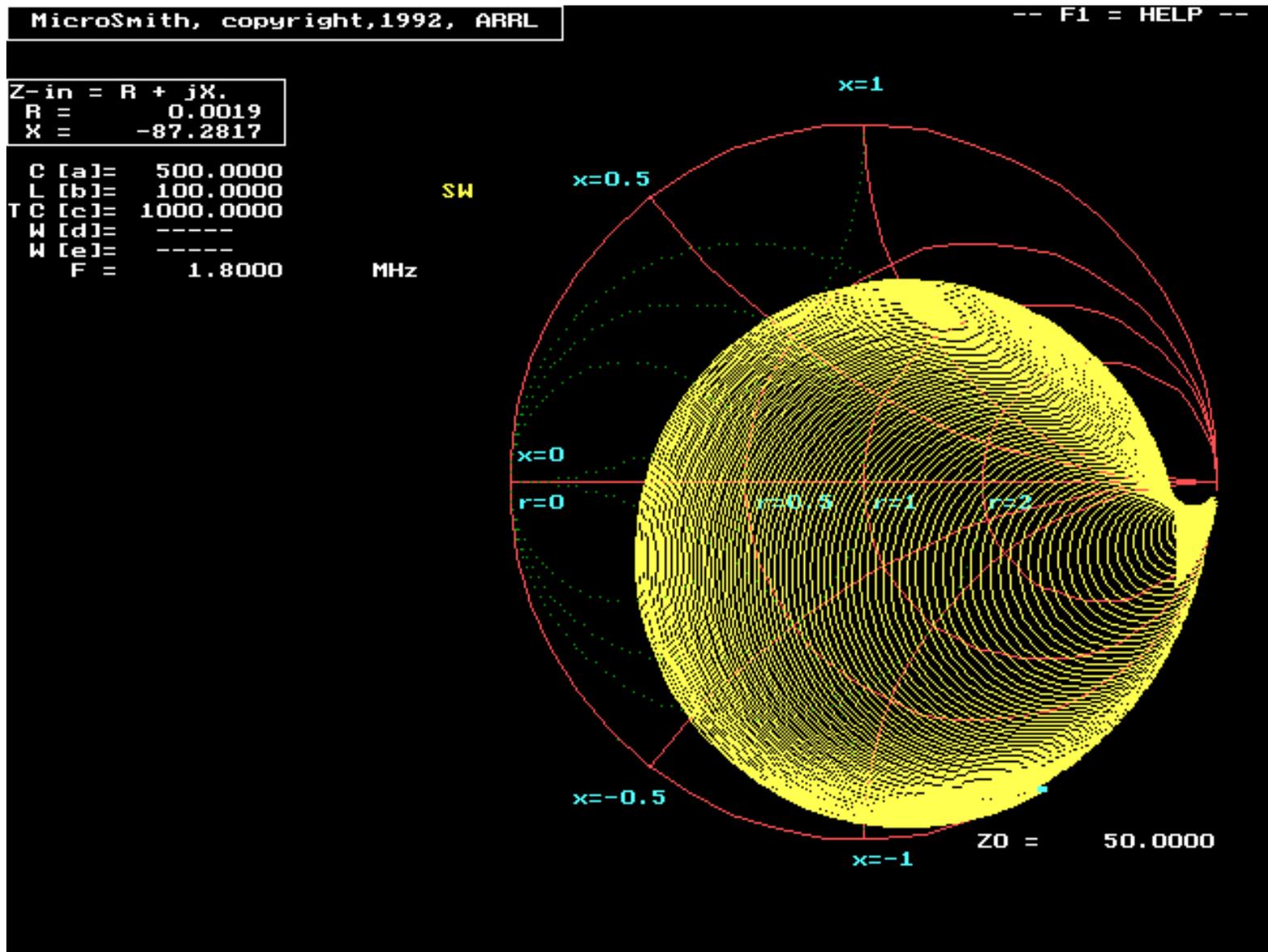
Matching range is greatly extended, and even covers capacitive loads.

Power Ratings

Heating is a long term problem that accumulates with time. Failure occurs when the heat causes a component to physically change from excessive heat. Heating is related to power dissipated by a simple formula $P := I^2 \cdot R$ The resistance is normally determined by knowing the component's Q (not the operating Q) and the current through the component.

The series resistance that dissipates energy as heat is found by dividing reactance by component Q. With 88 ohms and a Q of 5000 (typical of a good large air capacitor) Equivalent Series Resistance is 88/5000 or .0176 ohms. Heat is given by .0176 squared times 3.11 amperes. Heat is .0176*3.11*3.11 or .170 watts, not bad for an air variable capacitor. (Multi-layer chip capacitors can have Q's in the tens or hundreds of thousands, as can vacuum capacitors. Simple ceramic disc and small mica capacitors generally have Q's in the upper hundreds to many thousands.)

Current through C2 is found by dividing 274 volts by R-ant. If R-ant were 15 ohms the current would be 274/15 or 18.27 amperes. Voltage would be C2's reactance of 88 ohms times 18.26 amperes times the safety and peak factor of 4. Peak voltage would be 6500 volts for an 88-ohm reactance capacitor. This is a good example of why "T" antenna tuners arc with low load impedances while working fine at higher impedances, and handle more power when tuned to use maximum possible capacitance.

Arcing is an instantaneous voltage problem. In general a solid dielectric (insulation) punches through and fails almost instantly. Voltage rating is reduced with every arc, even brief unnoticeable arcs. As a general rule solid dielectrics suffer non-recoverable damage from any arc, even a very brief arc (such as a momentary low-current static discharge.

If the dielectric is a vacuum, liquid, or gas, and if the component's conductors or dielectric do not physically distort or change from heating, the insulation can "heal" and full performance is often restored. With an air capacitor, ionized air must circulate out of the capacitor.

If an arc causes a physical change, such as a hole, carbon path, sharpened or raised edge, breakdown voltage is almost always permanently reduced. One exception would be when an arc removes debris or sharp points, such as melting copper whiskers in a vacuum capacitor (copper can actually grow tiny whiskers in a high vacuum). Sharp points, even microscopic sharp points, greatly reduce breakdown voltage. If a sharp point is reduced and rounded by an arc, voltage breakdown will increase. If we actually melt the plate, voltage rating is reduced. Surfaces must be micro-polished, we cannot sand a capacitor plate to restore full breakdown levels if a capacitor is arc-damaged (variable capacitor plates are commonly polished by tumbling them in a very soft abrasive like walnut shells).

Capacitors in parallel with a known resistance. RMS sinewave voltage across the component is given by $E := \sqrt{P \cdot R}$ where P= maximum possible level of PEP applied. R=resistance the capacitor parallels.

Capacitors arc and fail from instantaneous voltage peaks, not from average or RMS voltage. We multiply the

RMS voltage across C1 times 1.414 and add an important safety factor. Multiplying RMS voltage by four normally works for outdoor mounted air-capacitors, as long as they are kept debris and moisture free.

A 50-ohm feedline with 1500 watts of absolute peak power (1500 watts CW or FM carrier or 1500 watts PEP SSB or AM) applied results in 274 volts RMS times 4 or about 1.1kV. Use a 1.1kV or higher voltage variable. This would be about .030" air gap in a typical construction air capacitor with normal manufacturing irregularities (generally very safe to assume .01" spacing for every 300 volts). With 50-ohm matched lines and a 1500-watt PEP (or CW carrier) transmitter, smaller transmitting or larger receiving air variables will work.

Heating

Current through resistance causes heating. The resulting heat over some period of time can cause physical changes that physically alter the component. In the short term, this can change the electrical parameters such as capacitance or voltage breakdown.

To find electrical energy converted to heat, we have to know the equivalent dissipative parallel resistance and voltage across the component or the equivalent series resistance (ESR) and current through the component. I generally work with current and ESR.

The current is a function of reactance and voltage across the capacitor, just like current through a resistance is found from voltage across the value of resistance. Assuming we have 1500 watts and 50 ohms resistive load directly across the capacitor, we have 274 RMS volts across the reactance of C1. This voltage, over the reactance in ohms, gives us the RMS (heating) current flowing through the capacitor.

Use the formula $I = E/X$. With a 1000pF capacitor (88 ohms on 160m) we have 274/88 or about 3.11 amperes.

Note: This is not exact because capacitors have complex series-impedances, but it is close enough for our questions as long as we have reasonably good capacitors. One common exception where this simplification will not work is with long coaxial stubs. Long coaxial stubs have a considerable amount of distributed series inductance. This causes a voltage increase as we move away from the feedpoint in an open stub, or a current increase as we move out on a shorted stub. This increase in voltage or current increases loss, decreases bandwidth, and increases effective loss resistance. Not only that, a stub (or linear loading) has higher distributed resistance because it has long conductors, rather than thick compact conductor area like a conventional component.

Loss and Component Failure

We often gauge system "loss" by the temperature of the component. We often assume a "cool" feeling component has low loss and a hot feeling component has very high loss. Using "touch or feel" temperature without considering size and ability to transfer heat is a good way to estimate component life, but it is not a good way to estimate system efficiency or component losses. Unless we carefully consider the complex issue of physical size and ability to transfer or dissipate heat to surrounding objects or air, we will have no idea how much power is actually lost in the component. Small components not only get hotter, they get hotter much faster than large components!

A physically long coaxial capacitor has a large surface area to dissipate heat. Even a very lossy coaxial capacitor

will have significantly less temperature rise than a small very low loss compact disk capacitor, because the capacitor concentrates all the heat into one very tiny area. We might form the impression we have an inefficient system because the small capacitor drifts in capacitance value or fails from heat, yet it is usually the cool-running stub that has much higher power loss!

This doesn't mean we shouldn't use a coaxial stub as a reactance if they are reliable, but it does mean we have to be careful assuming it is the best solution for loss.

Bandwidth

Bandwidth tells us nothing about efficiency, unless we understand the entire system in detail.

In general the largest bandwidth occurs with the fewest reactive components. Any unnecessary reactance will increase system operating Q. One exception is when multiple networks with low loaded (operating) Q are used to obtain a large phase shift, rather than a few components that would require high operating (loaded) Q.

Stubs act like a combination of many small series-inductances with multiple shunt-capacitors. Coaxial capacitors not only generally have more loss than reasonably-constructed lumped capacitors, they also have less bandwidth. A stub's reactance changes faster with frequency than reactance in a compact lumped component, because it contains "unnecessary" distributed reactances. The stub also has higher internal voltages and currents, because of the distributed reactances. A coaxial stub may be cheap and "feel cold" in operation, but we shouldn't delude ourselves into thinking it is less lossy than a small capacitor. Remember it isn't only power dissipated that determines temperature rise, we must consider the total area heat is distributed over. 1 watt of heat feels very hot in a component the size of a pencil eraser, while 100 watts of heat could be undetectable by touch in a long thick piece of coaxial cable!

Transmitting

160 Meter Transmitting Antennas at W8JI

200-Foot 12-inch face Tower

160-meter vertically polarized transmitting antennas center around this 200-foot tower.
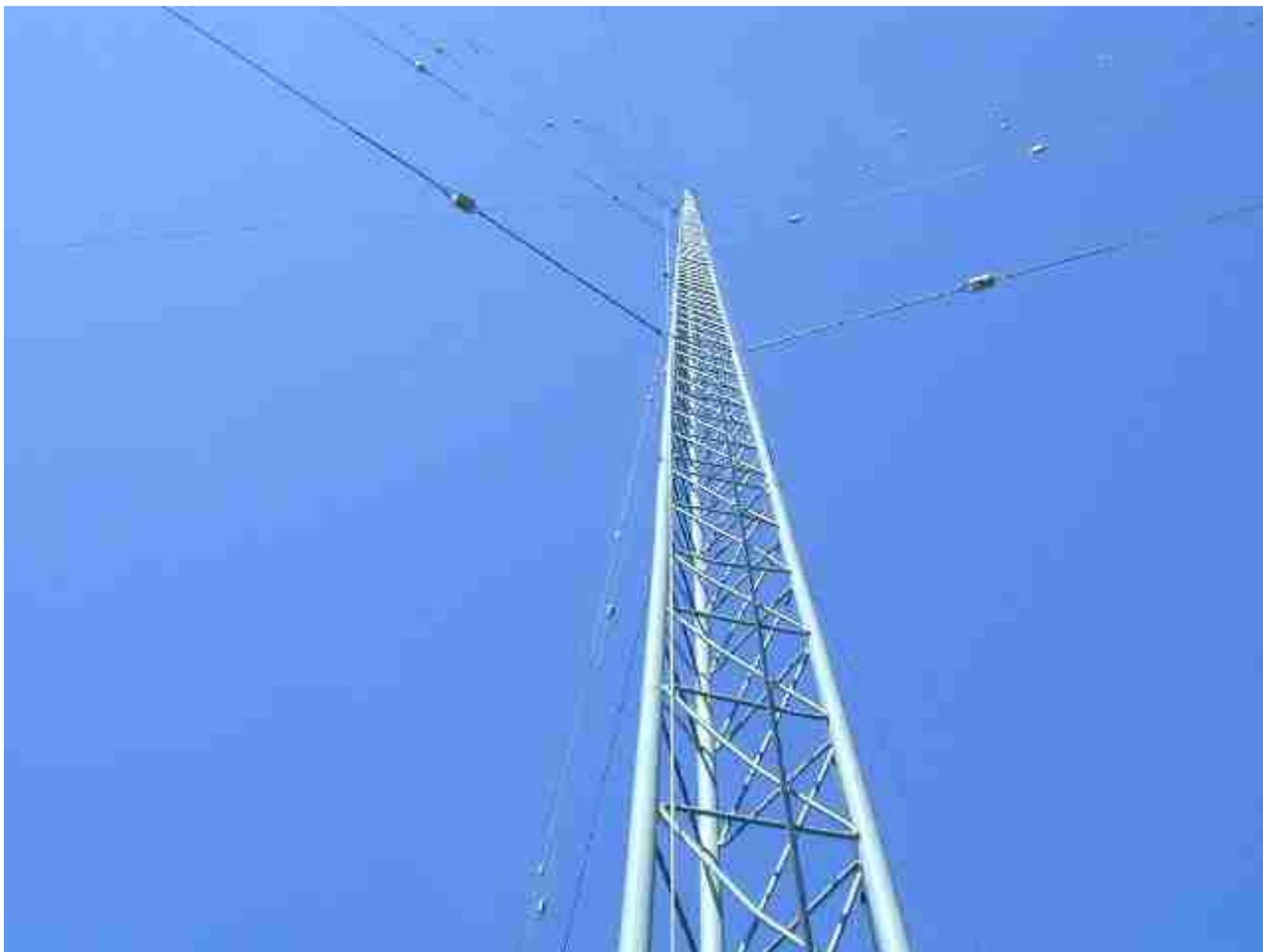
When I moved here in December 1998, this tower was hastily erected as a quick way to get on the air. A single 0.5" diameter 50-ohm feedline and 8-pair control line run to the tower. A large outdoor storage box at tower base holds 160-meter phasing and matching systems.

OMNI Pattern

This 200-foot tower is the main omni-directional antenna. It is series-fed against a ground system of about one-hundred 200-ft radials. The radials are shallow buried 16-gauge bare solid copper wire known as "buss wire". A small low-pass L-network using a vacuum capacitor and variable inductor matches the tower. A 50-foot transmission line coiled around an insulated form between the L-network and RCS-8V antenna relay detunes this tower when the four-square is used. The four-square is detuned when the omni-tower is used.

318-foot 18-inch face Tower

High and Low Dipoles are supported by this 318-foot tower.

Most of my dipoles are supported by this "dipole support tower". This tower is 318-foot tall and sits on a base insulator.

This tower started as an elevated radial test tower and grew from there. Because of the early elevated radial experiments, the 318-foot tower has the remains of 60 1/4-wl buried radials below it.
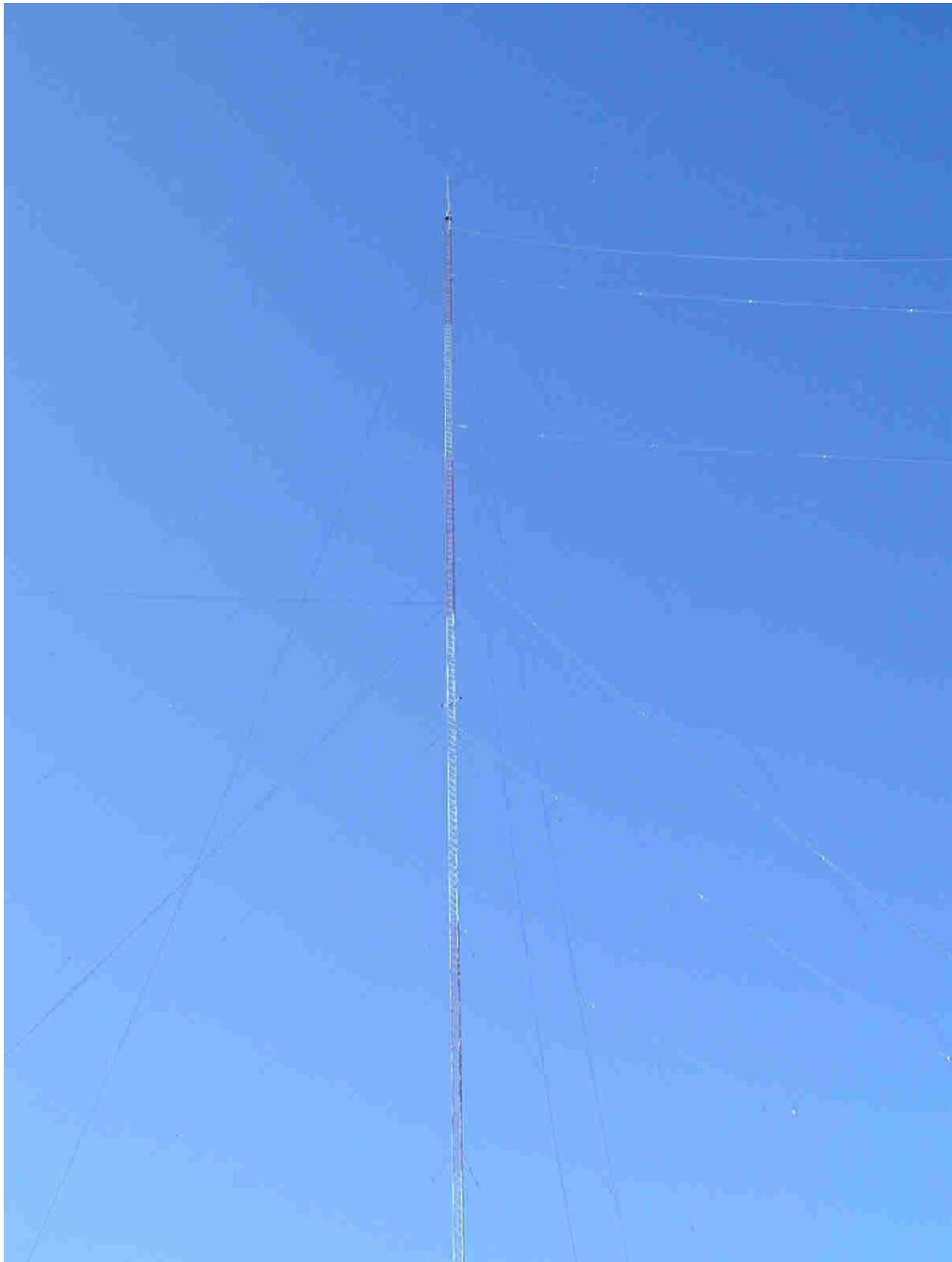
At the top: Our two-meter repeater antenna

Three nylon U-V resistant 7/16-inch ropes over pulleys for pulling up antennas.

Guy lines are EHS steel, with sectionalizing insulators. There are three anchor distances. Look at guyline slope.

At the 170-foot level, two more ropes over pulleys. Just below that, the middle marker lights.

Want to climb up?

View from 318-foot tower

We are only at around 250 feet here, and looking towards the east. The far right horizon has a water vapor cloud from a power plant near Juliette, Georgia about 15 miles away. Nearby is the Whistle Stop Cafe, from the movie Fried Green Tomatoes.

Ground height is about 800 feet above sea level, but the important thing for VHF and UHF is the ground slowly rolls away in three directions. The effective height above average terrain for your view for the next 20 miles is around 300-feet, even though you are only 250ft above ground.

The rope in this picture comes from the end of one of my 318-ft high apex Inverted Vee Dipoles for 160 meters, and goes out over those tiny 80 to 100-foot tall trees about 700 feet away.

My 28 Beverages and 12 vertical receiving antennas are out in the pastures, spread over a 1500 by 3000-foot area. They are controlled by BCD data over underground control cables. Eight coaxial trunk lines feed the receiving antennas by running to switching hubs, where smaller antenna feedlines leave.

For stereo recordings of 160-meter signals, [click here.](#)

Higher Bands

I'm embarrassed. Most of my work is on 160 meters, so I neglect the other bands. I'll work on 80-10 meters someday soon, I hope! At the moment, I only have a single rotating 160-foot tower that supports Yagis for 40 meters up through 6 meters. The hardware comes from K0XG, and has been trouble-free. It has sixty 140- foot long buried radials for lightning protection and to help with low NVIR antennas like dipoles. There are more antennas and feedlines on that tower now.

TX_four_square

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Four-Square

The 200-ft tower has five catenary lines, each about 500 feet long. Four of these lines, running NE, SE, SW, and NW support four 1/4-wavelength wire verticals used in a 160-meter four-square antenna.

Each four-square element is made from a "bow-tie" of #16 wire, which fans out 12 feet at the top. The elements are fed with 50-ohm RG-8 type coaxial cable. Each element has about fifty 1/4-wavelength radials. These radials are bonded to the large radial system of the 200-foot tower at points where more than two wires cross. This makes the entire grounding system for the 200-foot and four-square elements more similar to a large screen rather than just than individual radials.

Four-Square Design

One item overlooked throughout the history of the four-square is the effective element spacing. Effective element spacing is not the same as the distance along each side of the square. The effective spacing is less than that distance, because the array fires from corner to corner. The effective array length is 1.414 times the spacing along each side, not twice that length. With 90 degree per side array size, we really only have a 127 degree long end-fire array when firing across diagonal corners. The correct phasing for an optimum null at zero degrees elevation directly in line with the array is not 90 and 180 degrees, but is actually around 117 and 234 degrees! With the traditional 90 and 180 degree phasing, the array should be stretched to a corner-to-corner distance of 180 degrees, making each side 127 degrees long. Unfortunately feedlines of 1/4-wl electrical length won't reach that distance with velocity factor and connection lengths involved, so the obvious choice is to use greater than 90-degree phase shift. As an additional benefit, the directivity and gain are higher with closer spacing and larger phase shift. Performance is better for both receiving and transmitting.

The phase and spacing used are obviously not optimum, and the popular coupling units almost certainly do not do what we expect. They sacrifice gain because any four-square 1/4 wl on a side really does not "want" 90- and 180-degree phase shift, and they really are not designed for the real impedance presented by the feedlines from the elements. Why do we use such a design? Probably because it was originally used, and no one ever looked at the array in any detail. It was easy to copy the original concepts, and no one thought to check and see if the phasing systems were optimized.

Array Transmission Lines

(the following assumes lines have negligible loss)

I use 1/4-wavelength 50-ohm lines feeding each four-square element. 1/4-wavelength lines have a unique property. When ***odd-quarter-wave*** lines are sourced from ***equal voltages***, the load ends of the line ***will source equal currents into any load impedance.***

There is a second important often overlooked property. When a line is misterminated and has standing waves, total phase shift along the line does not equal the electrical length of the line. An exception occurs when the line is any multiple of 90-degrees.

Just as impedances invert every 1/4 wl, the fixed source voltage at one end becomes a fixed source current at the other end. Not only that, phase shift is constant even though the lines are terminated by different load impedances. Many arrays use this concept to advantage, and a current sourced 1/2 wl line would work just as well as the 1/4 wl lines if we fed the line with equal currents instead of equal voltages. (As a matter of fact that concept is behind a two-band transmitting array I designed.)

Current Distribution

A two-element directional system requires *equal current* at the current maximum of each directional element for maximum F/B ratio and gain. A three-element array would require a binomial 1:2:1 ratio for best pattern. The center element, in effect, has to supply a field for each of the two outer elements to work against, it is doing double duty. The four-square is a novel approach to a three-element array, the center two elements are fed in phase, and act as one common element. If all four elements have equal current, total center element current is twice the current in each end. The 1:2:1 current distribution is satisfied.

With elements series-fed at the point of current maximums, each element requires equal current at the feedpoint. (If the elements were voltage-fed at the voltage maximum, they would require equal voltages at the feedpoint to have equal currents in the elements!) This makes transmission-line type phasing systems easy to design. We simply design the system so each phasing-line has unity standing wave ratio, and tie the elements directly to the phasing lines. Because voltages are equal everywhere in the phasing system, element currents are equal at the end of each 1/4-wavelength feedline.

Phasing Systems

My design operates the phasing line without standing waves. Since the phasing line is not a multiple of 1/4-wl, if the delay line were misterminated phase shift would not equal the line electrical length.

Voltage at each end of the delay line would also be different. Since the line is matched, phase shift will be equal to the electrical length of the line and voltages will be equal at each end of the delay line.

There really isn't any need for "dump" resistances or hybrids in any phasing system. If you look at the systems, the dump resistances really don't guarantee performance. A four-square system never requires controlled "power splitting". As a matter of fact, equal power in each element would guarantee poor F/B ratio and gain.

Keep in mind that each element, other than the two in parallel, has a different feed impedance. With different load impedance at each element current can't be equal in each feedline element group if power is divided evenly. The four-square system, at least one with 1/4 wl feedlines, actually requires equal voltages. Equal power insures a poor pattern, while equal current at the element (or equal voltage at the start of the 1/4 wl feedlines) guarantees optimum patterns. Most four-squares work as well as they do only because the array is somewhat forgiving. They are a good example of reasonable performance through oversight, rather than careful planning.

I solved these problems in my system by placing the phasing common-point in the direction of maximum radiation and using crossfire phasing. This is very similar to the feed-system of a log periodic, except I optimized phase shift and current distribution.

A crossfire delay line system feeds the center two elements in my array. This is similar to the scheme I use in my phased receiving antennas. Each center element has around 25 ohms of operating impedance, and that produces 100 ohms at the end of the 1/4-wavelength 50-ohm feedline that goes to each element. The center two elements, when paralleled, have a combined impedance of 50 ohms.

The delay line from the common point to the center two element junction is 60-degrees of RG-8 coax. The 180-degree "flip" comes by using a choke balun. This is similar to the technique I use in receiving antennas. The primary difference is in core size and materials, my high-power inverting transformer uses eight 2-inch diameter 65-material cores and Teflon insulated coaxial cable. This choke allows the shield and center to be transposed at one end of that delay line. Loss is low in the delay system, because it is a matched 50-ohm line. It is also only 60-degrees long and only handles only about 2/3 of total system power (almost 1/3 of the system power is applied directly to the front element).

The rear element offers a slight problem, because that element has a very low impedance. It almost has enough current, through mutual coupling alone, to be fully functional. As a consequence, it requires very little common point power. With a 1/4-wavelength feedline and parallel reactance canceling, the rear element presents several hundred ohms of resistance to the phasing network. My solution for feeding the rear element was to build an artificial transmission line using lumped components. For the rear delay line, I used multiple L/C sections to simulate a transmission line

with a characteristic impedance matching the rear element. This has three advantages:

1.) Q is low, making phase-shift much less frequency critical.

2.) Voltage is equal at each end of the artificial transmission line, allowing the delay line to all parallel each other at the conmmon point.

3.) I can easily "tweak" delay line characteristics with a few adjustments and optimize the array null.

In effect, I created a high-impedance transmission line from individual L/C components, and only needed to use 120 degrees of lag (the 240 degree lead I needed is electrically the same as a 120 degree lag) in the network. This is all very similar to the system in my [small receiving four-square](#). Because current is low, components can be modest sized. The end result is more bandwidth, more stability, and less loss than a simple one-stage network.

The Results

While this may sound like a complex system, it is very simple in practice. This array has exceptional bandwidth, exhibiting useful F/B ratio and gain over the entire 160-meter band. This also means a similar array for 80-meters would cover almost all of 80 and 75 meters. Additionally, this array has very low loss regardless of operating frequency. It does not divert power into a dump resistance.

My phasing design produces higher gain than conventional 90/180 degree phased four-squares through a tighter pattern and less loss.

Becuase the pattern is "tight", I added a simple system that provides a wide East, South, West, or North pattern. That pattern nearly equals the gain from a traditional four-square, and provides equal or better gain than traditional four-squares in all directions over a wider frequency range.

Here is the pattern of a traditional four-square at 15 degrees elevation, assuming perfect distribution and phase:
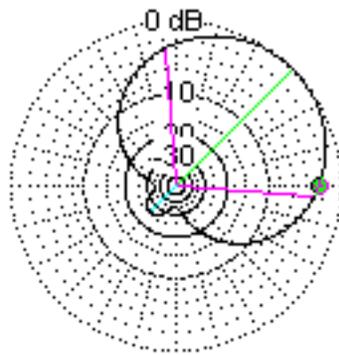
EZNEC

1.8 MHz

Azimuth Plot		Cursor Az	0.0 deg.
Elevation Angle	15.0 deg.	Gain	5.52 dBi
Outer Ring	8.0dBi		-2.48 dBmax

Slice Max Gain	8.0 dBi @ Az Angle = 45.0 deg.
Front/Back	25.61
Beamwidth	98.4 deg.; -3dB @ 355.8, 94.2 deg.
Sidelobe Gain	-17.61 dBi @ Az Angle = 225.0 deg.
Front/Sidelobe	25.61 dB

Note the gain is 8dBi, and the 90-degree azimuth gain is -2.5dB maximum or a net gain of 5.5dBi in the "between" directions

Here is the pattern of the same array with 120/240 degree phasing:
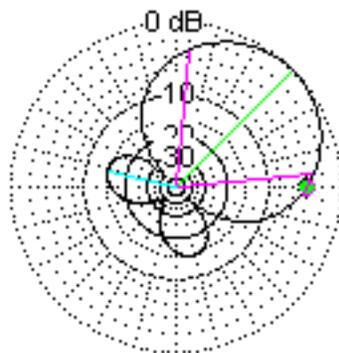
EZNEC

1.8 MHz

Azimuth Plot		Cursor Az	0.0 deg.
Elevation Angle	15.0 deg.	Gain	5.13 dBi
Outer Ring	9.14dBi		-4.01 dBmax

Slice Max Gain	9.14 dBi @ Az Angle = 45.0 deg.
Front/Back	37.28
Beamwidth	78.8 deg.; -3dB @ 5.6, 84.4 deg.
Sidelobe Gain	-5.48 dBi @ Az Angle = 166.0 deg.
Front/Sidelobe	14.62 dB

The gain is 9.14dBi, and the cursor azimuth gain is -4 dB maximum for an "in between" gain of

5.13dB. You only give up .4dB FS at the worse point for a net gain of 1.14dB in primary directions.

Gain increase through the phasing change is just over 1 dB through pattern changes, and is more than that through decreased loss (the model neglect phasing system and feed system losses. F/B ratio change is quite noticeable, as is the ability of the antenna to receive.

My actual working array is even more flexible, allowing the choice of eight directions plus omni.

In concert with a large ground system (the array sits in an area that has ground systems extending for at least a wavelength in all directions) and reduced loss in the phasing system, overall performance has been enough to offset salt-water path and distance advantages of northeast stations using smaller systems. Better F/B while achieving higher gain and more bandwidth is a definite advantage, although any one of the three might have caused me to use this system (besides the satisfaction of making something work as well as possible).

Baluns on Log Periodic Antennas

[<u>Home</u>] [<u>Up</u>]

Some antenna manufacturers place baluns at the incorrect location in LPDA arrays. If we consider what the balun does we can see how the mistake occurs.

In the case of antennas, a balun isolates the shield of the cable from excitation by voltage differences between the feedline's termination and ground. In one case the termination might be a balanced dipole, with each terminal connection terminal having voltage to "earth". If the dipole is balanced and has 100 volts across the feed terminals, the feedpoint would have about 50 volts to an imaginary ground at the antenna.
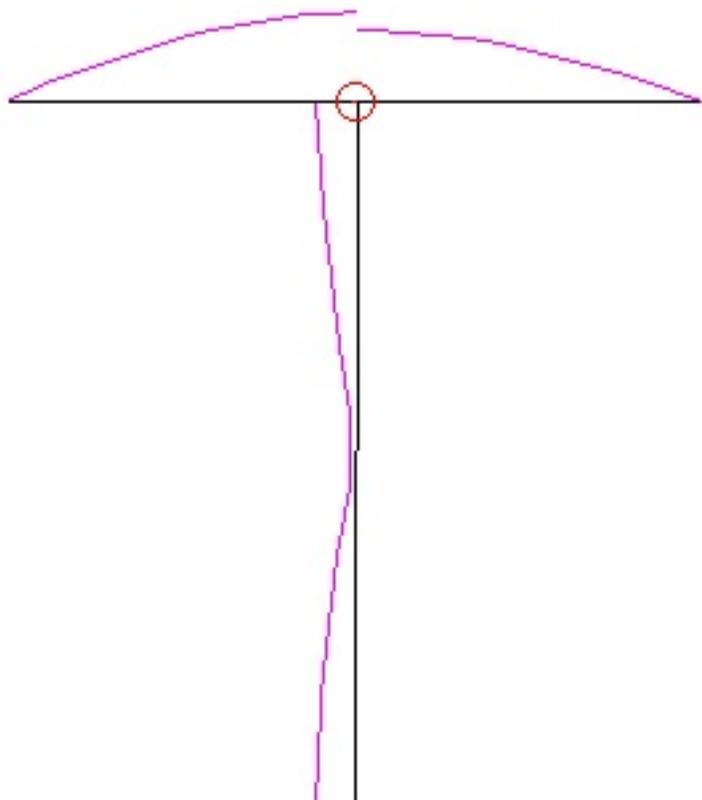
The above current is for a 1/2 wave dipole 1/2 wl above ground, with coaxial feed. You can see the

feedline shield has significant current. In this case about 40% of the antenna's maximum current!

We can simulate a choke balun by adding a current source in series s with the shield, and setting current for zero amperes. The voltage across that current source will indicate the common mode voltage exciting the feedline.
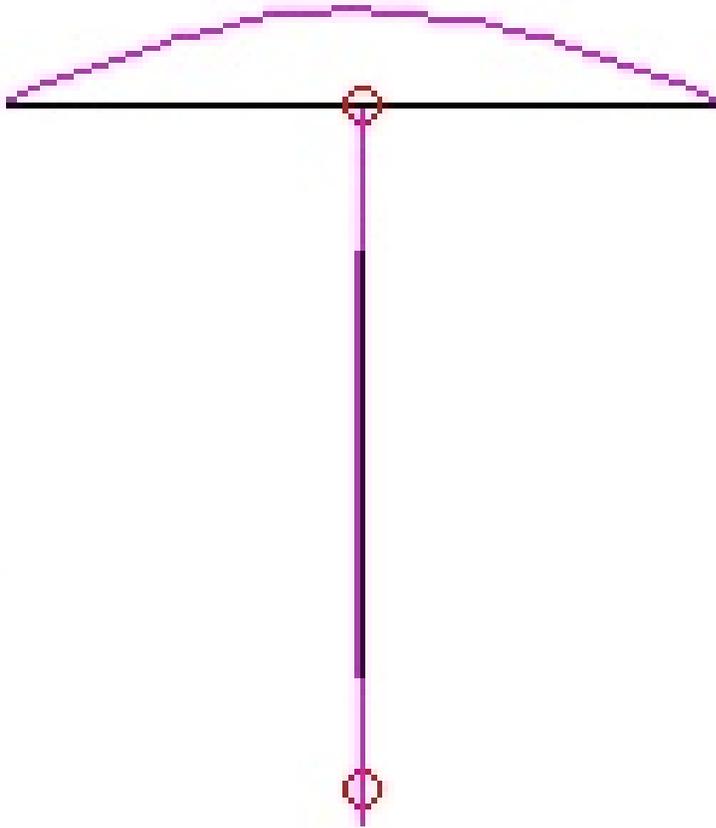
Adding a perfect choke balun, current on the feedline shield goes to zero and the voltage across the choke balun is now found from the source menu:

Source 1 Voltage = 61.02 V. at -0.01 deg.
Current = 1 A. at 0.0 deg.
Impedance = 61.02 - J 0.009274 ohms
Power = 61.02 watts

SWR (50 ohm system) = 1.220 (450 ohm system) = 7.375

Source 2 Voltage = 37.42 V. at 173.27 deg.
Current = 0 A. at 0.0 deg.
Impedance is infinite
Power = 0 watts
SWR (50 ohm system) > 100 (450 ohm system) > 100

Source 1 is the actual terminal excitation of the dipole, while source 2 is the source that cancels common-mode feedline current. We can see the voltage across the perfect balun is indicated by source 2 as 37 volts, slightly more than 1/2 the feedpoint differential voltage. (We could also use a very high impedance load here, and do the same thing.)

What's Wrong with the LPDA Feed?

Some antenna manufacturers have you tape or otherwise attach the coaxial feedline along the length of a boom that is electrically hot, the boom feeding half the elements!! The very fact the boom feeds elements of the antenna requires the boom to have significant VOLTAGE along its length. Any cable parallel to the boom is excited by boom currents and voltages. Cables routed against the hot boom will have whatever the voltage is between the boom and the tower trying to excite current in the feedline shield.
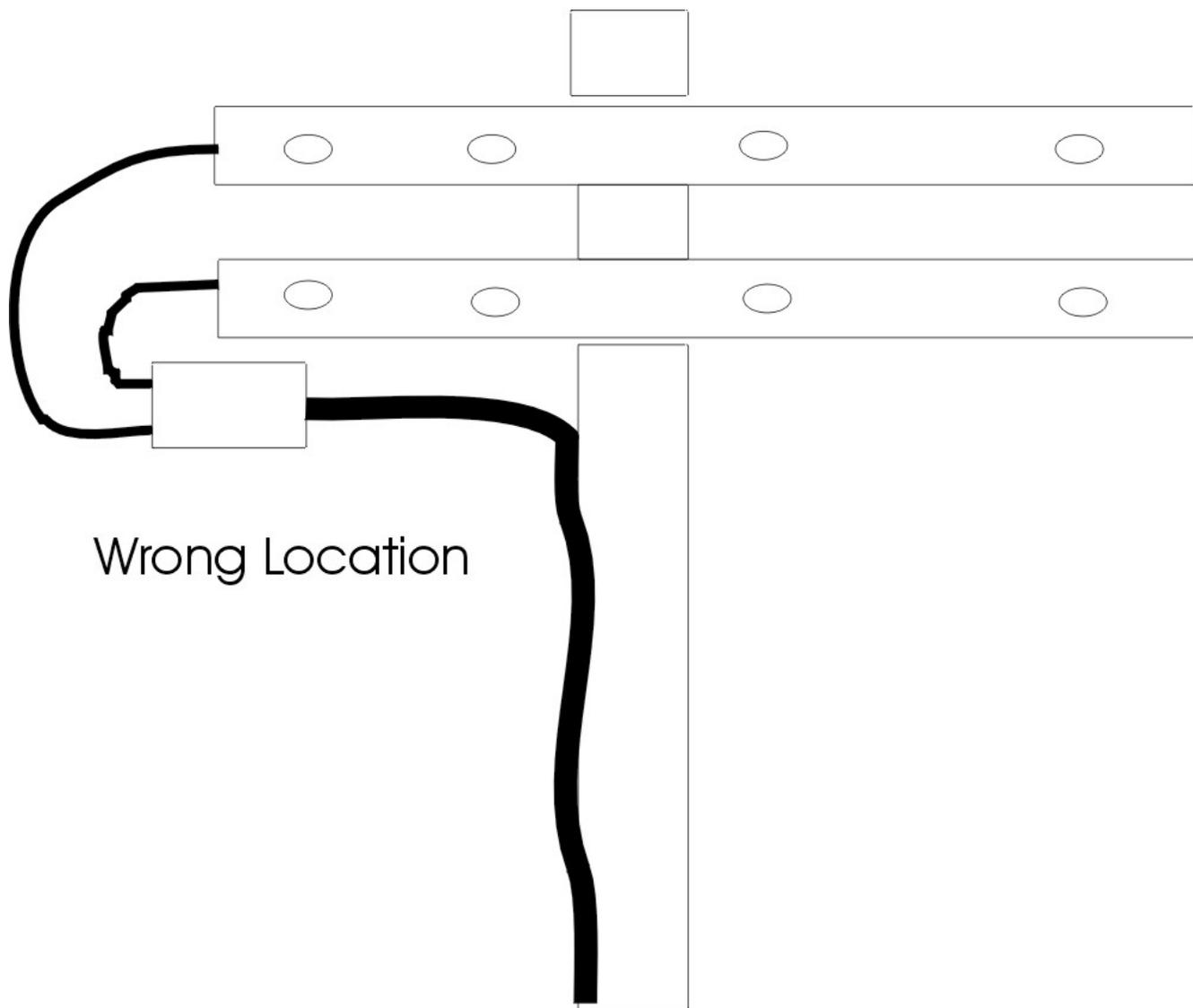
Wrong Location

If the log has a 1:1 balun at the feedpoint, you can simply tape the coax to the boom that the shield connects to without any balun at the log feedpoint. The best thing to do is to ground the coax shield to the hot boom at the exit point where the coax leaves the boom, and install a balun at that point! You can do this by installing a barrel feedthrough connector at the exit point and clamping the connector to the boom, then waterproofing the assembly. The balun goes between that connector and the feedline exiting the antenna boom and going to the station.

Center to top boom

Shield to bottom boom

Right Location
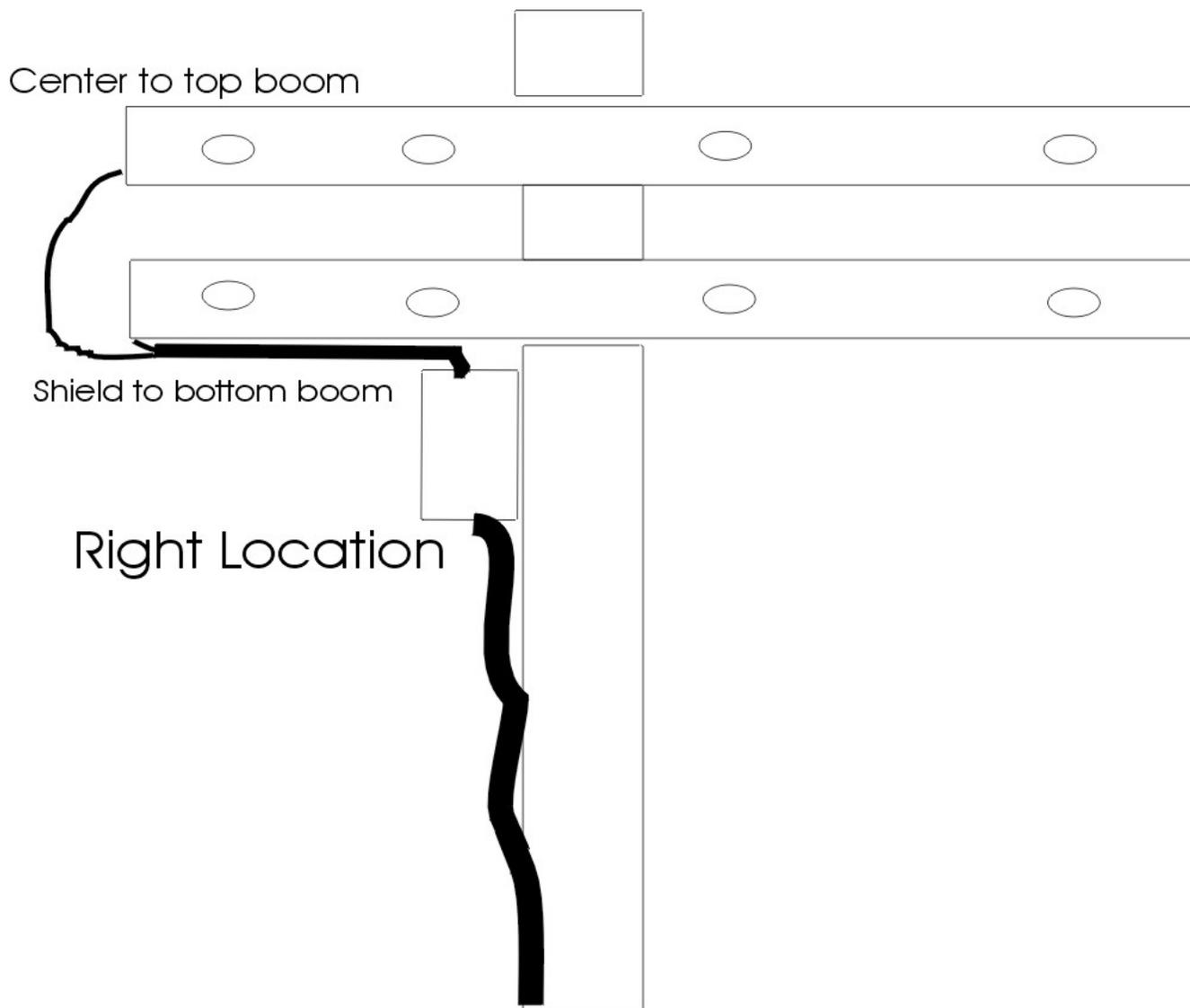
If the beam has an impedance matching balun at the feedpoint, or if you simply want to leave the balun at the feedpoint, you can leave the balun at the feedpoint. The feedline has to be suspended a few feet below the hot booms.
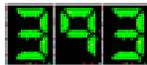
since 6/21/04

Snap on core material data

[Home] [Up]

Can anyone help me identify this core?

It is just over 1" long and about 1/2 inch diameter and snaps easily over RG58 or RG59 cables.

Please reply to W8JI at contesting dot com thanks!

Short dipoles and problems

[**Home**] [**Up**]

Recently I've been fielding problems from people using what has come to be called a "Cebik dipole". I also see people recommending this antenna to people with limited space for antennas. The problems I hear about generally center around tuner damage or failure of a tuner to match the antenna. An analysis of various installations has shown that the Cebik dipole is too short to be an effective antenna on the lowest band without some form of low-loss reactance compensation in, at, or near the antenna.

While pattern looks excellent on higher bands, lowest band performance of a short dipole (in this case 88 feet long on 80 meters or 44 feet long on 40) certainly leaves a great deal to be desired when feedline and tuner losses are included. It isn't the efficiency of the antenna that is in question, it is feedline efficiency and high voltages in the matching system that become problems.

The text below actually applies to any short dipole. If you don't like the dB or so loss of a G5RV system (including tuner and feedline losses), you *really* won't like an antenna shorter than a G5RV.

Loss in a Short Dipole

Power loss in short dipoles primarily comes from compensating reactances and matching to 50 ohms. This is because VAR (volt-amperes reactive) power is very high.

In a system with reactance, current and voltage are not in phase or in step. Because maximum current does not occur at the same time as maximum voltage, the simple product of current times voltage (I*E=P) can be much higher than the actual useful power. The higher reactive current causes increased current squared times resistance (I^2*R=P) heating and loss. The higher reactive voltage, in a similar fashion, causes increased dielectric losses.

In the worse case power loss caused by increased voltage and current in reactive systems can actually cause component failure.

Short dipoles and verticals will almost always model with very good efficiency, because a model is a perfect lossless system if we do not include feedline losses and matching losses.

Let's look at an example by looking at the suggestion of shorting a dipole to approximately .33 wavelengths.

Dipole Impedance

Eznec shows a 44-foot dipole to have an input impedance of

The model is shown here:

Wires

Wire   Create   Edit   Other

☐ Coord Entry Mode   ☐ Preserve Connections   ☐ Show W

Zoom

▲   ▲

▼   ▼

Display   Current

Reset   Reset

Wires

No.	End 1				End 2				Diameter
	X (ft)	Y (ft)	Z (ft)	Conn	X (ft)	Y (ft)	Z (ft)	Conn	(in)
1	0	0	50	W2E1	0	-22	50		#16
▶ 2	0	0	50	W1E1	0	22	50		#16
＊									

Source Data

File   Edit   Search   Format

2D Plot: Cebik 40M and up Dipole

File   Edit   View   Options   Reset

```
            EZNEC+ ver. 4.0

Cebik 40M and up Dipole          7/25/04      7:02:29 AM


-------------- SOURCE DATA ----------------


Frequency = 7 MHz

Source 1        Voltage = 656.2 V. at -87.34 deg.
                Current = 1 A. at 0.0 deg.
                Impedance = 30.51 - J 655.5 ohms
                Power = 30.51 watts
                SWR (50 ohm system) > 100  (450 ohm system) = 46.099
```

* Total Field

0 dB

-10

-20
-30
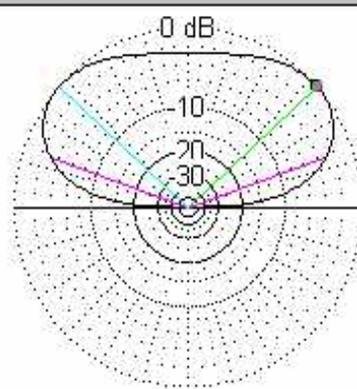
Elevation Plot
Azimuth Angle   0.0 deg.
Outer Ring   6.26 dBi

Slice Max Gain   6.26 dBi @ Elev Angle = 42.0 deg.
Beamwidth   141.6 deg.; -3dB @ 19.2, 160.8 deg.
Sidelobe Gain   6.26 dBi @ Elev Angle = 138.0 deg.
Front/Sidelobe   0.0 dB

From this we see the impedance is 30 ohms resistive, not too bad. The problem is the reactance of -j655. This results in an SWR of over 47:1 on a typical "window ladder line", or 46:1 on a real open wire 450 ohm line.

I've verified the results of TLA (supplied with ARRL Antenna Handbooks) for dry new transmission lines, and it's pretty close. Unfortunately 450 ohm window line is a little worse than TLA predicts, and the impedance is generally lower. But we can consider TLA to be close.

Using TLA we find the following feedline losses for a 75 foot long feedline:

```
 MS
 TLA
 Auto ▼  □  🗐  📋  ✛  🖼  🖨  A

        TLA (Transmission Line), Copyright 1993-1997, ARRL -- by N6BV
                        Ver. 1.0, Mar 05, 1997

    450-Ohm Window Ladder Line
    Length of line:    75.00 ft.
    Frequency:       7.000 MHz
    Transmission line characteristic impedance:  450.0 - j  1.37 Ohms
    Matched-line loss, dB per 100 ft.:     0.125 dB
    Velocity factor of transmission line: 0.950
    Maximum voltage rating of transmission line: 10000.0 V
    Matched-line attenuation =      0.094 dB
    Resistive part of impedance at load:   30.5
    Reactive part of impedance (- cap., + induct., ohms) : -655
    SWR at load:        43.22
    SWR at line input:       29.50
    Additional line loss due to SWR:    1.661 dB
    Total line loss:    1.755 dB   ( 33.2%)

    At line input, Zin = 20.92 - j 294.93 Ω = 295.67 Ω at -85.94°
    At     1500 W, max. rms voltage on line:    4424.6 V
    Distance from load for peak voltage =     55.0 ft.
    Maximum rms voltage rating of cable: 10000 V estimated

    Impedance (Z), Frequency (F), Main Menu (M), Antenna Tuner ([T]), Exit
```

Note the following from the above:

- Line loss is now just under 2dB for a perfectly dry new feedline (it's actually just over 2dB if you use real losses, rather than TLA). This does not include tuner losses or less than perfect installation.

- The maximum RMS voltage on the transmission line is a staggering 4424 volts with 1500 watts. This is over 6250 volts peak voltage. (Even with 100 watts peak voltage would be 1600 volts!)
- Line input impedance is 21 -j295, this is what the tuner must match.

Suppose we add a typical "3KW" tuner to this system. Typical tuners have the following component values and Q's:

- Capacitors: 4.5kV, 250 pF, Q=5000 at mid-setting
- Inductors: maximum inductance 38uH with Q=200

We will assume a perfect balun without any loss. (It makes very little difference by the way if the balun is on the tuner input. Don't think moving a current or choke balun to the input of a tuner is a solution. The balun is under the same common mode stresses.)

With 250pF maximum capacitance, matching the odd impedance of this short dipole requires a tuner with the above component limitations to operate with a Q of 30.4. The resulting loss is 233.6 watts, and maximum voltage across a capacitor is 4250 volts! You are almost at the voltage rating of 4.5kV, so any small imperfections will cause a capacitor to arc. The inductor in the tuner would also dissipate 192 watts, which would damage most inductors unless duty-cycle was very low.

Some tuners (like MFJ Tuners) would be worse than the above values, Dentron tuners would be close to the above values, and many tuners would not even be able to match the load impedances. Automatic tuners and pi-network tuners would have a difficult time matching an impedance like this.

Conclusion

 The lower limit in size of a multiband dipole, before feed system and matching losses move to the edge of severe problems, is about 200 feet on 160 meters, 100 feet on 80 meters, 50 feet on 40 meters, and so on.

A good rule of thumb is to multiply the band in meters by 1.25. The result is the minimum dipole length you can use without using a good matching system in, at, or near the antenna!

The G5RV length of feedline and antenna is the lower limit in size. A normal G5RV system, including tuner, typically has about 1dB of loss on 80 meters and less than 2dB *system* loss (*including loss from coax and matching*) on 80,40, and 20 meters. People seem to hate G5RV's, yet they now seem to be willing to further shorten the G5RV and recommend others do the same!

As an antenna is shortened from that length, losses in the feed system (even what Hams consider a good one) climb rapidly. 88 feet is just too short for an 80 meters antenna, because as you see above it is at the limit of what most tuners will match. It also places most tuners at their power limit at a few hundred watts of applied power.

The optimum length for a multiband dipole is near 1/2 wl on the LOWEST band, and the optimum open-wire feedline length is any odd multiple of 1/8th wavelength on the lowest band. This means an optimum 80-meter dipole would be about 125ft long, and the feedline would be 25-30ft, 75-90ft, or 125-150ft long. The longer the feeder, the more likely you are to having to trim it for optimum tuner performance.

Most tuners like to see impedances HIGHER than 50 ohms, and inductive loads at low impedances. Pi's and L's are NOT a solution to matching problems. They actually are significantly more restricted in matching range than a conventional T using the same general style and quality components!

Feedline voltage is also a good way to estimate wet-weather performance of "window" ladder line. If voltages are more than 1000 volts RMS at 100 watts, operation in wet weather will certainly cause tuning or loss problems. Use TLA and other tools as a way to plan antennas. Remember, there are more important things than pattern! A good pattern is useless if you cannot feed power to the antenna,

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End-fed vertical j-pole and horizontal zepp

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[Verticals and baluns](#)

The rule:

There is a rule that applies to end-fed antennas that cannot be broken. The rule is:

Current flowing into the antenna must be equaled by current flowing into a ground or counterpoise.

This means if you have any type of end-fed antenna without a radial or counterpoise system, the outside of the coax shield must carry an equal amount of current to that flowing into the antenna! This is true no matter how many series traps or choking devices you add at the feedpoint.

End-fed Myths

There are many incorrect ideas and claims surrounding end-fed vertical antennas (like the I-Max 2000), end fed horizontal antennas, Zepp, and J-pole antennas. All of the above antennas are actually quite similar. The J-pole is closer to the original Zepp antenna (Zeppelin antenna) than the bent antenna we commonly call a Zepp. I'll examine the most common and troublesome myth or false claim, that such antennas do not require a ground or will not have troublesome ground currents flowing over conductors into the station ground.

Some myths have been perpetuated or re-enforced by use of incomplete models. The most common modeling mistake is omitting the feedline or, if included, the coupling device (tuning or matching unit) properties are omitted. The major problem often isn't feedline radiation, it is common mode currents from the feedline. This article will explore why those currents occur, and give worse and best case examples of things affecting common mode currents.

Origination of J-pole and Zepp Antenna

This form of antenna originated as a practical solution to feeding a wire antenna trailing behind a dirigible. This type of antenna is a *Marconi antenna*, and it requires a counterpoise of some form.

(Marconi, by definition, means an antenna that is excited against a counterpoise or ground. An end-fed antenna is really a Marconi antenna, since it is impossible to force current into the end of an antenna without equal current flowing in some form of counterpoise.)

One way to create an effective ground system or counterpoise would be to use metal structural parts of the airship. This could create a problem if an RF-induced arc occurred where a mixture of hydrogen and oxygen was present. Running high RF current levels around inside a large bag of hydrogen with a pod of people hanging below was considered poor engineering.

The solution was use of a trailing counterpoise wire. With a second trailing wire was a counterpoise, the antenna system cold be floated from the airship frame. By floating the antenna and ground from the structure, most of the RF currents would be confined to the trailing wire and trailing counterpoise. The initial trailing wire antenna was 1/2 wavelength long, and the counterpoise 1/4 wl long. This set up a high impedance antenna feed working against a low-impedance counterpoise. Most of the current was in the antenna, as seen in this model:

EZNEC+
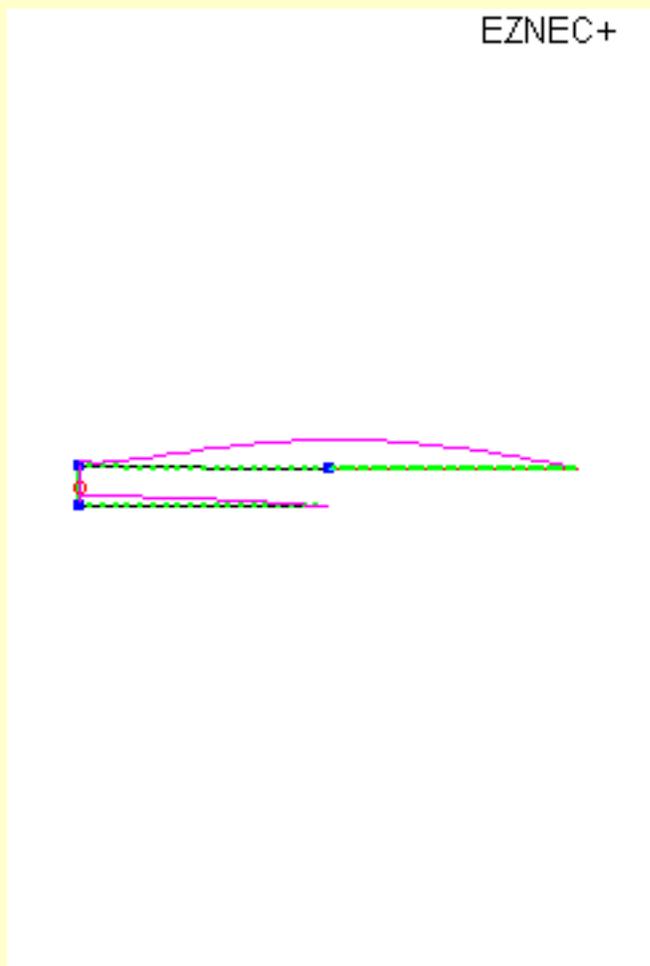
Note the current curves in this model. Maximum current in the counterpoise wire is much less than the maximum current in the 1/2 wave section. Currents at 500 watts are:

- .71a into the 1/2 wl wire
- .71a into the counterpoise wire
- 2.32a maximum at the middle of the 1/2 wl wire

(Pattern is similar to a 1/2 wl dipole at the same height.)

Notice how currents fit the rule I mentioned at the start of this article. At the feedpoint the counterpoise wire, because there is no other path, carries exactly the same current as the antenna wire.

This model, like the models almost everyone uses, has an omission.

The flaw in this model is the same modeling omission K6STI included in his Zepp article, and the same omission repeated in "A Tiger by the Tail". The models do not include real-world effects of the antenna coupling system! All three examples feed the antenna with a *perfect* ground independent current source. That source that has no path to ground or to the station wiring.

When you include an imperfect source, the tiger turns around and bites you squarely on the rump. Here is what happens when a worse-case ground path is added:
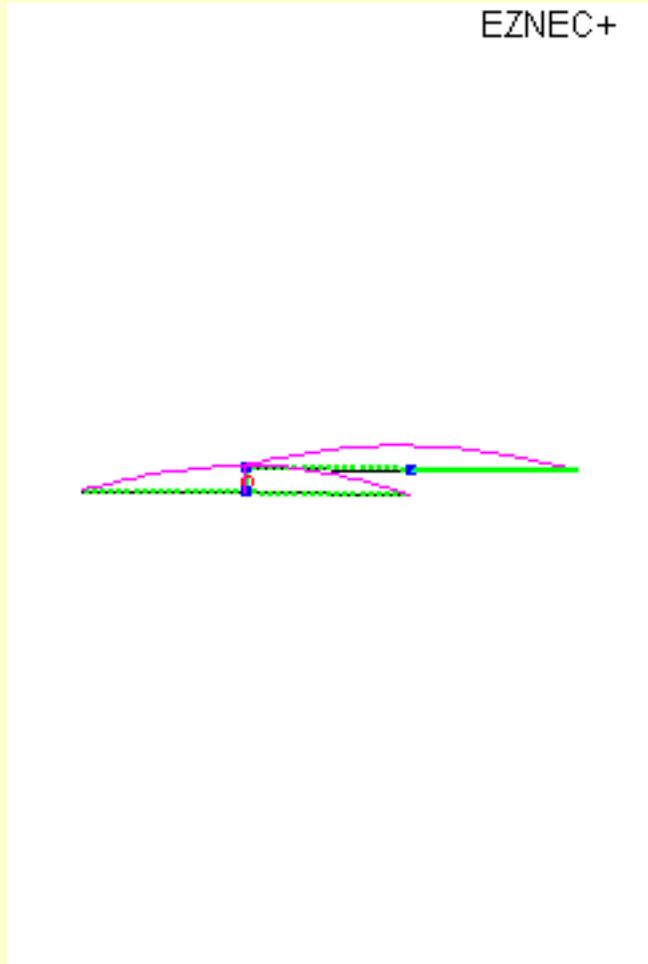
In this case, I added a 1/4 wl wire to represent a low-impedance common mode path from the counterpoise side of the antenna connection. We see ground currents almost equal maximum antenna current!

In reality, most antennas will not be fed this poorly. A typical GOOD tuner has a few thousand ohms common-mode impedance. (There isn't any way around this, it is an impedance limitation inherent in the devices. You especially can't get around the problem by moving a balun to the input of a tuner!)

500-watt system currents when a typical maximum common-mode impedance is added are:

- .71a flows into end of antenna
- . 1a into ground
- .61a into antenna counterpoise

It is the .1a that causes all the problems. If you do not have a very good low-impedance RF ground on your tuner, this significant 100mA of unwanted RF current will flow into you, the power line,

the power supply, the microphone, and everything else connected to the radio and tuner! Of course the current will divide by the ratio of impedances, but this is precisely why, contrary to what other authors might claim, RF problems are common with Zepps. As a matter of fact, K9STH (a respected person on QRZ forums) uses a Zepp in an non-ideal configuration and has become a big proponent of extraordinary [station RF grounds](#). This isn't meant to criticize anyone, but it is necessary to call attention to the problem of model sources being perfect while real-world systems have limitations.

The fact is, *if an antenna is fed properly there is absolutely no need for a station RF ground! The station RF ground is a band-aid for poor antenna system design.*

If an antenna is modeled properly to include feedlines and tuners, you will see problems that idea-source models overlook! Conclusions about system behavior change when tuners and transmission lines are included in models.

The Next Zepp Step

Let's look at the next evolution of the Zepp (this applies to end-fed verticals like the I-max 2000 although I will cover end-fed 5/8th wave verticals in detail at a later time). In this case we will model a 1/2 wl end-fed antenna fed with a 1/4 wl transmission line (a conventional year-2000 Zepp antenna), and see what happens when the 1/4 wl feedline is tied to ground through the common-mode impedance of a typical better-grade antenna tuner.

The ideal case for Zepp or J-pole (they are the same except for the bend in the Zepp at the feed) is when the radiator is very thin, exactly 1/2 wl long, and fed with a parallel conductor stub that is exactly 1/4 wl long.
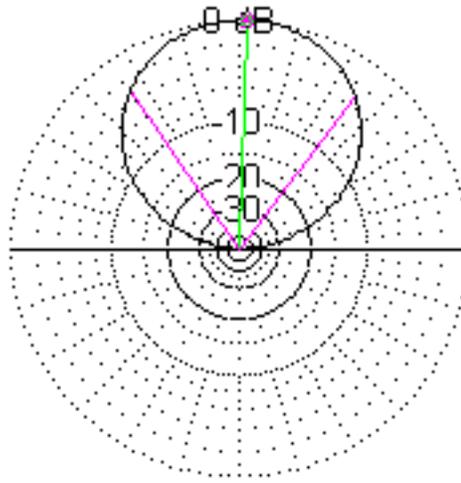
Here is the antenna view:

EZNEC+
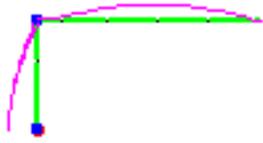
Ideal antenna shows no pattern distortion:

* Total Field   EZNEC+

                                                                 3.8 MHz

Elevation Plot   Cursor Elev   88.0 deg.
Azimuth Angle   0.0 deg.   Gain   10.77 dBi
Outer Ring   10.77 dBi   0.0 dBmax

Slice Max Gain   10.77 dBi @ Elev Angle = 88.0 deg.
Beamwidth   72.0 deg.; -3dB @ 52.7, 124.7 deg.
Sidelobe Gain   < -100 dBi
Front/Sidelobe   > 100 dB

Although high feedline currents appear, they are closely balanced and out-of-phase. This is why both W4RNL and K6STI concluded feedline radiation is not a problem, and I agree *for this specific perfect case!*

The real world adds a few variables, however. Real-world sources are not ground independent, real-world antennas are not always 1/2 wl long, and real-world feedlines are not 1/4 wl long.

Let's explore what happens when a the real world enters the perfect model!

Adding a worse-case ground path to a perfect antenna, we have:
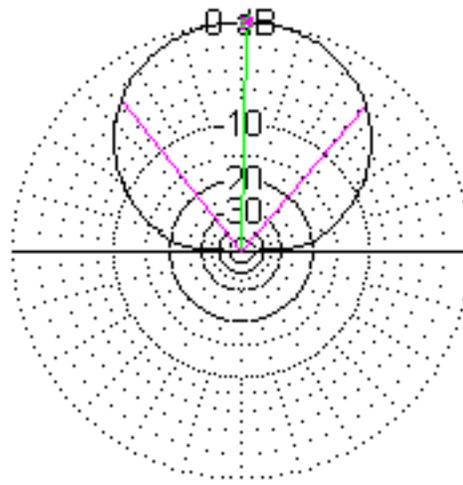
```
* Total Field                                    EZNEC+

                          0 dB
                         -10
                         -20
                         -30

                                                 3.8 MHz

Elevation Plot                  Cursor Elev   89.0 deg.
Azimuth Angle   0.0 deg.        Gain          8.45 dBi
Outer Ring      8.45 dBi                      0.0 dBmax

Slice Max Gain  8.45 dBi @ Elev Angle = 89.0 deg.
Beamwidth       79.4 deg.; -3dB @ 49.6, 129.0 deg.
Sidelobe Gain   < -100 dBi
Front/Sidelobe  > 100 dB
```

Now we see a slight flattening of the bottom of the pattern. While the pattern does not change much, we have the following currents at 500 watts:

- 6.1 amperes feedline wire 1
- 5.35 amperes feedline wire 2
- .75 amperes ground path

Even with the small pattern error, we now has a very significant .75 amperes flowing to earth through the station equipment and perhaps eventually the operator. This is with an IDEAL antenna, and a normal worse-case antenna feed.

What is a worse case feed? The common J-pole antenna fed with coax! .75 amperes would represent the undesired shield currents and supporting pipe or mast current total when a J-pole is fed with coax.

This problem gets worse when the Zepp feedline is not a 1/4 wl long, and the antenna is not 1/2 wl long. The I-MAX 2000 CB vertical is one example. The antenna has a random feedline (your coaxial feedline and mast) as a counterpoise, and is not a 1/2 wl end-fed vertical! This is why so many people complain about RF feedback with Zepp and end-fed 1/2 wl antennas.

If time permits I'll add examples of worse length combinations, but this example of a best length proves common-mode currents are an issue with a perfect Zepp connected to a real tuner. This is why people with Zepp are generally people who believe good RF grounds are required. They are correct in that belief. if you have a poor antenna design you need a good RF ground at the station!

Ground Plane and J-Pole Problems and Patterns

The J-pole is a worse-case situation for examining unwanted feedline radiation. In the case of the J-pole without radials, all of the common mode current created by the poor feed system flows over the coax shield and mast. The mast and coax is generally vertical and as high and clear as the actual thing we think is the antenna, and so unwanted radiation from the feedline is much more apparent than in the low-band Zepp where both feedline and antenna are near earth.

We will use this as a reference. Here is the pattern of a *1/4 wl groundplane antenna* without proper feedline decoupling:
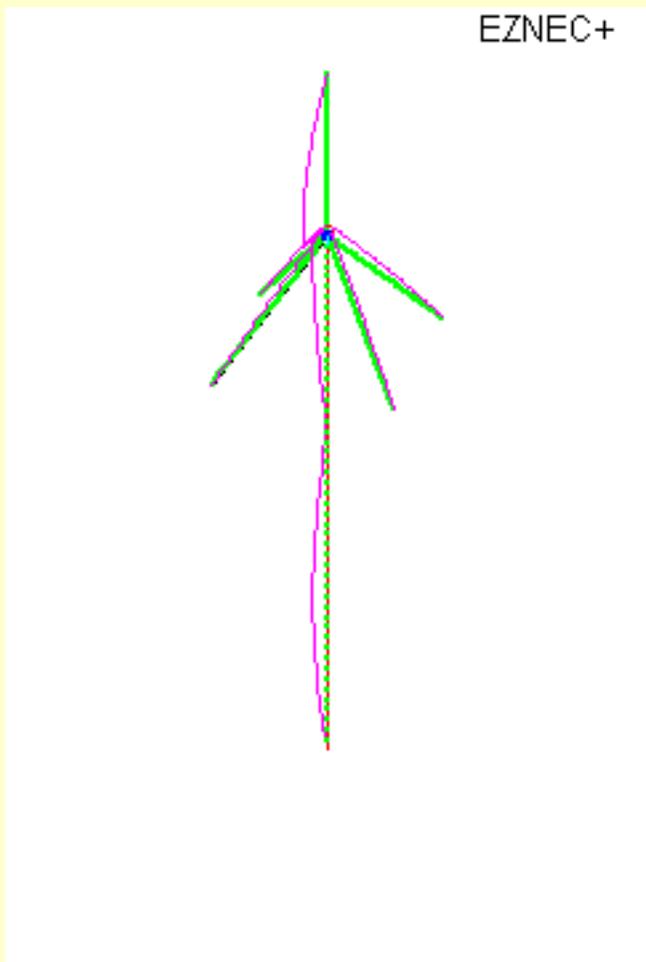
You can see some current exists on the feedline. The amount of current is actually significant, almost 1 ampere compared to 2.5 amperes in the antenna (500 watts applied)! A 1/4 wave groundplane with four radials actually needs a balun or some other form of feedline decoupling.

The pattern in freespace of this antenna is:
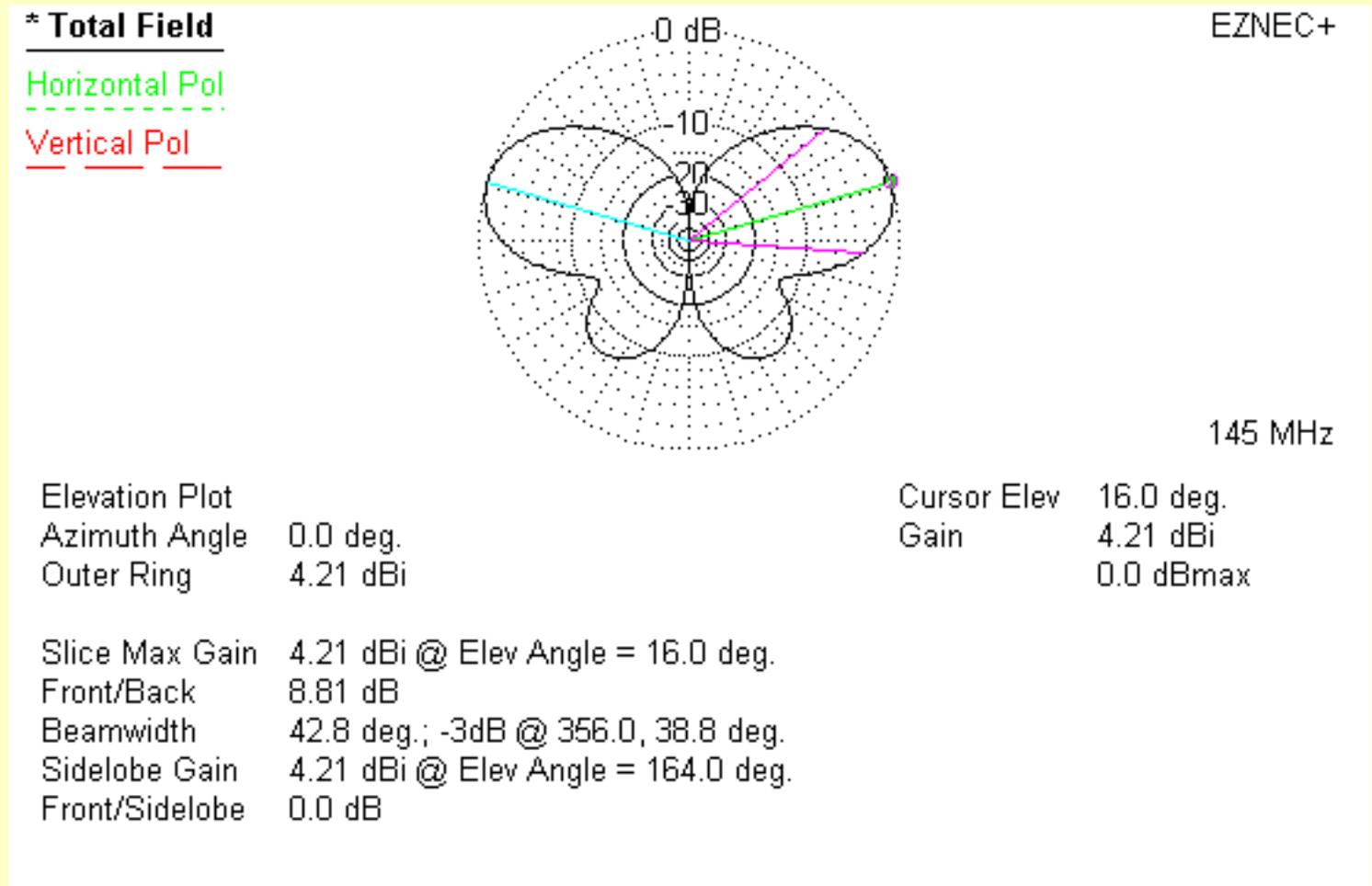
The pattern distortion is caused by feedline common-mode current. If you think this is bad, imagine what happens with an end-fed antenna (even a 1/2 wave antenna) and NO radials! In that case all of the antenna current flows over the feedline shield! Many antenna designs actually use the feedline and mast radiation that others dismiss as "insignificant" to increase antenna gain. In some cases, the antenna designers really don't even understand what they did to create a "magic" antenna.

Here is the same 1/4 wl groundplane with a feedpoint balun with a common-mode impedance of 3000 j3000:
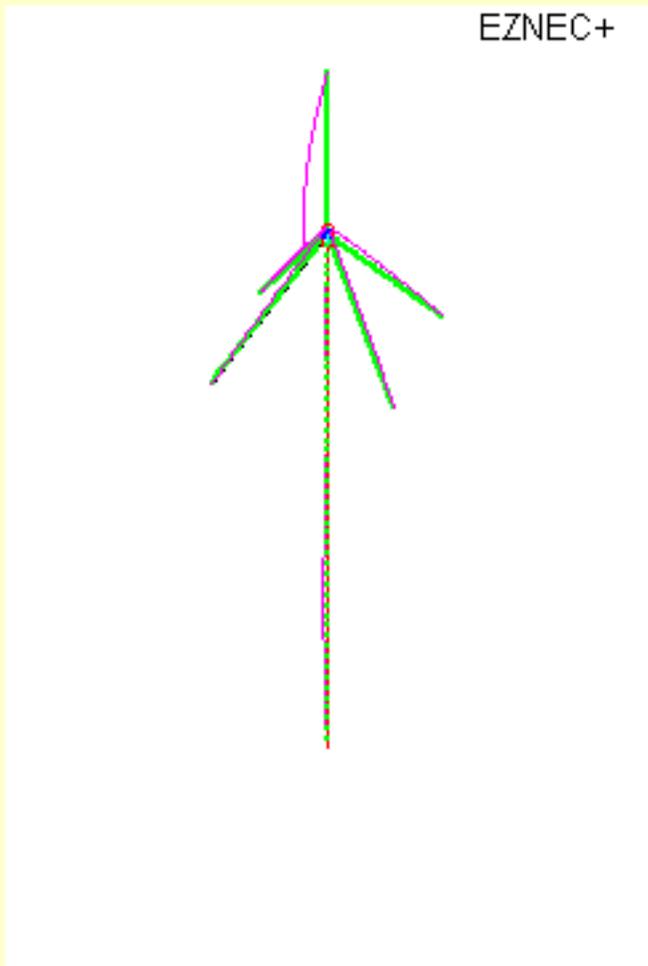
In this case feedline currents are under .4 amperes, and antenna current increases to 3 amperes. pattern is:

See how much cleaner the pattern is? The pattern is much cleaner and more like what we expect from a groundplane far above earth. We still don't have a good system, because the balun needs a ground on the "unbalanced" side to be effective. (This is why I recommend grounding the unbalanced side of baluns to antenna booms of Yagi antennas. That coil of wire hanging in the air might not be enough without a ground!)

With a perfect balun and optimum de-coupling (a few extra radials below the balun or an extra-ordinary balun), the pattern looks like this:

* Total Field

Horizontal Pol
- - - - - - - -

Vertical Pol
___ ___ ___

EZNEC+
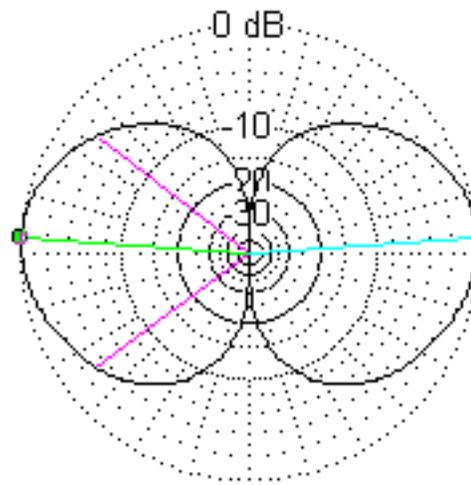
145 MHz

Elevation Plot		Cursor Elev	176.0 deg.
Azimuth Angle	0.0 deg.	Gain	2.69 dBi
Outer Ring	2.69 dBi		0.0 dBmax

Slice Max Gain   2.69 dBi @ Elev Angle = 176.0 deg.
Front/Back   0.18 dB
Beamwidth   74.5 deg.; -3dB @ 142.6, 217.1 deg.
Sidelobe Gain   2.69 dBi @ Elev Angle = 4.0 deg.
Front/Sidelobe   0.0 dB

This is the ideal pattern of a 1/4 wl groundplane, and it takes extraordinary care to obtain this pattern. Many people think they have this pattern, because they model antennas ***without*** *the feedline attached!*

To obtain this pattern I had to:

- Use a good balun just below the radial tip level
- Ground the feedline shortly below the balun to a small groundplane
- Slope the radials downwards more, to a 75 degree angle.
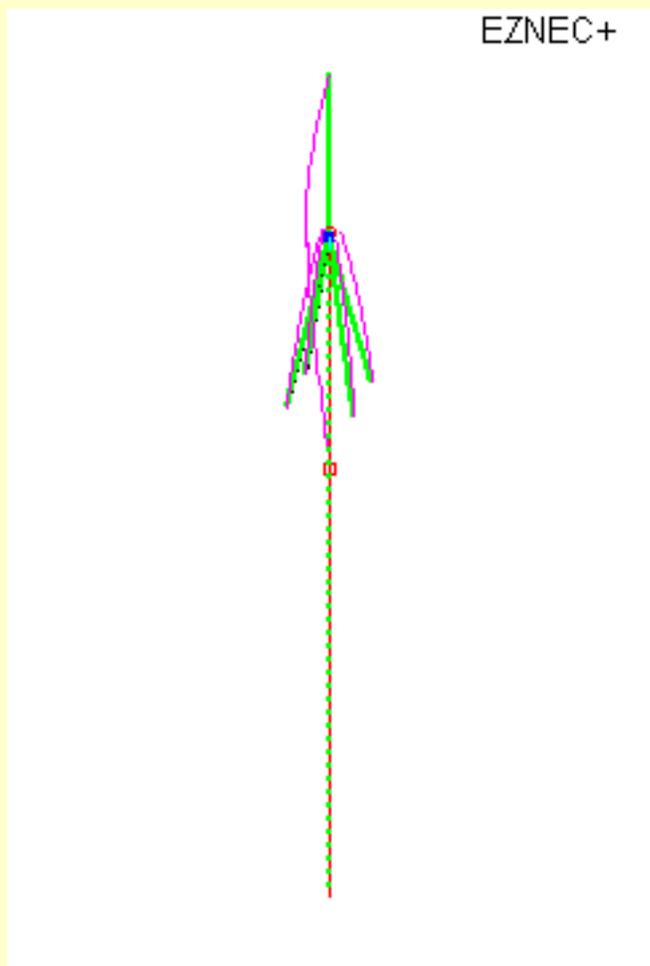
It is easily possible to do this in the real world. Instead, we have numerous articles that ignore the feedline and assume we don't have a problem with common mode current upsetting antenna pattern.

Worse yet, some articles claim the groundplane antenna only needs one or two radials!! Why does this happen? Because the person who made that assumption never modeled the mast or feedline. They eliminated the real-world problem through an omission in the model.

If an antenna model does not include the feedline, losses in the feedline, or imperfections in the source it much more likely than not is a flawed model. The real-world antenna often will be much different if the model ignores feedline loss and feed system source or matching system problems. Short dipoles are one example.

The J-pole and other end-fed Hertz antennas as prime examples of antenna with severe common mode current problems. The coax shield has to be at zero volts potential and have exactly equal and opposite currents flowing into and out of the load and source, otherwise the feedline radiates. Feedline length or weather changes that affect feedline moisture between the outer jacket and the support for the feedline will often change SWR. The severe common-mode feedline problems of end-fed 1/2 wave antennas is why some people swear by them, and other people swear at them.

Here is a zoom of a correct model of a J-pole with a vertical feedline and/or mast attached.
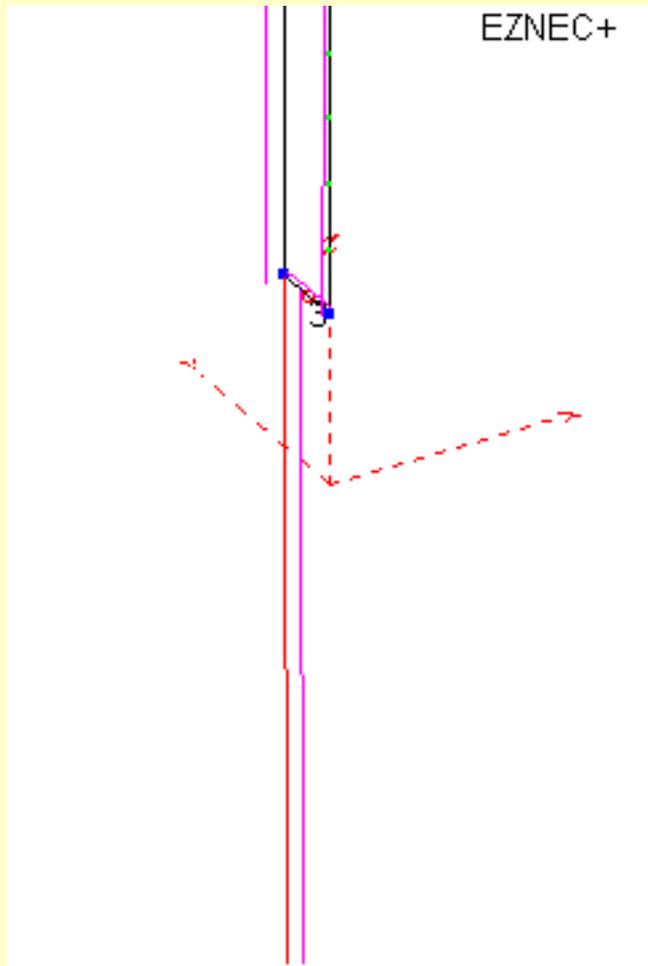
You'll see the feedline or mast grounds directly to what everyone assumes is a "zero voltage" point. This is the electrical equivalent of any J-pole with the coax connected in series with the feedpoint, and the shorter leg connected to the shield. The shield can be connected to any supporting mast with any change in results. Here is the resulting pattern:
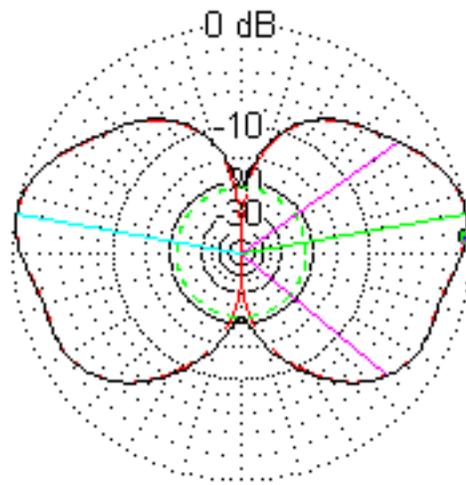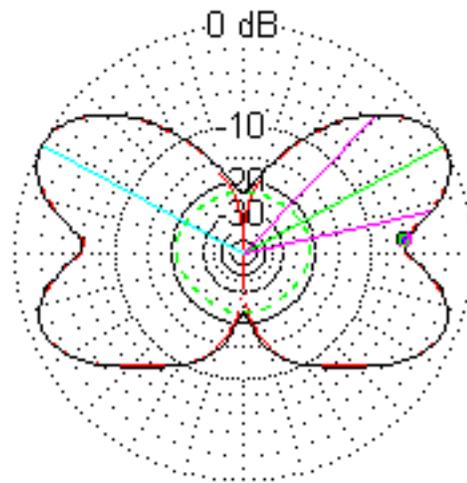
The gain is now 2.37 dB at 4 degrees elevation compared to 2.69 dB for the 1/4wl groundplane. This isn't the worse feed arrangement....it is actually the best for the J-pole! Here is the pattern with the feedpoint reversed, the shield is connected to the longer element, and the center conductor to the short element:

Notice the low-angle gain dropped about 5dB with just a simple reverse of feedline connections! If I didn't model the feedline, the model would never show this problem. In ALL cases, the SWR stays near 1:1, yet gain at low angles changes 5dB!

The following model is an I-Max 2000 vertical with a vertical feedline. In this case I picked the worse feedline length:

EZNEC+
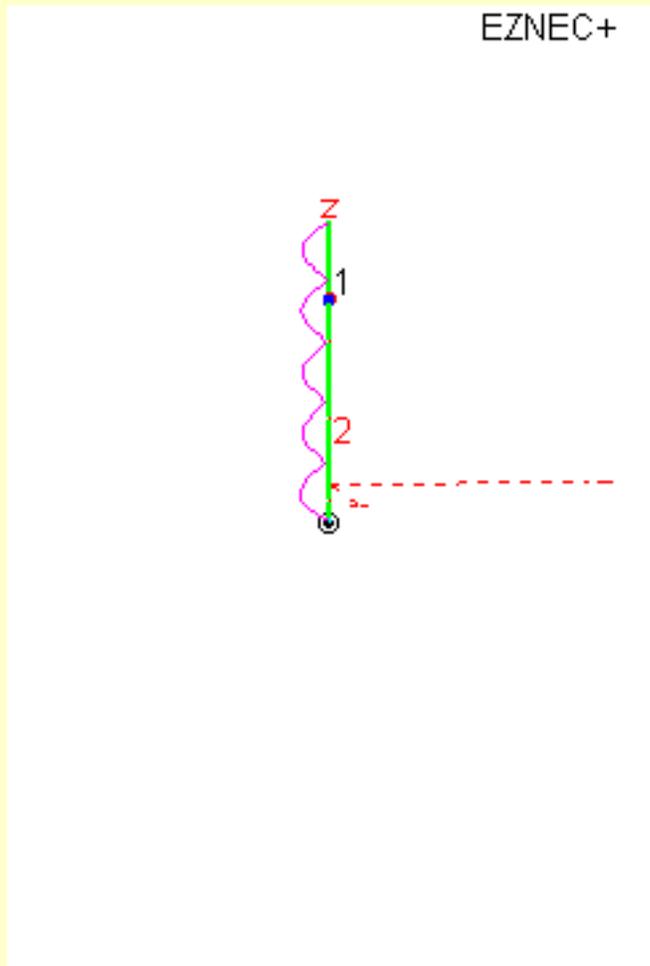
Feedline current is 100% of antenna current. This illustrates why so many people complain about SWR problems and RF in the shack with end-fed verticals like the I-MAX 2000!

Here is the pattern of an antenna that copies the I-MAX dimensions and feed system:

This is a NEGATIVE gain antenna at low angles. A 1/4wl groundplane would seriously out-talk the I-MAX 2000 or any other 5/8th wl antenna that does not have a large groundplane.

Even if we use the optimum feedline and mast length, here is the very best the end-fed antenna will do:

* Total Field

Horizontal Pol

Vertical Pol

0 dB
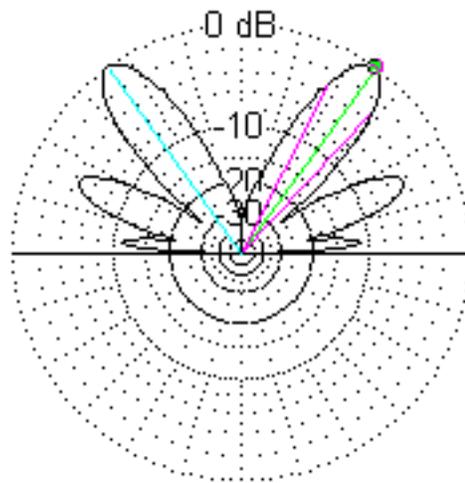
EZNEC+

28 MHz

Elevation Plot		Cursor Elev	8.0 deg.
Azimuth Angle	0.0 deg.	Gain	2.67 dBi
Outer Ring	3.2 dBi		-0.53 dBmax

Slice Max Gain   3.2 dBi @ Elev Angle = 6.0 deg.
Beamwidth   10.9 deg.; -3dB @ 2.8, 13.7 deg.
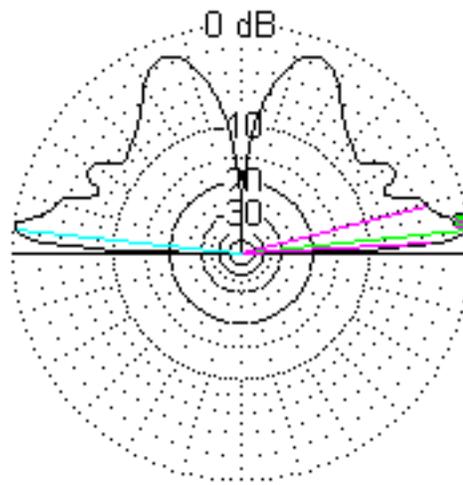Sidelobe Gain   3.2 dBi @ Elev Angle = 174.0 deg.
Front/Sidelobe   0.0 dB

In this case we now have 2.67 dBi, which is actually a little less than a 1/4wl groundplane will do! The severe common-mode mast and feedline currents make "no-radial" verticals extremely sensitive to mounting height, mounting structure, feedline length, and grounding. This is NOT normal for antennas, it is a sign of a design problem.

Summary End-feds Without Grounds

ANY END-FED ANTENNA REQUIRES A LARGE GROUNDPLANE OR OTHER EXTRAORDINARY ISOLATION METHOD OR METHODS TO PREVENT FEEDLINE OR MAST COMMON MODE CURRENTS!

This is true for 5/8th waves, Zepp antennas, R7's, R5's, or even common J-poles. End-feeding antennas is bad news unless you have a large well-established ground at the feedpoint. Even 1/4wl groundplanes have common mode problems.

Some manufacturers have wised-up.

Cushcraft, in their Ringo-Ranger, eventually added a separate additional groundplane below the antenna to tame the wild common-mode currents of the Ringo. Even that solution is barely

acceptable.

The Isopole antenna used multiple sleeve sections to decouple the feedline, and it probably was one of the best antennas available for immunity to feedline coupling problems.

This problem gets worse when the element is 5/8th wave long. Think of that when you read claims of "no-radial" CB antennas with "3dB gain" and a low wave angle. They actually have negative gain at desired DX angles over a conventional 1/4-wave groundplane! Instead concentrating your signal towards the neighbor's TV set or an airplane flying overhead. They cause your entire antenna system to be critical for feedline grounding, routing and length and even allow moisture on the feedline jacket to change performance of the system!

If you find this page interesting, please pass it along. It's time to get the feedline into models!

This page has **00567** **hits since creation on July 31, 2004**

Two or even three skirts may be added to the mast as shown in Fig. 22-5c, but reduced excitation of the lower sections diminishes the effectiveness of each additio skirt.[6]   Thus, while the multiple-skirt coaxial antenna resembles a collinear arra in-phase half-wave elements, its gain is not as great.   Typical gain values are ta lated in Table 22-2, but since no test data are available for this type of antenna, values given are merely engineering estimates.

(a)   (b)   (c)

FIG. 22-5. Evolution of multiple-skirt coaxial antenna.

The relatively large diameter-to-length ratio of a practical skirt produces an effect which requires that the exterior length of the skirt be reduced by a factor of 0.8 to 0.9 and the interior length by from 0.95 to 0.98.[6] This is accomplished cutting the skirt to the proper exterior length and inserting a dielectric slug in skirt to increase its electrical length.

diversity receiver and transmission

[Home]

Diversity Receivers

We sometimes are told a quad antenna, by virtue of the vertical side and horizontal top and bottom wires, is "diversity". That just isn't true at all. Neither is an intentional mix of vertical and horizontal reception from one antenna (like a Carolina Windom), or even two separate hori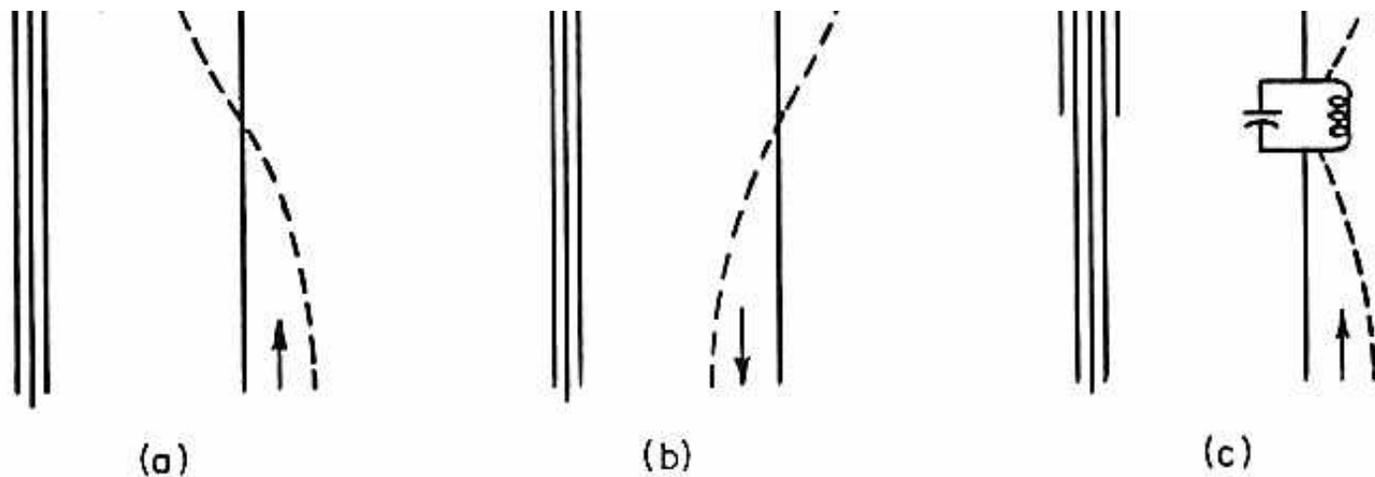zontal and vertical antennas when combined into one receiver signal input. I've actually never even found it to work very well even in stereo or with a voting system, as long as the antennas are within a wavelength or two of each other.

Despite what we are sometimes told, we can't simply adjust two receivers to the same basic frequency, pipe one output into each ear, and have diversity. This is especially true with unlocked receivers that are even a fraction of a Hz different in frequency.

There are articles describing locking the dials of the FT1000D so the sub-receiver tracks the main receiver. This does work in the FT1000, because the oscillators share a common time base. The main problem with the FT1000 is the second receiver isn't a "good" receiver.

Testing For Diversity Capability

You can test your system for correct phase-lock by tuning in a carrier (like WWV) and mixing the two outputs together. When audio levels are equal, you should hear no warble or vibrato in the tone. You should hear something between a full or peak that stays perfectly locked. If your system fails this test, it will deduct greatly from possible diversity advantages. Click here for a failed test. Click here for a passed test.

Remember any warble or slow fade variation indicates the receiver is unacceptable.

True vs. Stereo Diversity

I use a loose form of diversity reception on 160 and 80 meters I call "stereo diversity". It really isn't true diversity where the receivers vote and the best S/N ratio captures the audio output, it is a system that requires your "brain" to do the voting.

I've found this system very good for substantial improvements in readability of noise-floor signals. The difference can be worth as much as a signal being nearly readability 5 (perfect) in stereo to readability 2 without. When a signal is marginal, it is all the difference in the world.

I accomplish stereo diversity by phase locking two separate receivers together so audio outputs are exactly in phase. The receivers are virtually identical, even to the point where I hand select crystal filters for equal group delay change over the filter passband. Every oscillator in the receiver is common to both receivers.

This allows me to successfully combine two antennas or antenna arrays that are several wavelengths apart into one system, a task otherwise impossible. Mixing two quiet widely-spaced antennas causes noise to appear smoother and more "hollow sounding", while the signal actually stands out more. Some of the recordings on my DX Sound Files page are in stereo. You can listen to this effect with stereo headphones by changing your computer's volume control settings to stereo or mono.

Transmitting

One common thought or claim is by transmitting both vertical and horizontal via skywave, we would have the best of two worlds. Worse yet, there is a common idea or belief we can build antennas producing two independent polarizations, and that resulting dual polarization will provide the best of two worlds and reduce transmitted signal fading. Several obvious flaws with this concept are outlined below.

Generating Two Polarizations

When speaking of polarization, we are talking about the direction of the imaginary flux lines in the electric field. When dealing with the far-field effect called EM radiation, *the imaginary electric and magnetic flux lines are conveniently at right angles to each other*. While either could have been used for the reference standard of polarization, the electric field became the polarization reference.

The imaginary flux lines represent the force caused by any and all electric fields. They "exist" only at one angle in one small portion of space at any instant of time. We *can not generate two polarizations at the same time at any reference point in space* when broadcasting our signals, not with any antenna! The idea we can have dual-polarization transmissions probably comes from misunderstanding what antenna modeling programs are showing us, or a flawed or limited imagination causing an incorrect mental picture of what actually occurs with antennas.

Modeling programs only show two perfectly filtered views of the actual field. They do this out of

necessity, because the actual electric field is far too complex to display in its entirety. While we could see a slice of the field showing true polarization at any given angle and distance, I'm not aware of any commonly available programs that provide such useful information. A typical pattern display generally shows the response that would be observed through perfect vertical and horizontal filters. Modeling programs generally do not tell us the phase relationship between the intensities we see displayed, so we have no idea what the actual polarization is.

The bottom line is this....we don't know, when looking at the display, what the actual polarization is unless it is 100% vertical or 100% horizontal.

A Simple Tilted Dipole

Visualizing an actual antenna might help us picture an antenna pattern correctly, and understand what we commonly perceive incorrectly.

Imagine we have moved back a considerable distance from the center of an extremely high dipole that was installed tilted at 45-degrees. We move back from the dipole center without changing height, and observe the electric flux lines near us. To us, the distant antenna appears tilted at a 45-degree angle from our lower left to upper right. Each end of the antenna is exactly the same distance from us. In other words, this is a "side view" of a perfect sloping dipole.

If we could actually see flux lines near us representing the electric field, the lines would appear to parallel the distant antenna. Yet a view on a modeling program would show the field intensity of the electric field to be an exactly equal mix of vertical and horizontal fields!

Many of us would (incorrectly) describe this antenna as producing equal vertical and horizontal polarization in the direction where we view the antenna. The logical conclusion would probably be a vertical or a horizontal would respond equally well to that field, and that is correct for a perfect vertical or horizontal.

What we probably would fail to understand or visualize is the actual polarization. The peak response would be to a dipole antenna tilted 45-degrees in the same slope direction as the radiator, from lower left to upper right. Most important and most often missed is the simple fact that another dipole tilted 45-degrees opposite, from lower right to upper left (even though broadside towards the distant source) would have no response! It would be cross polarized, and response would be minimal.

This idea we have both polarizations is the root of the misunderstanding, and misunderstanding always seems to breed voodoo antenna claims and snake oil solutions. The most common false conclusion would be thinking sloped antennas reduce polarization related fading. The "equal-V-and-H-antenna" would be assumed to provide the best of two worlds, transmitting or receiving when either vertical or horizontal polarization or anything in between is required for an optimum signal.

Ionosphere Propagated Signals

Any distant signal arriving via the ionosphere is constantly changing in polarization. The ionosphere is a poorly aligned soup of ions, and that soup is constantly being stirred. It is not a flat perfectly aligned mirror. The ionosphere also provides multiple modes and paths for signals, particularly on frequencies well below the maximum usable frequency. The phase and level of the same signal arriving from each path constantly changes. Because of this, arriving signals tilt and rotate.

While there are some statistical odds that more time will be spent centered around one effective polarization than another, the fact remains that very little time is spent at one distinct polarization. The same effect holds true when transmitting.

Because of the random nature of polarization, the signal just as likely would be tilted 45-degrees left on on sloper as 45-degrees right. It is just as likely to fall into a cross-polarization null with the sloper as with any other angle of radiator, except one centered at the optimum tilt. We can easily see the idea sloped wires, Inverted L's, and even Windom antennas with "leaky" baluns reduce fading by providing "diversity" is pure rubbish. The same holds true for intentionally mixing two polarizations from two separate antennas, even if each is fed from separate amplifiers.

Statistically, we are actually MORE likely to have deep fades when we transmit with two very different systems than with one! The reason is simple, we excite the multiple paths better and increase multipath propagation. Since the phase delay is random and constantly changing, any attempt at circular or dual polarization would greatly increase fading when more than one path makes it to the receiver.

This is actually the reason 5/8th wavelength verticals fell out of use in broadcast work. The small high-angle lobe of the 5/8th wavelength antenna created severe deep fading and phase distortion at receivers in the fringe areas. A mix of horizontal and vertical antennas, even if properly phased to provide a rotating wave, would be even worse.

Reducing Fading

The best solution is to have two separate antennas, selecting the best antenna at any given instant of time for the path. This is true for both receiving and transmitting. Commercial sites do this by employing some form of voting, based either on signal-to-noise or absolute signal level.

Dealing with weak signals provides special problems. Good CW operators can copy code that is actually below noise floor. Because S/N ratio is near zero, noise detectors in a voting system would become overly complex and unreliable. Perhaps someone can develop a DSP system that allows

voting, but my attempts have been largely unsuccessful.

My solution is to use a stereo system with phase-locked receivers, and process the audio in my head. With antenna separations over a few wavelengths, the background "white noise" takes on a distinct hollow sound. Signals are easier to pick out, and the ability to copy CW below the noise floor is greatly enhanced. The end effect of this is reduced fading.

For transmitting, the only useful approach is having a variety of antennas available and picking the antenna generally more optimum for the particular distance, direction, and time of day. Without feedback from the receiver, it is all a guessing game. One thing I do know is that mixing my antennas directly never resulted in improved signal strengths or reduced fading in many dozens of blind tests.

How to Check Signals

[**[Home](#)**]

This page still under construction revised 1/8/2004!!

Related pages: [Receiver tests](#)   [Key Clicks](#)   [Checking Transmitters](#)

I noticed W2VJN has proposed using absolute signal level at a fixed test spacing for click measurements. While I agree with George's suggestion we establish a test method, I strongly disagree that a 15Hz BW signal level measurement of the peaks accumulated in dozens or hundreds of scans is any more useful that listening off-frequency on a known good receiver.

Some radios, like the FT-1000 series and other Yaesu radios, click on both make and break. Radios that do that are much more disruptive than radios that click only on make or break. The slope of the roll-off is also very important.

What we really need for click measurements is a peak and average power measurement on the adjacent frequency.

My local wintertime 350Hz BW noise (after preamplifier) compared to a sample of signals on one night was:

Noise -127dBm

9H1BM -122dBm

OM0WR -95dBm

DF2PY -88dBm

WA8OLN -78dBm

W3GH -60dBm

W4ZV -42dBm

The dynamic range between noise and W4ZV on one occasion was over 85dB! It also illustrates how important antennas, location, and propagation are rather than power. The dB difference between signals from the same area is profound. Many signals run at or near noise floor. This may not be typical of every night, but it shows how large the signal level variations between weak DX and strong signals are.

W8LRL and I have heard each other while running about 30 micowatts!

There are three main reasons we hear clicks and splatter off-frequency:

- The receiver is not up to the job, but it could be
- The transmitter is not up to the job, but it could be
- The signal is just too close in frequency, and reasonable use of technology won't cure the problem

Dynamic Range (IM3)

The most useful specification for receivers is called **dynamic range**. Dynamic range (DR) tells us the ratio in dB (anything described in decibels is a ratio) of the weakest signal that can be heard to the level where problems start.

Receivers

IM3 DR would be the ratio of two equal level signals (creating a third-order product by unwanted mixing) to the noise floor of the receiver.

Imagine we have two strong CW signals spaced 1kHz apart, one at 1840kHz and another at 1839 kHz. As the level is increased, the receiving system has a increasingly non-linear response. The 2nd harmonic of 1840 can mix with the fundamental of 1839, and the result would be a new signal at 2*1840-1839=1841kHz. Another signal, if the two original signals are equal strength, appears at 2*1839-1840=1838kHz.

(There are also sums, but they fall outside the filters and tuned circuits. We normally can't hear them, so we generally just ignore them.)

The response of this product is non-linear. The level of the mixing product increases faster than the level of either individual "real" signal.

When we can hear that "phantom signal" above the noise floor of the receiver, it adds interference

to other weak signals we might be trying to hear. We reference the main signal level this occurs at to the noise floor, because that is the level where it would start to be noticeable.

Always remember actual overload is the result of the vector sum of signals in the passband of the system at that point. One way to look at it is that overload is an accumulated power problem, not an absolute level problem with one signal. This means a great number of weaker signals can cause the same problem as a few strong ones.

Transmitters

Transmitters are a special case partially covered in a checking transmitters page.

Higher Order IM

We can have higher order IM products, but they are always odd sums or differences. For example the 2nd harmonic of one signal can mix with the third harmonic of another, and we have 5th order products (2*f1 minus 3*F2 called fifth because 2+3=5).

Although they can affect CW receivers, higher order products create most of our SSB problems. This occurs because higher-order products fall well outside the filter passband of a typical SSB receiver or transmitter.

Blocking

Blocking is the point where we can detect a change in sensitivity or gain from a single strong unwanted signal. Blocking "pumps" the receiver gain and can actually make false clicks. We always have to sure any click reports are given with signal levels well below the blocking point!

Blocking again is a ratio referenced to noise floor, since loss of sensitivity (either through increased noise or decreased gain) can affect readability of weak signals.

Transmitter Splatter

[<u>Home</u>]

Related pages: <u>Checking signals</u> <u>Receivers</u>   Emission rules <u>Part 97.307</u>

Under construction 1/17/2004

Transmitter Intermodulation Distortion

A common culprit behind wide SSB signals (assuming the receiver is good) is high order distortion products. Transmitters process multiple frequencies at once. Mixing between these signals caused by amplitude response non-linearity can generate new frequencies. The offending products are those with odd-orders.

In a typical SSB transmitter the *2nd RF harmonic* of a modulating tone from one pitch can mix with the *third harmonic* of the RF caused by another modulating tone. This creates a new frequency that is called the *5th-order product* (2*F1 minus 3*F2 is called fifth because it is 2+3=5). We also can have a lower order product called the *third-order product* where the 2nd harmonic of one tone mixes with the fundamental of the other. We call that the third-order product because 2*F1 minus 1*F2 makes the new signal.

A 1850kHz LSB transmitter modulated with two tones of 500Hz and 3000Hz would have main signals at 1849.5 and 1847kHz. The third-order IM products would fall at:

(1849.5*2)-1847=1852kHz

(1847*2)-1849.5= 1844.5kHz

The fifth-order would fall at:

(1849.5*3)-(1847*2)=1854.5kHz

(1847*3)-(1849.5*2)=1842kHz

You can see every increase in order spreads the signal another (F1-F2) up and down the band. In

the above case F1-F2 is 2.5kHz. The 7th order product would be 2.5kHz above and below the 5th order products. Any odd product adds bandwidth to the signal.

If the product is direct mixing or even-order mixing, mixing would be (1*F1)-(1*F2), (3*F1)-(1*F2), (2*F1)+(2*F2), etc. The "harmonic order" in the mixing would total an even number. Let's try that:

(1849.5*1)-(1847*1)=2.5kHz outside the passband of the transmitter and antenna!

(1849.5*1)+(1847*1)=3696.5 outside TX passband.

(1849.5*3)-(1847*1)=3701.5 kHz again outside the passband of the antenna and transmitter's RF section. This was the fourth order.

Even-order isn't a problem in RF transmitters. This is why push-pull RF amplifiers don't help audible distortion and don't help bandwidth. They do cut down on harmonic distortion making band-filter requirements less.

Keep these rules in mind:

- *Any increase in frequency difference between the highest and lowest modulation frequency increases BW greatly.*
- *An increase in level increases the strength of the IM product in even greater proportion than we might expect.*

Higher odd-order products create most of our SSB problems. This occurs because high odd-order products fall well outside the normal passband of a typical SSB transmitter, spreading wasted energy to adjacent voice channels.

Three Greatest Sins

The three greatest sins creating unnecessary bandwidth are:

1. *Turning up a radio's internal power or "drive limit" pot.*
2. *Enhancing Bass and Treble at the same time*
3. *Under-loading an amplifier*

Radios

Some modern radios are especially poor, even when operated at rated power levels. For example,

the TS-2000 and IC-756 series radios are particularly bad on 160 meters. I can hear some of these radios, when signals are strong, producing weak spurious emissions (splatter) 10-20kHz away from the operating frequency. If you look at ARRL test reports of transmitters, you will see many radios are just a bit above class-C performance levels. (Remember the ARRL uses dB below PEP, which improves results by 6dB compared to commercial test methods.)

IM3 levels of -30dB are really very poor. An old KWM2 I tested was -47dB using ARRL standards. Compared to something like an IC-756, the Collins had about *50 times LESS power* in total adjacent channel distortion products!

Compression and Processing

Any type of speech processing or compression (even ALC) makes the IM problem worse, because the average level of F1 and F2 increase. It is the multiple tones in our voices that create the "original" tones that mix, so the more we put in the more that come out!

Processing is a double-edged sword, however, because it also limits the level of peaks. Decreasing the level of peaks compared to average power loads bias and power supplies more steadily, and can keep stages after filtering out of limiting or distortion. While it might bring the level of "low-level nasties" up, it also decreases the likelihood of high-level overdrive problems.

Light or modest processing is actually beneficial in reducing splatter!

ALC

ALC is normally slow and plagued with inherent problems. Filters in radios add group delay (the signal takes noticeable time to move through filters), and the ALC loop adds a time-delay of its own. The result is leading edge signal overshoot, which often shows up as an adjacent channel "spit" or "pop" on leading or rising edges of voice or CW.

Some rigs like the early 775DSP and IC706 are plagued with very high levels of overshoot. A new 775DSP I had actually overshot to around 300 watts on leading edges for a few milliseconds. Kenwoods and other radios also have this problem. This problem is not only harmful for bandwidth, it also can damage amplifiers. The problem often gets *WORSE* as power level is turned down!

Gain should be set so ALC just starts to take effect, if a drive control is available. Rigs like the FT1000D include a "Drive" adjustment.

WideFi or Enhanced Audio

Enhanced audio is a generally bad idea, since it adds and boosts unnecessary lows and highs. The audio processing brings the level of the extreme low and high frequencies up to abnormal strength. Since the frequency spread of the lows and highs is wider and the level is stronger, IM products also are wider and stronger. Enhanced or Hi-fi audio, even with perfect "brick wall" filtering, is wider than communications audio for a given RF amplifier quality.

Many of the radios popular with the HiFi crowd are among poorer radios for IM performance! My own opinion is HIFi is OK on emptier bands, but we should do all we can to discourage enhanced audio on crowded bands or near weak signal areas.

Transmitter Tests

The standard transmitter test is IM3 or higher order products. The general test uses two tones of equal level. If it is a radio, the two tones are fed into the audio port. If it is an amplifier, the tones must generally be from two separate transmitters generating steady carriers. The carriers are mixed through a combiner and used to drive the amplifier. The reason we use two separate transmitters in the test is most amplifiers, especially cathode driven triodes, are far cleaner than most modern radios. The test radio's two-tone IM would establish the IM distortion limit (not the amplifier, in most cases.

Confusing Conflicting Standards

In most commercial tests, we find the maximum power of one tone and compare that level to the level of third-order spurious signals created in the transmitter. The ARRL, for some reason, adopted a different reference. They compare the PEAK power of both tones to the spurious, rather than the level of one test tone. This inflates transmitter IM results, making everything appear 6dB better than it really is in the standard test used commercially.

If you look at Eimac data sheets, the IM specs are dB-below-one-tone of the two tones. If you look at data sheets from other sources, they might use dB below PEP. One manufacturer's US importer of Russian tetrodes used dB below PEP to compare the quality of their product to Eimac.

Unfortunately Eimac used dB below one tone, while the other tetrode test used dB below one tone, so the tetrode manufacturer wrongly proclaimed their tetrodes "cleaner" when in fact they were not. If you look at QST tests of ETO, QRO, and Ameritron amps you will see the 3CX800's in the AL800H greatly surpass the 4CX800's in the other amps for IM3 and IM5. This confusion is a clear example of how mixed standards gets us in trouble.

It's a Poor Test Anyway

Two-tone IM3 (or higher orders, like IM5) transmitter tests generally show us the very ***BEST*** a

transmitter will do. In general, two-tone tests are pretty poor tests for system designed to process speech. The two-tone test uses two steady signals (normally with wide spacing), but actual modulation has many frequencies that vary at syllabic rates.

In two-tone tests, the test spacing is generally a few kHz. The varying load on power and bias supplies is at the separation of the two frequencies. Small capacitors filter the time-varying load, while the long term dynamic load remains constant. Two-tone tests do *NOT* show power supply deficiencies.

Slow variations in speech level load and unload power and bias supplies. This causes supplies to "wobble around". The conclusion of some is that screen or bias regulation is "unimportant", but that conclusion is mostly rooted in the fact they use a very poor test method that does not show low frequency dynamic regulation problems.

A Better Test

There are two tests that are better. One test is an adjacent channel power tests, with normal voice modulation of the transmitter, another test I developed uses a three-tone signal.

The three tone test injects a third low-frequency tone into the system. Is anything from a warble to a low-pitched hum variation in power levels of the two major tones. The analyzer reads the peak amplitudes of mixing in higher frequency tones while the level is varied at a syllabic to low pitched audio rate. The low pitch amplitude modulation is varied in frequency until the worse case IM is produced.

The variation causes the power to change at a speech rate, testing supply regulation effects on wide spaced distortion at all important frequencies for speech.

The adjacent channel power test simply uses normal voice operation and compares the long term peak power in an adjacent channel to the peak power in the desired channel.

Either test gives a much more reliable indication of transmitter bandwidth than a two-tone test.

Remember, a two-tone test is generally a "best case" scenario.

Summary

A normal two-tone test is not really very effective in measuring a SSB signal, because there is no slow dynamic change typical of a voice. Voltage regulation problems are masked and do not show up when a two-tone test is used, because load currents all average the same amount. Filter capacitors and other energy storage components mask any voltage regulation problems. We can, as

a general rule, be confident the actual SSB voice performance is LESS than a two-tone test indicates.

A large improvement occurs with a three-tone test, when levels are varied at syllabic rates as well as higher frequencies in the voice range. This tests voltage regulation problems otherwise hid in a two-tone test, because all speech frequencies ranges are included. The most accurate test, of course, is with actual speech. The FCC now requires some commercial radios used on congested bands to be tested with actual speech.

We can help ease spectrum pollution by:

- **Strongly discouraging use of enhanced bass and treble on crowded bands. It has no place on crowded band or near weak signals. Wide-Fi is selfish and inconsiderate even when it uses a 3kHz filter because of the increase in level and frequency spread of IM products.**

- **Discouraging and chastising people would turn the power limit or drive limit control inside radios up. This is CB behavior! There isn't a radio made that will tolerate a user-increase in power limit without a serious degradation in IM performance. If you have a friend who peaks up the power control inside a radio, tell him why it is bad. Radios are bad enough without removing even more headroom. Even the best transistors cannot be driven more than about half of saturated power before IM becomes unacceptable.**

- **Making sure we use processing and ALC, but only at modest levels. We should just see the needles start to show compression.**

- **Making sure we [tune amplifiers correctly](), and avoiding 12 volt transistor amplifiers or grid-driven tetrode amplifiers whenever possible.**

MFJ-259B Calibration

[Home]

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Revised 8/24/03

History:

I worked with another contract engineer (my friend JB) in designing the MFJ 259 and 269. No one else outside of MFJ was involved, and certainly no one named Ted Hart.

This information is here because it is the **correct** *way to calibrate the MFJ-259B analyzer. This work is all donated. K1BQT took an instruction set supplied by MFJ and re-wrote it. I modified, expanded and edited that work. This page is the result.*

It is best that no one copy this, and start handing it out in mass. The only reason for this request is there must be a point of control of information, so it can be corrected or expanded as errors or omissions are found.

I am not aware of any other source that gives correct calibration procedures. It is important that the 259 be calibrated by these steps, even if they sound complex. If you don't do it right, don't do it. Without following these steps, many special functions may not work correctly even if the unit tests properly on calibration loads!

How This Type of Device Works

This type of analyzer contains an RF oscillator, a linear amplifier to increase power, and an internal resistor bridge in a conventional Whetstone bridge configuration.
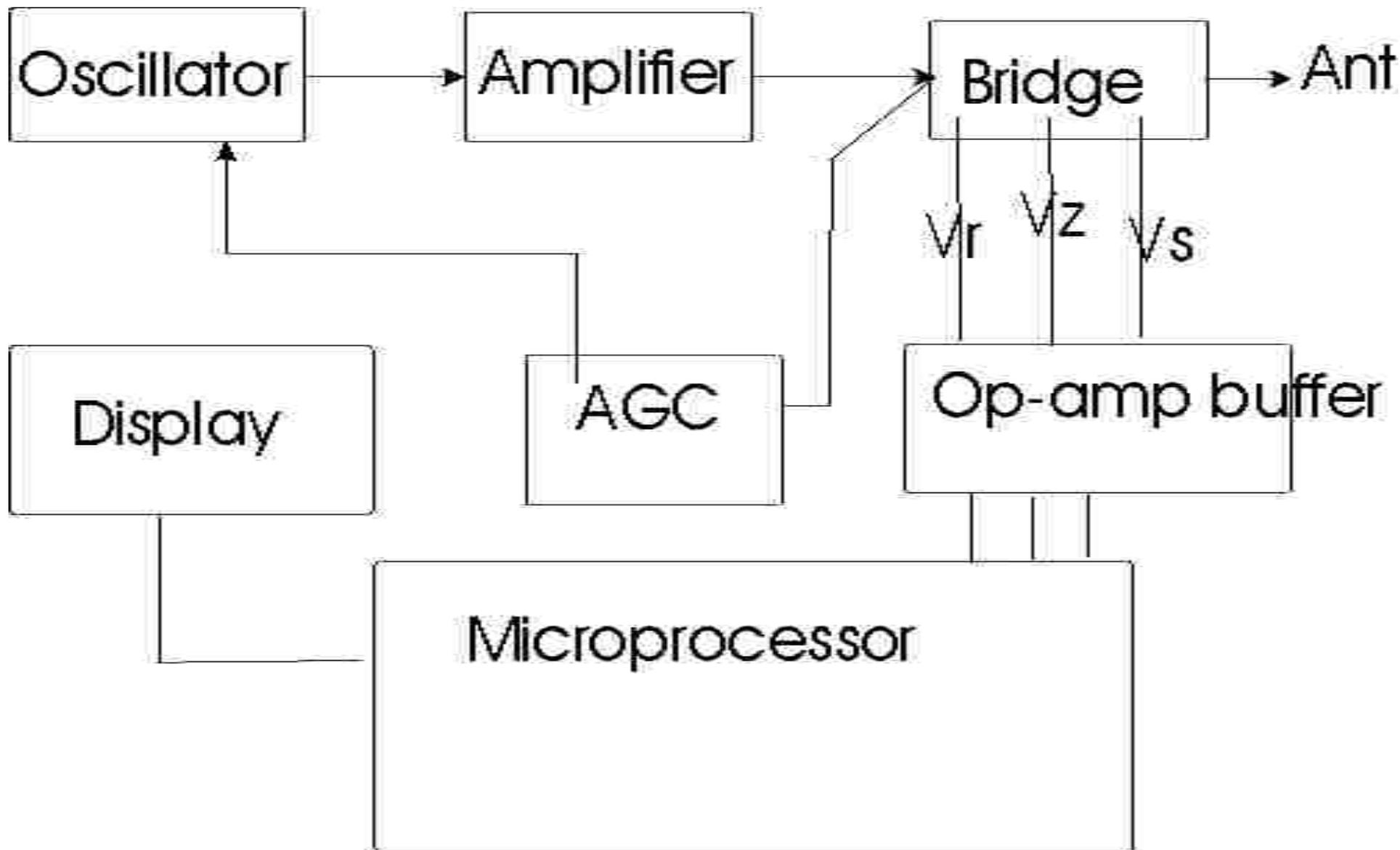
Since it is designed to be inexpensive, there are a few shortfalls with this system.

The bridge is dc-coupled from an internal resistor bridge to the antenna port. The bridge detectors are NOT frequency selective, and respond to anything from minor dc offsets through microwave signals. This causes a potential problem if there is *any* voltage appearing across the antenna port, from dc through microwave. (This is also true for competing analyzers from other manufacturers.) There are multiple reasons why, at the time of design, these units were dc coupled with broadband detectors. Hopefully someday a higher cost-design with selective detectors will become available, but for right now

this is all that is available for amateur use from any manufacturer.

The second shortfall is the internal amplifier must be linear and have very low total harmonic content. Total harmonic power, at the lowest load impedance, must be down at least 25dB and preferably 35dB. This is true for *ANY* antenna analyzer, since you do not want the analyzer to measure the load at two frequencies!

Because the detector is broadband and because it is dc coupled to the antenna, any external voltage across the input port causes measurement errors. It is the **accumulated voltage of multiple sources** that is most important, not the strength of any individual signal. Because of that, large antennas should be tested at times when propagated signals in the range of the antenna's response are at minimum strength.

A definite RFI improvement occurs with a bandpass filter, but multiple-section bandpass filters cause impedance measurement problems. Multiple-section filters behave like transmission lines of random line impedances, loss, and lengths as frequency is varied. The best solution is to use a single-stage bandpass filter and dc isolation on large arrays or with long feedlines. I often use a good 1:1 isolation transformer for measurements, and often find a parallel L/C filter (like the MFJ-731 Filter) useful.

Most Likely Failures

Other than manufacturing errors, the **detector diodes clearly stand out as the most common problem. They are the most easily damaged devices in the analyzer.** If you have a sudden problem, it is most likely a defective detector diode. Diode damage almost always comes from accidentally applying voltage on the antenna port.

Why are the diodes so sensitive? In order for the detectors to be accurate within a fraction of a percent (one bit), detector diodes must have very low capacitance and a very low threshold voltage. This means the diodes, through necessity, must be low-power zero-bias Schottky microwave detector diodes. The same characteristics that make them accurate and linear cause the diodes to be especially sensitive to damage from small voltage spikes. *ALWAYS discharge large antennas before connecting them to the analyzer! Never apply external voltages greater than 3 volts to the antenna port!*

Technical Support

From time to time MFJ gives incorrect advice on the 259B and other analyzers. One bad piece of advice that has come to my attention concerns measuring stubs. Someone in MFJ support has been telling customers that the manual is wrong, but the manual is actually correct! The person in support is wrong.

A problem with measuring stubs and cable lengths occurs if the harmonics are high in the 259 or 269, or if you do not tune for lowest X and minimum impedance and store that impedance, and then move to the VERY NEXT minimum X and lowest impedance. Whatever anyone tells you at MFJ, that is how it has to be done.

How This Unit Works

This is a rough outline of how this unit works:

The MFJ-259B, and other digitized MFJ antenna analyzers, compare three major voltages in a 50-ohm bridge circuit. They are:

Vz= Voltage across the load. This voltage is called Z in the alignment display menu

Vr= Voltage indicating bridge balance. This voltage is called R in the alignment display menu

Vs= Voltage across a 50-ohm resistor between the RF source and the load. This voltage is called S in the alignment display menu

 All voltages are converted through an eight-bit A-D converter to a 256-bit digitized output with a test-display range of 0-255 bits. By knowing the ratio of these voltages, as compared to the regulated RF source voltage, many different load parameters can be calculated. An antenna analyzer could calculate everything (except sign of reactance) from measuring only Vs and Vz, but at certain impedances any small error in either Vs and Vz becomes critical. This is especially true when voltage is digitized into a 256-bit format (~0.4% steps). At certain impedances, an almost immeasurable voltage change will cause a sudden large jump in the measured impedance parameters.

To reduce display impedance jumps, SWR is weighed into the calculation of reactance and resistance at low SWR values. (An SWR bridge is most accurate when the load is closest to 50 ohms, which is a primary measurement area where impedance measurements through Vz and Vs become critical.) By factoring in a direct SWR measurement from an internal bridge, the analyzer can check and "correct" any small level errors in Vs or Vz. This reduces the impedance jump that would occur with a one-bit jump in voltage. This also why bits must be calibrated for near-perfect accuracy. a one-bit error can cause a resistive load to appear reactive (the total of Vs and Vz must always be 255 bits or less for a load to be considered resistive).

Calibrating the MFJ-259B Antenna Analyzer

This calibration procedure is the correct procedure for later MFJ-259B's. Disregard any other information. Since MFJ-259B firmware has several versions under the same model number, you may find some final performance verification steps not valid. These steps will involve parameters that do not appear on the display.

Be sure you have printed a copy of the board layout showing adjustment points, have read all this, and have suitable loads before proceeding.

Adjustments

This unit has tracking and gain adjustments for Vz, Vs, and Vr. Tracking is set at low voltages (low bits). Gain is set at high voltages or bits. Together they make the detector voltage output closely track the actual RF voltage.

This unit also has meter calibration adjustments. The analog meters almost certainly suffer from some scale linearity problems, they will be somewhat less accurate than the digital display. These adjustments only affect analog meter readings. The meter adjustments do not affect the display.

Quiescent current (bias) in the RF amplifier section is adjustable. This adjustment directly affects output signal harmonic content. Harmonics are worse with low supply voltages, and with low impedance loads. ***Be sure you check the harmonics as outlined below, with a 1/4 wl open-circuit stub!!***

Excessive harmonics can cause severe errors in measurement of frequency-selective loads, even when dummy-load SWR tests appear perfect. Loads most sensitive to harmonic-induced errors include, but are not limited to, antenna tuners, tank circuits, very short resonant antennas, and distance to fault and stub length measurements. If you notice something "funny" going on with a stub measurement, it may be a fault of incorrect bias.

Warning: **Never calibrate around a sudden "problem" that appears. If a detector suddenly shifts voltage, the problem is almost certainly a defective detector diode. If the meter is recalibrated with a defective (leaky) diode, the meter will probably NOT track correctly with frequency.**

Alignment

Tools and Equipment:

[ ] #2 and #1 Phillips-head screwdrivers

[ ] Digital meter or accurate analog meter for checking supply voltage

[ ] Small set of non-metallic alignment wands for coils, or small jeweler's screwdrivers for controls

[ ] Power supply, regulated to 12-volts + - 5%

[ ] General-coverage receiver with level meter or Spectrum Analyzer (these are optional with additional work and use of a stub)

[ ] ~10 MHz 1/4wl open-stub, for example 15' good-quality solid-dielectric RG-8, UHF connector at on end, open on other (not needed with analyzer or receiver)

[ ] 2.2-ohm 1/4 or 1/2 watt film resistor (not needed with stub)

[ ] Accurate load set to include:

    A.   Short

    B.   12.5-Ω load

C. 50-Ω load

D. 75-Ω load

E. 100-Ω load

F. 200-Ω load

Note 1: Loads must be constructed using physically small 1% carbon-film resistors.

- **DO NOT use large resistors. Acceptable results will be obtained when load resistors are mounted in the very bottom of a UHF-male connector.**

- **The ideal load resistors are surface-mount precision-resistors, but other styles will work. It is acceptable to parallel multiple resistors to obtain low resistances, but don't series connect more than two resistors!**

- **Never use physically large resistors, such as 1-watt or larger resistors, unless you are absolutely positive they are composition-types (very rare).**

- **Since the loads are used to set the number of bits in critical calculations, the maximum reactance error will always be worse than the percentage of resistive load error. *A one bit error in calibration (~ .4%) can cause a purely resistive load to read reactive.***

Quick-connect loads can be made with surface mount resistors on a BNC male chassis mount connector with the bayonet removed. This makes a "quick connect" connector that will slide directly into a type-N female. In this case, use a good UHF to BNC female adaptor for the 259 units. With a 269, the load will plug directly in to the N female.

Note 2: The power source should be the LOWEST expected operating voltage. DO NOT use a standard "wall-wart" or batteries! You can reduce voltage from a conventional 13.8v regulated supply by adding a few series diodes. Silicon diodes will normally drop about 0.6volts or so per diode. Three or four diodes will reduce place the voltage below 12 volts.

WARNING: The MFJ-1315 AC adapter or other "wall-warts" should NOT be used to power the unit for most alignment steps.

Step 1

Visual Inspection: Before, during, and after calibration, be mindful of physical condition. Watch for missing or loose hardware. Do not tug, stress, or repeatedly flex leads, or carelessly flop or toss things about. Keep your bench clean. Follow these rules the entire time you have the unit apart!

Step 2

Battery Tray Removal: This step provides access to trim-pots and most inductor adjustments.

[ ] Remove last two batteries at each end of the tray.

[ ] Remove two screws (right side) and extract the tray.

[ ] Always position the battery tray to minimize strain on wires.

Refer to the board layout pictorial for specific control locations.

Step 3

Band Overlapping: Each band should overlap the next by a small amount to ensure gap-free coverage from 1.8 MHz to 170 MHz. While viewing the *LCD Frequency Display*, wiggle the bandswitch from side-to-side gently. Watch for any display or meter dropout. Check each band as follows:

[ ] **114-170 MHz:** Oscillator tunes from below 114.0 MHz to above 170.0 MHz. Check tune for dead spots.

[ ] **70-114 MHz:** Oscillator tunes from below 70.0 MHz to above 114.0 MHz

[ ] **27-70 MHz:** Oscillator tunes from below 27.0 MHz to above 70.0 MHz.

[ ] **10-27 MHz:** Oscillator tunes from below 10.0 MHz to above 27.0 MHz.

[ ] **4-10 MHz:** Oscillator tunes from below 4.0 MHz to above 10.0 MHz.

[ ] **1.8-4 MHz:** Oscillator tunes from below 1.8 MHz to above 4.0 MHz. Check tune for dead spots.

*While verifying overlap, at least check the lowest and highest bands carefully for dead spots. The LCD Display will indicate 000.000MHz if a dead spot occurs. Dead spots generally indicate a defective tuning capacitor (**TUNE**).*

If you find wiggling the switch causes a dropout, the switch may have dry or dirty contacts. Less likely are poor solder joints, but check solder joints first. If you must clean and lubricate the switch, be aware it is a difficult task. The entire board needs to be lifted from the case front. Dirty band-switch contacts may be restored with spray tuner cleaners. The best place to spray the switch is from the front side (shaft side), right below the nut. You must remove the switch indexing tab retainer nut and the metal switch retainer (stop) under the nut. Be sure the stop goes back exactly as removed.

To correct overlap problems, locate and retune the appropriate VFO coil (see Pictorial for coil locations). Note that L1-L4 are slug-tuned and require an insulated hex-head tuning wand. If you use the wrong size wand or a worn wand, it will break a slug!

Inductors L5 and L6 are located on the component side of the board and are compression-tuned (press turns closer together to lower frequency or spread apart to raise frequency). Make only very small corrections--*especially to L5 or L6*--and recheck the band you are adjusting. You should also check the *next lower band* after each adjustment to ensure that the lower band hasn't moved excessively.

Important Warning: VFO coils MUST be aligned from highest frequency to lowest frequency. The next higher range affects next lower band the greatest amount. Do not attempt VFO coil adjustment unless you are experienced working with VHF-LC circuitry or complex alignment procedures.

Step 4

Harmonic Suppression/ Bias: Connect the analyzer *exactly* as shown below.

- The impedance of the cable to the measurement device should match the impedance of the measurement device.
- The "T" must be connected either directly to or placed within a few inches of the analyzer.
- The power source must be the lowest expected operating voltage.
- The measurement device must be well-shielded, and not pick up any substantial signal from the analyzer when the "T" is disconnected from the analyzer.

Stub directly to analyzer!
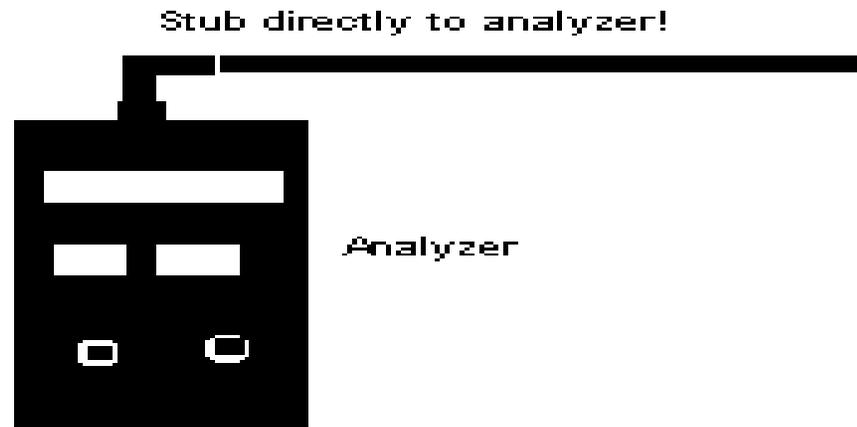
Analyzer

Test Setup Using Internal V2 for Harmonics

Any length

To Measurement Device

2.2 ohm
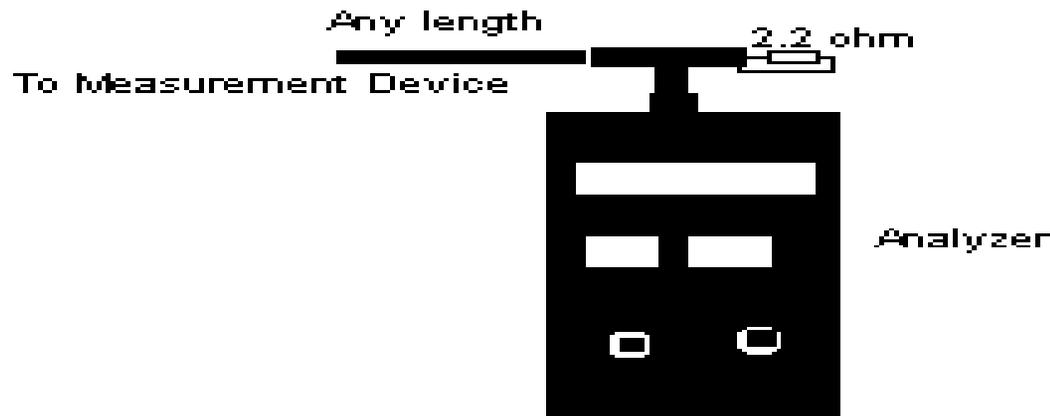
Analyzer

Test Setup using External Analyzer and resistor for Harmonics

Step 5

Harmonic Suppression (bias R89): This adjustment reduces oscillator harmonics. Harmonics will cause incorrect readings under some load conditions.

WARNING: *Incorrect adjustment of R89 will NOT show when checking with resistive dummy loads!!! The unit will appear to calibrate correctly, but will produce errors in stub length, distance-to-fault, and other frequency selective functions.*

When R89 is set properly, harmonic suppression of –30 to –35dBc should be possible across most of the analyzer's tuning range. This particular

adjustment must be made at the lowest expected operating voltage. Proper alignment requires a 12.0-volt regulated supply as a power source. NEVER use an AC adapter or any supply voltage higher than 12-volts when making this adjustment. A calibrated spectrum analyzer works best for monitoring harmonic output, but a well-shielded general-coverage receiver with signal-level meter will also work. The receiver MUST be "T'd" into the analyzer just as the spectrum analyzer is, and the Tee and resistor must be located at the analyzer connector. If you do not have a good-quality receiver or spectrum analyzer, use the test mode of the analyzer with a stub. Watch MFJ analyzer test-mode Vz. Test-mode Vz will roughly indicate total harmonic voltage when the analyzer is set at the stub's exact resonant frequency. Entering the test mode is described in Detector Calibration (Step 6).

[ ] **a.** Install either a 15' RG-8 open stub, or resistor and measurement device, and tune analyzer to approximately **10 MHz**.

[ ] **b. (stub and internal Vz use only)** Observing Vz on the data display (analyzer test mode), adjust frequency until the *lowest* fundamental output reading (or lowest impedance) is obtained. You should clearly see the MFJ analyzer's fundamental frequency output voltage (Vz) go through a deep null.

[ ] **c.** Observe the analyzer frequency reading. This is the approximate resonant frequency of the stub, and the <u>test frequency</u>.

[ ] **d.** Without changing the analyzer <u>test frequency</u> setting, observe the second harmonic level. This harmonic will be at twice the MFJ analyzer frequency counter reading**..**

[ ] **e.** Adjust **R89** for **lowest 2nd harmonic meter reading** on the receiver, lowest Vz test-menu reading, or lowest harmonic level on the spectrum analyzer. Be SURE the fundamental frequency level remains nulled in the analyzer.

WARNING: Always repeat steps (b) through (e) at least one extra time when relying on display Vz. The original null point of any stub will shift if there is a substantial reduction in harmonics after R89 is adjusted. The original stub frequency, as observed at (c), will probably change slightly. It is NOT necessary to recheck when doing a resistor load test with a good-quality spectrum analyzer or receiver. With a resistor, exact test frequency is NOT critical.

NOTE: If you have a poorly performing spectrum analyzer or receiver with limited dynamic range, use a stub with the spectrum analyzer or receiver instead of a 2.2 ohm resistor. If you have a reasonable quality spectrum analyzer or receiver (at least 50dB dynamic range) use a 2.2-ohm non-inductive resistor in lieu of the stub, resistor adjustment is easier and more accurate.

Detector Calibration

Step 6:

This critical sequence calibrates A-D conversion for various load conditions. If you know your unit has been tampered with, preset trim pots **R88**, **R89**, and **R90** to their center positions before continuing. If you find any control bottoms-out in adjustment, you almost certainly have installed an incorrect load or the analyzer has a defective detector diode.

To prepare for detector tracking alignment, place the analyzer in *Test Mode*. Entering test mode may be tricky with some units, and it may take practice. To enter *Test Mode*:

[ ] Turn power off.

[ ] Hold down *MODE* and *GATE* buttons while restoring power.

[ ] As display comes up, slowly (about 1 second period) rock between pushing the **MODE** and **GATE** switches alternately (the best method is to use two fingers, and rock your hand from side to side between the two buttons)

[ ] Confirm analyzer has entered test mode (it may take more than one try).

[ ] Using the **MODE** button, advance display to the **R-S-Z** screen (shown below).

Note: **If you go past the R-S-Z screen, you can still see R-S-Z by pushing and holding the MODE button.**

10.000 MHz

Rxxx   Sxxx   Zxxx
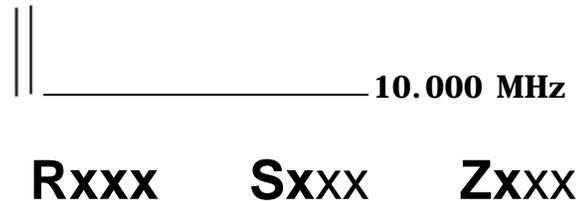
[ ] Tune analyzer operating frequency to approximately **10.000 MHz**

[ ] Leave antenna connector **Open**

[ ] Set **R72** for **Z=255**

[ ] Install the **Short**

[ ] Set **R73** for **S=255**

[ ] Install **12.5-Ω** load

[ ] Set **R90** for **Z=051**

[ ] Set **R53** for **R=153**

[ ] Install **200-**Ω load

[ ] Set **R88** for **S=051**

[ ] Set **R72** for **Z=204**

[ ] Install **75-**Ω load

[ ] Set **R89** for **R=051**

[ ] Install **12.5-**Ω load

[ ] Reset **R90** for **Z=051**

[ ] Set **R73** for **S=204**

[ ] Reset **R53** for **R=153**

[ ] Install **200-**Ω load

[ ] Reset **R88** for **S=051**

[ ] Verify or set **Z=204**

[ ] Install **75-**Ω load

[ ] Reset **R89** for **R=051**

Important Note: Small single-turn trimpots are touchy to adjust and tracking settings are somewhat interactive. If specified readings aren't obtained on the run-through, repeat the sequence a second time (accuracy counts). When the sequence is complete, turn power off to remove the analyzer from *Test Mode*.

Be particularly mindful of the total bits of Vz and Vs. If the sum of these bits ever exceeds 255 with a resistive load, the analyzer

will indicate reactance.

Display Test and Analog Meter Calibration

Step 7:

This step sequence checks meter calibration and verifies accuracy of the *LCD Display* information.

Remove and re-apply power and enter the *Real-Imaginary* impedance mode **R-X**. Readings + or - 10% of reading or + or - 5 ohms of display, whichever is larger, are considered within design specifications. Typically digital readings are almost perfect with proper detector calibration. Analog meter readings may be outside that range, and as much as 20% off with some load values.

[ ] Install **75-Ω** load

[ ] Verify reading of **R= 75 X=0** on *LCD Display* (±10%)

[ ] Install **50-Ω** load

[ ] Verify reading of **R=50 X=0** on *LCD Display* (±10%)

[ ] Set **R67** for reading of **50** on the *Impedance Meter*.

[ ] Verify reading of **1.0** *SWR Meter* (no deflection).

[ ] Install **Open** load

[ ] Verify reading above **400** on *Impedance Meter*

[ ] Install **100-Ω** load

[ ] Verify reading of **R=100 X=0** on *LCD Display* (±10%)

[ ] Verify reading of **100** on *Impedance meter* (approximate).

[ ] Set **R56** for a reading of **2** (2:1) on the *SWR Meter*

[ ] Install **12.5-Ω** load

[ ] Verify a reading of **4:1** SWR on *LCD display* (3.8-4.2 good)

[ ] Verify reading of **>3** (greater than 3:1) on *SWR Meter*

[ ] Install **200-Ω** load

[ ] Verify reading of **4:1** SWR on *LCD display* (3.8-4.2 good)

[ ] Verify reading of **>3** (greater than 3:1) on *SWR Meter*

Capacitance Mode Check

Step 8:

If you have a few precision capacitors, you can verify the calibration between the ranges of 100 and 5000 pF. Read the analyzer manual for details of capacitor measurement.

[ ] Install no load

[ ] Set *Mode* to *Capacitance*

[ ] Set VFO to **70 MHz**

[ ] Verify **4-6 pF** reading on LCD Display

Frequency Counter Check

Step 9:

These steps verify accuracy of the counter. Note that the counter's clock isn't user-accessible, so no adjustments will be made. To conduct this test, use a general-coverage receiver in AM mode.

[ ] Tune in **WWV** on 5.0,10.0,15.0, or 20.0 MHz (frequency with best reception).

[ ] Install a short clip lead or wire in the analyzer's *Antenna* jack.

[ ] Turn on the analyzer and **zero-beat the WWV signal** as closely as possible.

[ ] Compare *LCD Display* reading to the WWV frequency being used.

[ ] Verify agreement is within ±**5 kHz.**

Advanced Modes Check

Step 10:

This sequence verifies operation of the analyzer's advanced features. To enter *Advanced Mode* menu:

[ ] Turn unit off.

[ ] Hold down the MODE and GATE switches while turning power on.

[ ] Verify *"Advanced"* appears on the *LCD Display*.

[ ] Install **Open** load

[ ] Tune VFO to >**170 MHz** (top end of coverage range)

[ ] Verify **Z** = <**650** Ω with (about) **90°** phase shift

[ ] Install RG-8 open stub

[ ] Tune VFO for minimum **Z** reading (around 10 MHz)

[ ] Verify **Z-min** = **0** to **2** Ω

[ ] Install **50-Ω** load

[ ] Set VFO to **1.8 MHz**

[ ] Verify **Z = 50 Ω**, θ **= 0°**, and **SWR = 1** (±10%)

[ ] Enter **RL** *Mode* (return loss)

[ ] Verify **RL** = **>42 dB**, ρ **= 0**, **SWR = 1**

[ ] Enter *Match Efficiency Mode* (skipping *DTF Mode*)

[ ] Verify **ME** ≅ **100%** (approximate)

[ ] Press and hold MODE and GATE buttons to restore *Main Modes*

[ ] Remove load and verify **Z = >650** on *LCD Display*

Conclusion:

Step 11:

[ ] Reinstall battery tray

[ ] Confirm charger jumper is set for type of batteries used (disable for alkaline).

[ ] Reinstall cover

 This completes calibration.

MFJ-259B Calibration Checklist

Make a copy and check each box as you proceed down the calibration list.

Physical Condition

[ ] Hardware, batteries okay

Harmonic Check

[ ] Suppression -35 dBc or better

Overlapping

[ ] All bands have sufficient overlap

Check Digital Calibration at 10 MHz in test mode.

[ ] <u>**Open**</u>

[ ] R72 for Z=255

[ ] <u>**Short**</u>

[ ] R73 for S=255

[ ] <u>**12.5-Ω**</u>

[ ] R90 for Z=051

[ ] R53 for R=153

[ ] <u>**200-Ω**</u>

[ ] R88 for S=051

[ ] R72 for Z=204

[ ] **75-**Ω

[ ] R89 for R=051

[ ] **12.5-**Ω

[ ] R90 for Z=051

[ ] R73 for S=204

[ ] R53 for R=153

[ ] **200-**Ω

[ ] R88 for S=051

[ ] Verify Z=204

[ ] **75-**Ω

[ ] R89 for R=051

Analog Calibration: 10 MHz, values approximate

[ ] 75-Ω

[ ] Verify R= 75 X=0

[ ] **50-**Ω

[ ] Verify R=50 X=0

[ ] Set R67 for 50 on Imp Meter.

[ ] Verify 1.0 on SWR Meter

[ ] **Open**

[ ] Verify >400 on Imp Meter

[ ] **100-Ω**

[ ] Verify **R=100 X=0**

[ ] Verify 100 on Imp Meter

[ ] R56 for 2 (2:1) on SWR Meter

[ ] **12.5-Ω**

[ ] Verify 4:1 on LCD (3.8-4.2)

[ ] Verify >3 on SWR Meter

[ ] **200-Ω**

[ ] Verify 4:1 on LCD (3.8-4.2)

[ ] Verify >3 on SWR Meter

Capacitance Mode Check

[ ] **Open**

[ ] Set VFO to 70 MHz

[ ] Set Mode to *Capacitance*

[ ] Verify C≅4-pF

Counter Check

[ ] Counter Okay

Advanced Modes

[ ] Tune to 170 MHz

[ ] **<u>Open</u>**

[ ] Verify <650, Phase ≅ 90°

[ ] **<u>3' RG-58</u>**

[ ] Tune for Z-min (≅150 MHz)

[ ] Verify Z= 0-2 Ω

[ ] **<u>50-Ω</u>**

[ ] Tune to 1.8 MHz

[ ] Verify Z=50 Ω, θ=0°, SWR=1

[ ] Advance to Return Loss

[ ] Verify RL=>42dB, ρ=0, SWR=1

[ ] Advance to Match Efficiency

[ ] Verify ME $\cong$ 100%

[ ] Restore Main Modes

[ ] **<u>Open</u>**

[ ] Verify Z=>650

End of Procedure

Pictorial Diagram of Analyzer Board

Locations for trimpots and inductors

Charger
On Off

R90   R89

Binary Gain
Levels

R73

R72   R88

R53

Impdance
Meter Sens   SWR Meter
Sens

R67

R56

Harmonic
Suppression

R84

L6
(114-170)

L5
(70-114)

VFO Coils
(Adj Overlap)

L4

(27-70)   L3   L2   L1

(10-27) (4-10) (1.8-4)
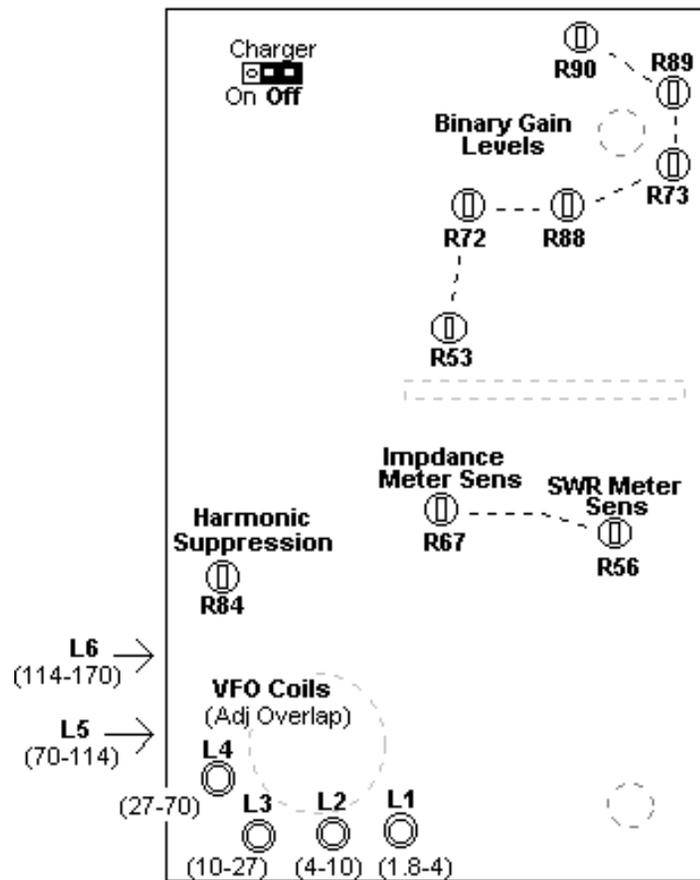
Loads Using Standard-Value Resistors
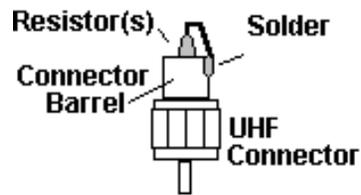
- Install resistors all the way down in the connector, the goal is zero lead length

.

- Use precision 1% carbon or metal film 1/8th-1/4 watt resistors

12.5Ω = (4) 50-ohm or a single 15Ω and 82Ω 1% in parallel

50Ω = 49.9-ohm or 100Ω and 100Ω in parallel

75Ω = 75-ohm or 150Ω and 150Ω in parallel

100Ω = 100Ω

200Ω = 200-ohm or 100Ω + 100Ω in series

Important Note: *These simple HF loads will not always be accurate for SWR checks above 30 MHz. Only precision terminations should be used in the VHF region, and even then there can be some errors. The MFJ-259B does not correct for connector impedance bumps or the electrical length between an external load and the bridge inside the unit.*

RF Noise Powerstroke Diesel

[[Home](#)]

Related pages:

[mobile antennas](#) [Loading coils](#) [Field Strength Measurements](#)

RFI measurements showed HF noise at a distance of 5 meters from my 2003 Power Stroke was 15-20dB less than my 1995 F-250. I never could get all the noise out of my 7.3L 1995 F-250 HD PSD. The measurements of noise field a small distance from the truck gave me great hope I could have a better HF mobile with minimal work.

I am delighted to report my 2003 F-250HD PowerStroke diesel has virtually no RFI with only very simple and fast corrections. Be aware this is for my antenna mounting location, and my style of truck. This is a new old-stock truck I found at a small rural dealer for a *very* good price!

Antenna Mounting Location

I mounted the antenna in the stake hole of the bed that is located just behind the cab on the driver's side. I use this location because:

1. I do not want to drill holes in a new truck
2. I have no topper or camper shell to mount the antenna to
3. I want maximum performance
4. I would like to remove the mount if necessary

Antenna Mount

My antenna mount is home made. It is formed from a scrap sheet of stainless steel that looks to be about 7 gauge. In order to make the mount, I went though the following process:

1. bent two angle-sticks of metal that match the rectangular dimensions of the stake hole
2. TIG welded a 3/4inch nut inside one L to match the height of the existing Ford bolt hole
3. TIG welded the L's into a rectangular tube
4. TIG welded a few sharp points on the tube near the bolt hole to break the paint for a ground connection
5. bent a wide long plate into an L that matched the size of a Tarheel mount
6. added holes to match the Tarheel plate
7. TIG welded the bent plate to the end of the tube
8. added gussets to brace the bend, since that was a weak area that allowed the antenna to flop around too much
9. cut a 4" square load spreader out of 3/16 inch thick scrap stainless
10. drilled a hole in the load spreader plate
11. cut square of rubber as a buffer to prevent damage to the trick's bed rail top

You can see where I get my shield ground connection. I feed the antenna with the single wire to the lug normally used to mount the impedance correction coil. I mount my impedance correction coils on PL-259 plugs so I can change them easily. I screw them into the SO-239 connector normally used as a feed!

Noise Noise Noise

When the antenna was installed, I had S-7 injector noise on my IC-706. I cured it with the following:

I removed the Tarheel motor line choke bead because it was far too low an impedance. I was receiving on that lead, and it was bringing RF back into the cab, I use my own jumbo size 44 mix bead.

You will probably find most frame-type vehicles mount passenger compartments on rubber mounts to reduce noise and vibration. The cab is only grounded at the front. The bed is bolted solidly to the frame, and this means all RF currents must travel down the bed to the frame, follow the frame forward to the front of the cab, flow back along the cab towards the antenna. This actually makes a very short thick antenna out of the truck frame and cab of the truck! The common mode currents excite all the wiring under the vehicle, increase ground losses under the truck, and increase noise ingress from truck wiring into the antenna.

To cure the common mode current problems, I added a ground strap between the bed and the cab.

Bridge across this gap!

You can make the connection UNDER the chassis at the locations shown in the following picture:

Note that I heavily tin the edges of the 1" braid. I do this insure all of the braided wires stay fully in contact all of the time. The flex area must remain solder free.

I use stainless thread cutting screws, and stainless external tooth star washers under the braid in the tinned area to cut through the paint.

Adding this single strap made the following changes:

The base resistance of the antenna dropped from 45 ohms to 25 ohms on 7 MHz. This mean I greatly reduced radiation from the truck body or frame.

Noise level dropped 2-3 S units

RF in truck wiring dropped substantially

Remaining Noise

A small amount of noise remained audible after the grounding. I traced this noise to radiation from the exhaust system. The engine block was

hot with RF, the injector control modules are mounted on the block and have leads that leave to other wiring in the truck. This lets the injector modules "push" against the wiring in the truck, making the poorly grounded engine block pump up and down with pulse voltages. Since the exhaust system bolts to the engine with a direct conductive path through the turbocharger, the tailpipe is actually excited just like an antenna.

I cured ALL of the remaining noise with the addition of one more ground strap between the exhaust and the frame just ahead of the muffler.

You can see the widening of the pipe as it expands into the muffler, with the fuel tank and transfer case in the background. The strap grounds under a nut used for the exhaust hanger. As before, I tin the leads and use stainless star washers to help maintain connections. Notice I clamped the strap under the factory exhaust joint clamp.

I now have absolutely no noise at all from the diesel engine on any band.

781 **since July 5 2004**

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My Shack

[Home]

FT1000D

IC751A's

Modified R4C's (phase locked to each other)

AL800 on 6 meters

AL1200 for FT1000

AL1500 for IC751A

My Shack

Skin depth and Conductors

[Home]

The following short article is about Litz wire, braid, and skin depth. It is still being expanded.

The best explanation of skin depth I have seen is in Circuits and Networks (by Koehler) on page 196:

196	TRANSMISSION-LINE PARAMETERS	8

conductor with return path at a distance equal to several times the diameter of the conductor. Assume that an emf is applied between the two ends of the conductor and further assume that the conductor is divided into many concentric cylinders of equal cross-sectional area. Now if the current density were uniform over the entire cross section, the total current in each cylindrical shell would be the same. The magnetic flux encircling the innermost cylinder would be larger than that of the outermost cylinder. Since the voltage drop along each of the cylinders is made up of a resistance drop (IR) plus a reactance drop (jIX), the reactance drop along the innermost cylinder would be higher than that of the outermost cylinder because of the greater flux. Obviously this cannot be true because the voltage drop across all cylinders must be the same, i.e., the impressed voltage. Consequently, the current density changes in such a manner as to equalize all voltage drops. This results in a higher current density toward the outside surface of the conductor. This effect, called skin effect, is a function of the frequency, permeability, conductivity, and diameter of the conductor. It causes the conductor to have a higher ac resistance than dc resistance because it effectively reduces the size of the conductor by virtue of the decrease in current density from the outside to the inside portion of the conductor. The current not only decreases in magnitude but also changes in phase with reference to the current at the surface. At high frequencies, dependent upon the size of the conductor and its electrical properties, most of the current may flow only in a very thin outer cylinder.

An exact mathematical analysis of current distribution in a cylindrical conductor shows that the current density is given in terms of Bessel functions of a complex argument that involves the distance from the center divided by δ where $\delta = 1/\sqrt{\pi f \mu \sigma}$; μ is the permeability and σ is the conductivity. When δ is very small compared to the radius of the

the conductivity. When δ is very small compared to the radius of the conductor the ac resistance of the conductor is substantially the same as the dc resistance of a conductor with a cross-sectional area equal to $2\pi r\delta$, where r is the radius of the conductor.

When δ is not small compared to r, it cannot be assumed that the entire current flows in a cylinder of radius r and thickness δ. For large values of δ/r, an exact solution* for the current distribution results in the curve of Fig. 8-3, where R_{ac}/R_{dc} is shown as a function of r/δ. This curve indicates that when $r/\delta \leq 1$, the ac and dc resistances are substantially equal. Thus if it desired to have the resistance of a conductor

So we see, skin effect occurs because the outer areas of a conductor have less magnetic flux surrounding them than the inner layers. While we often consider the effect on materials without weaves, think about how this affects woven conductors like braiding or Litz wire. Any individual conductor has HIGHER impedance when it moves into the center areas, and if current finds an alternate path it moves out to the lower-impedance outer layer.

If strands touch with high resistance connections, it effectively adds unnecessary resistance to the path. The fact the surface is rough also DECREASES effective surface area for current for a given occupied physical area, because there are non-conductive air gaps in the occupied area!

The increased resistance (and impedance) is why braiding should never be used for high frequency high current applications unless it is oversized to compensate for increased impedance and dissipation. Parallel stranding decreases effective current carrying surface area at high frequencies, but at least avoids the weave problems.

As a matter of fact the weave is what causes most of the loss increase when typical coaxial cables are contaminated by water. The problem occurs, even after the cable is dry, because the hundreds of weave contact points corrode and make poor connections, increasing the resistance. Something similar happens when we "unpack" a braid from a jacket. The pressure between weaves decreases, and series resistance (and impedance) increases!

We could insulate strands, and once we establish a basic loss in conductivity caused by replacing useful conductor area with insulation we would find the impedance slope much flatter with frequency. If the strands are insulated (Litz wire) and frequency is increased, eventually we reach a point where the conductor resistance increases and we are much worse off than we would be with braided or stranded wire. The upper limit point is generally around 1 MHz for Litz wire.

From Radio Engineering by Terman:

Litz Wire.—The effective alternating-current resistance of a conductor can be made to approach the direct-current resistance at low and moderate radio frequencies, by forming the conductor from a number of strands of small enameled wires connected in parallel at their ends, but insulated throughout the rest of their length, and thoroughly interwoven. If the stranding is properly done, each wire will, on the average, link with the same number of flux lines as every other wire, and the current will divide evenly among the strands. If at the same time each strand is of small diameter, it will have relatively little skin effect over its cross section, so all of the material is equally effective in carrying the current. Such a stranded cable is called a *litz* conductor.

Practical litz conductors are very effective at frequencies below about 1000 kc, but as the frequency becomes higher the benefits disappear. This is because irregularities of stranding, and capacitance between the strands, cause a failure to realize the ideal condition at very high frequencies.

In reality, the real advantage of Litz wire is NOT reduced resistance per unit length for a given diameter. The advantage is less slope in resistance with frequency and reduced eddy current when the conductor is in a multi-layer coil or transformer. The individual strands are like laminations in a transformer core, and below a certain frequency they greatly reduce eddy currents by decreasing the "short circuit" path distance for magnetic flux induced currents that are not in the normal current flow direction.

My own measurements have shown noticeably lower ESR (equivalent series resistance) for a given conductor size when using a **SOLID** wire as compared to any Litz wire sample I have tried above 300kHz (the lower limit of my main vector network analyzers).

For a given wire diameter and form factor, I've always been able to achieve higher Q with solid round or ribbon conductors.

Also see mobile and loaded antennas.

stacking broadside collinear

[[Home](#)]

Popular rumor is gain doubles each time elements are doubled. In other words, adding a second vertical above another vertical would add 3dB gain. Adding two more after that (total of four) produces 6dB gain over a single element.

The same concept is often applied to stacked Yagi or dipole antennas.

In neither case is it true. Gain comes from forcing nulls in areas of high radiation. This removes energy from the null areas, and that energy moves to enhance gain in other directions. This effect is sometimes called ***pattern multiplication***.

Before antenna modeling software was commonly available, we often used pattern multiplication to estimate patterns and gain. It still is a useful tool.

The graphs below, from Jasik's Antenna Engineering Handbook, shows the gain of various couplets or elements when placed either end-to-end (Collinear) or parallel above each other (stacked broadside).

Collinear Gain

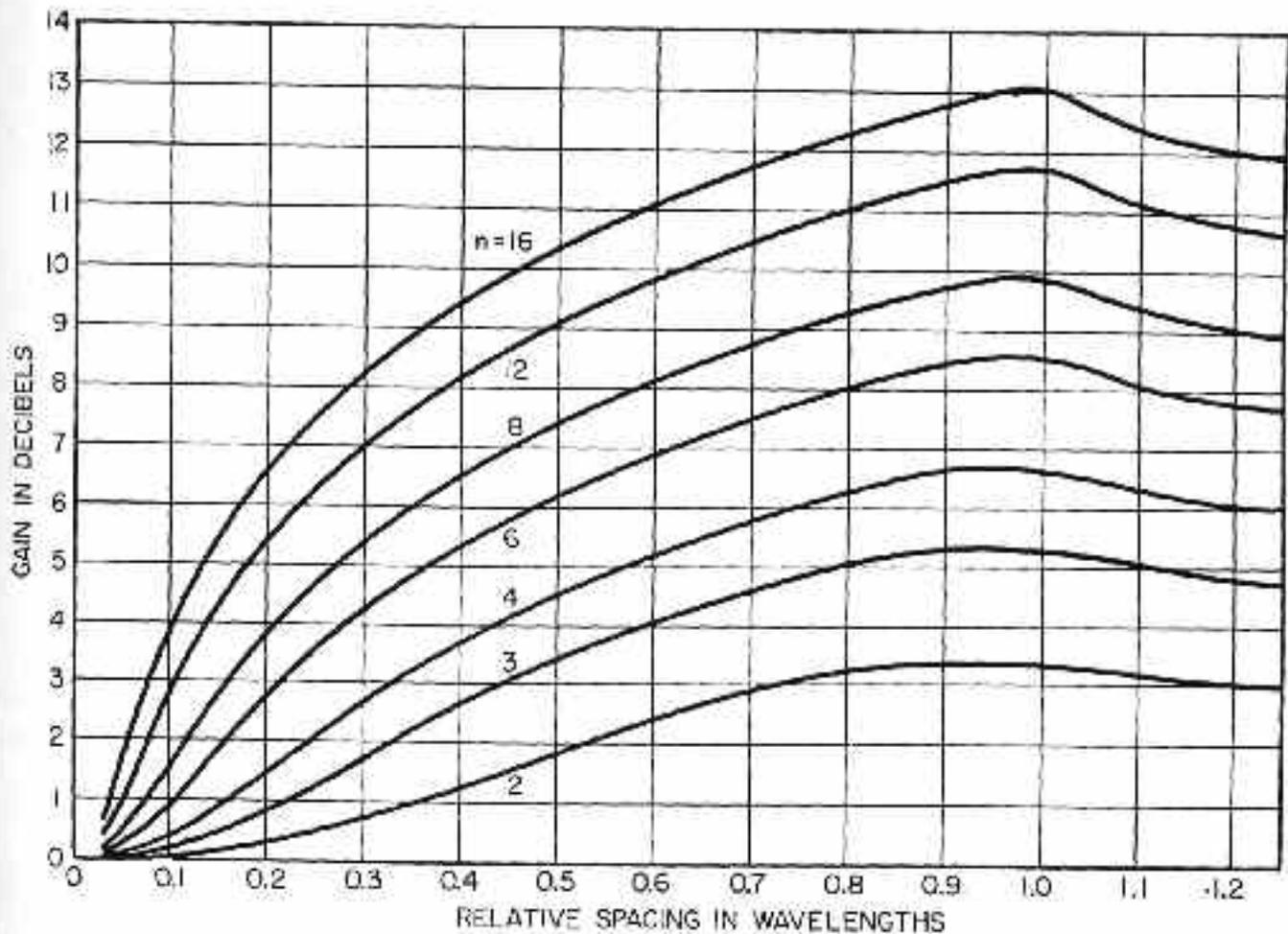First we have the end-to-end element placement gain.

FIG. 5-22. Gains to be expected from a collinear (omnidirectional) array of short-dipole elements. The gain is relative to that of a single element.

We can see the gain for two elements peaks at .9 wavelength spacing. This spacing is the current maximum spacing of the elements, NOT the end spacing. With a 1/2 wl dipole in each element, the end-to-end spacing would be .9 - .25 -.25 = .4 wavelengths. The overall array length would be .9 + .25 + .25 = 1.4wl

With two dipoles end-to-end the center-to-center spacing of current maximums would be .25+.25=.5wl. The absolute maximum gain would be found on the graph above at the crossing of the vertical .5 relative spacing line and intersection of curve 2 (two elements). The gain would be less than 1.9dB in ANY collinear antenna.

We would achieve 3dB at about .71-.25-.25= .21wl tip-to-tip spacing, or 1.21wl total collinear element array length.

To double gain again (adding 3dB more) the array would have to be four elements with .75wl center-to-center spacing in elements! The array would be 3*.75=2.25wl (this is number of elements minus 1 multiplied by required spacing) for the total spacing of four elements). The end-elements would extend .25 wl each from the current maximum, so the array length from end-to-end would be

2.25+.25+.25=2.75wl.

To go from three to six dB requires we change antenna length from 1.21wl to 2.75wl!

We cannot simply double length to double gain! That concept is wrong.

Note we can get more than 6dB gain by using more than .75wl element center-to- element center spacing. We really should have .95wl element center-to element center spacing, making the array N-1*S + El = L where N=number of elements, S=spacing, El = element length. The result is an array of four elements would have optimum gain of 6.7dB with a length of 3.35wl.

The longer the array is, the wider the individual elements should be spaced for optimum gain.

Broadside

Broadside gain applies to elements that are parallel and one above the other. This could be Yagis or dipoles. This is the *OPTIMUM or maximum gain*, not the actually gain you might have.

Optimum broadside stacking distance increases with more directive elements, so a pair of three-element Yagis stacked requires wider spacing than a pair of dipoles, and less maximum stacking gain is possible.

Here is the optimum gain graph for dipoles:
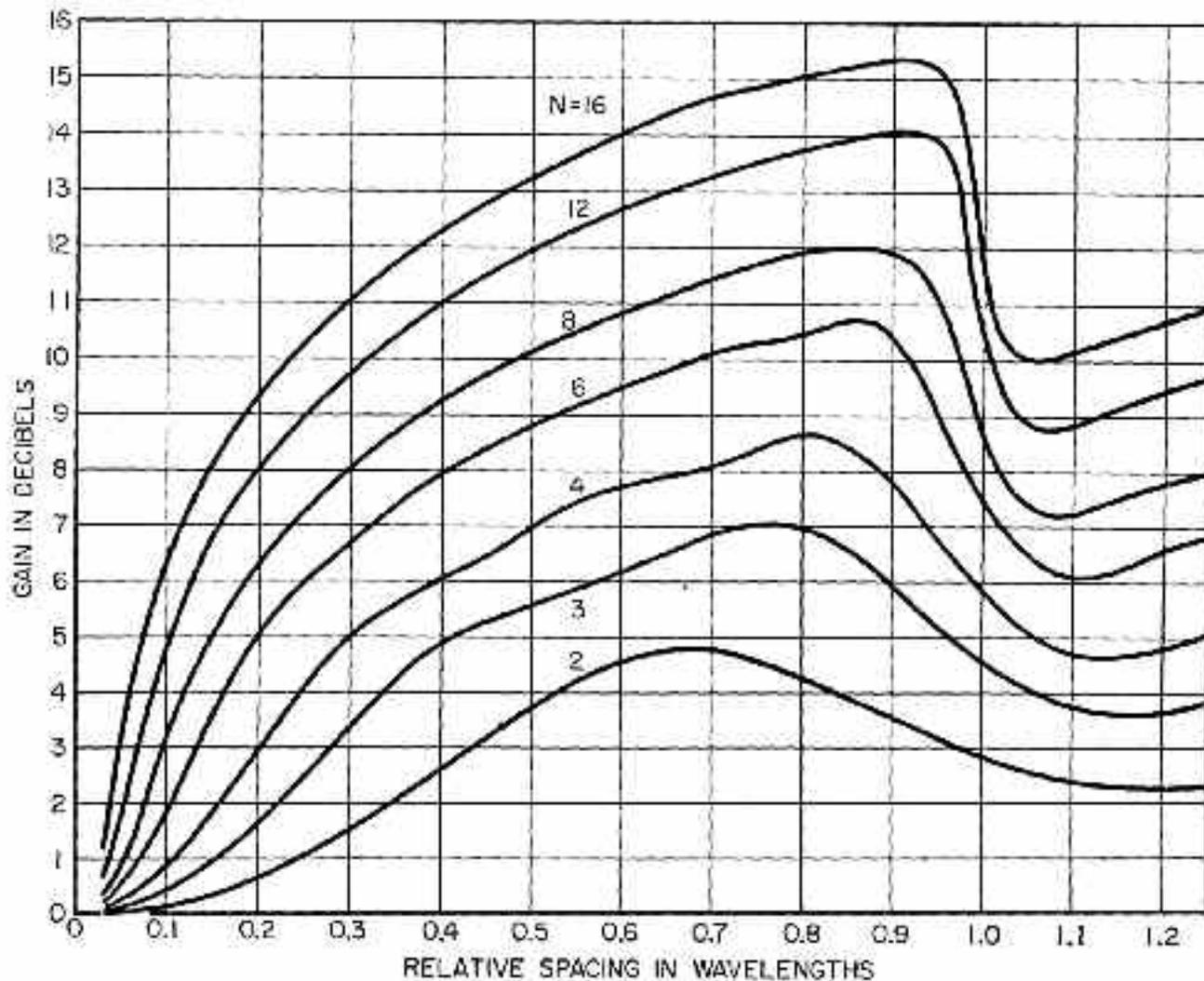
Fɪɢ. 5-10. Gains* to be expected from a single array (no reflector) of broadside elements.

You can see maximum gain occurs at .675wl stacking height. The stacking gain is 4.8db, not 3dB as we see claimed claim. Again the more elements, the wider the stack spacing must be between elements.

I hope these graphs help dispel the myth that doubling element numbers doubles gain! It just doesn't work that way.

A Practical Application?

Combination Broadside/Collinear/ End-fire

If all goes as planned, a new 3.5-10MHz array will be installed later this year in the SE USA using four 80-meter 1/2wl elements fed with open wire line with optimum spacing in broadside-endfire. This array will be 300ft tall and 400 feet wide. A reflector grid of wires will be spaced 1/8th wl behind the array.

The 80 meter gain is 4.8dB broadside + 3.2dB collinear for a total of 8dBd. A screen reflector adds 3dB gain, for a total of 11dBd gain on 80 meters. This is 11dB over a dipole at the same mean height! With ground reflection gain the total dBi gain is about 18dBi.

On 40 meters, the same array will produce 21dBi (or 14dB gain over a dipole at the same mean height). In contrast, a typical 3-element full size Yagi on 80 meters would have about 7dBd gain. The curtain is roughly like four stacked three-element Yagis on 80, and eight stacked on 40.

Curtain pattern can be slewed in both E and H planes to focus on the target. Such arrays are actually practical, and have been used for years in SW BC work.

W4AN

Trust Fund Information

There have been many information requests for the trust fund for Graeme and Erik, Bill and Dana's sons. The family is overwhelmed by the outpouring of kind words by fellow amateurs. If you would like to contribute to the W4AN Memorial Fund for Bill and Dana's children, the trust fund account information is:

Account number 20005913

W4AN Memorial Account for Graeme and Erik Fisher

North Atlanta National Bank

10500 Old Alabama Road Connector

Alpharetta, GA 30022
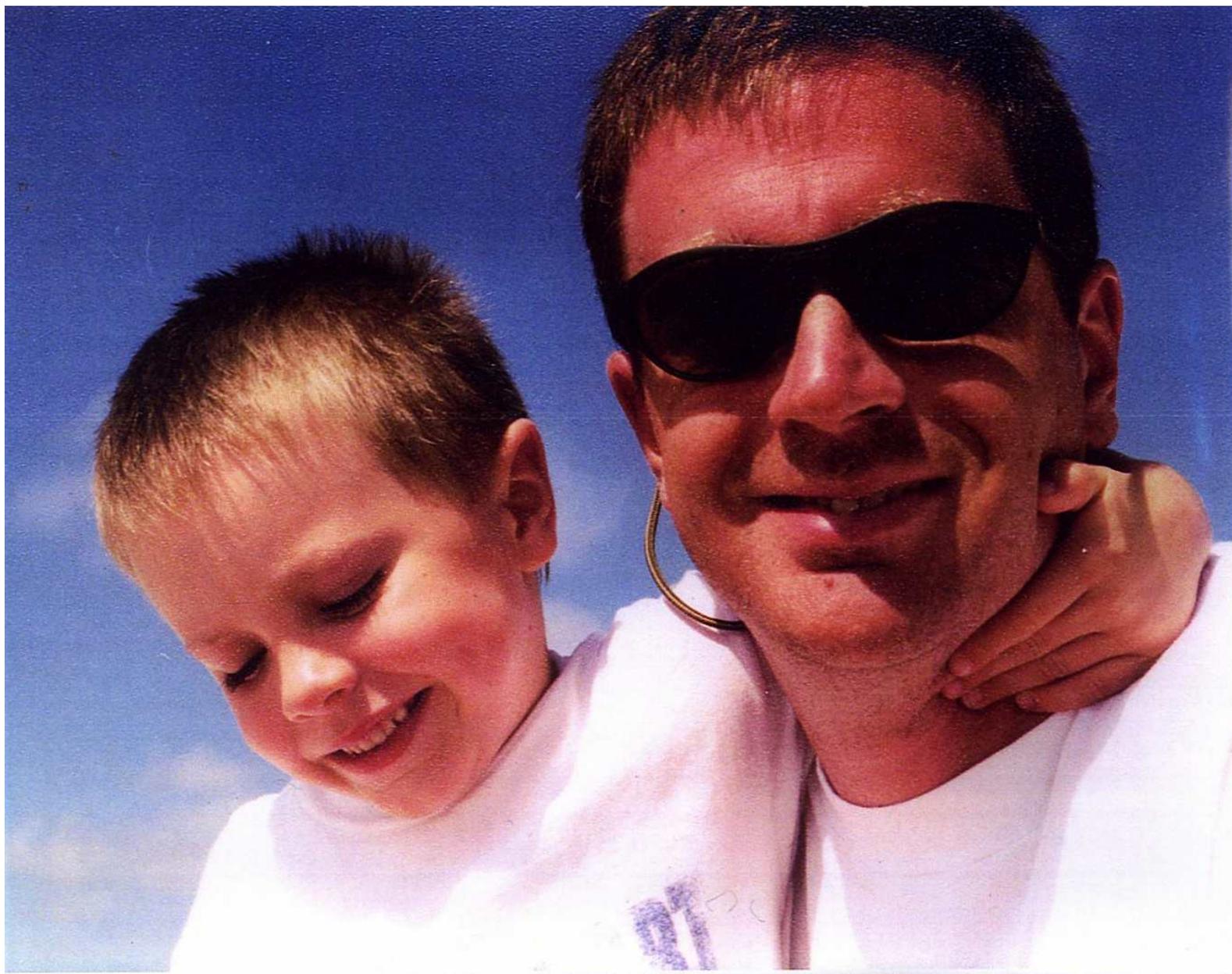
Bank Phone (678-277-8400)

If you wish to make a contribution, any amount will help. Make checks payable to:

W4AN Memorial Account.

You can include the account number "**20005913**" or "**Graeme and Erik Fisher**" in the "for" line.

I will post more information after the Dayton Hamvention rush is over and I recover. Briefly, the basic plan is that Graeme and Erik will be able to draw from this fund for worthwhile activities like school, a first home, or some valid emergency. Having a headache, buying new mag wheels or a dirt bike would not be a good reason. College tuition, books, or housing would be. It will be a 50/50 split unless one doesn't require funds. If neither requires funds, then it will go to some other similar circumstance involving an amateurs young children.

The fund is administered by several well-known members of the amateur community. All of this will be released in final form with more detail a short while after Dayton. Sorry, but anyone who works in the industry knows what a huge time-sink Dayton is!

Graeme and Bill Fisher

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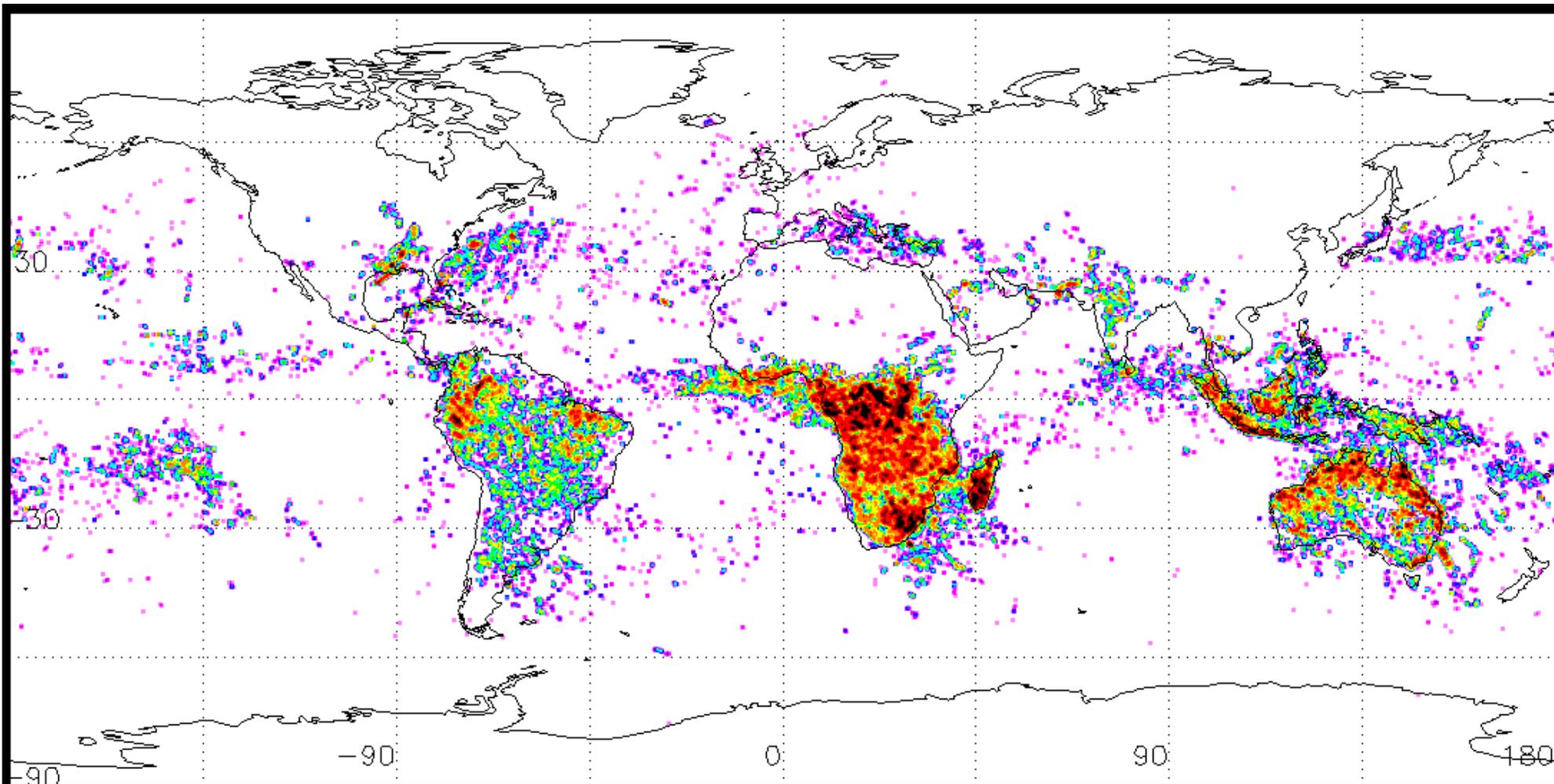
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Areas	27117
Flashes	137810
Groups	683597
Events	1463457
(Created : 02/14/100)	

Flash scale

| 1 | 2 | 3 | 4 | 5 | >5 | >10 | >15 | >25 | >50 | >100 | >150 |
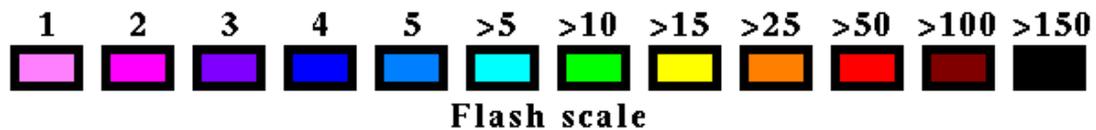
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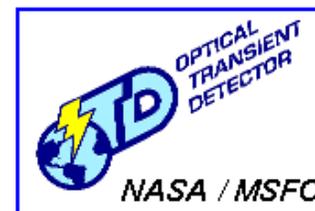
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Receiver Test Data

Device Under Test	Noise Floor	AGC Thrshld	dB	100kHz Blocking	Sensitivity	LO Noise Spacing	kHz	Front End Selectivity	Filter Ultimate	Dynamic Range Wide Spaced	kHz	Dynamic Range Narrow Spaced	kHz
Drake R-4C/CF-600/6	-138 dBm	0.7 uV	3	130dB	0.15 uV	135 dBc	10	A-Preselector	130dB	85dB	20	84 dB	2
AOR AR-7030	-122/-128*	2.2	3	130	0.5/0.22	130	10	D Hi /Lo Pass	90	100	20	82	2
Atlas 350-XL	-131	1.0	11	117	0.2	125	4	C Band Pass	95	81	20	81	2
Kenwood TS-830/YK88	-129	1.5	3	122	0.1	114	2	C Preselector	85^	84	20	81	2
Added 2/27/04 Ten-Tec Omni VI+	-135	0.7	3	145	0.2	137	20	B Band Pass	80	97	20	80	2
Yaesu 901-DM	-135	1.6	3	124	0.15	109	2	C Preselector	85	87	20	80^	3
Collins R-390A	-137	N.A.		130	0.2	130	2	A+ Trk Presel	85	81	20	79	2
Ten-Tec Corsair	-131#	0.1	14	130	0.2	132	5	C Band Pass	90	93	20	79	3

Icom IC-720A	-137	1.6	3	138	0.15	117	10	C 0.5 Octave	80	93	50	78	3
Kenwood TS-820S	-137	0.4	3	115	0.2	125	10	C Preselector	80	79	20	78	3
JRC NRD-515	-138	3.5	4	103	0.1	118	10	C 0.8 Octave	80	95	20	77^	2
Ten-Tec Omni V	-134	1.2	6	135	0.18	134	10	C Band Pass	100	89	20	76	2
Atlas 210/215X	-120#	N.A.		123	0.5	N.A.		C Band Pass	95	76	20	76	2
Drake R-7	-135/-140*	1.0/0.4*	3	145	0.28/0.15*	114	10	B 0.5 Octave	85	97	100	75	2
Drake TR-7	-134	1.3	3	146	0.5	116	10	B 0.5 Octave	90	99	100	75	2
Heath SB-104	-123	N.A.		92	0.5	N.A.		C Band Pass	75	79	20	75	4
WJ HF-1000	-129/-136#*	0.11	3	123	0.23/0.13*	115	10	D Wideband	80	99	20	75	5
Ten-Tec Omni-B	-136	0.2	25	129	0.15	130	10	C Preselector	80	87	20	74	2
Icom IC-730	-140	1.5	3	135	0.1	118	10	C 0.5 Octave	80	92	50	74	3
Kenwood R-820S	-125	4.0	3	125	0.35	123	10	C Preselector	75	74	20	74	4
Collins 75-S3B	-146	1.1	15	122	0.1	120	4	B+ Preselector	85	88	20	74	2
Icom IC-781	-127/-138*	2.0/0.5*	3		0.5/0.18*	129	10	B 0.5 Octave	90^	94	20	73	2
Stock 781	-128/-135	2.4/0.7	3	131	0.5/0.22			B 0.5 Octave	90	98	20	78	2
781 Pin Diodes	-126/-134	3.6/1.15	3	134	0.54/0.21			B 0.5 Octave	90	98	20	72	2
Kenwood TS-930S	-135	2.0	3	143	0.15	115	10	B- 0.5 Octave	80^	86	20	73	3
Icom IC-701	-129	5.5	6	130	0.3	125	10	C Band Pass	75	81	50	73	4
Collins 75S-3C	-141	1.3	12	121	0.14	120	4	B+ Preselector	95	85@	20	72	2
JRC NRD-525	-132#	0.9	3	123	0.2	120	10	B Trk Presel	65	95	50	72	5

Drake R-8	-128/-131#*	0.6/0.3	3	130	0.25/0.18*	115	10	C 0.5 Octave	75^	90/85*	20	71	5
Icom IC-R72	-127/-135#	3.1/1.2*	3	129	0.28/0.11*	122	10	C 0.5 Octave	75	87*	20	71	5
Icom R-9000	-131#	0.8	3	129	0.15	128	10	B 0.5 Octave	90	93	20	71	5
Added 2/27/04 Elecraft K2	-135	2.6	15	118	0.22	123	10	C Band Pass	80^	95	20	70	2
JRC NRD-535	-135#	0.9	3	114	0.1	117	10	B Trk Presel	70	92	50	70	5
Kenwood TS-830S	-136#	0.9	3	122	0.1	113	2	C Preselector	80	84	20	70	3
Icom IC-761	-131/-139*	2.0/0.7*	3	145	0.4/0.17*	129	10	B- 0.5 Octave	90^	87	20	70	2
Lowe HF-150	-126#	0.7	3	126	0.3	113	10	F No Bandpass	75	84	20	69	5
Kenwood TS-430S	-136#	0.6	3	134	0.1	102	10	C 0.5 Octave	70	78	20	69	5
JRC NRD-545	-130#	2.0	6	127	0.2	118	10	B 0.5 Octave	65	96	100	66	5
Signal/One CX-11A	-122#	0.6	17	109	0.6	119	50	C 0.5 Octave	105	90	50	68	5^
Kenwood TS-180S	-139	0.9	3	115	0.15	120	10	C Preselector	80	70	20	68	3
Drake TR-4C	-124#	1.2	3	105	0.4	130	10	C Preselector	80	74	20	68	2
Icom IC-735	-126/-133*	1.5	12	135	0.35/0.18*	123	10	C 0.5 Octave	90^	83	20	68	2
Added 1/21/04 Icom IC-R75	-123#/-130*	3.5/1.3* uV	3	119	0.5/0.2*	109	10	B 0.5 Octave	80	95	50	67	5
Drake SW8	-127#	0.9	3	125	0.32	113	10	N.A.	70	92	20	67	5
Racal 6790 GM	-128	0.3	1	145	0.7	130	10	D Broadband	85	95	20	66	2
Lowe HF-235	-126#	0.8	3	129	0.35	117	10	D Octave	80	71	20	66	5

AOR AR3030	-131#	2.0	10	130	0.16	117	10	C 0.5 Octave	85	90	20	66	5
Yaesu FRG-100	-133#	0.9	3	127	0.13	112	10	C 0.5 Octave	70	99	50	65	5
Kenwood R-5000	-131#	0.4	3	134	0.2	120	10	C 0.5 Octave	80^	86	20	65	5
Added 2/27/04 Palstar R-30	-123	2.6	3	130	0.35	116	10	C Octave	90	88	20	64	5
Yaesu FRG-7700	-130#	1.3	3	123	0.2	100	10	D Octave	65	83	50	64^	5 t
Kenwood R-1000	-130#	0.9	3	119	0.2	107	10	D Octave	70	76	20	64^	3 t
Heath SB-303	-134	N.A.		104	0.5	N.A.		C Preselector	70	66	20	64	4
Collins KWM-380	-127#	1.1	5	123	0.3	99	10	B 0.5 Octave	70^	94	50	64^	2
Icom IC-751	-127/-133*	6.3	3	138	0.4/0.2*	127	10	B- 0.5 Octave	90^	84	20	64	2
Kenwood TS-520	-139	N.A.		116	0.15	N.A.		C Preselector	70	63	20	63	3
Yaesu FT-One	-135	1.0	3	130	0.2	99	10	C 0.5 Octave	80^	91	50	63^	2
Collins 75-S3 Wing	-145	1.0	14	105	0.1	N.A.		B Preselector	75	75	20	63	3
JRC NRD-93	-141	1.6	3	128	0.15	133	10	A+ Trk Presel	80	94	20	63	2
Yaesu FT-980	-136	1.8	3	140	0.12	106	10	C 0.5 Octave	62^	96	50	63	2
Icom IC-R70/R-71A	-129/-135*	3.1/1.4*	3	132	0.4/0.2*	128	10	B- 0.5 Octave	90^	86	20	62	3
Grundig Satellite 700	-127#	1.6	3	106	0.3	118	10	N.A.	85	76	20	62	5
KWZ-30	-130	1.0	3	120	0.8	118	10	D	80	100	20	60	5
Collins 51S1	-134	1.0	7	117	0.13	146	10	A Trk Presel	100	84 <>	100	60	5

Icom R-8500	-135#	0.45	3	132	0.11	131	10	B 0.5 Octave	75	81	20	59	5
Yaesu FT-101E	-141	N.A.		102	0.15	N.A.		C Preselector	70	60	20	59	3
Drake R-4C Stock	-139	0.7	3	130	0.15	135	10	A-Preselector	70	85	20	58	2
Yaesu FT-757	-120/-134*	1.6	3	130	0.7/0.15*	109	10	C 0.5 Octave	70^	86	20	56	3
Ten-Tec 340	-123/-133#*	0.5/0.13	3	109	0.4/0.14	113	10	B 0.5 Octave	70	93	100	46	5
Kenwood R-2000	-130#	1.4	3	115	0.15	105	10	D Octave	70	71	20	45	5
Kenwood R-600	-130#	0.8	3	109	0.2	99	10	D Octave	65	68	20	F.L.	5
Yaesu FRG-8800	-132#	0.6	3	122	0.18	N.A.		D Octave	70	87	20	F.L.	5
AOR 5000	-124/-130#*	0.9/1.8	3	118	0.2/0.35	103	10	B 0.5 Octave	60	58	50	^	5

Notes on Receiver Table Legend:

One (1) dB blocking (gain compression) test done at 100 kHz to eliminate phase noise interaction.

N.A. = Data not available.

= Measured with SSB filter.

***** = Built-in preamp actuated.

F.L = Filter limited, no measurement was possible due to signal leakage around filter.

t = Readings would have been lower if 2-kHz spacing had been possible.

^ = Measurement was phase-noise limited rather than intermodulation limited.

@ = Dynamic range is 90 dB at 100 kHz spacing.

<> = 20kHz dynamic range is 66dB.

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Spectral Analysis of a CW keying pulse

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1 Introduction

This problem was brought to my attention by Tom Rauch, W8JI, who, in his note to me, had described his experience and correctly pointed out all the main features that govern the bandwith. These notes are an expanded version of my reply giving the mathematical explanation.

To fit the most CW signals into the available spectrum, we need to limit the bandwidth taken up by the signals. It is therefore useful to see how the energy in a dot or dash pulse is distributed around the carrier frequency. Here I give some notes on how to make this analysis. The main result is that the spectrum for many keying shapes is given by the product of the spectrum of a square pulse times the spectrum of the slope of the rise and fall behavior of the pulse.

It seems from my experience reading morse, that the rise time should be the main factor in producing code that can be read by ear comfortably. Since the rise time dominates the bandwidth for the usual CW signal, the analysis shows that to get a nearly optimal bandwidth to rise time, the keying pulse shape should have a gaussian slope.

In the next section I review basic Fourier analysis of amplitude modulation. I then calculate the spectrum of a pulse with an exponentially shaped rise and fall as would be produced by simple RC networks. The results suggest the more general analysis in the following section, with the conclusion that a pulse with gaussian slope, i.e. error function rise and fall shapes, will have an optimal bandwidth and rise time.

It seems likely that all of this would have been worked out by radio engineers in the early 1900s when CW signals were first employed.

2 Fourier analysis for amplitude modulation

To analyze the spectrum generated by keying a transmitter let's look at a single ``dot.'' If we imagine we have a

carrier with angular frequency ω_0, and we amplitude modulate it with an envelop A(t), we get the amplitude of the signal from the transmitter is

$$f(t) = A(t)\cos(\omega_0 t) \tag{1}$$

To calculate the energy spectrum we Fourier transform this to get

$$\tilde{f}(\omega) = \int_{-\infty}^{\infty} dt\, A(t)\cos(\omega_0 t)e^{i\omega t} = \frac{1}{2}\left[\tilde{A}(\omega - \omega_0) + \tilde{A}(\omega + \omega_0)\right] \tag{2}$$

where

$$\tilde{A}(\omega) = \int_{-\infty}^{\infty} dt\, A(t)e^{i\omega t}. \tag{3}$$

In the usual case, the modulation $A(t)$ contains frequency components much smaller than the carrier frequency. Therefore the $\tilde{A}(\omega + \omega_0)$ is negligible and can be ignored. The energy spectrum of the amplitude modulated signal is therefore given by

$$P_0(\omega) = \frac{1}{4\pi}|\tilde{A}(\omega - \omega_0)|^2 \tag{4}$$

and since $A(t)$ is real, $\tilde{A}(\omega) = \tilde{A}^*(-\omega)$, and the sidebands are symmetric around the carrier frequency.

It is convenient to write

$$P(\omega) = P_0(\omega + \omega_0) \tag{5}$$

so that ω here is the frequency difference from the carrier, so that $P(\omega)$ gives the energy distribution for an

angular frequency of ω from the carrier angular frequency. I will call $P(\omega)$ the sideband energy density.

3 Application to Keying Bandwidth

To get an explicit result, I'll assume an explicit form for a keying waveform. A simple form where the Fourier transforms can be calculated analytically is the case where the wave builds up exponentially (as in the usual RC circuit) to the carrier value when the key is pressed, and then decays exponetially to zero when the key is released. That is

$$A(t) = \begin{cases} 0 & t < 0 \\ 1 - e^{-t/\tau} & 0 < t < T \\ (1 - e^{-T/\tau})e^{-(t-T)/\tau} & t > T \end{cases} \tag{6}$$

where T is the keying pulse width and τ is its time constant. A plot of this waveform is shown in figure 1 for the cases where τ is 4 milliseconds, and T is 20 and 50 milliseconds.
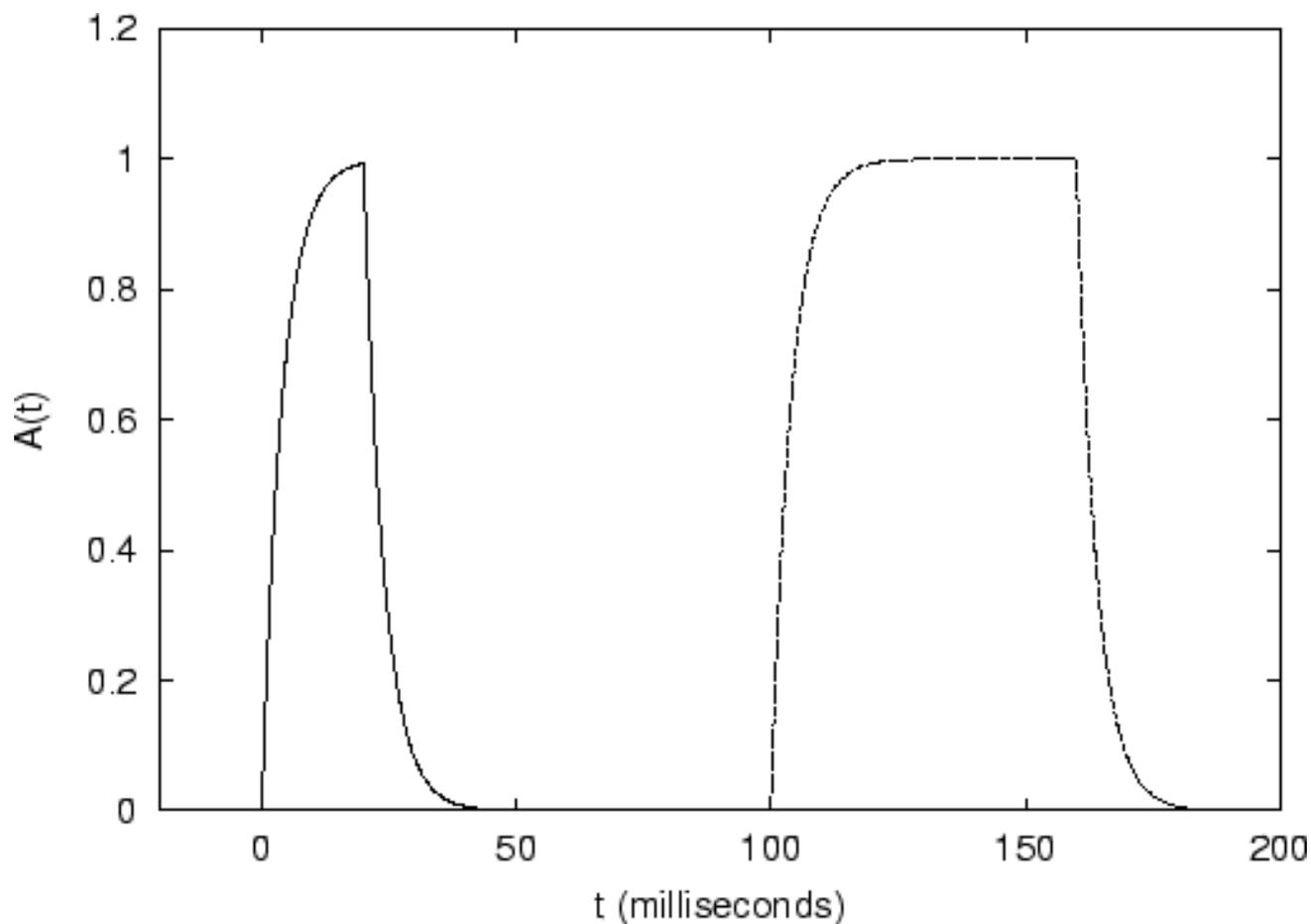
Figure 1: The exponential keying waveforms for a time constant τ of 4 milliseconds and durations of 20 and 50 milliseconds. The 50 millisecond pulse begins at $t = 100$ milliseconds to separate it from the 20 millisecond pulse.

The fourier transform integral of the amplitude is straightforward and gives

$$\tilde{A}(\omega) = - \left[e^{i\omega T} - 1 \right] \frac{1}{i\omega \left(i\omega\tau - 1 \right)} \tag{7}$$

so that the sideband energy density) becomes

$$P(\omega) = \frac{1}{\pi} \frac{\sin^2 \left(\frac{\omega T}{2} \right)}{\omega^2} \frac{1}{\left(1 + \omega^2 \tau^2 \right)} . \tag{8}$$

So the sideband energy density has two factors. If we measure it in dB relative to some fixed value, we add the logarithms of the factors. The only dependence on T comes from the first factor. The $\sin^2$ function is always less than or equal to 1, so this will subtract from the other factor which only depends on τ the time constant which determines the rise time.

The main features of the sidebands will therefore be given by the rise time, while the length of the pulse will modify those features somewhat.

Figures 2 and 3 show the energy density in dB referenced to the carrier energy density for T of 20 milliseconds. The curves are plotted together in figure 4. Each curve is plotted versus frequency $f = \omega/2\pi$.
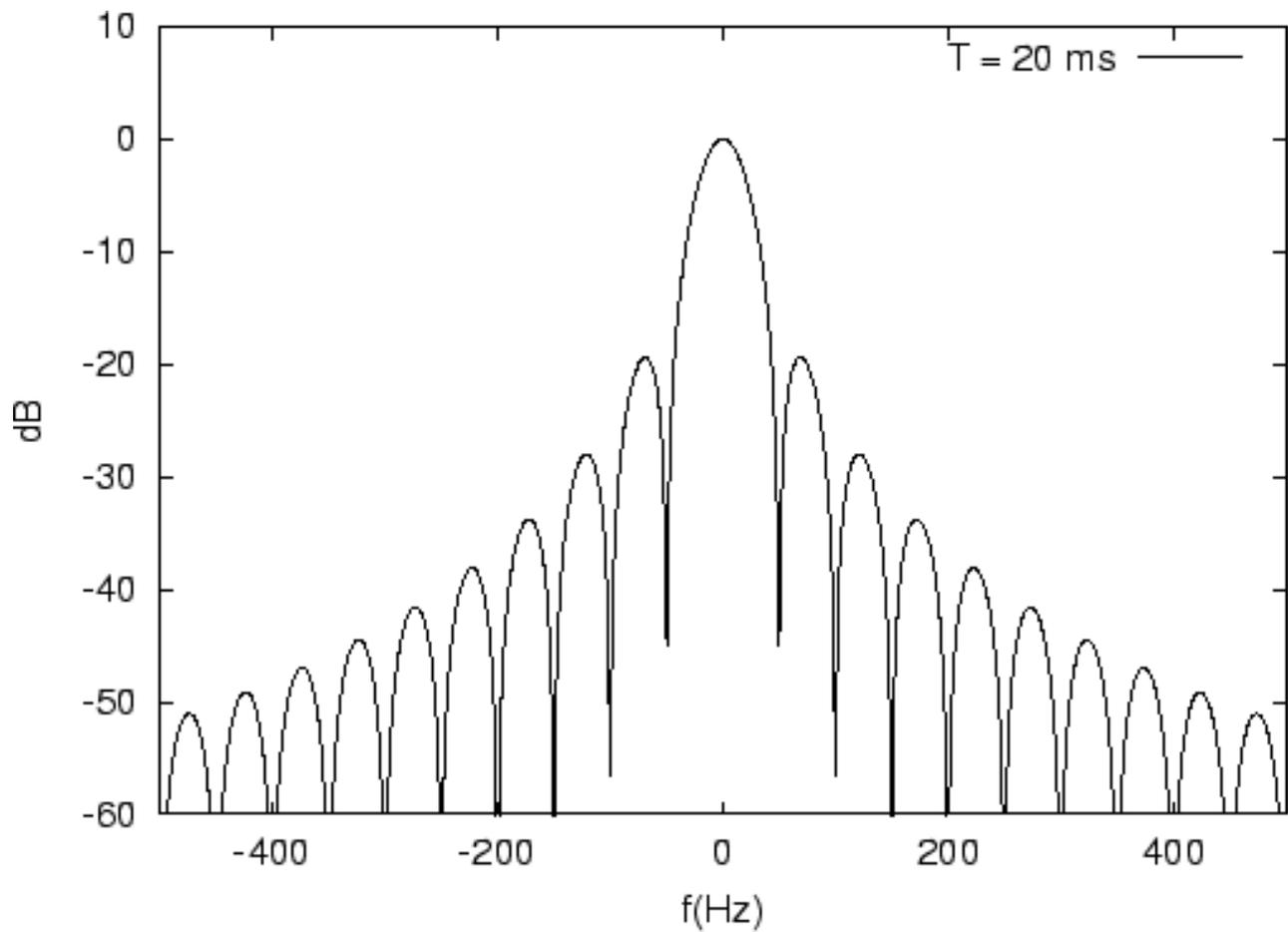
Figure 2: The sideband energy density for an exponential keying wave form with $\tau = 4$ milliseconds and T of 20 milliseconds. The carrier energy density is set to 0 dB.
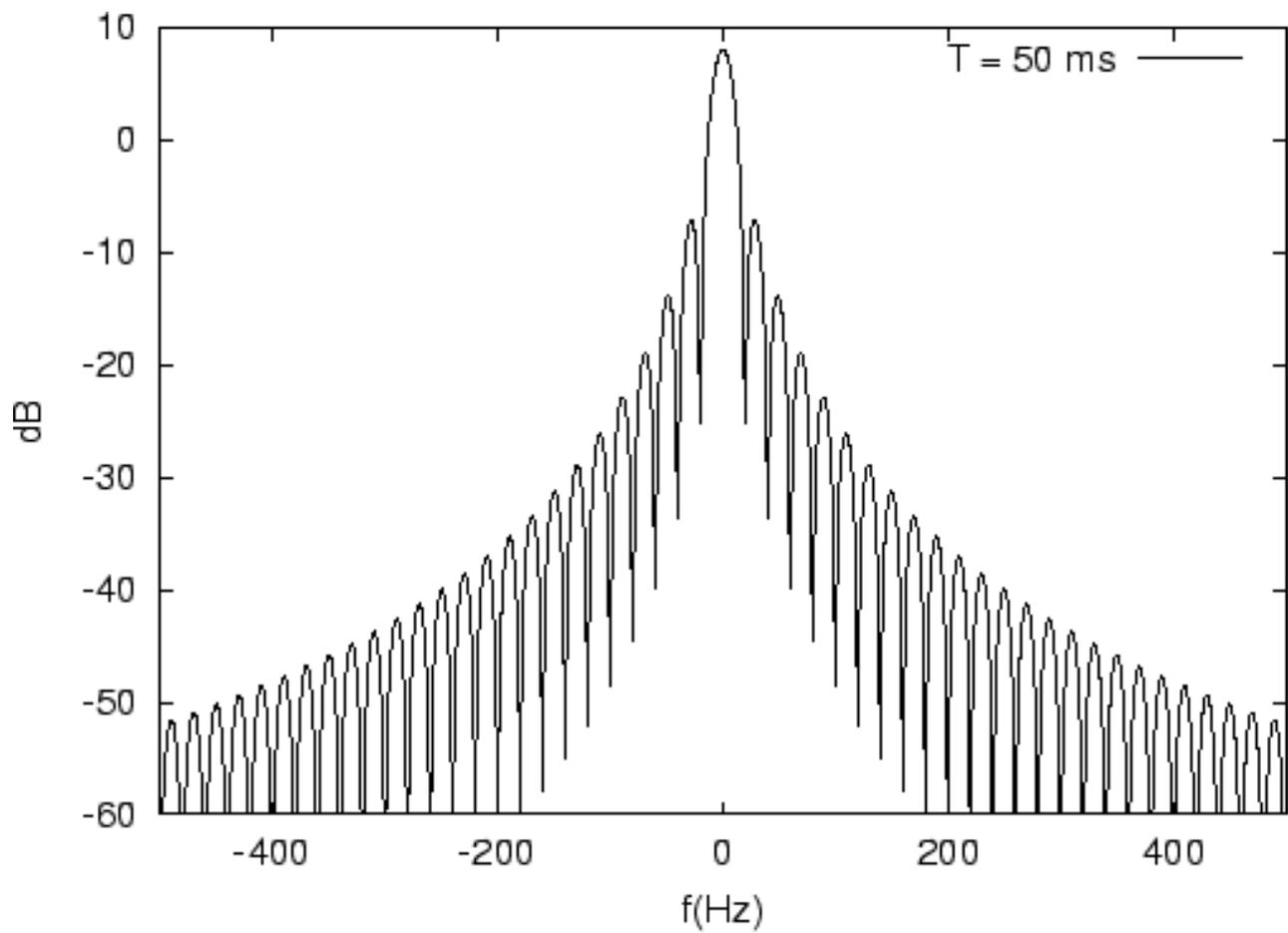
Figure 3: The sideband energy density for an exponential keying wave form with $\tau = 4$ milliseconds and T of 50 milliseconds. The scale is the same as figure 2.
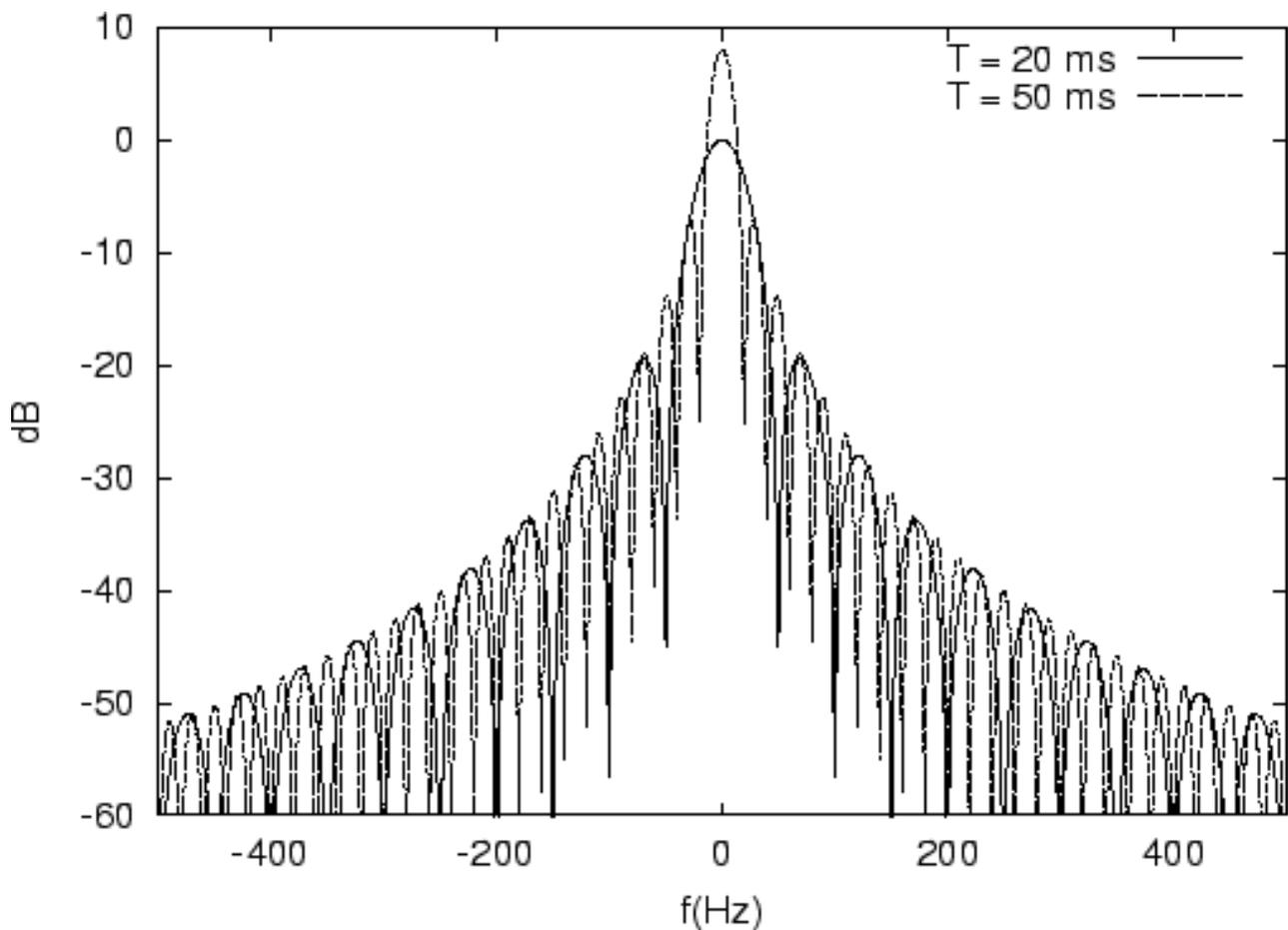
Figure 4: The plots of figures [2] and [3] combined.

Notice that the differences between the 20 millisecond and 50 millisecond pulses are first the energy near the carrier frequency is larger for the longer pulse as needed since it has about 2.5 times as much energy, and second the ``ringing'' has more oscillations for the longer pulse as expected. The sidebands fall off 12 dB per octave once we are at frequencies beyond about $1/\tau$.

The effect of the keying speed on the bandwidth as long as the rise time is small compared to pulse length is the change in shape of the central peak. It does get narrower for slower keying and wider for faster keying, however, the keying speed does not effect the overall bandwidth.

4 General Pulse Shape

I can write a general pulse shape as

$$A(t) = E_r(t + T/2) - E_f(t - T/2) \qquad (9)$$

where E_r and E_f describe the rising and falling edges of a pulse and are positive functions that go to zero at

large negative t and to 1 at large positive t. The Fourier transform of A is the sum of the transforms of the two terms. If the rising and falling edges have the same form, we can write

$$A(t) = E(t+T/2)-E(t-T/2) = \int_{-\infty}^{\infty} dt'[\delta(t+T/2-t')-\delta(t-T/2-t')]E(t') \quad (10)$$

and integrating by parts gives the result

$$A(t) = -\int_{-\infty}^{\infty} dt' S(t-t')\frac{dE(t')}{dt'} \equiv -\int_{-\infty}^{\infty} dt' S(t-t')E'(t') \quad (11)$$

where $S(t)$ is a square pulse of width T,

$$S(t) = \begin{cases} 1 & |t| < \frac{T}{2} \\ 0 & |t| > \frac{T}{2} \end{cases} \quad (12)$$

and I have defined E' to be the derivative of E.

Since $A(t)$ is written as a convolution, its Fourier transform is now the product

$$\tilde{A}(\omega) = \tilde{S}(\omega)\tilde{E}'(\omega). \quad (13)$$

For keying wave forms this has the nice interpretation that the spectrum is given by the spectrum of a square pulse of length T multiplied by a factor that is the Fourier transform of the slope of the rise and fall wave form.

Except for an unimportant change of the zero of time, the exponential case calculated above can be written as

$$E(t) = \begin{cases} 0 & t < 0 \\ 1 - e^{-t/\tau} & t > 0 \end{cases} \quad (14)$$

The Fourier transforms of these are

$$\tilde{S}(\omega) \quad = \quad \frac{2\sin\left(\frac{\omega T}{2}\right)}{\omega}$$

$$\tilde{E}'(\omega) \quad = \quad -\frac{i}{1+i\omega\tau} \tag{15}$$

and the sideband energy density is exactly as before.

As we saw for the exponential case, the bandwidth is dominated by the rise and fall time. Therefore it seems reasonable to try to optimize the rise and fall waveform of the keying pulse. In terms of the function $E'(t)$ whose integral is the rising and falling wave form, we want to simultaneously make its width in real time and in frequency small. One measure of this is the product of $\Delta\omega$ and Δt where they are defined as the variances in frequency and time

$$(\Delta t)^2 \;=\; \int_{-\infty}^{\infty} dt\, t^2 E'(t) - \left[\int_{-\infty}^{\infty} dt\, t E'(t)\right]^2$$

$$(\Delta\omega)^2 \;=\; \int_{-\infty}^{\infty} d\omega\, \omega^2 \tilde{E}'(\omega) - \left[\int_{-\infty}^{\infty} d\omega\, \omega \tilde{E}'(\omega)\right]^2 . \tag{16}$$

This problem is well known in optics and quantum mechanics where it goes by the name of the minimum uncertainty wave-packet[1] . The solution is a gaussian

$$E'(t) = \frac{1}{\sqrt{\pi\tau^2}} e^{-\frac{t^2}{\tau^2}} . \tag{17}$$

The Fourier transform of this gaussian is

$$\tilde{E}'(\omega) = e^{-\frac{\omega^2\tau^2}{4.}} \tag{18}$$

and the keying wave form with this $g(t)$ has $E(t) = 1/2[1 + \operatorname{erf}(t/\tau)]$

$$A(t) = \frac{1}{2}\left[\text{erf}\left(\frac{t+T/2}{\tau}\right) - \text{erf}\left(\frac{t-T/2}{\tau}\right)\right] \tag{19}$$

where $\text{erf}(x)$ is the error function[2] defined to be

$$\text{erf}(x) = \frac{2}{\sqrt{\pi}}\int_0^x du\, e^{-u^2}. \tag{20}$$

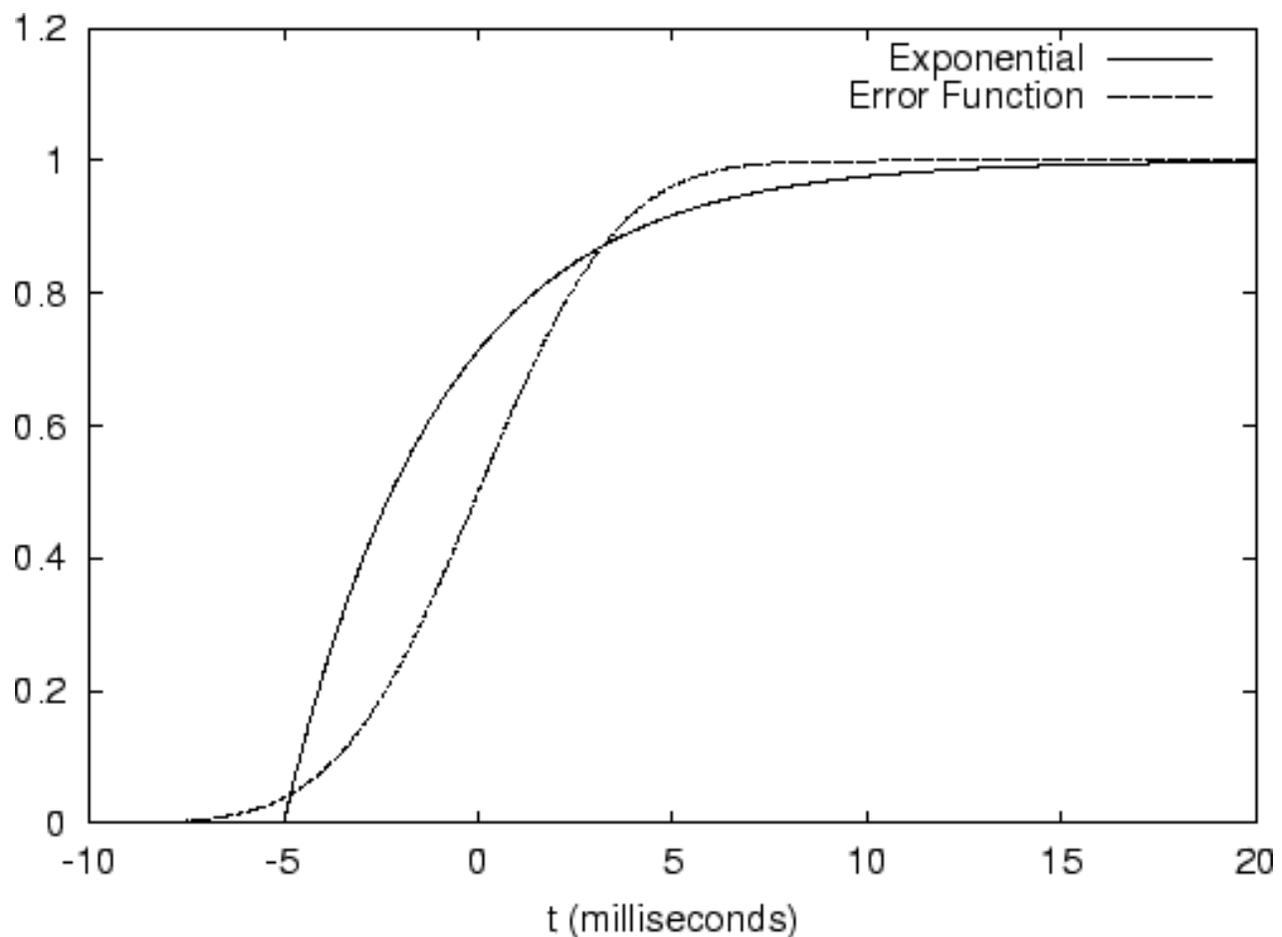The differences in the keying wave form are shown by the turn on shape of the keying pulses in figure 5.

Figure 5: A comparison of the error function turn on with the exponential turn on, both for their respective τ values of 4 milliseconds.

In figure 6 I show the keying wave form for $\tau = 4$ milliseconds and $T = 20$ and 50 milliseconds as in figure 1 for both the exponential and optimized wave form. Notice that the abrupt changes in the exponential form are absent from the error function form.
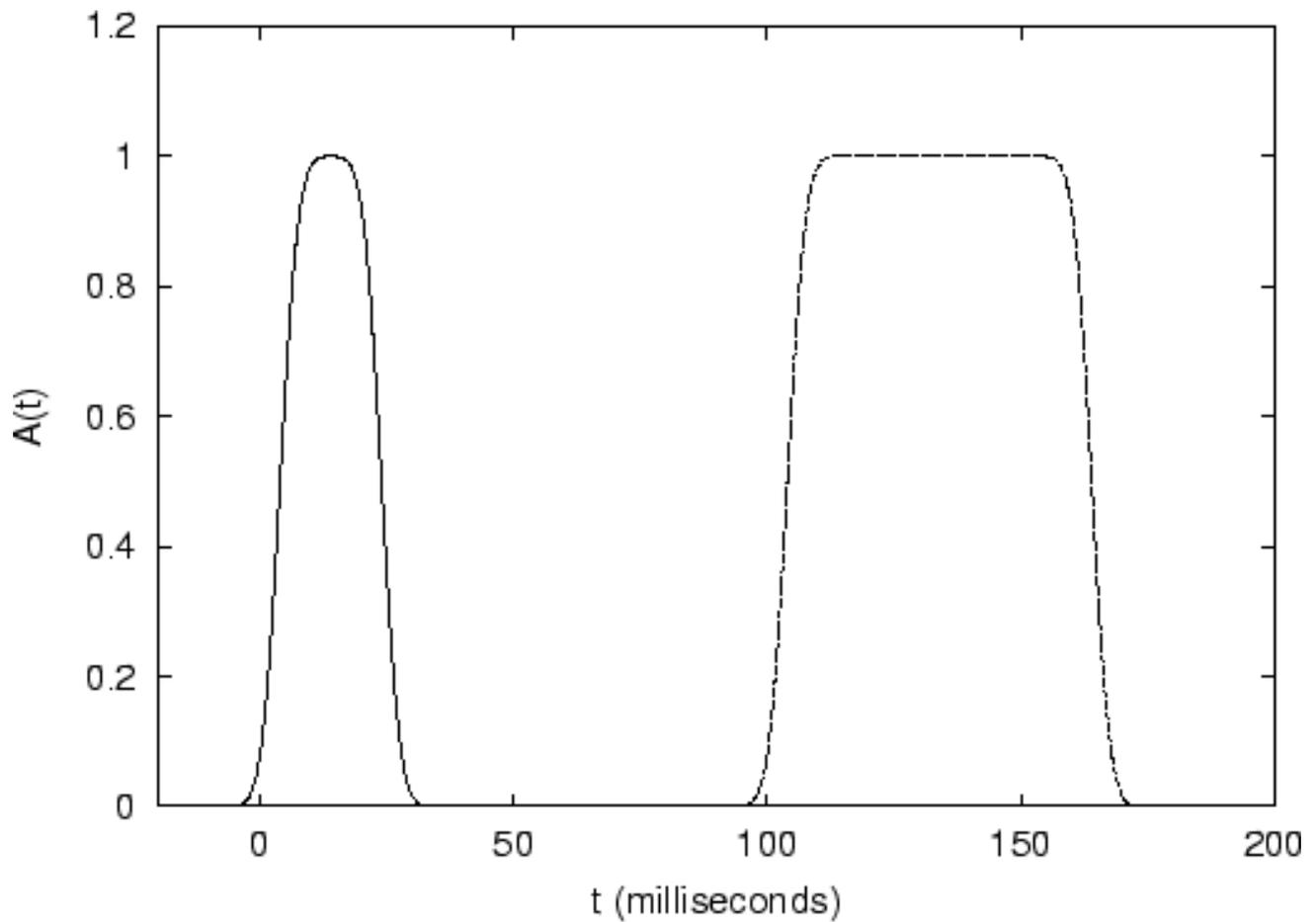
Figure 6: The error function keying waveforms for τ of 4 milliseconds and durations of 20 and 50 milliseconds. The 50 millisecond pulse begins at $t = 100$ milliseconds to separate it from the 20 millisecond pulse.

Figures [7] and [8] show the sideband energy density.

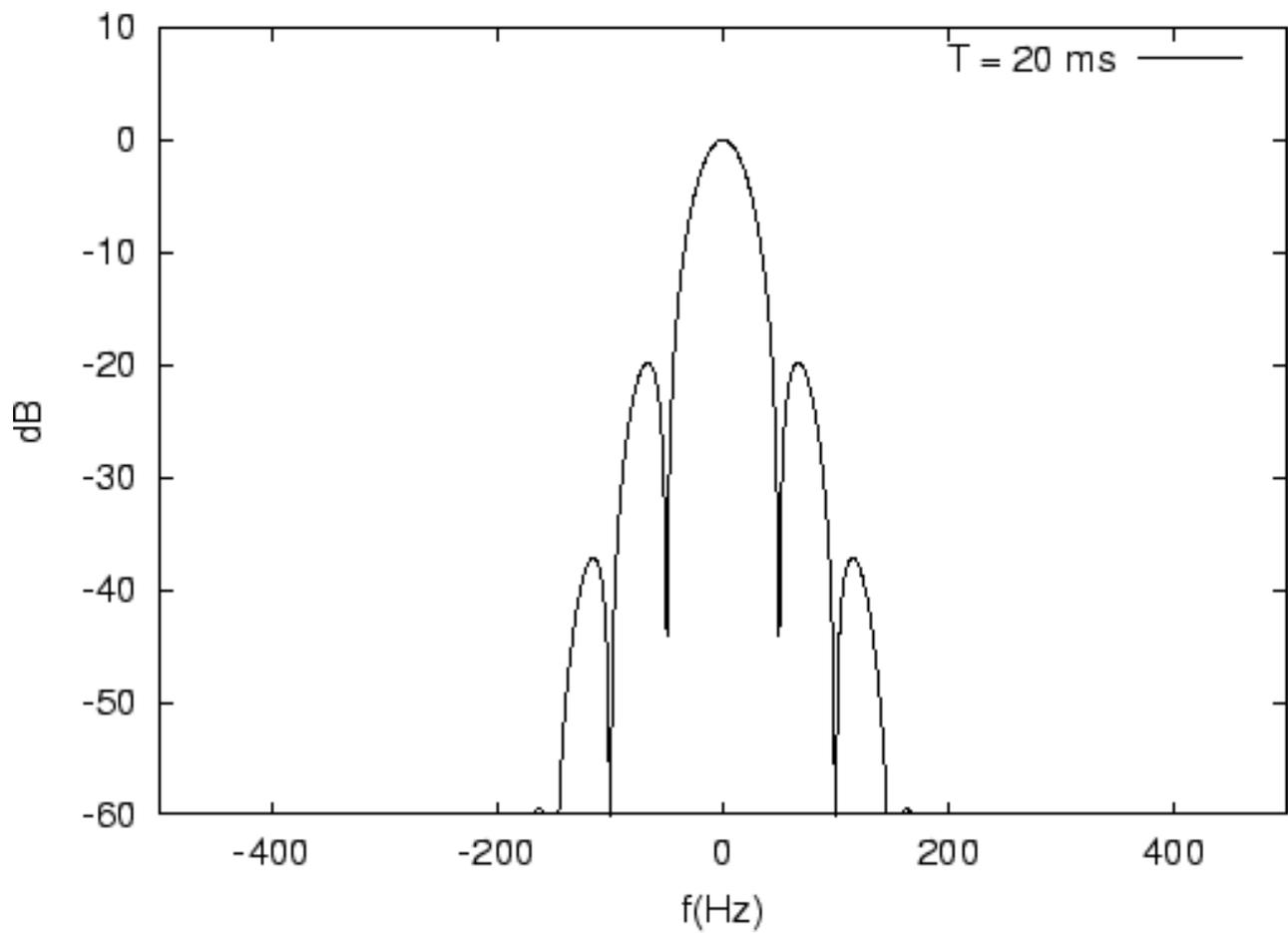**Figure 7:**The sideband energy density for an error function keying wave form with $\tau = 4$ milliseconds and T of 20 milliseconds. The carrier energy density is set to 0 dB.
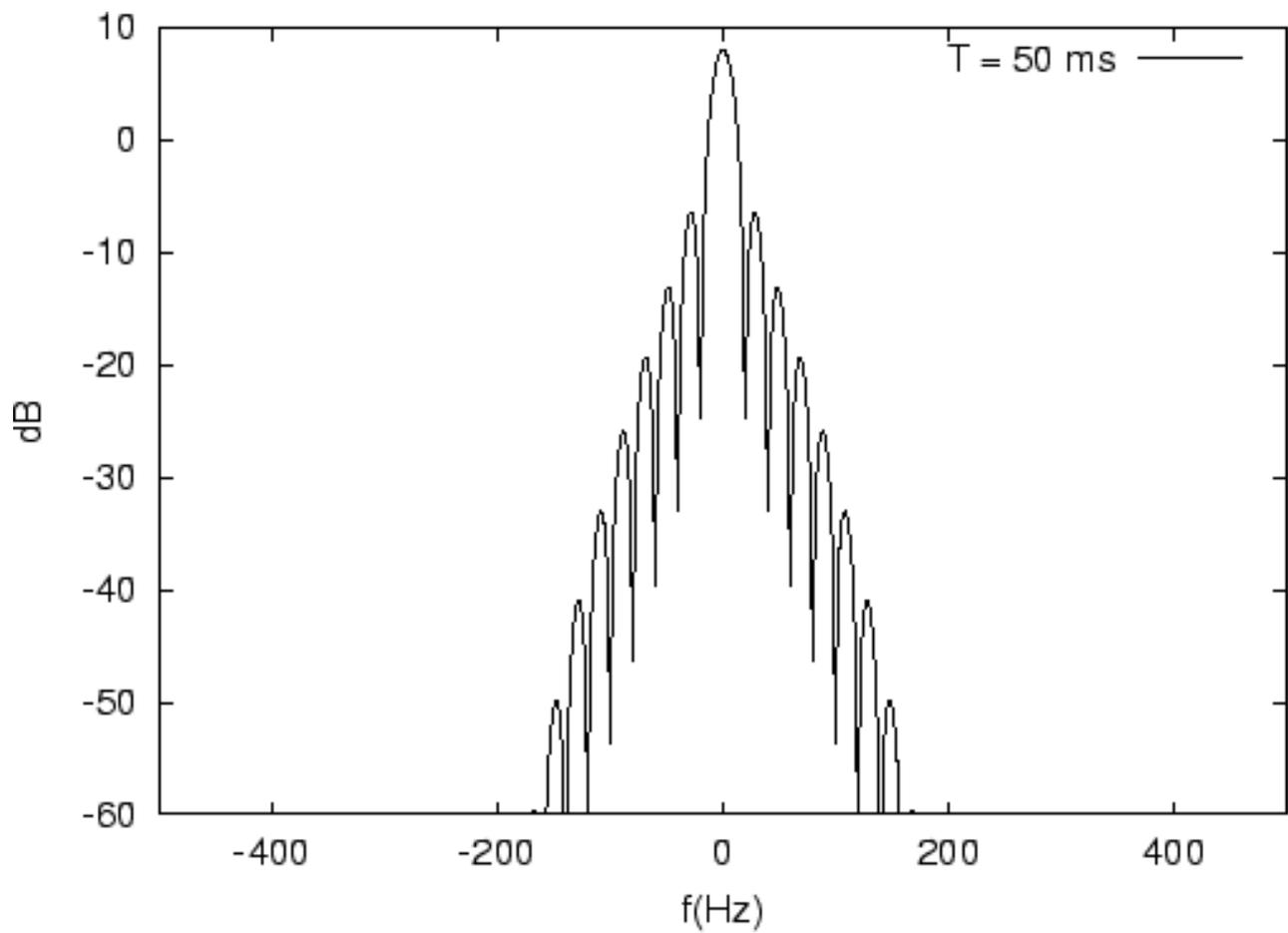
Figure 8: The sideband energy density for an error function keying wave form with $\tau = 4$ milliseconds and T of 50 milliseconds. The scale is the same as figure [7](#).
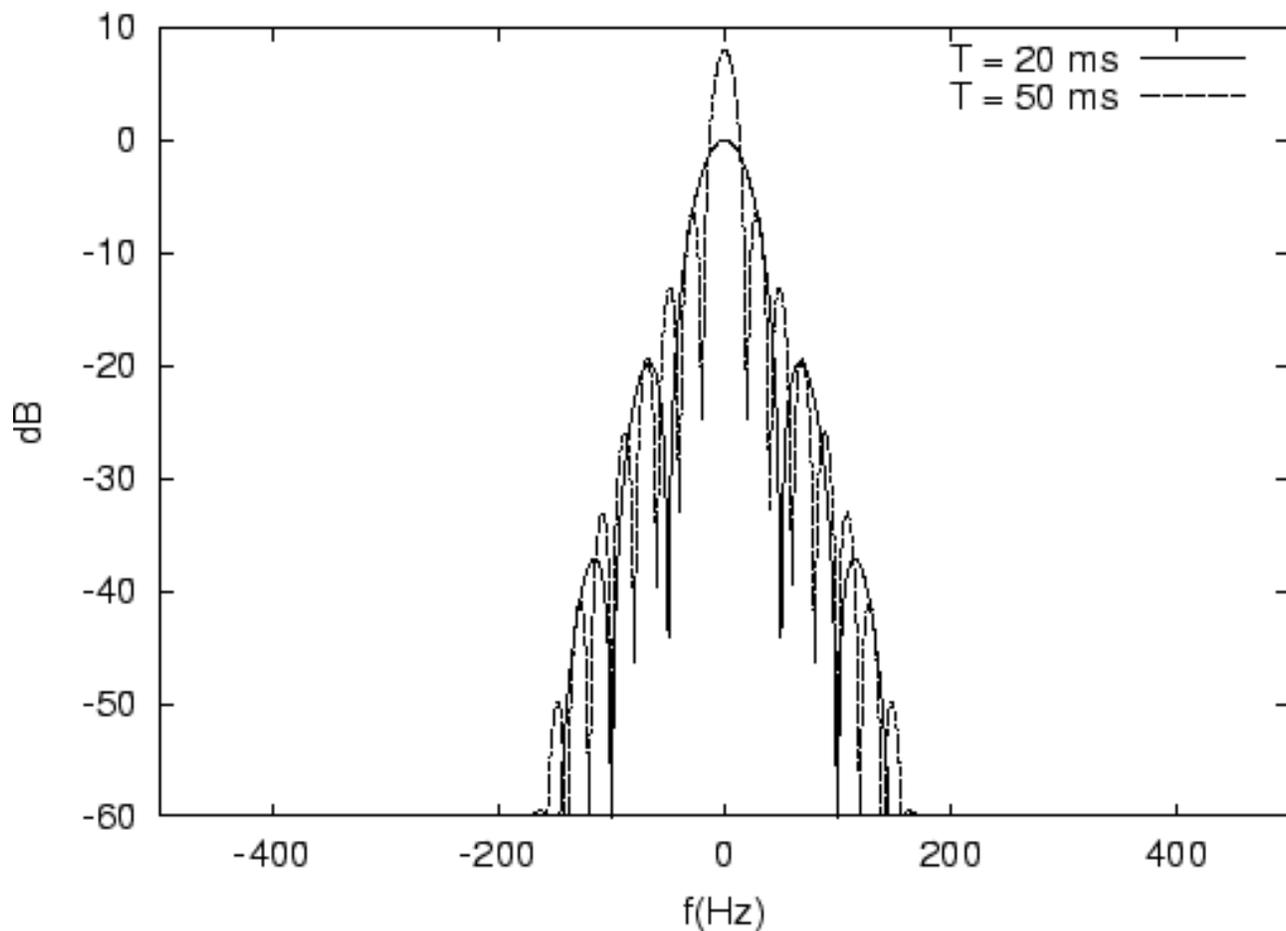
Figure 9: The plots of figures [7] and [8] combined.

5 Extensions and Conclusions

The sideband energy density for many pulse shapes factorizes. The first of the two factors is proportional to the Fourier transform squared of the square pulse and the second by the Fourier transform squared of the slope of the rise and fall. The analysis can be easily generalized to an arbitrary sequence of pulses. The Fourier transform of the single square pulse simply needs to be changed to the Fourier transform of the sequence of square pulses. For a given rise time, the error function shape for the rise and fall will attenuate unnecessary interference away from the carrier frequency much better than exponential keying.

Bibliography

1

      E. Merzbacher, *Quantum Mechanics*, 3rd Ed. (John Wiley and Sons, New York, 1998).

2

      M. Abramowitz and I. Stegun, *Handbook of Mathematical Functions* (National Bureau of Standards, New York, 1964).

Spectral Analysis of a CW keying pulse

Up: [W9CF Links](#)

About key clicks

At International Radio we get a lot of questions about the FT-1000MP series key clicks and the effect of the Inrad mod. There is a lot of data out there, some of it taken with spectrum analyzers and some taken with second radios. There is no standard for comparison in the spectrum analyzer data, so it is difficult to draw any conclusions from it. Data taken with a second radio is very subjective at best and is essentially meaningless for comparison purposes.

We have taken data on unmodified FT-1000MPs (original version) and also on radios modified with the Inrad circuitry. There have been no changes made in the keying circuitry in the Mark V or the Field versions. The data was taken on the same spectrum analyzer, with the same settings and with the analyzer settings clearly shown on the plots. In order to compare the MP performance with some other radios, we also are showing data taken under the same conditions, with the analyzer set the same way, etc. on several other radios. We will post data on our site on any one radio of a type which we can get for testing.

Here is the technical explanation for the measurements. The radios are being keyed at 40 WPM by an external keyer. The spectrum generated by a CW transmitter is similar to amplitude modulation where the modulating frequency is the keying rate. Sidebands are produced on both sides of the carrier. If the modulating source were a perfect sinewave, there would be only one sideband on each side of the carrier. However, the modulating source is nearly a square wave and is very rich in harmonics. So the sidebands created are spaced by the modulating frequency and have an envelope which is a function of the waveshape of the keying source. Note that the envelope does not change for different keying speeds. The number of sidebands within the envelope change as the keying speed is changed, but this does not alter the bandwidth of the composite signal.

The really important thing here is the shape of the keying signal.

If we pick a certain offset from the carrier and take a dB down reading, we can have a figure of merit for comparison purposes. One kHz is a good place to make this measurement as the sideband level is generally way out of the noise

and is easy to read from the plots. These numbers are posted next to the analyzer plots.

Several things are readily apparent when viewing the plots. First, the Inrad modification makes a substantial reduction in the FT-1000MP spectrum. Also, the modified MP is quite good when compared with the other radios. As a matter of fact, it's narrower than some radios for which popular opinion claims good keying.

The "bad name" which the MP has developed in recent times probably has something to do with the number of radios in the field. In a major contest there are more MPs in use than any other radio. Imagine, for example, if everyone in the last 160 meter contest was using an Omni VI+ or a K2 driving a KW amplifier. The popular press would then be saying that something has to be done about those radios. In truth, something needs to be done with most of the currently popular radios.

The Inrad mod

What is it and how does it work? The mod consists of 3 small components, two go on the IF board and one goes on the RF board. On the IF board, a 0.047 capacitor and a 100 K resistor in parallel are placed across C2148. On the RF board, a 0.1 uF cap is placed across C1004 (Mark V and Field) or C1216 in the original MP. (If you don't have a service manual, Louis, W7DZN has posted pictures on the VA3CR web site.)

In each case, these capacitors are used to slow the waveform which is used to key the gate voltage on FETs. These stages are doing the CW keying. The resistor on the IF board is required to keep the duty cycle of the keyed waveform at 50% by rebiasing the gate of the FET.

Some people have asked about routing the RF through the 500 Hz filter instead of the SSB filter. It seems like a good idea, however, the keying is done beyond the filters and this change would have no effect on the keyed waveform. Note that this is done in the Omni VI+ and the keying is distorted by the ALC loop. Another way to improve the keying would be to use a gausian or a transitional filter to shape the keying waveform. In a perfectly linear system this would work very well. However, the FETs being keyed are far from linear. Thus it is not possible to improve the output spectrum beyond what the Inrad mod does without substantial changes in the radio.
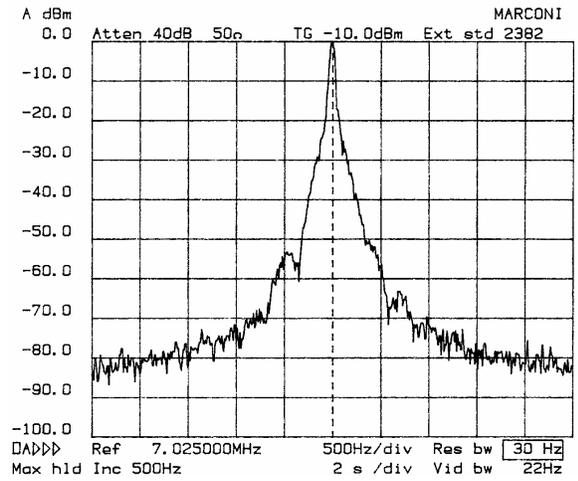
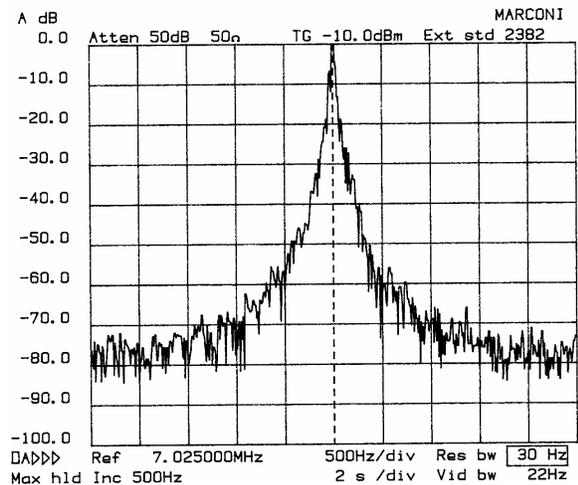*(continued)*

Unmodified MP

Level at +/-1 kHz= -62 dB
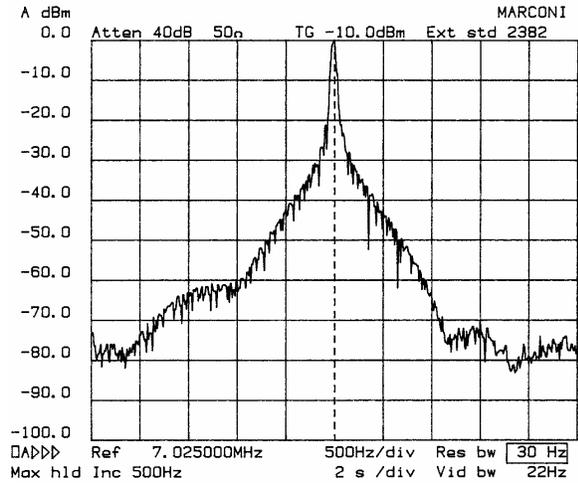
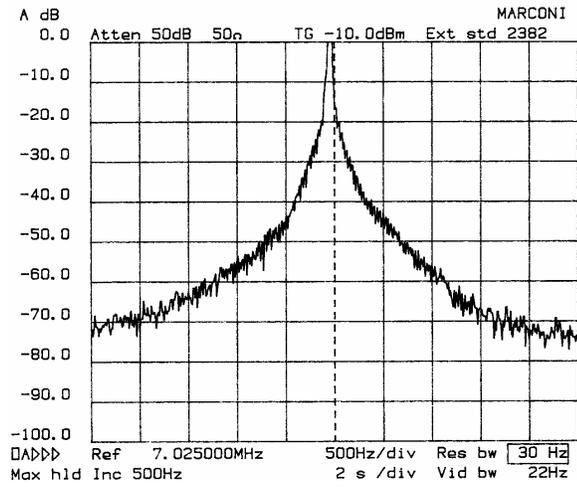**Modified MP-1**

Level at +/-1 kHz= -73 dB

Modified MP-2

Level at +/-1 kHz= -72 dB

IC-765

Level at +/-1 kHz= -58 dB

A dB
0.0 Atten 50dB 50Ω   TG -10.0dBm Ext std 2382   MARCONI
-10.0
-20.0
-30.0
-40.0
-50.0
-60.0
-70.0
-80.0
-90.0
-100.0
⊡Δ▷▷▷   Ref   7.025000MHz   500Hz/div   Res bw   30 Hz
Max hld Inc 500Hz   2 s /div   Vid bw   22Hz

K2

Level at +/-1 kHz= -54 dB

A dB
0.0 Atten 50dB 50Ω   TG -10.0dBm Ext std 2382   MARCONI
-10.0
-20.0
-30.0
-40.0
-50.0
-60.0
-70.0
-80.0
-90.0
-100.0
⊡Δ▷▷▷   Ref   7.025000MHz   500Hz/div   Res bw   30 Hz
Max hld Inc 500Hz   2 s /div   Vid bw   22Hz
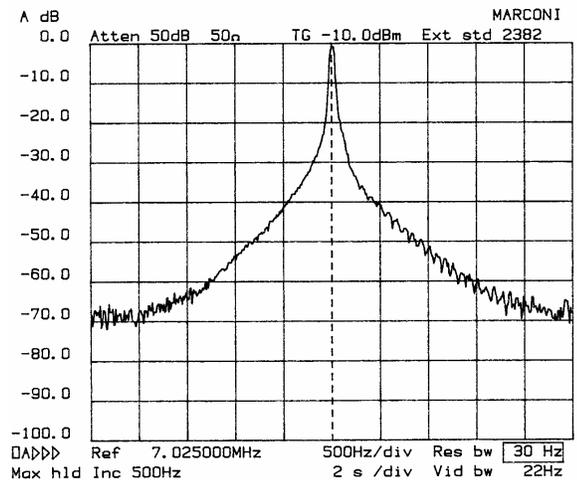
Omni VI+

Level at +/-1 kHz= -56 dB
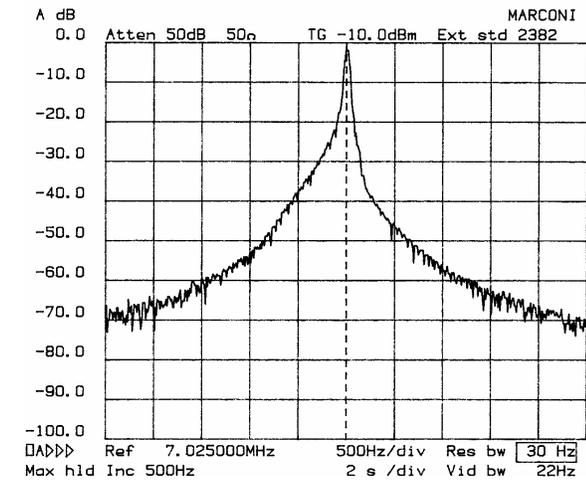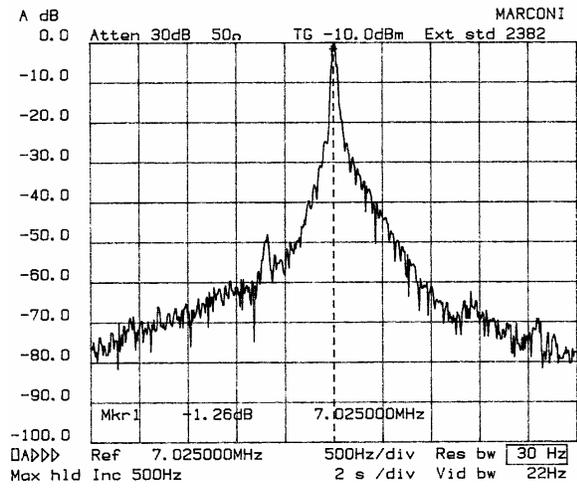
A dB
0.0 Atten 50dB 50Ω   TG -10.0dBm Ext std 2382   MARCONI
-10.0
-20.0
-30.0
-40.0
-50.0
-60.0
-70.0
-80.0
-90.0
-100.0
⊡Δ▷▷▷   Ref   7.025000MHz   500Hz/div   Res bw   30 Hz
Max hld Inc 500Hz   2 s /div   Vid bw   22Hz

FT-817

Level at +/-1 kHz= -61 dB

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Beverage Antenna Products

This equipment provides the best possible receiving system performance of any design! This is the same equipment design proven to make top scores in 160m contests year after year. Fully isolated feeds prevent common mode noise and unwanted signals. This equipment has excellent reliability and is housed in metal cases for improved life and shielding. Units use RF tight type F connectors for superior reliability and weather characteristics.

W8JI Beverage Antenna Construction Article

Reversable Beverage System

The RBS-1 is a two-direction, reversable Beverage array designed and used by W8JI. It consists of the reversable feed point matching system (RBS-1FP) and a reflection transformer (RBS-1RT).   The unit can switch directions of reception by having a 10-18 Vdc voltage applied through the coaxial feed. The antenna between the two units is standard 450 Ohm ladder line (not included). Operating range is a broad 0.2-30 MHz. Fully isolated grounds are used to prevent common-mode noise and unwanted signals.

Specifications:
Power Requirement: 9-10 Vac or 12-15 Vdc
Switching Current: less than 80 mA
Antenna Impedance: 300-600 Ohms (450 Ohms recommended)
Feedline Impedance: 50-75 Ohms (75 Ohm recommended)
Reflection Transformer Ratio: 1:1

Metal housings are used for shielding and improved life.

View Assembly Instructions for DXE-RBS-1P (314K pdf)

Right Click, "Save Target As"

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Part #	Price	Add to Cart
DXE-RBS-1P	$129.00	

Receiver Preamplifier, 0.3-35 MHz

This is the best HF low-noise amplifier available. The RPA-1 is optimized for 0.3-35 MHz operating range. It's push-pull amplifier design and robust components enable it to with stand high signal levels and operate when you need it most.

Gain: 16 dB, 0.3-35 MHz (+1.5-0 dB over this range)
Input Third Order Intercept: 23 dBm
Noise Figure: 3.5 dB
One dB Compression: +26 dBm (~0.4 W output)
500 Hz BW IM3 Dynamic Range: 110 dB or greater
Power Requirement: 10-18 Vdc @ 140 mA maximum

Suitable for indoor or outdoor installation with the option of being powered through the coaxial feed. Metal housing provide shielding and improve life. RCA and Type F connectors are used on the input and output of the RPA-1. The unit comes with a relay that automatically bypasses the amplifier when dc power is removed.

Part #	Price	Add to Cart
DXE-RPA-1	$109.00	

Beverage Feed System

This Beverage Feed System is designed and used by W8JI. This receive only antenna pulls signals out of the noise that most antennas don't even pick up. The BFS-1 has fully isolated feeds to prevent common mode noise. The transformer used in this receive only antenna handles 25 watts of power so it won't go into saturation even with the strongest of signals.

Specifications:
Operating Frequency: 0.1-30 MHz
Antenna Impedance: 400-500 Ohms
Feedline Impedance: 30-100 Ohms (75 Ohms recommended)
Common Mode Isolation: 20-30 dB (antenna to feedline)
Power Handling: 20 watts continuously or 200 watts for 5 seconds
Termination Resistance: 400-500 Ohms

Utilizes stainless steel hardware and metal housing for long life. Utilizes type F connectors for superior reliability and weather characteristics.

View Assembly Instructions for DXE-BFS-1(316K pdf)
Right Click, "Save Target As"

Part #	Price	Add to Cart
DXE-BFS-1	$35.00	

470-Ohm Non-Reactive 2 Watt Resistor
Pack of 10 Resistors

This non-inductive resistor can absorb extreme amounts of power for short periods without damage. It has the highest surge immunity of any 2-watt leaded (with leads) carbon or metal resistor. Metal-film and carbon-film resistors easily fail from even minor electrical disturbances. When used as a Beverage termination, the DXE resistor withstands nearby lightning strikes significantly better than hard-to-find carbon composition resistors. It is superior to carbon composition resistors in real life, retaining resistance value despite exposure to heat and power surges.

Applications:

- Beverage Antenna Terminations
- Parasitic Suppressors
- Dummy Loads

Sold Only as a Pack of 10 resistors

Part #	Price	Add to Cart
DXE-ECM-R470-2	$20.00	

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Customer Showcase

Anna Hill
KG8YL
MI
USA

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Downloadable Files

NOTE: I place no restrictions on the use of any of these files except that they may not be sold. I do not warrant the programs in any way, and no support is available for them.

All downloadable files with .EXE extensions are self-extracting ZIP files. After downloading a file, go to the directory containing the file (first starting DOS if you're using a Windows system). Then type the name of the file. (You don't need to type the '.EXE'.) This will cause the file to expand into two or more other files. One of these will be a documentation file such as READ.ME or INFO.TXT containing instructions. Windows users can generally read these files with Notepad. Files with extension .ZIP must be un-ZIPped with PKWare's PKUNZIP program, widely available as shareware. Again, a documentation file will appear when the file is un-ZIPped.

The "Field Day Special" Antenna

The Field Day Special is a two-element wire beam with a unidirectional pattern. It's similar to the "ZL" special, except that its elements are the same length. This makes it easy to reverse the pattern by using two feedlines, one at the center of each element. The original antenna was described in June 1984 QST. An antenna like this can't directly be scaled for other bands unless the size of the twinlead it's made from is scaled in proportion. However, using the program available here, you can design your own for the frequency band of interest. By changing the various dimensions, you can trade gain/beamwidth, feedpoint impedance, and bandwidth. Improvements in the program have been made since the article was published, so an optimal design for 20 meters will have dimensions slightly different than the ones shown in the article. The currents in the elements of one design were measured and found to agree very closely with those predicted by the program. Quite a number of these have been built by various people for several bands, with universal reports of operation as predicted. I use a 20 meter version nearly every year for Field Day, and occasionally 15 and 40 meter versions as well. 20 and 15 meter Field Day Specials have been my primary home QTH antennas for these bands for a number of years. Download **FDSP~.EXE**, then follow the directions under "Downloadable Files", above. More information about the antenna, including construction and feeding, is included in one of the expanded files. The program is a DOS program and can be run from the Windows 95/98/NT DOS prompt.

Current Baluns

Baluns are a continuing source of discussion on the Internet newsgroups. In response to some questions about "current" or "choke" baluns, I wrote an analysis of their operation. In the past, I had believed that placing the balun at the input of an unbalanced tuner would allow the balun to work properly when the output of the tuner sees a high impedance load. The analysis shows this to be incorrect. A brief analysis is given in **IBALBRF.TXT**, and a more detailed analysis in **IBALUN.TXT**. Both are plain ASCII text files and can be viewed with Notepad or a similar application.

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